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A DEMODULATOR FOR COMPLEX-MODULATED PULSES

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ABSTRACT

A Complex-Modulated-Pulse Demodulator has been developed to demodulate pulse signals with respect to pulse amplitude, pulse position, pulse width, pulse period, and pulse frequency over a pulse-repetition-rate range of 20 cycles to 1 Mc and a pulse-width range from 0.1 to 100,000 μsec. This range is covered by a two-position switch; no tuning is required within the ranges.

A gating system is included which, in conjunction with a multiple slave-sweep type display, enables the operator to visually select for demodulation any desired pulse from a pulse train. Slicing circuits are provided for reducing the effects of receiver noise so that information from tangential signals can often be obtained. The output signal from the demodulator, derived from a boxcar generator having a linear frequency response from 2 cycles to 800 kc, is an accurate reproduction of the modulation envelope carried by the original video pulse signals. It is expected that this demodulator will be an invaluable tool in detecting information carried by complex-modulated-pulse signals, and it is probable that future secured guided missile and communications systems will utilize some kind of pulse modulation that can be handled by this demodulator.

PROBLEM STATUS

This is an interim report; work is continuing.

AUTHORIZATION

NRL Problems R06-04 and R06-16
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A DEMODULATOR FOR COMPLEX-MODULATED PULSES

INTRODUCTION

For several years there have been available pulse analyzers of the slave-sweep type that are capable of displaying complex-modulated-pulse signals. The utility of these analyzers is, however, severely limited when information from the pulse envelope is desired. Therefore, a Complex-Modulated-Pulse Demodulator (CMPD) has been developed to detect the information carried by the pulse envelope itself.

The general problem of the detection of information carried by complex-modulated-pulse signals will for convenience be arbitrarily divided into three general phases. The first phase deals with the detection of the pulse envelope impressed on a radio-frequency carrier and the subsequent amplification of this envelope by a video amplifier of suitable bandwidth. The second phase involves the demodulation of the information carried by the pulses themselves. The third phase treats the analysis of this information.

When the CMPD is used in conjuncation with a previously developed analyzer[1,2] of the slave-sweep type, it may offer a solution to the second phase of the complex-pulse-demodulation problem. If the pulse signals to be analyzed are modulated to a sufficiently high percentage, the general type of modulation can be observed directly from the slave-sweep-analyzer display. However, an analysis of modulation frequencies cannot usually be determined from such displays. Furthermore, if the percentage of pulse modulation is small, it becomes impossible to determine even the type of modulation with the resolution available on these analyzers. Tests have indicated that pulse modulations of such small percentage,that they are not detectable on currently available analyzers,can be demodulated with a very favorable signal-to-noise ratio using the CMPD. On the basis of this test alone it becomes obvious that in addition to the present analyzers more equipment is needed to indicate whether or not modulation even exists.

The major tool used in developing the CMPD was a complex-pulse-modulation simulator[3] which had been designed for the purpose. This generator simulates a two-channel pulse system where either or both pulses can be individually modulated in amplitude, width, period, frequency, or position with respect to a sync pulse. It is the function of the demodulator to select the desired pulse and reproduce its modulation envelope over a wide range of pulse repetition rates and pulse widths with a minimum of "tuning" adjustments.

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[1] Munzer, E. N., "Improved Countermeasures Pulse Analyzer Techniques," NRL Report 3847 (Confidential), August 29, 1951

[2] Markell, J. H., "Refinements in Countermeasures Signal Analyzer Techniques," NRL Report 4341 (Confidential), April 27, 1954

[3] Holmes, J. C., "An Improved Complex Pulse Modulation Simulator," NRL Report 4184 (Confidential), July 30, 1953

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The actual analysis of this envelope represents the third phase of the problem. The CMPD reproduces the modulation envelope of a modulated pulse signal. The problem of deciding whether the signal is a missile-control signal, a coded voice communication channel, or something else, has yet to be fully investigated.

GENERAL DESCRIPTION

The Complex-Modulated-Pulse Demodulator(Fig. 1) serves a threefold purpose. First it enables the operator to select at will for demodulation any desired pulse from a train of video pulses such as one might encounter in a multichannel pulse communications system. Second, it reproduces the modulation envelope from the modulated video pulse, whether the pulse be frequency, position, period, width, or amplitude modulated. Third, it has provisions for eliminating the effects of low level noise on a signal and improving greatly the envelope signal-to-noise ratio for very noisy signals. In addition, the operator has the use of an AN/SLA-2, APA/74, or similar type analyzer on which he can display the raw amplified video as it is fed to the input of the CMPD, the selected pulse in a train to the exclusion of all others, or the selected pulse after it has passed through the noise-slicing circuits.

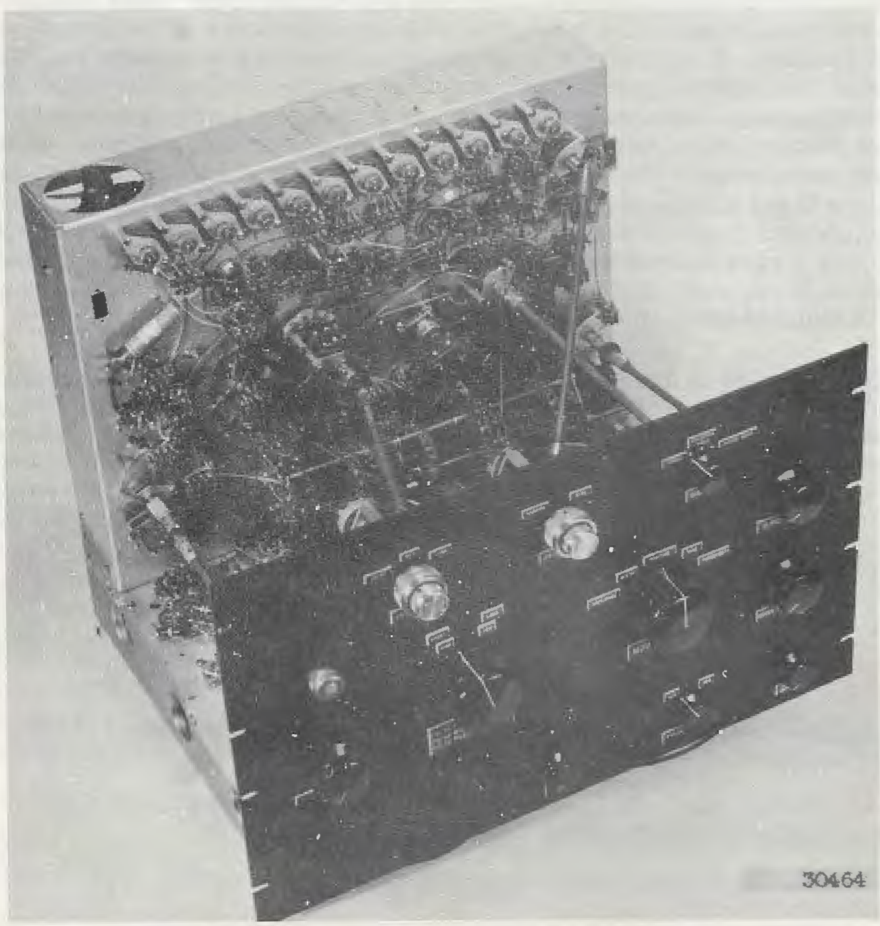

Fig. 1 - Complex-Modulated-Pulse Demodulator

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Figure 2 is a block diagram and Plate 1 attached at the back of this report is a complete circuit diagram of the CMPD. The raw video signal from the video amplifier is fed to a display switch, the video pulse selector, and to the input of the sync separator. The sync separator produces a narrow trigger pulse for every sync pulse coming in. This trigger pulse initiates a delay gate, the width of which is controlled by the operator. The back edge of this gate initiates another gate whose width is likewise adjustable. This second gate, the pulse selection gate, is sent to the video pulse selector. A video pulse appears at the output of the video pulse selector only if it occurs during this selection gate. The pulse selection gate also is fed to one input of an addition circuit; the sync-separator trigger-pulse output is fed to a second input. If the DISPLAY switch is set to the RAW VIDEO position, the raw video (Fig. 3a), the sync-separator output, and the selection gate are all added and appear at the output of the three-input addition circuit. This signal is fed to the slave-sweep analyzer where it might appear as shown in Fig. 3b, which represents a two-channel pulse system with sync pulse and shows the superimposed negative-polarity selection gate. Figure 3c shows the selection-gate delay and width so adjusted that the selection gate just includes the third pulse in the two-channel pulse system.

The pulse thus selected is sent from the video pulse selector to the display and modulation switches and to the noise-slicing circuit where the top and bottom are literally "sliced" from the pulse to eliminate all noise pulses except those of sufficient amplitude to extend into the slice. With the exception of amplitude-modulated pulses the modulation is preserved upon passing through the slicer, so a path by-passing the slicer is provided for the amplitude-modulated pulses.

Period or frequency-modulated sliced pulses are distributed to a special pulse stretcher. Pulses whose position with respect to the sync pulse is modulated (position modulation) are fed to a circuit that produces a gate whose width is equal to the time interval between the sync pulse and the information pulse. The output pulses from the pulse-stretching circuit, the position-to-width converter, and the noise-slicing circuit then pass through the modulation switch to a Schmitt circuit (output Schmitt circuit) which changes the pulses to ones of constant amplitude but leaves the pulse-time and width relationships unchanged.

At the input of the time-to-amplitude converter now appear pulses whose width, period, or frequency is modulated. The time-to-amplitude converter changes this width in the case of pulse width or pulse position modulation, or it changes the period in the case of pulse period or pulse frequency modulation to a pulse whose amplitude varies linearly as the respective pulse widths or periods. In Fig. 2 it can be seen that all pulses, regardless of the original type of modulation, have been transformed into amplitude-modulated ones. Those incoming video pulses that were originally amplitude modulated are by-passed around the time-to-amplitude converter over the same path that passed them around the noise-slicing circuit.

The amplitude demodulator produces from these amplitude-modulated pulses a dc output whose level is always equal to the amplitude of the last pulse to enter the amplitude demodulator. Figure 4a shows an amplitude-modulated pulse as it is fed to the input of the amplitude demodulator, and Fig. 4b shows the resulting envelope produced at the output. The short negative spikes appearing at each "step" in the output envelope will be explained later in this report.

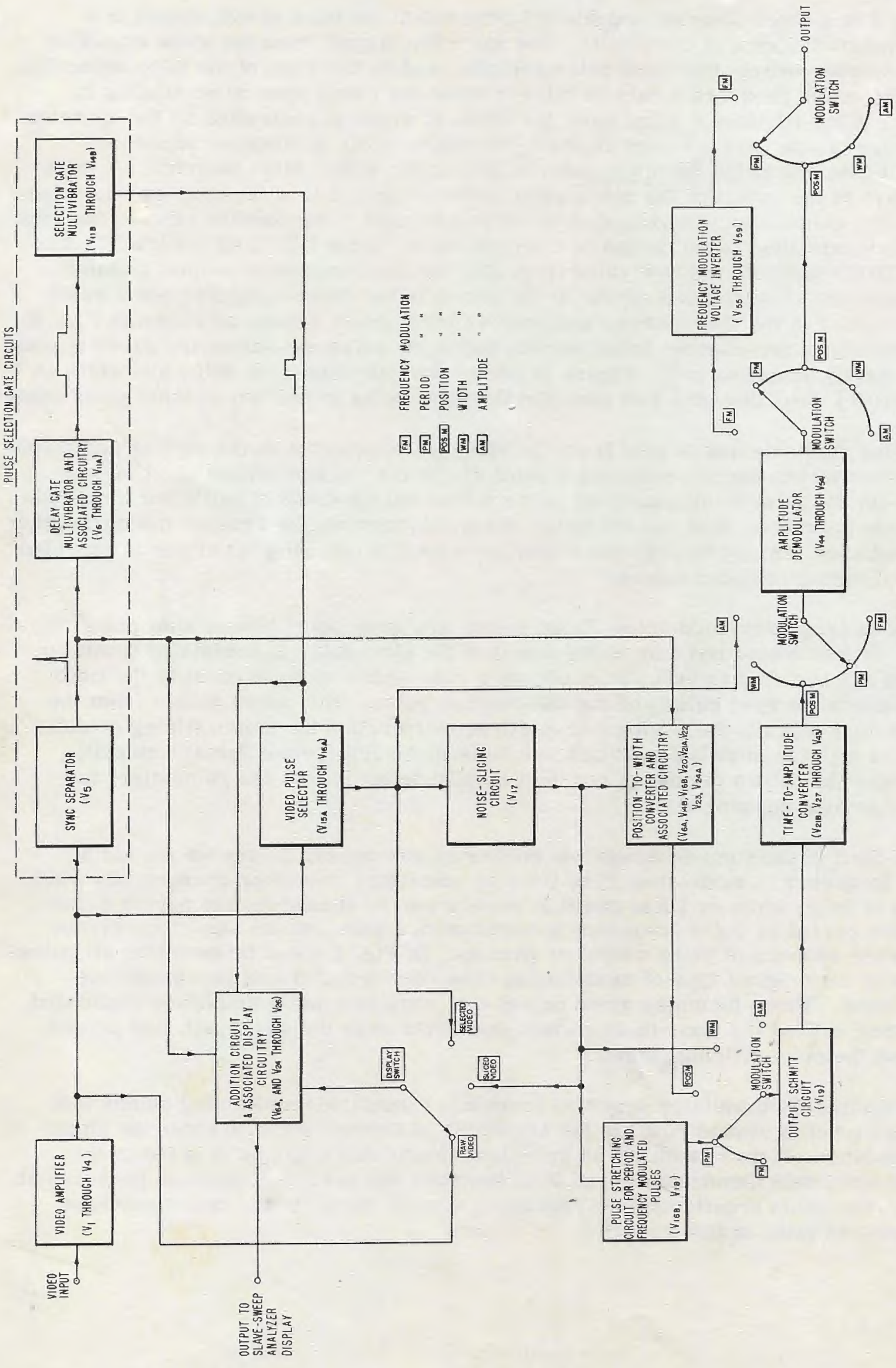

Fig. 2 - Block diagram of Complex-Modulated-Pulse Demodulator

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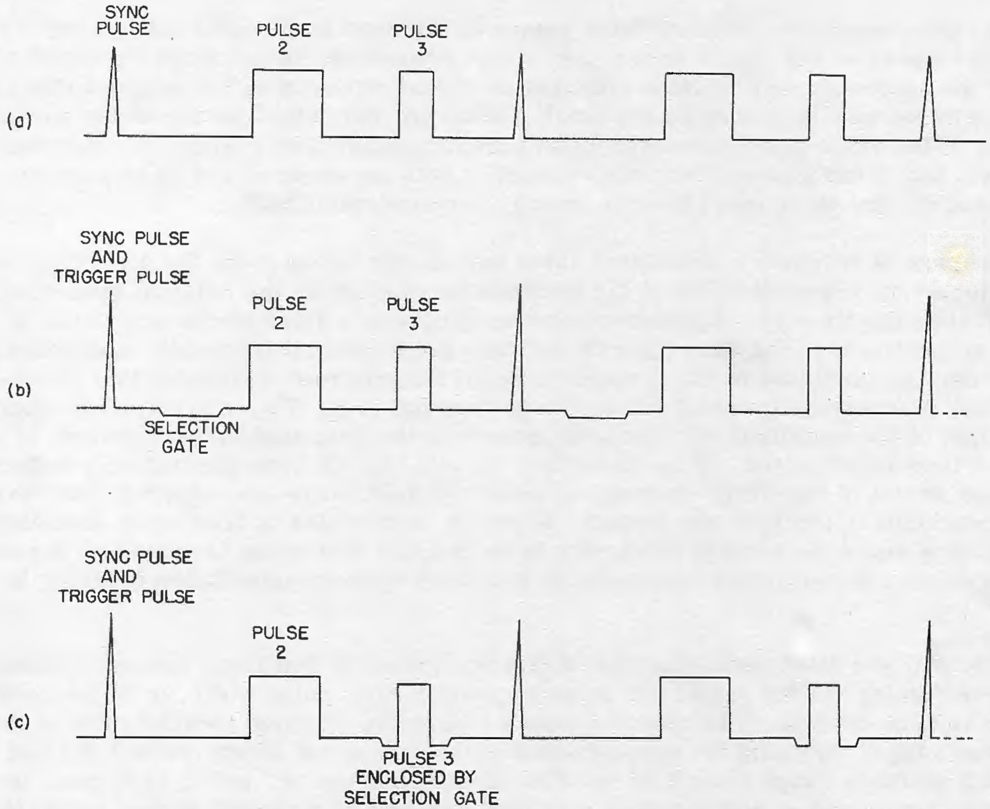

Fig. 3 - Typical displays with display switch set to RAW VIDEO position

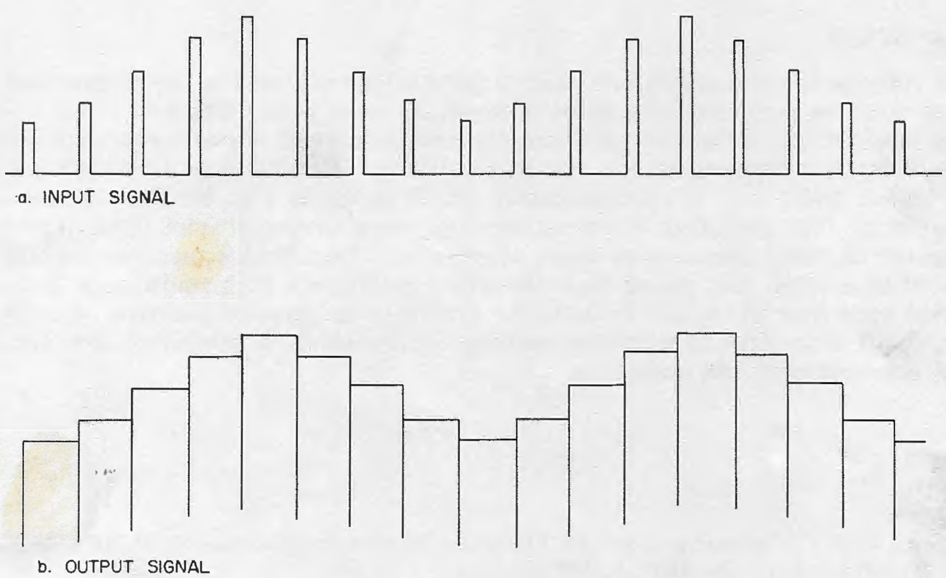

Fig. 4 - Typical signals from amplitude demodulator

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In the case where the original video pulses at the input to the CMPD are simply modulated with respect to any single parameter except frequency, the envelope produced at the output of the amplitude demodulator contains all the information in the original video pulse plus extra harmonics produced by the small spikes and the "step" nature of the output envelope. If the video pulses are modulated simultaneously with respect to more than one parameter, and if the modulations with respect to both parameters are to be demodulated simultaneously, the pulse must be sent through two separate CMPD's.

In the case of frequency-modulated video pulses, the output from the amplitude demodulator is not a true representation of the information carried by the original modulated signal because the time-to-amplitude converter produces a pulse whose amplitude is directly proportional to the time interval between the individual frequency modulated pulses, and inversely proportional to the frequency. A circuit has been developed that produces a voltage that is inversely proportional to the voltage fed to it. This circuit, when connected to the output of the amplitude demodulator, produces the true modulation envelope of the frequency-modulated pulses. This "inverter" circuit has not been permanently included in the present model of the CMPD because of technical difficulties encountered with respect to long term drift in the inverter circuit. However, in the case of frequency-modulated signals having small percentage frequency deviation, the distortion produced by the amplitude demodulator is negligible (Appendix A) and the frequency-modulation inverter is unnecessary.

