A STUDY OF A SIMPLIFIED PULSE POSITION DEMODULATOR

By

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PREFACE

In the past decade advances in the utilization of radio frequencies have resulted in complex problems. So many uses have been found for radio that the radio spectrum has become increasingly crowded with transmitters and corresponding interference problems despite careful allocation of frequencies. The public demand for frequency allocations has prompted many investigations as to methods of better utilization of the usable electromagnetic spectrum. These studies have resulted in complex transmission systems. Greatest utilization of the spectrum has resulted from the development of pulse sampling techniques used in many multiplex systems.

The principle research and development work in this field has been done by Federal Telecommunications Laboratory, Inc., Nutley, New Jersey. Articles published by employees of this firm are the principle source of information in this field. In addition to using the commercially available equipment in their work they have developed special electron ray commutator tubes for mixing and separating the pulse train. In the systems on which literature is available, the pulse train repetition rate has been relatively low so that it is impossible to transmit an audio frequency of greater than 12 kilocycles and even this frequency is demodulated with very poor fidelity. The present study was initiated with the idea of improving the quality of transmission of the higher audio frequencies with particular reference to cable transmission using no radio frequency carrier although the method studied could be used in connection with an ultra-high frequency transmitter and broad-band receiver.

At the suggestion of Prof. A. L. Betts, a study of pulse modulation was begun by C. W. Merle and John A. B. Bower in February, 1949. A 100 kc. pulse sampling frequency was selected since this would provide an upper limit of 50 kc. for the modulating frequency and thus should allow excellent quality at frequencies up to 20 kc. Early designs of a pulse modulation system were made by these men and Thomas King. In February, 1950 R. D. Kelly and I. Edward Lynch undertook further studies of the system. In the early stages of demodulator construction it was decided that redesign of of the system could provide improved circuit operation. Kelly and Lynch concluded their study in January, 1951 at which time both transmitter and receiver had been constructed and tested.

In June, 1950 the author commenced a study of the system with the intention of determining the degree of crosstalk between adjacent channels. In the early stage of testing the operation of the receiver indicated that some modifications would be necessary in order to provide adequate data. With the concurrence of Professor Betts it was decided to design and construct a completely new demodulator. In the present demodulator design the author has attempted to simplify the circuitry as much as possible while providing demodulation of as high quality as practicable.

The writer wishes to express his gratitude to Professors A. L. Betts and H. T. Fristoe of the Department of Electrical Engineering, Oklahoma Institute of Technology for their cooperation, assistance, and guidance in this study. He is also greatly indebted to R. F. Buck, of the Research Foundation, Oklahoma A. & M. College, whose advice in pulse techniques has been of great aid.

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PART I

INTRODUCTION

In recent years the radio art has pushed the usable electromagnetic spectrum to ever higher frequencies so that radio waves now vary in wavelength from 3,000 meters to 0.5 centimeter. In the higher frequency region effective use of the wide bandwidth available allows pulses of extremely fast rise and decay time to be transmitted. It can be shown that the required bandwidth is equal to the reciprocal of twice the pulse rise or decay time, whichever is smaller.[1] According to this rule if we are interested in a pulse rise time of 0.1 microseconds, a bandwidth of 5 megacycles would be necessary. Bandwidths of this order are most easily achieved by using carrier frequencies of 100 megacycles or more. This bandwidth is independent of the number of pulses transmitted in a given time interval. Therefore, if intelligence modulates some pulse characteristic, and if these pulses can be separated within the receiver and the intelligence extracted it is possible to use a single radio frequency channel to transmit a multiple pulse train which can carry several different modulating signals.

Pulse modulation methods may be classified into four general groups as follows:

a. Pulse-time modulation (PTM) in which the instantaneous value of the sampled modulating wave causes a proportional

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[1] C. W. Earp, "Relationship between Rate of Transmission of Information, Etc.", Electrical Communication, XXV (June, 1948) p. 184.

variation in the occurrence time of some pulse characteristic.

b. Pulse-amplitude modulation (PAM) in which the instantaneous value of the sampled wave modulates the peak amplitude of the pulse.

c. Pulse-code modulation (PCM) in which the instantaneous samples of the modulating wave are quantized into a discreet number of step functions in amplitude or into a presence or absence of a pulse in time.

d. Composite methods combining above mentioned methods.[2 3 4]

Pulse-time modulation may be divided into three general types as follows:

a. Pulse-position modulation (PPM) in which the instantaneous value of the sample modulates the position in time with respect to a master pulse.

b. Pulse-duration modulation (PDM) in which the pulse duration is proportional to the instantaneous value of the sample. This also is sometimes called pulse-width modulation (PWM) or pulse-length modulation.[5]

c. Pulse-frequency modulation (PFM) in which the modulating wave modulates the repetition frequency of a pulse train.[6]

It should be mentioned that no standard definition has yet

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[2] Reference Data for Radio Engineers, Third Edition, p. 285.

[3] E. M. Deloraine, "Pulse Modulation", Proceedings of the IRE, XXXVII (June, 1949), pp. 703-704.

[4] "IRE Standards on Pulses", Proceedings of the IRE, XXXIX, (June, 1951) pp. 624-626.

[5] Ibid., p. 625.

[6] Reference Data for Radio Engineers, loc. cit.

been accepted for pulse-frequency modulation. This term has been used to refer to a frequency modulated radio frequency carrier which is transmitted only in pulses.[7] It is also defined as the frequency-modulation of a carrier wave consisting of a series of direct current pulses.[8]

Early studies in pulse modulation centered around the use of pulse-amplitude modulation to increase the transmitter power efficiency by reducing the duty cycle. Later the emphasis changed to utilize pulses to decrease the noise level. Since pulse-amplitude modulation offers no signal to noise improvement and also requires amplitude linearity in the receiver, other methods of pulse modulation were investigated. Because pulse-code modulation introduces a distortion of quantization which resembles noise the principal interest has been in pulse-time modulation, which provides a great improvement in signal to noise ratio.[9]

Pulse frequency modulation (under the definition herein used) is somewhat more difficult to use in multiplex, and pulse-duration modulation results in a variable duty cycle which imposes more stringent requirements on radio frequency equipment and power supply regulation. Pulse-position modulation uses pulses of constant duration and amplitude which are modulated

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[7] H. Goldberg and C. C. Bath, "Multiplex Employing Pulse-Time and Pulse-Frequency Modulation", Proceedings of the IRE, XXVII (January, 1949), pp. 22-28

[8] Reference Data for Radio Engineers, loc. cit.

[9] S. Moskowitz and D. D. Grieg, "Noise Suppression Characteristics of Pulse-Time Modulation", Electrical Communication, XXVI (March, 1949) pp. 46-51; Proceedings of the IRE, XXXVI (April, 1948), pp. 446-450.

in time with respect to a marker pulse which is fixed. A major advantage of this system is that any link circuits need not have linear amplitude characteristics.[10] If the pulse has an instantaneous rise or decay time amplitude noise variations which appeared only on the peaks would be removed by limiting or clipping action and, since the detector is affected only by time variations, would not appear. Thermal noise or shot effect which causes random variations in triggering time within the circuitry would produce noise in the detected signal. Since these effects are generally of low order and may be reduced by proper circuit design the overall noise level of a pulse-position modulation system is low.

A major problem in PPM multiplex is the reduction of crosstalk between channels. This may be reduced by increasing the channel pulse spacing, but this reduces the number of channels which can be placed between marker pulses. Circuit design which results in a minimum rise and decay time for marker and channel pulses will also reduce crosstalk.[11]

In the present study pulse position modulation was selected because of its relative simplicity and the high order of improvement of signal to noise ratio inherent in this system. The demodulation of a pulse position modulated wave consists first, of the separation of the marker pulse from channel pulses; second, of the separation of the channel pulses; and third, of the

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[10] E. M. Deloraine, op. cit., p. 703.

[11] S. Moskowitz, L. Diven, and L. Feit, "Cross-Talk Considerations in Time-Division Multiplex", Proceedings of the IRE, XXXVIII (November, 1950), pp. 1330-1334; Electrical Communication, XXVIII (September, 1951), pp. 209-213.

translation of a time variation into an amplitude variation in order to obtain the original modulating frequency. The problem of channel separation may be solved in several ways after the marker pulse has been selected. A delay pulse may be formed, usually by a multivibrator or delay line, which is fixed in relation to the marker pulse and determines the leading edge of a time selection gate pulse. A delay multivibrator is the technique utilized in the present system. This system has the advantage of using readily available components, but when a dozen or more channels are multiplexed an extremely bulky system results. The Cyclophon commutator and decommutator tubes were developed to overcome this difficulty by Federal Telecommunications Laboratory.[12]

The link selected for this test was a 50 ohm coaxial cable (RG-8/U). The selection of this link in no way precludes the use of this system as a pulse modulator for a ultra high frequency radio link, but the choice was made in order to simplify testing of the circuits involved. Any length of cable may be used so long as video amplifiers can boost the signal to the proper level for proper triggering of the demodulator.

The 100 kc. repetition rate for the marker pulse is unusually high. The highest repetition frequency utilized in system described in published literature prior to the present series of theses was 24 kc. with most systems utilizing sampling frequencies of around 8 kc.[13] It may be shown that the sampling frequency

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[12] D. D. Grieg and A. M. Levine, "Pulse-Time Modulated Multiplex Radio Relay System-Terminal Equipment", Electrical Communication, XXIII (June, 1946), p. 159.

[13] Loc. cit.

must be at least twice the highest frequency it is desired to transmit.[14] Intuitive reasoning also indicates this when we realize that it is impossible to synthesize a single cycle of a sine wave with less than two samples. This indicates that the maximum frequency usually transmitted in pulse modulation is 4 kc. The present system should show a response up to 50 kc. if the audio system was adequate. This means that at the high frequency cut-off of the human ear (around 15 to 18 kc.) this system should provide excellent response.

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[14]C. E. Shannon, "Communication in the Presence of Noise", <u>Proceedings of the IRE</u>, XXXVII (January, 1949), p. 10.

