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RCA-NBC TELEVISION PRESENTS A POLITICAL CONVENTION AS FIRST LONG DISTANCE PICK-UP

By

O. B. HANSON

Vice President and Chief Engineer, National Broadcasting Company, Inc.

Summary—*During the week of June 24th, 1940, the National Broadcasting Company's television field pick-up equipment was installed in Convention Hall in Philadelphia for the televising of the Republican National Convention of 1940. The signals were transmitted from the NBC's convention control room via the Bell System's Philadelphia-New York coaxial cable to NBC's main control room in the RCA Building, Radio City, New York, and thence to Station W2XBS, the NBC television transmitter atop the Empire State tower. It was estimated that approximately 40,000 people viewed by television the proceedings in the Convention Hall on some 4,000 receivers scattered throughout the vicinity of New York in homes and restaurants, etc. The signal from Station W2XBS was intercepted by the General Electric Company by means of a specially constructed receiving system on Helderberg Mountain and rebroadcast through the General Electric Company's television transmitter W2XB to the television audience in that vicinity. The following article tells the story of the televising by NBC of this national event.*

THE televising of the Republican National Convention proceedings at Convention Hall, Philadelphia, June 25th to 28th inclusive, marked the first time that network facilities for television were put to a practical operating test in this country. Scenes of Convention floor activities were picked up by NBC's television cameras and transmitted by coaxial cable from Philadelphia to New York, a distance of 104½ miles, before being radiated from the NBC television transmitter atop the Empire State Building in New York City.

The choice of Philadelphia by the Republican National Committee as the city in which their 1940 National Convention was to be held gave NBC the opportunity of televising a news event of national importance, but had this event been staged in any other city exclusive of New York metropolitan district, it could not have been brought to the screens of the New York television viewers.

During the past 15 months many news events occurring within twenty-five miles of Radio City have been picked up by the NBC telemobile unit, relayed to Radio City and broadcast from Station W2XBS atop the Empire State tower, and viewed in the homes of some 4000 television set owners. It was NBC's desire to extend the range of pick-up to more distant points, but facilities for transmission of television signals from city to city are not yet in existence, with the one exception of an experimental coaxial cable installed by the Bell System, between Philadelphia and New York. This cable, while primarily designed for the study of multiple carrier telephone transmission, has such electrical characteristics that it would permit the transmission of television image signals.

The engineers of the Bell System were equally anxious to test out their coaxial cable as a television transmission medium and willingly spent several months preparing the cable, its repeaters and terminal equipment to tie in to the NBC television equipment. With such a facility available, NBC made the necessary arrangements with the Republican Convention Committee to enable the tens of thousands of viewers in the New York area to witness the convention scenes occurring a hundred miles away in Philadelphia.

As one might expect, many new problems arose in the contemplation of this first network television pick-up. The coaxial cable terminated in the Bourse Building in downtown Philadelphia, 3½ miles short of the Convention Hall. In New York its terminal equipment was located at 32 Sixth Avenue, 4 miles short of Radio City. At first it was thought that a radio link would have to be provided at the Convention Hall to transmit the signal to the Bourse Building terminal of the coaxial cable, and another radio link to complete the circuit from 32 Sixth Avenue to Radio City. However, by careful selection of telephone circuits, the installation of wide band repeaters at intervals along these circuits provided satisfactory extension of the coaxial transmission circuit direct from the Convention Hall to the NBC terminal room in the RCA Building at Radio City. Numerous test transmissions were conducted over this circuit during the weeks preceding the Convention.

Within the Convention Hall itself, NBC was faced with a series of interlocking problems that required solution. Among these were the providing of suitable electric power for the operation of the telemobile unit, the proper location of cameras, of the large mobile unit itself, of adequate and properly directed lighting on points of interest, the installation of cables and many intercommunicating circuits, etc. As at all conventions, radio and television do not have the field to themselves. The newsreel cameramen and newspaper photographers have their special problems, and our lighting and operating requirements had to be coordinated with theirs. Lighting particularly required considerable coordination in order that television might have a continued source of adequate illumination on the speakers' stand and at other points of interest. The many other engineering details to be worked out before we were ready to transmit a program were taken up point by point as the Republican Committee's plans for the event unfolded.

The equipment available for field use consisted of Telemobile Units 1-A and 1-B (Figure 1), and the newly developed transportable equipment (Figure 2). Mobile Unit 1-B contains a 400-watt television transmitter, W2XBT, licensed for use in the frequency band 156-162 megacycles. Inasmuch as wire facilities were to be provided between Convention Hall and the Philadelphia terminus of the New York coaxial

Fig. 1—Telemobile Unit 1-A and 1-B operating at RCA Building, New York City.

cable, a distance of 3½ miles, this unit was eliminated from consideration. Had there not been time to provide and thoroughly test the specially engineered telephone link to the coaxial cable terminal, this unit might have been used as the local radio link.

Telemobile Unit 1-A contains the video and sound equipment necessary for the operation of the two television cameras associated with it. The unit is equipped with two complete camera chains plus a synchronizing generator and many other pieces of auxiliary equipment. One camera chain is associated with a standard Iconoscope, mosaic size 4.81 x 3.62 inches. The other chain is used with an experimental Orthicon tube, mosaic size 2.33 x 1.74 inches. The characteristics of the Orthicon tube as distinguished from the standard Iconoscope will not be gone into here as they have been covered in an article appearing in the October 1939 issue of the RCA REVIEW.

This equipment is mounted on nine racks extending down the center line of a twenty-six foot, ten ton custom built vehicle. The cameras may not be located further than 250 feet from the mobile unit. This restriction was imposed by the length of camera cable normally used in routine program operation in the New York City area. The cable is 1⅞ inches in diameter and weighs 1¼ pounds per foot, and includes four coaxial lines and 28 other normal conductors.

Fig. 2—Transportable equipment operating at Convention Hall. NBC video engineer R. W. Clark at controls.

The transportable equipment consists of a synchronizing generator and two camera chains designed for the small Iconoscope, mosaic size 3 x 2.75 inches. This equipment has been designed in eleven separate units, each of which are approximately the size of a medium suitcase. One thousand feet of camera cable, broken down into fifty, one hundred, and two hundred foot lengths is available. By combining transportable equipment with the telemobile unit equipment, four cameras were thus made available.

Program plans called for television coverage on as wide a scope as was consistent with equipment and personnel limitations. Obviously, the speakers' platform with a small area surrounding it was the focal point of interest. Secondary to this point was the arena floor where 1001 delegates and their alternates were seated. The possibility of presenting prominent citizens and nationally known news commentators as well as the events in the main hall to the television audience was recognized as an important phase of televising the Convention. Such complete coverage called for the construction of a suitable studio adjacent to the main hall. In addition it was deemed essential to be able to show arriving dignitaries, crowd scenes, and traffic on the highway fronting the building. Such a pick-up could be achieved by locating a television camera on the sidewalk at the most used entrance to the hall. An inventory of equipment and personnel available indicated that such program plans could be carried out.

The most important consideration for the ultimate success of this venture was to secure illumination at the speakers' rostrum which would insure a picture being obtained by both the Orthicon camera with long focal length lens, and the Iconoscope camera with a faster, wide angle lens.

Lighting at this Convention had to be considered from the newsreel standpoint as well as that of television. As distinguished from newsreel operation, television operation is a continuous performance and thus the lighting for television coverage had to be continuous. It is the standard practice of newsreel cameramen to take only excerpts of even the most important speeches. Armed with an advance script, they "shoot" only those portions in which they are interested. For such operation it is not necessary that continuous high level lighting be maintained.

Sufficient lighting for television was obtained by suspending ten 5 kw solar spot lamps from the ceiling, and locating two on each balcony. These overhead lamps were mounted on a frame which was approximately forty feet above and thirty feet in front of the speakers' stand. The light from these lamps was directed towards the stand at an angle of approximately fifty degrees from horizontal. This source of light provided an illumination of 800 footcandles, incident at the speakers' platform.

Eight 150-ampere d.c. arc lamps were installed at points on the balcony so that they could be directed at any spot on the floor. These lamps illuminated the state delegations and alternates. At times they were directed on the speakers' stand when the newsreels wished extra illumination in that vicinity. When they were not so used, television's threshold of continuous illumination was maintained by the fixed lights. When the center of interest was transferred from the rostrum to the floor, such as during the balloting for candidates, the television cameras followed the beams of the arc lamps. When not in use for newsreel work, the latter were maintained for television's use on the floor of the hall.

The wisdom of obtaining a threshold of illumination directed on the speakers' rostrum was borne out on the fourth day of the Convention, for the first three days the weather had been ideal. The temperature outside the hall had maintained itself at a comfortable sixty-eight to seventy-eight degrees. In the hall, the temperature was

Fig. 3—NBC television cameras focussed on speakers' platform.

higher, but the delegates had not found the heat from the d.c. arcs too uncomfortable. On the fourth day, the outside temperature rose to a point in the eighties. The delegates were seated in such a manner that some states' delegations comprising 60 to 70 delegates had been supplied with only fifty or so chairs. The rest doubled up. The delegates, preserving decorum, sat with their coats on. The result was inevitable. After the first ballot had been taken, the delegates petitioned the Chairman to order the arc lamps extinguished. The Chairman recognized the plight of the delegates and acted promptly. The arc lights were extinguished and the illumination dropped to less than threshold value and thus there were to be no more scenes for television showing the arguments of the delegates as they massed under their state banners. The speakers' stand remained adequately illuminated, however, and television was able to carry through its planned coverage for the one and one-half remaining days.

The speakers' rostrum was situated eight feet above the main floor level and on its center line in the long dimension. A balcony eleven feet above the main floor ran around three sides of the arena. The eight by five foot television camera platform (Figures 3 and 4) was suspended off the south balcony, at a point sixty feet from the

Fig. 4—Camera platform in main hall. Orthicon camera is in foreground.

speakers' stand. From this position, the television cameras faced the speaker at an angle of forty-five degrees. The newsreel camera platform, a much larger structure, was located along the same balcony at a point fifteen feet further from the speaker than the television platform. The position of the television camera platform was chosen as a compromise between the angle of the view and the distance between the speaker and the camera. Had the stand been advanced along the balcony another twenty feet toward the stage the distance from the speaker would have been reduced by approximately fifteen feet, but the result would have been almost a profile view. The forty-five degree angle position was retained as presenting the most desirable viewing angle possible. The fifteen-foot spacing between the television stand and the newsreel stand was arrived at as the point beyond which the television stand would obscure the view of the newsreel man occupying the innermost position on his stand, when he was using a thirty-degree lens.

The standard Iconoscope camera and the Orthicon camera were set side by side on the NBC camera platform. A 19¾-inch focal length, f.4.5 treated lens was used on the Orthicon camera. The angle of view included a space approximately three feet either side of the speaker.

Fig. 5—Television Studio, Convention Hall. Announcer Ray Forrest (left) interviewing Mr. A. H. Morton, NBC Vice President in charge of Television.

When viewed on a twelve-inch Iconoscope, a bust view of the speaker occupied five inches vertically. The Orthicon was also used to pick up the floor scenes during the balloting and demonstration parades after a candidate had been placed in nomination. The Iconoscope camera was equipped with an eight-inch focal length, f.2 treated lens. The wide angle of this lens included almost the entire width of the stage when focussed on the speakers' stand. When directed on the floor, this lens permitted a view of more than one-half of the state delegations. The sensitivity of the Orthicon enabled it to be used satisfactorily in picking up scenes having lower light levels than that provided on the speakers' stand. The lenses on both cameras had been surface treated with magnesium fluoride. Such processing, by reducing reflection losses, gave an increased lens efficiency of approximately forty per cent.

The telemobile unit containing the video equipment associated with the main hall Iconoscope and Orthicon cameras was parked on a freight loading ramp inside the Convention Hall and one floor below the main arena. It remained parked there for the duration of the Convention. The electrical constants of the NBC mobile unit equipment require that

Fig. 6—Street position at main entrance to Convention Hall. Announcer (left) interviewing Mayor Lamberton of Philadelphia.

the camera cables be limited to 250 feet in length. Fortunately, it was possible to so locate the cameras and mobile unit that this length was not exceeded.

A number of problems entered into the choosing of a location for a television studio. It was desirable that the following factors be satisfied: Accessibility to staircases and elevators; accessibility to speakers' platform, state delegations and other points of activity; size, shape, and general technical adaptability; satisfactory conditions for ventilation and air conditioning to overcome the heat generated by 15 kw of portable lighting units; proximity to adequate source of power supply for lighting, air conditioning units, etc. A dressing room on the third floor was chosen as our studio location wherein television interviews were to take place (Figure 5). The wardrobe room associated with the dressing room was chosen as the control room where the transportable equipment could be used to best advantage (Figure 2). This room was windowless and therefore was easily darkened, a desirable feature for viewing pictures on a Kinescope.

For the televising of scenes on the street in front of Convention Hall, a camera location was set up on the sidewalk in front of the main entrance to the building (Figure 6). From this vantage point,

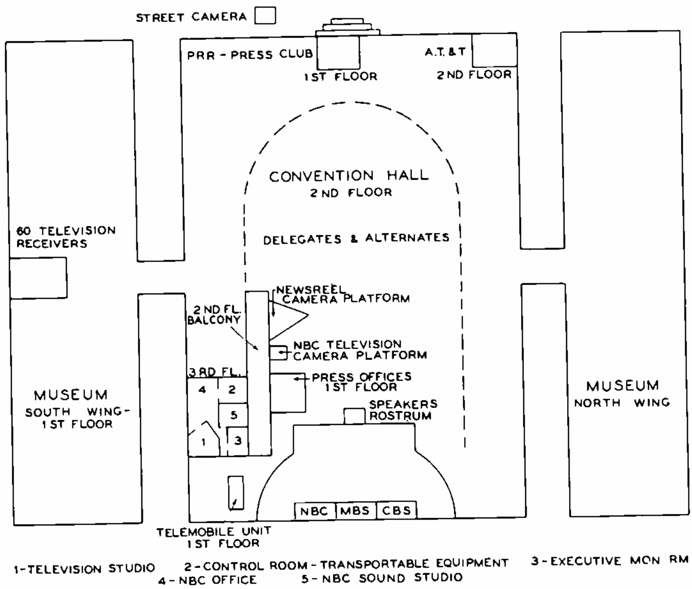


Fig. 7

one of the small type Iconoscope cameras showed activities occurring there during daylight hours, when there was little or no business being transacted in the Convention Hall.

For the use of the general public, the RCA Manufacturing Company installed sixty television receivers in a large room in the south wing of the Museum Buildings associated with the Convention Hall. These receivers were supplied to permit those who were unable to obtain seats in the Hall to view, by television, the events occurring within the Convention Hall. Television receivers were also installed in the Convention Hall offices of the three leading Press Associations and in the PRR Press Club room. The picture signal and sound were transmitted to these receivers by means of special cables.

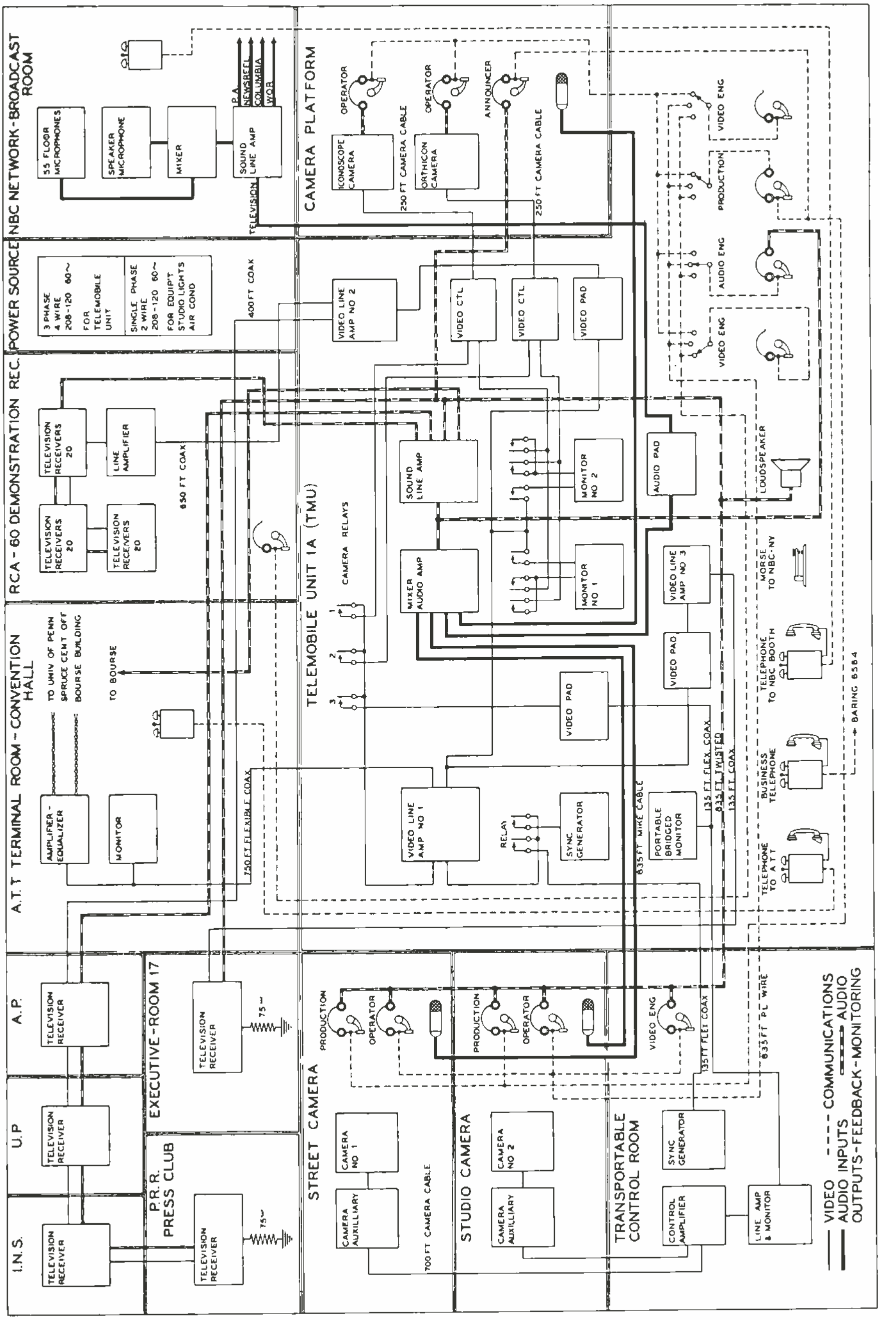
Reference to Figure 7 shows the relative positions of the camera platform, the mobile unit, NBC sound network booth, television studio, sound network studio, NBC viewing room, RCAM public viewing room, offices of the AP, UP, INS, and PRR Press Club and the Bell System telephone terminal room and power transformer room. Reference to Figure 8 shows how these widely separated television locations within the Convention Hall were interconnected. Over two miles of wire were installed in the Convention Hall by NBC to supply the necessary video, audio, monitoring feedback and intercommunication circuits. These circuits were exclusive of facilities installed by the Telephone Company, such as the audio, Morse, private line and business telephone circuits within the building. Of the figure mentioned, 2155 feet consisted of flexible coaxial cable and 1250 feet was represented by the various camera cables. The remainder was divided among communications, audio, and power wiring facilities. This is mentioned in order to acquaint the reader with the magnitude of this temporary television pick-up.

With three main pick-up points available, it was necessary to provide switching facilities to permit rapid transfer from one camera to another. This meant that a rapid switching system had to be devised whereby the signal outputs of the transportable and mobile units equipment could be transmitted to the NBC main control room in Radio City, New York, over the Bell System coaxial cable, as well as to the local monitoring receivers and the sixty receivers in the public viewing hall.

Signal switching, and provisions for multiple feeds involving two separate types of pick-up apparatus are much more complicated for video signals than for audio signals. The complete video signal contains not only picture signals but also blanking and synchronizing pulses. Both types of equipment in use were complete within themselves. Each was controlled by its individual synchronizing generator. The mobile unit was equipped with an electronic type of synchronizing generator and the transportable equipment made use of a rotary generator for originating the primary pulses. Thus these two systems were not synchronous one with the other.

Of course, there would have been nothing to prevent a system being devised wherein the basic pulses from one generator could be used to control the other so that the two systems would have been synchronous. The time at our disposal did not permit the design or construction of such a system and it would hardly have been worth while for such a temporary installation. Reference to Figure 8 shows the system used.

The Kinescope blanking pulse complete with the video frequencies comprising the picture intelligence and the synchronizing signal developed by the transportable equipment were divided at their respective sources and then fed to the mobile unit on separate flexible coaxial


NBC NETWORK-BROADCAST ROOM
55 FLOOR MICROPHONES
SPEAKER MICROPHONE
MIXER
SOUND LINE AMP
P.A.
NEWSREEL
COLUMBIA
WOR
TELEVISION
POWER SOURCE
3 PHASE 4 WIRE 208-120 60~ FOR TELEMOBILE UNIT
SINGLE PHASE 2 WIRE 208-120 60~ FOR EQUIP'T STUDIO LIGHTS AIR COND
RCA-60 DEMONSTRATION REC.
TELEVISION RECEIVERS 20
LINE AMPLIFIER
650 FT COAX
400 FT COAX
A.T.T. TERMINAL ROOM ~ CONVENTION HALL
TO UNIV OF PENN
SPRUCE CENT OFF
BOURSE BUILDING
TO BOURSE
AMPLIFIER-EQUALIZER
MONITOR
750 FT FLEXIBLE COAX
CAMERA PLATFORM
OPERATOR
ANNOUNCER
ICONOSCOPE CAMERA
ORTHICON CAMERA
250 FT CAMERA CABLE
TELEMOBILE UNIT 1A (TMU)
VIDEO LINE AMP NO 2
VIDEO CTL
VIDEO PAD
SOUND LINE AMP
MONITOR NO 2
MONITOR NO 1
AUDIO PAD
MIXER AUDIO AMP
CAMERA RELAYS
VIDEO LINE AMP NO 3
VIDEO LINE AMP NO 1
RELAY
SYNC GENERATOR
PORTABLE BRIDGED MONITOR
635 FT MIKE CABLE
135 FT FLEX COAX
635 FT TWISTED
135 FT COAX
VIDEO ENG
PRODUCTION
AUDIO ENG
LOUDSPEAKER
MORSE TO NBC-NY
TELEPHONE TO NBC BOOTH
BUSINESS TELEPHONE
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STUDIO CAMERA
CAMERA NO 1
CAMERA NO 2
CAMERA AUXILLIARY
700 FT CAMERA CABLE
TRANSPORTABLE CONTROL ROOM
SYNC GENERATOR
CONTROL AMPLIFIER
LINE AMP & MONITOR
635 FT PL WIRE
VIDEO
AUDIO INPUTS
OUTPUTS-FEEDBACK-MONITORING
COMMUNICATIONS
AUDIO

Fig. 8

cables. At the mobile unit, a portable monitor equipped with bridging input was connected across the picture signal line. Both the blanking and the synchronizing pulses were terminated at a camera relay which was installed in addition to the two camera relays normally used for switching either of the mobile unit cameras to the line amplifier input. The synchronizing signal originating at the mobile unit was connected so that it was retained whenever camera relays associated with the mobile unit were operated. These mobile unit synchronizing pulses were disconnected when the third relay associated with the transportable was closed. The synchronizing pulses from the transportable equipment were then being fed into the line amplifier. The picture signals associated with these synchronizing pulses were also fed into this line amplifier through a separate camera channel relay. The ratios of synchronizing, blanking and picture signal voltages of both these systems were compared at the mobile unit master control. Adjustments were made at the camera channel control points so that when either set of equipment was switched to the line amplifier, the voltages would be comparable. The overall output voltage was in turn adjusted to check with the input voltage requirements of the Bell System equipment situated in the telephone terminal room.

Problems in connection with the sound accompanying the picture presented little difficulty by comparison with those presented by the video signal. The audio pick-up from the speakers' rostrum and the delegates' microphones was supplied to the television control room from the NBC network control room. Microphones associated only with television were located at all camera positions for the use of the announcers in giving running comments or interviewing personalities. The sound network feed and television microphones were brought into a four-position mixer on the Telemobile unit 1-A.

The communication facilities (Figure 8) installed for the engineering and program department coordination surpassed any set-up used previously for audio broadcasting at other pick-ups of this character. Telemobile Unit 1-A was designed as the master control point. Three television engineers and the program director constituted the personnel at this point. The private line system aboard the mobile unit was set up so that one video engineer was in constant communication with the cameras associated with the mobile unit. The other video engineer communicated with the video engineer at the controls of the transportable equipment. This latter position also mainained communication with camera operators in the studio and on the sidewalk. The audio engineer monitored a private line to the RCAM public viewing room. In addition he operated the Morse wire to New York. When the audio engineer was engaged with the RCAM circuit, the second video position engineer took over the Morse instrument, which was placed between the two men. Three other telephone systems were strategically placed so that any member of the engineering staff had access to them. These were: a business telephone, on which calls were received from the press offices, the Press Club, National Committee members, and the NBC office; a private line to the Bell System terminal room which was used occasionally to check television signal voltage levels, reports of noise, or other matters pertinent to the transmitted signal; a private line to the NBC sound network booth for conversations concerning the audio feed from the state delegations, and the speakers' stand. On all circuits other than the latter three, a key switching system allowed any of the engineering personnel in the mobile unit to transfer his phone equipment to another circuit.

Program producers were stationed at all camera positions with the exception of the camera platform in the main hall. At this point the announcer acted as his own producer. The program department was interconnected with its own private line system which enabled it to carry on conversations concerning program matters without interfering with technical discussions on the engineering department's communication circuits.

It is customary in sound broadcasting for an announcer, in his running comment, to present word pictures of the events taking place around him. However, in television the radio audience can see the event on its home screens, which considerably changes the technique of announcing. The running commentary is unnecessary. At the Convention, the predetermined sound coverage which followed a routine set-up by the Republican National Committee dictated the movements of the cameras. The program director, watching the picture on the Kinescope in the mobile unit, instructed both television camera operators and announcers as to the scenes to be televised, as the center of interest shifted in the hall.

During the balloting, the cooperation between announcer, producers, control engineers and camera operators was at its peak. By means of the private line facilities installed, producers and engineers directed the camera operators both from the engineering and program standpoints as to what events and sequences of scenes were to follow. This may best be explained by citing an incident which occurred during the balloting. The camera was trained on a state delegation which was about to cast its vote. The scene was self-explanatory, and the announcer remained silent. He noticed two members of another delegation engaged in a heated argument. This delegation's vote was to be cast within the next two or three minutes. The announcer broke in to describe the argument which was rapidly coming to a point where the two men were about to exchange blows. The camera operator, his eyes fixed on the scene encompassed by his own view finder was unaware of what was transpiring until he heard the announcer's description. He immediately panned his camera to the spot indicated by the announcer. The resulting picture was one of the television highlights of the Convention.

The system used to keep all points informed as to what was transpiring on the sound system is known as "feed back monitoring." Each engineering and program position is equipped with a monitoring device. In the case of the camera operators, the audio control engineer, producers and announcers, sound is monitored by means of head receivers. A loudspeaker system in the mobile unit is used by the video engineers and program director. In the case of a producer desiring to transmit information to camera operators when there is no time to relay such information by means of the engineering department's private line facilities, he may cut into the feedback monitoring system, a special microphone which permits everyone to hear his instructions on the system.

Two months before the plans of the National Committee had materialized sufficiently to enable specific locations for the mobile unit and studio to be determined, the engineers of the Bell System started work on the installation of special circuits, repeaters and equalizers to extend the Philadelphia-New York coaxial cable circuit from the Bourse Building in Philadelphia to the Convention Hall. The circuit was terminated in the telephone terminal room at the northwest corner on the second floor of the Convention Hall Building. Later, when the location for the mobile unit was selected at the southeast corner of the first floor, it was necessary to interconnect these two points. Seven hundred and fifty feet of flexible coaxial cable was needed to complete the connection.

Figure 9 shows the route which the picture signals took from the various cameras located in the main hall, the television studio and street, through the local control rooms to the Bell System and thence through the cable to the RCA Building in New York, and eventually to the home receivers through the NBC Empire State transmitter. At the telephone terminal room in the Convention Building, the Bell Laboratories set up a monitor, an amplifier and video equalizer. Two picture circuits were provided between this point and the Bourse Building, the second being held as a spare. Repeaters were located at the University of Pennsylvania, the Spruce central telephone office, the Bourse Building and the Convention Hall. These locations are approximately one and one-eighth circuit miles distant from each other. The picture circuits from the Convention Hall to the Bourse Building were each approximately three and one-half miles in length, and were capable of transmitting a band of frequencies from 30 to 2,800,000 cycles per second.

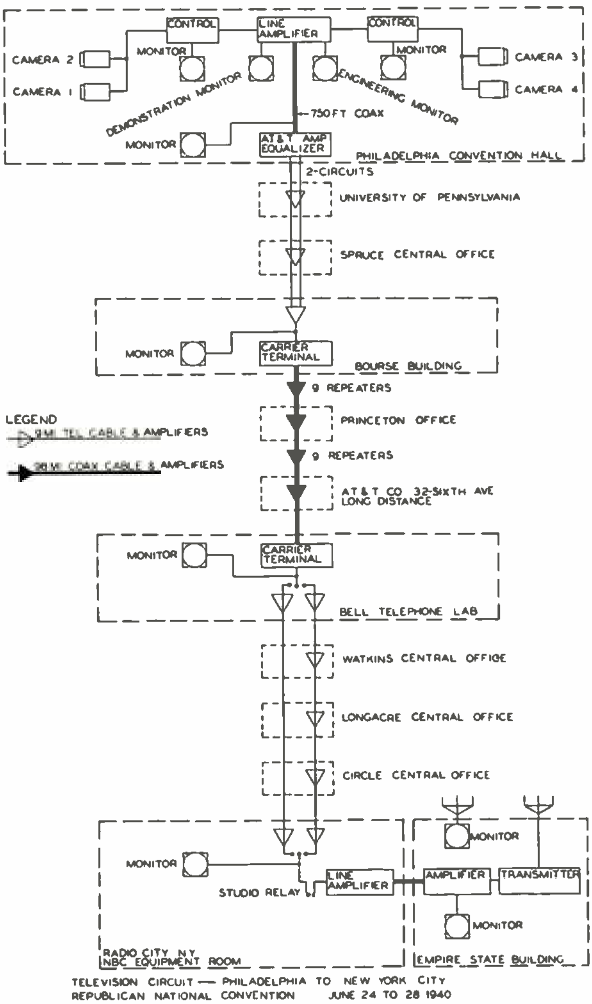

Fig. 9

At the Bourse Building coaxial terminal, the video signal was transposed to a higher range of frequencies, the lowest of which was of the order of 300,000 cycles, for transmission over the coaxial cable to New York. At the New York terminal of the cable, the signal was restored to the normal video range.

A NEW HIGH-QUALITY SOUNDHEAD

BY

R. H. HEACOCK

Engineering Department, RCA Manufacturing Company, Inc., Camden, N. J.

Summary—*The methods of filtering out the irregularities of motion introduced by the intermittent sprocket of the projector as well as those produced by the rest of the projector mechanism and the film take-up are briefly described. The new "shock-proof drive" is described in detail. Other less important design problems such as limitation of static pick-up, immediate replacement of burned-out exciter lamps, and the greater ease of optical adjustment of sound-optical lens barrel are outlined. supplemented by a description of a modern high-quality soundhead in which these problems have been solved.*

THE most vital and difficult problem facing sound engineers in the reproduction of sound on film has been the elimination of irregularity of film motion introduced by the intermittent sprocket of the motion-picture projector. Since the motion-picture industry was well established before the advent of sound, designs of projector, magazines, pedestal, take-ups, and other accessories had all been developed with no thought of the requirements of the equipment for reproducing sound. In these designs, smooth rotating motion was applied to the upper pull-down sprocket which pulled the film down from the upper magazine and also to the lower hold-back sprocket which held the film back against the pull of the film take-up located in the lower magazine. Between these two smooth-running sprockets was the intermittent sprocket which, as its name implies, would move the film from one frame to the next and would then hold the frame stationary in the picture aperture for projection of that frame. In the film path the intermittent sprocket and film gate were isolated from the smooth running sprockets by means of an upper and lower film loop. The quick pull-down of the intermittent sprocket acted merely to move momentarily some of the film from the upper loop to the lower loop. For silent-picture projection, the intermittent motion of the film did not have to be entirely eliminated before it reached the lower hold-back sprocket since some slight irregularity of rotative speed at this sprocket had no effect on the operation of the lower take-up mechanism.

About the time sound on film was introduced, the number of frames projected per second was increased to a standard of 24. This was done for three main reasons, (a) to provide smoother action in the projected picture, (b) to simplify the recording of high frequencies, and (c) to reduce flicker.