An unusual and most useful feature of this equipment is that there are no circuits that necessitate "tuning" to the particular pulse repetition rate, pulse width, or pulse position that is to be demodulated. The operator sets a single two-position switch to one of the two overlapping ranges covering the pulse-repetition-rate range of 20 cycles to 1 Mc and the pulse width/position range from 0.25 to 1000 μsec. The "tuning" within each range is all done automatically and is completed in less than one second after the proper range is selected. These ranges overlap to such an extent and the automatic "tuning" circuits are so effective that it is most improbable that an operator would set the range switch to a range where the automatic "tuning" circuits could not operate properly.

VIDEO AMPLIFIER

Before video pulse signals can be sent to the various circuits in the demodulator proper, they must be amplified to a value of about 20 volts peak. The raw video signals found at the video output terminals of currently available countermeasures receivers are often of the order of a fraction of one volt in amplitude. A video amplifier having a rise time of 0.04 μsec and a gain of approximately 100 is shown in Fig. 5. With the exception of the output stage this amplifier is essentially the same as one previously described in an NRL report[4] on countermeasures video amplifiers. Two 404A's, used as amplifiers, are separated by a split-load phase-inverter which provides a high impedance output for the first 404A amplifier stage and enables the amplifier to produce positive polarity pulse output with either a positive or negative polarity input signal. A miniature low-capacity relay (SW1) accomplishes the switching.

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[4] Hendrickson, A. H., "Countermeasures Video Amplifier Improvements," NRL Memo. Report 71 (Confidential), October 1, 1952

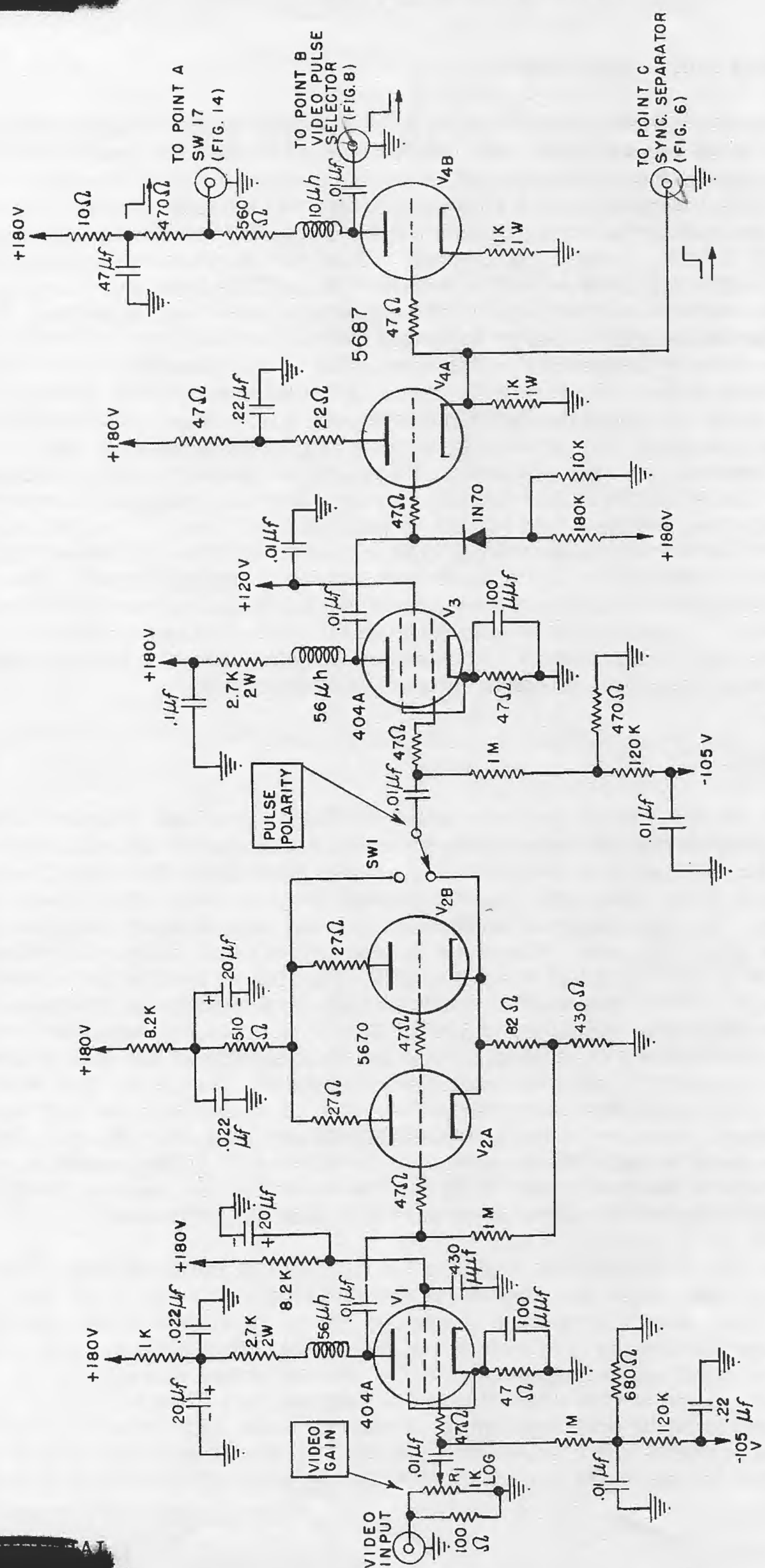

Fig. 5 - Video amplifier circuit

PULSE SELECTION GATE CIRCUITS

The pulse selection gate circuits (Fig. 6) provide a gate whose width and delay with respect to a sync pulse are adjustable and which when fed to the video pulse selector selects for demodulation only that pulse which occurs during this selection gate. The positive polarity pulse output from the video amplifier (V4A cathode signal) is connected to the grid of a sync-separator Schmitt circuit (V5A, V5B). By varying the bias on this grid with the SYNC LEVEL control, the Schmitt circuit can be made to commutate only on pulses whose peak amplitudes exceed a certain voltage. The sync pulse produced by the complex-pulse-modulation simulator is of somewhat greater amplitude than that of the other two information pulses, and by adjusting the bias on the Schmitt circuit grid, the circuit can be made to commutate on the sync pulse alone, the other pulses being of insufficient amplitude to pass the threshold levels of the Schmitt circuit. (With signals whose sync pulses are not of greater amplitude than the information pulse amplitudes, other circuits must be added to provide a sync pulse of greater amplitude than the others before the signal reaches the sync separator.) The positive polarity constant amplitude pulse produced by the sync-separator Schmitt circuit is fed to a cathode follower (V6A) to provide both a high-impedance load for the Schmitt circuit output and a low-impedance driving source for the circuits to be driven. The output pulse from the cathode follower drives an amplifier (V6B) having as its plate load a damped ringing circuit. This ringing circuit differentiates the plate current pulse producing a 0.12-μsec negative polarity pulse for every positive polarity pulse coming from the sync-separator Schmitt circuit. This pulse triggers the multivibrators that produce the delay gate, the trailing edge of which in turn initiates the pulse-selection-gate multivibrator gate.

Delay Multivibrators

The design of the delay multivibrators posed a difficult problem. Figure 7 illustrates a typical signal situation that dictates rather stringent requirements on this multivibrator's design. Suppose for example it is desired to select and demodulate the seventh pulse of the illustrated pulse train. The time interval between the sync pulse (No. 1) and pulse 7 is about 23.8 μsec. The time interval between the leading edge of pulse 7 and the following sync pulse is about 1.2 μsec. Therefore the duty cycle of the delay multivibrator must not be less than $100 \times 23.8/(23.8 + 1.2) = 95\%$. This figure is difficult to achieve with a standard, dual-triode, monostable multivibrator, so a cascade multivibrator was designed having a duty-cycle capability of 100%. This duty cycle was achieved by simply connecting two multivibrators in tandem so that the trailing edge of the gate produced by the first multivibrator triggered the second multivibrator. Now if the gate width of the combination is defined as the time interval between the initiation of the first multivibrator gate and the completion of the second multivibrator gate, then the duty cycle of each multivibrator need be only 50% in order that the duty cycle of the combination be 100%. It is the trailing edge of the gate produced by the second half of the cascade delay multivibrator that finally triggers the video pulse selection gate multivibrator.

Multivibrator V7A, V7B, and V8A produces the first half of the delay gate. It is plate-triggered by the negative pulse provided by the damped ringing circuit in the plate of V6B. The gate width is controlled by switching in various timing capacitors (SW2) and adjusting the time constant by varying the grid-return resistance R4. This potentiometer (R4) has a logarithmic taper which makes adjustment at the low resistance end (narrow gate widths) less critical. The cathode of this multivibrator is returned to a negative voltage (-105 v) to provide long term stability with tube aging. Clamping diode V8A limits the normally "on" current in V7B and provides a low impedance path for the discharge of the timing capacitor during the multivibrator recovery period. Two separate plate voltages for V7A and V7B were

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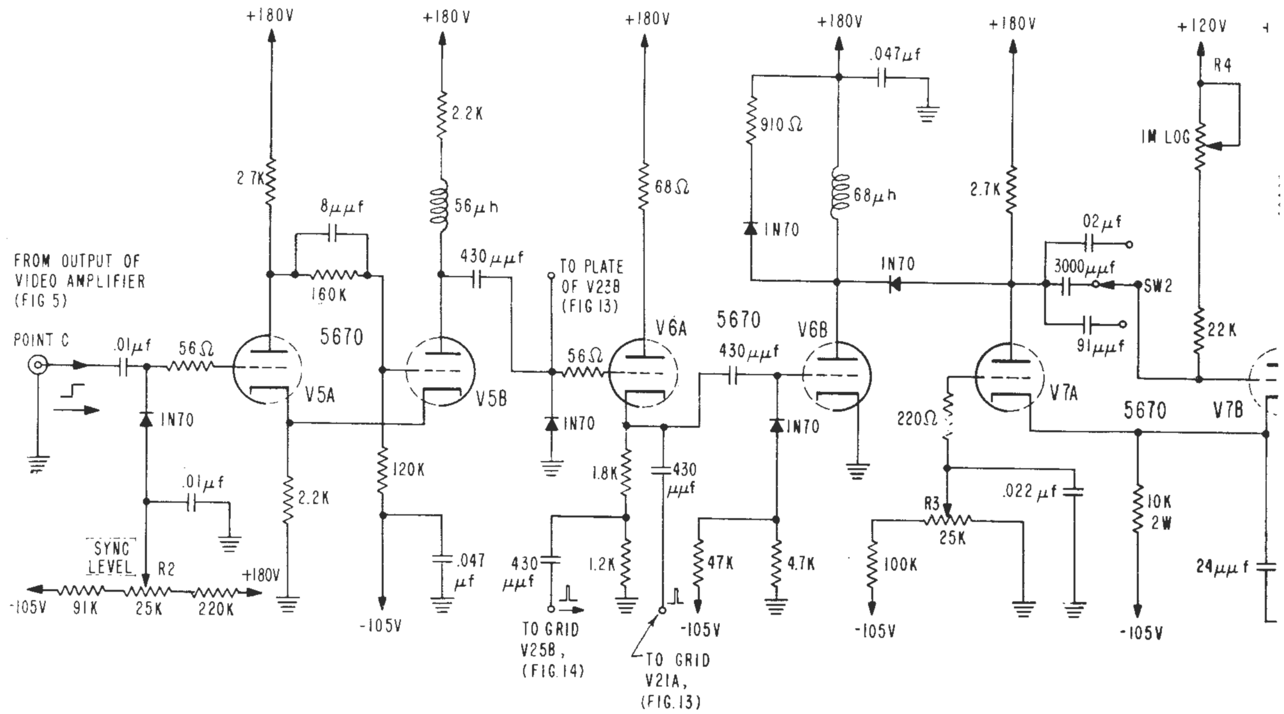

(SW2 AND SW3 GANGED TOGETHER) → SELECTION GATE DELAY RANGE SWITCH

SW4 → SELECTION GATE WIDTH RANGE SWITCH

(R4 AND R7 GANGED TOGETHER) → SELECTION GATE DELAY CONTROL

R10 → SELECTION GATE WIDTH CONTROL

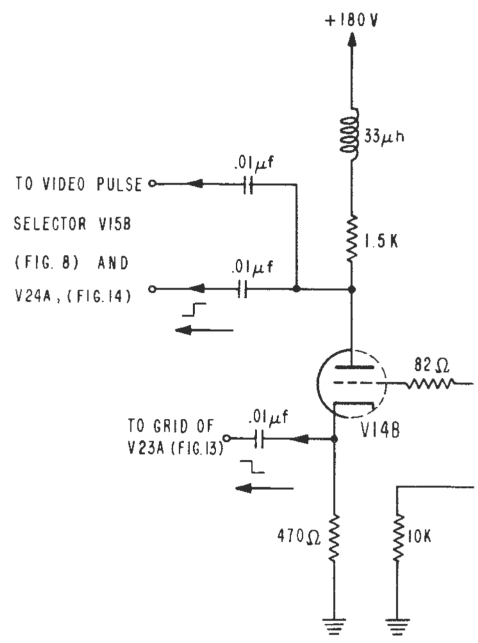

Fig. 6 - Pu

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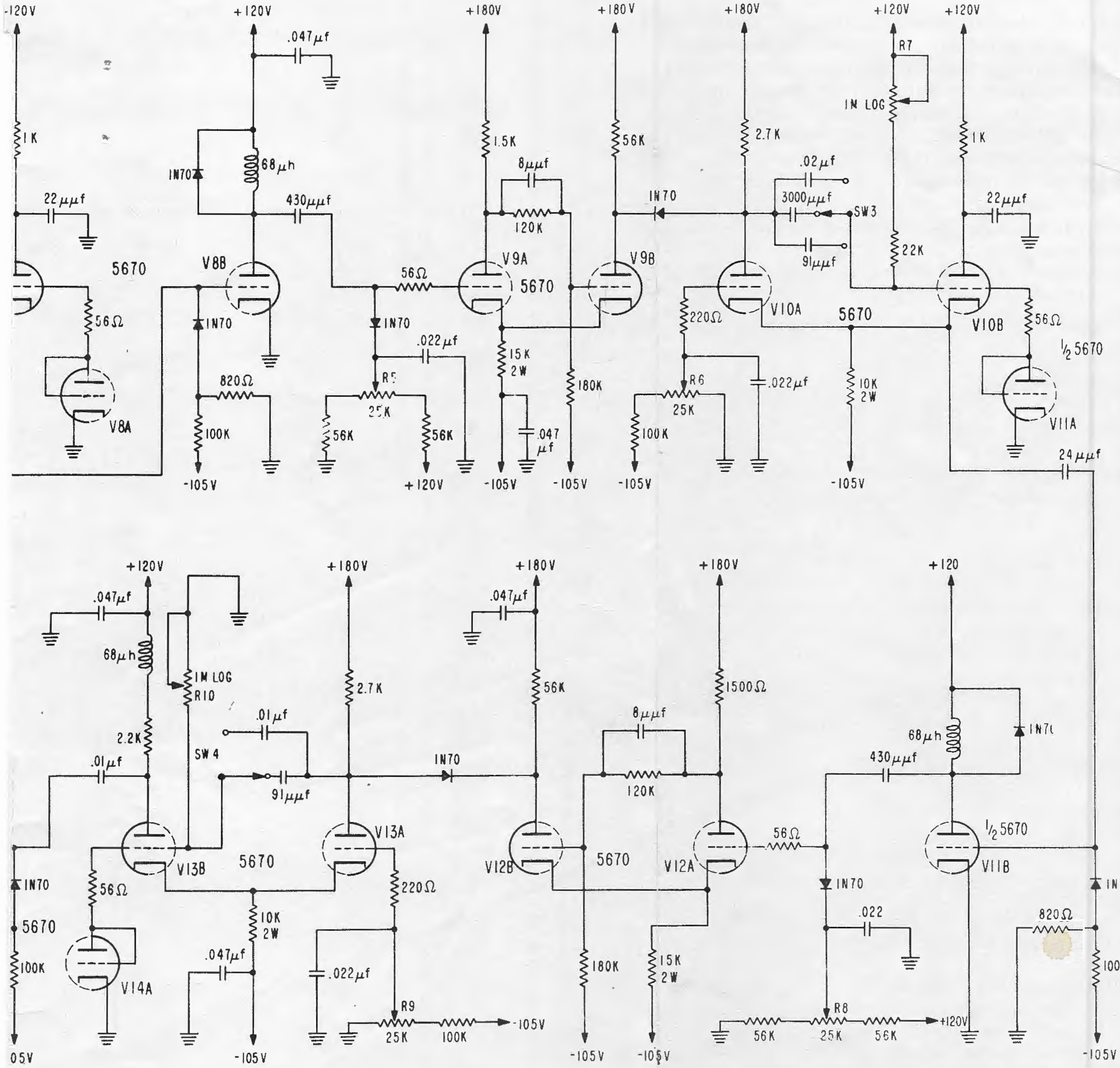

ılse-selection-gate circuit

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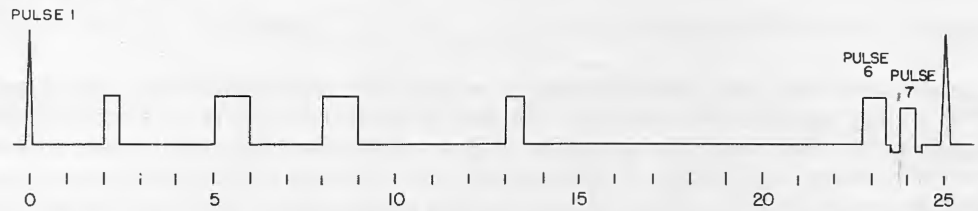

Fig. 7 - Seven-channel pulse system

chosen to get the best rise time and pulse amplitude consistent with tube dissipation ratings. The 25K potentiometer (R3) in the grid circuit of V7A adjusts the minimum gate width and is set so that the gate width produced by this multivibrator is about 0.13 μsec with R7 at its minimum resistance setting. Since the signal for the following stage is derived from the multivibrator cathode, a 22-$\mu\mu$f capacitor was installed to by-pass the plate of V7B to ground and to improve the cathode rise time.