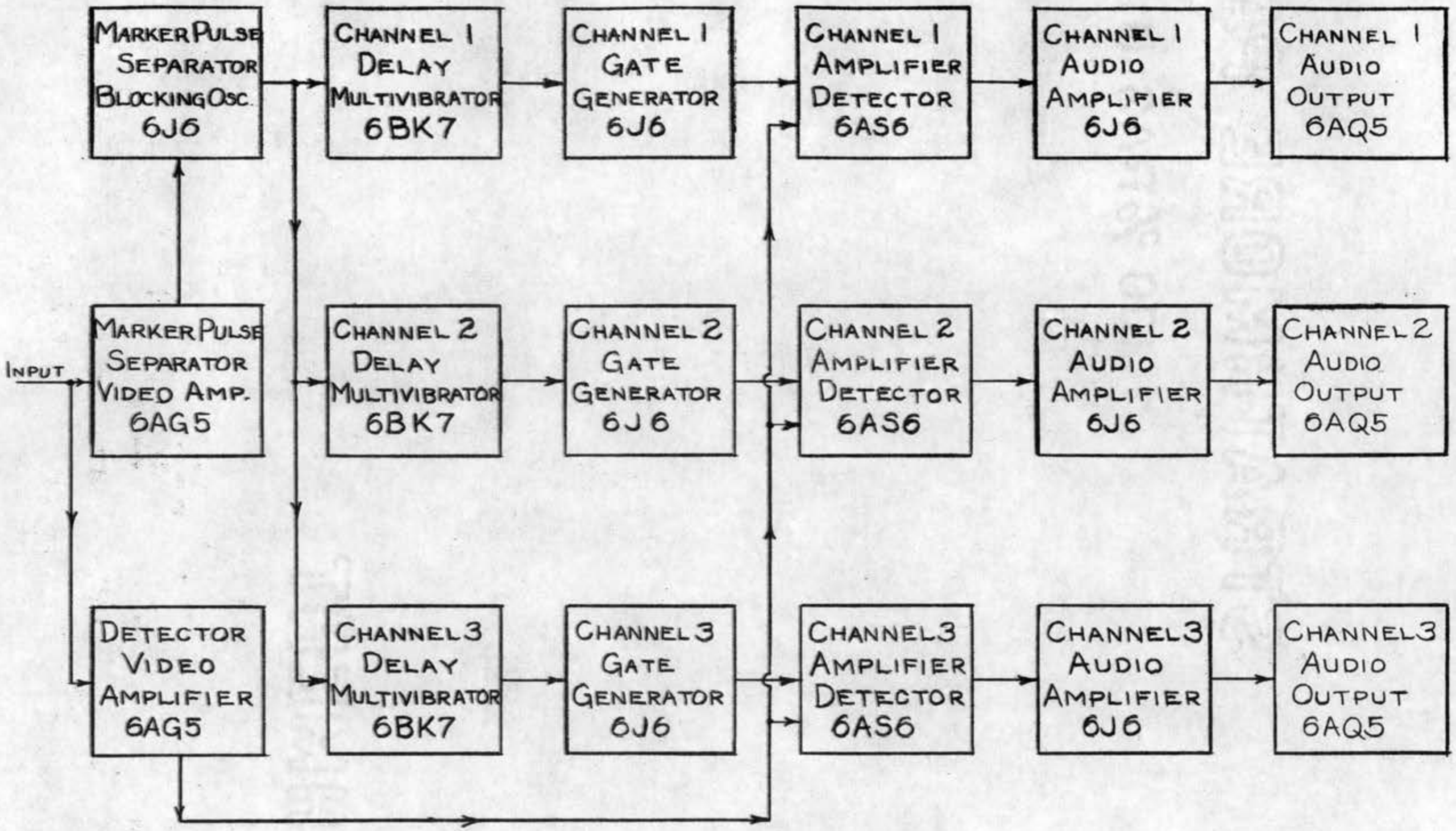


Figure 1
Demodulator Block Diagram

PART II

METHOD OF DEMODULATION

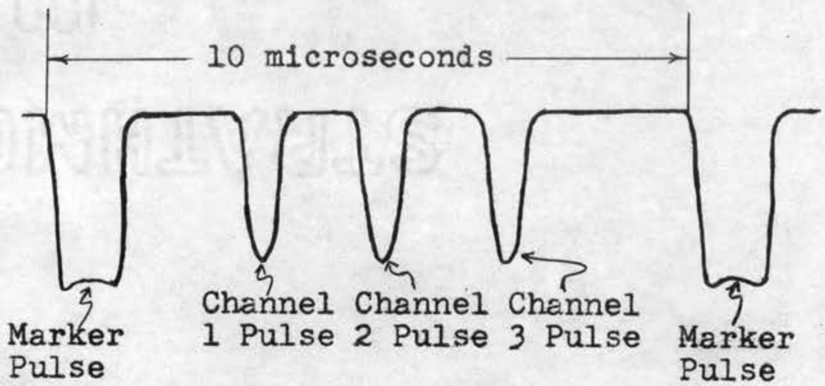


Figure 2
Input Pulse Train

The general problem involved in this study is the recovery of the modulating waveform from the pulse train. The available pulse train is shown in Figure 2. It consists of a marker, master, or synchronizing pulse of 5.25 volts amplitude and 1.4 microseconds duration and three channel pulses of slightly lower amplitude and approximately 0.8 microseconds duration. Figure 2 shows the pulse train with no modulation applied. When modulation is applied to a channel pulse the time between the marker pulse and each channel pulse is varied in proportion to the modulating voltage applied.

The demodulation of this waveform consists of the following steps:

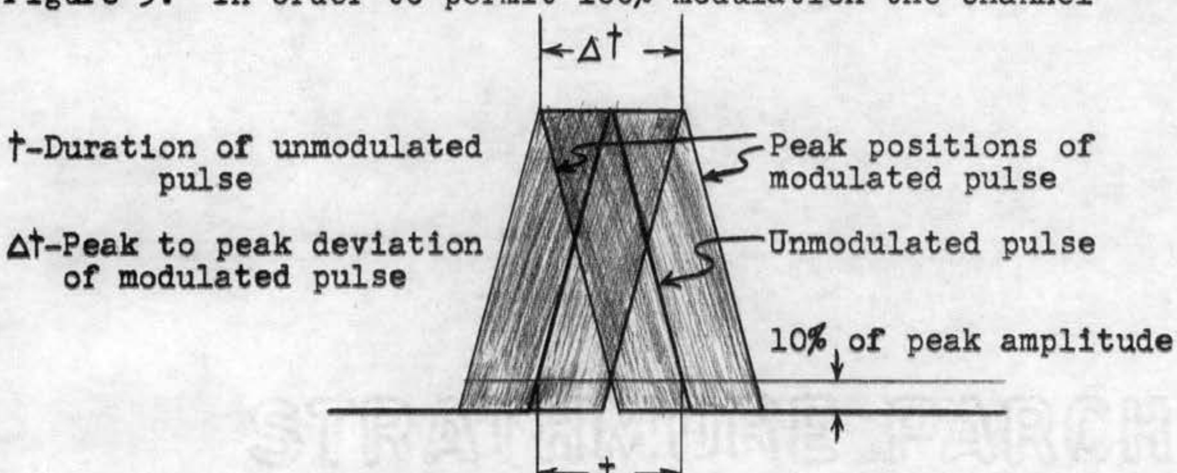
1. The separation of the wave train into the three channels.

2. Individually converting each channel pulse from position modulation to amplitude modulation.

3. The individual detection and amplification of each

modulated wave train to recover the modulating waveform.

The marker pulse must be different from the channel pulses either in amplitude, duration, or polarity so that it can be separated from the channel pulse in order to provide a fixed time reference in the demodulator. From the marker pulse three different fixed delay times are established and at the end of these delays a discreet time gate or separator pulse is generated. These separator pulses are set so that one separator pulse occurs simultaneously with one channel pulse and allows only this channel pulse to reach succeeding circuits. This separator pulse must be of sufficient time duration to allow maximum modulated deviation of the channel pulse. The definition for 100% position modulation is a peak to peak deviation equal to the pulse width. This definition is illustrated by a triangular pulse shown in Figure 3. In order to permit 100% modulation the channel



$$\text{Percent Modulation} = \frac{\Delta t}{t}\,100$$

Figure 3
Position Modulation Percentage

separator pulse must be equal in width to the channel pulse. After the channel pulses are separated the position modulation must in some way be changed to amplitude modulation. This is usually done by combining the position modulated pulse with a sawtooth voltage waveform. It should be noted that this technique will permit amplitude variations to affect output, but since the amplitude may be clipped in stages previous to combining with the sawtooth voltage this characteristic is not a disadvantage. Various methods of accomplishing the change from position modulation to amplitude modulation may be utilized, but some form of sawtooth voltage is probably the simplest and most reliable. The amplitude modulated pulse train is then detected by some conventional form of detector and after the pulse carrier has been removed by a filter the recovered modulation waveform is amplified through conventional audio amplifiers.

The demodulation system evolved for this study is shown in block diagram form in Figure 1. The first step necessary to demodulate the pulse train is the separation of the marker pulse which synchronizes the succeeding circuits. In order to achieve separation the marker pulse separator amplifier first inverts and amplifies the pulse train. The resulting positive pulses are then integrated so that the marker pulse, which has a longer time duration than the channel pulses, will be of greater relative amplitude. The integrated pulses are used to trigger the marker pulse separator blocking oscillator which generates a pulse of large amplitude when the integrated marker pulse rises to proper triggering voltage. This blocking oscillator pulse triggers the channel delay multivibrators. Each of these multivibrators generates a square wave every time it receives a positive triggering pulse. The duration of the square wave is adjustable in each multivibrator and is set so that the trailing edge of channel 1 delay multivibrator occurs just before channel 1 pulse, the trailing edge of channel 2 delay multivibrator occurs just before channel 2 pulse, and the trailing edge of channel 3 delay multivibrator occurs just before channel 3 pulse. The trailing edge of these multivibrators triggers a gate generator for each channel. This gate generator supplies a sawtooth voltage to the suppressor grid of the amplifier detector causing the tube to act as a variable amplifier with gain proportional to suppressor voltage. The inverted input pulse train is applied to the control grid of the gated amplifier, but output at the plate will occur only when the suppressor grid allows the tube to conduct. Since the channel pulse is position modulated along the sawtooth suppressor waveform an amplitude modulated pulse train of 100 kc. repetition rate appears at the plate. This amplitude modulated pulse train is detected by a diode detector and fed to a cathode follower audio amplifier where a rejection filter eliminates the 100 kc. pulse carrier. The resultant audio signal is fed to an audio power output stage.

PART III

ANALYSIS OF CIRCUITS

The operation and analysis of the individual circuits necessary to achieve demodulation will now be considered. In the following discussion the d. c. voltages were obtained with a vacuum tube voltmeter (Hewlett-Packard Mod. 410-A). The waveforms were traced from a Du Mont Type 256-D synchroscope. Time and amplitude measurements of these waveforms were obtained from the same synchroscope. Most waveforms were obtained using a probe with a series 3.9 megohm resistor in parallel with a 10 micromicrofarad capacitor. This probe minimizes the effect of cable capacity and attenuates the waveform by a factor of 15.