In order to obtain a sound take-off location at which the film would be moving uniformly it was necessary that severe intermittent motion of the film at the picture gate be eliminated in a film length of approximately 15 inches. The slightest irregularity in film speed at the point of sound take-off would be very undesirably reproduced in the form of "wow". Various methods of filtering out irregularities in film speed

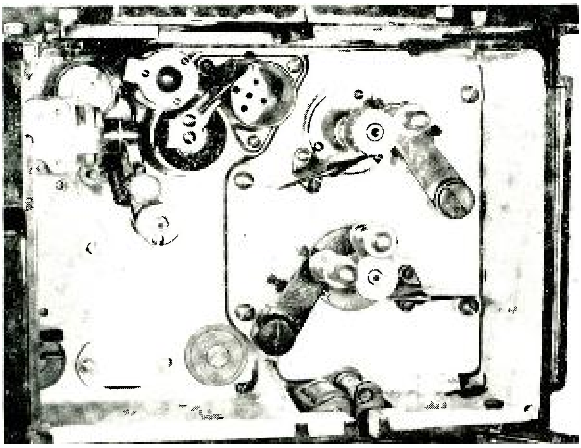

Fig. 1

were used in early soundheads. At the present time all of the better-type soundheads make use of the conventional rotary stabilizer.

ROTARY STABILIZER

This rotary stabilizer is an ingenious device which filters out slight irregularities in film motion such as may be caused by very small disturbances coming from the gear train or by the sprocket teeth engaging with the film. This unit, since it is the heart of the modern high-grade soundhead will be described in detail.

Figure 1 shows an inside view of the operating compartment of the soundhead. The rotary stabilizer is attached to the film drum. When the film is pulled around the drum by the constant-speed sprocket, the film rotates the drum which in turn causes the outside shell of the rotary stabilizer to rotate. Figure 2 shows a cross section of the rotary stabilizer. It should be mentioned that the outside shell is made of very light material so that it will not place undue load on the film when the soundhead is started. Located inside the external housing is a heavy flywheel which is mounted upon a large ball bearing. The clearance between the outside diameter of the flywheel and the inside of the shell is only 0.008 inch. The rotary stabilizer is completely filled with oil so that when the light outer shell rotates, it transmits its rotative motion through the thin oil layer to the heavy flywheel. After the flywheel has been brought up to speed, it tends to stabilize and filter out variations in speed of the outer shell.

The film between the projector hold-back sprocket and the sound-head constant-speed sprocket is drawn tight and then slackened off two sprocket holes so that the drum floats on a two-hole loop between these sprockets. In this way the rotary stabilizer can filter out minor

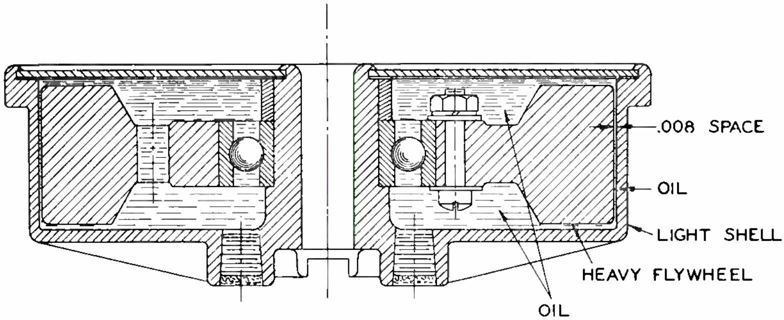

Fig. 2

speed variations caused by meshing gears in the motor-drive gear train and sprocket teeth engaging with the film-sprocket holes.

In addition to the major problem of providing uniform film speed at the sound take-off position, many other problems presented themselves for solution. Some of these are covered in the following pages.

The outlook of exhibitors has changed very materially since the advent of sound. At first the exhibitor had to satisfy only the public demand for the little-known and brand-new "talkies". No one was unduly concerned about either "wows" or "response curves".

Now, however, things have changed very materially. The exhibitor has been forced to have high-grade equipment which will give high-quality sound or the cash customer goes across the street where, in his opinion, "the sound is better". This, naturally, has forced the manufacturer to incorporate in equipments the most improved features to guarantee satisfactory sound to the patrons of the theaters in which the equipment is located.

Shock-Proof Drive

When equipment is made for optimum performance, cost should not be a limiting factor in providing smooth and "wow"-free operation. It is obvious that the intermittent projector mechanism places a high, sharp peak load on the driving members which is additional to the erratic loading created by inexpensive chain-driven take-ups, improper belts, bent reels, and poorly adjusted take-up mechanisms. Each contributes its part to supplying an objectionable source of irregular and erratic rotational motion.

Since the constant-speed sprocket is immediately adjacent in the film path to the sound take-off location on the rotary-stabilized drum, two spiral gears have been added in the equipment described in this article to improve the method of picture-head and lower take-up drive. On other commercial soundheads, the motor is connected to the "motor pinion" by means of the flexible coupling. The "motor pinion" drives the "constant-speed shaft gear" and the "hold-back shaft gear" simul-

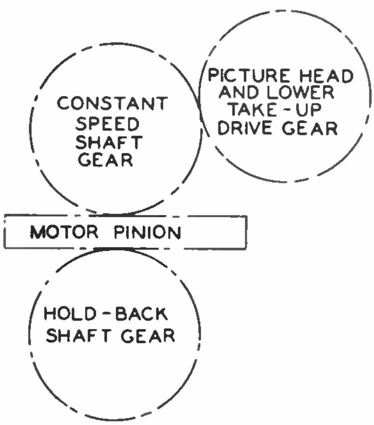

Fig. 3

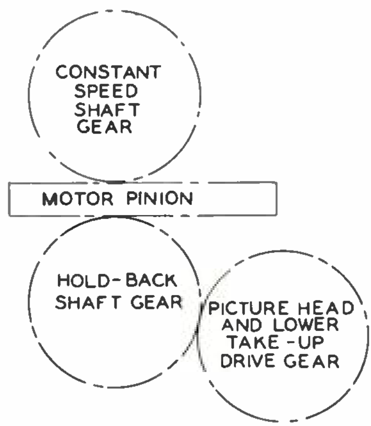

Fig. 4

taneously. Constantly meshing with the "constant-speed shaft gear" is the "picture head" and "lower take-up drive gear". It may be seen from Figure 3 that all of the speed variables introduced by either the picture-head or the lower take-up mechanism are impressed directly upon the constant-speed sprocket. These variable loads necessarily result in "wow" in the reproduced sound. The upper shaft (constant-speed sprocket shaft) is a long member, on the operating end of which is located the constant-speed sprocket. Keyed to the opposite end of this member is the gear and pulley assembly which drives both the picture head and lower take-up. The lower shaft is a short one and has only the hold-back sprocket located on its operating end and no driving member on its other end.

In the soundhead described in this paper, this set-up has been reversed to produce "shock-proof" drive. Figure 4 shows that the "motor pinion" simultaneously drives the "constant-speed shaft gear" and the "hold-back shaft gear". The variable loads of the picture-head and lower take-up mechanism are continuously impressed upon the "hold-back shaft gear". Due to the steep angle of the meshing teeth of the "motor pinion" and mating helical gears, this variable load cannot be transmitted from "hold-back shaft gear" through "motor pinion" to "constant-speed shaft gear". It is to be noted, however, that the upper shaft (constant-speed sprocket shaft) is, this time, the short member and has attached to its operating end the constant-speed sprocket. Its other end terminates in a cap and acts only as a support member. The lower shaft for the new design is the long one with the

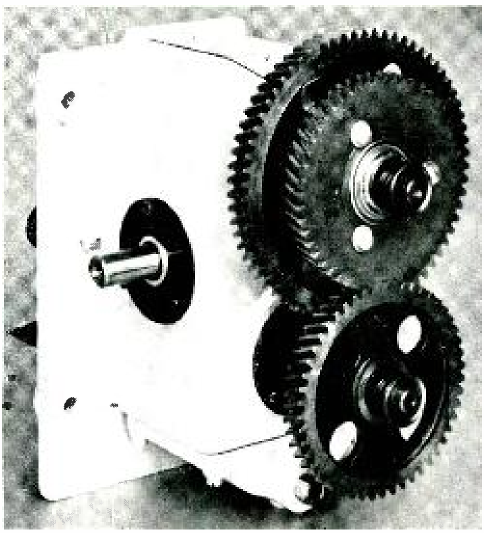

Fig. 5

hold-back sprocket located on its operating end, while one of the new spiral gears already referred to is keyed to its opposite end. This spiral gear meshes with a second spiral gear with a 1:1 ratio. This second spiral gear is directly assembled to the picture-head drive gear and the lower take-up belt pulley. These three members rotate as one unit on two sealed, dust-proof, self-lubricating ball bearings which rotate on a stationary stud projecting from the back of the gear-case housing.

It is evident that the varying loads from both the picture-head and the lower take-up mechanism are impressed through the gear train on the hold-back sprocket shaft. They cannot go from the lower spiral gear through the motor-pinion shaft to the upper spiral gear due to the steep angle of these gears. Consequently, the constant-speed sprocket shaft is isolated and is effectively "wow"-filtered for variables of picture-head and lower take-up drive.

On this new soundhead, the complete drive gear mechanism is assembled in a single removable unit as shown in Figure 5.

Although the new method of drive cannot eliminate entirely irregularities or "wows", it does very effectively reduce them. Comparative tests made in the laboratory indicate that the use of "shock-proof" drive will reduce by approximately twenty per cent the "wows" introduced by average picture-head and average lower take-up mechanisms.

Mounting and Oil Collection Plate

It must be possible to mount any of the standard picture projectors on the soundhead. Accessories for driving the picture projector must

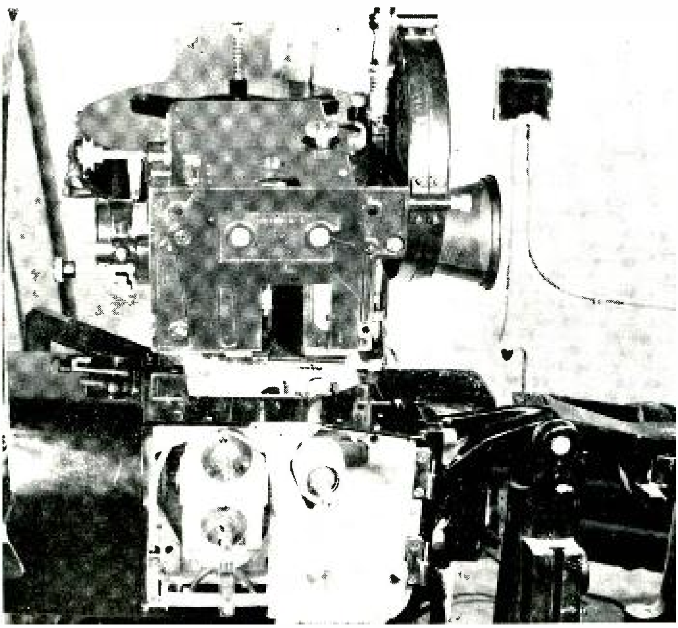

Fig. 6

be simple and easy to install. Figure 6 shows the projector and mounting plate tipped back, away from the soundhead. In order to facilitate the attachment of various standard picture projectors to the soundhead, two different types of mounting plates have been designed. For simplicity, only that plate which is used for Simplex Projectors will be described.

The picture-head, to which the mounting plate is rigidly attached, is set upon the top of the soundhead so that three pins which project from the bottom of the mounting plate are in engagement with three corresponding slots in the top of the main case casting of the soundhead. Two of the pins on the bottom of the mounting plate engage two slots which are parallel to the operating side of the main case casting. These pins working in the slots permit motion toward or away from the screen without permitting lateral motion. The third pin, already referred to, engages a lateral slot in the top of the main case casting which is at right angles to the two slots. This third pin is mounted eccentrically with relation to a large hex-headed cap screw. If this hex-headed screw is rotated, the complete picture-head mechanism may be moved smoothly and accurately either toward or away from the screen so that exact mesh of picture-head drive gears may be readily established.

A fool-proof method of oil collection must be provided so that oil from the projector will not get into the soundhead and smear the lens system of the sound reproducer. In addition to acting as a picture-head mounting plate, this same member acts as an oil-collection plate

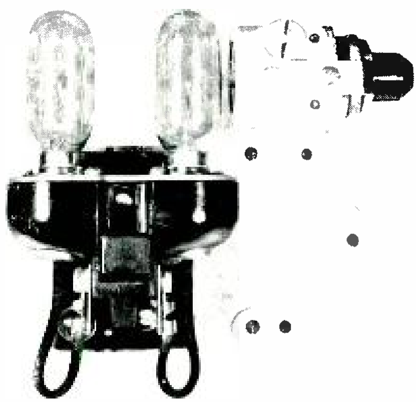

Fig. 7

which conveys the oil completely through the soundhead to a transparent, unbreakable, removable receptable.

A uniform brilliant light must be supplied in a narrow aperture in order to scan the sound recording on the film. Means for accurately focusing this light must be provided. Figure 7 shows the double-exciter-lamp mounting and the optical-lens barrel provided for this purpose. An adjustable stop is incorporated so that there is no lost motion between the focusing nut and its contacting members.

In the past it has been difficult to properly locate the slot for the lug on the optical lens barrel in such a manner that the light slit would be exactly at right angles to the sound track on the film. This difficulty has been overcome by making the slot in a loose sleeve in which is inserted the optical lens for proper rotational position in the sound bracket. The sleeve is then located in position at the factory so that it cannot be adjusted in the field. In this way, a more accurate and less expensive assembly is available.

Double Exciter Lamp Holder

In the soundhead described prefocused bulbs are used which are similar to those in present use in the automobile industry. The skirt on the base of the lamp is attached last. A careful adjustment is made so that the filament is held with relation to the three indexing points on this skirt to within ±.010 inch in both lateral planes and in the vertical plane. This tolerance is so much closer than has been possible

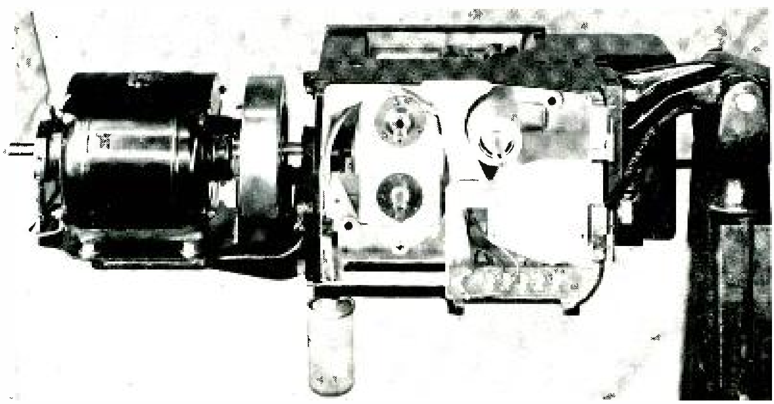

Fig. 8

with the older type of lamp that it eliminates the necessity for vertical and lateral adjustment of a new exciter lamp.

For easy replacement a new double-exciter lamp holder is now used which is not of the turret type. Positive and sure electrical contacts are established by means of spring-wiping action on two rigid posts. One of these posts is displaced laterally with relation to the other so that if the double-lamp socket is removed, rotated 180 degrees, and reinserted in the soundhead, electrical contact is automatically established for the lamp adjacent to the lens barrel. This double-exciter lamp socket makes it possible to position a new exciter lamp practically instantly.

Criticism has been made that some soundheads in the field are supported in such a manner that the complete assembly of the soundhead, projector, and magazines is not rigid enough and that the picture on the screen is not steady. To prevent the possibility of such a fault in this soundhead, a new and heavy main bearing bracket with heavier clamping screws has been designed. The main case casting has also been designed to give adequate strength and rigidity.

The lateral guide and pressure roller can be adjusted laterally to provide very accurate positioning of the film with relation to the sound aperture and insure smooth passage of the film about the drum with a minimum of buckling.

All pad rollers which are used in the soundhead contact the film only at the outside edges beyond the sprocket holes.

The drum and sound take-off parts must be cushion-mounted with relation to the rest of the soundhead in order to filter out motor-armature and gear-train vibrations. As may be seen in Figure 1, the complete centerplate is cushion-mounted in sponge rubber in relation to the main case casting.

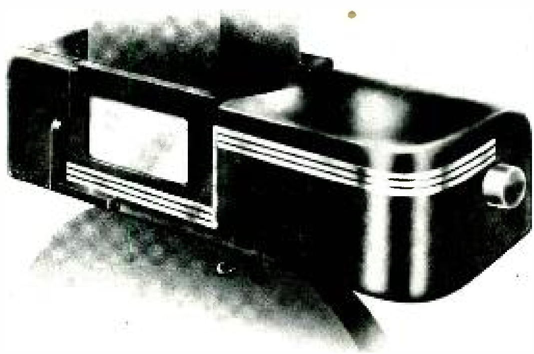

Fig. 9

A completely shielded phototube circuit including socket, leads, and transformer must be provided in order to eliminate static pickup from the mechanism.

Figure 8 shows the motor to which is rigidly attached a flywheel. The flywheel increases, to approximately three seconds, the time required for the overall system to reach full speed thus keeping film mutilation to a minimum.

The modern broadcast studio has influenced the style of current theater equipment. Ten years ago, it was widely felt that the theater booth did not have to present an attractive appearance. Following the lead of many of our modern broadcasting studios, our foremost theaters are rendering their projection rooms more attractive. A motor cover has been added and the general styling changed as shown in Figure 9 to make the soundhead look massive and powerful when assembled with a suitable pedestal, 18-inch magazine, and a large size arc-lamp housing.

The effect of planned styling on the soundhead is evident in the method of taking care of an indicator for the exciter-lamp. On earlier soundheads, a small "bull's-eye" was located on the exciter-lamp compartment door. When the lamp went on this "bull's-eye" was illuminated in the conventional manner. On this soundhead, some of the light from the exciter-lamp passes through a slit in the exciter-lamp shield. This light strikes the edge of the glass window and by means of edge lighting illuminates the manufacturer's trade mark. In this way an ornamental use of the light has been made.

Too great emphasis cannot be placed upon ease of servicing. If component parts or assemblies are difficult to reach, it naturally follows that these parts of the soundhead frequently do not get the necessary attention from the projectionist. In time, lack of attention

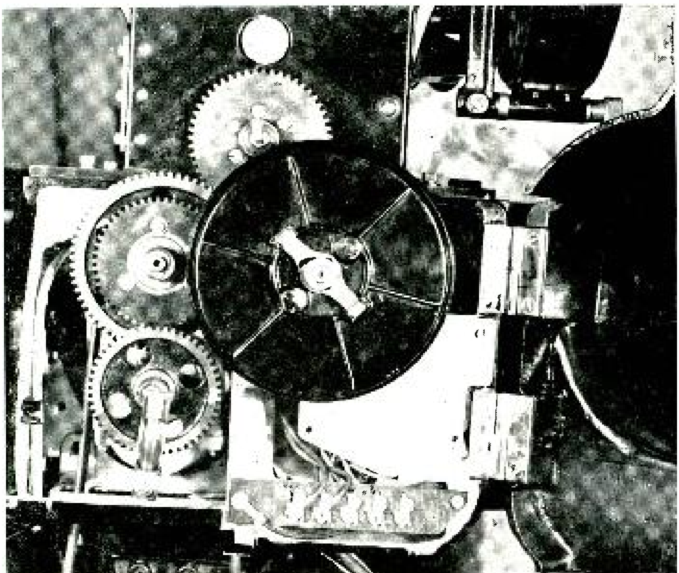

Fig. 10

will result in faulty sound and dissatisfaction on the part of the customer. Consequently, great care has been taken to facilitate servicing of all parts of the soundhead.

The gear-box assembly is readily removable so that immediate replacement of this complete unit can be made should any trouble develop in the motor-drive gears. The complete center-plate assembly may also be readily removed from the soundhead as a unit.

Figure 10 shows the terminal board which is mounted below the rotary-stabilizer and all drive gears. This permits easy inspection of connections by opening the gear cover which is mounted on hinges to the main case casting and by removing the small terminal board cover which is held in place by means of two knurled-head screws.

In this equipment, the thoughts, suggestions, and constructive criticism of many sales and service engineers working in close cooperation with development and design engineers have been coordinated to produce this high-grade soundhead.

STABLE POWER SUPPLIES FOR ELECTRON MICROSCOPES

BY

A. W. VANCE

RCA Manufacturing Company, Inc., Camden, N. J.

Summary—*A description is given of the power supplies, developed in the Electronic Research Laboratory of the RCA Manufacturing Company, for the operation of electron microscopes. A high direct voltage (20 to 100 kilovolts) of exceptional constancy is obtained by applying an electronic regulator to a transformer and rectifier driven from the ordinary 60-cycle power line. Similar electronic regulators supply currents of great stability for operating the magnetic lenses of the microscope. Data are given showing the variation of voltage and current with time under ordinary power-line conditions. These supplies are so stable as to impose no limitations on the resolving power of the electron microscope.*

THE electron microscope has presented electrical engineers with new power supply problems. The required voltage varies between 20 and 100 kilovolts, with a maximum load of less than 1 milliampere at any voltage. Three currents ranging up to 400 milliamperes through load resistances of about 700 ohms are required. The currents and potentials must have great constancy over short periods, determined by the maximum exposure time of the photographic plate used to record the images. The longest exposure is about 30 seconds. The short-time constancy required is determined by the resolution of the microscope, the degree of misalignment of the electron lens system, and the nature of the specimen. Theoretical calculations based on a resolving limit of 10 Angstrom units indicate that the most critical current must not vary more than 0.004 of one per cent, and that the high voltage should not vary more than 0.0066 of one per cent. It is desirable that the voltage should be as independent of load as possible to minimize variations in voltage due to change in the filament emission of the electron microscope and to any variable leakages or corona discharges.

The regulated high-voltage d-c supply described in this paper is powered by a conventional voltage-doubling rectifier, oil immersed in a steel tank with its output brought out through a rubber insulated shockproof cable.

A schematic diagram of the regulated high-voltage system showing only the essential components appears in Figure 1. The high-voltage rectifier is enclosed within dotted lines and is marked "A". A 1000-

megohm 100,000-volt carbon resistor, forming the high-voltage portion of the voltage divider and shunted by a compensating condenser, is mounted in the generator tank under oil. The high voltage leaves the generator through a shockproof cable and enters another oil-filled tank containing the resistance-capacitance filter shown. The wire-wound resistor element of this filter serves as a current limiter for the system and prevents serious overloading of the generator in case of breakdowns and discharges in the microscope. The condenser is made up of an oil-filled condenser plus the capacitance of 50 feet of shock-proof cable leading from the high-voltage vault. The lower voltage portion of the voltage divider is adjustable between 5 and 25 megohms for a voltage range of from 100 down to 20 kilovolts. This resistance ap-

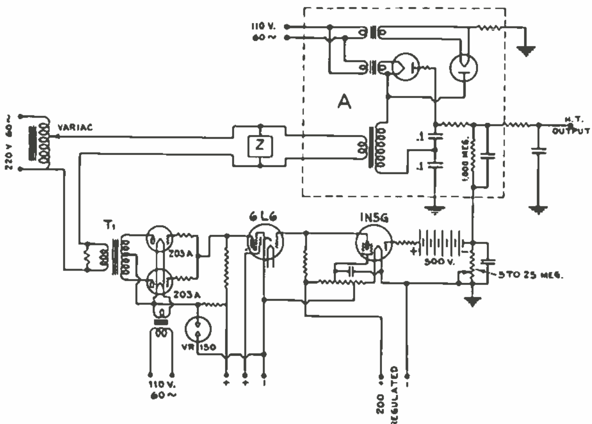

Fig. 1

pears at the remote control position and is shunted by a condenser in order that the capacitance ratio of the divider approximates its resistance ratio.

As is shown in Figure 1, the divided high voltage is opposed by a 500-volt dry battery as the standard and the difference is impressed on the first tube (1N5-G) in the d-c amplifier system. This tube is d-c connected to a 6L6 d-c amplifier actuated by a floating d-c supply. The 6L6 drives in parallel fashion the grids of two 203-A power tubes, whose plates are connected in push-pull across the secondary of a plate transformer. The primary is shunted by an impedance-limiting resistor and is connected in series with the a-c input supply to the main generator. The 203-A tubes act as a variable resistance across the secondary and so vary the primary impedance in accordance with changes in common grid voltage. Since the grids of the 203-A's are driven somewhat positive when the lowest impedance is desired, it is necessary that the grids be isolated from each other by the resistors shown in their grid circuits. This is necessary to prevent the grid of that tube, whose plate voltage is negative, from limiting, because of grid current, the positive swing on the other grid. The VR150-30 glow tube serves to maintain the cathodes of the 203-A's at 150 volts positive with respect to the 6L6 cathode. The whole system is an inverse-feed-back circuit with one point heavily by-passed, and producing all the high-frequency attenuation. Since the loop gain of this system passes through unity at about 90° phase shift and has no lower cut-off, it is inherently stable.

The network "Z" shown, connected across the input to "A", is for the purpose of power factor correction, and maintaining the input impedance of the generator within reasonable limits over the wide range of required output voltages. These wide voltage changes are accomplished by means of an electromechanical system which automatically "follows up" the regulator. The regulator shown in Figure 1, therefore, is required to handle a narrow range sufficient to correct for line-voltage changes or load changes. The follow-up regulator will be described later. For the present, the circuit in Figure 1 will be considered as operating at fixed output voltage.

The constancy of the output voltage in Figure 1 is limited by a number of factors. On a straight loop-gain basis the regulator should reduce the effect of the line-voltage variations by several thousand times. Very fast line-voltage variations are reduced much less effectively, since the regulator is limited in speed by the power-line frequency. The standard battery has been measured to have less than three parts in one million variation within the 30-second exposure time and so does not limit the constancy. The voltage divider, so far as the resistors themselves are concerned, is also stable to at least 3 parts in one million over a 30-second period. However, the circulation of the oil about the resistor, due to thermal and electrical gradients, causes the effective ratio of the voltage divider to vary. The capacity divider is affected by oil circulation also. The effect is not fully understood and seems to be related to the movement of the oil. Dividers of negligible d-c leakage enclosed in wax exhibit great stability. However, in this unit the divider is oil-immersed and one of the principal causes of instability is the variation of the effective ratio of this divider. Drifting of the d-c regulating amplifier is a possible source of output variations. This effect has been reduced to negligible proportions by supplying plate, screen, and filament voltages for the input 1N5-G from a small electronically regulated power pack which maintains its voltage constant to less than 0.1 of one per cent at all times. Since the gain of this stage is about 100, the drifting in the following stages is negligible. Ripple constitutes a continual voltage variation. The peak ripple of this unit, as measured by means of a capacitance divider, is

0.0016 of one per cent. The regulator maintains the voltage across the voltage divider essentially independent of load and thus the output impedance at the end of the resistance-capacitance filter is nearly equal to the value of the filter resistor. With proper control of corona and leakage at the microscope and storage battery filament supply, no serious variations due to change in load current have been observed.

The measurement of the voltage variations of a high-voltage supply of this nature is difficult, because suitable standards are not available. Either a high-voltage supply of great constancy or a very high-resis-

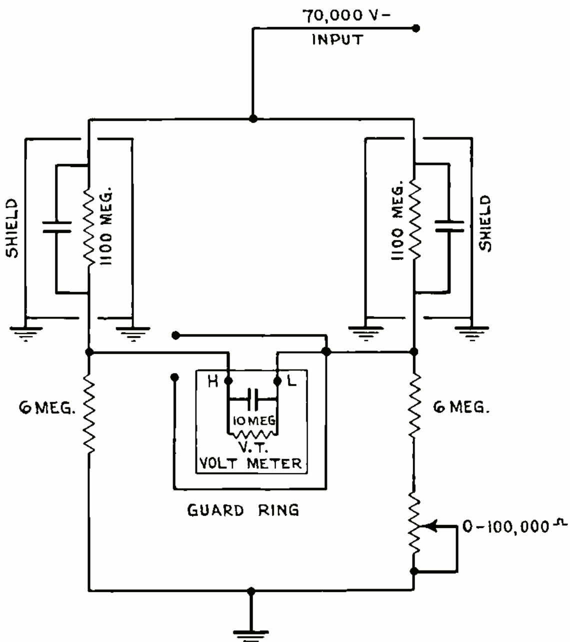

Fig. 2

tance potentiometer of equally great stability is necessary. The standard high-voltage supply is out of the question since it would mean a battery bank supplying up to 100,000 volts. The potentiometer cannot be wire wound to the right order of resistance (1000 megohms) at reasonable cost. The problem, however, is simplified by the facts that only short-time stability is required in the standard potentiometer and that the absolute magnitude of the voltage to be measured is relatively unimportant. The problem was solved by constructing two identical potentiometers using 1100-megohm, 100,000-volt carbon resistors enclosed in metal cases filled with high-resistivity wax for the high-voltage portion and 6-megohm wire-wound resistors for the lower voltage portion. A hard rubber insulated compensating condenser is included in the wax-filled case with the high-voltage resistor. The low-voltage connection is brought out in a coaxial lead, and the high-voltage connection enters through a hard rubber bushing. The two wax-enclosed resistors are placed in an oil-filled tank with two 6-megohm wire-wound resistors connected by coaxial leads to form the bridge circuit shown in Figure 2.

The indicator is a vacuum-tube microammeter similar to the RCA Ultra-Sensitive Meter, with 0.1-volt full scale range at 10 megohms resistance. The meter has a shunting condenser which keeps the ca-

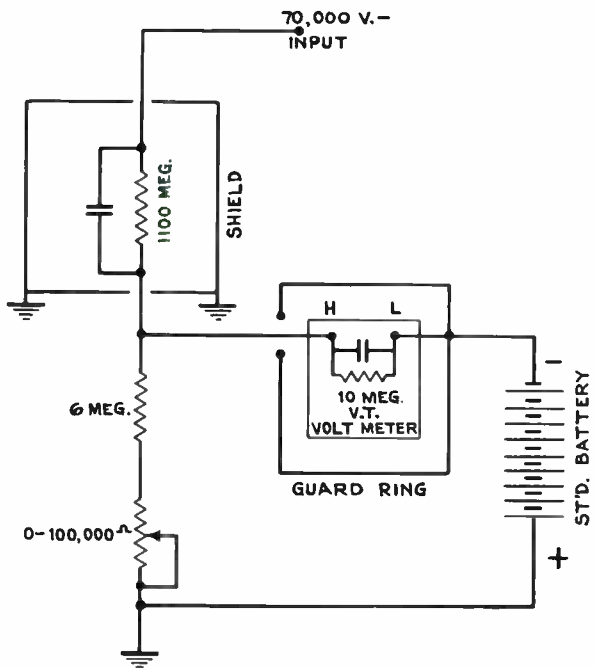

Fig. 3

pacitance ratio of the bridge nearly equal to the resistance ratio. 70,000 volts were applied to this bridge from the regulated supply and the meter was adjusted to zero by adjusting the 100,000-ohm resistance box shown in one leg of the bridge. The maximum deflection observed, within any 30-second interval during a twenty-minute period, was 0.001 peak-to-peak volt. This corresponds to a stability of 0.00026 of one per cent for each potentiometer. One of the potentiometers was then disconnected and a dry battery substituted as shown in Figure 3. The dry-battery voltage was adjusted so that the meter read 0.05 volt. The meter was observed for 20 minutes and the maximum deflection during any 30-second interval was 0.014 volt. The average deflection appeared to be one-half of this, or 0.007 volt; occasionally the voltage stayed within 0.003 volt for 30 seconds. The dynamic ratio of the potentiometer, allowing for the 10-megohm load of the meter, was 294. The fractional maximum percentage of variation, therefore, is about $\frac{0.014 \times 294}{70{,}000}$, or 0.0059 of one per cent; the average variation, 0.003 of one per cent; and the best condition, 0.0012 of one per cent. The ripple percentage must, of course, be added to these figures, and thus the maximum variation becomes 0.0075 of one per cent. This is reasonably close to the 0.0066 of one per cent theoretically required for a resolution of 10 Angstrom units.

The follow-up regulator which takes care of large variations is shown in schematic form in Figure 4. This regulator acts to maintain

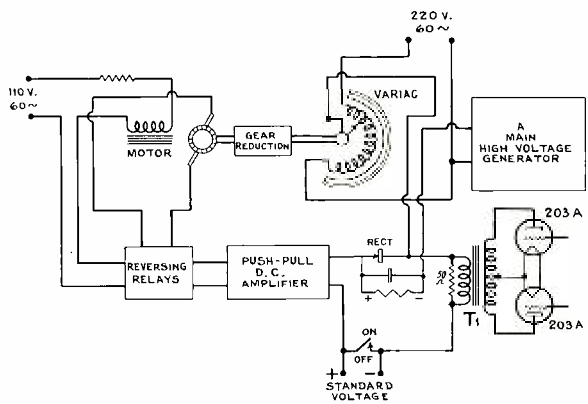

Fig. 4

a constant a-c voltage across the primary of the regulating transformer. The a-c is rectified and compared to a standard d-c voltage; the difference is applied to a push-pull d-c amplifier of conventional design. The output of this d-c amplifier drives a relay system which controls the motor, shown geared to the variac, in such a way that when the standard voltage exceeds the rectified voltage across the regulating transformer, the motor drives the variac up to increase the voltage. When the rectified voltage exceeds the standard voltage, the motor drives the variac down. The d-c amplifier is over-biased in the output stage, and thus has a "dead spot" condition where the motor circuit is open. This "dead spot" is large enough that the motor seldom moves except when the voltage is being adjusted. This follow-up regulator is used to turn the voltage on and off without disturbing the exact voltage setting as determined by the ratio of the voltage divider in the main regulator. This is accomplished by short circuiting the standard voltage so that the motor drives the variac down to zero. The voltage is, therefore, never changed rapidly over any great range, minimizing transients due to switching. The actual installation includes several routine interlocks, limit switches, and circuit breakers, which will not be discussed here.