The negative polarity gate on the multivibrator cathode is differentiated and sent to an amplifier (V8B) whose plate current pulse is further differentiated by the damped-ringing-circuit plate load. A negative polarity pulse is thereby produced by the trailing edge of each multivibrator gate. Because the voltage pulse amplitude from a damped ringing circuit falls with increasing pulse repetition rate, and because the gate width of a multivibrator for narrow pulse outputs is a function of the trigger pulse amplitude, a Schmitt circuit (V9A and V9B) is used to produce from the varying-amplitude ringing-circuit pulse a constant amplitude pulse of the same polarity. It is this pulse that triggers the second multivibrator of the cascade delay multivibrator circuit. This Schmitt circuit has a negative voltage cathode return to improve the circuit stability with respect to tube aging. The bias adjustment in the grid circuit of V9A is necessary to allow for the tolerances of the resistor network coupling the plate of V9A to the grid of V9B, the Schmitt-circuit dc threshold levels being quite sensitive to the voltage dividing ratio determined by these resistors. This Schmitt circuit is unique in that the plate of V9B shares the plate load resistor of the following stage V10A. Multivibrator V10A, V10B, and V11A produces the second half of the cascade multivibrator delay gate. This gate is initiated by the negative pulse from the Schmitt circuit V9A and V9B. Since V9B is normally nonconducting, there is no need for a separate plate load resistor. The 56K plate resistor of V9B is used only to return the plate of V9B to the plate supply voltage so that the plate of V10A will not have to recharge the plate capacity of V9 upon the termination of the multivibrator gate. Multivibrator V10A, V10B, and V11A is identical in circuitry and operation to multivibrator V7A, V7B, and V8A. The second multivibrator timing capacitor switch (SW3) is ganged to SW2. The respective timing potentiometers R7 and R4 are likewise ganged so that the gates produced by each half of the combination cascade multivibrator are equal. With a three-position timing-capacitor switch (SW2 and SW3) for the delay multivibrators, a delay range from 0.25 to 10,000 μsec is covered with plenty of overlap between the individual ranges. The negative gate from the cathode of V10B is differentiated, amplified by V11B, again differentiated by the damped-ringing-circuit plate load of V11B, and the negative pulse thus obtained at the termination of the gate is sent to a Schmitt circuit V12A and V12B (identical in circuitry and operation to Schmitt circuit V9A and V9B). Finally, the negative polarity pulse from the Schmitt circuit (V12A and V12B) triggers the pulse-selection-gate multivibrator V13A and V13B.

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Pulse-Selection-Gate Multivibrator

The pulse-selection-gate multivibrator is similar in operation to those described above. The timing potentiometer R10 is returned to ground to provide a somewhat wider range of gate widths than would be available with a positive grid return. Positive grid returns are used with the cascade delay multivibrators because of the increased gate-width stability gained thereby. The timing stability requirements are less stringent in the case of the pulse-selection-gate multivibrator than in the case of the delay multivibrators. The multichannel pulse signal (Fig. 7) illustrates this point. Where it is desired to select the seventh pulse from the train, it can be seen that the delay multivibrator gate must have an absolute stability of approximately $\pm 1/6$ μsec in order to neither include any part of pulse 6 or exclude any part of pulse 7. Percentagewise the $1/6$-μsec drift out of 23.5-μsec total delay time dictates a $100 \times (1/6)/23.5 = 0.7\%$ gate-width stability for the cascade delay multivibrator. However, the width of the pulse selection gate in the illustrated case can vary some ± 0.25 μsec without either cutting off any of pulse 7 or including the following sync pulse. Since the pulse-selection-gate width is about 1 μsec, a stability of only $100 \times 0.25/1 = 25\%$ is required. Cases where the stability requirements for the pulse-selection-gate width would be greater than for the selection-gate delay are unlikely. When a single-pole two-position low-capacity miniature relay is used as the timing-capacitor switch for the pulse-selection-gate multivibrator, a pulse-selection-gate width range from 0.25 to 10,000 μsec is available. The positive-polarity pulse selection gate from the plate of V13B is inverted by amplifier V14B and then sent to the video pulse selector.

VIDEO PULSE SELECTOR

The video pulse selector (Fig. 8) is a switch that permits the raw video signal to pass through it only during the pulse-selection-gate interval. A dual triode (V15A and V15B) has both cathodes tied together to a common cathode resistor which is returned to the negative 105-volt supply. Two 1N70 crystal diodes are tied in parallel between the dual triode common cathode and ground with the diode cathodes grounded. The triode grids are each normally biased a fraction of a volt below ground so that with no signals applied to either grid sufficient triode cathode current is drawn to bring the cathode voltage up to ground level with an excess current of about 12 ma flowing through the crystal diodes. Under normal operation the negative polarity raw video signal from the video amplifier is fed to the grid of V15A, and the negative-polarity pulse selection gate from the plate of V14B is sent to the grid of V15B. The amplitude of the negative-polarity pulse selection gate is greater than that of the peak video signal. The output signal from the video pulse selector is taken from the common triode cathode. In the absence of a pulse selection gate, the negative-polarity video signals drive the grid of V15A farther below ground and lower the plate current in this tube. However, the grid of V15B is biased sufficiently high so that the plate current in V15B alone is enough to hold the common cathode at ground level, and no video signal appears at the output. When the pulse selection gate arrives in the absence of a video pulse, the V15A plate current holds the cathode at the ground level until a video pulse arrives and allows the cathode to drop and follow the grid of V15A as in a standard cathode follower. In a previous circuit a 1N56 low-impedance crystal diode was used in place of the two 1N70's in the present circuit. Trouble was encountered when the negative 105-volt supply voltage was applied before applying the positive supply voltages, because the full 105 volts appeared across the back impedance of the crystal thus burning out the crystals. The 1N70's will stand this voltage, so two are used in parallel to keep the forward impedance low. Provision is made for disabling the selector so that signals having no sync pulse to trigger the sync separator can still pass through the video pulse selector without the benefit of a pulse selection gate. Coupled to the SYNC LEVEL (Fig. 6) control shaft is a double-pole single-throw switch (SW5 and SW6) that is actuated at one

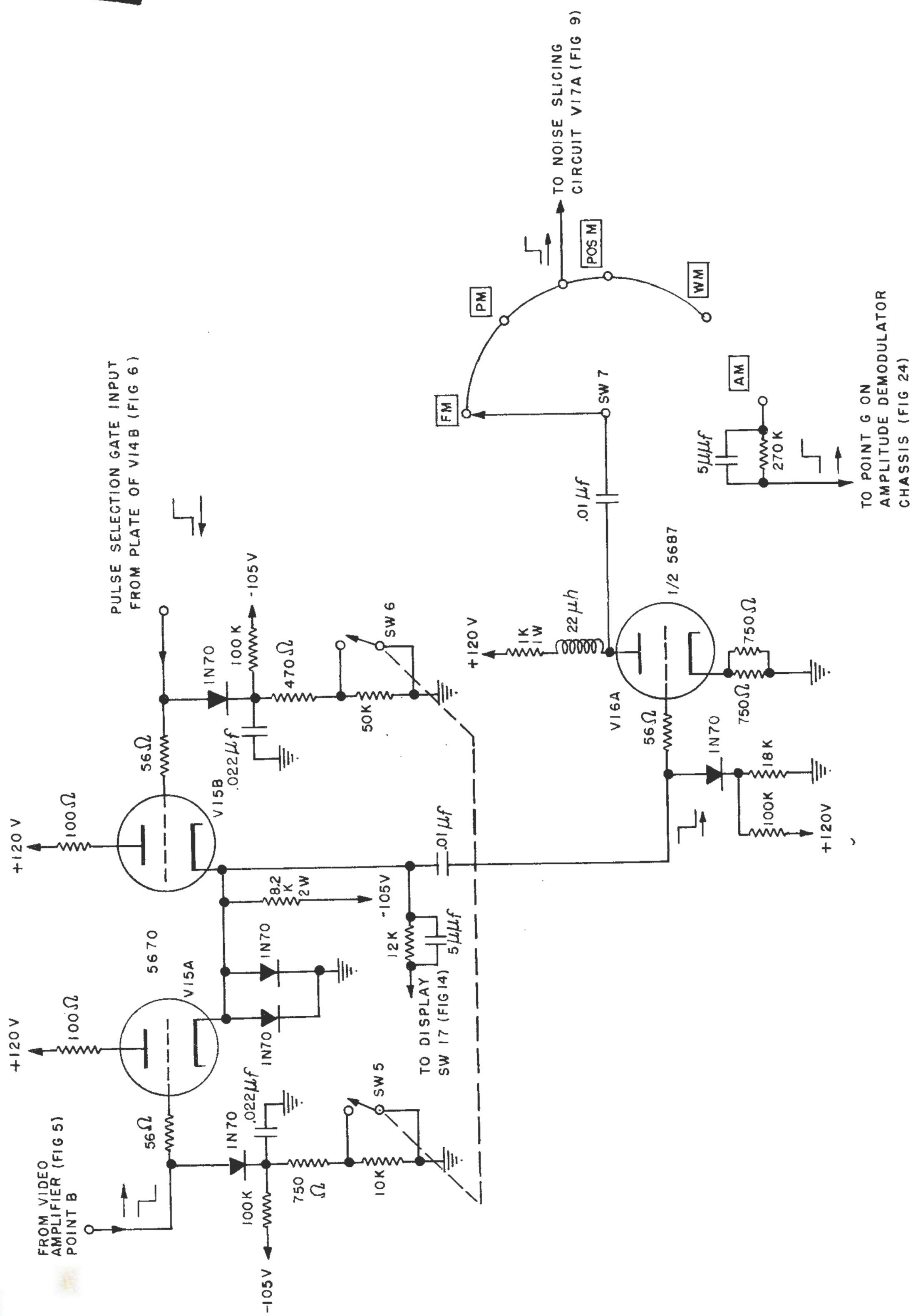

Fig. 8 - Video pulse selector circuit

of the extreme positions of the control shaft. When SW6 is opened, about 35 volts of negative bias is added to that already present on the grid of V15B so that this tube will be completely cut off even during the period when video signals appear on the grid of V15A. Switch SW5 adds about 9 volts of bias to the grid of V15A to bring the cathode down from its clamped position into a region where the plate dissipation of V15A will not be exceeded. With SW5 and SW6 open, all video signals pass undistorted through the video pulse selector.

The output signal from the video pulse selector is coupled to an amplifier (V16A) where it is inverted; the video signals now have a positive polarity. This amplifier utilizes one half of a high perveance 5687 dual triode which accommodates the large grid voltage swing without straying too far from the linear amplification region and at the same time produces a low-impedance 20-volt signal output. The linearity of amplitude-modulated pulses is determined by the distortion present in this amplifier. The amplifier output signal is then sent via a 5-position, single-pole switch (SW7) to the amplitude demodulator in the case of amplitude-modulated pulses and to the noise-slicing circuits for all other types of modulation.

NOISE-SLICING CIRCUIT

The noise slicer shown in Fig. 9 is a Schmitt circuit designed so that the two threshold voltages are separated by a very small voltage interval of about one or two volts. A manual bias control (SLICE POSITION control) sets the voltage on the Schmitt-circuit input grid, thereby controlling the position of the threshold voltage interval (slice) relative to the base line of the incoming pulse. Figure 10a shows a noisy pulse as it comes from the plate of amplifier V16A. Assuming that the Schmitt circuit will commutate whenever the input grid passes both threshold levels in either direction, the bias control (R11) is adjusted so that the input grid to the Schmitt circuit normally rests about 6.5 volts below the lower threshold level, as shown on the voltage scale to the left of the pulse in Fig. 10a. With the base line of the incoming pulse clamped to this voltage, it can be seen from the diagram that the threshold voltage interval "slices" the rise and fall of the pulse at voltage points between which lie no noise pulses of sufficient amplitude to cross both threshold levels. Thus the output signal (Fig. 10b) from the plate of V17B contains no noise of an amplitude-modulated nature, and the width or the repetition rate of the pulse is unchanged except for whatever period or width jitter is present because of noise on the rise and fall of the pulse. The slicer, of course, becomes increasingly less effective as the noise on the pulse approaches and passes the tangential value. For example, when noise peaks become of sufficient amplitude to cross the threshold interval, the output pulse from the Schmitt circuit contains much width jitter, and width modulation information has considerable competition with the noise.

PULSE-STRETCHING CIRCUIT FOR FREQUENCY AND PERIOD MODULATED PULSES

Before frequency or period modulated pulses from the slicing circuit are sent to the time-to-amplitude converter, they are sent to a pulse-stretching circuit where they are converted to pulses of constant widths. This circuit accomplishes two purposes: the first is to eliminate from a frequency or period modulated pulse any width modulation that may be present; the second is to provide pulses of sufficient width to operate the time-to-amplitude converter properly when the pulse width to pulse period ratio of a signal becomes very small. If a pulse, in addition to being frequency or period modulated, is also width modulated and is sent to the time-to-amplitude converter without first removing the width modulation, then the time-to-amplitude converter cannot properly demodulate the information carried by the frequency or period modulation envelope. Also, for the proper demodulation of frequency or period modulated pulses where the pulse width to pulse period ratio is very small, it is necessary to stretch the width of the pulse before it reaches the time-to-amplitude converter.

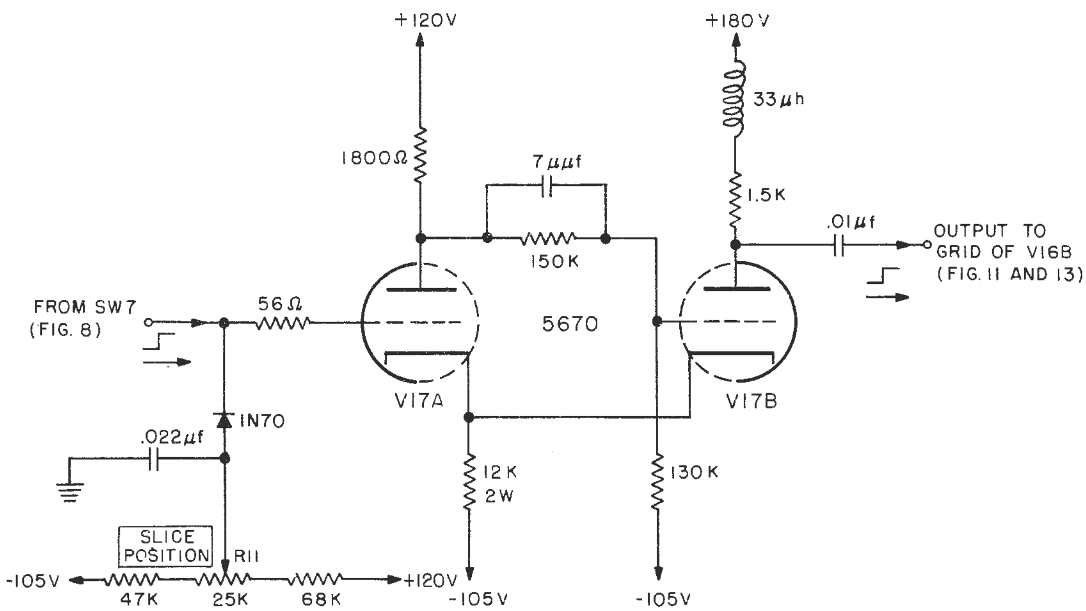

Fig. 9 - Noise-slicing circuit

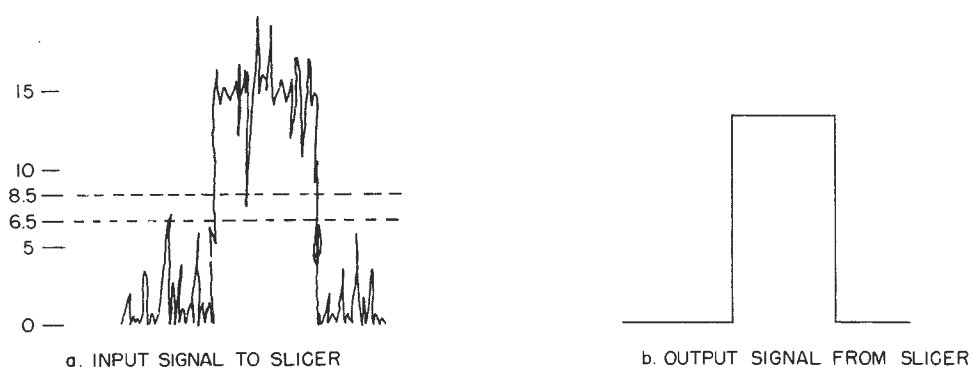

Fig. 10 - Slicer operation with noisy signal

Figure 11 is a diagram of the multivibrator pulse-stretching circuit V18 with its associated differentiation amplifier V16B. The output pulse from the slicing circuit is inverted by amplifier V16B, and the plate-current pulse is then differentiated by the damped-ringing-circuit plate load to produce a negative-polarity 0.1-μsec pulse from the leading edge of each pulse coming from the slicer. This pulse triggers a conventional one-shot multivibrator that produces a gate whose width is determined by whichever of the two timing capacitors is connected in the circuit by switch SW15. The smaller capacitor (24 $\mu\mu$f) gives the multivibrator a 0.15-μsec gate, and the 390-$\mu\mu$f capacitor when shunted across the 24-$\mu\mu$f capacitor produces a 12-μsec gate. These two gate widths are selected by the CMPD RANGE switch, the shorter gate corresponding to the high pulse-repetition-rate range and the longer gate to the lower range.

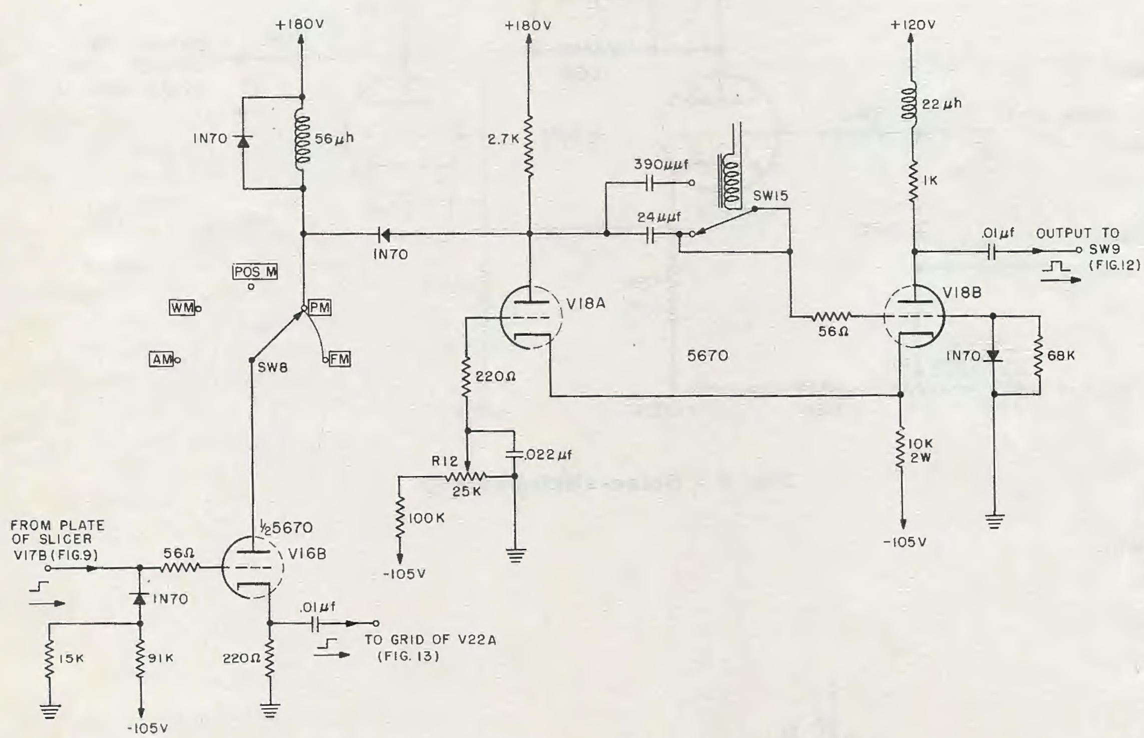

Fig. 11 - Multivibrator pulse-stretching circuit

Amplifier V16B is also used in connection with pulse-width and pulse-position demodulation, and in these connections different plate loads are required from that used in triggering the pulse-stretching multivibrator. These loads are switched by a five-position switch SW8 which is ganged to SW7 (Fig. 8) on the same shaft. The bias on the grid of V18A is set so that a 0.15-μsec pulse is produced when the smallest timing capacitor is in the circuit. The output signal from the plate of V18B is now a positive-polarity pulse of one of two preset widths.