The circuits will be analyzed in the following order:

A. Marker Pulse Separator Video Amplifier

B. Detector Video Amplifier

C. Marker Pulse Separator Blocking Oscillator

D. Delay Multivibrator

E. Gate Generator

F. Amplifier Detector

G. Audio Amplifier

H. Audio Output

In the following figures capacitors will have values stated in micromicrofarads unless otherwise designated. Resistors will have values stated in ohms unless otherwise designated by K. for kilohms or M. for megohms. Waveforms are drawn in proper time scale and the potential scale is designated for each waveform.

A. MARKER PULSE SEPARATOR VIDEO AMPLIFIER

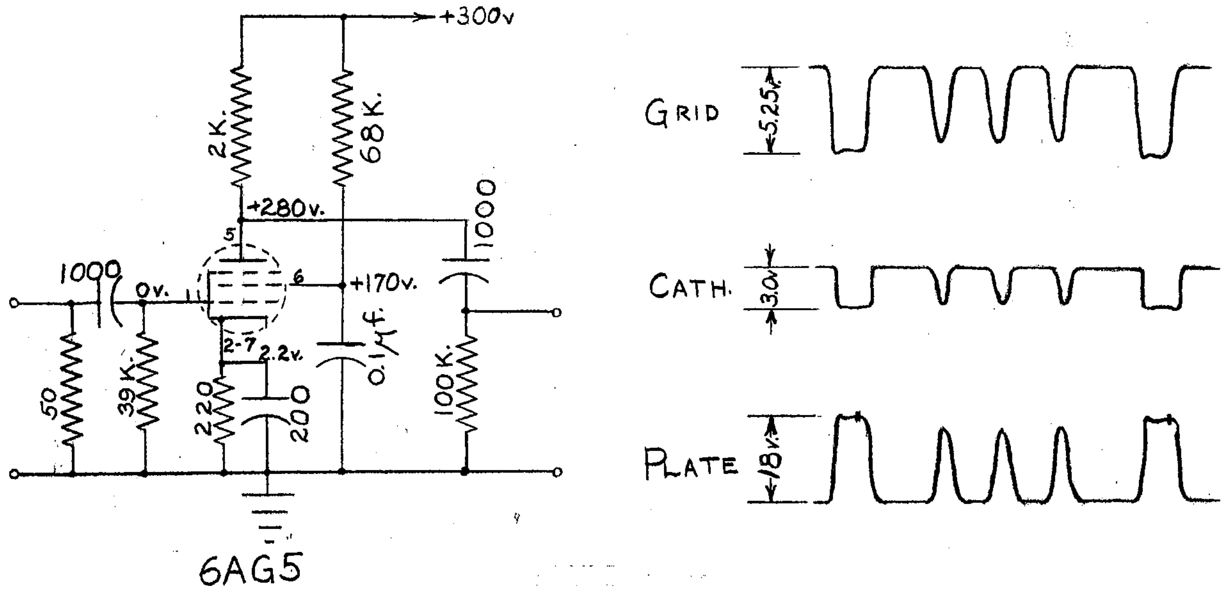


Figure 4
Schematic, Voltages, and Waveforms for
Marker Pulse Separator Video Amplifier

The marker pulse separator amplifier shown in Figure 4 is a video amplifier of conventional design which serves to invert, amplify, and isolate the input pulse train. A small plate load resistor is used with a pentode so that integration of the plate waveform will not occur. With the operating potentials used the transconductance of the 6AG5 is around 5000 micromhos so that with a plate load of 2 K. a tube gain of 10 is possible. The actual tube gain of 8 is lower than the theoretical because the grid signal is so large that the tube is driven very close to cut-off at the negative peak and full transconductance is realized only under conditions of a small signal. The input waveform of 5.25 volts is reduced by cathode degeneration to an effective grid signal of 2.25 volts. With a tube gain of 8 the plate waveform has a peak amplitude of 18 volts. The stage gain is therefore only 3.43. The cathode bypass capacitor is a critical component chosen to give proper signal amplification by selective degeneration of the high frequency components in the fast waveform. If this capacitor is too large the high frequency components will not be degenerated and overshoot of the leading and trailing edges will occur. If the capacitor is too small the lower frequency components will have insufficient gain and sloping of the leading and trailing edges will result. The output voltage of this stage is used to feed the marker pulse separator blocking oscillator.

B. DETECTOR VIDEO AMPLIFIER

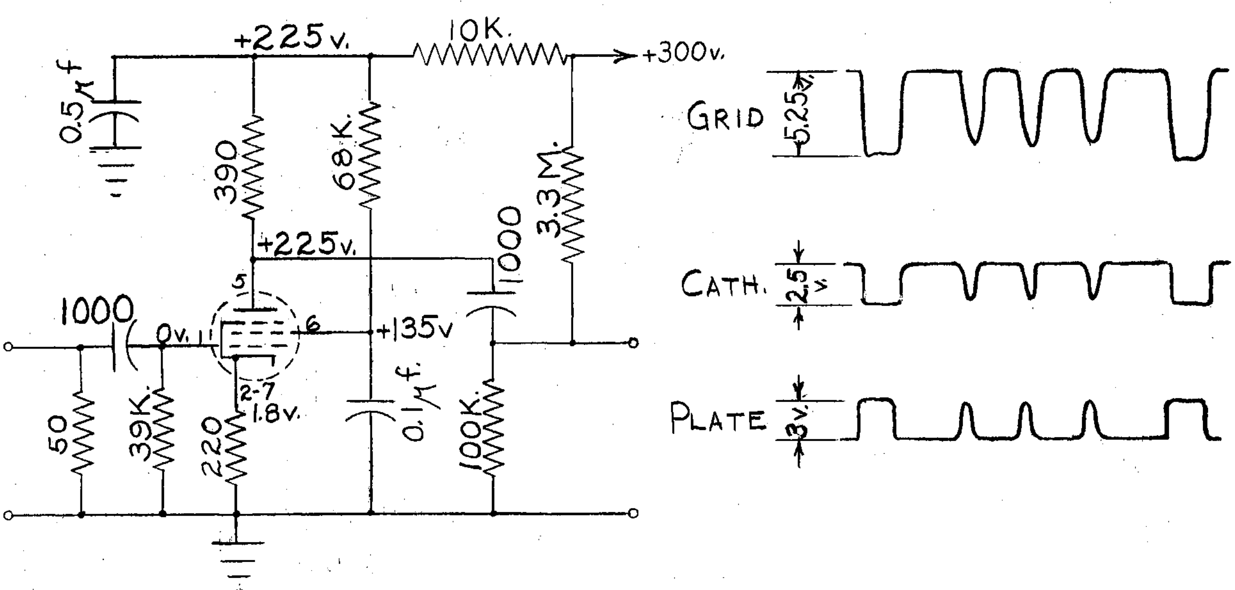
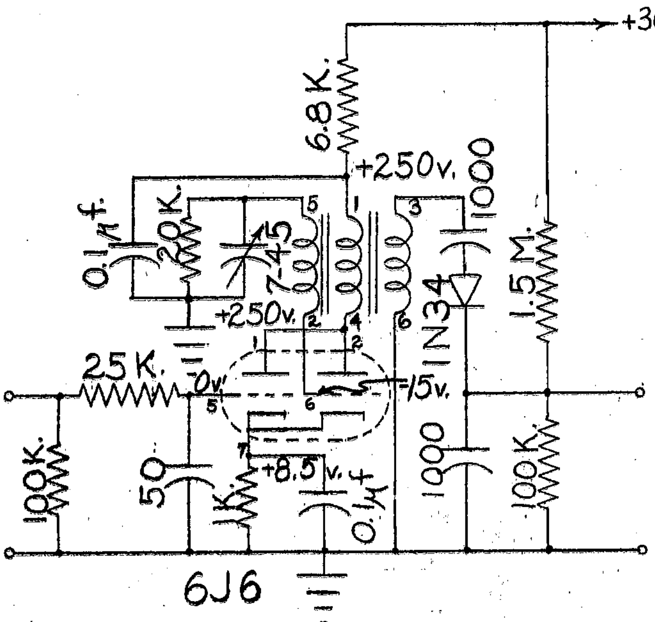


Figure 5
Schematic, Voltages, and Waveforms
For Detector Video Amplifier

The detector video amplifier shown in Figure 5 is a video amplifier similar to the circuit previously described. This circuit is used principally as an isolating and inverting stage. A 390 ohm plate load resistor is used since gain is not desired. With the pentode transconductance of around 5000 micromhos a theoretical tube gain of 1.95 is possible. The grid signal is of such large amplitude that the tube is driven near cut-off and full theoretical gain is not realized and actual tube gain is only 1.1. Cathode degeneration reduces the grid signal to 2.75 volts so that the output voltage is 3 volts or the net stage gain is only 0.57. Because of the low gain of the stage selective degeneration in the cathode is not necessary. Because of the low amplitude of the plate signal additional filtering of the B supply is necessary and is accomplished by a 10 K. resistor and 0.5 microfarad condenser. The output waveform of this amplifier is applied to the control grids of the three amplifier detector stages. The bleeder resistors for these grids (100 K. and 3.3 M.) are included in the circuit diagram.

C. MARKER PULSE SEPARATOR BLOCKING OSCILLATOR



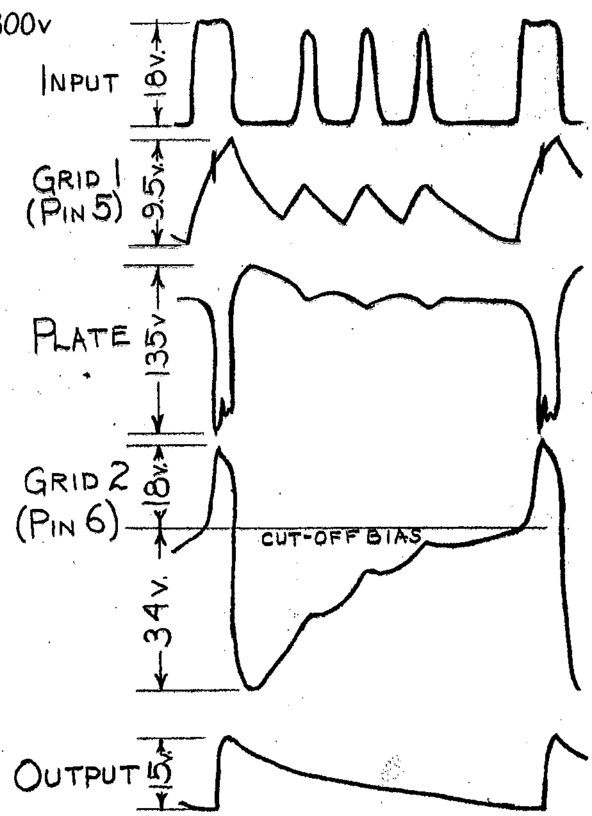


Figure 6
Schematic, Voltages, and
Waveforms for Marker Pulse
Separator Blocking Oscillator