The current supplies, while demanding even greater stability than the voltage supplies, are less of a problem because of the low voltages involved. In Figure 5 is shown the circuit for one of these supplies. The source of power is a well filtered conventional rectifier. The main

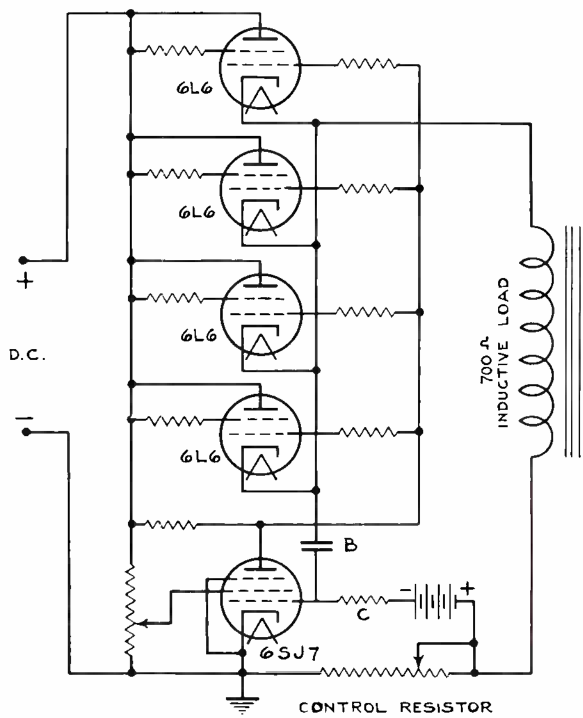

Fig. 5

control tubes are 6L6's connected as triodes, with resistors in the screen-grid circuits to prevent excessive screen dissipation. The load passes through a control resistor to the ground side of the rectifier. This resistor is of the low-temperature coefficient, wire-wound type. The voltage drop across the resistor is compared to the standard battery and the difference impressed on the grid of the 6SJ7 d-c amplifier tube which drives the grids of the 6L6's. Thus a simple feedback current regulator is formed. The circuit acts as a current regulator for very low frequencies, and by means of the condenser marked "B" and the grid resistor marked "C" also provides constant voltage regulation for high frequencies, such as hum frequencies. Due to the large inductance of the load, the feedback gain for hum frequencies is very large and negligible hum is obtained in the load. The constant-current feature at low frequencies minimizes variations in output current due to temperature rise in the electromagnetic lens load. At low frequencies the feedback gain is not enough to give the desired output current constancy against line-voltage changes. There exists in this circuit a bridge action due to the screen-grid connection of the 6SJ7. As the input voltage increases, the screen voltage increases, tending to increase the plate current of the 6SJ7, and so bias off the 6L6's. The higher the screen voltage applied to the 6SJ7, the greater this effect because by setting the screen voltage at a certain value, the screen action can be made to compensate exactly for the lack of sufficient gain in the feedback part of the regulator and make the output current independent of input voltage over small ranges. If the screen voltage is too high, the system over-regulates; if too low it under-regulates. With this adjustment properly set, the circuit of Figure 5 holds the output constant to 0.002 of one per cent for ordinary line-voltage conditions.

In conclusion it may be said that the theoretical requirements based on a 10-Angstrom unit resolving limit have been closely approximated. These supplies have been used with electron microscopes continuously for one year without any indication that they limited their operation in any way. Work on regulated current and voltage supplies is being continued in the research laboratories of the RCA Manufacturing Company in order that supplies will be available to meet any new requirements of the constantly improving electron microscope.

The writer wishes to express his gratitude to Dr. V. K. Zworykin, under whose direction this project was undertaken, for his valuable encouragement and advice. The writer wishes to thank Mr. J. M. Morgan for his aid in constructing, installing, and testing the apparatus.

A VESTIGIAL SIDE-BAND FILTER FOR USE WITH A TELEVISION TRANSMITTER

By

George H. Brown

RCA Manufacturing Co., Inc., Camden, New Jersey

Summary—*The television transmitter standards adopted by the R.M.A. place the carrier of the picture transmitter at a point which is 1.25 megacycles above the lower edge of a six-megacycle channel. The television receiver characteristic permits reception of the carrier, the upper side-bands, and those lower side-bands which lie within 0.75 megacycle of the carrier. Any other lower side-bands which might be transmitted would not be accepted by the receiver, and would therefore play no part in furnishing picture information. Since these side-bands which are not accepted by the receiver lie outside of the assigned channel, they must be suppressed at the transmitter in order that there may not be any interference caused to other services operating on adjacent channels. This paper describes a filter which has been built for this purpose.*

The filter is placed in the transmission line between the power amplifier and the antenna. In order to insure the absence of reflected energy on the transmission line leading to the filter at any frequency generated by the transmitter, the filter is so designed that the input impedance of the filter is practically a constant for both the pass and rejection band, while the reactance remains essentially zero.

In order to secure the constant-resistance feature throughout the rejection band, the rejected energy is dissipated in water-cooled resistors of a special type which have zero reactance and constant resistance throughout the required band.

The equivalent circuit of the filter is shown. Because of the high carrier frequency as well as the extreme selectivity required, the circuit elements are sections of concentric transmission line. The factors governing the lengths and diameters of these sections are discussed.

The vestigial side-band filter described here has been placed in a practical television installation and has been operating satisfactorily since March, 1939. The tests and observations made at the time of this installation are described.

I. Introduction

THE channels allotted to television stations in this country are limited in width to six megacycles. An early form of channel layout was as shown in the top diagram of Figure 1. The sound carrier was located one-quarter megacycle below the upper edge of the band. The picture carrier was located 2.5 megacycles above the lower edge of the band. With the picture carrier located at this point, the maximum possible modulation frequency was 2.5 megacycles. Any higher frequency modulation would cause side-bands which would lie outside the assigned channel. The receiver was operated in a semi-single-side-band fashion, having the response characteristic shown in the top diagram of Figure 1. It is to be noted that side-bands which lie more than 0.75 megacycle below the picture carrier are not accepted by the receiver and hence serve no useful purpose in forming the picture. This point is of importance in explaining the new channel arrangement.

In the new arrangement, the sound carrier remains unchanged in its position, but the picture carrier is now placed 1.25 megacycles above the lower edge of the band, thus providing an additional space of 1.25 megacycles between picture and sound carriers. The receiver characteristic is modified as shown in the lower sketch of Figure 1. If the transmitter is now modulated with frequencies up to four megacycles, all the frequencies in excess of 1.25 megacycles will cause lower side-bands to be formed which will lie below the lower edge of the assigned band. These side-bands may easily cause interference with other services which may operate in this region. Since these side-bands are not accepted by the receiver, it seems obvious that they may be filtered off at the transmitter.

It is the object of this report to describe a filter which has been built for this purpose. The filter is placed in the transmission line somewhere between the antenna and the power amplifier of the transmitter. In a conventional high-pass filter, without losses, the input impedance would become practically a pure reactance in the rejection band. If this filter were placed directly at the transmitter, this build-up of reactance in the lower side-band region would cause unsymmetrical operation of the final stage, with ensuing generation of transients. Further, if the filter were placed some distance from the transmitter, the reactance in the rejection region would cause reflected energy to be present on the transmission line leading to the filter, with consequent multiple images in the received picture. The filter discussed here has been so designed that the input resistance of the filter is practically a constant for both the pass and rejection band, while the reactance remains essentially zero.[1] Because of this fact, the transmission line between transmitter and filter is terminated in its characteristic impedance for all frequencies, pass and rejection. Then the transmitter looks into a pure resistance over both upper and lower side-bands, and the transmitter generates a double side-band signal, quite unaware of the fact that much of the lower side-band energy is not radiated.

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[1] For an excellent mathematical treatment of constant resistance networks, the reader is referred to "Constant Resistance Networks with Applications to Filter Groups" by E. L. Norton, Bell System Technical Journal, April, 1937, and to U. S. Patent 2,076,248 issued to E. L. Norton.

The lower side-band energy that is not radiated is absorbed in water-cooled resistors which form an integral part of the filter.

The receiver characteristic for the new arrangement is shown in the lower sketch of Figure 1. (Curve *R.*) Here also is shown the desired characteristic of the transmitter filter. (Curve *T.*) Roughly, the requirements are that the filter shall pass all frequencies above a point which is itself 0.5 megacycle above the lower edge of the band,

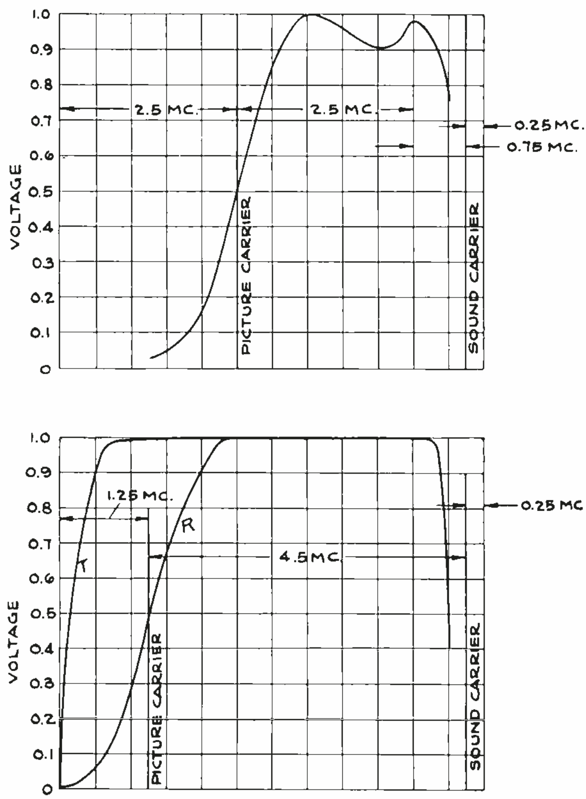


Fig. 1.

and that the attenuation of all frequencies below the lower edge of the band be as great as possible.

It was found that it was not practical at the time to secure this characteristic in a single filter stage. Actually two distinct types of filter are used, with three sections of each type.

II. Type A Filter

The Type A filter is shown schematically in Figure 2. In discussing this filter, we will consider the channel to extend from 44 megacycles to 50 megacycles, since that is the channel for which the first filter was built. Turning then to Figure 2, we may specify the values of the inductances and capacitances shown. It is assumed that the resistance of the antenna (or the input to the antenna transmission line) is 70 ohms. Then the dissipative resistor is made to be 70 ohms. The inductance L_1 is so chosen that it has a reactance of 70 ohms at 44 megacycles, while C_1 is a capacitor having a reactance of 70 ohms at 44 mc. L_2 and C_2 are so proportioned that the combination of the two in a

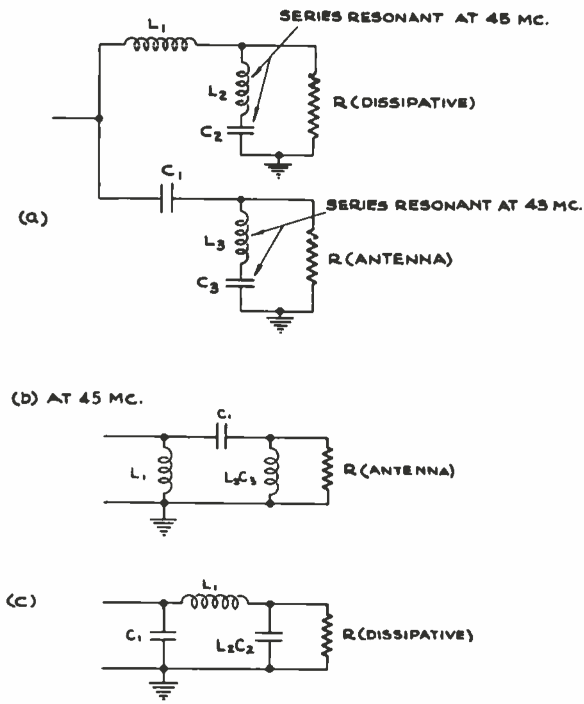


Fig. 2.

series circuit becomes resonant (zero reactance) at 45 megacycles, and at the same time, reaches a capacitive reactance of 70 ohms at 43 megacycles. L_3 and C_3 are chosen so that they yield zero reactance at 43 megacycles, and rise to an inductive reactance of 70 ohms at 45 megacycles.

Let us now examine the circuit at 45 megacycles. The dissipative resistor is short-circuited by the combination of L_2 and C_2, so that no energy goes to the resistor. The reactances of L_1 and C_1 have changed only slightly from 70 ohms. The combination of L_3 and C_3 has placed an inductive reactance of 70 ohms in parallel with the antenna resistance, yielding the equivalent circuit of Figure 2b. Since each of the arms of the Pi network has an impedance of 70 ohms, we have simply a one to one impedance transfer, with an input impedance of 70 ohms, pure resistance, and with a 90-degree phase advance between output and input voltages. All the energy put into the system comes out at the antenna terminals.

At 43 megacycles we have exactly the reverse procedure. The antenna resistance is short-circuited, and the Pi section of Figure 2c is obtained. The input impedance is still 70 ohms, pure resistance, but all of the energy is passed to the dissipative resistor.

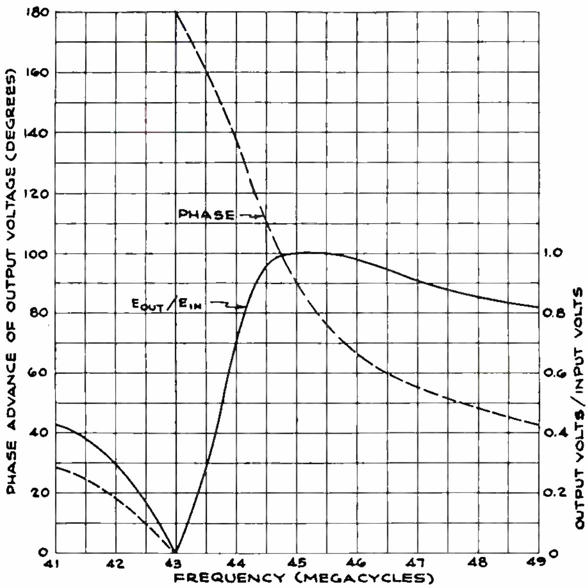


Fig. 3.

We have thus illustrated that the input impedance of the filter is a pure resistance of 70 ohms at 43 and 45 megacycles. As a matter of fact, calculations show that this condition exists for all the frequencies in question.

Figure 3 shows the phase and amplitude characteristics of the Type A filter. We see that we get very great attenuation in the region of 43 megacycles, but that the curve rises rapidly so that at 41 megacycles the attenuation is not very great. To pull down the output in the region from 41 to 43 megacycles, it becomes necessary to place three Type A filters in series. This point will be taken up later in the paper.

Let us turn now to the actual construction of the filter. If we were to use lumped inductances and capacitances, the required values of each element would be as shown in Table I. The construction of elements of this magnitude, accurately tuned to the proper frequency, and able to stand the transmission of power of the order of 10,000 watts, becomes a serious problem, which was briefly considered before turning to the use of concentric-transmission-line elements.

TABLE I

L_1	0.253	microhenries
C_1	51.75	micromicrofarads
L_2	2.72	microhenries
C_2	4.6	micromicrofarads
L_3	2.845	microhenries
C_3	4.81	micromicrofarads

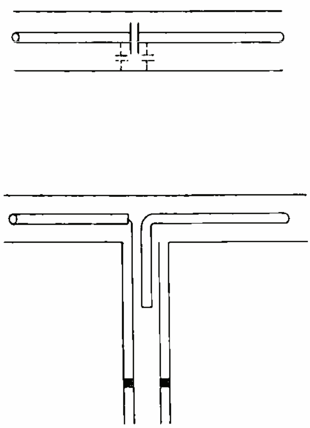


Fig. 4.

Fig. 5.

The capacitance, C_1, is to be inserted as a series element, that is, both terminals are above ground potential. It is well known that a section of concentric transmission line, of 70 ohms characteristic impedance, open-circuited at the far end, and of length equal to one-eighth wave, has an input impedance which is capacitive and has a value of 70 ohms. However, such an element could not be placed in the hot lead because of the capacitance from the outer conductor to ground. To get around this difficulty, the outer conductor of this capacitance element is extended until its length is one-quarter wave, and the remote end is connected directly to ground. This procedure makes the impedance of this outer conductor to ground very high and effectively floats the series capacitance element. The construction is shown in Figure 4.

The inductance element, L_1, is obtained in exactly the same manner except that a shorting plug is placed at the end of the eighth-wave rod.

The series resonant elements which shunt the antenna resistance and the dissipative resistor are formed by single sections of concentric lines, with shorting bars at the far end, as shown in Figure 5. It is known that the input impedance of a transmission line which is shorted at the remote end becomes zero when the length of transmission line is any integral multiple of a half-wave length. As the frequency increases from this critical value, the input impedance becomes

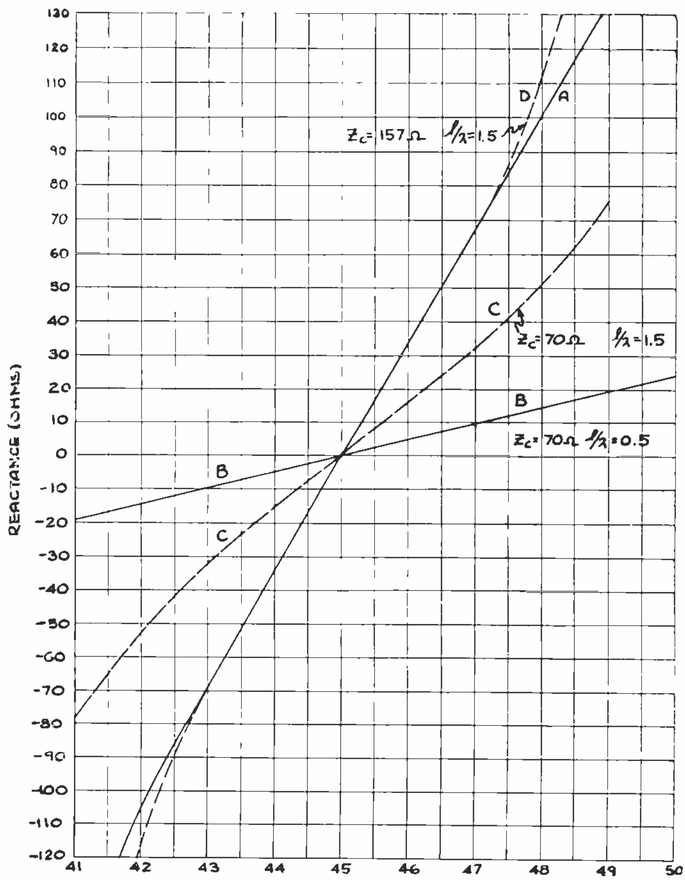


Fig. 6.

inductive, and at frequencies below the critical frequency, the input impedance becomes capacitive. We will consider the case of L_2 and C_2, which have zero reactance at 45 megacycles and reach a capacitive reactance value of 70 ohms at 43 mc. Figure 6 shows several reactance curves, one of which (Curve *A*) shows the calculated curve obtained when L_2 and C_2 have the lumped constants shown in Table I. Curve *B* is the input reactance curve obtained when the line section has a char-

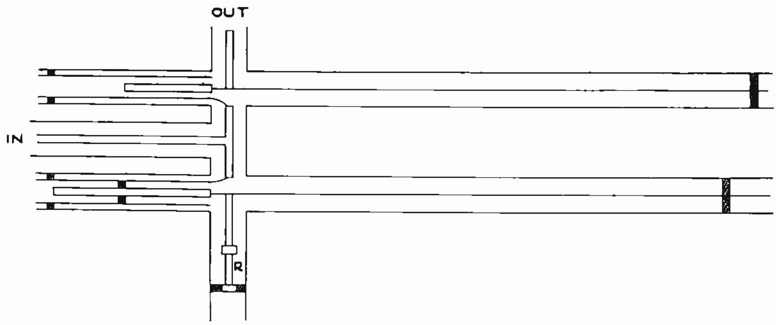


Fig. 7.

acteristic impedance of 70 ohms, and has a length equal to one-half wave. We see that the capacitive reactance at 43 megacycles is only 9.85 ohms. Curve *C* shows the input reactance when the line section has a characteristic impedance of 70 ohms, and is 1.5 waves long. The capacitive reactance at 43 megacycles is now 31.2 ohms. Curve *D* is the input reactance when the line section has a characteristic impedance of 157 ohms and is 1.5 waves long. This curve coincides almost exactly with the desired curve. Therefore, the line element chosen to simulate L_2 and C_2 consists of a shorted transmission line which is 1.5 wave lengths at 45 megacycles (32.8 feet) and which has a characteristic impedance of 157 ohms. The inner diameter of the outer pipe is 4.5 inches, while the inner rod has a diameter of 0.34 inch.

The constructional layout of a single Type A filter is shown in Figure 7. When three of these units are used in series, the point marked "out" on Figure 7 is connected to the input of the following filter.

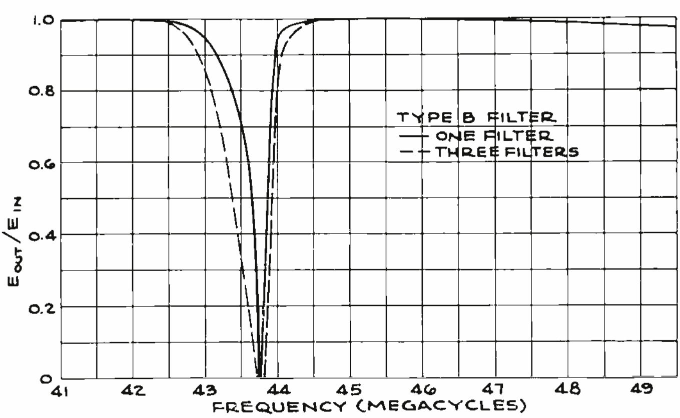


Fig. 8.

III. Type B Filter

When three Type A filters are connected in series, the resulting attenuation curve is satisfactory except for the small region around 43.75 megacycles, where it was felt that the attenuation was not great enough. Accordingly, it becomes desirable to use a filter which has a maximum attenuation point at 43.75 megacycles. In fact, it seemed desirable to simply cut a notch in the pass characteristic at this point.[2] Figure 8 shows the attenuation of such a notching filter (Type B) for the case where a single Type B filter was used and where three such filters were placed in series. These are experimentally determined curves.

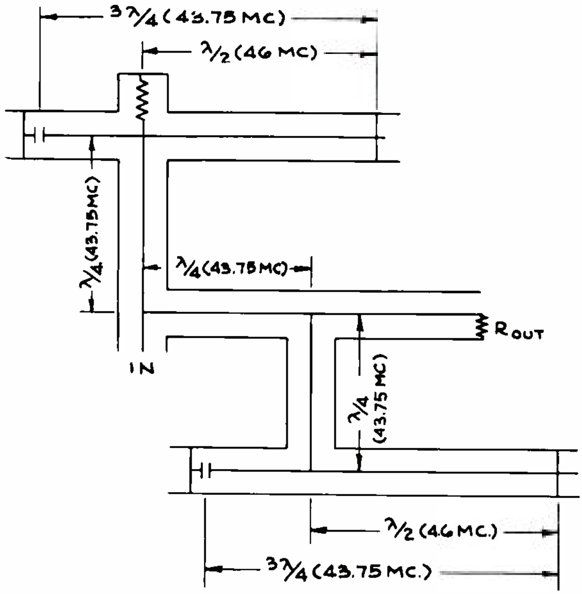


Fig. 9.

The Type B filter is likewise constructed of concentric transmission lines. The filter utilizes a water-cooled resistor in each unit, and each unit possesses the constant input resistance feature. The filter arrangement is shown in Figure 9. The element lengths are shown on this sketch.

IV. Filter Tests

During the early development work, a single Type A and a single Type B filter unit were built and tested in Camden. The measured

[2] This notching filter becomes increasingly important when used on the 50 to 56 megacycle channel, for then the maximum attentuation point would occur at 49.75 megacycles, the sound carrier of the 44 to 50 megacycle television channel.

amplitude characteristic of the Type A filter was so close to the calculated characteristic shown in Figure 3 that it seems unnecessary to repeat it here. As stated above, Figure 8 shows the measured response of the Type B filter. The input impedance of these two units in series was found to be quite satisfactory. Tests were made in which the filter

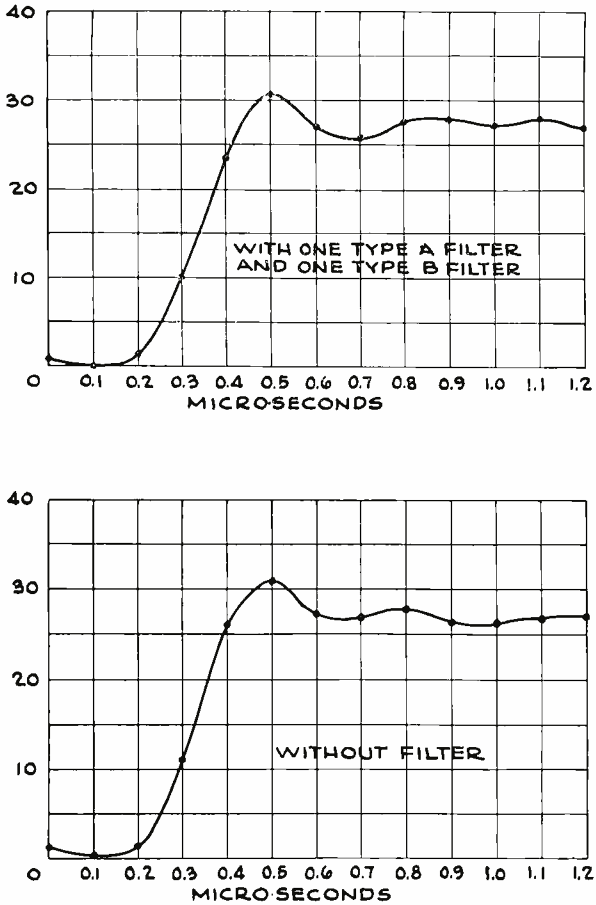


Fig. 10.

combination was fed with a signal generator which was modulated by means of a standard test pattern. A television receiver was placed at the output of the two filters. Observations were made of the received test pattern. The filter was next removed from the system, and the signal generator fed directly to the receiver. The test pattern without the filter was then observed. Most observers agreed that there was no essential difference between the two pictures.

The signal generator was modulated with a square wave, and the response at the second detector of the receiver recorded on a cathode ray oscilloscope. The filter was then placed in the circuit, and the square-wave response again observed. The results for the two cases are shown in Figure 10.

The tests on these single filter units were satisfactory in every respect. Of course, the attenuation, while in strict agreement with the calculated values, was not as great as necessary for the problem at hand. Accordingly, two more Type A filters and two more Type B filters were constructed.

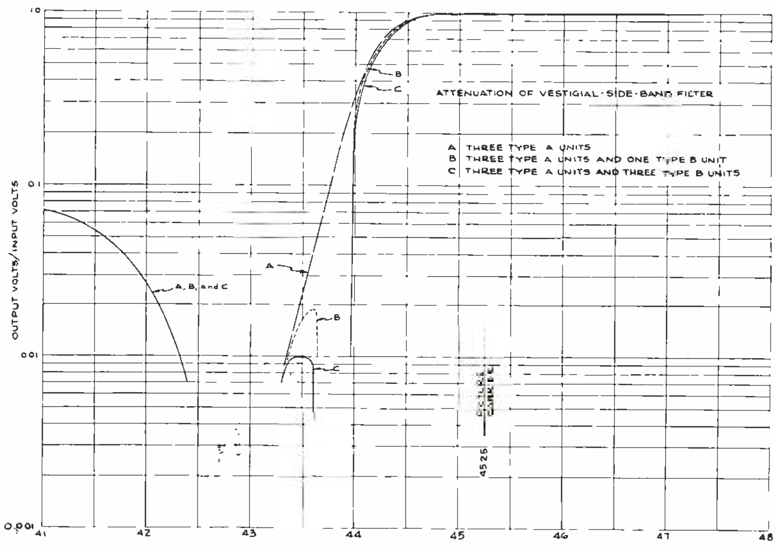


Fig. 11.

With three filters of each type, the response to a square wave was again determined and the test pattern was observed. These tests were again satisfactory. The overall attenuation characteristic as measured is shown by Figure 11. The results shown by this curve were considered to be adequate.

Observations of standing waves on the 70-ohm feed line leading from the test oscillator to the filter showed that the filter system offered a very good termination to the transmission line. The observed reflection on the feed line is shown as a function of frequency in Figure 12.

The entire set of filters as set up in Camden is shown in Figure 13. The input to the filter is at the extreme left rear and is not visible in the picture. Three Type A filters, stacked one above the other, are shown on the left, while the three Type B filters are on the right.

After a thorough series of tests in Camden, the filters were dismantled and reassembled on the eighty-fifth floor of the Empire State Building, New York City, where tests could be made using the television transmitter of the National Broadcasting Company. Many tests

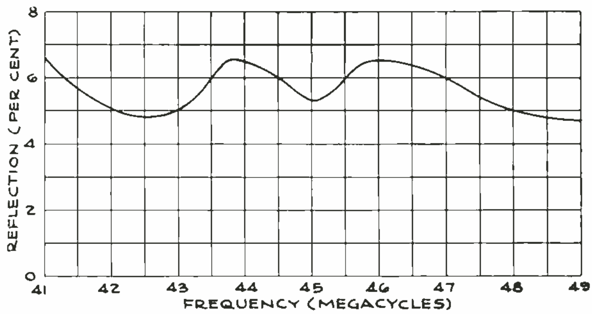


Fig. 12.

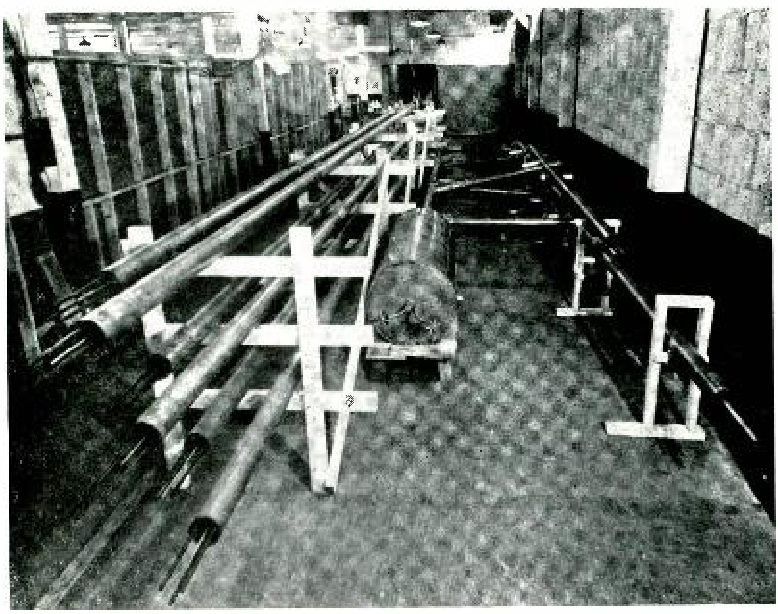


Fig. 13.

were made in 1938 and early in 1939. Field observations were made for a time using a temporary antenna located on the side of the building at the same level as the eighty-fifth floor. During March, 1939, the filter system was connected to the new antenna[3] which had just

[3] Nils E. Lindenblad, "Television Transmitting Antenna for Empire State Building," RCA Review, April, 1939.

been constructed on the top of the Empire State Building. On April 30, the new system went into regular operation.

The new top antenna is fed by two 55-ohm lines operated in push-pull. However, the output of the filter is arranged to feed into a single 70-ohm line. Accordingly, a matching drum was constructed to transfer from a single-ended 70-ohm line to the two push-pull 55-ohm lines. A sketch of this drum is shown on Figure 14. We see from the left hand (a) sketch that it would not be possible to drive the two push-pull lines with equal and opposite voltages unless means are provided to effectively isolate the end of the outer conductor of the 70-ohm

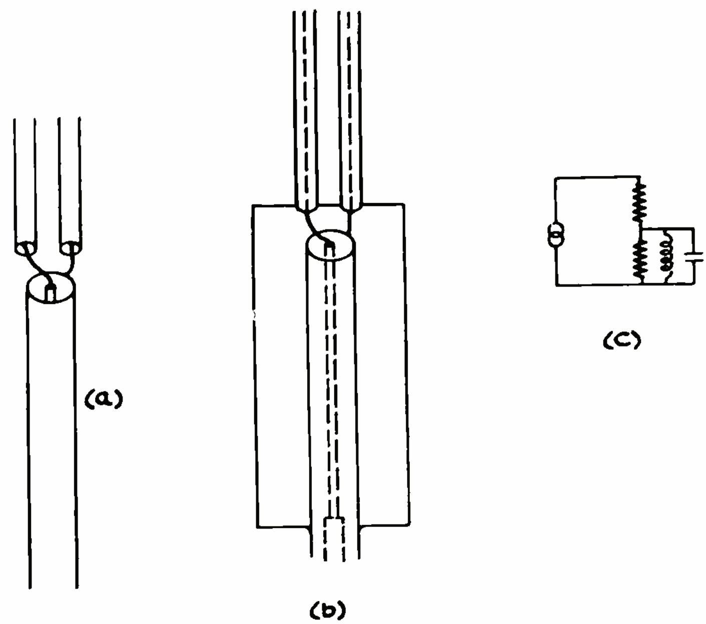


Fig. 14.

line. A quarter wave isolating section is placed around the 70-ohm line, as shown in the middle (b) sketch of Figure 14. The equivalent circuit, which determines the degree of equality of the voltages on the two 55-ohm lines, is shown on the right side (c) of Figure 14. This type of converting drum is described by Lindenblad.[4] In our particular case, the diameter of the outer drum was made large in order to secure high impedance for this shunting or isolating circuit over a wide band of frequencies. Figures 15 and 16 show views of the drum which is used at the Empire State Building. The drum may also be seen in

[4] RCA Review, April, 1939, page 401.