Before being sent to the time-to-amplitude converter, the stretched frequency or period modulated pulse is sliced by another Schmitt circuit V19 (Fig. 12). All signals with the exception of amplitude modulated ones are ultimately sliced by this output Schmitt circuit so that the pulses sent to the time-to-amplitude converter, whether they come from the frequency/period modulation pulse stretcher, the position-to-width converter, or direct from the noise slicing circuit, are then of the same amplitude. Switches SW9, SW10, and SW11 are ganged on the same shaft with SW7 (Fig. 8) and SW8 (Fig. 9). Switch SW9 provides the proper bias for the output Schmitt circuit (V19) for the various input signals. Switches SW10 and SW11 likewise provide the proper biases required by the input circuits of the time-to-amplitude converter.

CIRCUITRY FOR CHANNELING PULSE WIDTH AND PULSE POSITION MODULATED PULSES

Pulse Width Modulation Channel

Upon passing through the slicer and amplifier V16B, width-modulated pulses are fed directly to the output Schmitt circuit where they are sliced and sent to the time-to-amplitude converter. The dotted lines in Fig. 12 show the various switch positions for proper channeling of width-modulated signals from the output of V16B through the output Schmitt circuit. With SW7 set to the pulse width demodulation position, a resistive load is switched in to the plate circuit of V16B which now serves as a straight amplifier inverter for the positive-polarity signals fed to it from the slicer. The negative-polarity signals from the plate of V16B are then sent directly via SW8 and SW9 to the output Schmitt circuit where they are sliced and fed to the time-to-amplitude converter as negative-polarity pulses. In this case the output Schmitt circuit has its input grid biased somewhat above the upper threshold voltage, whereas in the case of frequency- or period-modulated positive-polarity pulses the grid is biased below the lower threshold voltage, thereby allowing the output Schmitt circuit to slice pulses of either polarity and at the same time provide output pulses of constant amplitude independent of polarity. The reason for feeding to the time-to-amplitude converter pulses of different polarities for the different types of modulations is explained on page 33.

Pulse Position Modulation Channel (Position-to-Width Converter)

As stated previously, for the case of position-modulated pulses, the pulse position with respect to the sync pulse must be transformed to a pulse whose width is equal to the time interval between the sync pulse and the modulated pulse itself. This is done because the time-to-amplitude converter will only operate on pulses whose width, frequency, or period is the variable parameter. It will not operate directly on a pulse-position modulated signal. The pulse position-to-width converter with its associated circuits is shown in Fig. 13. The converter itself (V20) consists of a Schmitt circuit modified so that it can be made to commutate in one direction when a negative-polarity pulse is applied to the grid of V20A, and commutate the other way with the application of a negative pulse to the V20A cathode. The sync pulse from the sync-separator circuit (V5), Fig. 6, is sent through a cathode follower (V6A) and is then amplified and inverted by V21A, arriving at the grid of V20A as a negative-polarity pulse. The selected and sliced video pulse (assumed to be position modulated) is fed to the grid of the V16B amplifier. A positive-polarity pulse appears across the 220-ohm cathode resistor and is applied to the grid of a grounded-cathode triode (V22A), the plate of which is connected to the cathode of V20A. The grid of V22A is normally biased

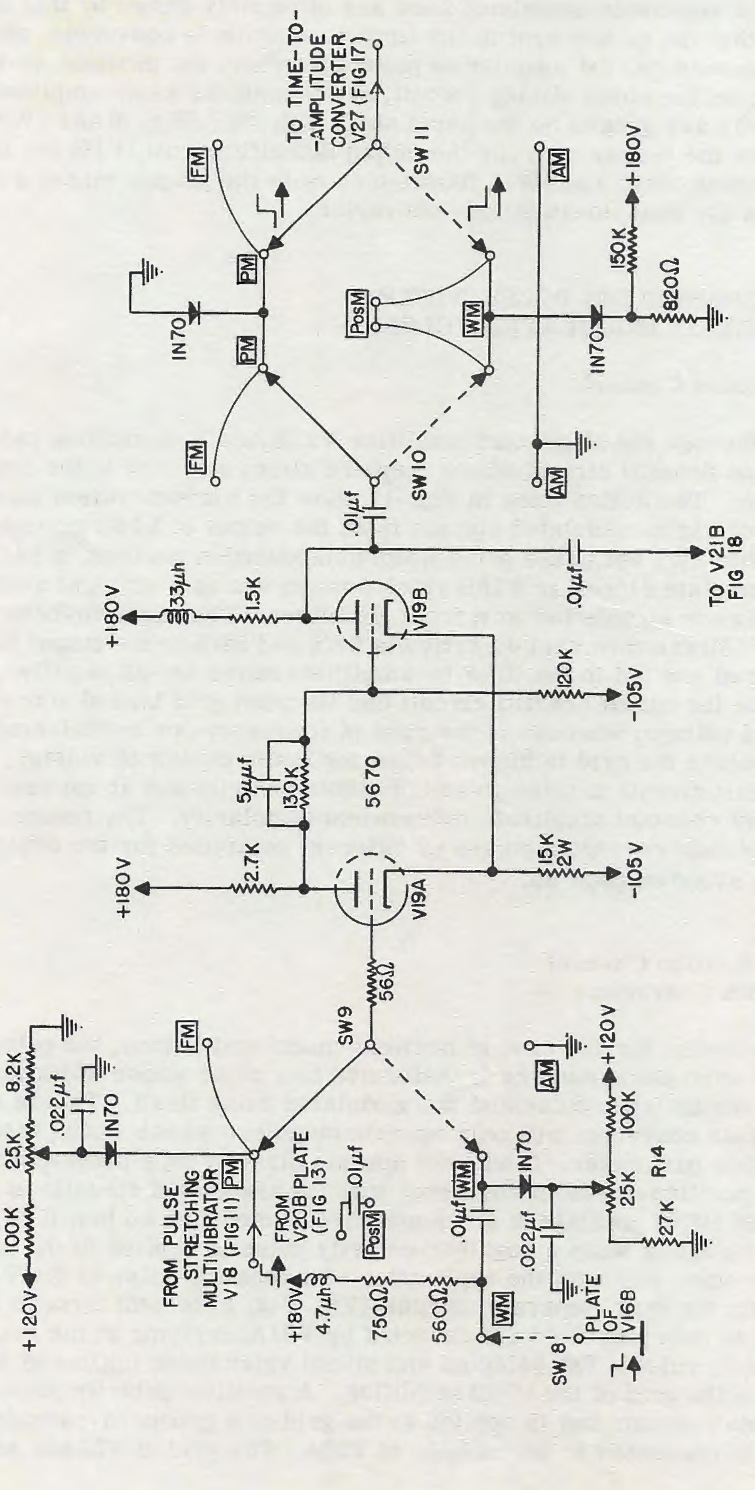

Fig. 12 - Output Schmitt circuit

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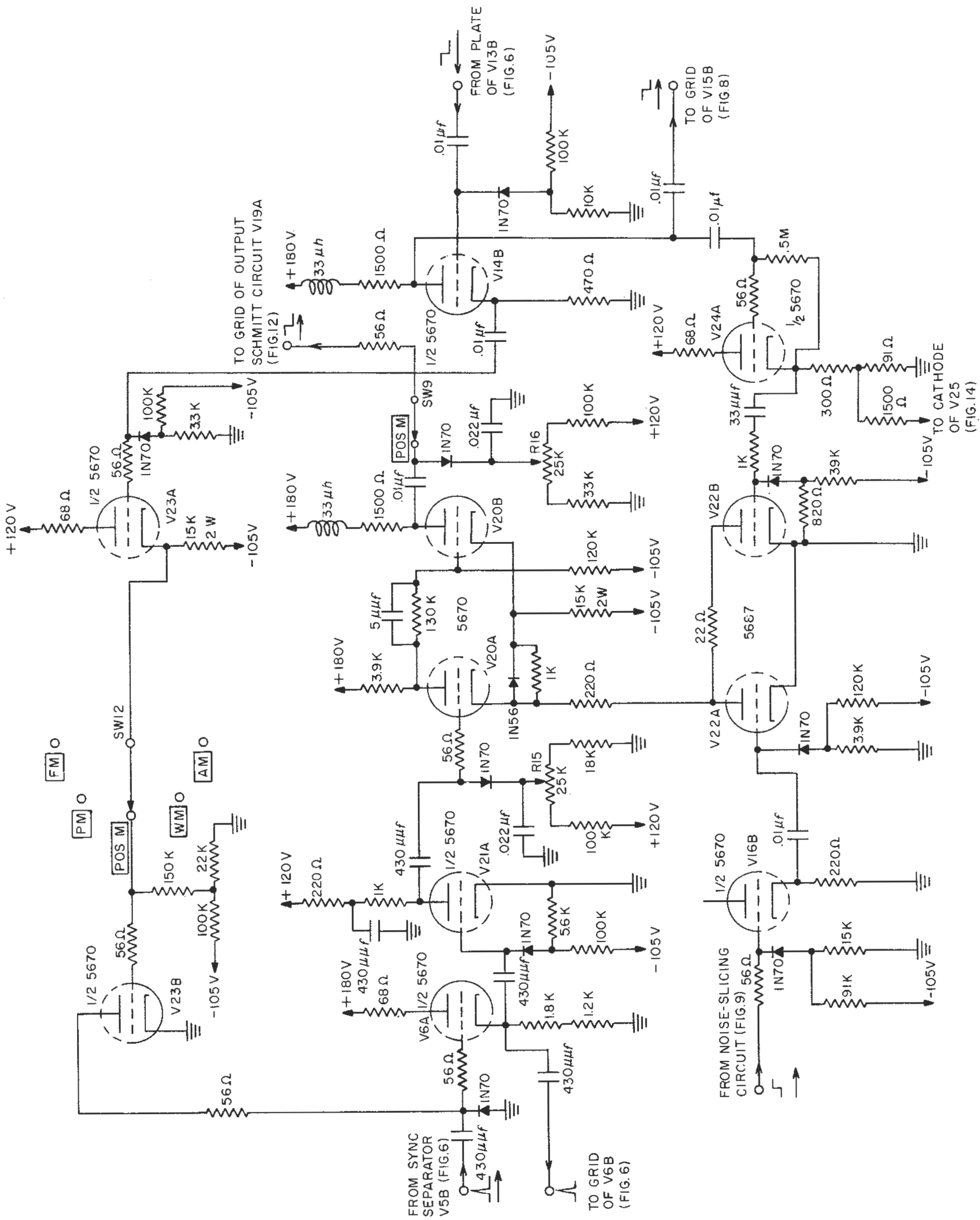

Fig. 13 - Pulse position-to-width converter circuit

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to cutoff. The application of the positive-polarity pulse from the cathode of V16B to this grid causes the plate impedance of V22A to become very low, which essentially grounds the cathode of V20A through this low plate impedance. The negative-polarity sync pulse appears on the grid of V20A, and the negative-polarity information pulse appears on its cathode. This Schmitt circuit (V20) was designed so that the threshold voltages were about 8 volts apart, the lower one being some 10 volts above the ground level. The bias control on the grid of V20A is set so that the grid normally rests about half way between these two threshold levels. When the sync pulse arrives, it drives this grid well down below the lower threshold voltage cutting off V20A and causing V20B to conduct. The circuit remains in this state until the selected information pulse arrives at the grid of V22A. The plate impedance of this tube is suddenly lowered to bring the cathode of V20A close to ground, thus lowering the V20A grid bias enough to make the circuit commutate in the opposite direction. A negative-polarity pulse is produced at the plate of V20B the duration of which equals the time interval between the sync pulse and the information pulse. This negative-polarity pulse is sent via switch SW9 (Fig. 12) to the grid of the output Schmitt circuit (Fig. 12) where it is sliced and sent to the time-to-amplitude converter as a negative-polarity width-modulated pulse. The crystal-resistor combination in the cathode circuit of V20 allows the cathode of V20A to be driven negative without changing appreciably the voltage on the cathode of V20B. If both cathodes were simultaneously driven negative, then the grid of V20B would be unable to reach cutoff and the circuit could not commutate reliably from the pulses applied to its cathodes.

Three problems with respect to noisy signals were encountered in an earlier model of this position-to-width converter. First, if the video signal from the video amplifier was very noisy, it was possible for noise peaks on top of information pulses to be of sufficient amplitude that when added to the amplitude of the information pulse the combined peak voltage of the information pulse plus noise pulse equaled or exceeded that of the sync pulse. This situation resulted in an additional trigger pulse output from the sync separator and a selection gate being generated from each such extraneous noise pulse that met the above specifications. Thus pulses other than the desired information pulses were included by an unwanted selection gate.

Secondly, if a noisy information pulse produced an unwanted trigger pulse, then both the unwanted trigger pulse and the information pulse would arrive simultaneously at the position-to-width converter with the negative-polarity trigger pulse being applied to the (V20A) grid and the negative-polarity information pulse being applied to the cathode. If by chance these two pulses were of nearly the same width, there would be hardly any net grid-to-cathode voltage change, and the Schmitt circuit would not commutate at all.

Thirdly, trouble was met if the noise on an occasional information pulse was of such a polarity and amplitude that the information pulse amplitude, when added to the noise amplitude, produced a pulse of insufficient amplitude to make the noise-slicing circuit (V17) (Fig. 9) commutate. Consequently, no information pulse reached the position-to-width converter.

The first two problems were solved by the addition of a lockout circuit to prevent any extraneous trigger pulses generated during the pulse-selection-gate interval from either triggering off the pulse-selection-gate delay multivibrator or triggering the input grid of the position-to-width converter. The pulse selection gate from the cathode of V14B (Fig. 13) drives a cathode follower V23A which is biased so that its cathode normally rests about one volt below ground. This cathode is connected directly through switch SW12 (ganged to SW7, SW8, SW9, SW10, and SW11) to the grid of triode V23B. The cathode of V23B is grounded, and the plate is connected directly to the grid of cathode follower V6A.

Throughout the duration of the pulse selection gate the grid of V6A is shorted to ground through the low plate impedance of V23B whose grid is held hard positive by cathode follower V23A. This lockout prevents any extraneous trigger pulses from getting through cathode follower V6A to either the pulse-selection-gate delay multivibrator or the trigger pulse amplifier V21A during the pulse selection gate.

The second and third problems mentioned above are both situations where the position-to-width Schmitt circuit (V20) is turned on by a trigger pulse and then for one of the two mentioned reasons is not turned off. Either two pulses arrive simultaneously at the position-to-width converter or no information pulse arrives. In either of these two cases the position-to-width converter produces a pulse of width equal to the pulse period interval plus the interval between sync pulse and information pulse. Ordinarily this occasional "long" pulse would be several times longer than the interval between the sync pulse and information pulse. Because of the nature of the automatic "tuning" circuitry in the time-to-amplitude converter, this occasional "long" pulse is undesirable (page 32); it resulted in a complete blanking of the time-to-amplitude converter for about 3/4 second. One of these extraneous pulses per second makes the output signal-to-noise ratio of the CMPD approach zero.

When an occasional information pulse from the noise slicer is actually missing, a circuit comprising V24A and V22B automatically provides a pulse that turns the position-to-width multivibrator off at the termination of the pulse selection gate. Therefore, V20 is prevented from producing a pulse that is long enough to blank the time-to-amplitude converter. The negative-polarity selection gate from the cathode of cathode follower V24A is sent via a differentiating circuit to the normally cutoff grid of triode V22B. This triode is shunted from the cathode of V20A to ground in exactly the same way as was V22A. At the grid of V22B there appears a negative-polarity voltage spike at the initiation of the pulse selection gate and a positive-polarity one at the termination. Since the negative spike simply adds more bias to the already cutoff grid of V22B, nothing happens. However, the positive-polarity spike turns V22B hard on with a resulting negative-polarity pulse appearing on the cathode of V20A. If no information pulse has come through the noise slicer during the pulse selection gate, then the positive voltage spike produced by the trailing edge of the pulse selection gate will turn off the position-to-width Schmitt circuit (V20). The pulse thus produced by V20 will appear as a discontinuity in the final output envelope of the CMPD.

DISPLAY CIRCUITS

Provision is made to display at will on an AN/SLA-2 or AN/APA-74 type indicator the raw video signal as it comes from the video amplifier, the selected video signal as it leaves the video pulse selector, or the sliced video pulse from the noise slicer. Figure 14 shows the various circuits connected with channeling these signals to the indicator. Switch SW17 selects the negative-polarity raw, selected, or sliced video for display and feeds it to the grid of amplifier V24B. The positive-polarity signals from the plate of V24B are then sent to one of three inputs to an addition circuit. This addition circuit consists of a dual triode (V25A and V25B) with cathodes connected to a common cathode resistor. Input signals to this circuit are connected to each of the two grids and to the bottom end of the cathode resistor. The algebraic sum of these three signals appears at the common cathode connection of V25A and V25B. The signal from V24B is connected to the grid of V25A. The trigger pulse output from the sync separator is sent via cathode follower V6A to the grid of V25B, and the negative-polarity selection gate is sent through a divider network in the cathode circuit of V24A to the bottom of the cathode resistor

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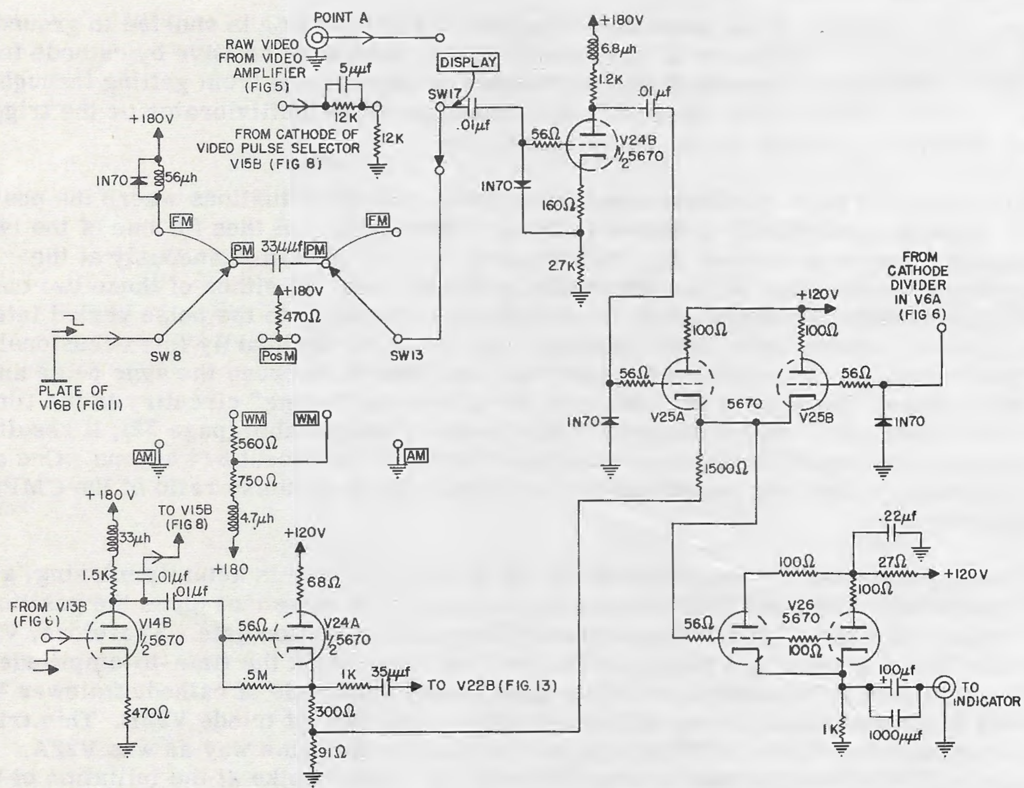

Fig. 14 - Display circuits

for V25. The V25 cathode signal, which is the sum of these three input signals, is sent through a low output impedance cathode follower (V26) to the indicator input.