The marker pulse separator blocking oscillator shown in Figure 6 separates the marker pulse from the channel pulses and provides a synchronizing pulse for succeeding stages. The amplified and inverted pulse train is applied to an integrating circuit having an effective time constant of approximately 1.85 microseconds. This allows the marker pulse to rise to 53% of its maximum value as the marker pulse is 1.4 microseconds wide. Since the channel pulses are only 0.8 microseconds wide they would only rise to 35% of maximum value. The resulting integrated waveform is applied to one grid of a dual-triode having common cathodes and plates. When the grid voltage reaches a point where conduction occurs the plate voltage drops causing a more positive voltage to be applied to the second grid which causes the second half of the tube to draw more current and further lowering the plate voltage. This continues until the tube is drawing saturation current and the incremental decrease of current causes a negative voltage to be applied to the second grid further decreasing current and lowering grid potential until it is at a potential far more negative than cut-off and both halves of the tube are non-conducting. The recovery of the second grid is controlled by a parallel RC circuit having an adjustable time constant which is adjusted so that the free running period of the blocking oscillator is slightly over 10 microseconds. Then the integrated marker pulse applied to the first grid serves to synchronize the blocking oscillator. The plate circuit is decoupled from the B supply by a 6.8 K. resistor and a 0.1 microfarad capacitor to prevent the high amplitude voltage pulses involved in this stage from causing interference in other circuits. The output waveform is taken from a third winding of the pulse transformer to provide d.c. isolation and to obtain a positive pulse for triggering the three channel delay multivibrators. The common grid circuit of these multivibrators is included in the partial schematic.

D. DELAY MULTIVIBRATOR

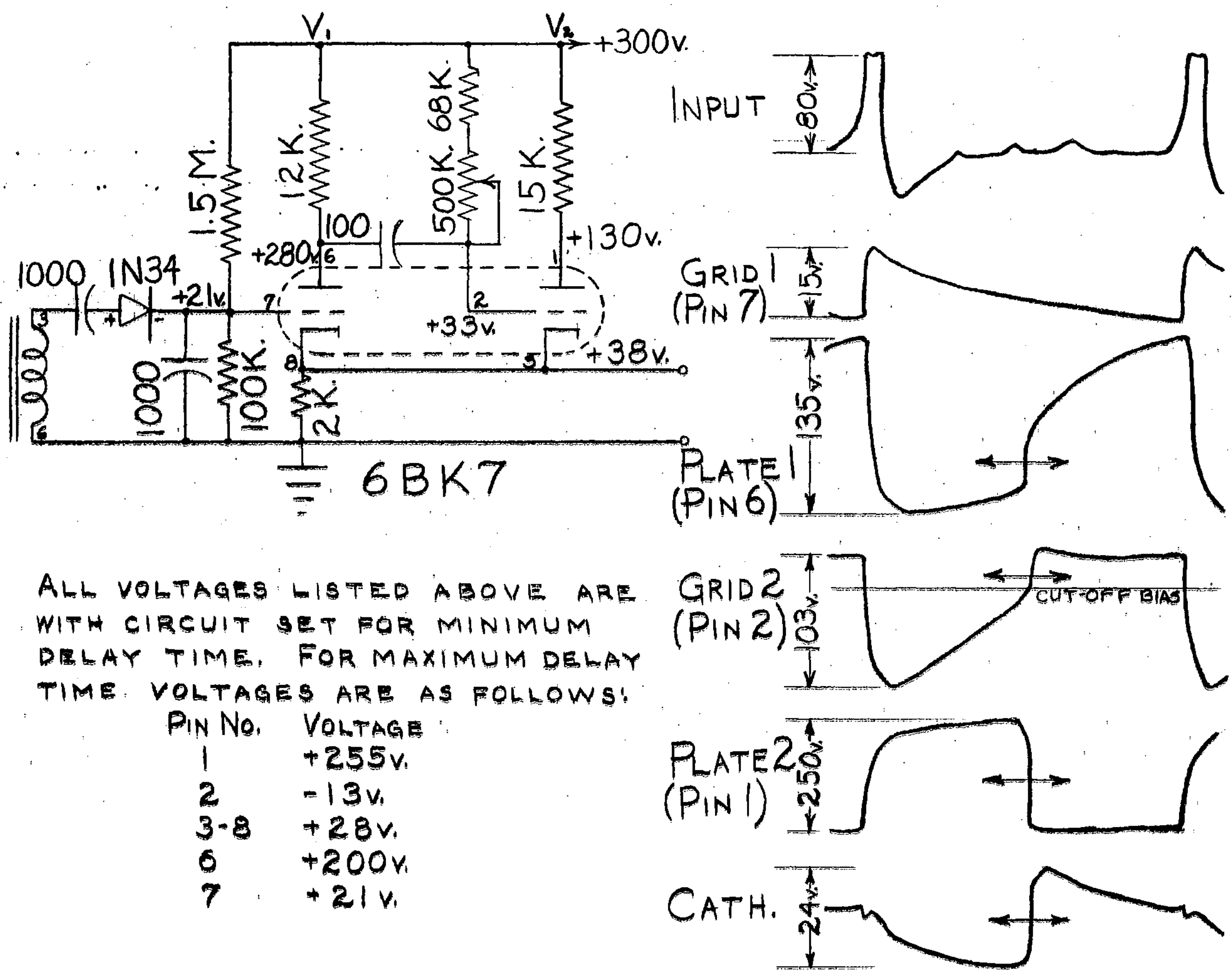


Figure 7
Schematic, Voltages, and Waveforms for Delay Multivibrator

The channel delay multivibrator shown in Figure 7 is used to supply a time delay reference after the beginning of the marker pulse. This is a monostable multivibrator in which a trigger pulse is supplied to the grid by the marker pulse separator blocking oscillator. A monostable multivibrator is described as having one stable state to which it returns after it has been set into operation by a triggering pulse. Thus it supplies one output waveform for every triggering waveform it receives.[1]

The blocking oscillator waveform is applied to the anode of a 1N34 germanium diode through a coupling capacitor for d.c. isolation. This diode prevents negative pulses from being applied to the grid which might cause the multivibrator to shut off before the end of its natural period. A bleeder network is used to fix the d.c. grid potential for proper triggering and the 1000 micromicrofarad capacitor from grid to ground slows the recovery of the grid to prevent spurious triggering of the multivibrator.

In the stable state V_2 is conducting with the grid held slightly positive with respect to the cathode by the return circuit to the B supply. This sets the cathode at a potential of around 30 volts which holds V_1 at cut-off.

When the triggering pulse is applied to the first grid V_1 begins to conduct and the plate voltage drops. This drop in plate voltage is coupled to the second grid, and due to the gain of V_1 causes this grid to reach a potential far beyond cut-off. The circuit is now in an unstable state with the two triodes reversed in action. V_1 is conducting heavily and V_2 is cut off. In this state there is no waveform coupled from the plate of V_1

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[1]B. Chance, V. Hughes, E. F. MacNichol, D. Sayre, and F. C. Williams. <u>Waveforms</u>, New York McGraw-Hill Co. 1949 p. 162.

to the second grid so it is allowed to recover toward the potential of the B supply. This recovery is controlled by the RC circuit of 100 micromicrofarads and the resistors from the grid of V_2 to the B supply. When the second grid reaches a potential where V_2 begins to conduct the cathode potential rises, V_1 is once again cut-off, and the circuit has returned to its stable state.

The channel delay multivibrator is the most critical circuit involved in this demodulator because of the relatively wide range of period over which it must operate. It is very necessary in this circuit to use a tube with high transconductance. The 6BK7 which is used has the highest transconductance of any triode available (8500 micromhos). A high transconductance is necessary in order to supply sufficient signal on the grid of V_2 while using a low value of plate load resistor for V_1. The low plate load resistance is necessary to prevent integration of the waveform which would slow the transition process. It would be more desirable to use a larger capacitor between the plate of V_1 and the grid of V_2 in order to couple more energy to the grid. This is not feasible, however, because of the 100 kc. repetition rate involved. When the circuit returns to its stable state this capacitor tends to retain some of its charge from the unstable period. The recovery circuit for this capacitor is through the cathode resistor, the grid of V_2, and the plate load resistor of V_1. Since other circuit considerations dictate the choice of cathode and plate load resistors it is feasible only to reduce the capacitor in order to shorten the recovery time. If this capacitor has not recovered from the unstable state the multivibrator will tend either to act as a count-of-two multivibrator (two triggering pulses are required to cause circuit action) or to have successive periods of different length. When the period of the multivibrator reaches a point where the trailing edge is near to the next triggering pulse it is impossible to keep the multivibrator from having successive periods which are different. As a result of these considerations the size of this capacitor is a compromise between a large capacitor for transfer of sufficient energy for proper multivibrator action and a small capacitor for fast recovery.

The choice of grid potential for V_1 is also very critical. As increase of bias on this tube lowers the amplitude of the plate waveform and may prevent V_2 from triggering. A decrease of bias increases the amplitude of plate waveform and may set the grid of V_2 at such a low potential that recovery of this grid cannot be rapid and the minimum unstable period is increased. The potential of V_1 is adjusted so that the shortest possible period can be attained. If an adjustable potential for the grid were provided the multivibrator could be varied in this way rather than by controlling the time constant of grid recovery for V_2.[2]

A low plate load resistor is necessary for V_2 to prevent integration of the waveform. This transition period of the waveform is used to provide synchronizing pulses for the next circuit.

The cathode waveform is coupled to the gate generator because the leading edge is slow and the trailing edge is fast.

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[2]*Ibid*, p. 169.

The rise of the trailing edge then synchronizes the gate generator.