Figure 13. Our tests showed that this method of going from single-ended to push-pull was quite satisfactory, even for the wide range of frequencies required in a television system.

Reference has been made earlier in the paper to the water-cooled resistors used in the filters. These resistors were designed especially for this application. Essentially, the resistors form the inner con-

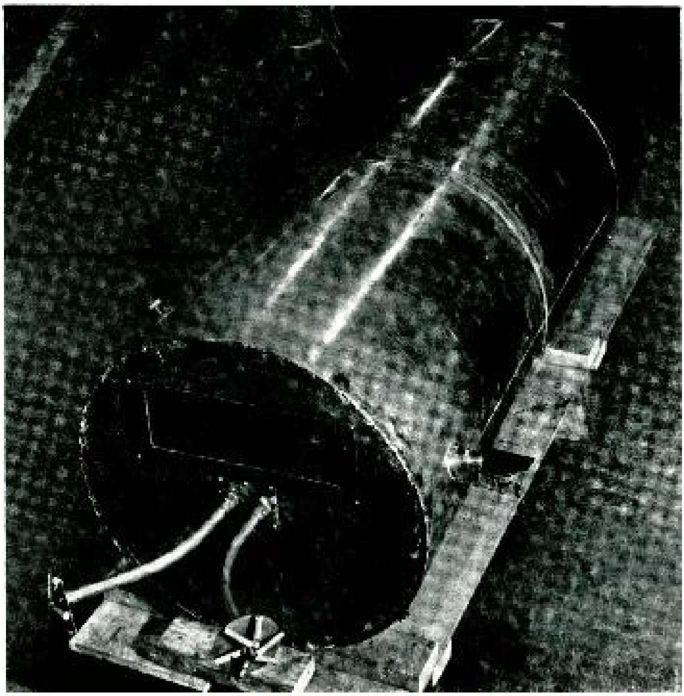


Fig. 15.

ductor of a transmission line. One end of the resistor is shorted to ground. Water is sent through this shorting plug, through the middle of a ceramic tube which supports the resistive film, and back over the surface of the resistor tube. The resistor unit, together with the shorting plug, is shown in Figure 17. As may be seen, the resistor is encased in a glass tube so that the water will return over the surface of the resistor. The theory behind the design of this type of resistance termination is given elsewhere.[5] These resistors form a very good termination for the 70-ohm line. In addition, a rather large amount of power may be dissipated with a small amount of water flow. Tests

[5] G. H. Brown and J. W. Conklin, "Low-Reactance High-Frequency Resistors."

have shown that the resistors will safely handle a power of one kilowatt with a water flow of slightly more than one gallon per minute. Six of these resistors have been in regular use at the installation in the Empire State Building, without any failures to date.

Early in March, 1939, the transmitter was operated into the top antenna with the filters removed from the circuit. A short time later,

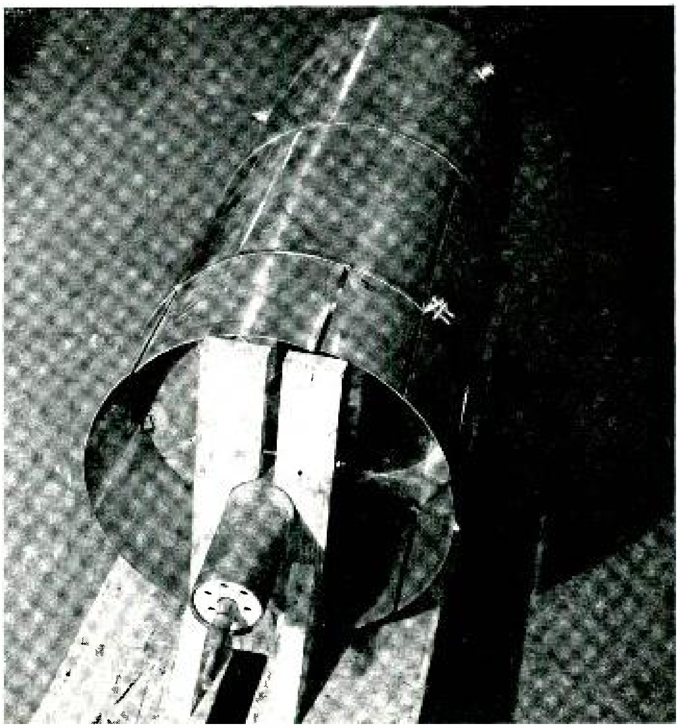


Fig. 16.

the filter was placed in the system, with a resulting transmitted picture that was unimpaired by the insertion of the filter.

V. Conclusion

The filter systems described in this report make possible the use of higher modulation frequencies, that is, the limited assigned band is utilized to the best advantage. The filter was built and tested in the laboratory, where almost ideal conditions existed. Later, the filter was installed in a high power television system, and is now in operation in a practical system.

APPENDIX I

A Consideration of the Energy Distribution in the Side-Bands of a Television Signal

While our general experience and observations indicate that the major portion of the radiated energy of a television picture lies in the frequency spectrum very close to the carrier frequency, and that the energy dissipated in the filter resistors is small, it seems desirable to examine this distribution for a few typical cases. We will consider the case of a picture which consists of alternate vertical black and white bars of equal width. The superimposed synchronizing signals will be neglected. Then the radio frequency signal will be as shown in Figure 18. This signal consists of a sine-wave voltage which has a

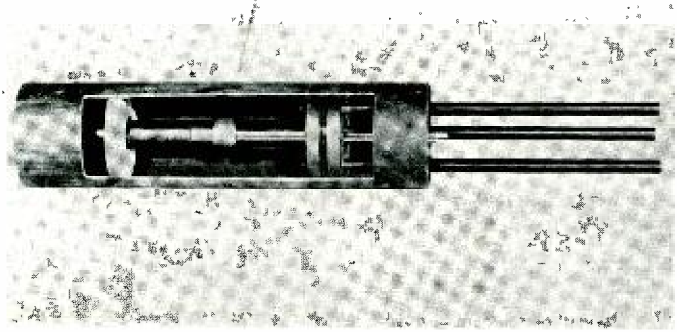


Fig. 17.

frequency equal to the frequency of the carrier. This sine wave of voltage exists for a period of time T, becomes zero for an equal period, and then repeats. This signal may be analyzed by the method of Fourier. It will then be found[6] that the signal may be represented by a carrier frequency and upper and lower side-bands equally distributed on both sides of the carrier. The amplitudes of these side-bands are represented by a simple relation.

If the voltage of the intermittent sinusoidal voltage is E, the carrier voltage, of frequency f, has a magnitude of $\frac{E}{2}$. The first upper side-band has a frequency of $f + \frac{1}{2T}$ and a magnitude of $\frac{E}{\pi}$. The next

[6] L. J. Peters, "Theory of Thermionic Vacuum Tube Circuits," McGraw-Hill, 1927. Pages 124-127.

upper side-band has a frequency of $f + \frac{3}{2T}$ and a magnitude of $\frac{E}{3\pi}$.

The lower side-bands follow similar relations. The pertinent relations are shown in the following table. (If T is expressed in microseconds, the frequencies will be given in megacycles.)

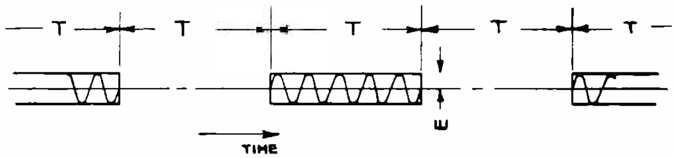


Fig. 18.

TABLE II

Frequency	*Voltage Magnitude*	*Watts in each side-band when peak power is 10,000 watts (see note)*
$f - \frac{15}{2T}$	$\frac{E}{15\pi} = -0.0212E$	4.5
$f - \frac{13}{2T}$	$\frac{E}{13\pi} = 0.0245E$	6.0
$f - \frac{11}{2T}$	$\frac{E}{11\pi} = -0.0289E$	8.35
$f - \frac{9}{2T}$	$\frac{E}{9\pi} = 0.03535E$	12.5
$f - \frac{7}{2T}$	$\frac{E}{7\pi} = -0.0455E$	20.7
$f - \frac{5}{2T}$	$\frac{E}{5\pi} = 0.0637E$	40.6
$f - \frac{3}{2T}$	$\frac{E}{3\pi} = -0.1062E$	112.8
$f - \frac{1}{2T}$	$\frac{E}{\pi} = 0.3185E$	1012.0
f (carrier)	$\frac{E}{2} = 0.5E$	2500.0

$f + \dfrac{1}{2T}$	$\dfrac{E}{\pi} = 0.3185E$	1012.0
$f + \dfrac{3}{2T}$	$\dfrac{E}{3\pi} = -0.1062E$	112.8
$f + \dfrac{5}{2T}$	$\dfrac{E}{5\pi} = 0.0637E$	40.6
$f + \dfrac{7}{2T}$	$\dfrac{E}{7\pi} = -0.0455E$	20.7
$f + \dfrac{9}{2T}$	$\dfrac{E}{9\pi} = 0.03535E$	12.5
$f + \dfrac{11}{2T}$	$\dfrac{E}{11\pi} = -0.0289E$	8.35
$f + \dfrac{13}{2T}$	$\dfrac{E}{13\pi} = 0.0245E$	6.0
$f + \dfrac{15}{2T}$	$\dfrac{E}{15\pi} = -0.0212E$	4.5

Note: By a peak power of 10,000 watts, we mean that E is of such a value that the power in the wave would be 10,000 watts if the sinusoidal wave were continuous, not interrupted for the period, T. Because the wave is cut off half of the time, the actual total power for the wave shown in Fig. 2D will be 5000 watts for a peak power of 10,000 watts. If, in Table II, we sum the power in the carrier, the first eight upper side-bands, and the first eight lower side-bands, we find a power of 4939.9 watts.

The power absorbed by the filter resistors may be readily computed from Table II. We assume that the carrier frequency is 45.25 megacycles. Then we assign a definite value for T. We then turn to Table II and see which side-bands lie between 41 and 44 megacycles and sum up the power which is assigned to these side-bands. The results of this summation are shown in Figure 19 as a function of the time interval T. We see particularly that for vertical bars which last more than 0.4 microsecond, the filter resistors must handle less than 200 watts for a peak power of 10,000 watts.

Evidently, the severest condition occurs when the transmitted picture consists of vertical bars which are between 0.36 and 0.4 microsecond wide. Then, the filter resistors must be able to handle 1125 watts. It seems reasonable to assume that average picture conditions will be closer to the 200-watt region.

It is interesting to note that 100 per cent modulation of the 2500-watt carrier, at a modulating frequency of 2.25 megacycles, causes only 625 watts to be sent into the filter resistors.

We now have sufficient data to examine the interesting case of two television transmitters operating in adjacent bands. Specifically, the No. 1 channel[7] lies from 44 to 50 megacycles with picture carrier at 45.25 megacycles, while No. 2 channel lies between 50 and 56 megacycles with picture carrier at 51.25 megacycles. From Table II we can

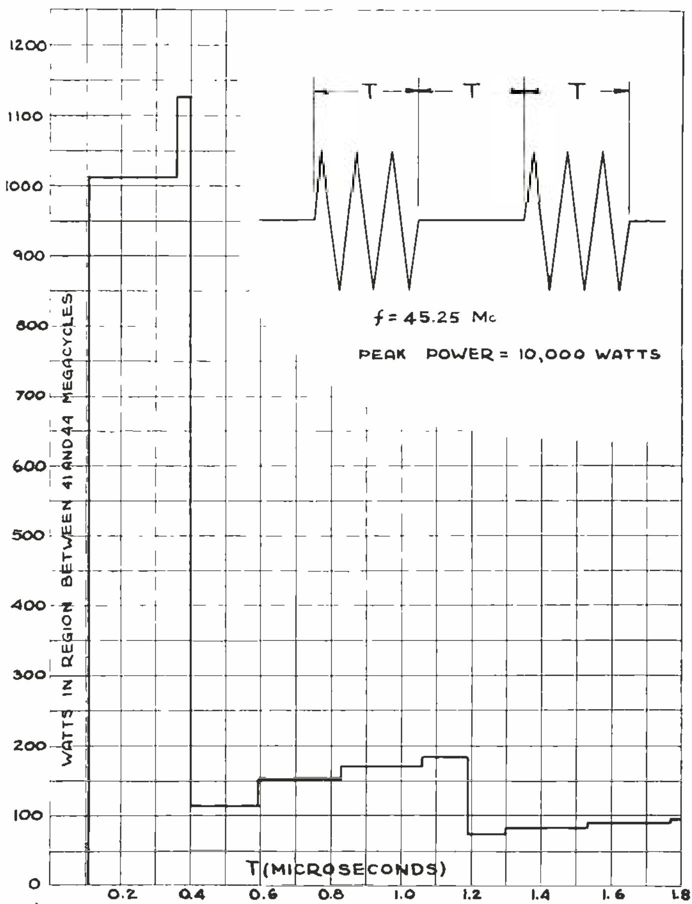


Fig. 19.

easily obtain the side-band energy distribution of either transmitter. We will consider each transmitter to have a peak power of 10,000 watts. In constructing Figure 20 we assumed each transmitter to be

[7] Since this paper was written, the television channel assignments have been changed. All references to channels in this paper are based on the old channel assignments that existed in 1939.

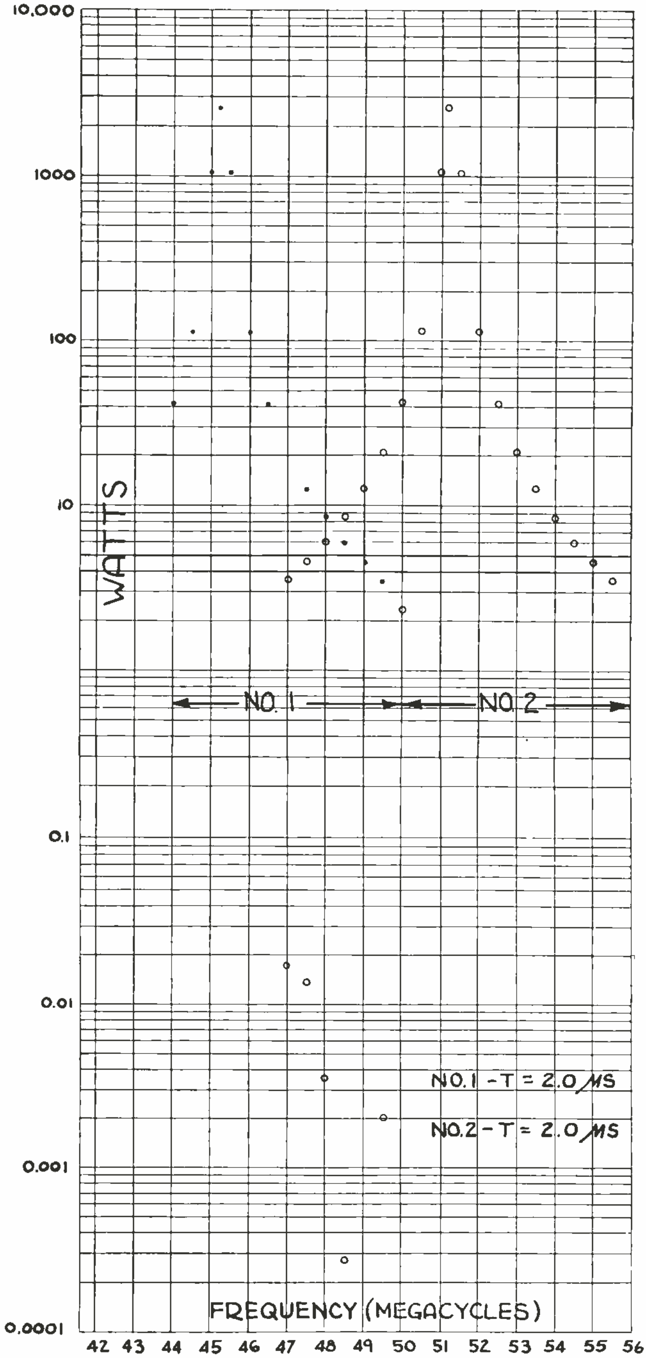

10,000
1000
100
10
0.1
0.01
0.001
0.0001
WATTS
NO. 1
NO. 2
NO. 1 - T = 2.0 MS
NO. 2 - T = 2.0 MS
FREQUENCY (MEGACYCLES)
42 43 44 45 46 47 48 49 50 51 52 53 54 55 56

Fig. 20.

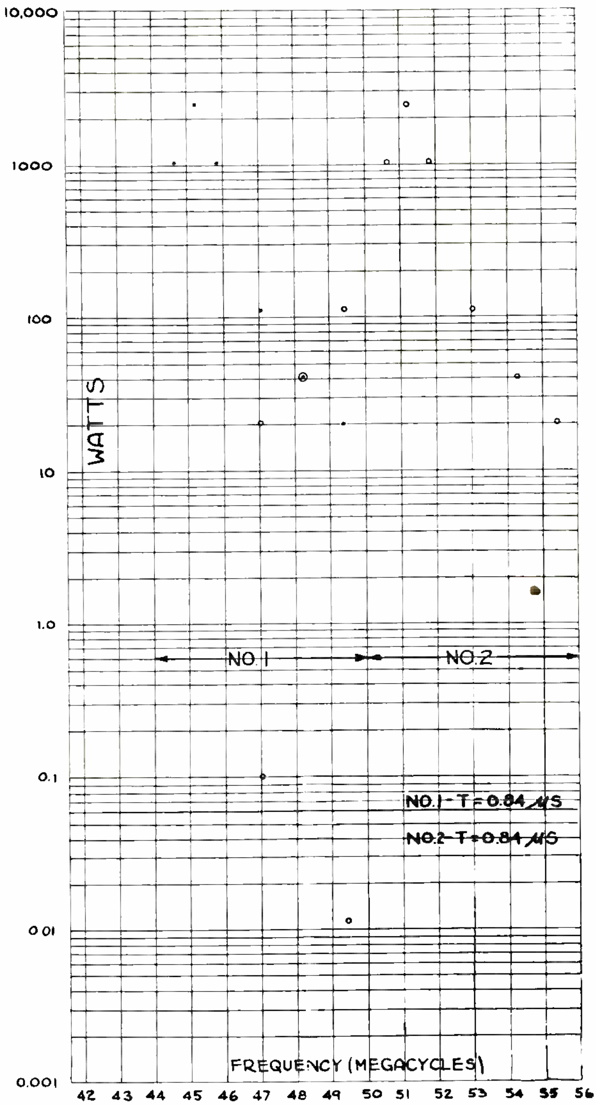


Fig. 21.

sending out a picture made up of vertical alternate black and white bars, the duration of each bar being 2.0 microseconds. The dots on Figure 20 indicate the energy distribution throughout the No. 1 channel due to the No. 1 transmitter. The circles show the energy distribution due to the No. 2 transmitter. We see that the No. 2 transmitter would radiate a strong signal in the No. 1 channel. The lower set of

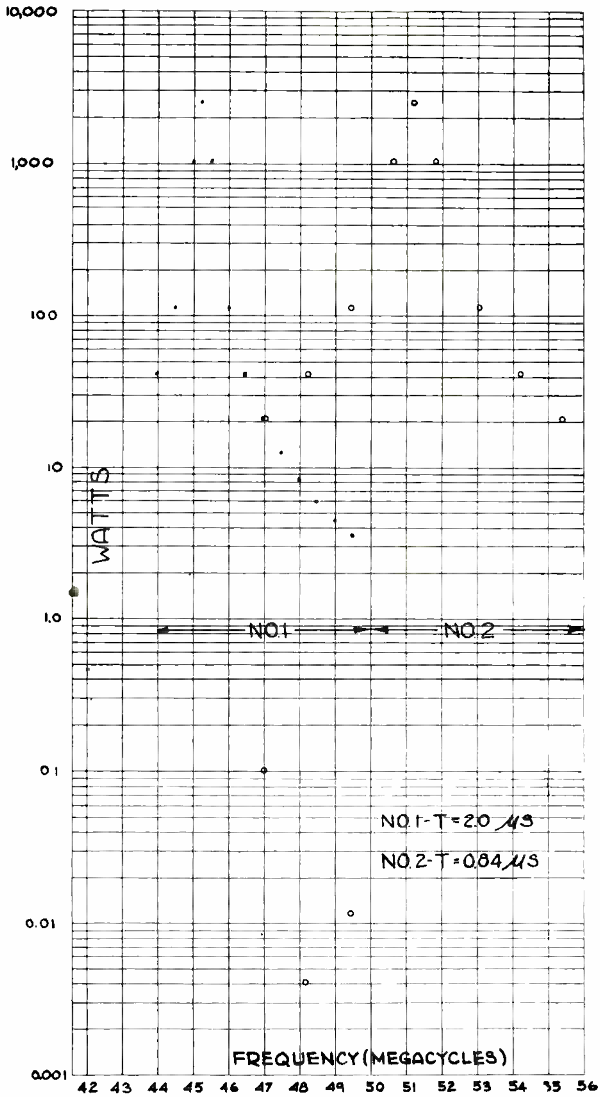


Fig. 22.

circles on Figure 20 shows the disturbing energy when the side-band filter is in operation. In this case there seems to be ample protection.

Figure 21 is similar to Figure 20 except that the width of the modulating bars is now 0.84 microsecond. In Figure 22 the width of bars for the No. 1 transmitter is 2.0 microseconds while No. 2 transmitter is modulated with 0.84 microsecond bars. The need for a side-band filter is quite evident from these figures.

APPENDIX II

Input Impedance of Type A Filter

For ease of reference, the Type A filter schematic diagram is repeated in Figure 23. We have already shown how transmission line elements may be used to take the place of the lumped elements shown in Figure 23. To compute the various circuit relations we use the following equations.

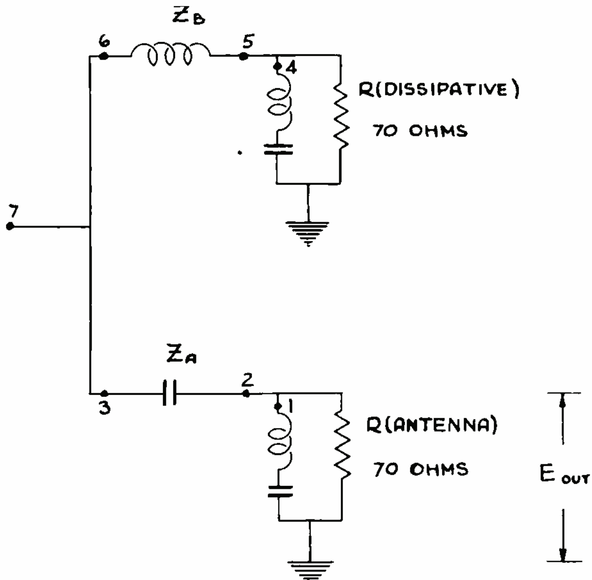


Fig. 23.

$$Z_A = -j70 \cot\left(\frac{f_{mc}}{44} \times 45°\right)$$

$$Z_B = +j70 \tan\left(\frac{f_{mc}}{44} \times 45°\right)$$

$$Z_1 = +j157 \tan\left(\frac{f_{mc}}{43} \times 540°\right)$$

$$Z_4 = +j157 \tan\left(\frac{f_{mc}}{45} \times 540°\right)$$

We may then compute the impedance at the input or at any other point. At the same time we may compute the attenuation through

TABLE III

f_{mc}	Z_A	Z_B	Z_1	Z_2	Z_3	Z_4	Z_5	Z_6	Z_7	Per Cent Reflection	$E_{out}/\sqrt{P}$
41	—j77.8	+j63	—j70	35—j35	35—j112.8	—j174.5	60.4—j24.2	60.4+j38.8	70—j2.73	1.95	0.42
42	—j75.2	+j65.2	—j35	14—j27.9	14—j103.1	—j119	52.2—j30.7	52.2+j34.5	68.5—j3.5	2.7	0.30
43	—j72.5	+j67.6	0	0	—j72.5	—j70	35—j35	35+j32.6	65.5+j1.9	3.65	0
44	—j70	+j70	+j35	14+j27.9	14—j42.1	—j35	14—j27.9	14+j42.1	70.2+j0	0	0.707
45	—j67.6	+j72.5	+j70	35+j35	35—j32.6	0	0	+j72.5	65.5—j1.9	3.65	1.0
46	—j65.2	+j75.2	+j120.5	52.4+j30.25	52.4—j35	+j35	14+j27.9	14+j103.1	68.5+j3.5	2.7	0.98
47	—j63	+j77.8	+j186.5	61.8+j23	61.8—j40	+j70	35+j35	35+j112.8	71.6+j3.6	2.8	0.91
48	—j61	+j80.5	+j295	66.2+j15.8	66.2—j45.2	+j114	50.7+j31	50.7+j111.5	73+j2.08	2.4	0.85
49	—j59.7	+j82.2	+j588	69.5+j8.2	69.5—j51.5	+j174.5	60.5+j24.2	60.5+j106.4	75+j1.2	3.5	0.815

TABLE IV

f_{mc}	Z_1	Z_2	Z_3	Z_4	Z_5	Z_6	Z_{in}	Per Cent Reflection	$E_{out}/\sqrt{P}$	$E_{res}/\sqrt{P}$
41	0.0428∠+87.5°	18.55∠+88°15′	+j0.0428	+j18.55	1∠+3°	1∠—1°	1.0∠+2°	1.7	1.0	0.04
43	0.259∠+74°50′	3.86∠—74°50′	+j0.269	+j3.72	0.965∠—15°	1.035∠+15°	1.0∠—5°	4.3	0.965	0.26
43.5	0.638∠+50.5°	1.565∠—50.5°	+j0.825	—j1.21	0.77∠—39.5°	1.298∠+39.5°	1.0∠0°	0	0.775	0.64
43.75	1.0∠0°	1.0∠0°	∞	0	0	∞	1.0∠0°	0	0 (0.025)	1.0
44	0.717∠—44°10′	1.392∠+44°10′	—j1.03	+j0.97	0.697∠+45°50′	1.435∠—45°50′	1.0∠0°	0	0.63	0.73
46	0	12.7∠—90°	0	—j12.7	0.998∠—4.5°	1.0∠0°	0.998∠—4.5°	3.9	1.0	0
49	0.135∠+82.5°	3.16∠—86.5°	+j0.135	—j3.14	0.952∠—17.5°	1.17∠+15°35′	1.18∠—6°	9.65	0.99	0.14

the filter in a very simple manner. The power into the antenna resistor is

$$P_{\text{out}} = \frac{E^2_{\text{out}}}{70}$$

Also, this power is equal to the power fed to this network at Point 3 so that

$$P_{\text{out}} = E^2_{\text{in}}\, G_3$$

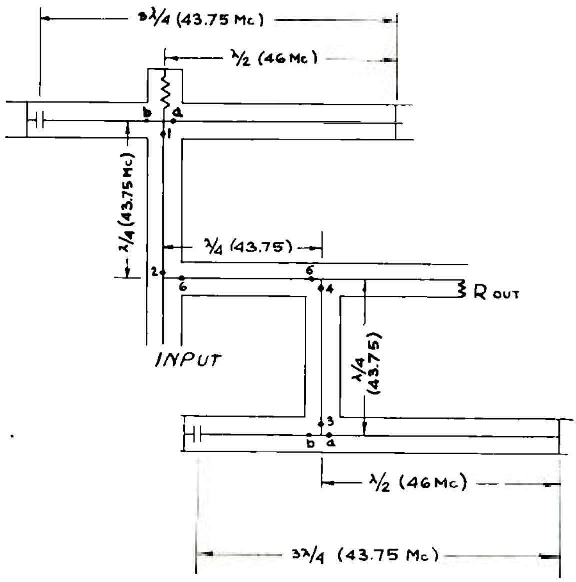


Fig. 24.

where G_3 is the conductance at Point 3. Then

$$E_{\text{out}}/E_{\text{in}} = \sqrt{70G_3}$$

The power into the total sytem is

$$P_{\text{in}} = E^2_{\text{in}}\, G_7$$

For a constant power input,

$$\frac{E_{\text{out}}}{\sqrt{P}} = \sqrt{\frac{70G_3}{G_7}}$$

We see that Z_7 is practically a pure resistance of 70 ohms, so that G_7 is close to 1/70 in value. Under this condition a constant input voltage corresponds to a constant input power so we may use either equation to determine the output voltage. The results of these calculations are given in Table III. Also, we have included in this table, a column labelled "Per Cent Reflection." This is the amount of reflected voltage wave given in terms of the incident voltage wave on the concentric feed line to the filter.

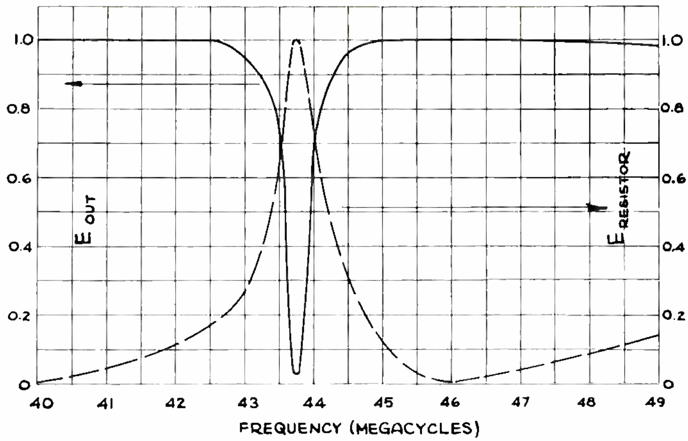


Fig. 25.

APPENDIX III

Input Impedance of Type B Filter

The Type B filter is shown in Figure 24. In computing the impedances of this filter, we may simplify the work by remembering that the resistors have the same value as the characteristic impedance of all the line elements. Then we may take the resistor value and the characteristic impedance as unity. Table IV shows the various impedances involved, in terms of unity starting resistance. For 70 ohm lines and resistors we simply multiply the values in Table IV by 70. Figure 25 shows the calculated values of the voltage on the output and the voltage on the dissipative resistor as a function of frequency. In making these computations, we made use of the relations

$$Z_a = +j1.0 \tan\left(\frac{f_{mc}}{46} \times 180^\circ\right)$$

$$Z_b = -j1.0 \cot\left(\frac{f_{mc}}{43.75} \times 99^\circ\right)$$

A TRANSMITTER FOR FREQUENCY-MODULATED BROADCAST SERVICE USING A NEW ULTRA-HIGH-FREQUENCY TETRODE

By

A. K. Wing

RCA Manufacturing Company, Inc., Harrison, N. J.

and

J. E. Young

RCA Manufacturing Company, Inc., Camden, N. J.

Summary—*This article describes a 1-kw transmitter developed for the new frequency-modulated broadcast service recently opened up by the Federal Communications Commission. A description of the circuits, tubes and mechanical arrangement of the apparatus is included. The electrical performance is described.*

THE following is a description of a 1-kw frequency-modulated transmitter. The method used to obtain stabilized frequency modulation was developed by M. Crosby of the R.C.A. Communications, Inc., and has the advantage of simplicity combined with a high degree of performance. Figure 1 is a simplified schematic diagram of the complete transmitter. An 807 tube is used as an oscillator. Its frequency is varied by two 807's to accomplish the desired frequency modulation. The oscillator tube is designated as the modulated oscillator in the diagram. It will be observed that two modulator tubes are connected with their plates in parallel, while their grids are supplied with radio-frequency energy from the oscillator tank circuit with a phase difference of 180 degrees from grid to grid. The audio-frequency modulating voltage is also introduced into the grid circuits 180 degrees out of phase so that under quiescent conditions the modulator tubes draw equal and oppositely phased currents from the tank circuit. An audio-frequency signal disturbs this balance, causing one tube to draw more current and the other less, thereby changing the effective reactance across the oscillator tank circuit. The phase of the currents flowing in the plate circuits of the modulator tubes is controlled by the grid tank and link coupling circuit and is adjusted to be 90 degrees out of phase with the oscillator tank voltage producing the current. The modulators, therefore, look like a reactive load to the tank circuit and this reactance varies with the amplitude of the audio-frequency voltage impressed on the modulator grid.