With the display selector switch (SW17) set so that the raw video signal is sent via V24B to the grid of V25A, the display on the crt of the indicator will show not only the raw video signal, but the pulse selection gate as well. The pulse selection gate appears as a low-amplitude negative-polarity gate that when properly adjusted will include the pulse from which modulation information is desired. Figure 3c shows a display in which the pulse selection gate has been adjusted to include the last pulse in the three-pulse train. The sync pulse will appear in the display to be of greater amplitude than it actually is because the addition of the sync-separator output trigger pulse with that of the raw video sync pulse produces an artificial "sync" pulse of greater amplitude than the information pulses. This arrangement was made so that the operator of the CMPD can tell at a glance which pulse triggers the sync separator by simply observing which video pulse in the display "jumps" in amplitude as the SYNC LEVEL control is advanced. Also, the indicator sweeps will be assured of firing on the sync pulse if it is of considerably greater amplitude than all the others.

With switch SW17 connected to the output of the video pulse selector, only the selected video pulse is sent to the grid of V25A. The indicator display in this case will show the selected video pulse, the pulse selection gate, and the trigger pulse from the output of the sync separator. The trigger pulse in this case is fed only to the grid of V25B with the

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result that the trigger pulse sent to the indicator will be of such low amplitude that the display sweeps will trigger on the selected video information pulse rather than on the trigger pulse. This is not inconvenient since it is the selected video pulse that the operator desires to see when he sets switch SW17 to that position.

With switch SW17 in the third position, the grid of V24B is connected through the proper voltage divider networks to the output of the sliced-pulse amplifier V16B. The indicator then displays the sliced pulse, the pulse selection gate, and the trigger pulse. As in the case of the selected video pulse display the trigger pulse appears to be of lower amplitude than the sliced information pulse. Switch SW13 (ganged to SW7, SW8, SW9, SW10, SW11, and SW12) connects the grid of V24B through the proper voltage dividers to the various plate loads of V16B. These voltage dividers are necessary in order that all signals on the indicator crt display will be of approximately the same amplitude, thus eliminating the need for adjustment of the indicator gain control.

TIME-TO-AMPLITUDE CONVERTER

The time-to-amplitude converter (Fig. 15) takes the frequency, period, or width modulated pulses from the output of Schmitt circuit V19 (Fig. 12) and produces pulses whose amplitudes vary linearly with either the pulse period or the pulse width, whichever is chosen. For the purposes of lucid explanation this converter will be discussed with respect to period-modulated pulses alone. It will then be shown how with a simple switching circuit the same circuitry may be used with equal success to perform the same operation with respect to width-modulated pulses. No switching whatever is involved in the time-to-amplitude converter in changing from pulse period modulation to pulse frequency modulation. The converter produces a pulse whose amplitude is proportional to the pulse period in both cases.

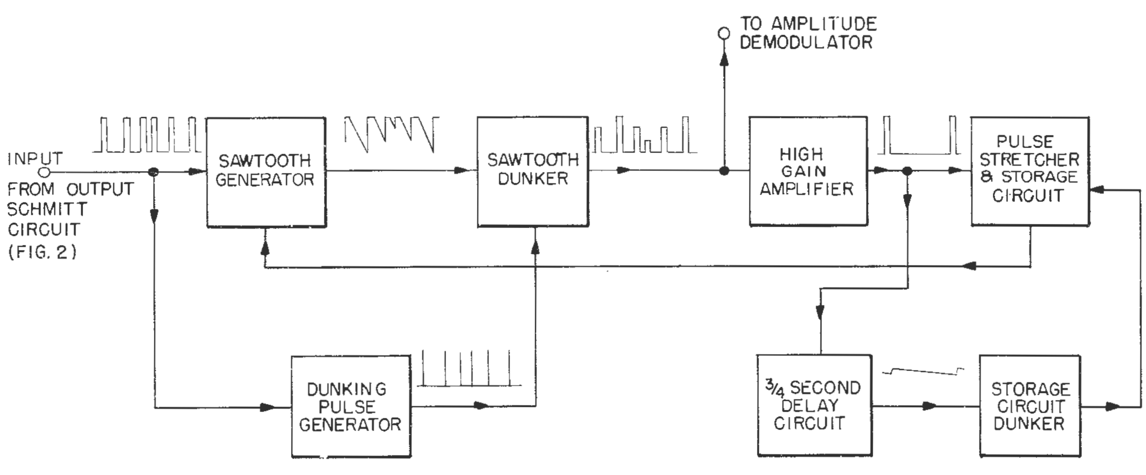

Fig. 15 - Block diagram of time-to-amplitude converter

The period-modulated positive-polarity pulses charge a capacitor which in turn discharges through a constant-current pentode producing a linear, constant-slope, negative-polarity sawtooth voltage waveform whose duration is equal to the entering pulse period.

If this period is modulated (Fig. 16a), the sawtooth amplitude will be likewise modulated (Fig. 16b). At the termination of each pulse entering the converter, a 0.1-μsec dunking pulse is produced by the dunking pulse generator. This pulse is sent, along with the sawtooth voltage, to the sawtooth dunking circuit where the negative-polarity amplitude-modulated sawtooth voltage is transformed to a positive-polarity amplitude-modulated pulse of width equal to that of the pulse entering the time-to-amplitude converter. (This amplitude-modulated pulse is sent to the amplitude demodulator for final demodulation.) Next, the amplitude-modulated pulse is sent to a high-gain amplifier that is adjusted so that only those pulses having amplitudes exceeding a certain threshold value are amplified. The output pulses from this amplifier are rectified, stretched, and filtered, and the peak value of the pulse voltage is stored by a 2-μf capacitor in an RC circuit having a time constant of approximately three minutes. This voltage controls the bias on the grid of the pentode discharge tube in the sawtooth generator. If at any time the amplitude of the pulses entering the amplifier exceeds a certain threshold value, the voltage across the 2-μf capacitor in the storage circuit is increased. This increase in voltage causes an increase in the negative grid bias of the sawtooth discharge pentode, and the pentode plate impedance (sawtooth discharge path) is raised. The slope (dE/dt) of the sawtooth voltage waveform becomes less, thus lowering the peak amplitude. By the use of this negative feedback loop, the peak amplitude of the pulses entering the high-gain amplifier is kept at a constant value. Because of the three-minute time constant in the storage circuit, the bias on the discharge pentode remains at a sufficiently constant value to permit proper demodulation of pulse signals having repetition rates as low as 10 cycles per second. At pulse repetition rates below 5 cycles per second there is a noticeable change in sawtooth slope that takes place during the interval between pulses, indicating discharge of the 2-μf storage capacitor between pulses and resulting in distortion.

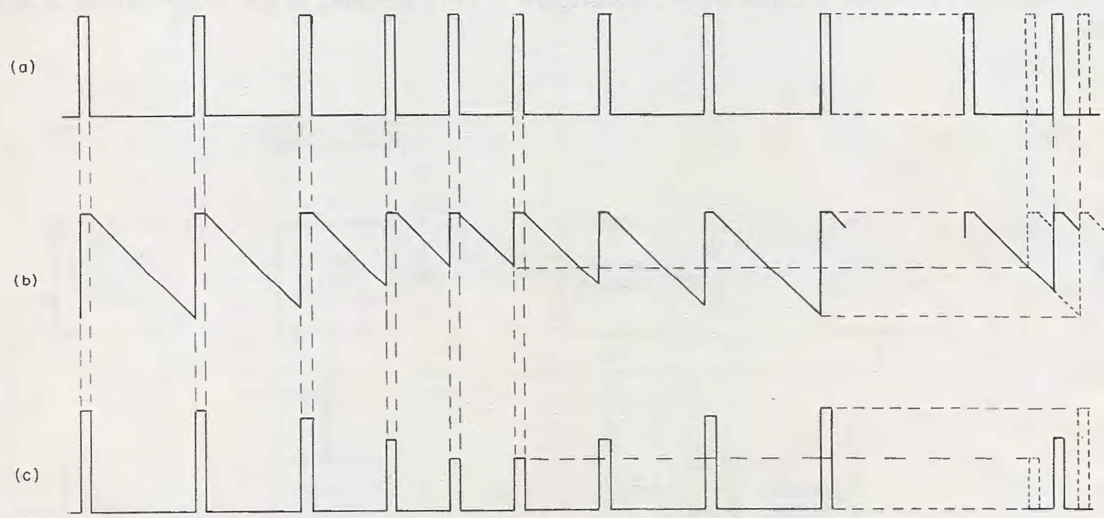

Fig. 16 - Time-to-amplitude-converter waveforms for period-modulated pulses

This negative feedback system is unique in that the peak value of the sawtooth voltage, and consequently that of the amplitude-modulated pulses appearing at the input to the amplifier, can be kept constant over a very wide range of pulse repetition rates. For high repetition rates the bias on the pentode discharge tube is low, permitting high current discharge of the sawtooth capacitor. At low repetition rates the opposite is true since the high bias forces the sawtooth capacitor to discharge slowly.

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In a previous model it was found that if the converter were suddenly switched from a low repetition rate to a high one, it took one or two minutes for the 2-μf storage capacitor to discharge to a voltage that provided the proper bias for the pentode discharge tube consistent with proper operation at the new repetition rate. In switching from a high repetition rate signal to a low one, the problem disappears; the 2-μf storage capacitor is being charged by a low output impedance cathode follower, and the process is almost instantaneous. A circuit was developed that would instantaneously discharge the 2-μf storage capacitor in the event that in the preceding 3/4 second no pulse signal from the sawtooth dunker was of sufficient amplitude to reach the high-gain amplifier threshold voltage. Thus, the storage capacitor is always in a discharged state within one second after the signal output drops below the amplifier threshold voltage value, and the sawtooth discharge pentode plate impedance is at its minimum value. Upon application of a pulse input signal the pentode will discharge the sawtooth capacitor very rapidly, producing a high-amplitude sawtooth and a high-amplitude output signal from the sawtooth dunker. The consequently high-level signal output from the amplifier is rectified and charges the 2-μf storage capacitor to the voltage that provides the proper bias for the pentode discharge tube so that the peak sawtooth voltage just reaches the amplifier threshold voltage. Figure 15 shows the block diagram of the circuitry added to provide delayed dunking for the 2-μf storage capacitor. The output from the high-gain amplifier is rectified and used to charge an RC circuit having a time constant of about 2.5 seconds. The voltage across the capacitor of this RC circuit controls the bias on a dunking tube that is shunted across the 2-μf storage capacitor. If after a time lapse of 3/4 second no pulses have been rectified, the voltage on the normally cutoff grid of the storage-capacitor dunking tube rises from cutoff to cathode level making the tube conduct and discharging the 2-μf storage capacitor. As soon as a signal appears at the output of the amplifier, the 3/4-second delay capacitor is recharged and the 2-μf storage-capacitor dunking tube is cut off until such time as signals are missing for a period of 3/4 second or more.

Sawtooth Generator and Sawtooth Dunking Circuits

The positive-polarity period-modulated pulses are sent from output Schmitt circuit V19 (Fig. 12) through a low-output-impedance cathode follower V27 (Fig. 17) to the plate of diode V28A. A 68-$\mu\mu$f capacitor is connected between the cathode of this diode and ground. To the ungrounded side of this capacitor is also connected the plate of the 6AN5 (V29) pentode discharge tube. Each positive pulse entering the cathode follower charges the 68-$\mu\mu$f capacitor through the low impedance of diode V28A to the peak voltage of the pulse. After the termination of the pulse, the cathode of the charging diode is free to drop, and the 68-$\mu\mu$f capacitor begins a constant current discharge through the plate impedance of the pentode which lasts until the next pulse enters cathode follower V27 and recharges the capacitor. The plate of the discharge pentode is also directly coupled to the grid of cathode follower V30A, providing a low impedance output for the sawtooth voltage waveform.

A 100-$\mu\mu$f capacitor connects the cathode of V30A to both the cathode of a grounded-plate clamping diode (V28B) and the plate of a grounded-cathode dunking triode (V31A). At the termination of each pulse entering cathode follower V27, a 0.1-μsec positive-polarity dunking pulse is produced by the dunking pulse generator (Fig. 15). This pulse arrives at the normally cutoff grid of the dunking triode V31A just as the linear discharge of current through V29 begins. The 100-$\mu\mu$f capacitor coupling the cathode of V30A to the plate of the dunking tube is "instantly" discharged with the result that the plate of the dunking tube and the cathode of the clamping diode are now resting at ground potential. As the negative-going sawtooth voltage on the grid of V30A continues to drop, the 100-$\mu\mu$f capacitor charges through the 12K cathode resistor of the cathode follower and the forward plate impedance of the clamping diode. This clamping diode prevents the signal on the

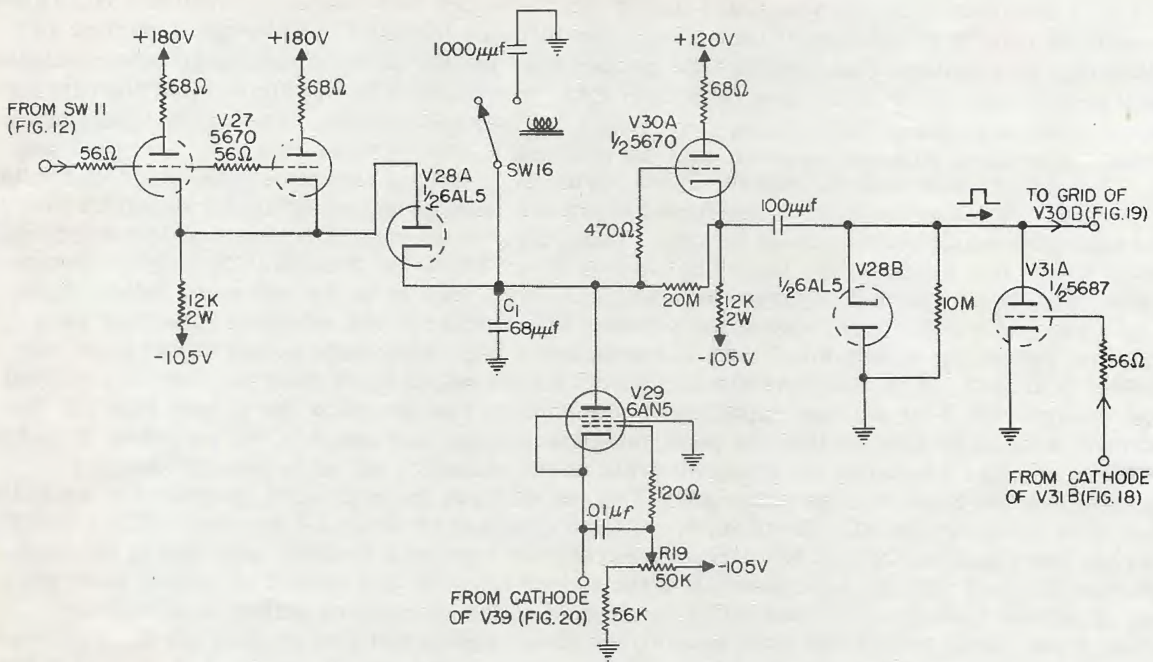

Fig. 17 - Sawtooth generator and sawtooth dunking circuits

plate of the dunking tube from going below ground level. When the next pulse enters cathode follower V27, the sawtooth is terminated, and the cathode of cathode follower V30A is suddenly raised an amount equal to the amplitude of the sawtooth waveform. This voltage rise is coupled through the 100-$\mu\mu$f capacitor to the plate of the dunking tube, and upon termination of the pulse entering V27 this voltage is again dunked to ground level to produce a pulse whose amplitude is equal to that of the sawtooth waveform and directly proportional to the time interval between the information pulses entering V27.

The grid of cathode follower V27 is clamped to ground (Fig. 12) when receiving period-modulated pulses. This places the cathode of V27, the plate of V28A, the cathode of V28A, and the plate of V29 above ground. The grid of the pentode discharge tube is connected to an adjustable negative voltage (R19), and the cathode of the pentode is connected to a negative voltage source whose potential is controlled by the voltage across the 2-$\mu\mu$f storage capacitor in the storage circuits. By controlling the pentode cathode-voltage, the relative grid-to-cathode voltage is changed and the plate current through the pentode is thereby controlled. The cathode resistors of cathode followers V27 and V30A are returned to the negative 105-volt supply to provide good linearity and wide dynamic range for the pulse signals.

As the pulse repetition rate entering this circuit is lowered, the feedback circuits appropriately raise the negative bias on the discharge pentode so that the peak sawtooth amplitude will remain constant with increasing sawtooth width. When the plate impedance of the pentode reaches such a high value that the various leakage resistances across the 68-$\mu\mu$f capacitor become a deciding factor in determining the value of capacitor discharge current, the sawtooth waveform no longer retains its constant slope but becomes more and more exponential in shape as the pulse repetition rate is lowered. In order that the plate impedance of the discharge pentode will never have to reach such an extreme value, a two-position switch (SW16) is provided to add a fixed capacitance across the 68-$\mu\mu$f capacitor for the proper demodulation of pulses of low repetition rates. Each position of this switch corresponds to one of the two overlapping bands of pulse repetition rates, each of which can be successfully covered by the plate impedance range of the discharge pentode. This switch is ganged to SW15 (Fig. 11). The combination switch consisting of SW15 and SW16 comprises the RANGE switch and selects the range within which the particular pulse repetition rate falls. The pulse stretcher previously described (Fig. 11) simply stretches the period-modulated information pulses to a width sufficient to permit complete charging of the 1000-$\mu\mu$f capacitor shunted across the 68-$\mu\mu$f capacitor with the RANGE switch set to the LOW repetition rate range.

The main leakage path shunting the 68-$\mu\mu$f capacitor is found to exist between the heater and cathode of the 6AL5 (V28A) charging diode. Because of troubles with 60-cycle pickup through the heater-to-cathode impedance, the heater of V28A and V28B is run from a 6-volt dc supply. Because filtered dc supplies are difficult to insulate, it was impossible to completely isolate the cathode of V28A from ground. A 20-megohm resistor was installed between the plate of the discharge pentode and the cathode of cathode follower V30A. This resistor appears to the plate of the pentode to be a constant current source, and it supplies to the capacitor shunting the pentode more than enough current to cancel out that which is flowing out of the capacitor through the cathode-to-heater leakage resistance of diode V28A. The small extra current supplied by the 20-megohm resistor will not seriously distort the linearity of the sawtooth waveform because the resistor with its cathode-follower termination is essentially a constant current source. Any extra current that tends to raise the voltage on the capacitor will be taken care of automatically by an equal constant current flowing out through the pentode.