E. GATE GENERATOR

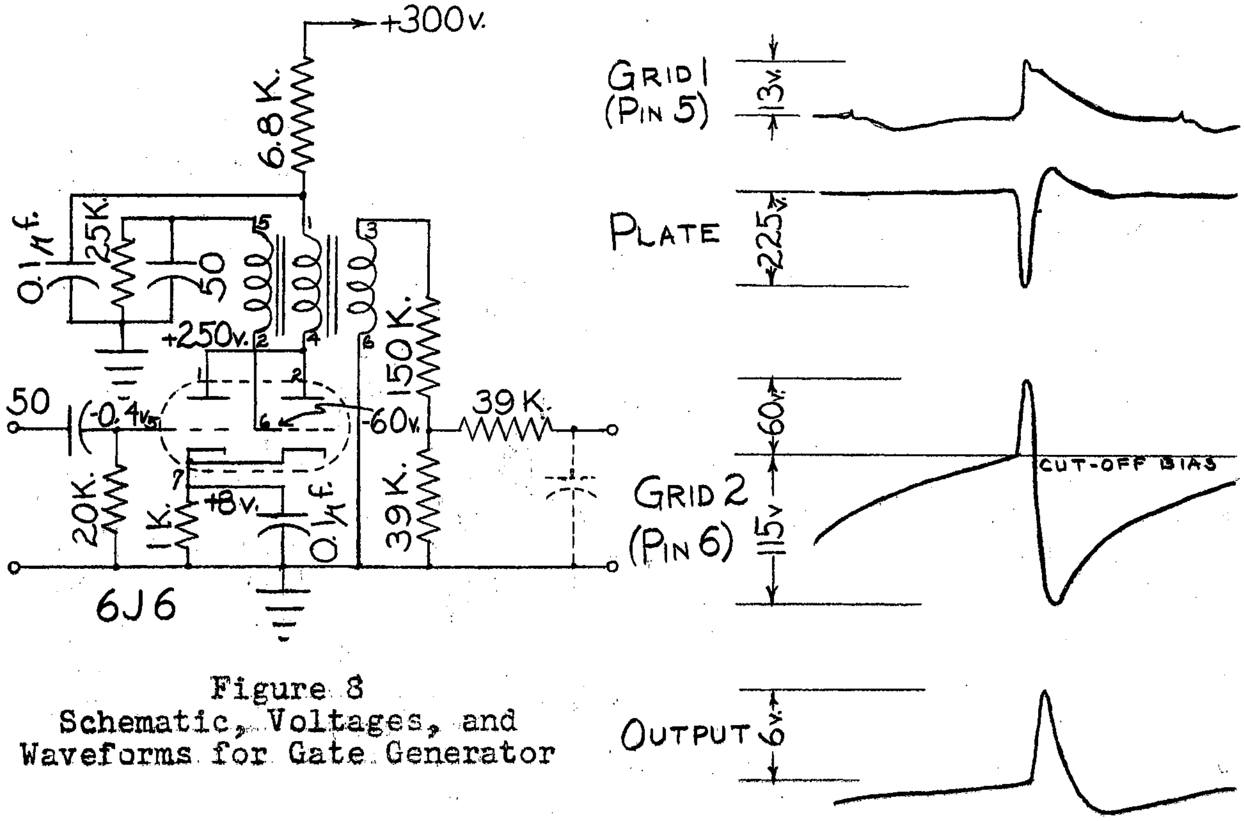

Figure 8
Schematic, Voltages, and Waveforms for Gate Generator

The gate generator shown in Figure 8 is a blocking oscillator circuit similar to the marker pulse separator blocking oscillator. The positive trailing edge of the cathode waveform of the channel delay multivibrator is applied to one grid of a 6J6 dual triode. This causes current to be drawn and the common plate voltage to drop. The connection polarity of the pulse transformer is such that this plate voltage drop causes a rise in the potential of the other grid causing the second half of the tube to draw more current and the plate voltage to drop further. This process continues until saturation current is drawn through the tube and the polarity of the grid signal reverses causing the plate potential to rise and the grid potential to decrease. This drives the grid to a potential far below cut-off bias. Controlled by the R C time constant of 25 K. and 50 micromicrofarads the grid potential slowly decays until it is slightly below cut-off when the next triggering waveform arrives. The output waveform from the third pulse transformer winding is resistance divided to 21% of full value by the 150 K. and 39 K. resistors. This reduced waveform is integrated by the 39 K. resistor and the input capacitance of the suppressor grid of the amplifier detector. The resultant suppressor grid waveform is a sawtooth voltage of fast rise and slow decay time.

F. AMPLIFIER DETECTOR

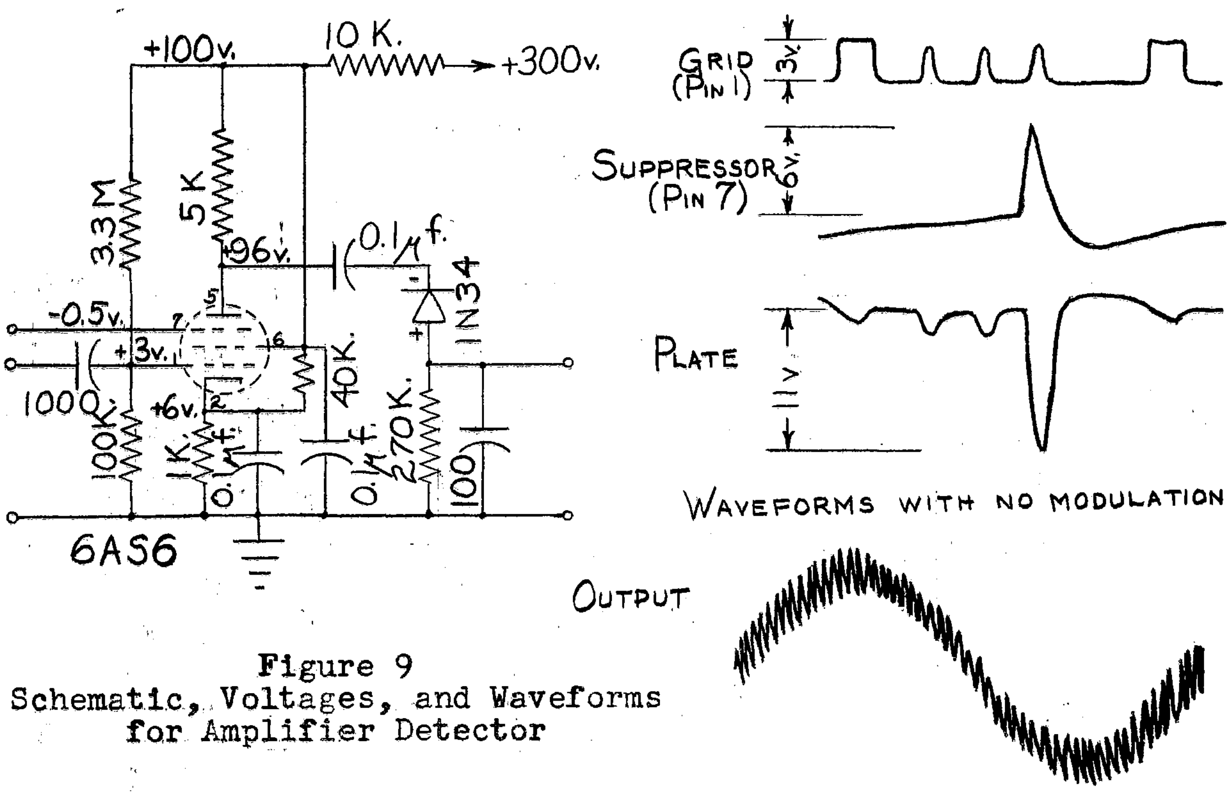

Figure 9
Schematic, Voltages, and Waveforms for Amplifier Detector

The amplifier detector shown in Figure 9 is usually called a time selector, gated amplifier, or coincidence circuit. The essential characteristic of this type of circuit is that a plate signal occurs only when signals are applied to two grids simultaneously. In this circuit the two signals are applied to the control grid and the suppressor grid. The signal applied to the suppressor grid is variously termed a "gate", "pedestal", or "selector pulse".[3]

The 6AS6 used for this circuit is a tube which is designed especially for this type of operation. Within a limited region the gain of the tube can be considered to be proportional to the instantaneous potential of the suppressor grid. Usually the selector pulse applied to the suppressor grid is a square wave so that the tube is cut off until the time of the selector pulse and during the selector pulse operates at constant gain. This preserves the selected control grid waveform at the plate. In this circuit a sawtooth waveform selector pulse is applied to the suppressor with the result that the tube gain varies during the period of the selector pulse. When a time modulated pulse train is applied to the control grid during the time of the selector pulse the plate waveform becomes an amplitude modulated train due to the variable gain of the tube. It should be noted that, due to insufficient suppressor bias the actual plate waveform contains all channel and marker pulses, however only the desired channel pulse which occurs during the time of the selector pulse will become amplitude modulated. The change from time modulation to amplitude modulation may occur either on the rise or decay of the sawtooth selector pulse. The rise time is

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[3]Britton Chance, "Time Demodulation", <u>Proceedings of the IRE</u>, XXXV (October, 1947) p. 1046.

0.3 microseconds and the decay time is 1 microsecond. This means that modulation deviations not greater than $\frac{0.3}{0.8}$ = 37.5% may be detected on the rise of the sawtooth waveform and deviations not greater than $\frac{1.0}{0.8}$ = 125% may be detected on the decay of the sawtooth. In normal operation the selector pulse should be set so that the channel pulse occurs during the slower decay time. This decay time however has a slight curvature which will cause the output audio to have some harmonic distortion. The fast rise time is more strictly limited. The ideal selector pulse should have a linear rise time of 0.8 microseconds and decay time of less than 0.05 microseconds.

The negative polarity amplitude modulated pulse train on the plate of the 6AS6 is coupled to the cathode of a 1N34 germanium diode which acts as a peak detector feeding into an R C circuit with a time constant of 27 microseconds. This allows the pulse to decay to 69% of its peak value by the time of the next pulse, essentially maintaining the peak value. The time constant of this circuit must be carefully chosen with consideration to desired frequency response of the system and subsequent filtering. The use of a long time constant will more nearly preserve peak amplitude and thus filter out more of the 100 kc. component. If a longer time constant had been used for this circuit the distortion of 20 kc. audio signals would be increased.

The demodulated audio waveform across this RC circuit is fed to the grid of a cathode follower audio amplifier.