To maintain the degree of frequency stability required by the regulations of the Federal Communications Commission covering frequency-modulation stations, means are provided to hold the average carrier frequency within very close limits. The regulations require that the stability be better than plus or minus 2000 cycles in the frequency band between 42 and 50 megacycles. This equipment will hold the average carrier frequency to within better than 0.0025 per cent of its assigned value. This is accomplished by the circuits shown to the left of the modulator tubes in the schematic diagram. Briefly, a separate quartz crystal-controlled oscillator is arranged so that its output excites one grid of a Type 1613 mixer tube. The other grid of this tube is supplied with energy from the radio-frequency stage following the oscillator. The plate circuit of the mixer is tuned to the difference frequency between these two signals. The frequency of the crystal is chosen so that this difference frequency will be 1.0 megacycle. The output of the mixer is coupled to a Type 6H6 rectifier tube through a discriminator circuit. The output of the rectifier is in turn connected into the grid circuit of the modulator tubes. The discriminator circuit is so set up and tuned that no controlling voltage is obtained on the grids of the modulator tubes so long as the mixer output frequency is exactly 1.0 megacycle. If the oscillator frequency varies, the beat note between the highly stable oscillator and the modulated oscillator will vary in the same way, thus setting up a voltage on the grids of the modulator tubes which tends to minimize this change. The control ratio is such that the net frequency change is approximately 1/20 of the change which would have resulted had the oscillator been uncorrected.

The use of two tubes as a differential modulator gives the circuit a compensating effect and tends to balance out circuit disturbances or irregularities which exist simultaneously in both modulator tubes. Thus, the effect of alternating-current filament supplies or a variation in plate potential due to either transients or rectifier ripple are almost completely cancelled out. An effect somewhat similar to that existing in a push-pull audio system also exists in that even harmonic distortions tend to cancel out and the fidelity of the system is therefore improved.

Two audio-frequency input circuits are provided, one flat from 30 to 15,000 cycles, for high-fidelity audio transmission, the other flat from 30 to 25,000 cycles to provide for the multiplexed transmission of facsimile on a sub-carrier frequency, as well. Filter circuits are provided in the automatic-frequency-controlling circuit to give this circuit sufficient time lag so that it will not be affected by the modulating signal. The automatic-frequency-controlling circuit, therefore, only tends to maintain the average carrier frequency constant and does not come into operation on the frequency excursions caused by modulation.

The radio-frequency circuits used are, in the most part, conventional, although the frequency range is from 30 to 108 megacycles. By careful positioning of the circuit elements it was possible to use lumped constants in all of the tank circuits. Inductive coupling is used between each pair of stages except between the tripler and intermediate power amplifier where a combination of inductive and capacitive coupling is used. The arrangement of the circuits associated with the output tubes is somewhat unique and can best be described in connection with the tubes themselves.

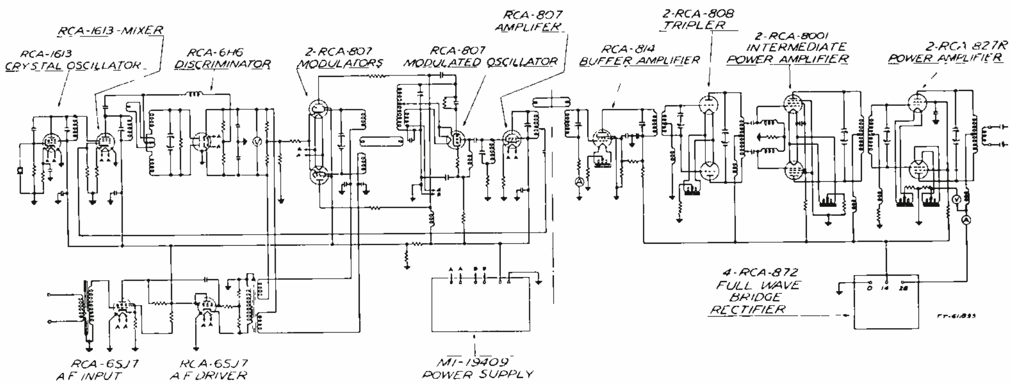


Fig. 1

The output tube used in this transmitter is the RCA-827R which was developed to provide 500 watts of radio-frequency output over the frequency range, and is capable of 100 per cent amplitude modulation under this condition of power output. It was first designed for use in the sound channel of a 1-kw television transmitter. However, it fitted so well into the requirements of the 1-kw frequency-modulated transmitter that it was also used for this purpose. Figure 2 is a photograph of the tube. Use is made of an external anode structure equipped with a fin assembly for forced air-cooling. This construction minimizes plate lead inductance. Forced air-cooling allows high anode dissipations to be obtained as compared with conventional air-cooled designs; the stream of cooling air also allows the glass envelope to be made smaller than would otherwise be possible. This permits the tube to be compact, allows high-frequency operation, yet permits adequate power capability. The general arrangement of the tube terminals can be seen in Figure 2. Two control-grid terminals and two filament terminals are provided. The screen grid is connected to the ring seal at the top of the glass envelope of the tube.

The tube was designed to have the characteristics of a beam tetrode and the principles of electron optics were utilized to minimize screen current and to provide characteristics approaching those of a pentode. A novel screen-grid construction was used to reduce the inductance of the screen grid and the screen-grid connections. The various features of mechanical construction by which these characteristics are obtained may be observed by reference to Figure 3 which is a cross-section of the tube and shows the construction in more detail. The basis of the construction is the use of a metal header in place of the more conventional glass stem or dish. This header is drawn from "Kovar,"

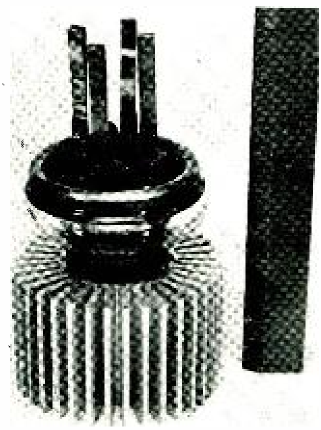


Fig. 2

shaped as shown to provide suitable flexibility at the outer edge for satisfactory sealing and to prevent deformation under atmospheric pressure when the tube is exhausted. This header serves as a low-inductance terminal for the screen which is mounted directly on it by means of a continuous conical support. External contact is made to the edge of the header by a series of springs which slide down over the top of the header section.

The connections to the control-grid and filament are brought through the header by smaller "Kovar"-to-glass seals of the type shown in the cross-section. It should be noted that all electrodes are supported from the header alone, no solid insulation being used between elements within the tube. Dielectric losses which would impair the performance of the tube are thereby eliminated. Rigidity of the electrodes is assured by the short, large diameter leads supporting them. This lead structure aids in minimizing lead inductance and losses; connecting the two control-grid leads in parallel further reduces such effects. External connection to the control-grid and filament is by flexible copper ribbons.

The construction of both the control and screen grids utilizes parallel, vertical wires. The orientation of the two grids is such that the wires of the screen grid are located in the shadow of the wires of the control grid, thus forming electron beams between the wires which reduce considerably the current collected by the screen grids as compared with a structure having random alignment. A novel feature of the construction of the control grid is the use of a graphite spacer at the free end of the grid. This takes the form of a washer with a series of small holes around its edge to receive the grid wires. Three of the wires support the ring while the remainder are free to move longitudinally in the holes. In this way each wire can expand independently of the others so that any tendency toward unequal expansion does not cause the wires to buckle, thus eliminating possible internal shorts or changes in characteristics. The heat of the filament is concentrated because of its compact arrangement, and since the spacing between the grid and the filament is small, special precautions

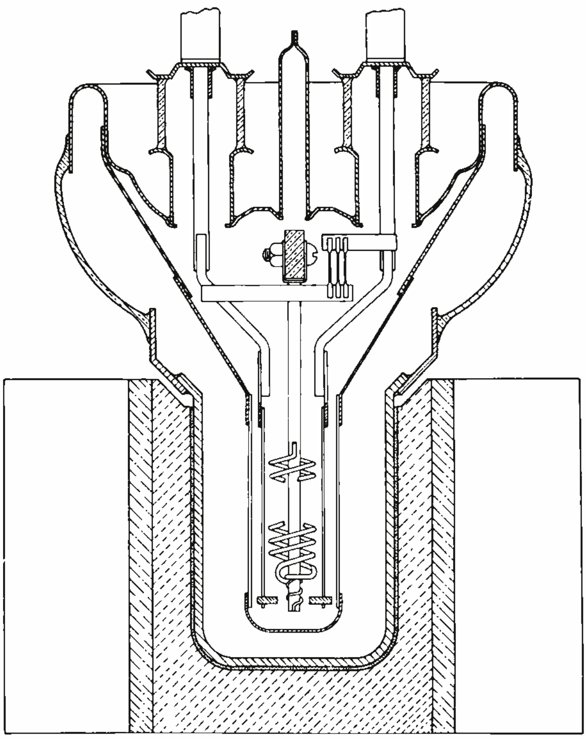


Fig. 3

are required to avoid undesirable emission from the control-grid. This grid is coated with zirconium to improve its heat-radiating properties and to reduce the tendency toward primary emission from the surface.

The filament of the tube is of thoriated tungsten and is helical in form. The filament-to-grid and grid-to-screen spacings are approximately 50 and 120 thousandths of an inch, respectively, resulting in short electron transit time. The spacing between screen-grid and anode is approximately 360 thousandths of an inch, providing a potential minimum due to space charge in this space. At a frequency of 100

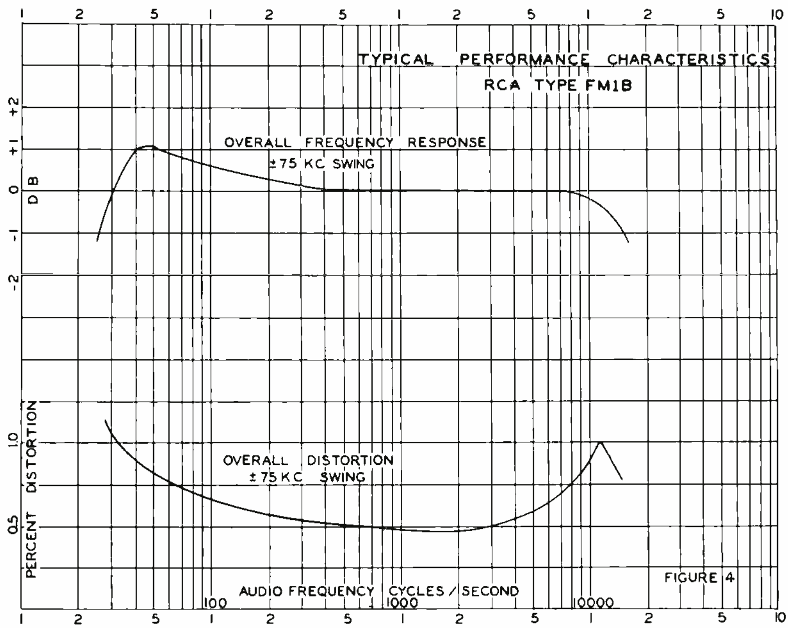


Fig. 4

megacycles and under conditions of peak positive-grid swing at the carrier for Class C plate-modulated telephony the transit angles in the three spaces are 14, 5, and 32 degrees, respectively.

Two other details of the construction are worthy of mention. The exhaust tubulation is made of metal tubing allowing the completed tip-off to be kept small and out of the way of the connections to the tube, while at the same time removing the hazard of breakage. The other detail is in the means of flashing the getter in the tube. A ribbon getter is connected between the two control-grid terminals just inside the header. During the exhaust process a radio-frequency voltage is connected to the two grid leads external to the tube. The difference in impedance of the two parallel paths—through the getter ribbon and through the grid—allows the current to heat the getter sufficiently to flash it with very little heating of the grid, and because of its location inside the screen-grid conical support, the getter flashes onto the support rather than onto the glass bulb wall. Thus, there is no getter deposited upon the bulb where it can cause leakage or losses.

The construction of the radiator assembly involves several new features. A machined copper core is shrunk into a section of a hollow aluminum extrusion which forms the fins. The anode of the exhausted

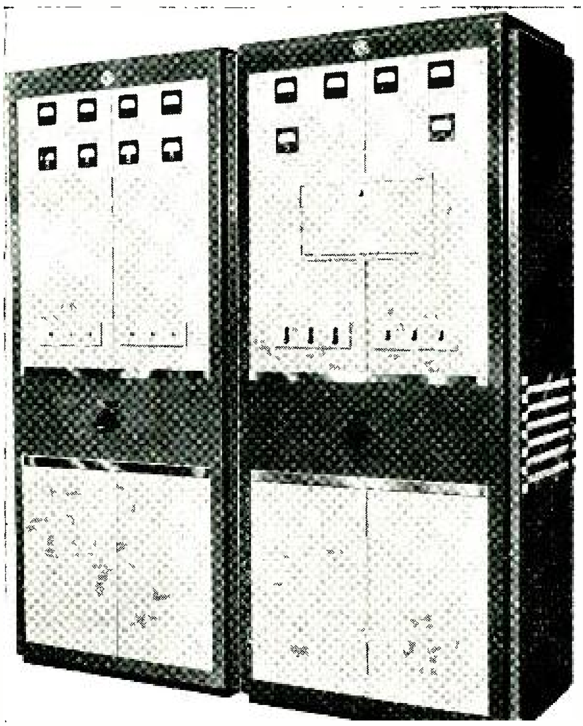


Fig. 5

tube is soldered into this core. This method of construction offers savings in manufacturing cost and in weight of the finished tube. Maximum rated total dissipation is somewhat less than 1300 watts.

The two 827-R's are mounted in metal shells on ceramic insulators which conduct the cooling air from the blower to the plate fins. The top of the tube is slightly below a horizontal cross-plate in the transmitter. Screen-grid and filament bypass capacitors form an integral part of this horizontal plate. The capacitors are built up by mounting one plate below and two plates above the horizontal cross-plate, each separated from it by a mica spacer. The whole assembly is clamped together by clamping screws mounted through Isolantite insulators.

The lower plate carries the ring of beryllium copper clips which make connection from the screen grid of each tube to the bypass capacitor plate. The filaments of the tubes are connected to the upper capacitor plates. The circuits between the filaments, the grid and plate return, and the screen grid are, therefore, all extremely short, resulting in very stable operation.

Plate-tank tuning of the power amplifier is accomplished by varying the capacitance between the metal shells into which the tube anodes are

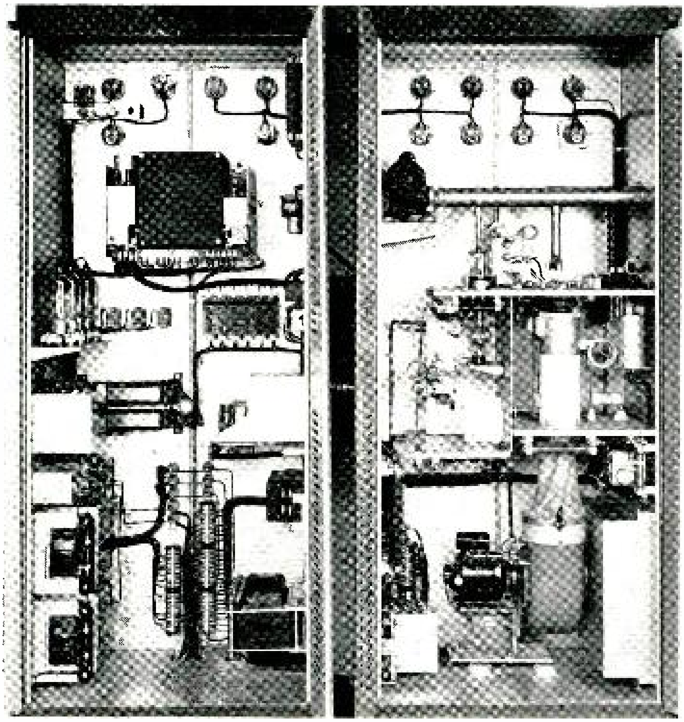


Fig. 6

mounted. The variation is obtained by sliding a brass block, shaped to fit into the space between the two jackets, up or down vertically in the space between them. At one end of the tuning range the brass block is adjacent to the ceramic blower pipe and as the supporting lead screw is rotated it moves up, increasing the capacity from plate-to-plate until the block is directly opposite the metal tube jacket, and the capacity is at a maximum. The variation in capacity secured by this means is approximately 14 micro-microfarads.

The plate-tank circuit is bolted directly onto the tube mounting jackets. At 108 megacycles it consists of a single rectangular section, while at the lower frequencies a conventional multi-turn tank coil is used. All of the parts of the power-amplifier plate-tank circuit which carry high-frequency energy are heavily silver-plated to minimize losses.

The load circuit is inductively coupled to the power-amplifier plate tank. Terminating insulators are provided in the top of the cabinet for connection to an open wire or concentric transmission line. An opening is provided in the bottom of the cabinet for an alternative concentric transmission-line connection.

Since frequency-modulation broadcasting is to be a high-fidelity service, every effort was made to make the performance of this transmitter exceed any reasonably anticipated requirements for fidelity. The distortion and frequency characteristics are shown in Figure 4. It will be observed that the distortion introduced by the entire transmitter is less than might reasonably have been expected from a single audio-frequency amplifier stage a few years ago.

A front view of the complete transmitter is shown in Figure 5. The left-hand unit contains the radio-frequency amplifiers while in the right-hand side are located the frequency-modulator unit and the high and low-voltage power supplies, control equipment, etc. Controls and meters are grouped at convenient levels. Figure 6 shows a rear view of the transmitter. The right-hand unit contains the radio-frequency amplifier chain which takes the output of the frequency-modulated exciter, approximately 5 watts, and amplifies it to a full 1-kw of output power. The first tube of this chain is a Type 814, which acts as an amplifier, the second stage consists of two Type 808's connected in push-pull and tuned to triple their exciting frequency. The output of this stage is at the output frequency of the transmitter and drives two Type 8001's which raise the power level of the excitation sufficiently to drive the two 827R's in the power amplifier. In the left-hand cabinet are mounted the frequency-modulated exciter unit, which may be seen in the upper center of the enclosure, the low-power rectifier mounted on a shelf on the right-hand side panel, and the main rectifier which has components distributed over the lower left-hand section of the chassis arranged for ready accessibility.

The tuning controls are grouped together and are operated by small wrenches. This prevents any possibility of tampering with these essential controls by a casual visitor or other unauthorized person. The tuning motion is carried to the operated units by straight shafts, combined with universals and beveled gears. At the high radio frequencies used in this transmitter it is vital that each tuning element be located correctly in the circuit from the electrical standpoint. The remote tuning drives accomplish this yet permit grouping of the controls. The external cabinet finish is of two shades of umber gray. The inside finish is silver gray.

The mechanical construction is somewhat unique in that the frame consists of a welded-up, box-like section with only a narrow cross-band placed horizontally approximately midway up the center front. The equipment is mounted on L-shaped vertical chassis. Two of these chassis fit into each frame forming the right and left portions of the front panel of each unit and the right and left sides. Virtually all of the equipment is mounted on the chassis and wired before the chassis are set into the transmitter frames. This facilitates assembly and wiring since it is possible for several men to work simultaneously on each unit. If it were necessary to do the assembly and wiring after the equipment was mounted in the cabinet, only one man could work on it at a time.

The RCA-827R tube and the radio-frequency portion of this transmitter were described in a joint paper by the authors in June, 1940, at the Boston Convention of the Institute of Radio Engineers. Much of the material presented here is taken from that paper.

A summary is given below of the electrical specifications of the Type FM-1-B transmitter.

ELECTRICAL SPECIFICATIONS

Frequency—Any single frequency between 26 and 108 mgacycles.

Power output into transmission line—70 to 600 ohms, 1000 watts.

Power input—4100 watts (approximately).

Power supply—230 volts, 60 cycles, single phase, instantaneous regulation 5 per cent maximum, slow-time voltage variation not to exceed limits of 210—250 volts.

Heater input—105 to 125 volts, 60 cycles, single phase, 0 to 350 watts, depending on ambient temperature.

Audio noise level—Amplitude-modulation noise: 60 decibels below carrier level. Frequency-modulation noise: 70 decibels down at plus or minus 75-kilocycle deviation.

Audio-frequency input level—For plus or minus 75-kilocycle deviation: Plus 8 vu. Average-program level: 0 vu.

Radio-frequency stability—Better than plus or minus 2000 cycles.

Audio-frequency characteristic—Plus or minus 1 decibel from 30 to 15,000 cycles.

Audio distortion—Less than 2 per cent rms at plus or minus 75 kilocycles deviation, including all harmonics up to 30 kilocycles, for any audio frequency from 30 to 15,000 cycles.

AN ANALYSIS OF CONSTANT-K LOW- AND HIGH-PASS FILTERS

By

Oliver S. Meixell

R.C.A. Institutes, Inc., New York

Summary—A development is made of expressions yielding complete insertion loss and phase shift as functions of only the frequency ratios: f/f_c (for the low-pass filter) and f_c/f (for the high-pass filter). In these ratios, f_c is the cut-off frequency of the filter and f is any frequency at which the attenuation and phase shift are desired. The analytic expressions are developed for single and two-section T or π structures terminated in a physical (or equivalent) resistance equal to $\sqrt{L/C}$.

IN DETERMINING the attenuation and phase shift caused by inserting an electric filter between a source of e.m.f. and a load it is ordinarily necessary to refer to the curves or analytic expressions of generalized filter theory. These are given on the basis of a filter terminated at both ends in an ideal impedance known as the image impedance. Under such conditions the attenuation constant α and phase shift β can be determined per section and then multiplied by the number of sections to give the total. On the other hand if the actual termination is a resistance and accordingly meets image requirements over only part of the frequency spectrum then the total insertion loss L and phase shift B must include, respectively, not only $n\alpha$ and $n\beta$ (where n is the number of sections), but also the effects of possible mismatch of the actual termination with the image impedance.

For certain networks such as the constant-K low- and high-pass structure it has been found possible to utilize a more direct approach than through generalized filter theory. The method to be demonstrated yields the total insertion loss L and the total phase shift B with the filter terminated at both ends in a physical (or equivalent) resistance equal to the frequently-used value of $\sqrt{L/C}$ where L and C are the element inductance and capacitance.

Before an illustration of the method is given consider the definition of complete insertion loss and phase shift. Refer to Figure 1 where a network is inserted between a source of e.m.f. with its impedance Z_A and a load of value Z_B.

Let I_1 be the current in Z_B before the insertion of the network and I_2 the current afterward. The vector ratio of the currents I_1 and I_2 is thus:

$$I_1/I_2 = \frac{|I_1| \underline{/\theta_1}}{|I_2| \underline{/\theta_2}} = \left|\frac{I_1}{I_2}\right| \underline{/\theta_1 - \theta_2} = \left|\frac{I_1}{I_2}\right| \underline{/B}$$

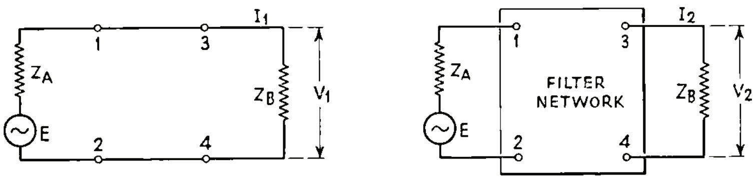
By definition of insertion loss and phase shift:

Total insertion loss in decibels $(db) = L = 20 \log_{10} |I_1/I_2|$ (1)

Total insertion phase shift $= B =$ Angle of I_1/I_2 (2)

This definition holds good regardless of the number of sections. If I_2 should be found equal to I_1 in magnitude and phase, then both the insertion loss and phase shift would be zero. In practice, the insertion loss of a well-designed filter section is negligible for most of the theoretical "pass" region.

In generalized filter theory the L defined above is evaluated as a summation of component attenuation constants together with reflec-



Before insertion of filter — After insertion of filter

Fig. 1

tion losses and other losses occasioned by possible mismatch of the actual with image impedance requirements. In a similar fashion the total phase shift B would be regarded as a summation of component phase constants and other phase shifts. While this approach is essential to define the behavior of single and multisection filters of diverse nature, it is apparent that any other valid means for obtaining L and B is satisfactory and even preferable in a particular case where the method involves less labor or a less extensive knowledge of filter theory.

With the foregoing in mind consider the application of mesh computation to constant-K low- and high-pass T and π structures terminated at each end in $\sqrt{L/C}$.* These structures are symmetrical ones and accordingly if the network is being inserted as shown in Figure 1, then $Z_A = Z_B = \sqrt{L/C} = R$. Furthermore, the analysis yields the ratio E/V_2 where E is the energizing voltage and V_2 is the voltage across the termination after the network's insertion. The necessary I_1/I_2 can readily be found in terms of E/V_2 by noting in Figure 1 that:

* For a T structure, the ideal or image impedance is given by

$$Z_T = \sqrt{L/C}\ \sqrt{1 - (f/f_c)^2}. \quad \text{(Low-pass)}.$$

Obviously if the actual termination is constant and equal to $\sqrt{L/C}$ then severe mismatch does not occur until after $f = 0.7 f_c$ at which point $Z_T = 0.71 \sqrt{L/C}$. In a similar way it can be shown that $\sqrt{L/C}$ is a practical though imperfect termination for the π structure.

Before insertion

$$I_1 = \frac{E}{Z_A + Z_B} = \frac{E}{2R} \qquad \text{whence } E = 2I_1R$$

After insertion

$$V_2 = I_2Z_B = I_2R$$

$$\therefore \frac{E}{V_2} = \frac{2I_1R}{I_2R} = 2\frac{I_1}{I_2} \text{ or } I_1/I_2 = \frac{1}{2}\frac{E}{V_2} \tag{3}$$

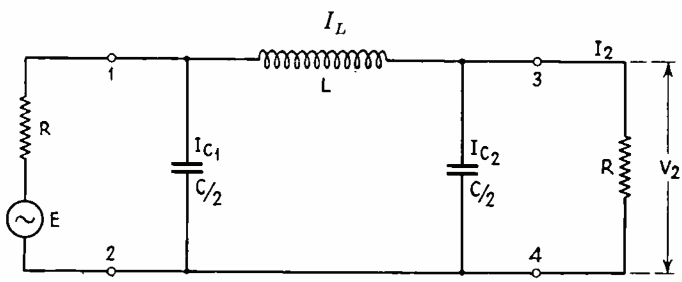


Fig. 2

The ratio I_1/I_2 will now be found in terms of $r = f/f_c$ for a single π section low-pass filter. Design theory gives the following element values:

$$L = \frac{R}{\pi f_c}$$

$$C = \frac{1}{\pi f_c R}$$

where $R =$ Termination resistance in ohms.

$f_c =$ Cut-off frequency in cps.

$L =$ Full-section inductance in henries.

$C =$ Full-section capacitance in farads.

Now that values of L and C have been found in terms of the physical termination R and the desired cut-off frequency f_c, the π structure is illustrated in Figure 2 with proper values of inductance and capacitance together with designated currents and voltages.

Allow a unit current (on the axis of reals) to flow in the load termination R, i.e., $I_2 = 1 + j0$.

Then

$$V_2 = I_2 R = R$$

Under this condition

$$I_{c2} = \frac{V_2}{Z_c} = \frac{R}{\dfrac{-j}{\omega \dfrac{C}{2}}} = j\frac{\omega CR}{2}$$

Further

$$I_L = I_2 + I_{c2} = 1 + j\frac{\omega CR}{2}$$

The voltage across the input condenser

$$E_{12} = V_2 + I_L Z_L$$

$$= R + \left(1 + j\frac{\omega CR}{2}\right) j\omega L = R - \frac{\omega^2 CLR}{2} + j\omega L$$

Whence

$$I_{c1} = \frac{E_{12}}{Z_c} = \frac{R - \dfrac{\omega^2 CLR}{2} + j\omega L}{-j\dfrac{1}{\omega \dfrac{C}{2}}}$$

$$= -\frac{\omega^2 CL}{2} + j\frac{\omega CR}{2} - j\frac{\omega^3 C^2 LR}{4}$$

$$I_o = I_L + I_{c1} = 1 - \frac{\omega^2 CL}{2} + j\omega CR - j\frac{\omega^3 C^2 LR}{4}$$

$$E = E_{12} + I_o R$$

$$= 2R - \omega^2 CLR + j\omega L + j\omega CR^2 - j\frac{\omega^3 C^2 LR^2}{4}$$

E is, therefore, the applied voltage which will produce $1 + j0$ current in the load.

Further

$$E/V_2 = E/R = 2 - \omega^2 CL + j\frac{\omega L}{R} + j\omega CR - j\frac{\omega^3 C^2 LR}{4}$$

This ratio, while established for an e.m.f. yielding unity current in the termination, will hold for any value of E. This is true since, for a given ω, the ratio is a function simply of the linear circuit parameters; L, C, and R.

From Equation (3)

$$I_1/I_2 = \frac{1}{2}\frac{E}{V_2} = 1 - \frac{\omega^2 CL}{2} + j\frac{\omega L}{2R} + j\frac{\omega CR}{2} - j\frac{\omega^3 C^2 LR}{8}$$

The quantities in the above equation can readily be expressed in terms of $r = f/f_c$ by remembering that for this type of filter

$$f_c = \frac{1}{\pi\sqrt{LC}} \text{ and } \mathrm{R} = \sqrt{\mathrm{L/C}}$$

whence also:

$$L = \frac{R}{\pi f_c} \text{ and } C = \frac{1}{\pi f_c R}$$

From these relations it follows that

$$\frac{\omega^2 CL}{2} = 2\,(f/f_c)^2 = 2r^2$$

$$\frac{\omega L}{2R} = f/f_c = r \qquad \text{where} \qquad r = f/f_c$$

$$\frac{\omega CR}{2} = f/f_c = r \qquad \text{where} \begin{cases} f = \text{any frequency} \\ f_c = \text{cut-off frequency} \end{cases}$$

$$\frac{\omega^3 C^2 LR}{8} = \frac{\omega^2 CL}{2} \cdot \frac{\omega CR}{4} = r^3$$

Accordingly

$$I_1/I_2 = 1 - 2r^2 + jr + jr - jr^3 = 1 - 2r^2 + jr(2 - r^2) \qquad (4)$$

The magnitude of Equation (4) is the magnitude of the ratio of the currents before and after the insertion of the filter section. The angle of Equation (4) is the total insertion phase shift and tells by what angle I_2 lags I_1. In accordance with filter convention an angle of lag (I_2 with respect to I_1) is a positive angle.

$$|I_1/I_2| = \sqrt{(1 - 2r^2)^2 + [r(2 - r^2)]^2} = \sqrt{1 + r^6}$$

$$B = \text{Angle of } I_1/I_2 = \tan^{-1} \frac{r(2 - r^2)}{1 - 2r^2}$$

Whence the polar form of the current ratio:

$$I_1/I_2 = |I_1/I_2| \,\underline{/B} = \sqrt{1 + r^6}\,\underline{/B}$$

By definition, (see Equations (1) and (2)):

Total insertion loss $L = 20 \log_{10} |I_1/I_2| = 10 \log_{10}(1 + r^6)$ (5)

Total insertion phase shift $B = \text{Angle of } I_1/I_2$

$$\text{i.e. } B = \tan^{-1} \frac{r(2 - r^2)}{1 - 2r^2} \qquad (6)$$

Equations (5) and (6) can be used at any frequency in either the pass or attenuation region. Suppose the frequency f takes on the values $0.5f_c$, f_c and $2f_c$ where f_c is the cut-off frequency. The following table shows the attenuation and phase shift:

f	$r = f/f_c$	$L = 10 \log_{10}(1 + r^6)$	$B = \tan^{-1} \frac{r(2 - r^2)}{1 - 2r^2}$
$0.5f_c$	0.5	0.067 db	60.3°
f_c	1.0	3.0 db	135.°
$2.0f_c$	2.0	18.1 db	210.°

Element values

Low Pass Filter: $L = R/\pi f_c$ and $C = 1/\pi f_c R$

High Pass Filter: $L = R/4\pi f_c$ and $C = 1/4\pi f_c R$

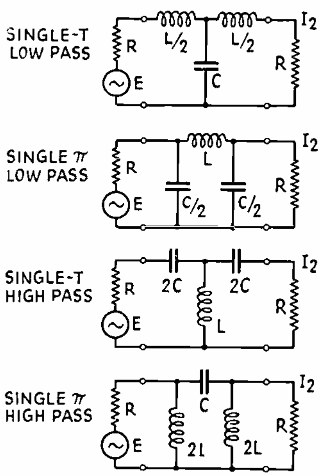
For both types: $R =$ Resistive termination in ohms.
$f_c =$ Cutoff frequency in c.p.s.

Configuration and Formulas for analysis at a frequency f

Notes: $I_1 =$ Current in load R before insertion of filter.
$I_2 =$ Current in load R after insertion of filter
($I_1 = E/2R$)

For Low Pass Filter:
$r = f/f_c$ and use (+) sign before quadrature term.

For High Pass Filter:
$r = f_c/f$ and use (—) sign before quadrature term.