Dunking Pulse Generator

The positive-polarity period-modulated pulse at the input of cathode follower V27 is also sent to the grid of a split-load phase inverter V21B (Fig. 18). The negative-polarity pulse from the plate of V21B is then sent through switch SW14 to the grid of Schmitt circuit V32A and V32B. (The purpose of the delay line for the cathode load of V21B will be explained later in connection with the demodulation of width-modulated pulses.) Switch SW14 is ganged together with switches SW7 through SW13. The input grid of Schmitt circuit V32 is biased by R17 above the upper Schmitt threshold level so that V32A is normally conducting. The incoming negative-polarity pulse then forces the Schmitt circuit to commutate producing a positive-polarity current pulse through V32B. The V32B plate load is a damped ringing circuit that differentiates this plate current pulse and produces a positive-polarity 0.1-μsec voltage pulse at the termination of each negative-polarity pulse entering the Schmitt circuit. This pulse is sent to triode V33A where it is amplified and inverted. The output pulse from this inverter varies in amplitude with pulse repetition rate, and there still remain on the base line small ringing transients that were not completely damped out by the crystal in the ringing circuit plate load of V32B. This signal is sent to a unique, one-tube slicer V33B that slices both the top and the base line from the pulse and at the same time provides a constant-amplitude positive-polarity output pulse. The cathode of this amplifier, normally resting several volts above ground, is returned through a resistor to the negative 105-volt supply. A crystal diode connected between the cathode and ground prevents the cathode from going below ground. This amplifier is highly degenerative when the tube is drawing current through the cathode resistor; therefore, the grid-to-plate voltage gain will be very small. This will remain the case until the grid signal

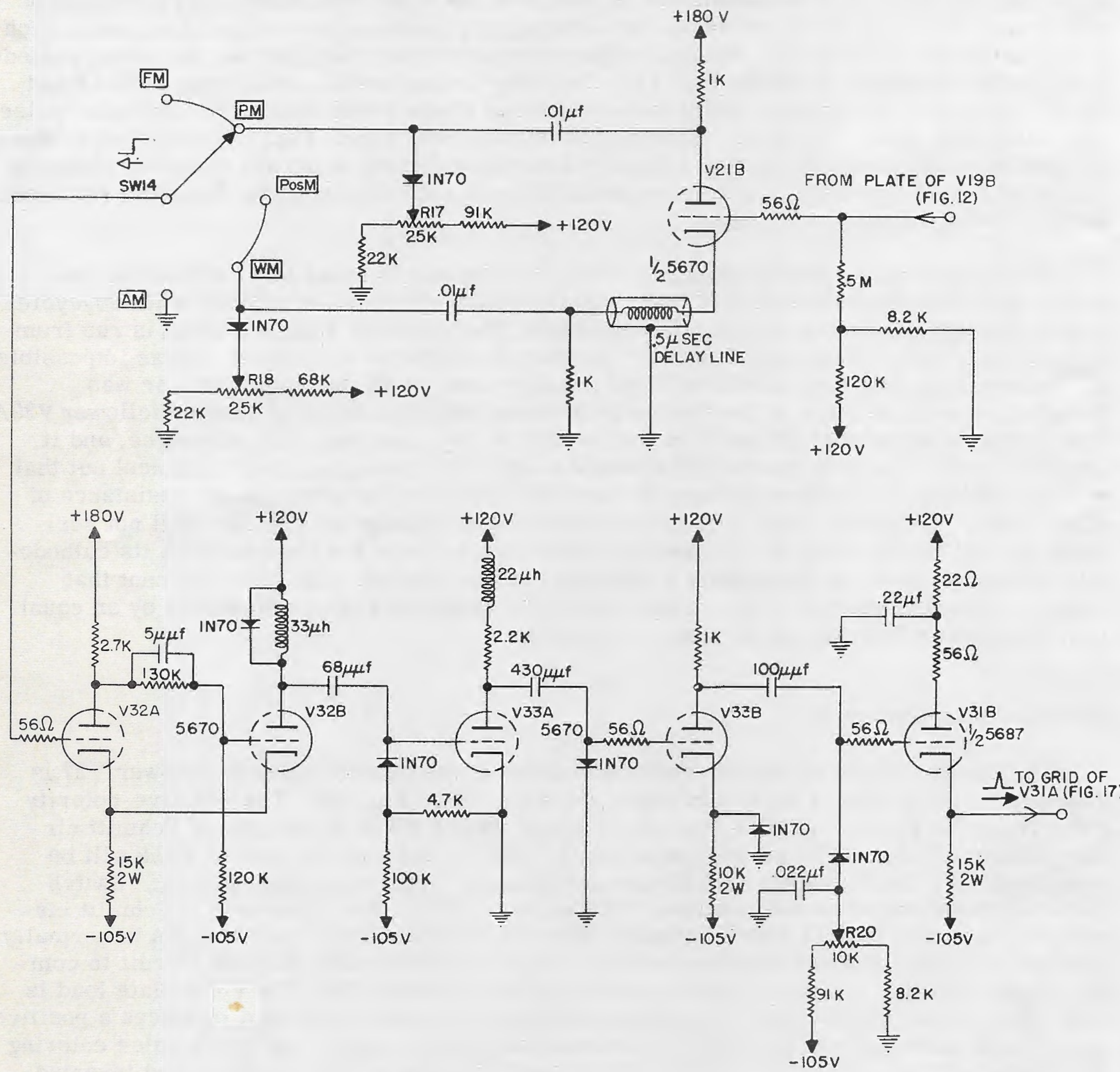

Fig. 18 - Dunking-pulse-generator circuit

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has driven the cathode down to ground level where the cathode current is suddenly shifted from the high-impedance cathode resistor to the low-impedance crystal diode. The grid-to-plate voltage gain jumps with the sudden loss of cathode degeneration, and the gain remains high until the grid approaches cutoff where the gain again reaches a low value. In this way only a slice out of the middle of the pulse is amplified by the full potential gain of the tube, since the base line of the pulse occurs in a very low gain region and the peak of the pulse drives the tube to cutoff. The output voltage from the plate of this amplifier is coupled to the grid of cathode follower V31B which provides the necessary low impedance to drive the positive-going grid of the dunking triode V31A (Fig. 17). The output of this cathode follower is dc coupled to the dunking tube grid, and the bias on this grid is determined by the bias setting (R20, Fig. 18) on the grid of the cathode follower. The V31B cathode resistor is returned to the negative 105-volt supply so that the grid of the dunking tube can be biased below ground.

High-Gain Feedback Amplifier

The amplitude-modulated pulse that appears on the plate of dunking tube V31A (Fig. 17) is fed to a split-load phase inverter (V30B) (Fig. 19). The cathode of V30B is the low-impedance output terminal through which the signal is sent to the amplitude demodulator. The negative-polarity amplitude-modulated signal from the plate of V30B is amplified by the triode phase inverter V34A and the resulting positive-polarity signal is coupled to the normally cutoff grid of another amplifier V34B. The bias on this grid is adjusted by setting R21 so that only signals whose amplitudes exceed a certain value will cause the tube to conduct; all smaller amplitude signals are in the grid cutoff region. The negative-polarity output signal from the plate of V34B (representing only those pulses whose amplitudes exceed the value of the bias set by R21) is then amplified by V35A.

Because of the high amplitude of the positive-polarity signals on the grid of V34B, there is considerable differentiation of these pulses through the grid-plate capacity of the tube, even in the absence of pulses having sufficient amplitude to make the tube conduct. Consequently, narrow positive- and negative-polarity voltage spikes appear at the plate of V34B as well as the negative-polarity pulses that represent actual conduction of V34B. These extraneous low-amplitude (less than one volt) spikes are prevented from reaching V35A by the network of three crystal diodes in the grid circuit. The grid of V35A is clamped to ground by diode 1 to help prevent the grid from making positive excursions. The signal from the plate of V34B is fed to the V35A grid through a high back-impedance diode (2) connected so that conduction occurs only with the application of negative-polarity pulses. Furthermore, a positive cathode-to-plate bias of about one volt is maintained on diode 2 by diode 3. Thus positive-polarity pulses appear across the low forward impedance of diode 3 and are prevented from reaching the grid of V35A by the high back impedance of diode 2. The positive-polarity signals that do manage to get through diode 2 are shunted to ground by diode 1. Because of the positive cathode-to-plate bias on diode 2, no negative-polarity signals will appear at the grid of V35A until the signal reaches an amplitude of at least one volt. This value is rarely exceeded by the amplitudes of the unwanted voltage spikes. The positive-polarity signals from the plate of amplifier V35A are sent to the grid of split-load phase inverter V35B, the cathode of which provides the low output impedance required to drive the pulse-stretching circuits.

The large RC filter networks in each amplifier-tube plate circuit act as combination 120-cycle filter and amplifier decoupling networks. It is most important that a minimum of 120 cycles power supply ripple voltage get through the amplifier since the peak output voltage from the amplifier determines the bias on the discharge pentode (V29, Fig. 17) and hence the slope of the sawtooth discharge waveform.

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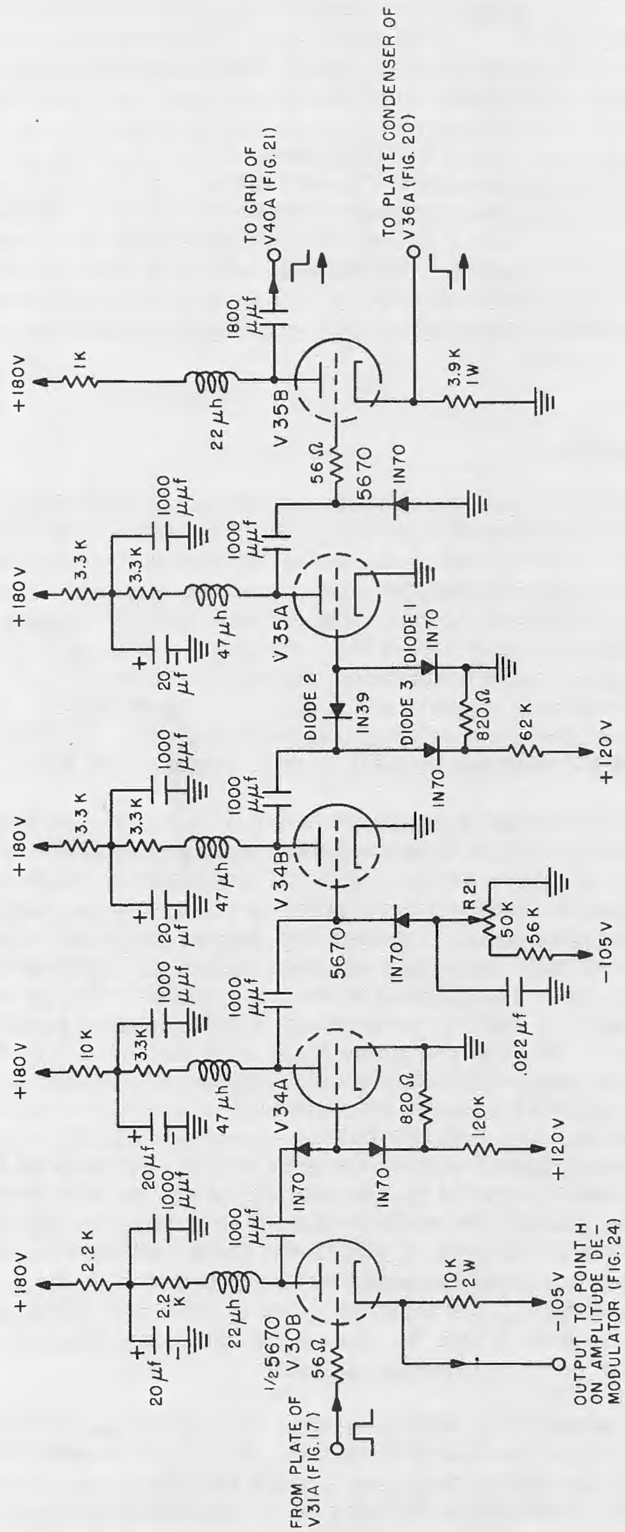

Fig. 19 - High-gain feedback amplifier circuit

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Pulse Stretcher and Storage Circuits

The positive-polarity pulses from the cathode of V35B are fed through diode V36A to charge a 250-$\mu\mu$f pulse-stretching capacitor (Fig. 20). The only discharge path for this capacitor is a 10-megohm shunting resistor returned to ground. A rectangular pulse is fed into the charging diode, and a sharp rise-time pulse having a long exponential decay (2500-μsec time constant) appears across the 250-$\mu\mu$f pulse-stretching capacitor. This stretched pulse is directly connected to the grid of a low-output-impedance cathode follower (V37A). The cathode of V37A is dc connected through a charging diode (V38) to one side of a 2-μf capacitor, the other side of which is grounded. This capacitor is the "storage" capacitor previously described. Each positive-polarity pulse appearing at the plate of charging diode V36A is stretched to a length sufficient for the complete charging of the 2-μf capacitor to a voltage almost equal to that of the peak value of the plate voltage of V36A. The ungrounded end of the storage capacitor is connected directly to the grid of cathode follower V37B so that the storage capacitor voltage can be sampled without discharging the capacitor. The cathode of this cathode follower is connected through a voltage-regulator tube (V39) to a 1.8K load resistor, the bottom end of which is connected to the negative 105-volt supply. This voltage-regulator tube serves as a low-impedance dc coupling network for transmitting the dc voltage changes appearing across the storage capacitor to the cathode of the pentode discharge tube V29 (Fig. 17). Across the voltage-regulator tube is shunted potentiometer R22, the arm of which is connected directly to the plate of diode V36B. The cathode of the diode is, in turn, connected directly to the plate of diode V36A. By adjusting potentiometer R22, the resting voltage on the grid of V37A is controlled, and since the grid of V37A is dc coupled through V37A, V38, and V37B to the cathode of V37B, the resting current through the voltage-regulator tube is thereby set. This feedback circuit from the cathode of cathode follower V37B through diode V36B back to the plate of diode V36A constitutes the circuitry that makes the stretching circuit a linear step charging circuit instead of an exponential one.

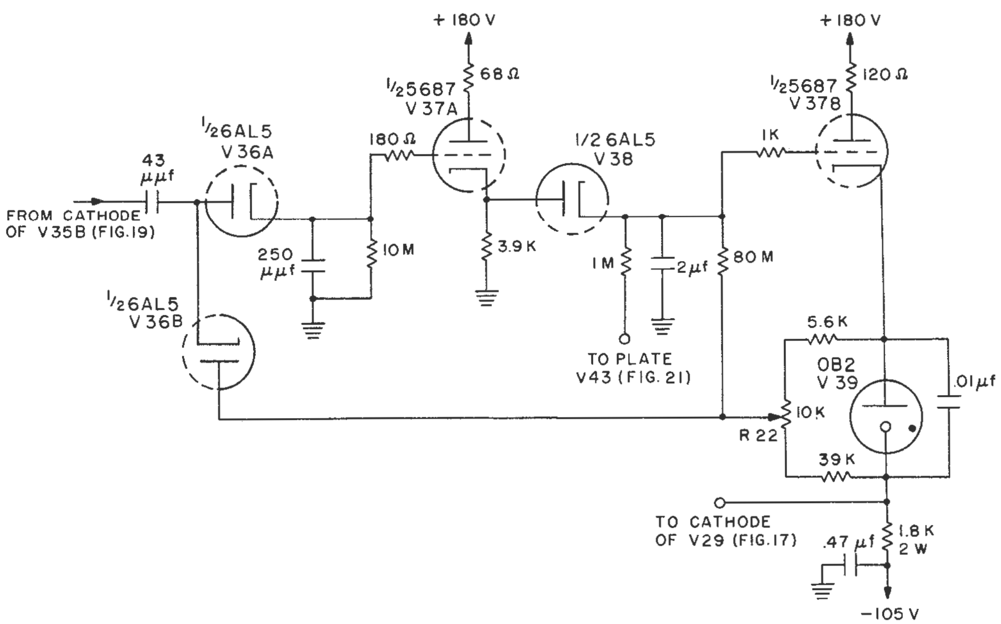

Fig. 20 - Pulse stretcher and storage circuits

[redacted]

The cathode of V35B is connected to the plate of charging diode V36A through a 43-$\mu\mu$f series capacitor. Assuming equal amplitude pulses, every positive-polarity pulse from the cathode of V35B places a small charge from this series capacitor on the 250-$\mu\mu$f pulse-stretching capacitor. Between pulses the series capacitor is recharged through V36B. Meanwhile, the voltage on the 2-μf storage capacitor has risen as a result of the small charge placed on the stretching capacitor. When the next pulse from the cathode of V35B arrives, the voltage on the plate of V36B has likewise risen so that its cathode is now clamped to a higher voltage. Between the next two pulses the series charging capacitor is recharged to this higher voltage with the result that a constant charge, independent of the voltage on the cathode of V36A, is placed on the 250-$\mu\mu$f capacitor. If the plate of V36B were simply returned to a fixed voltage, the charge placed on the 250-$\mu\mu$f capacitor would decrease with each succeeding pulse if the pulse repetition rate were high. If the repetition rate were low, the charge placed on the stretching capacitor would leak off between pulses, and each additional pulse would contribute nothing to raising the voltage on the storage capacitor. With the plate of the reference diode V36B connected as it is, the contribution of each pulse is stored and is used as a reference voltage to which the next charge is added.

To prevent the grid of V37B from drifting in a positive direction into the free-grid potential region, a bleeder was provided to keep the drift at a minimum, but at the same time it should have a sufficiently high resistance (80 megohms) to keep the drain current low. This bleeder is connected between the ungrounded side of the 2-μf storage capacitor and the R22 potentiometer tap. The current drain through this resistor was too high when it was returned to ground, resulting in a nonlinear sawtooth waveform from the sawtooth generator. Since the potentiometer tap rests somewhat above ground potential, and since it follows very closely the signal on the grid of V37B, the bleeder, when connected to this tap, acts as a very low constant-current drain on the storage capacitor.

Storage Capacitor Dunker and Associated Delay Circuits

Figure 21 shows the circuits that discharge automatically the 2-μf storage capacitor whenever the sawtooth signal at the plate of the pentode discharge tube (V29, Fig. 17) is of such low amplitude that it fails to pass through the high-gain feedback amplifier (Fig. 19). If at any time no signal appears at the cathode of V35B for a period of 3/4 second, a dunking tube automatically shunts a 1-megohm resistor across the 2-μf storage capacitor.