G. AUDIO AMPLIFIER

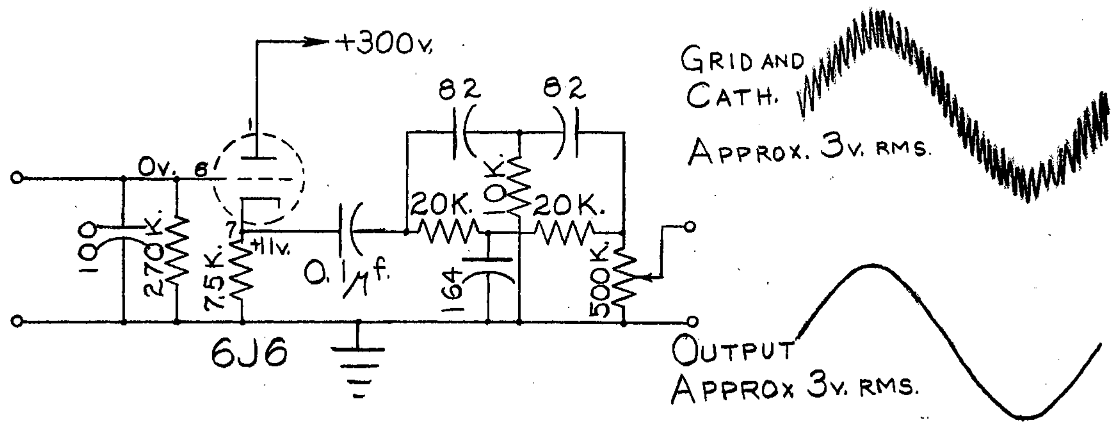

Figure 10
Schematic, Voltages, and Waveforms
for Audio Amplifier

The audio amplifier show in Figure 10 is a cathode follower circuit of conventional design used to present a high impedance load to the detector RC circuit which also serves as the grid circuit of the cathode follower. The cathode waveform of this stage is the recovered modulating voltage plus a 100 kc. component of considerable magnitude. The cathode is coupled to a rejection filter which is tuned to reject the 100 kc. pulse carrier. This filter is a twin-T null network which rejects the frequency at which $f = \frac{1}{2\pi RC}$. In the above circuit R is 20 K. and C is 82 micromicrofarads which gives rejection at 97 kc. The normal tolerance of the components used is such that rejection at 100 kc. is adequate. Several low-pass and rejection filters were tried using both RC and LC circuits, but the twin-T proved most effective in eliminating the unwanted frequency. A disadvantage of this circuit is that it has such wide bandwidth that it attenuates frequencies between 15 kc. and 20 kc. The loss of high frequency response in this system is due almost entirely to the characteristics of this filter circuit. The filtered demodulated waveform is applied to the grid of the audio output stage.

H. AUDIO OUTPUT

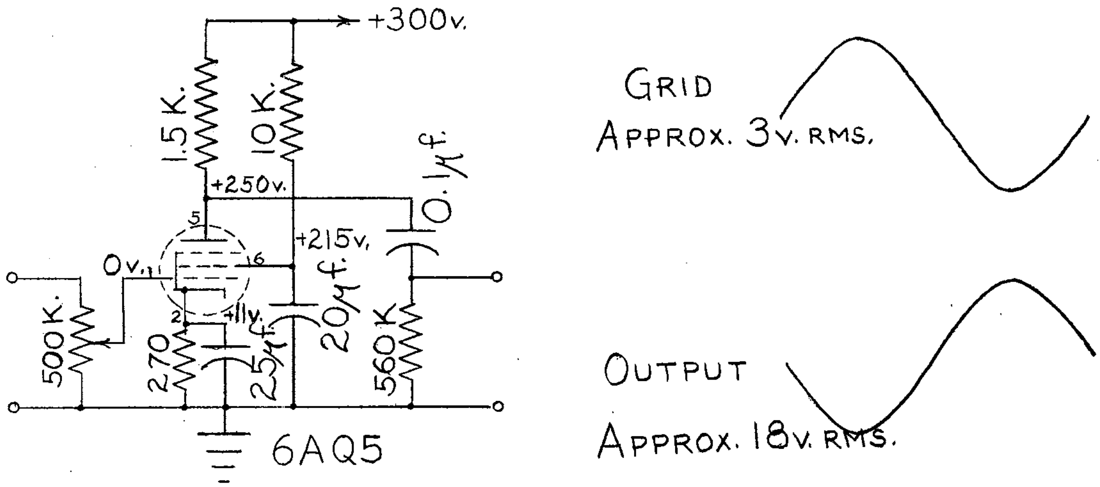

Figure 11
Schematic, Voltages, and
Waveforms for Audio Output

The audio output stage shown in Figure 11 is a conventional resistance coupled audio amplifier with cathode bias. This circuit operates with a transconductance of approximately 4000 micromhos and a plate load resistance of 1.5 K. which results in a gain of 6. The 6AQ5 is intended primarily for use as a transformer coupled amplifier and was used in this stage so that easy conversion to an audio output transformer could be made. A large screen bypass capacitor must be used to prevent screen grid signal from causing crosstalk since a common screen grid circuit is used. In order to develop proper bias a cathode resistor of 270 ohms must be used because of the large tube current in a power stage. This low cathode resistor is very difficult to bypass adequately at low audio frequencies and the loss of audio response below 200 cycles is quite noticeable even with a 25 microfarad capacitor.

If this stage were used as a single ended audio output stage with transformer coupling to a speaker a larger cathode resistor could be used because of the lower d.c. drop and higher a.c. reactance of the transformer. This cathode resistor could be bypassed more easily and low frequency response improved. As now used the audio output stage should not be loaded by a resistance of lower than 100 K. unless further low frequency attenuation can be tolerated. This stage may have an output voltage of 10 to 20 volts depending on the percentage of modulation and the gain control in the grid circuit of this stage.

The complete demodulator schematic is shown in Figure 12. The circuits for each stage are arranged in the same manner as the block diagram shown in Figure 1.

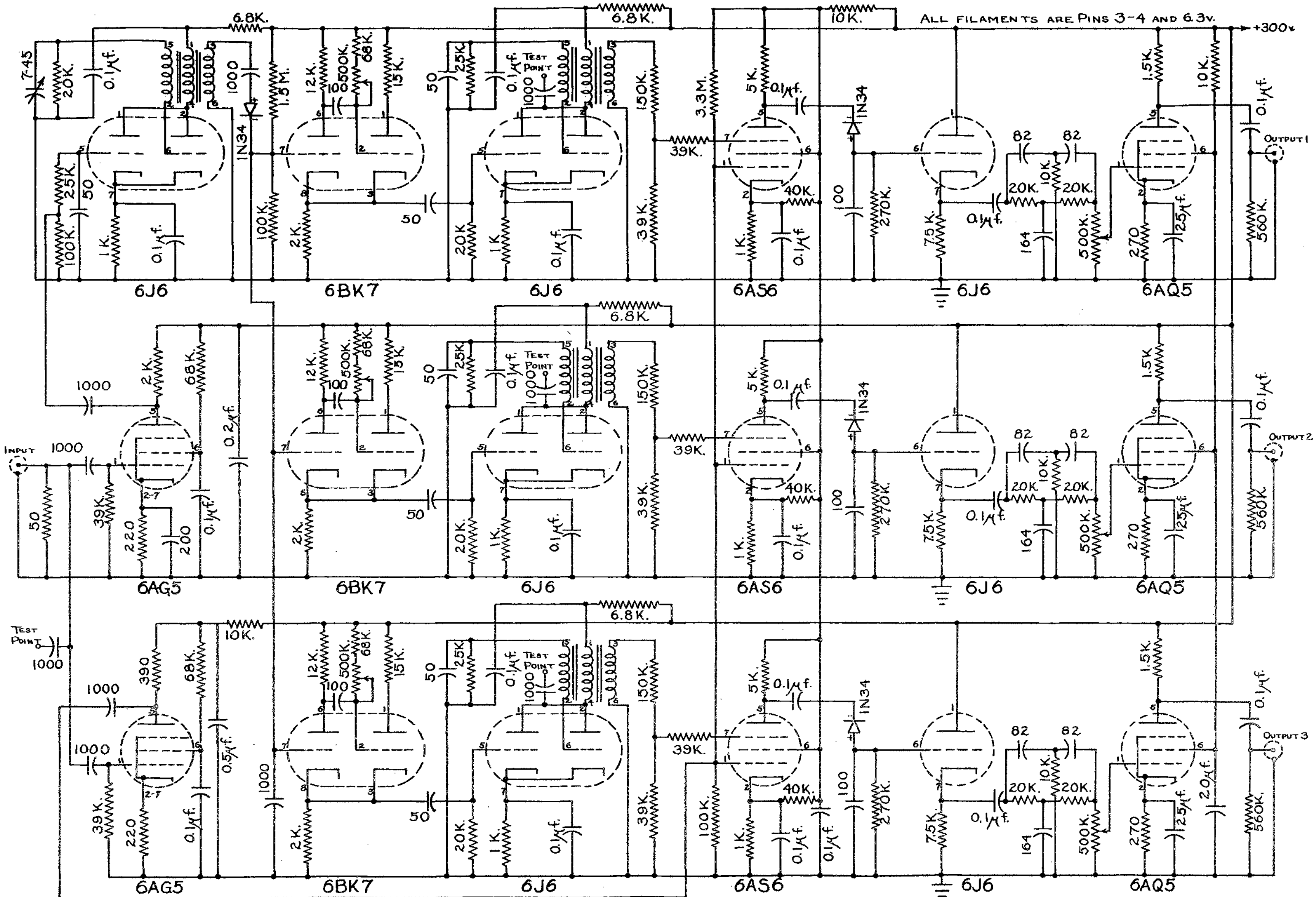

Figure 12
Demodulator Schematic

PART IV

TEST PROCEDURE AND RESULTS

Tests of frequency response, distortion, and crosstalk were performed in order to better understand the operation of the system. The tests of frequency response and distortion were performed simultaneously for each channel. One transmitter channel was modulated for these measurements by a Hewlett-Packard Model 200-B audio oscillator. Modulation deviations of 100%, 80%, 50%, and 30% were used at spot frequencies from 20 cycles to 20 kilocycles and distortion measurements made at check frequencies between 30 cycles and 15 kilocycles with a Barker and Williamson distortion meter. Percentage of modulation was measured by time measurement on a Du Mont Type 256-D synchroscope. This synchroscope could not be synchronized directly on the 100 kc. from the transmitter so it was necessary to build a blocking oscillator divider which would divide the 100 kc. down to a repetition frequency of around 1000 cycles. The first blocking oscillator triggers on the 10th pulse and the second blocking oscillator triggers on the 9th pulse. The output of the second blocking oscillator is used to synchronize the synchroscope. Because of this lowered repetition frequency on the test synchroscope a zero beat may occur at certain modulation frequencies so that the pulse presents the appearance of being unmodulated. If this situation is suspected it may be verified by making a slight change in the modulation frequency which should bring the pattern to a normal time modulated appearance. All output voltage measurements were made with a Hewlett-Packard Model 410-A vacuum-tube-voltmeter and audio output was monitored with a Du Mont Type 208-B oscilloscope.

The results of these tests are plotted on the following graphs and discussed individually. The data taken was so extensive that it could not be presented conveniently in tabular form. The graphs show the actual test data points plotted as a small circle with the average curve shown.