Formulas which apply (all sections)

$$I_1/I_2 = 1 - 2r^2 \pm jr(2 - r^2)$$
$$|I_1/I_2| = \sqrt{1 + r^6}$$

Total Insertion Loss in Decibels (db)

$$L = 20 \log_{10} |I_1/I_2| = 10 \log_{10}(1 + r^6)$$

Total Phase Shift (degrees or radians—as convenient)

$$B = \tan^{-1} \frac{\pm r(2 - r^2)}{1 - 2r^2}$$

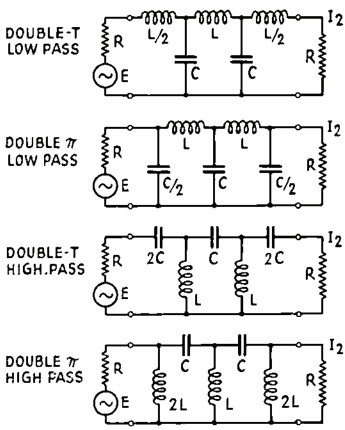
Fig. 3

If the filter were terminated on an image basis the attenuation would be 0 at both $0.5f_c$ and f_c. The table shows this to be substantially true at $0.5f_c$ where the actual termination is close to the image for this type of filter. Observe, however, how the attenuation rises to 3 db at f_c due to mismatch effects. The phase shift increases in the positive or counter-clockwise direction to 135° at f_c. With the filter terminated on an image basis, the phase shift would be 180° at $f = f_c$ and for all higher frequencies.

Using an approach like the one just outlined for the single π low-pass structure, formulas were developed for other single- and two-section constant-K filters terminated in a physical (or equivalent) resistance equal to $\sqrt{L/C}$. These are summarized in Figure 3 and Figure 4.

An examination of the summary brings out the interesting fact that the attenuation is identical for both T and π structures terminated in $\sqrt{L/C}$. In the same way the T and π phase shifts are alike. These identical attenuation and phase shifts are true in spite of the fact that a vastly different type of mismatch occurs for the T as compared with the π structures particularly in the vicinity of cut-off.

The expressions for the attenuation are readily used at any frequency in the spectrum. On the other hand, plots of the phase-shift



Formulas which apply
(all sections)

$$I_1/I_2 = 2(1-2r^2)^2 - 1 \pm j2r(2-r^2)(1-2r^2)$$

$$|I_1/I_2| = \sqrt{1 + 4r^6(1-2r^2)^2}$$

Total Insertion Loss in Decibels (db)

$$L = 10 \log_{10}[1 + 4r^6(1-2r^2)^2]$$

Total Phase Shift (degrees or radians—as convenient)

$$B = \tan^{-1} \frac{\pm 2r(2-r^2)(1-2r^2)}{2(1-2r^2)^2 - 1}$$

Fig. 4

characteristics are made on Figure 5 through $f = 2.25 f_c$. In this span the phase shift passes over practically its full range of values which is an aid in the proper determination of total phase shift. For example, say a single-section high-pass filter is used. The analytic expressions at $f = 0.5 f_c$ ($r = f_c/f = 2$) yield the insertion loss $L = 18.1$ db and the phase shift $B = -210°$ measured in the negative or clockwise direction and, therefore, an angle of lead of 210°. On the other hand for a two-section high-pass filter at $f = 0.5 f_c$ ($r = 2$), $L = 41$ db and $B = -30°$. This phase shift is really $-360° + (-30°)$ or $-390°$ or an angle of lead of 390° as an examination of the curves will show. A study of the phase-shift formulas shows that single-section high-pass filters terminated in $\sqrt{L/C}$ have a limit of 270° while the two-section filters reach a limit of 450°; both of these lead angles occurring, theoretically, at 0 frequency. Similarly, the low-pass single- and double-section structures have limiting lag angles of 270° and 450°, respectively, occurring at infinite frequency.

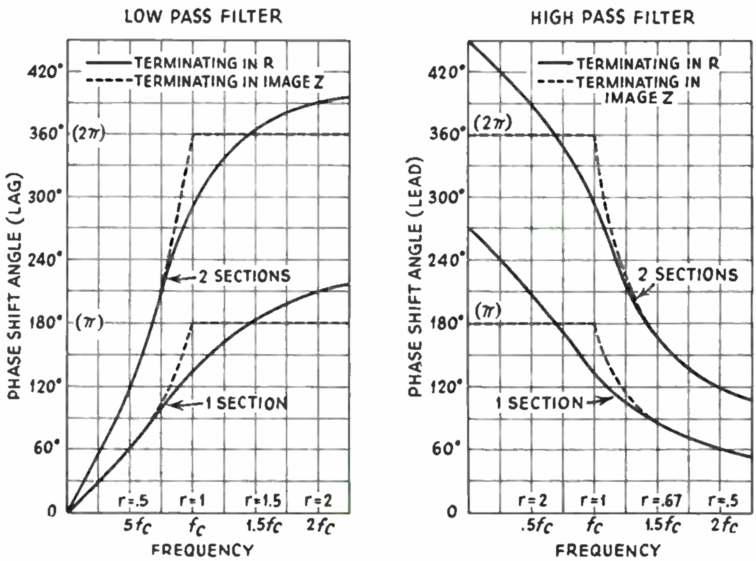


Fig. 5—Phase-Shift *B*

For purposes of comparison, Figure 5 likewise shows the phase shifts if the structures were terminated in their ideal or image impedances.

So far, no mention has been made of the possible influence of resistance in the filter's L and C elements. Since the Q value of good condensers is very high we need be concerned only with the resistance of the coils. Consider a double-section low-pass filter whose terminations are each 500 ohms and whose frequency of cut-off is 10,000 cps. Computations and the completed filter are shown in Figure 6.

$$\text{Element values: } L = \frac{R}{\pi f_c} = .01592h. \quad C = \frac{1}{\pi f_c R} = .0636\ \mu f$$

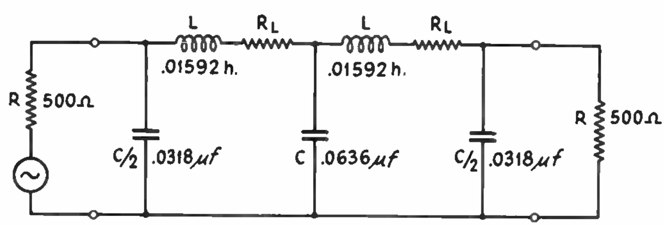


Fig. 6

If it is assumed that the condensers are resistanceless and that the coil Q value is 40 at 10,000 cps ($R_L = 25$ ohms), the following table shows the attenuation and phase shift calculated on this basis as com-

Frequency (cps.)	$r = f/f_c$	L (By formula; $R_L = 0$)	B (By formula; $R_L = 0$)	L ($Q = 40$ at 10,000 cps.)	B ($Q = 40$ at 10,000 cps.)
5,000	0.5	0.07 db	120°	0.55 db	120°
10,000 (f_c)	1.0	7.0 db	297°	7.5 db	292°
20,000	2.0	41.0 db	390°	41.0 db	389°

pared with the same quantities from the formulas of Figure 3 and Figure 4.

It is evident that the comparison of values shows little or no error through the use of the formulas except in the attenuation at low frequencies. Even this small departure can be reduced through the use of coils with a Q higher than assumed. The two-section filter has an attenuation and phase shift of 7.5 db and 292° lag, respectively, at the cut-off frequency. These values are substantially the same whether or not the coil resistance is considered. For an image termination the attenuation would be zero, but terminal mismatch rather than coil resistance causes the actual attenuation to be so high. Other low- and high-pass filters were similarly analyzed for the effect of coil resistance and the results in each case indicated that the formulas covering the applications illustrated in Figure 3 and Figure 4 can be used with accuracy provided the Q of the coils is reasonably high.

The structures chosen for analysis are not necessarily the only ones for which L and B can be found as functions of r alone. The reader can develop compact and easily used expressions in the case of the constant-K half section terminated in $\sqrt{L/C}$. This termination may be either a physical resistance or equivalent one (as through the use of a transformer). It is possible that certain of the "m"-derived sections can similarly be analyzed in terms of any frequency (f), the cut-off frequency (f_c), and the frequency at which infinite attenuation takes place (f_∞).

Irrespective of whether or not a simplified general solution can be found in the manner shown for the constant-K structure, it should be noted that the method of mesh computation (generalized in that case) can be applied readily to any ladder structure at a particular frequency. It is necessary only to compute the ohmic values of the impedances in symbolic form and then proceed in just the same way to get a vector ratio of input and output voltages. If this network is a filter then it is not difficult to determine also the attenuation and phase shift caused by the filter's introduction into the circuit at the frequency for which the impedances were determined.

CASCADE AMPLIFIERS WITH MAXIMAL FLATNESS

By

V. D. Landon

RCA Manufacturing Company, Inc., Camden, N. J.

Summary—*The theory of the conventional constant-K filter is not very satisfactory from the engineering standpoint. The theory is developed on the basis of terminations which match the surge impedance. Unfortunately, this calls for terminating resistors which vary with frequency.*

When a compromise-fixed resistor is used, it becomes difficult to calculate the performance of the filter accurately. Also, the curve is, in general, not the flattest which can be had with the available number of reactive circuit elements. In the system which is about to be described, the performance is very easy to calculate regardless of the number of circuits. Also, the curve is the flattest which may be obtained with a given number of reactive circuit elements. Conventional filter theory usually ignores the resistance of the component reactors. At any rate, the effect of resistance in the reactors is very difficult to calculate accurately. In the following system, each element is expected to have resistance of a required value and the effect is accurately known.

PART I

Low Pass Characteristics

A LOW pass characteristic may be obtained by the use of any number of reactance elements from one up to n. If only one is used, the circuit is one of those shown in Figure 1. If two reactive elements are to be used, it would be possible to build a two-stage cascade amplifier in which each stage took the form of one of the circuits of Figure 1. However, considerably better performance may be obtained by using one of the circuits of Figure 2. When "n" is the number of reactance elements to be used, a cascade amplifier may be built using the circuits of Figures 1 and 2 as coupling units.

In practice, Figures 1-a and 2-a are not suitable as interstage coupling networks. They are included for generality. They might be used as input circuits.

Circuits 1-b and 2-b involve the practical arrangement of the conventional amplifier in which the voltage output of one circuit is applied to the grid of a vacuum tube. As a result a current $I = E\ g_m$ is applied to the following circuit. In case the plate impedance of the tube is not so high as to be neglected, the modification required is too well known to warrant repeating here.

PERFORMANCE OF CIRCUITS OF FIGURES 1 AND 2

The performance of circuit 1-a or 1-b is:

$$\frac{E_1}{E_o} = A \frac{\beta}{j\omega + \beta}$$

(See Appendix 1)

$$\left| \frac{E_1}{E_o} \right| = A \frac{\beta}{\sqrt{\omega^2 + \beta^2}}$$

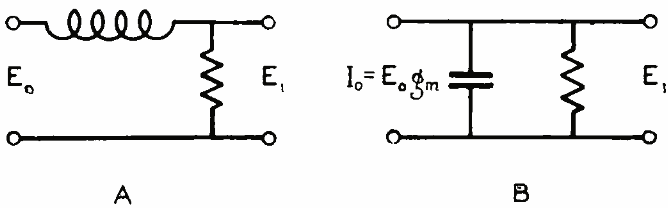


Fig. 1—Low pass coupling networks having one reactance element.

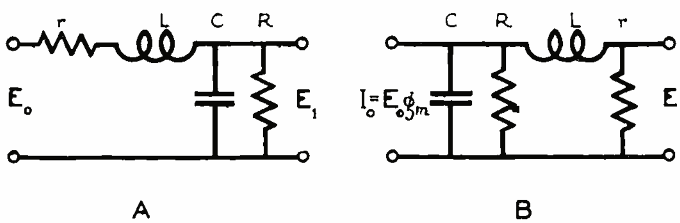


Fig. 2—Low pass coupling networks having two reactance elements

where

E_o is the input voltage

E_1 is the output voltage

A is the gain at zero frequency

$= \text{unity}$ in Figure 1-a

$= g_m R$ in Figure 1-b

β is the frequency where the response drops to $1/\sqrt{2}$. This definition applies for Figure 1 only.

$\beta = R/L$ for Figure 1-a

$= \dfrac{1}{RC}$ for Figure 1-b

$j = \sqrt{-1}$

ω is the angular frequency, variable.

The performance of circuit 2-a or 2-b is:

$$\frac{E_1}{E_o} = A \frac{\beta^2}{j\omega\alpha + \beta^2 - \omega^2}$$

(See Appendix 2)

$$\left|\frac{E_1}{E_o}\right| = A \frac{\beta^2}{\sqrt{\omega^4 + \omega^2(\alpha^2 - 2\beta^2) + \beta^4}}$$

where

A is the gain at zero frequency.

β is the angular frequency at which the coupling network produces a ninety-degree phase shift.

α is the damping factor.

Performance of Cascade Amplifier

When several stages of amplification are operated in cascade, using the circuits of Figure 1 and/or Figure 2 and a total of n reactance elements is used, the overall response is expressed by the following:

$$\left|\frac{E_1}{E_o}\frac{E_2}{E_1}---\right| = (A_1A_2---)(\beta_1\beta_2^2---)\frac{1}{\sqrt{\omega^{2n} + C_{n-1}\omega^{2n-2} + C_{n-2}\omega^{2n-4} --- + C_1\omega^2 + \beta_1\beta_2^2---}}$$

where C_{n-1}, $C_{n-2}---$, C_1 are constants determined by various circuit parameters.

The flattest overall response curve occurs when the first $(n-1)$ derivatives of $\left|\frac{E_1}{E_o}\frac{E_2}{E_1}---\right|$

(taken with respect to ω^2) are equal to zero at the point $\omega = 0$. This results in the condition that:

$$C_{n-1} = C_{n-2} = --- \qquad C_1 = 0. \quad \text{(See Appendix 3)}$$

Assuming this condition, we have

$$\left|\frac{E_1}{E_o}\frac{E_2}{E_1}---\right| = \frac{(A_1A_2---)(\beta_1\beta_2^2---)}{\sqrt{\omega^{2n} + (\beta_1\beta_2^2---)}}$$

If $\beta_1 = \beta_2 = \beta_3 - - - = \beta$

$$\left| \frac{E_1}{E_o} \frac{E_2}{E_1} - - - \right| = \frac{(A_1 A_2 - - -)\beta^n}{\sqrt{\omega^{2n} + \beta^{2n}}} = \frac{(A_1 A_2 - - -)}{\sqrt{(\omega/\beta)^{2n} + 1}} \; *$$

THE FACTORS OF $\omega^{2n} + \beta^{2n}$

The expression $\omega^{2n} + \beta^{2n}$ can be factored into n factors, $(\omega^2 - \beta^2\mu_1)$ $(\omega^2 - \beta^2\mu_2) - - - - (\omega^2 - \beta^2\mu_n)$ where $\mu_1, \mu_2, - - - \mu_n$ are the various nth roots of -1.

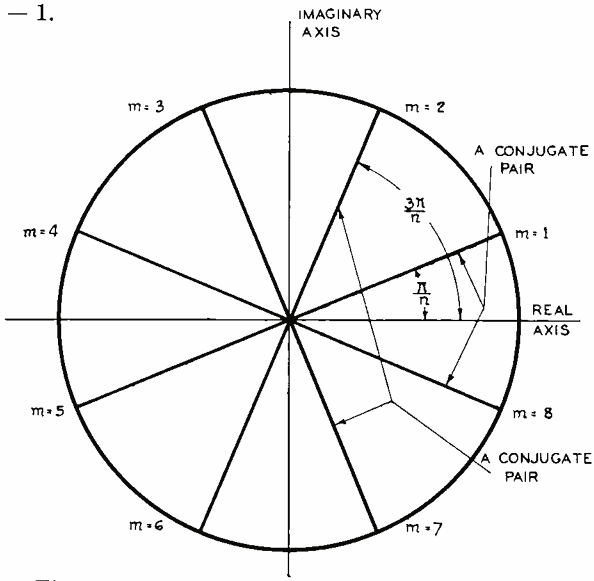


Fig. 3—Geometrical interpretation of the various nth roots of (-1) when $n = 8$.

If m is any number from 1 to n

$$\mu_m = \cos\left[\frac{\pi}{n}(2m-1)\right] + j\sin\left[\frac{\pi}{n}(2m-1)\right]$$ (See Figure 3 and Appendix 3)

If n is odd, then setting $m - \frac{n+1}{2}$ gives $\mu = -1$. Omitting this root, if present, there are an even number of other roots. The first half of these have positive imaginary parts. The second half are identical except that they have negative imaginary parts.

Choosing m less than $n/2$

$$\mu_m = \cos\left[\frac{\pi}{n}(2m-1)\right] + j\sin\left[\frac{\pi}{n}(2m-1)\right]$$

* A similar equation is given by S. Butterworth in *Experimental Wireless* and *The Wireless Engineer*, Oct. 1930.

The conjugate of this root is:

$$\mu_{n+1-m} = \cos\left[\frac{\pi}{n}(2m-1)\right] - j\sin\left[\frac{\pi}{n}(2m-1)\right]$$

Then $(\omega - \beta\mu_m)\ (\omega - \beta\mu_{n+1-m})$

$$= \left(\omega^2 - \beta^2\cos\left[\frac{\pi}{n}(2m-1)\right] + \beta^2 j\sin\left[\frac{\pi}{n}(2m-1)\right]\right)$$

$$\times \left(\omega^2 - \beta^2\cos\left[\frac{\pi}{n}(2m-1)\right] - \beta^2 j\sin\left[\frac{\pi}{n}(2m-1)\right]\right)$$

$$= \omega^4 - 2\omega^2\beta^2\cos\left[\frac{\pi}{n}(2m-1)\right] + \beta^4$$

This is the product of the mth pair of conjugate roots of the expression $\omega^{2n} + \beta^{2n}$.

Conditions Resulting in Maximal Flatness

The conditions can now be specified for making the overall performance of the amplifier correspond to the ideal equation:

$$\left|\frac{E_1}{E_o}\,\frac{E_2}{E_1}---\right| = \frac{(A_1 A_2 ---)\ \beta^n}{\sqrt{\omega^{2n} + \beta^{2n}}}$$

If n is even, there should be $n/2$ stages of the type of Figure 2 and the performance of the mth stage is expressed by

$$\left|\frac{E_m}{E_{m-1}}\right| = A_m \frac{\beta^2}{\sqrt{\omega^4 + \omega^2(\alpha_m - 2\beta^2) + \beta^4}} \qquad \text{(See Appendix 2)}$$

where $\omega^4 + \omega^2(\alpha_m{}^2 - 2\beta^2) + \beta^4 = \omega^4 - 2\omega^2\beta^2\cos\left[\frac{\pi}{n}(2m-1)\right] + \beta^4$

then $$\alpha_m{}^2 - 2\beta^2 = -2\beta^2\cos\left[\frac{\pi}{n}(2m-1)\right]$$

$$\alpha_m{}^2 = 2\beta^2\left(1 - \cos\left[\frac{\pi}{n}(2m-1)\right]\right)$$

$$\alpha_m/\beta = 2\sqrt{\frac{1-\cos\left[\frac{\pi}{n}(2m-1)\right]}{2}}$$

$$= 2\sin\left[\frac{\pi}{2n}(2m-1)\right]$$

If n is odd, there are $\frac{n-1}{2}$ stages of the type of Figure 2 and one stage of the type of Figure 1. The performance of the latter is expressed by

$$\left|\frac{E_1}{E_o}\right| = \frac{A\beta}{\sqrt{\omega^2+\beta^2}}$$ corresponding to

that root of $(\omega^{2n} - \beta^{2n})$ which is expressed by $(\omega^2 - \beta^2\mu_{\frac{n+1}{2}})$ because $\mu_{\frac{n+1}{2}} = -1$.

It should also be noted that β must have the same value for each stage.

For the Figure 2 type of circuit, β $\left(=\sqrt{\frac{r+R}{R}\frac{1}{LC}}\right)$ is that angular frequency at which the circuit produces a 90° phase shift. If either $r = O$, or $R = \infty$, then β is the true resonant frequency of L and C. Thus, if only one source of damping is used in each stage, the same value of $\frac{1}{\sqrt{LC}}$ is used in each stage. If both r and R are present, a smaller value of $\frac{1}{\sqrt{LC}}$ is required and the value is different in each stage.

If the one element stage is of the Figure 1-a type, then $\beta = r/L$. If it is of the 1-b type, $\beta = \frac{1}{RC}$. For the overall amplifier, β is the frequency where the response drops to $1/\sqrt{2}$ of that at zero frequency.

For the Figure 2 type of circuit, α is the damping factor. It is equal to the sum of the damping factors calculated separately for the series and shunt resistors. That is, $\alpha = \frac{r}{L} + \frac{1}{RC}$. (See Appendix 2.)

The fraction β/α might be called the effective q of any stage. If $R=\infty$, the fraction β/α reduces to $\frac{\sqrt{L/C}}{r}$ or the reactance at resonance divided by the series resistance. The value of β/α is also the gain of each stage at the frequency β relative to that at zero frequency. The value varies from stage to stage, according to the expression:

$$\beta/\alpha = q = \frac{1}{2 \sin\left[\frac{\pi}{2n}(2m-1)\right]}$$

This expression takes its largest values when $m=1$, in which case it is approximately equal to n/π for large values of n.

In the following table, the proper values for β/α are given for a number of values of n and m.

TABLE OF DESIGN CONSTANTS

n	m	$\phi = \frac{\pi}{2n}(2m-1)$	$\sin\phi$	$\beta/\alpha = \frac{1}{2\sin\phi}$
1	1	$\pi/2 = 90°$		
2	1	$\pi/4 = 45°$	.7071	.7071
3	1	$\pi/6 = 30°$	.5	1.0
	2	$3\ \pi/6 = 90°$		
4	1	$\pi/8 = 22.5°$	.3827	1.31
	2	$3\ \pi/8 = 67.5°$	.9239	.541
5	1	$\pi/10 = 18°$	.3090	1.618
	2	$3\ \pi/10 = 54°$	.8090	.618
	3	$5\ \pi/10 = 90°$		
6	1	$\pi/12 = 15°$	.2588	1.930
	2	$3\ \pi/12 = 45°$	.7071	.7071
	3	$5\ \pi/12 = 75°$	.9659	.517
7	1	$\pi/14 = 12\ 6/7°$	.2220	2.25
	2	$3\ \pi/14 = 38\ 4/7°$	.6233	.804
	3	$5\ \pi/14 = 64\ 2/7°$	.9010	.555
	4	$7\ \pi/14 = 90°$		
8	1	$\pi/16 = 11\frac{1}{4}°$	.1950	2.563
	2	$3\ \pi/16 = 33\frac{3}{4}°$	.5555	.900
	3	$5\ \pi/16 = 56\frac{1}{4}°$	.8315	.601
	4	$7\ \pi/16 = 78\frac{3}{4}°$	.9808	.504

Where $\phi = 90°$, a circuit of the type of Figure 1 is called for. The expression q, is meaningless in this case.

THE MEANING OF "MAXIMAL FLATNESS"

The term "Maximal Flatness" is admittedly a coined expression. It was chosen expressly to describe the family of curves defined by the equation:

$$\left| \frac{E_1}{E_o} \frac{E_2}{E_1} --- \right| = \frac{(A_1\, A_2 ---)}{\sqrt{(\omega/\beta)^{2n} + 1}}$$

These curves are in a sense the flattest curves which may be obtained by the use of n reactive circuit elements. However, it is necessary to add several modifying statements for strict accuracy. For example, it is necessary to rule out any circuit employing a parallel

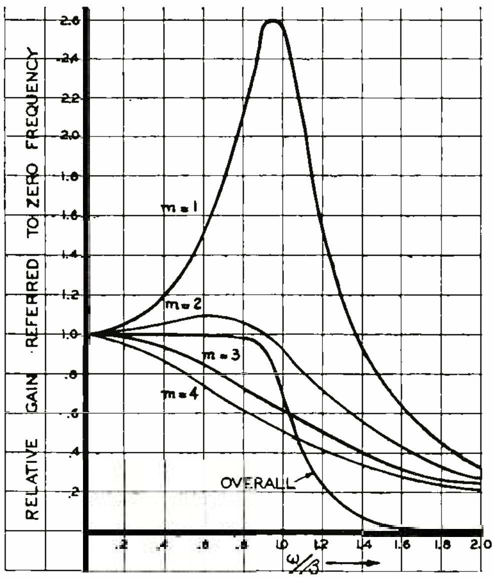


Fig. 4—Response curves for $n = 8$.

resonant circuit in a series arm or a series resonant circuit in a shunt arm. In filter parlance, the only structures discussed are constant K; m-derived sections are not covered.

It should also be remarked that the curves corresponding to the above equation are not curves of minimum deviation. For example, the gain at $\omega = \beta$ and at slightly lower frequencies could be improved by increasing the value of q for the stages of smaller m numbers. Curves can be produced with n combined peaks and valleys. The peaks can be made of equal height and the valleys of equal depth by juggling parameters. The "best" curve depends on the tolerance allowable between peaks and valleys.

Unfortunately, this case is very difficult to handle. Perhaps the chief merit of the equation for maximal flatness is its unusual simplicity. However, the improvement over amplifiers using identical stages is very real as shown later.

AMPLITUDE CHARACTERISTICS FOR $n = 8$

In Figure 4, performance curves are plotted for each of the four stages and for the overall amplifier when $n = 8$. The conventional

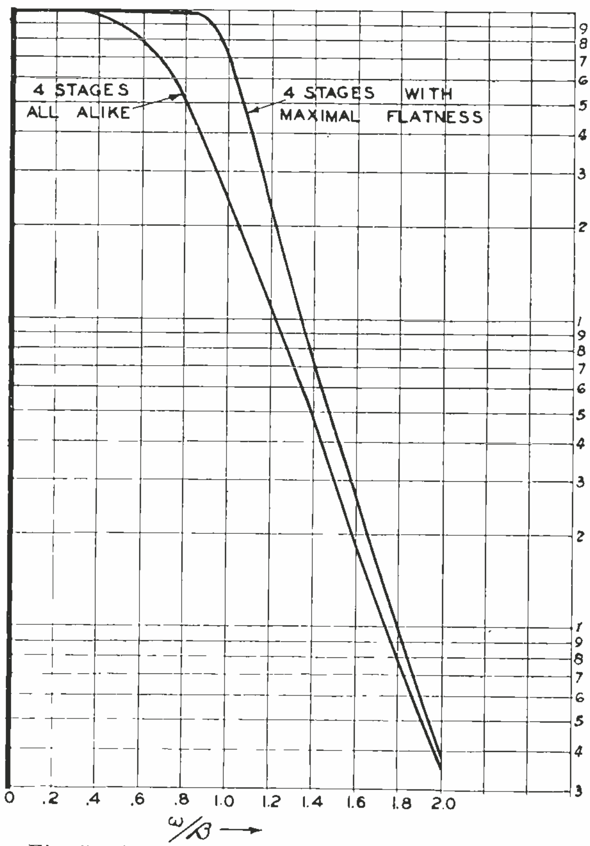


Fig. 5—Comparison of two four-stage amplifiers.

manner of operating a similar amplifier corresponds to using four stages all alike, each stage having the characteristics corresponding to $n = 2$ and $m = 1$. The two overall curves are compared in Figure 5.

It can be seen that the band width at 90 per cent is nearly twice as great in the case of the maximally flat amplifier. The gain per stage is the same in both amplifiers at zero frequency.

It should be noted that the overall performance of the maximally flat amplifier may be calculated directly from the simple formula

$$\left|\frac{E_4}{E_o}\right| = \frac{A_1 A_2 A_3 A_4}{\sqrt{(\omega/\beta)^{16} + 1}}$$

Time Delay Characteristics*

It appears to be necessary to calculate the time delay of each stage separately. The vector gain is:

$$\frac{E_1}{E_o} = \frac{1}{1 - (\omega/\beta)^2 + j\,\omega/\beta\;\alpha/\beta}$$

The angle of retardation is:

$$\Theta = \tan^{-1} \frac{\omega/\beta\;\alpha/\beta}{1 - (\omega/\beta)^2}$$

This angle Θ is $\pi/2$ in each stage at $\omega = \beta$, but varies from stage to stage for other values of ω.

The angle Θ (expressed in radians) divided by ω is the time delay in seconds. Multiplying this figure by $4f_c$ expresses the time delay in time units equal to $\frac{1}{4f_c}$ seconds. Time delay curves are shown in Figure 6 for each stage and for the overall amplifier for $n = 8$.

Band-Pass Characteristics

The circuits of Figure 2 may be converted into band-pass circuits by using the low-pass band-pass analogy.† This involves tuning the inductor to p_o (the mean angular frequency of the proposed pass band) by placing a capacitor in series having the value $\frac{1}{p_o^2 L}$. Also the capacitor of Figure 2 is tuned to the same frequency by placing an inductor in parallel having the value $\frac{1}{p_o^2 C}$. Now if p is any angular frequency, the reactance of L and $\frac{1}{p_o^2 L}$ in series is $pL - \frac{p_o^2 L}{p} = L \left(p - \frac{p_o^2}{p}\right)$.

* The expression "time delay" as used here means the delay of the sinusoidal components, and not the envelope time delay which is $d\Theta/d\omega$.

† See "The Low-Pass Band-Pass Analogy" by V. D. Landon, *Proc. I.R.E.*, December, 1936.

Similarly, the susceptance of C and $\dfrac{1}{p_o^2 C}$ in parallel is $C\left(p - \dfrac{p_o^2}{p}\right)$

Thus, if ω is set equal to $\left(p - \dfrac{p_o^2}{p}\right)$, all the equations for the low-pass amplifier apply to the band-pass amplifier. However, ω must now be interpreted as the band width between frequencies of equal attenuation. (See Appendix 5).

The conductively coupled circuits just described are not as practical as the circuit of Figure 7-a in which two resonant circuits are mag-

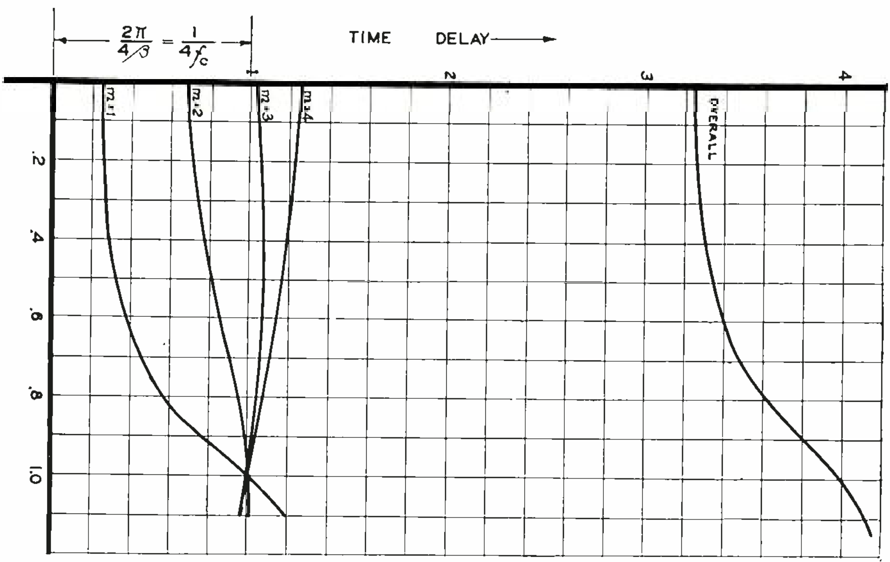


Fig. 6—Time delay curves.

netically coupled together, because in the latter case, the two L-C ratios can be made equal. Here again the same form of equations apply when the terms are properly re-defined. The relations are given in detail in Appendix 6.

If the circuit of Figure 7-b is used, the present theories apply only approximately because its response curve is fundamentally assymetrical in a manner favoring low-frequency response at the expense of high-frequency response.

If R_p were removed and the input voltage were placed in series with C, this circuit would become identical with Figure 7-a (with $r = O$). However, by the use of Thevenin's theorem, it can be shown that the input circuit used in Figure 7-b is equivalent to a series input voltage, this voltage having a value equal to the voltage drop which would occur if C were the only load on the current generator. Hence, the selectivity curve of Figure 7-b is a distortion of that of Figure 7-a, obtained by dividing by the frequency. This does not produce an important effect if β/p_0 is small. If β/p_0 is large, however, there is a tendency for the low-frequency side of the pass-band to be higher than the high-frequency side. This may be compensated for if α_p is not equal to α_s. This is done by tuning the circuit having the lower damping factor to a higher frequency. By this means, the two peaks (if any) in the pass-band may be made equal in amplitude. Nevertheless, the assymetry of the skirts of the curve remains.

In spite of its lack of symmetry, the circuit of Figure 7-b is one of the most practical. This is because the placement of C_p *and* C_s allows the plate capacitance and grid capacitance of vacuum tubes to be used in place of these circuit elements.

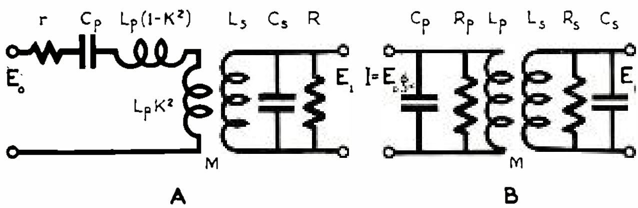


Fig. 7—Coupling networks utilizing coupling circuits.

The gain is $A = \dfrac{g_m}{\beta \sqrt{C_p C_c}}$ and is the same for each of the stages, providing all the damping is concentrated in either primary or secondary in each stage. (See Appendix 7)

Band-Pass Characteristics from Single-Tuned Circuits Cascaded

The results using the coupled circuits just described may be duplicated (from a selectivity standpoint) by the use of a single-tuned circuit per stage with staggered tuning. The number of stages becomes n rather than $n/2$. The type of amplifier used is thus a function of the optimum ratio of the number of tubes to the number of circuits.