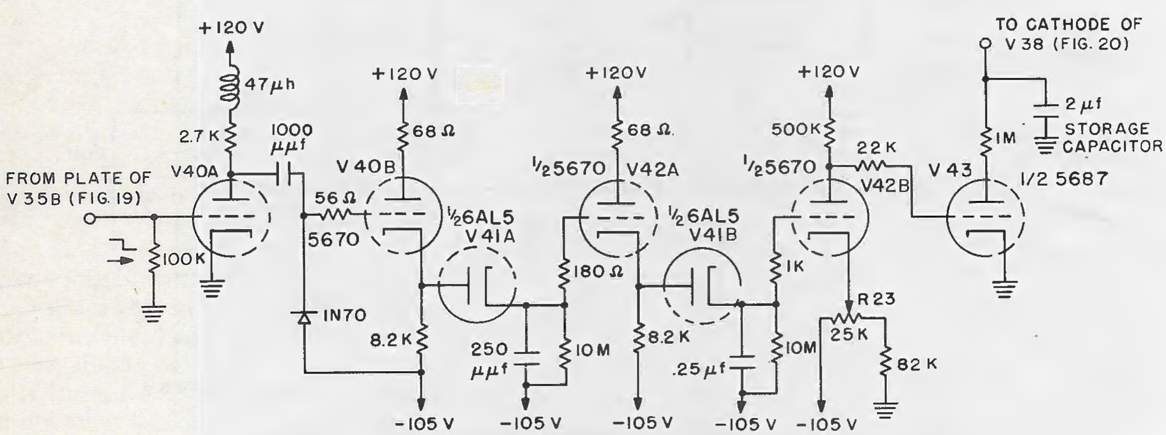

Fig. 21 - Storage capacitor dunker and associated delay circuits

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The negative-polarity signal from the plate of the split-load phase inverter V35B is sent to an amplifier (V40A) that produces a high-amplitude positive-polarity pulse which is sent to the grid of cathode follower V40B, the cathode of which is returned through an 8.2K resistor to the negative 105-volt supply. The grid is returned through a crystal diode to the same voltage so that the cathode normally rests about 100 volts below ground. Through diode V41A the positive-polarity cathode signals charge a 250-$\mu\mu$f pulse-stretching capacitor that is shunted by a 10-megohm resistor to the negative 105-volt supply. The positive-polarity stretched pulse appearing across this network drives a second cathode follower (V42A), which in turn charges a 0.25-μf pulse-stretching capacitor that is also shunted by a 10-megohm resistor to the negative 105-volt supply. This stretching network has a time constant of 2.5 seconds. Thus, whenever a positive-polarity pulse arrives at the grid of V40A, a positive-polarity pulse having a long exponential decay appears across the 0.25-μf capacitor. The voltage on this capacitor drives the grid of a triode, V42B, whose cathode is connected to the arm of a potentiometer, R23, which serves as a bias adjustment. The plate load for this triode is a 500K resistor returned to the positive 120-volt supply. By adjustment of the bias voltage on this tube, the plate voltage can be made to run 50 or 60 volts below ground, both grid and cathode normally resting only a few volts above the -105-volt level. The plate of V42B is connected directly to the grid of a triode dunking tube, V43. The cathode of this tube is grounded and the plate is connected to one end of a 1-megohm resistor, the other end of which is connected to the ungrounded side of the 2-μf storage capacitor (Fig. 20). In practice, potentiometer R23 is set so that with no signal applied to the input of the time-to-amplitude converter the plate voltage on V42B rests slightly above ground. With the application of signal, the 0.25-μf capacitor in the grid circuit of V42B is charged to a higher positive voltage resulting in a higher V42B plate current and a lower plate voltage. The lower plate voltage cuts off the grid of the dunking triode (V43) relieving the 2-μf storage capacitor of its 1-megohm load. The grid of V43 remains in the cutoff region as long as pulses arrive often enough to keep the 0.25-μf storage capacitor charged. If pulses are missing for any period longer than approximately 3/4 second, the voltage on the 0.25-μf capacitor will have dropped far enough again to turn on the storage-capacitor dunking tube. The storage capacitor is thus discharged so that it will be ready to be recharged to the proper voltage consistent with the repetition rate of the next signal that happens along.

Operation from Width-Modulated Pulses

The single-pole five-position switch SW14 (Fig. 18) provides the only switching necessary for the time-to-amplitude converter to operate from width-modulated pulses rather than period-modulated pulses. In the case of pulse-width or pulse-position modulated pulses a negative-polarity width-modulated pulse appears at the plate of V19B in the output Schmitt circuit (Fig. 12). This pulse is sent to the grid of the split-load phase inverter V21B (Fig. 18). The cathode load of V21B consists of a 0.5-μsec delay line, the output end of which couples a negative-polarity delayed pulse through switch SW14 to the input grid of Schmitt circuit V32. From the V19B plate of the output Schmitt circuit (Fig. 12) the undelayed negative-polarity width-modulated pulse is sent through switches SW10 and SW11 to the grid of cathode follower V27 (Fig. 17). This negative-polarity pulse (Fig. 22a) appearing at the cathode of V27 and the plate of diode V28A allows the sawtooth discharge capacitor to discharge through pentode V29 until the termination of the pulse when the sawtooth capacitor is recharged through diode V28A. The voltage waveform across the sawtooth capacitor appears in Fig. 22b. Because of the 0.5-μsec delay line in the dunking pulse generator, the dunking pulse (Fig. 22c) arrives at the grid of V31A 0.5-microsecond after the termination of the sawtooth with the result that a constant-width (0.5 μsec) pulse (Fig. 22d) whose amplitude is directly proportional to the width of the incoming pulse appears at the plate of V31A.

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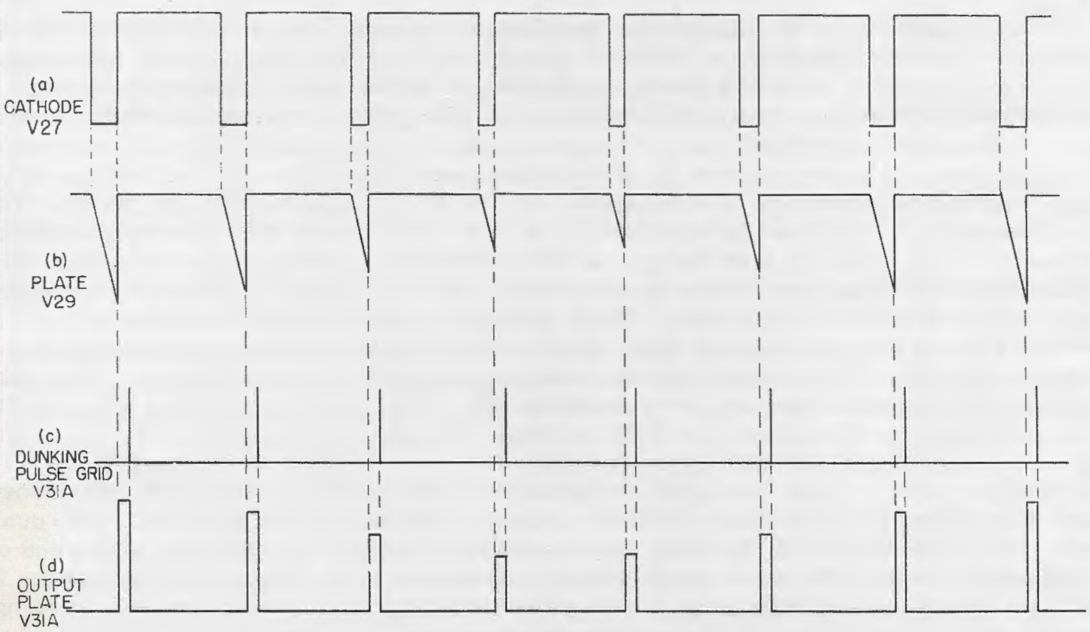

Fig. 22 - Time-to-amplitude-converter waveforms for width-modulated pulses (refer to Fig. 17)

The same range switch, SW15, that selects the pulse-period range now controls the pulse-width range so that pulses of widths varying from 0.5 μsec to 100,000 μsec may be linearly converted to amplitude-modulated pulses.

Adjustment

Adjustment of the various potentiometers in the time-to-amplitude converter is a simple matter if the following procedure is employed.

1. With no signal input to the converter, set R19 (Fig. 17), R20 (Fig. 18), R21 (Fig. 19), R22 (Fig. 20), and R23 (Fig. 21) so that the voltage on the potentiometer arm is most negative.

2. With a voltmeter connected between the cathode of voltage-regulator tube V39 (Fig. 20) and the negative 105-volt supply, adjust R22 so that the meter reads 20 volts. This adjustment assures sufficient current flow through the voltage-regulator tube.

3. Without disconnecting the meter, advance the R19 (Fig. 17) setting until the meter reads 24 volts. Potentiometer R19 adjusts the bias on the pentode discharge tube and determines the maximum current that can be drawn through this tube, consequently setting the minimum pulse period or pulse width for which the converter will operate properly.

4. Connect the voltmeter between the plate of V42B (Fig. 21) and ground. With V43 removed from its socket, adjust R23 so that the plate of V42B rests between 10 and 30 volts above ground. If this adjustment is properly made, the V42B plate voltage will automatically drop to about 50 volts below ground upon the application of signal.

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5. Reinsert V43 in its socket, and disconnect the voltmeter. Connect the input of an oscilloscope to the cathode of V31B (Fig. 18). Assuming that all the rest of the previously described CMPD circuitry is functioning properly, connect a pulse signal of repetition rate 100 kc and pulse width 1 μsec to the input of the CMPD. Set switches SW7 through SW14 for pulse-period demodulation. Adjust the bias on the dunking-pulse Schmitt circuit V32 with R17 until a rectangular positive-polarity 0.1-μsec pulse appears at the cathode of V31B.

6. Connect the voltmeter between the cathode of V31B and ground. Adjust R20 so that the cathode-to-ground voltage is about -4 volts.

7. Disconnect the voltmeter and connect the scope input to the cathode of V30A (Fig. 17). Adjust R21 (Fig. 19) slowly until a good stable sawtooth waveform first appears. Then, noting the amplitude of this sawtooth, continue turning R21 farther in the same direction until the amplitude of the sawtooth reaches about three-fourths of its original value.

8. Connect the input of the scope to the cathode of V30B (Fig. 19) and readjust R20 (Fig. 18) for proper dunking of the sawtooth voltage waveform.

AMPLITUDE DEMODULATOR

The function of the circuitry hereto described has been to select a desired pulse from a pulse train and convert its modulated parameter to an amplitude-modulated one. It is the purpose of the amplitude demodulator (Fig. 23) to produce from the amplitude-modulated pulses an envelope representing as closely as possible the information carried by the original video pulse as it entered the CMPD. The amplitude-modulated pulses from either the video selector or the time-to-amplitude converter arrive at the input of a split-load phase inverter. The negative-polarity output of the phase inverter is amplified twice and differentiated; the positive-polarity spikes are clipped off in the process. From the output of this differentiating amplifier and clipper comes a negative-polarity narrow pulse, the leading edge of which is coincident with that of the positive-polarity pulse entering the split-load phase inverter. However, the amplitude of this narrow trigger pulse is not constant, so it is sent to a slicing circuit where the top and bottom are sliced from the pulse to produce a clean, constant-amplitude, trigger pulse. This pulse then triggers a multivibrator to produce a positive-polarity 0.35-μsec dunking pulse, the leading edge of which is also coincident with that of the pulse at the input of the split-load phase inverter. This pulse is sent through a cathode follower to the boxcar generator.

The positive-polarity output signal from the split-load phase inverter is sent through a 0.5-μsec delay line to the boxcar generator. The boxcar generator is simply an amplitude-sensitive pulse stretcher having for all practical purposes an infinite time constant. The amplitude of its output signal is equal to the amplitude of the last pulse to enter the generator. One-half microsecond before each delayed amplitude-modulated pulse enters the boxcar generator the 0.35-μsec pulse from the dunking-pulse multivibrator is sent to discharge the infinite time-constant capacitor in the boxcar generator. About 0.15 microsecond after the dunking-pulse gate is over, the amplitude-modulated signal pulse charges this capacitor to a voltage equal to the amplitude of the pulse. The narrow negative-polarity spikes appearing in the output envelope result from the dunking pulse discharging the pulse-stretching capacitor.

In Fig. 24 a small, low capacity relay SW17 operating from a control deck ganged to SW7 through SW14 connects the input of the demodulator to the output of either the video selector or the time-to-amplitude converter, depending on how the original video pulse is

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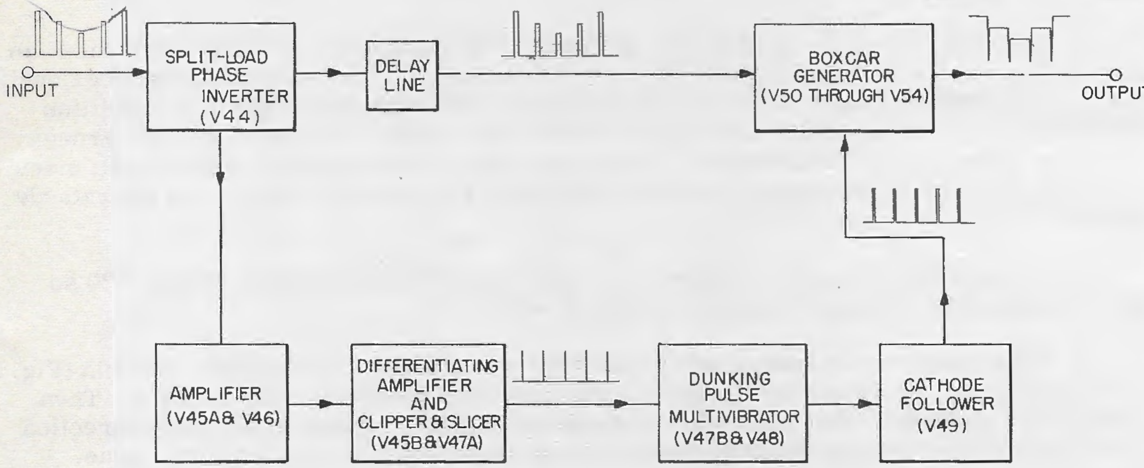

Fig. 23 - Block diagram of amplitude demodulator

modulated. The positive-polarity amplitude-modulated pulses are fed to the split-load phase inverter V44. The negative-polarity output signal from the plates is taken from the arm of a logarithmic 1000-ohm potentiometer (R24) which serves as the plate load of V44. This signal is fed to amplifier V45A. Because the pulses on the grid of V45A are of negative-polarity and can vary in amplitude from a fraction of a volt to 10 volts or so, the amplified positive-polarity pulses at the plate of V45A would likewise vary greatly in amplitude over a much greater voltage range if the clamping circuit consisting of V46 and a crystal diode were not present. It is the purpose of V45A to amplify the low-amplitude pulses so that they will be of sufficient amplitude to drive the differentiating amplifier V45B. If the clamping circuit were not present, the high-amplitude pulses would likewise be amplified with the result that if V45B was to have sufficient gain for the low-amplitude pulses, it would have insufficient dynamic range to accommodate the high-amplitude ones. The grid of V45B would then clamp to a negative voltage with the result that low-amplitude pulses arriving at this grid would be of insufficient amplitude to even drive the grid above cutoff. The crystal in the plate circuit of V45A clamps the plate to the voltage-regulator-tube voltage of approximately 105 volts. The plate of V45A normally rests at about 100 volts. Small negative-polarity signals on the grid cause the plate to move positively toward 120 volts which is the plate supply voltage for this tube. But the plate of V45A can never rise above 105 volts (the voltage-regulator-tube voltage) with the result that low-amplitude incoming pulses are amplified fully, and the high-amplitude ones cease being amplified after the output dynamic signal reaches a peak value of 5 volts. This signal is then sent to the grid of amplifier V45B, the plate load of which is a damped ringing circuit which produces a negative-polarity differentiated pulse from the leading edge of each positive-polarity pulse applied to the grid. This pulse varies in amplitude with repetition rate and with the amplitude modulation present on the pulse as it arrives at the grid of V45B. It must be clipped to remove both this amplitude variation and the small positive-polarity pulses not completely damped out by the ringing circuit crystal. A special one-tube amplifier-slicer (V47A)(described on page 27) performs both these functions. The resulting positive-polarity pulse of uniform amplitude from the plate of V47A is applied to the grid of V47B. The low plate impedance of this tube produced by the positive-polarity pulse on its grid is shunted between the plate and ground of V48A, triggering multivibrator V48. This multivibrator produces a constant-width gate of about 0.35 μsec. There is no bias adjustment on the grid of V48A, the bias being provided by a tap on the cathode resistor. The grid of V48B is clamped to ground through a 1N39 crystal diode to decrease the

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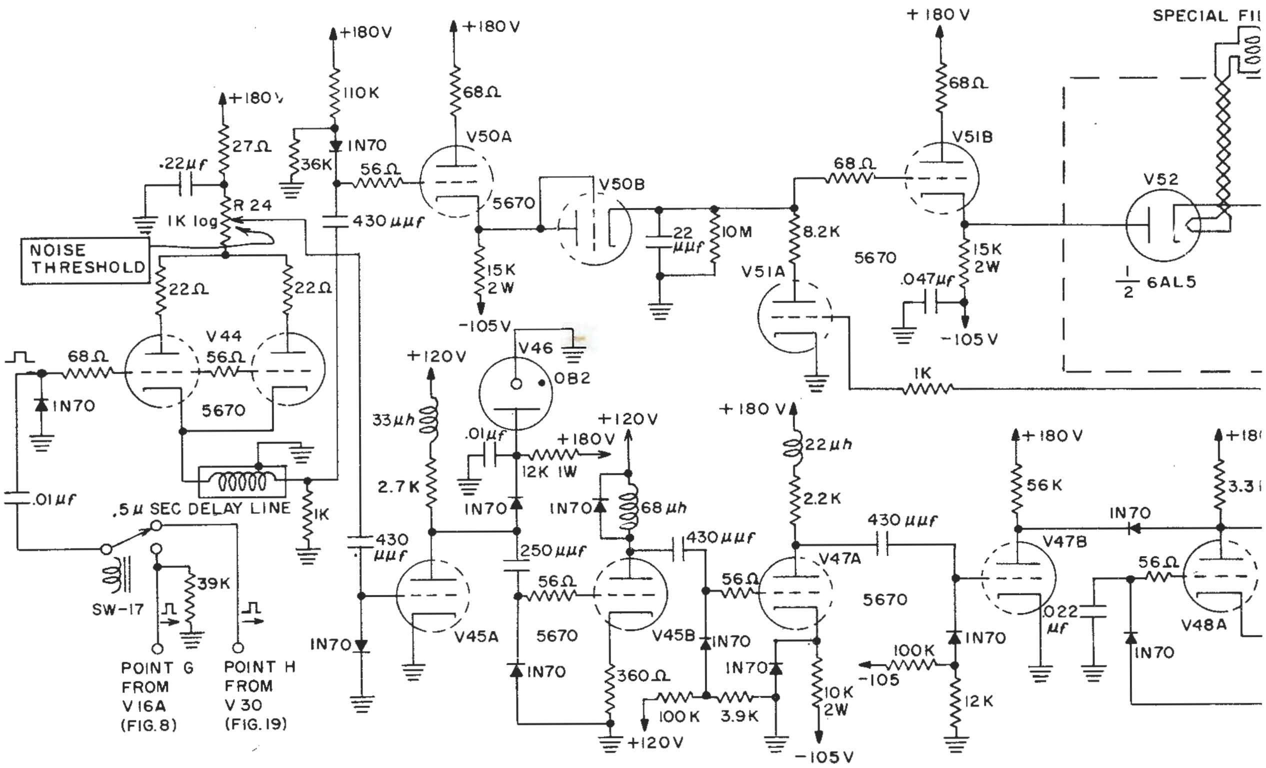