The frequency response characteristics of the system are shown in Figure 13. Each point plotted represents an average of 14 measurements. Data was taken at four different percentage of modulation in each of the 3 demodulator channels. In addition two percentages of modulation were tested with detection on the leading edge of the amplifier detector suppressor grid sawtooth voltage. The system shows response within 3 db. from 48 cycles to 16 kilocycles and within 1 db. from 140 cycles to 9 kilocycles. The loss of low frequency response is principally due to the inadequate cathode bypass capacitor of the audio output stage. The curve shown in dotted lines above the overall response curve show the low frequency response on the grid of the audio output stage and shows a 1 db. loss at 71 cycles and 3 db. loss at 19.5 cycles. Unless a very high quality speaker system were used the low frequency response of the system is quite adequate. The high frequency loss is due principally to the filter characteristics, but since the perceptible loss of response is at almost the high frequency cut-off of the human ear this loss of response should not be objectionable even if the speaker system were adequate to permit reproduction of these frequencies.

The results of distortion tests are shown in Figure 14. A separate curve is shown for each percentage of modulation.

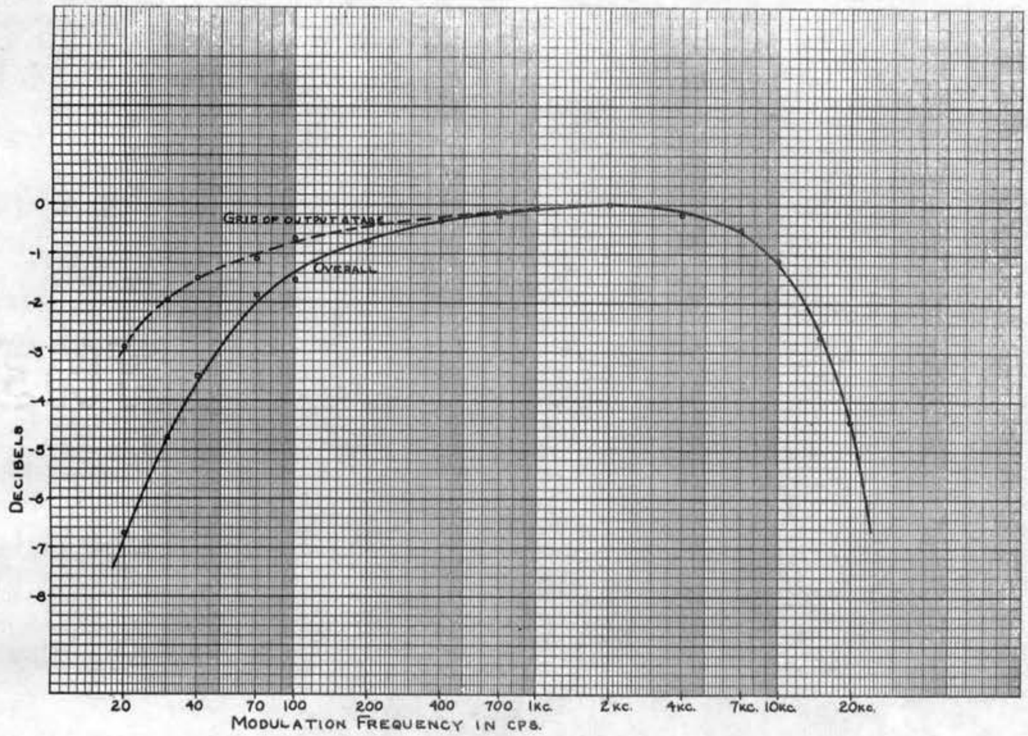

Figure 13
System Audio Frequency Response

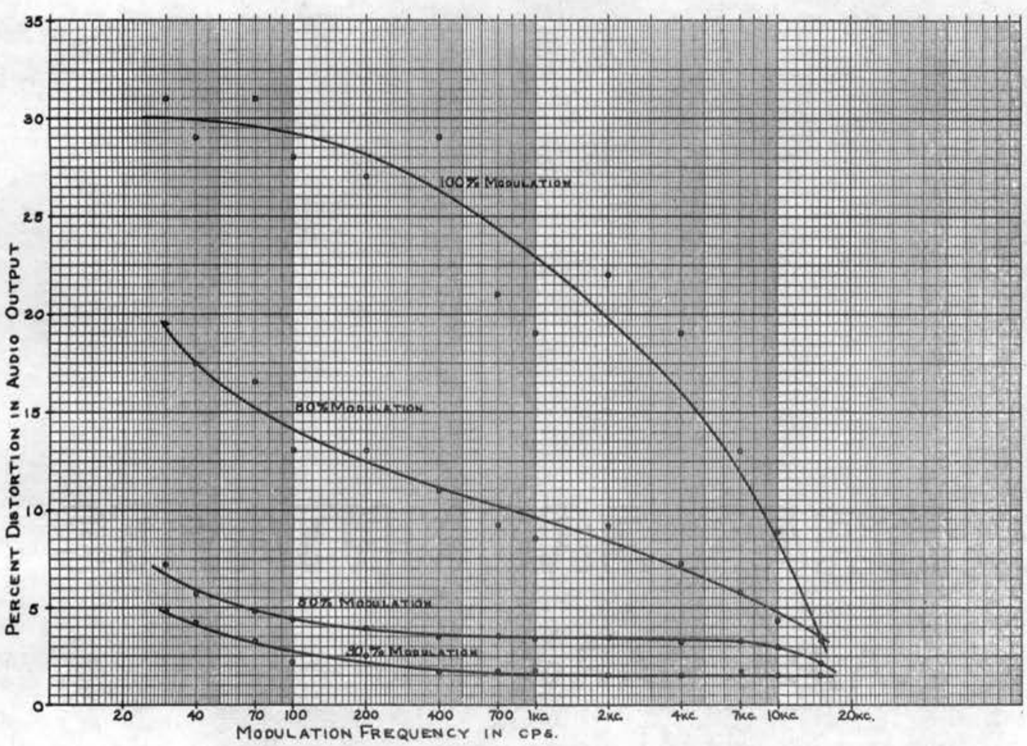

Figure 14
System Distortion Characteristics

Points plotted are the average for the three demodulator channels. The distortion at higher modulation percentages is quite objectionable particularly at lower frequencies. In analyzing this data due consideration must be given to the characteristics of the distortion meter. This meter measures total distortion, including noise, at frequencies up to 45 kc. Thus any 100 kc. distortion remaining in the demodulated audio would not be measured. The monitor oscilloscope indicates that this 100 kc. distortion is quite noticeable when 30% modulation is used although the measurement does not indicate this. If a low modulation percentage were used consistently it would be desirable to increase the time constant of the detector circuit. In fact the measurements of distortion at 15 kc. indicate that this time constant could be increased without noticeable effect on high frequency distortion since the measurable distortion is a minimum at higher audio frequency. Probably one reason for the lower measured distortion is that the distortion meter could only measure up to the third harmonic of 15 kc. but could measure much higher harmonics of lower audio frequencies. Figure 15 also indicates another cause for low frequency distortion. A distortion measurement at 60% modulation on the grid of the audio output stage shows distortion to be nearly constant at frequencies below 1 kc. so apparently the audio output stage is contributing to the low frequency distortion.

A test of distortion with detection on the leading edge of the amplifier detector suppressor sawtooth voltage is plotted in Figure 16. While these tests were not sufficiently extensive to provide conclusive evidence they seem to indicate that linearity