If n is odd, one stage is resonant to p_0. The remainder of the stages occur in pairs, one pair for each value of m. The two stages constituting a given pair, are tuned, one on each side of p_0, so that p_0 is the geometric mean of the two resonance points p_a and p_b.

$$p_a - p_b \approx \beta \cos \left[\frac{\pi}{2n} (2m-1) \right]$$

$$\alpha_a \approx \alpha_b \approx \tfrac{1}{2}\,\alpha_m \approx \beta \sin\left[\frac{\pi}{2n}(2m-1)\right]$$ (See Appendix 8A) (or 8B)

The symbol $\approx$ means, "is approximately equal to"

The gain per stage at $p = p_o$ is:

$$g_m \frac{1}{\beta C}$$ (See Appendix 8A)

This is about half the gain obtainable per stage with two tuned circuits per stage where the gain is:

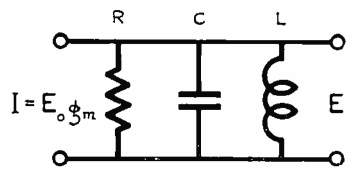


Fig. 8—Coupling network having a single tuned circuit.

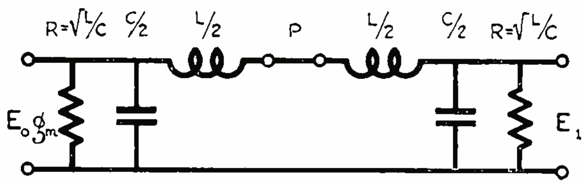


Fig. 9—A one-section low-pass filter.

$$g_m \frac{1}{\beta \sqrt{C_p C_s}}$$

because the minimum value for C is $C_p + C_s \approx 2\sqrt{C_p C_s}$ (if C_s and C_p are nearly equal).

Comparison to Conventional Filters

It is of some interest to compare the performance of maximally flat amplifiers to conventional constant K filters. A conventional one section low-pass filter terminated in a fixed resistance at each end, as shown in Figure 9, is found to have the same curve as the maximally flat cascade filter with $n = 3$. This seems to be somewhat of a coincidence because a two section low-pass filter with the same termination as shown in Figure 10, is inferior to a maximally flat amplifier, with $n = 5$, though it has the same number of reactance elements. These statements are proved in Appendix 9.

If the constants of the circuit of Figure 10 are changed to the proper values, it can no doubt be made maximally flat, although it is believed that finding the proper values would be very difficult. Additional resistors in other meshes would probably be required as well as changes in the values of inductance and capacity.

CONCLUSION

It has been shown that the performance curves of conventional cascade amplifiers may be improved by using unlike parameters for the various stages. Conditions resulting in maximal flatness have been described.

Unfortunately, the circuits employed in obtaining low-pass curves, do not involve shunt condensers on both the input and output. Since both are unavoidable in practice, this arrangement is only suitable

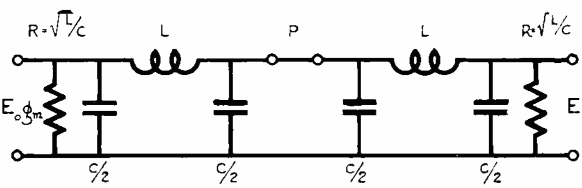


Fig. 10—A two-section low-pass filter.

when other circuit parameters are such that the shunt tube capacity may be neglected. Thus, the maximally flat low-pass arrangement is not suitable if the maximum product of gain times band width is desired. If the requirements are a reasonable gain, a flat pass-band, and a sharp cut-off with a minimum of circuit elements, then it may be the best arrangement.

The same defect applies to the band-pass arrangements with the exception of that shown in Figure 7-b. A maximally flat amplifier composed of stages like Figure 7-b does have the maximum product of gain times band width obtainable. Hence, it is one of the most practical of the circuits shown. If one tuned circuit per stage is used, the gain per stage is about one half of that obtainable with coupled circuits as in Figure 7-b.

The conception of maximal flatness applies equally well to filter-like structures in which the reactance elements are all connected together into a single network, but the mathematical simplicity is lost when an effort is made to find the proper values of the reactive elements.

ACKNOWLEDGMENT

The writer wishes to express his appreciation to Dr. J. L. Barnes and to Mr. D. Nason for valuable assistance in preparing the paper.

LIST OF SYMBOLS

n = the number of reactance elements in the low-pass case.
= the number of tuned circuits in the band-pass cases.

E_o = the input voltage of an amplifier, or of the stage under discussion.

E_1 = the output voltage of the first stage, or of the stage under discussion.

E_m = the output voltage of the mth stage.

I = the plate current due to applied grid voltage E.

g_m = the mutual conductance of amplifier tube.

A = the amplification at zero frequency for the low-pass case.
= the amplification at the mean frequency of the pass-band in the band-pass cases.

A_m = the value of A in the mth stage.

A_a, A_b = the gain of a given (mistuned) stage at its resonant frequency.

β = angular frequency band width of amplifier at $1/\sqrt{2}$ response.

α = damping factor.

ω = any angular frequency in low-pass case.
= distance between any two angular frequencies of equal attenuation, in band-pass case.

p = any angular frequency in band-pass case.

p_1 = any arbitrary value of p.

p_2 = the angular frequency at which the reactance (of a circuit resonant to p_o) is equal and opposite to that at p_1.

p_o = the mean angular frequency of the pass-band.

p_a, p_b = the resonant angular frequencies of a pair of single-tuned-circuit stages; one tuned on each side of p_o.

e = the base of the Naperian system of logarithms.

$C_1, C_2 -- C_{n-2}, C_{n-1}$ = constants.

$C_o = \beta / p_o$.

$\mu_1, \mu_2 --- \mu_m -- \mu_n$ = the various nth roots of -1.

$j = \sqrt{-1}$

m = any integer (between limits as defined where used).

q = ratio of reactance to resistance in low-pass case.
= ratio of reactance at band width frequency to resistance, in band-pass case.

Q = ratio of reactance at resonance to resistance.

$$\phi = \frac{\pi}{2n}(2m-1).$$

θ = angle of retardation at angular frequency ω.

$w = 2 \sinh x = p/p_o - p_o/p$.

$s = 2 \sinh y = p_b/p_o - p_o/p_b = p_o/p_a - p_a/p_o$.

$x = \ln p/p_o$

$y = \ln p_o/p_a$

$$K = \frac{M}{\sqrt{L_p L_s}}$$

r, R, L, C, L_p, C_p, R_p, R_s, M are defined by labeling in the figures.

$$b = \beta/p_o \cos\left[\frac{\pi}{2n}(2m-1)\right]$$

$f_c = \beta/2\pi$.

(To be continued).

BAND WIDTH AND READABILITY IN FREQUENCY MODULATION

By

Murray G. Crosby

R.C.A. Communications, Inc., Riverhead, N. Y.

Summary—*The signal-to-noise ratio characteristics of frequency modulation are considered with reference to the effect of band width on readability in voice communication. Results of listening tests are described which indicate the optimum deviation ratio to be unity for maximum readability and distance of transmission. Other tests show the superiority of frequency modulation over amplitude modulation with respect to readability.*

ONE of the first questions which arises when frequency modulation is chosen for a communication system is that of the band width or amount of frequency deviation to be used. Among the many factors which affect the choice of band width is the subject of quality or readability of signal required. It is the purpose of this article to consider that subject and to attempt to show the relationship between band width and readability.

When the term readability is mentioned, consideration is limited to types of service in which the primary object of the system is the bare transfer of intelligence. This confines the consideration to services where the desire is to transmit voice the maximum distance with full readability for, while a high signal-to-noise ratio is desirable in the reception of voice, increasing the ratio above a certain value does not improve the readability. Consequently if the high signal-to-noise ratio is obtained at the expense of the ability to receive weaker signals, which is the case when the band width is made too wide in a frequency modulation system, the system is not working at its best efficiency. As will be shown here, when the primary object is the transmission of intelligence, the maximum distance will be covered when the band width is made the minimum possible.

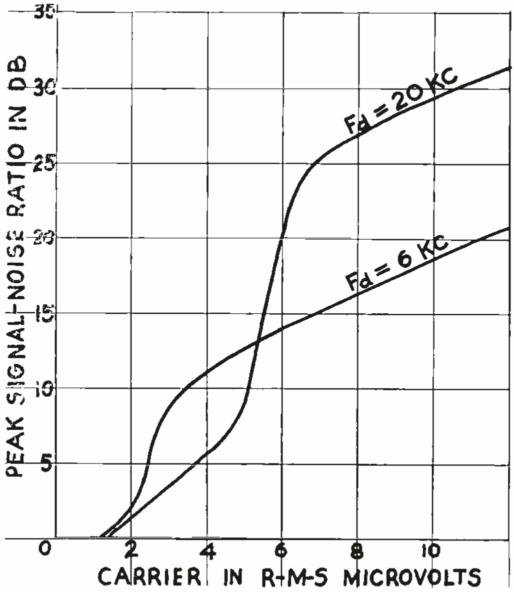


Fig. 1—Peak signal-to-noise ratio characteristics for frequency-modulation systems using maximum frequency deviations, F_d, of 20 and 6 kilocycles, respectively. The audio band width was 5 kilocycles. The noise was internal receiver noise and the signal-to-noise ratios were measured on an oscilloscope.

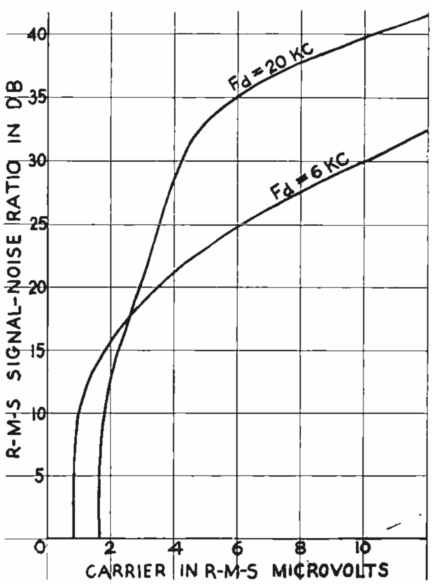


Fig. 2—Root-mean-square signal-to-noise ratio characteristics for the same systems used for Fig. 1. The signal-to-noise ratios were measured on a rectifier-type meter.

The reason for this superiority of a system using a low frequency deviation over one using a high frequency deviation, when maximum distance is the consideration, can best be shown by a study of the curves of Figures 1 and 2. These curves compare the peak and root-mean-square signal-to-noise ratio characteristics of two frequency modulation systems having maximum frequency deviations* of 6 and 20 kilocycles, respectively. The curves were taken by varying the carrier strength of the frequency-modulated signal generator and measuring the signal-to-noise ratio at the output of each receiver for the full frequency deviation that each receiver was capable of. The radio-frequency input circuit of the two receivers was common so that both receivers were on an equal basis as far as carrier strength and input noise were concerned. The noise consisted of the thermal agitation and tube hiss within the receiver. For the curves of Figure 1, the signal and noise in the receiver output were measured by means of an oscilloscope which indicates peak voltages. For those of Figure 2, an ordinary rectifier-type meter was used so that root-mean-square signal-to-noise ratios were obtained.

From these curves, it can be seen that, for carriers below a certain value, the low-deviation system produces a greater signal-to-noise ratio and is therefore more capable of "reaching down" into the noise to receive a signal. It will be noted that both systems give a signal-to-noise ratio which is approximately proportional to the carrier strength down to a certain carrier strength. Below that strength there is a rather sudden drop-off in signal-to-noise ratio. This drop-off is due to a phenomena which is peculiar to a frequency modulation system[1] and which is called the "improvement threshold" effect. Any frequency modulation system has an improvement threshold above which the frequency modulation gain or improvement is realized and below which the signal becomes submerged in the noise. The threshold occurs when the peak voltage of the carrier is equal to the peak voltage of the noise in the intermediate-frequency channel of the receiver. The full frequency modulation improvement is not realized until the carrier is about twice as strong as the noise.

It will be noted that the improvement threshold for the wider systems of Figures 1 and 2 occurs at a stronger carrier strength than that for the narrower system. Hence, other things being equal, the wider system requires more transmitter power to produce a signal which will be above the threshold. The reason for this will be apparent when it is realized that the wider receiver must have a wider

* "Deviation" in this paper refers to the amount of frequency shift to *one side* of the carrier.

[1] Murray G. Crosby, "Frequency Modulation Noise Characteristics", *Proc. of I.R.E.*, Vol. 25, No. 4, April 1937.

intermediate-frequency channel which inherently accepts a wider spectrum of the noise. The wider spectrum of noise has a larger peak voltage so that the improvement threshold distance for the wide system occurs at a higher carrier strength. Thus the threshold for the wider system of Figure 1 occurs at a carrier strength which is about twice the corresponding strength for the narrower system.

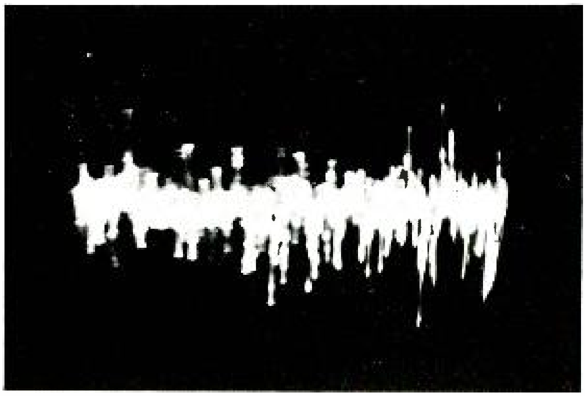
The difference between the shapes of the curves of Figures 1 and 2 is due to differences in crest factor (the ratio between the peak and root-mean-square voltage) of the noise. For carrier strengths above the improvement threshold, the crest factor is constant at a ratio of about 4.5. When the carrier is on the improvement threshold, the crest factor is higher by an amount which depends upon the deviation ratio (the ratio between the maximum frequency deviation and the maximum audio frequency of the system) of the frequency modulation receiver. This increase in crest factor is caused by the fact that the higher peaks of the noise approach equality with the carrier so that these peaks are at the threshold while the lower peaks produce a carrier-to-noise ratio which is above the threshold. When the peak voltages of the carrier and noise approach equality, the effective frequency variation of the resultant wave rises to very high values and the stronger of the two voltages assumes control of the receiver. Thus, if the noise is stronger than the signal the noise assumes control and depresses the signal. Hence as the carrier is lowered towards the threshold, the effective frequency deviation of the noise rises until the higher peaks begin punching holes in the signal. As the carrier is weakened still further, the weaker noise peaks also punch holes in the signal so that it is submerged in the noise. The point at which the highest peaks of the noise just begin to reach equality with the carrier peak voltage, produces a "sputtering" type of noise which changes the character of tube hiss or thermal agitation so that the improvement threshold is easily recognized. Figures 3A and B show the wave form of the noise at the threshold (which has been called the "sputter point") and above the threshold respectively.

When the noise is ignition or similar man-made noise the situation is similar to that shown by the peak signal-noise ratio curves of Figure 1. In addition, the frequency modulation receiver has inherent to it a noise-silencing action which is at least as effective if not better than the best amplitude-modulation noise silencer. This noise-silencing action is self-adjusting and is automatically adjusted for best operation as soon as the signal is tuned in. More detailed description of this action is considered elsewhere.[1,2]

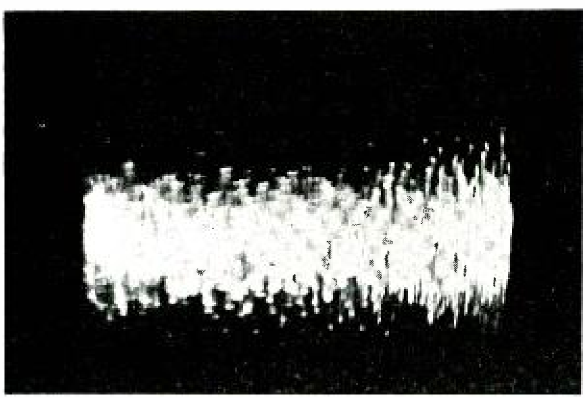
[2] Murray G. Crosby, "The Service Range of Frequency Modulation", *RCA Review*, Vol. 4, No. 3, January 1940.

Listening Tests

With the signal-to-noise ratio characteristics as described above, it can be seen that the answer to the question regarding the relative readability obtainable with systems using different maximum frequency deviations depends upon the magnitude of signal-to-noise ratio required for a given readability. For instance, taking the curves of Figure 1,



A



B

Fig. 3—Oscillograms of internal receiver noise appearing at the frequency modulation receiver output. A is with carrier-to-noise ratio at the threshold or "sputter point". B is with carrier-to-noise ratio above the threshold.

if readability is obtainable on the narrower system for the signal strengths which are below the threshold of the wider system, the narrower system will have a range of superiority in readability of weaker signals. As will be seen from the following curves, practically full readability is obtained right down to the improvement threshold of the narrower system. With the wider system, full readability is not obtainable until the signal is strong enough to reach the threshold of the wider system.

In order to determine the actual difference in signal readability for systems using different frequency deviations, a listening test was conducted on the two frequency-modulation systems which were used for Figures 1 and 2. A 5-kilocycle low-pass filter was inserted in the audio output of both receivers so that one had a deviation ratio of 1.2 and the other 4. The noise consisted of the thermal agitation and tube hiss originating in the radio-frequency circuits.

The curves of Figure 4 show the results of the listening tests. Readability numbers of the amateur RST system are plotted against

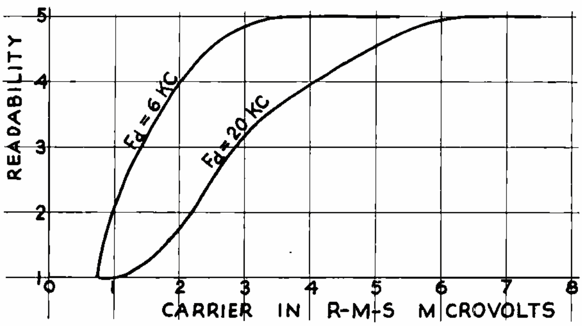


Fig. 4—Readability comparison of frequency modulation systems using frequency deviations of 20 and 6 kilocycles, respectively. The a.f. of both receivers cut off at 5 kilocycles. The ordinates are plotted in the readability scale of the RST signal-reporting system which is as follows:

1—Unreadable.
2—Barely readable, occasional words distinguishable.
3—Readable with considerable difficulty.
4—Readable with practically no difficulty.
5—Perfectly readable.

the microvolts output of the frequency-modulated signal generator. The points for the curves were taken from the averaged readings of three separate observers, namely A. M. Braaten, R. E. Schock, and the writer.

It is obvious from these tests that the narrower system is capable of "dipping down" deeper in the noise to receive a signal. For equal readability on the two systems, it appears that the system with a deviation ratio of 1.2 will receive a signal about one-half as strong as that possible with the system having a deviation ratio of 4. Hence the minimum readable signal strength decreases approximately proportional to the square-root of the ratio of the two deviation ratios. In the particular case of the systems used for Figure 4, changing the deviation ratio of the system from 4 to 1.2 is equivalent to an increase in power of about 4 times, as far as readability is concerned.

The curves of Figure 5 were taken in the same manner as those of Figure 4, but the two systems compared were the narrower frequency-modulation system of Figures 1 and 2 and its equivalent amplitude-modulation system. The maximum frequency deviation of the frequency-modulation system was 6 kilocycles and the audio band was 5 kilocycles. It can be seen that the narrower frequency-modulation system gives a readability which is always greater than that obtained on the amplitude-modulation system.

It is apparent from these tests that the optimum frequency deviation for a frequency-modulation system designed to obtain maximum distance for full readability is that which corresponds to a deviation ratio of one. Such a system produces very nearly the same intermediate-frequency and audio channel widths as the corresponding amplitude-

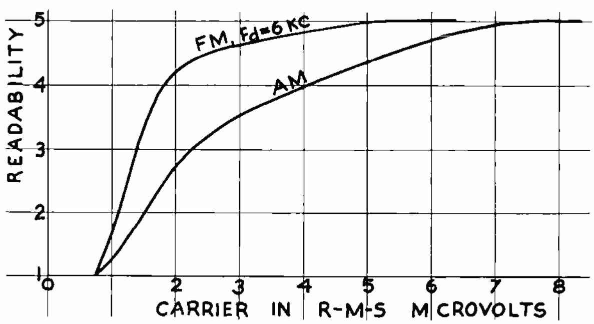


Fig. 5—Readability comparison of a frequency-modulation system and its equivalent amplitude-modulation system. Maximum frequency deviation = 6 kilocycles. Audio band = 5 kilocycles.

modulation system. To use a deviation ratio less than one would of course impair the efficiency without a compensating benefit in reduction of band width since it would be equivalent to an amplitude-modulation system which employed a modulation percentage less than 100.

The results of other listening tests[2] conducted by the writer have indicated that the use of pre-emphasis and de-emphasis, as is used in the present broadcast frequency-modulation systems, is of doubtful value when the object is the mere transmission of intelligibility instead of the enjoyment of a high-fidelity program. The peculiar triangular nature of the noise spectrum in the output of a frequency-modulation receiver is apparently more tolerable than the same noise spectrum after it has been made practically flat by the de-emphasis circuit. The results of the averaged observations of three observers showed that about 8 decibels more noise could be tolerated with the triangular frequency-modulation noise than with the rectangular amplitude-modulation noise. That is to say that when the intelligibility of voice is being received through fluctuation noise like tube hiss or thermal agitation, the same intelligibility may be received with 8 decibels more noise of the triangular characteristic obtained from the output of a frequency-modulation receiver than with the flat or rectangular noise characteristic obtained from the output of an amplitude-modulation receiver. This advantage is lost when pre-emphasis and de-emphasis are used because the triangular noise spectrum is converted to a rectangular spectrum.

FLUCTUATIONS IN SPACE-CHARGE-LIMITED CURRENTS AT MODERATELY HIGH FREQUENCIES

BY

B. J. THOMPSON, D. O. NORTH, AND W. A. HARRIS

RCA Manufacturing Company, Inc., Harrison, N. J.

PART IV — FLUCTUATIONS CAUSED BY COLLISION IONIZATION

BY

B. J. THOMPSON AND D. O. NORTH

Summary—*An expression is derived for the number of electrons released from the virtual cathode by each positive ion formed by collision. The plate-current fluctuations originating in the random formation of such ions are then formulated for a simplified, parallel-plane model approximating conventional sleeve-cathode structures. The formula does not require a knowledge of gas pressure, and thus is especially useful in connection with sealed-off tubes. Calculations from the theory, under the assumption that the chief constituent of residual gas is carbon monoxide, compare favorably with a series of measurements. It is concluded that the "gas" component of noise may generally be ignored when grid gas-current is less than a few hundredths of a microampere.*

INTRODUCTION

IN EVERY vacuum tube there are free gas molecules some of which are converted into positive ions by collision with electrons in those regions where the potential exceeds the ionization potential. Throughout its lifetime each ion affects the virtual cathode, if one exists, in such a way that the electron current is increased. Inasmuch as the formation of ions is a matter of chance, the increase in electron current caused by the ions must fluctuate at random. Thus, the plate current exhibits a "gas" component of noise in addition to the normal shot fluctuations investigated in Parts II and III. The results of an analytical and experimental study of this ionization noise are presented in this part.

It should be noted at the outset that no significant ionization noise is found in modern high-vacuum receiving tubes. The subject of this part of the series is, therefore, not of such serious import as it was so little as ten years ago. This is a direct tribute to present-day exhaust technique, particularly the use of effective new getters. Today's new tube with an estimated gas pressure of perhaps 10^{-4} micron has a grid current of less than 10^{-8} ampere, a value so small that it may be due as much to photoelectric emission as to the collection of positive ions. Under the circumstances, the objective of this investigation has been primarily to establish a sound quantitative understanding of the phenomena involved, a result of greater theoretical than practical interest.

In 1933 Ballantine[1] studied this same problem. Because he did not attempt to determine the number of electrons released from the virtual cathode by an ion during its lifetime, his analysis led to conclusions which can be quantitatively applied only to tubes in which both the gas

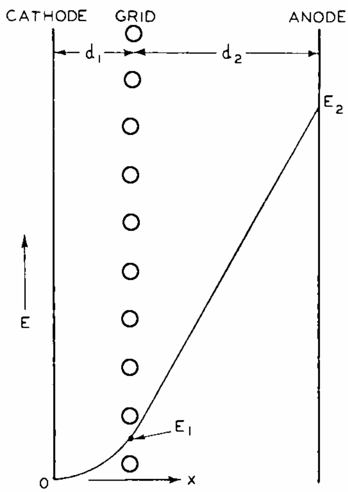


Fig. 1—Space potential in a model triode.

pressure and the increment in plate current caused by ionization are known. His qualitative predictions were well demonstrated in a series of careful observations of the noise produced by controlled amounts of mercury vapor, argon, and gas evolved from the tube structure.

The analysis below, begun in 1933, determines the number of electrons released per ion and leads to an end-formula which relates the noise to the grid gas-current, tube geometry, and electrode potentials, but, unlike Ballantine's, does not require a knowledge of the increment in plate current. Instead, this increment is also obtained as a function of the above parameters. One of the chief merits of this work is that it bears directly upon the performance of sealed-off tubes in which the gas pressure is not directly observed.

[1] S. Ballantine, "Fluctuation Noise due to Collision Ionization in Electronic Amplifier Tubes", *Physics*, Vol. 4, No. 9, pp. 294-306; September (1933).

The Analytical Model

For simplicity only the parallel-plane case is considered, and, unless otherwise specified, this will be a triode with negative grid bias. The space distribution of the gas is considered uniform in the regions where ionization occurs. It is assumed that each ion is formed with zero velocity, moves in a straight line under the acceleration of the electric field, and is collected (or neutralized) at first encounter with an electrode. Since carbon monoxide is known to be the chief component of residual gas in tubes with coated cathodes and since it has been found to have a representative probability of ionization, we shall confine our analysis to this gas. It is then proper[2] to ignore collision products other than the singly charged ion, CO^+; we can also ignore multiple collisions altogether.

The ionization potential of CO is about 14 volts[2,3]. With reference to Figure 1, the effective potential E_1 of the grid surface in conventional tubes is well below this figure. It will, therefore, be supposed that no ion is formed in the cathode-grid space and that the division of ion current between grid and cathode can be approximated in terms of the fraction α of open grid surface.

To avoid discussing transit-time effects, the analysis is limited to conditions in which the transit time of an ion is short in comparison with the period of the measuring circuit. Our conclusions will, therefore, be applicable only to frequencies less than about ten megacycles in conventional receiving tubes. However, the threshold effect of increasing transit angles is known to be a reduction of plate-current noise, so that we are, for small transit angles, calculating an upper limit.

As we have previously stated, ionization increases not only the fluctuations, but also the average electron current. It will be supposed that the ion-dependent increment in electron current is small, so that the actual operating parameters of the tube, such as transconductance, are satisfactorily approximated by their ion-free values.

The gas current to the grid will, of course, produce a shot voltage across the input circuit. In comparison with the plate-current fluctuations this source of noise can usually be ignored in radio-frequency amplifiers. It must be reckoned with in high-input-impedance d-c

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[2] A. L. Vaughan, "Mass Spectrograph Analysis and Critical Potentials for the Production of Ions by Electron Impact in Nitrogen and Carbon Monoxide", *Phys. Rev.*, Vol. 38, pp. 1687-1695; November (1931).

[3] John T. Tate and P. T. Smith, "The Efficiencies of Ionization and Ionization Potentials of Various Gases Under Electron Impact", *Phys. Rev.*, Vol. 39, pp. 270-277; January (1932).

amplifiers, where it may become the dominant noise source.[4] But since the calculation in the d-c case is straightforward, following the standard formula for temperature-limited shot effect, it will also be ignored at this time. Our model tube will have its grid at ground potential for currents at the operating frequency.

The space potential in the model is shown in Figure 1. Electrons are emitted with zero velocity. The familiar three-halves-power law is used in the cathode-grid space and the field is approximately uniform in the grid-anode space. Hence, for the symbols shown in Figure 1, the space potential E is determined by

$$\left.\begin{aligned} \frac{E}{E_1} &= \left(\frac{x}{d_1}\right)^{4/3} \qquad & o \leqslant x \leqslant d_1 \\ \frac{\partial E}{\partial x} &= \frac{E_2 - E_1}{d_2} \qquad & d_1 \leqslant x \leqslant d_2 \end{aligned}\right\} \qquad (1)$$

Because E_1 is small in comparison with the ionization potential, every ion which enters the cathode-grid space is assumed to move therein with a constant velocity equal to its terminal velocity at the cathode.

If the amplification factor (μ) is large enough, the grid may be regarded as a perfect shield (permeable membrane), inhibiting the influence which each ion exerts upon the virtual cathode until the moment when the ion has penetrated the grid surface. We have found, however, that μ's of conventional tubes are *not* large enough to permit this approximation, and thus the influence of both ions and electrons in the grid-anode space has to be admitted. The extended analysis is too lengthy to warrant presentation; hence the end-formulas alone will be listed. The analysis detailed below refers merely to the simplified, high-μ tube, but the procedure followed to extend it will be outlined.

Increment in Electron Current

Let I = average cathode (or anode) electron current in the absence of gas.

δI = increment in I as a consequence of ionization.

v_e = electron velocity at the grid.

v_i = terminal ion velocity at the cathode.

α = fraction of open grid surface.

[4] E.g., L. R. Hafstad, "The Application of the FP-54 Pliotron to Atomic Disintegration-Studies", *Phys. Rev.*, Vol. 44, pp. 201-213; August (1933).

I_i = total average ion current arriving at grid surface.
I_{ig} = total average ion current absorbed at the grid.
I_{ik} = total average ion current absorbed at the cathode.

Then we have $I_i = I_{ig} + I_{ik}$
and assume $I_{ik}/I_i \approx \alpha$.
Let m_i = mass of an ion.
m_e = mass of an electron.
M = molecular weight of an ion (28 for CO).

Then
$$\left(\frac{m_i}{m_e}\right) = 1833\, M \quad (5.13 \times 10^4 \text{ for } CO).$$

Our first objective is to determine the ratio of the increment in electron current caused by the ions to the ion current flowing to the cathode from a point of ion formation (potential E) in the grid-anode space. The calculation pivots around two principles. First, the net surface charge on the cathode must be zero (no field) at every instant. Second, the ion flow may be regarded as a movement of charge continuously distributed in space exactly as is customary for electron-current analyses.

Let us suppose that the cathode ion current I_{ik} consists of ions of just one common velocity (all formed at points having space potential E). The charge it induces on the cathode is

$$q_i = -\int_0^{d_1} \frac{I_{ik}}{v}\left(1 - \frac{x}{d_1}\right) dx = -\frac{1}{2}\,\frac{I_{ik} d_1}{v_i}.$$

The charge induced on the cathode by the electrons comprising the increment δI in electron current elicited by the presence of ions is

$$q_e = \int_0^{d_1} \frac{\delta I}{v}\left(1 - \frac{x}{d_1}\right) dx = \frac{\delta I}{v_e}\int_0^{d_1}\left(\frac{d_1}{x}\right)^{2/3}\left(1 - \frac{x}{d_1}\right) dx = \frac{9}{4}\,\frac{\delta I\, d_1}{v_e}.$$

If the field at the cathode is zero,

$$q_e + q_i = 0,$$

and

$$\frac{\delta I}{I_{ik}} = \frac{2}{9}\,\frac{v_e}{v_i} = \frac{2}{9}\left(\frac{m_i}{m_e}\right)^{1/2}\left(\frac{E_1}{E}\right)^{1/2} = 9.53\, M^{1/2}\left(\frac{E_1}{E}\right)^{1/2}$$

Since the ions are singly charged, the ratio likewise signifies the number of electrons $N(E)$ released from the virtual cathode by each ion of this terminal velocity,

$$N(E) = \frac{\delta I}{I_{ik}} = 9.53\ M^{1/2} \left(\frac{E_i}{E} \right)^{1/2} \qquad (2)$$

indicating approximately 10 electrons for each CO ion which reaches the cathode.

Equation (2) shows the increment in electron current under the assumption that all ions have the same velocity (formed at potential E), but it also shows the increment in electron current actually caused by the ions in each velocity class (formed between potentials E and $E + dE$). Knowing the number of ions formed at each potential, we may then determine the total increment in electron current. Let $P(E)$ be the average number of ions formed by each electron of energy E for each centimeter of its path[5]. The total increment in electron current as a consequence of ionization is, therefore,

$$\delta I = \alpha I \int_0^{E_2} N(E)\ P(E)\ dx = \frac{\alpha I}{\left(\dfrac{\partial E}{\partial x} \right)} \int_0^{E_2} N(E)\ P(E)\ dE. \qquad (3)$$

Here $\left(\dfrac{\partial E}{\partial x} \right)$ is the uniform field in the grid-anode space, and the lower limit to the integration is set at zero merely for convenience. There is, of course, no contribution to the integral from potentials below the ionization potential.

Similarly, the *total* ion current is

$$I_i = \frac{I}{\left(\dfrac{\partial E}{\partial x} \right)} \int_0^{E_2} P(E)\ dE,$$

and the grid gas-current is

$$I_{ig} = (1 - \alpha) I_i = \frac{(1 - \alpha) I}{\left(\dfrac{\partial E}{\partial x} \right)} \int_0^{E_2} P(E)\ dE. \qquad (4)$$

[5] Tate and Smith, Reference 3, have measured this quantity for a number of gases. Their data for CO will be used.

By combining (3) and (4), and using (2), we obtain

$$\delta I = 9.53\, M^{1/2} \frac{\alpha}{1-\alpha} E_1^{1/2} \frac{\displaystyle\int_0^{E_2} \frac{P(E)}{E^{1/2}}\, dE}{\displaystyle\int_0^{E_2} P(E)\, dE} I_{ig}.$$

Let us define a quantity

$$E_t^{1/2} = \frac{\displaystyle\int_0^{E_2} P(E)\, dE}{\displaystyle\int_0^{E_2} \frac{P(E)}{E^{1/2}}\, dE}. \tag{5}$$

We then find

$$\delta I = 9.53\, M^{1/2} \frac{\alpha}{1-\alpha} \left(\frac{E_1}{E_t} \right)^{1/2} I_{ig}. \tag{6}$$

E_t is, in a sense, a "mean" potential at which ions are formed. Since $P(E)$ appears in both numerator and denominator of (5), E_t is independent of the gas density. The experimental values of $P(E)$ published by Tate and Smith are shown for CO in Figure 2. The data refer to a gas density corresponding to a pressure of one millimeter and a temperature of zero Centigrade. E_t was computed from these data and is plotted as a function of E_2 in Figure 3.

The Fluctuations

It is illuminating to express the plate current fluctuations in terms of the temperature-limited diode current which exhibits equal fluctuations. In a diode current J the fluctuations are expressed[6] by

$$\overline{i^2} = 2eJ\Delta f.$$

Since the ion current exhibits random fluctuations they can be described in the same way.

We have observed that a given velocity class of ions produces an increment in electron current proportional to the ion current within the class and depending on the velocity of the ions [see Equation (2)].

[6] See Part II, p. 450; April (1940).

The ion current within each velocity class will fluctuate at random. The increment in electron current caused by each class will fluctuate by a proportional amount. Therefore, we may represent that portion of the plate-current fluctuations due to ion effects by a temperature-limited diode current J_1 as follows:

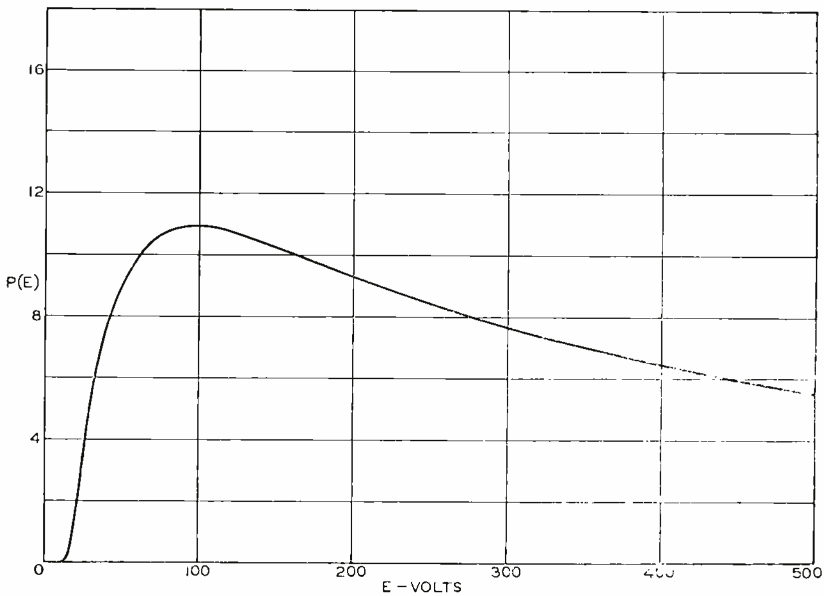

Fig. 2—Average number of $CO+$ ions formed by each electron of energy E for each centimeter of its path in carbon monoxide at one millimeter pressure, zero degrees Centigrade. Data from Tate and Smith.

$$J_1 = \alpha I \int_0^{E_2} N^2(E)\ P(E)dx = \frac{\alpha I}{\left(\dfrac{\partial E}{\partial x}\right)} \int_0^{E_2} N^2(E)\ P(E)dE. \tag{7}$$

By combining (3) and (7) and defining a quantity,

$$E_s = \frac{\displaystyle\int_0^{E_2} P(E)dE}{\displaystyle\int_0^{E_2} \frac{P(E)}{E}\, dE}, \tag{8}$$

we find

$$J_1 = 90.8\, M \frac{\alpha}{1-\alpha} \left(\frac{E_1}{E_s} \right) I_{ig}\,. \tag{9}$$

E_s is, in *another* sense, a "mean" potential at which ions are formed. It is also plotted in Figure 3 for CO.

Formally, this simplified analysis of high-μ structures is now complete, and can be tested by inserting a comparator diode across the

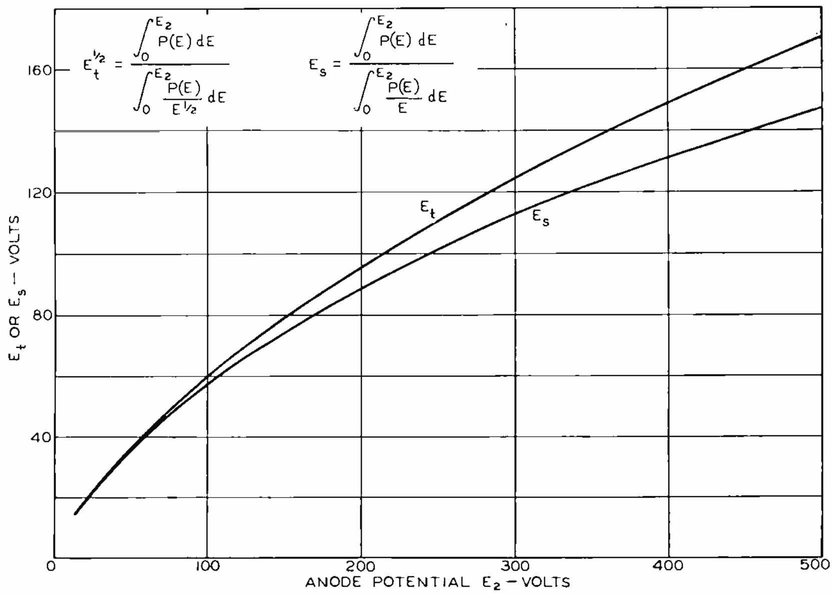

Fig. 3—Effective potentials of ion formation in carbon monoxide, as a function of anode potential.

plate circuit and determining J_1 experimentally. The quantities appearing on the right hand side of (9) can all be determined to a fair approximation. E_1 cannot be measured directly, but, in view of the three-halves-power law, can be expressed in terms of current and transconductance by

$$E_1 = \frac{3}{2} \frac{I}{g_m} \tag{10}$$

Extension of Analysis to Structures with Finite μ

The space-charge-free electrical equivalent of a three-electrode tube structure is a combination of three capacitances which may, with equal generality, be represented in either delta or star connection. The delta concept is commonly employed in tube engineering because the capacitances so defined (Maxwell's coefficients of induction) are easily measured and blend without struggle into networks attached to the tube. But for all problems relating to the distribution of the images of a charge moving through the structure, the star connection is far superior[7]. In Figure 4 the central connection A is not accessible, but plays an important rôle in that its potential corresponds to the "equivalent-diode potential". When a tube approximates the ideal parallel-plane structure, the capacitances (per unit area) are closely approximated by the following simple expressions:

$$C_k = \frac{1}{4\pi d_1}, \quad C_p = \frac{1}{4\pi d_2}, \quad C_g = \mu C_p \tag{11}$$

Adaptation of these expressions to simple structures without plane symmetry is functionally obvious.

A charge e in transit may now be regarded as moving, first, through a space (C_k) bounded by the cathode and a hypothetical "permeable membrane" at A, then through a similar grid-anode space (C_p). Its images (q) in the three electrodes can therefore be written with ease. For the parallel-plane structure they are

$$\begin{array}{l} -\dfrac{q_k}{e} = \\[2ex] -\dfrac{q_g}{e} = \\[2ex] -\dfrac{q_p}{e} = \end{array} \left\{ \begin{array}{c} \dfrac{(\mu+1)(d_1-x)+d_2}{(\mu+1)d_1+d_2} \\[2ex] \dfrac{\mu x}{(\mu+1)d_1+d_2} \\[2ex] \dfrac{x}{(\mu+1)d_1+d_2} \end{array} \right\} \begin{array}{c} = \\[2ex] = \\[2ex] = \end{array} \left\{ \begin{array}{c} \dfrac{d_1+d_2-x}{(\mu+1)d_1+d_2} \\[2ex] \mu\dfrac{d_1}{d_2}\dfrac{d_1+d_2-x}{(\mu+1)d_1+d_2} \\[2ex] \dfrac{\mu\dfrac{d_1}{d_2}(x-d_1)+x}{(\mu+1)d_1+d_2} \end{array} \right\} \tag{12}$$

$$o \leqslant x \leqslant d_1 \qquad\qquad d_1 \leqslant x \leqslant d_2$$

With these formulas, the extension of the simple ionization analysis is readily accomplished. To express the end-formulas concisely, we need a few additional parameters which are defined as follows:

[7] See, e.g., F. B. Llewellyn, "Operation of Ultra-High-Frequency Vacuum Tubes", *B.S.T.J.*, Vol. 14, p. 659; October (1935).

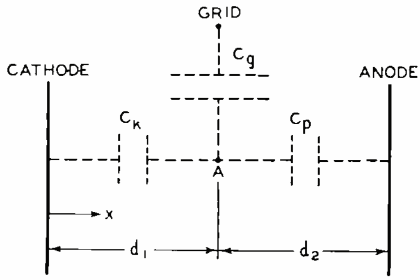

Fig. 4—Electrostatic representation of a triode.

$$\left.\begin{aligned}
h &= \frac{\tau_2}{\tau_1} = \frac{\text{grid-anode electron transit time}}{\text{cathode-grid electron transit time}} \\
\sigma &= \frac{\mu}{1+\mu+\dfrac{4}{3}\dfrac{d_2}{d_1}(1+h)-\dfrac{1}{3}h^2(6+4h+h^2)} \\
&= \frac{g_m}{g} = \frac{\text{transconductance}^8}{\text{conductance of equivalent diode}} \\
\lambda &= \frac{1+\mu+2\dfrac{d_2}{d_1}}{1+\mu+\dfrac{4}{3}\dfrac{d_2}{d_1}(1+h)-\dfrac{1}{3}h^2(6+4h+h^2)} \\
\beta &= \frac{4}{3}\frac{\left(\dfrac{d_2}{d_1}\right)^2}{1+\mu+2\dfrac{d_2}{d_1}}
\end{aligned}\right\} \qquad (13)$$

We shall also use the following dimensionless functions, which are plotted in Figure 5 for CO:

[8] Cf. Part II, p. 469.

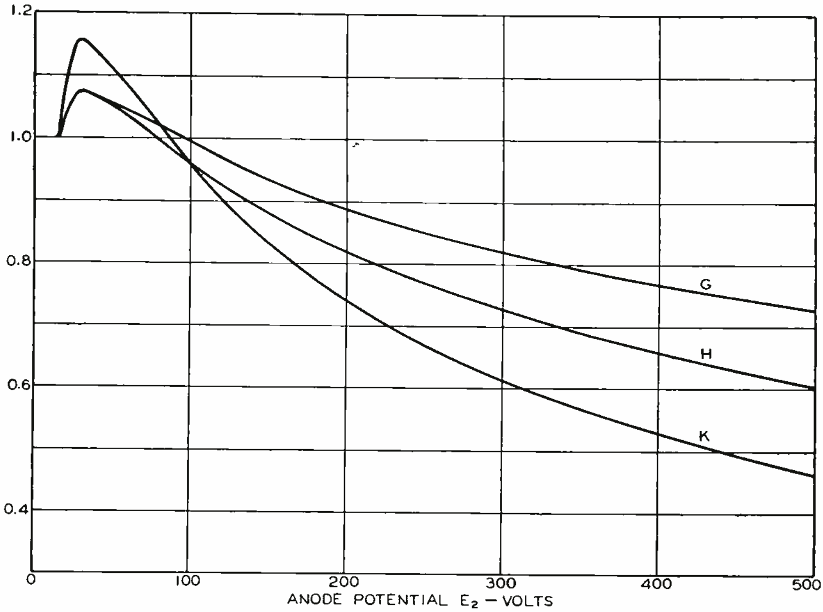

Fig. 5—Functions G, H, K evaluated for carbon monoxide.

$$
\left.
\begin{aligned}
G(E_2) &= \frac{\displaystyle\int_o^{E_2} \left(\frac{E}{E_2}\right)\left(3-2\frac{E}{E_2}\right)\frac{P(E)}{E^{1/2}}\,dE}{\displaystyle\int_o^{E_2}\frac{P(E)}{E^{1/2}}\,dE} \\[2ex]
H(E_2) &= \frac{\displaystyle\int_o^{E_2} \left(\frac{E}{E_2}\right)\left(3-2\frac{E}{E_2}\right)\frac{P(E)}{E}\,dE}{\displaystyle\int_o^{E_2}\frac{P(E)}{E}\,dE} \\[2ex]
K(E_2) &= \frac{\displaystyle\int_o^{E_2} \left(\frac{E}{E_2}\right)^2\left(3-2\frac{E}{E_2}\right)^2\frac{P(E)}{E}\,dE}{\displaystyle\int_o^{E_2}\frac{P(E)}{E}\,dE}
\end{aligned}
\right\} \qquad (14)
$$

In these terms, the number of electrons released from the cathode for each ion which originates at some point of potential E in the grid-anode space, and *ends its career at the grid,* is

$$N_g(E) = 9.53\, M^{1/2} \left(\frac{E_1}{E} \right)^{1/2} \cdot \lambda\beta \left(\frac{E}{E_2} \right) \left(3 - 2\frac{E}{E_2} \right). \qquad (15)$$

But, if the ion *goes through to the cathode,* the total number is

$$N_k(E) = 9.53\, M^{1/2} \left(\frac{E_1}{E} \right)^{1/2} \cdot \lambda \left[1 + \beta \left(\frac{E}{E_2} \right) \left(3 - 2\frac{E}{E_2} \right) \right]. \qquad (16)$$

Superseding (6), the increment in electron current becomes

$$\delta I = 9.53\, M^{1/2} \frac{\alpha}{1-\alpha} \left(\frac{E_1}{E_t} \right)^{1/2} I_{ig} \cdot \lambda^2 \left[1 + \frac{\beta}{\alpha} G \right]. \qquad (17)$$

Superseding (9), the plate-current fluctuations are given by

$$J_1 = 90.8\, M \frac{\alpha}{1-\alpha} \left(\frac{E_1}{E_s} \right)^{1/2} I_{ig} \cdot \lambda^2 \left[1 + 2\beta H + \frac{\beta^2}{\alpha} \mathrm{K} \right] \cdot \qquad (18)$$

Superseding (10), E_1 is now to be computed from

$$E_1 = \sigma \cdot \frac{3}{2} \frac{I}{g_m}. \qquad (19)$$

In general, the extended formula indicates a several-fold increase in J_1 over that derived in (9). The reason is that, although the electrical shielding action of the grid may be great, the ions spend most of their lives approaching the grid and but very little of their lives (if any) in the cathode-grid region.

The extension of this analysis to multi-collector structures is straightforward, following the pattern of Part III. But so little material of additional value accrues that it does not invite elaboration.

The experimental work described below is a direct test of (18). However, in making an advance rough estimate of the fluctuations, the rather large number of unsubstantiated premises and approximations which enter into the calculations probably does not justify all of the refinement. For such a purpose it is accurate enough to assume that λ, β, σ, G, H, and K all have the value unity, and that $\alpha = 0.9$. We then have, for CO gas,

$$J_1 \approx 1.4 \times 10^5 \frac{I}{g_m E_s} I_{ig} . \qquad (20)$$

Reference to Figure 3 suggests that 70 volts is a representative figure for E_s. If, in addition, we suppose I to be about 5 milliamperes, and g_m about 2,000 microhmos,

$$J_1 \approx 5 \times 10^3 I_{ig} .$$

From the treatment of the gas-free space-charge-limited triode in Part II, we know that the normal shot effect in such tubes can be expressed in terms of a temperature-limited-diode current J_o:

$$J_o = \Gamma^2 I \qquad (21)$$

where[9]

$$\Gamma^2 \approx \frac{1.29}{\sigma} \frac{g_m V_e}{I} \qquad (22)$$

and $V_e = \frac{kT}{e}$, T being cathode absolute temperature, k Boltzmann's constant. In typical tubes, Γ^2 is about 0.05. For the representative tube under consideration the ratio of mean square ion-elicited fluctuations to the ion-free space-charge-reduced shot fluctuations is

$$\frac{J_1}{J_o} \approx 10^5 \frac{I_{ig}}{I} = 2 \times 10^7 I_{ig} .$$

If the ion noise is to be kept below, say, twenty per cent of the normal shot effect, the grid gas-current should not rise above one hundredth microampere. So low a figure suggests a basic reason for the poor experimental agreement amongst studies of space-charge-limited shot effect before the advent of modern getters. Even today it represents a severe manufacturing tolerance. It should be pointed out again that, for a constant gas-current, ion noise decreases at higher frequencies. The expectation that large ion-transit-angles will reduce the theoretical value of J_1 has already been confirmed by a few rough comparative measures of J_1 at one and ten megacycles.

Ion-free noise in multi-collector tubes runs higher than in triodes[10]. The mean-square plate fluctuations in receiving-type pentodes run two to four times greater than the figure for triodes expressed by (21). For pentodes, then, the threshold gas currents are proportionately larger.

Measurement

The standard RCA-56 structure was used. This is a cylindrical mount having a 0.050-inch cathode, a 0.255-inch anode, and an oval

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[9] Part II, p. 470, Eq. (43b).
[10] See Part III, pp. 244-260; October (1940).

grid with a minor diameter of 0.095 inch and a major diameter (between side rods) of 0.170 inch. It was anticipated that this structure, representative of many in common use, would correspond closely enough to that of the parallel plane model to permit comparisons, particularly since the bulk of the current is concentrated in a sector in the vicinity of the minor grid diameter, and d_1 and d_2 can, therefore, be estimated with confidence despite grid ellipticity. The structural parameters, (13), were then assigned the following approximate values on the basis of design data:

$$\alpha = 0.88,\ \beta = 0.81,\ \sigma = 0.72,\ \lambda = 1.16.$$

The last two, although functions of the transit-time ratio h, are virtually constant over the range of operation, and were computed for $h = o$.

Noise was measured with the equipment described in Part II at a frequency of about one megacycle, over a band of about ten kilocycles. Original plans called for use of a comparator diode, but it was found difficult to maintain sufficiently high impedance in diodes carrying saturated currents in the order of ten milliamperes. The current fluctuations in such cases are believed to obey the standard formula for true shot effect. But in the presence of the strong fields necessary to produce saturation the emission becomes field-sensitive (Schottky effect), so that accurate work demands recognition of the finite contribution of the diode to the total circuit impedance. We found it less awkward, therefore, to use the signal substitution method, also described in Part II.

Into some of the tubes CO was introduced by heating calcium oxalate in a side arm; $Ca(COO)_2 \rightarrow CaO + CO + CO_2$. A liquid air trap removed the CO_2. After being flushed with about 2 centimeters of gas, the tubes were sealed off at a pressure of about 5 microns. Gas currents, read directly after sealing off, were always several microamperes. All such tubes showed a tendency to "clean up" during operation, but the rate was not sufficient to interfere with measurement. Indeed, the effect could be reversed by overheating the cathode or the plate a small amount, a welcome phenomenon for studies of performance versus gas density.

Into the other tubes studied no gas was introduced. But, after sealing off, high-frequency heating was applied until the gas current rose to the desired value. Gas so evolved was presumably representative of the residual gas in commercial tubes.

Exact interpretation of the data requires a measure of the normal space-charge-limited fluctuations which exist in the absence of gas. To this end both groups of tubes were provided with getters which, at the end of the ionization studies, could be relied upon to clean up the gas.

Measurement of the residual fluctuations agreed well enough with the theoretical expectation (21) on the two occasions in which this procedure was followed, so that, for the rest, the measurement was not considered necessary.

In every respect the performance of the tubes with evolved gas coincided with that of the tubes into which CO had been introduced. This tends to confirm the findings of many gas analyses that CO is a chief constituent of residual gas. There is, however, the possible alternative, that the constituents of the evolved gas had such an average molecular weight and ionization probability[11] as to simulate CO. The question was not pursued.

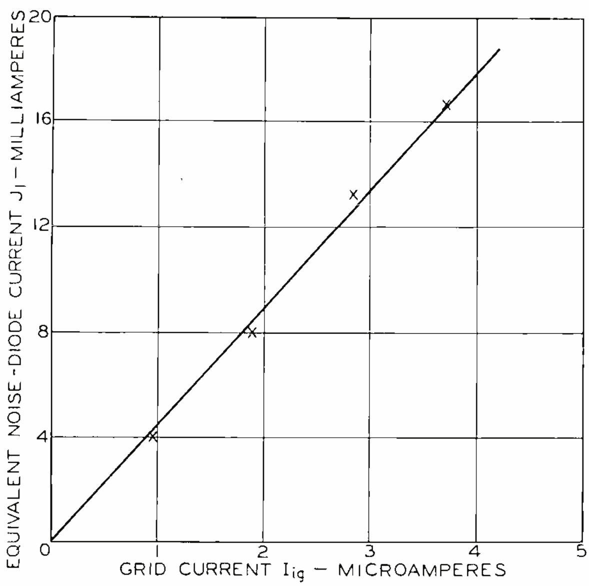

Fig. 6—Measured values of ionization noise versus ion current to the grid of a triode containing carbon monoxide

Juggling the gas content, as described above, yielded the data of Figure 6, demonstrating that ion-provoked noise is proportional to gas current, as one would expect. In this instance, gas was released by raising the cathode temperature in steps from an estimated 980°K to 1180°K, meanwhile holding anode potential at 250 volts, and maintaining a constant cathode current of 8 milliamperes by means of minor adjustments of the grid bias.

Figures 7 and 8 illustrate a representative set of observations of the kind used to test the analysis. This tube contained evolved gas. The anode potential was fixed at 250 volts. Cathode current was varied from 0.5 to 9.0 milliamperes by running the grid bias up from —17 to

[11] E.g., Any mixture of CO, N_2, NO, and O_2; cf. Tate and Smith, loc. cit.

—10 volts. Grid current, transconductance, and noise were measured simultaneously[12]. The experimental values of J_0 were obtained after the run by flashing a getter. Failure to obtain a closer fit to the theoretical curve for J_0 may be due largely to incomplete gettering; the residual gas current amounted to 0.009 microampere for the largest cathode current. Plotted values of J_1 are measured noise corrected by subtraction of the theoretical J_0. From (17), the largest fractional increment in ion-free cathode current was estimated to be 5 per cent, corresponding to a δI of 420 microamperes for an I of 9.0 milliamperes. It, therefore, appears justifiable to ignore the influence of the ions upon transconductance, space potential, etc.

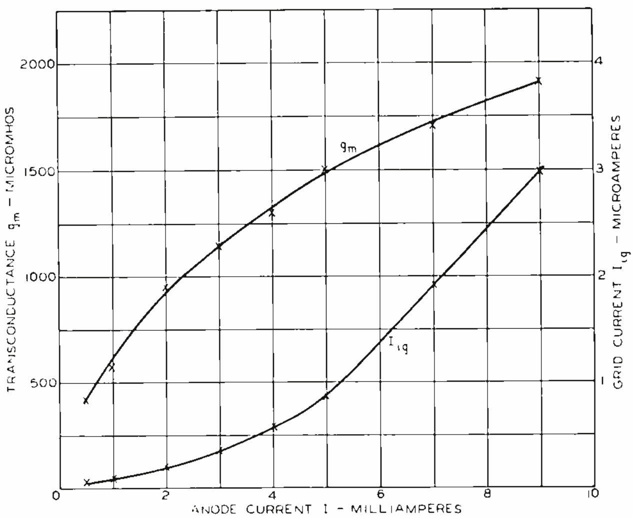

Fig. 7—Transconductance and evolved-gas current to the grid of a triode as a function of anode current. Anode potential fixed at 250 volts.

Measured mean-square noise always exceeded prediction in the manner typified by Figure 8. The most reliable comparison of theory and measurement is made for large J_1, where the measuring sensitivity was best, experimental error least likely, and the importance of the correction J_0 negligible. For the lowest point, the ratio of observed to calculated J_1 is 2.5; for the highest, 1.2. In consideration of the many obvious analytical departures from reality, the fit along the upper, most accurate part of the curve is considered a rather good agreement.

It should particularly be noticed that the expression for J_1 depends greatly upon the ratio d_2/d_1. In actuality, the virtual cathode in close-

[12] It is certain that the gas density was not constant, but increased with current. Precise information on this point was not essential and was not sought. However, by putting the analysis into reverse, one can estimate from (4) that the gas pressure ($O°C$) rose from about 0.2 to about 0.7 micron.

spaced tubes may be so positioned that the proper d_1 differs significantly from the cathode-grid spacing. A correction of this sort would further improve the agreement. While provisions for this, as well as for other refinements, could be arranged, they appear hardly justified at present.

CONCLUSIONS

We have attempted to provide a sound quantitative understanding of the basic fluctuation phenomena in conventional vacuum tubes con-

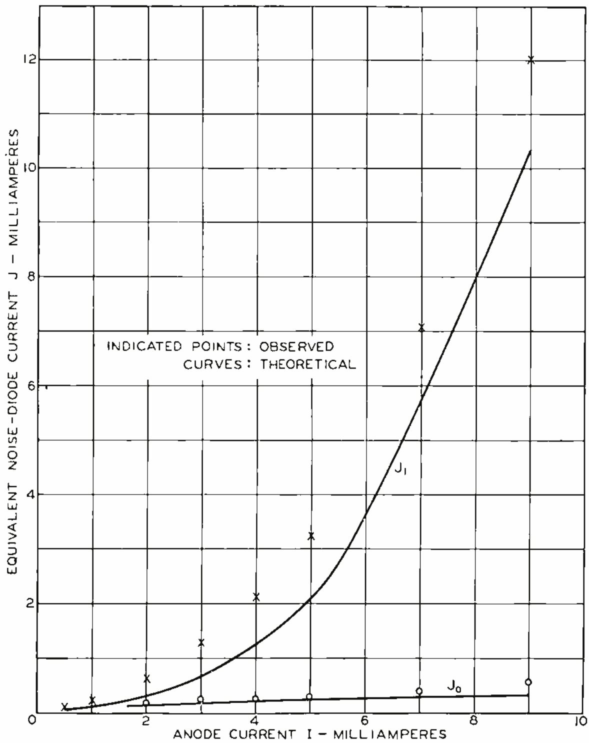

Fig. 8—Fluctuations (J_1) in the anode current of the triode of Figure 7 containing evolved gas, corrected by subtraction of the residual fluctuations (J_0) measured after flashing a getter. Curves calculated from theory.

taining small amounts of gas. We conclude that it is proper, at least for coated-cathode types, to regard CO as a representative gas for this purpose, and that it is possible hereby to predict the noise performance of such tubes in terms of structural constants, normal electrical characteristics, and the control-grid gas current. Both calculation and measurement suggest that the ionization component of noise may generally be ignored when the grid gas-current is less than a few hundredths of a microampere.

(To be continued)

OUR CONTRIBUTORS

GEORGE H. BROWN (A'30) was born October 14, 1908, at North Milwaukee, Wisconsin. He received the B.S. degree from the University of Wisconsin in 1930, the M.S. degree in 1931, and the Ph.D. degree in 1933. From 1930 to 1933 Dr. Brown was a Research Fellow in the Electrical Engineering Department of the University of Wisconsin, and since 1933 he has been in the Research Division of the RCA Victor Division of the RCA Manufacturing Company. He is a member of Sigma Xi and the American Association for the Advancement of Science.

MURRAY G. CROSBY joined the branch of the Radio Corporation of America which is now R.C.A. Communications, Inc., in 1925. He was engaged in the operating and design departments of that branch until 1926, when he took a leave of absence and returned to the University of Wisconsin for one semester and received his degree of B.S. in Electrical Engineering. Since that time he has remained in the research and development division of R.C.A. Communications, Inc. Mr. Crosby is a member of the Institute of Radio Engineers and a fellow of the Radio Club of America.

O. B. HANSON, Chief Engineer of the National Broadcasting Company, has, since the inception of broadcasting, been a very large contributor to its technical development. Many of the achievements made in that field have been due largely to his knowledge and efforts. Mr. Hanson's radio career began in 1912 when he attended the school now known as RCA Institutes. Upon completing his course he went to sea as a radio operator and was later transferred to the testing department of the Marconi Company, where he became Chief Testing Engineer. In 1920 he took another turn at sea. When radio broadcasting came into being Mr. Hanson became associated with WAAM, a pioneer station in Newark. In 1922 he accepted a position as assistant to the Plant Engineer of WEAF, then owned and operated by the American Telephone and Telegraph Company. With the formation of the National Broadcasting Company in 1926, Mr. Hanson went with the new company and since that date has directed technical operations and engineering activities for NBC. He supervised the designing and construction of the NBC studios at 711 Fifth Avenue (vacated when NBC moved to Radio City), and the NBC Chicago studios. When the time came for creating the NBC studios at Radio City. Mr. Hanson was well equipped to carry out the ambitious plans. The results of his work have attracted world-wide interest both on the part of technicians and laymen.

RALPH H. HEACOCK was graduated in Civil Engineering from Swarthmore College in 1918. He served as a hydrophone officer in submarine detection during the World War. He was an aeronautical engineer at the Naval Aircraft Factory until 1922, when he joined the Engineering Department of the Victor Talking Machine Company, serving as Assistant Superintendent of Inspection until 1930, then transferring to the Engineering Department of RCA Manufacturing Company. Since then he has worked on automatic phonographs, word spotters for broadcast use, and talking picture equipment. Mr. Heacock is a member of Sigma Tau.

VERNON D. LANDON, following his graduation from Detroit, Jr. College, became a Westinghouse engineer and for six years was in charge of their radio frequency laboratories. In 1929 he went with the Radio Frequency Laboratory in Boonton, N. J., as Assistant Chief Engineer, and in 1931 became Assistant Chief Engineer for Grigsby-Grunow. Since 1933 he has been with the Engineering Department of RCA Manufacturing Company.

O. S. MEIXELL is a graduate of Yale with a B.A. degree in 1927, and post-graduate study in Electrical Engineering at the Massachusetts Institute of Technology. For about five years thereafter, he was in the Engineering Department of the Western Electric Company at Kearny, New Jersey. Since March, 1936, he has been instructing in circuit theory and electronics at RCA Institutes, where he is head of the Department of Electrical Engineering.

DWIGHT O. NORTH received his B.S. degree from Wesleyan University in 1930 and his Ph.D. degree from the California Institute of Technology in 1933. Since 1934, Dr. North has been with the Research and Engineering Department of the RCA Manufacturing Company at Harrison, N. J., engaged principally in research studies of tube and circuit noise. He is a member of The Institute of Radio Engineers and a member of the American Physical Society.

BROWDER J. THOMPSON received a B.S. degree in electrical engineering from the University of Washington in 1925. From 1926 to 1931 he was engaged in vacuum tube research and development for the General Electric Company at Schenectady. From 1931 to the present he has been in charge of the Research Division, Research and Engineering Department, RCA Manufacturing Company, Inc., Harrison, N. J., engaged in television, ultra-short-wave, and other tube research problems. In 1936 Mr. Thompson was awarded the Morris Liebmann Memorial Prize by the Institute of Radio Engineers for important contributions to the ultra-short-wave field of radio. He is a Fellow of the Institute of Radio Engineers and a member of the American Physical Society.

ARTHUR WILLIAM VANCE was born on July 15, 1904. He was graduated from Kansas State College in 1928 with the degree of B.S. in electrical engineering. From 1929 to 1930 he was a television and facsimile research engineer at the Westinghouse Electrical and Manufacturing Company. Since 1930 he has been employed as engineer in charge of circuit research in the Electronics Research Division of the RCA Manufacturing Company.

A. KYLE WING received his B.S. degree in electrical engineering from Sheffield Scientific School, Yale University, in 1930, and his M.S. degree from Massachusetts Institute of Technology in 1931. During the summers of 1929 and 1930 he was employed in the vacuum tube department of the General Electric Company. Following work on broadcast receiver development with the Kolster Radio Company, he spent three years on transmitter tube development with the Federal Telegraph Company. Since 1934, Mr. Wing has been engaged in development work on transmitting tubes in the Research and Engineering Department of the RCA Manufacturing Company at Harrison, N. J. Mr. Wing is an associate member of the Institute of Radio Engineers.

J. E. YOUNG, was born in West Chester, Pa., May 18, 1906. He attended primary schools at Chicago, Ill., and West Chester, Pa., and attended high school at West Chester, Pa. He graduated from Drexel Institutes, 1928, with a degree of B.S. in E.E. and worked with General Electric Company, Radio Department, Schenectady, from 1928 to 1932, in high-power transmitter development and installation. He has been in RCA Manufacturing Company since 1932 to date, doing transmitter development, design, and installation work. He has been in charge of low-power transmitter product design from 1937 to date.

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