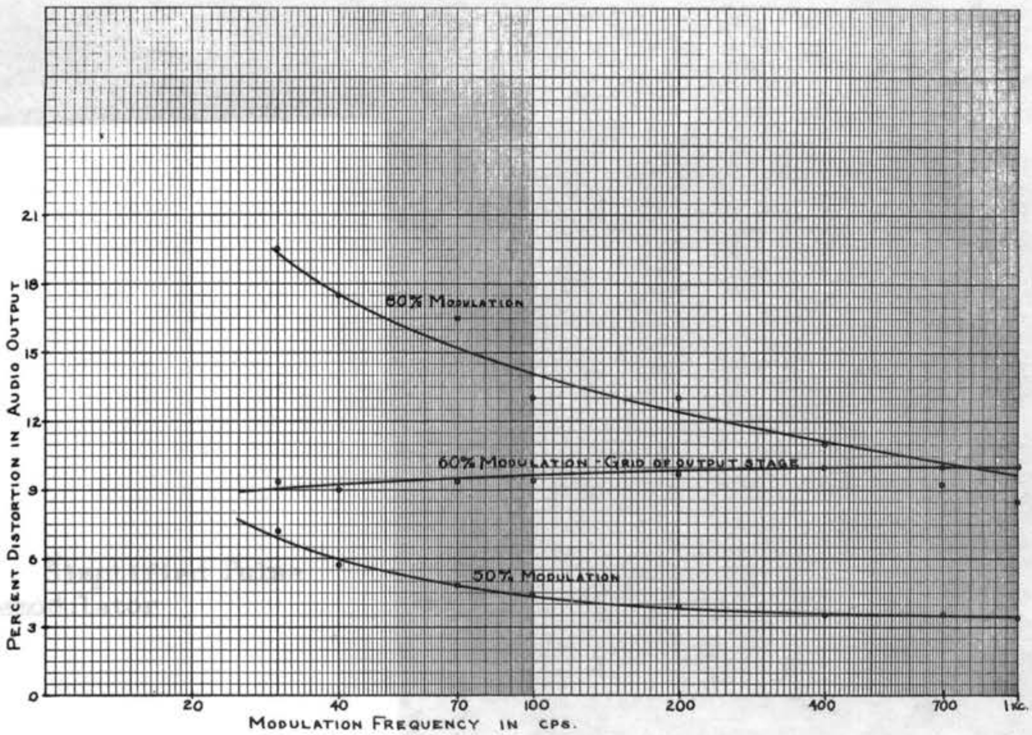

FIGURE 15
DISTORTION IN AUDIO OUTPUT STAGE

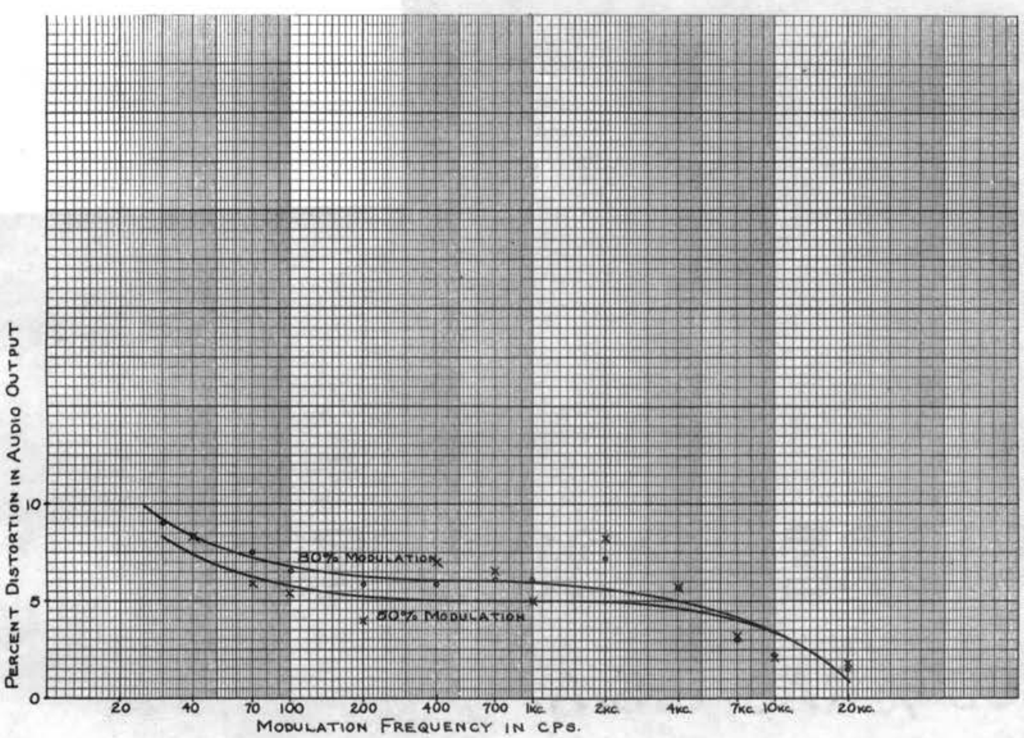

FIGURE 16
DISTORTION WITH LEADING EDGE DETECTION

of the suppressor sawtooth voltage can greatly increase the system quality in regard to distortion. The slight difference between distortion percentages at 80% and 50% modulation is unusual especially since the theory indicates that modulation percentages of greater than 37.5% should cause very great distortion due to clipping when leading edge detection is used. The cause for this discrepancy between theory and practice could not be discovered. Most probable sources of error are in measurement of modulation percentage by time measurement or change of the effective suppressor grid waveform on the amplifier detector depending on whether or not an oscilloscope probe is connected to this point. Tests at lower modulation percentage were not performed due to the low output voltage at 50% modulation containing a considerable amount of 100 kc. interference.

Test of crosstalk were performed by detecting a transmitter channel with 6 kc. audio modulation and presenting the detected output on a Du Mont Type 208-B oscilloscope. The adjacent channel pulse was modulated at 300 cycles and the spacing between these pulses varied. The amplitudes of desired signal and adjacent channel signal were measured directly on the oscilloscope using a low sweep rate (approximately 30 cycles). A slight modulation due to 60 cycles was noted but the amplitude was less than the adjacent channel modulation. Apparently this 60 cycle interference was due to a 60 cycle amplitude modulation in the transmitter. The synchroscope transmitter monitor showed a slight amplitude modulation of the transmitter channel pulse which had not been noted before. This test was the last one performed and this 60 cycle modulation of audio output was not perceptible in earlier tests. In conducting these tests the desired channel pulse was set near the middle of the period and the adjacent modulated channel moved closer from both directions with both pulses at 50% modulation. With the adjacent channel pulse occurring after the desired channel pulse the adjacent channel crosstalk was greater than 20 db. below the desired channel at spacings from 0.8 microseconds to 5 microseconds. When this spacing was decreased to 0.7 microseconds the adjacent channel interference became 4.6 db. below the desired channel even though 50% modulation could not be attained. At this point the pulses were also interfering in the transmitter so it is obvious that the cross talk would greatly increase. With the adjacent channel pulse preceding the desired signal pulse approximately the same results were obtained with one exception. At the extreme spacing which could be obtained (3.5 microseconds) the adjacent channel was 19.4 db. below the desired signal. Since transmitter delay multivibrator characteristics made it necessary to set the adjacent channel pulse very close to the marker pulse it is possible that modulation of this channel caused a slight modulation of the marker pulse which increased the crosstalk. At all other pulse spacings from 0.8 microseconds to 3 microseconds the adjacent channel interference was 20 to 30 db. below the desired signal. At 0.7 microseconds spacing the adjacent channel was 17.7 db. below the desired signal and interference in the transmitter was occurring. On the basis of this crosstalk test channel pulses could be spaced every 0.8 microseconds and crosstalk not be objectionable.

Checks were also made to determine the time available for channel pulses. Due to the demodulator characteristics a channel pulse must be set at least 3.5 microseconds after the beginning of the marker pulse. This time is used as follows:

Integration time	0.9	microseconds
Multivibrator minimum delay	1.7	
Gate rise time	0.4	
½ Gate decay time	0.5	
Total	3.5	microseconds

Actual measurement of maximum time could not be made because the transmitter channel pulse which reached the long delay time could not be modulated properly. A measurement of available delay range on the demodulator delay multivibrator indicates a delay from 2.6 microseconds to 8.4 microseconds. This leaves 5.8 microseconds of the total of 10 available for channel pulses. If these pulses were spaced every 0.8 microseconds, 7 channels could be used.

PART V

CONCLUSIONS AND RECOMMENDATIONS

The results obtained show that a simple demodulator for position modulated pulses is not only feasible but practical. While the results did not indicate that the existing demodulator gave the highest quality of reproduction they did much to indicate improvements which could be made.

While the demodulator proved reliable during the test procedure the greatest deficiency of the system was in respect to distortion. The increased distortion at high percentages of modulation shows that this is largely due to non-linearity of the sawtooth waveform applied to the suppressor grid of the amplifier detector. The feasibility of this general technique of demodulation seems to have been proved. In order to obtain a more nearly linear sawtooth waveform the circuit should be modified to include a gate generator which would generate a negative square wave of the same width as a channel pulse at the end of the multivibrator delay time. This square wave could then be applied to a sawtooth generator of the bootstrap type. It would possibly prove desirable to make such modifications as might be necessary to cause the present gate generator to have a pulse of slightly longer duration and then by diode clipping use this as the square wave generator to supply the gate waveform for a bootstrap sawtooth generator which could use a dual triode tube. The 6BK7 tube used for the delay multivibrators would probably be suitable for this circuit since it has separate cathodes and also high transconductance. If constant response over the entire audio spectrum were desired it would be of great advantage to use a minimum of audio amplification on the demodulator chassis proper. This is entirely feasible since the level of the audio signal at the grid of the present audio output stage is in excess of 1 volt and appears at a low impedance source. This audio output could then be amplified by a high quality amplifier and in turn feed a high quality speaker system.

While the 100 kc. interference is not objectionable except at low modulation percentage it could be reduced to an even lower level by using filter components which could be adjusted to give a rejection frequency of exactly 100 kc. instead of using stock components. The 100 kc. distortion could also be reduced by increasing the time constant of the detector circuit. It seems desirable with the existing system to use a low modulation percentage and if this were done the time constant could be increased considerably. If desired, a recycling detector could be used but the results obtained with the simplest form of detector seem to be adequate, so the increased complexity of this type of detector hardly seems warranted.

One of the most severe limitations of this system lies in the small number of channels that can be transmitted. While the number of channels could be increased by narrowing the width of the pulses it is necessary to have a considerable time deviation in order to keep the signal to noise ratio within reasonable limits. Thus the actual limit to the number of channels is set not by the pulse width, but by the minimum allowable time deviation. This minimum time deviation is in turn determined by the inherent "jitter" or noise due to random variations in triggering time of the pulse circuits involved. This noise level can be decreased by proper circuit design, but this would result in more complicated circuitry. Hence, the number of tubes does not increase directly as the number of channels increases because a large number of channels results in different design criteria. In order to increase the time available for channel pulses it would be very desirable to decrease the minimum multivibrator period. This could be done most easily by using different multivibrator circuit designs with a more limited range of period. The best way to decrease the minimum delay is by adjustment of the grid potential. This could be done by using separate bleeder resistors for each grid and separate coupling condensers for d.c. blocking. Under proper conditions the basic multivibrator design used in this system will operate properly at periods of less than a microsecond. The circuit used had a reasonably low minimum delay and a full range of adjustment. By decreasing this minimum delay one additional channel could be added or a total of 8 channels. If the integration time could be eliminated this would also permit an additional channel. This could be done by using amplitude or polarity to distinguish the marker pulse. If amplitude was used to distinguish the marker pulse the demodulator would then have to be modified to eliminate the integration time delay though the same basic circuits could be used. If a marker pulse of polarity opposite to the channel pulses were desired both transmitter and demodulator would require redesign. The transmitter would require modification so that the marker pulse would be of positive polarity since the channel pulses are negative. The demodulator would then be modified to synchronize on a positive pulse rather than a wide negative pulse which requires integration.

The operation of the 6AS6 amplifier also has possibilities of further development work. Time did not permit an extensive analysis of this stage to see if the circuit used was the optimum design. Potentials used were selected with due consideration to the tube curves shown in a tube handbook, but screen potential or plate load resistance changes might result in this circuit having increased gain or improved output waveform.

During the final series of tests operation of the transmitter seemed to indicate that it should either be modified or redesigned. The position of the pulses was not stable and occasional shifts in position were noticeable on the monitoring synchroscope. The extent of interaction between controls is also troublesome since both pulse amplitude and modulation percentage vary with position adjustment. The method of modulation also has many difficulties. By feeding the modulation into a high impedance point the impedance of the modulation source affects not only the mean pulse position but also the pulse width. The modulation source must also have d.c. continuity in order to permit circuit operation. This is also a distinct disadvantage since it requires transformer coupling of modulation when capacitor coupling would often be more desirable. It is believed that a redesign of the transmitter would be warranted in order to eliminate these difficulties as well as reduce transmitter complexity. By using blocking oscillators with diode clipping circuits for pulse generators the need for the large number of amplifiers would be eliminated. A delay multivibrator similar to the design used in the demodulator could probably be used in the transmitter with the modulation voltage capacitor coupled to the common cathodes in order to cause position modulation. This type of modulation has not been tried but the theory could be easily verified on the demodulator chassis, by coupling an audio signal to the cathode of a delay multivibrator. It is obvious that this technique would not permit the cathode to be used for the output waveform, but the plate of this tube is free and may be used for obtaining the delayed output pulse. If this method of modulation could be used it would permit the audio modulation to be inserted into a low impedance point of the circuit and could be operated with ground as one terminal. If simplicity of testing is desirable the transmitter should include a synchronizing pulse divider since few synchroscopes will synchronize at a 100 kc. repetition rate. These changes in transmitter design should permit a circuit having approximately the same number of tubes as the present demodulator.

This study shows that a simplified design of a pulse position demodulator for a 100 kc. sampling rate is practical. The results of the tests indicate that the choice of this sampling rate is justified only if a wide audio bandwidth is necessary and if a small number of channels is adequate.

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THESIS TITLE: A STUDY OF A SIMPLIFIED PULSE POSITION DEMODULATOR

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