Fig. 24 - Amplitude demodulator (

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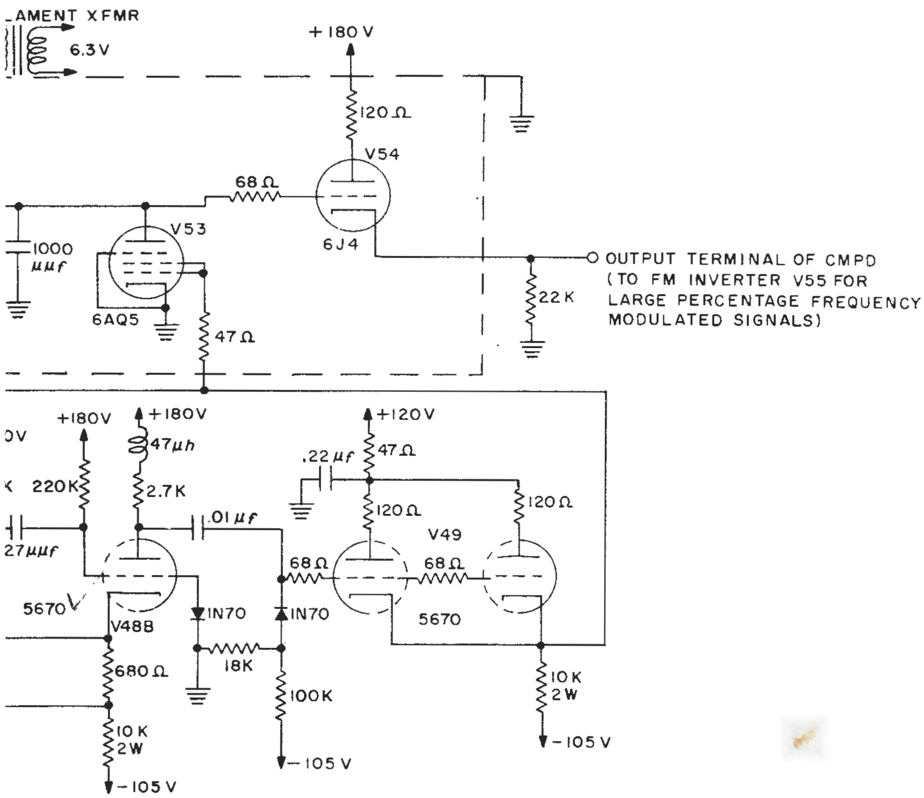

_AMENT XFMR
6.3V
+180 V
120 Ω
V54
68 Ω
V53
1000
μμf
6J4
6AQ5
47 Ω
OUTPUT TERMINAL OF CMPD
(TO FM INVERTER V55 FOR
LARGE PERCENTAGE FREQUENCY
MODULATED SIGNALS)
22 K
+180V
+180V
47μh
220K
2.7K
.01 μf
27μμf
+120V
.22 μf
47 Ω
120 Ω
120 Ω
V49
68 Ω
68 Ω
5670
V48B
IN70
IN70
5670
680 Ω
18K
100K
10 K
2W
-105 V
10 K
2W
-105 V
-105 V

circuit

multivibrator recovery time. The high diode-back-impedance shunted from grid to ground plays a very small role in determining the multivibrator gate width compared to the 220K timing resistor that is returned to the positive 180-volt supply, so that even with the wide variations in diode back impedance with temperature rise, the multivibrator gate width is quite stable. The positive-polarity gate from the plate of V48B is sent to a cathode follower V49. This provides the necessary low output impedance to drive the tube that dunks the capacitor in the boxcar generator.

The boxcar generator includes tubes V50 through V54. Tubes V50 and V51 comprise an amplitude-sensitive pulse stretcher that feeds the boxcar generator proper (V52, V53, and V54). The positive-polarity amplitude-modulated pulses from the cathode of cathode follower V44 are sent through a 0.5-μsec delay line to the grid of cathode follower V50A. The low-output-impedance cathode follower charges, through diode V50B, a 22-$\mu\mu$f capacitor to the peak voltage of the incoming pulse. The charge on this capacitor discharges exponentially through the 10-megohm shunt resistor. About 0.5 microsecond before the next video pulse appears at the output of the delay line, the positive-polarity dunking pulse from cathode follower V49 arrives at the grid of triode V51A. This tube is connected across the 22-$\mu\mu$f pulse-stretching capacitor, and upon the arrival of the positive-polarity dunking pulse the tube conducts very hard in order to discharge completely any charge left on this capacitor. About 0.15 microsecond after the termination of the dunking-pulse gate the delayed video pulse arrives and again charges the 22-$\mu\mu$f capacitor to the peak value of the pulse voltage. The 8200-ohm resistor in the plate of V51A prevents discharging the stretching capacitor to such a low voltage that diode V50B would carry excessive current, which would lower the V50A cathode voltage to a point where grid current would be drawn. This negative-polarity grid-current pulse would be sent back down the delay line to the cathode of V44 where it would appear as a positive-polarity grid-voltage pulse, often of sufficient amplitude to retrigger the dunking-pulse multivibrator. Figure 25a shows the amplitude-modulated pulse signal entering the pulse stretcher, and Fig. 25b shows the stretched-pulse signal as it appears across the stretching capacitor. This stretched pulse is then sent through cathode follower V51B which charges, through diode V52, a 1000-$\mu\mu$f capacitor. This capacitor has no shunt discharge path except that provided by the cutoff plate impedance of V53. The full charge then remains on this capacitor until the positive-polarity dunking pulse from cathode follower V49 arrives at the grids of the dunking tube (V53). During the 0.35-μsec dunking gate the charge on the 1000-$\mu\mu$f storage capacitor is shunted to ground by the plate impedance of the 6AQ5 dunking tube. The stretched video pulse from the pulse stretcher arrives 0.15 μsec after the conclusion of the dunking pulse gate to charge the 1000-$\mu\mu$f storage capacitor to the voltage amplitude of the video pulse. The voltage waveform (Fig. 25c) across the 1000-$\mu\mu$f capacitor drives the grid of cathode-follower output tube V54.

The video pulse stretcher is necessary to provide pulses of sufficient width to allow time for the 1000-$\mu\mu$f capacitor to charge to the peak amplitude of each video pulse. It takes approximately 0.2 to 0.3 microsecond for this capacitor to charge; the pulse stretcher allows the boxcar generator to function properly on video pulses as narrow as 0.1 microsecond.

In an earlier model trouble was encountered in the 1000-$\mu\mu$f storage capacitor circuit. Heater-to-cathode leakage in the charging diode (V52) caused undesirable 60-cycle ripple on the output waveform. A special heater transformer with high interwinding leakage resistance as well as electrostatic shielding was wound. Both windings of this transformer operate at 6.3 volts; the primary is fed from the 6.3-volt filament line and the secondary is connected directly to the heater of diode V52. This transformer is well shielded from the rest of the circuitry, and all heater leads are twisted and kept as short as possible.

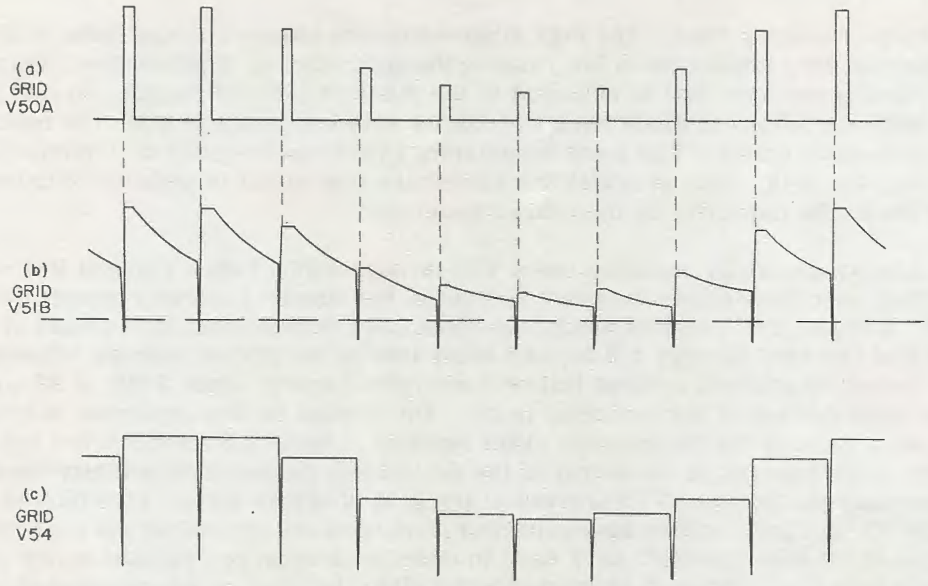

Fig. 25 - Amplitude demodulator waveforms (refer to Fig. 24)

After the transformer was installed and the above-mentioned precautions were taken, the 60-cycle ripple became negligible.

Leakage across the 1000-$\mu\mu$f storage capacitor, which was also a serious problem, was traced to two sources. Moisture from a person's breath provided a leakage path for the discharge of the storage capacitor thus making the equipment useless below pulse repetition rates of 200 cycles, and the "step" shaped output waveform degenerated into a sawtooth one. It was also found that the 5687 triode, then being used as the dunking tube, had an insufficiently high plate impedance at normal grid-cutoff voltages. Various other types of triodes were tried in place of the 5687's but with equally unsuccessful results. Tubes of the same type (5687's and 2C51's) picked at random from the shelf required anywhere from 12 to 80 volts of negative grid bias to cut off a devastating residual plate current that sometimes amounted to several thousandths of a microampere. Most of the triodes tested would not cut off completely regardless of the amount of negative bias applied.

The moisture problem was solved by installing V52, V53, V54, and the associated circuitry inside a shielded airtight box (indicated by dotted lines in Fig. 24). No more trouble from moisture has been encountered; however, the box could be evacuated and filled with a dry inert gas if necessary.

By substituting a pentode for the triode in the dunking circuit and by normally running both the screen grid and the control grid at about 15 volts below the cathode it was found that no sawtooth sag in the output voltage could be detected at repetition rates as low as 10 cycles per second. This was true of all pentodes that were tried at random from stock. The dunking pulse drives both the screen and control grids simultaneously to a potential about 5 volts above the grounded cathode. Bias for the 6AQ5 grids is determined by the grid bias on cathode follower V49, the cathode of which is directly coupled to the 6AQ5 grids. The 1K resistor in the grid lead of V51A allows cathode follower V49 to drive the two 6AQ5 grids a little harder than if there were three grids to drive directly. The grid of V51A does not have to be driven very hard compared to those of V53, since there is only a 22-$\mu\mu$f capacitor to discharge in the pulse stretcher compared to a 1000-$\mu\mu$f one to be discharged by V53.

The no-signal, dc output level of the output cathode follower (V54) is about 40 volts above ground. This voltage is determined by the bias on the grid of V50A; all connecting circuits between the grid of V50A and the cathode of V54 are dc coupled. It is necessary to keep the resting voltage on the plate of the 6AQ5 dunking tube high enough so that the on plate current through the dunking tube will discharge the 1000-$\mu\mu$f storage capacitor to its resting (no-signal) value in the 0.35 microsecond allotted to the dunking pulse.

Because of slight gas current in the 6J4 output cathode follower (V54) the grid in the absence of dunking pulses will slowly rise over a period of several seconds until it reaches its free-grid potential. This effect is not serious as long as there is no necessity for demodulating modulated pulses having repetition rates of less than 25 per second. This free-grid potential is not to be confused with the no-signal resting potential, which is that potential to which the voltage on the 1000-$\mu\mu$f capacitor would return if there were no signal entering the pulse stretcher but at the same time there were dunking pulses regularly discharging the 1000-$\mu\mu$f capacitor so that the grid of the 6J4 cathode follower could not drift.

While an attempt was being made to demodulate noisy amplitude-modulated signals, another problem arose. The fact that amplifiers V45A through V47A produce trigger pulses of equal amplitude from very low amplitude input pulses as well as high amplitude ones was disadvantageous. It was found that noise pulses on the base line would often produce trigger pulses of sufficient amplitude to trigger the dunking gate multivibrator, with the result that the unwanted noise pulses punched jagged "holes" in the output envelope. A gain control was installed as the plate load for V44, and the amplitude of the input signal to V45A was thereby controlled. In the event of a noisy signal this gain control is backed off until the noise pulses on the base line are of insufficient amplitude to trigger the dunking gate multivibrator. This being done, the noise pulses on the base line will not appear in the output envelope at all. Of course, those information pulses of amplitudes below that of the noise will not trigger the dunking gate, and those information pulses will also be without representation in the output envelope.

PULSE-FREQUENCY-MODULATION INVERTER

Theory of Operation

As explained previously, the output signal from the amplitude demodulator will not be a true representation of the original modulation envelope in the case of pulses that are originally linearly modulated with respect to frequency rather than period. The time-to-amplitude converter produces from frequency-modulated pulses as well as period-modulated ones constant-width pulses whose amplitudes vary linearly with pulse period. The envelope produced by the amplitude demodulator in the case of frequency-modulated pulses will be the reciprocal of the true envelope because the pulse frequency is equal to the reciprocal of pulse period. If the original pulse frequency modulation were of the form $E = \alpha(1 + \beta \sin \omega t)$, the output envelope from the amplitude demodulator would be of the form $E_0 = 1/\alpha(1 + \beta \sin \omega t)$. If the modulation factor β is small, then distortion in the output envelope is small (Appendix A); but as the modulation factor increases, the distortion in the output envelope also increases. In order to demodulate without serious distortion those frequency-modulated pulse signals having modulation factors greater than 0.1, circuitry has been developed to produce from a given dc input voltage the reciprocal of that voltage. This new voltage then represents the true modulation envelope when demodulating frequency-modulated pulses.

Figure 26 is a block diagram of a closed-loop voltage inverter. Voltage x, whose reciprocal is to be taken, is fed to one input of a multiplying circuit. The voltage sent by the output of the feedback network to the other input of the multiplying circuit will be called y. The voltage from the output of the multiplying circuit is αxy, the product of x and y multiplied by the product gain factor α of the multiplying circuit. This voltage is sent to a subtraction circuit where it is subtracted from a fixed voltage $\alpha x_0 y_0$ (where x_0 and y_0 are the quiescent values of x and y). The resulting voltage difference $(\alpha x_0 y_0 - \alpha xy)$ is sent to a high-gain amplifier of gain A. The output voltage from this amplifier is then added to a fixed voltage y_0. The resulting voltage $A(\alpha x_0 y_0 - \alpha xy) + (y_0)$, which is applied to the y input of the multiplying circuit, is by definition equal to y. The equation $y = A(\alpha x_0 y_0 - \alpha xy) + y_0$ becomes after division by A, $\alpha(x_0 y_0 - xy) = (y - y_0)/A$. If the gain A of the amplifier is large, the term $(y - y_0)/A$ is very small (provided the voltage variation $(y - y_0)$ is not extreme) and will be considered negligible. If k is substituted for $x_0 y_0$, the equation becomes $y = k/x$; therefore, the voltage y is proportional to the reciprocal of x, the input voltage.

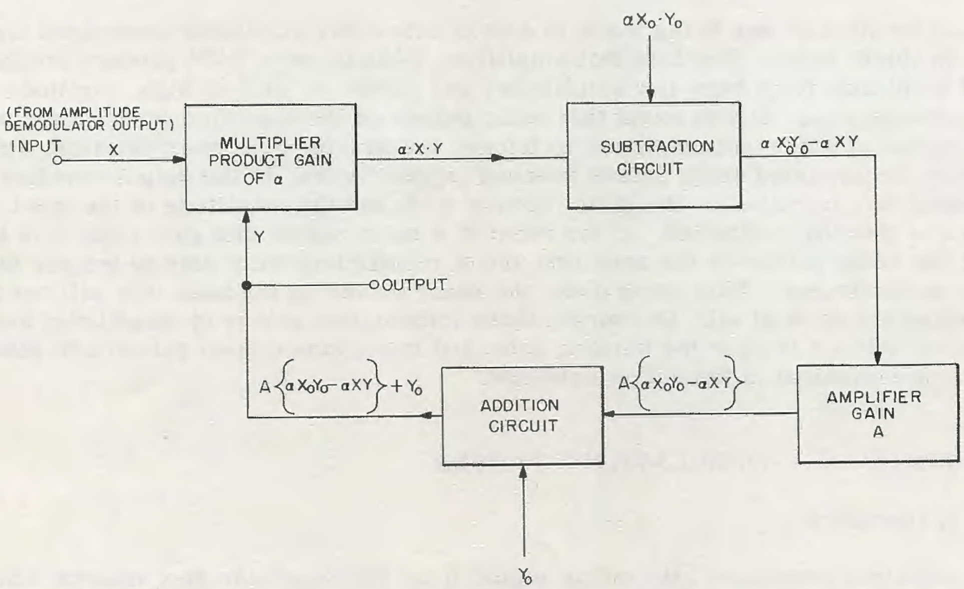

Fig. 26 - Block diagram of voltage inverter

Circuitry

The multiplying circuit utilizes the characteristic plate curves of a 5915 pentode. In this tube the suppressor grid as well as the control grid has a high grid-to-plate transconductance. Either grid by itself is capable of cutting off the tube plate current by applying sufficient negative bias. The interesting property of the 5915 pentode is that the plate current can be made proportional to the product of the voltages on the control and suppressor grids. Figure 27a is a plot of the plate current against various voltages on the control

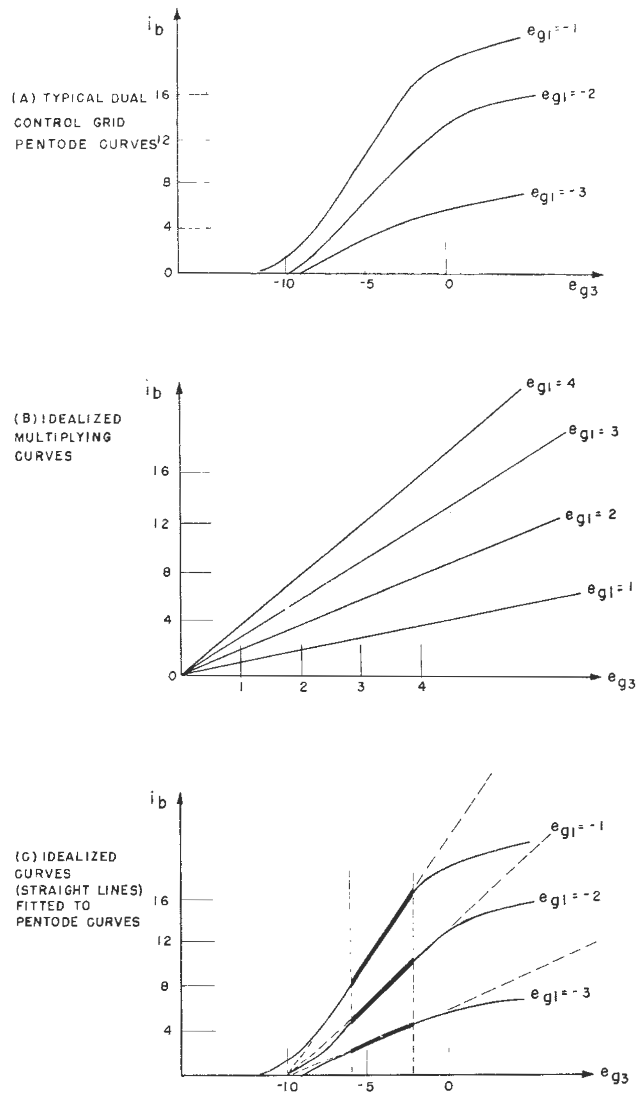

Fig. 27 - Characteristics of multiplying tubes

and suppressor grids for a pentode of this type. With the plate current as ordinate and the suppressor grid voltage as abscissa, the plate current curves for various set values of control grid voltage are plotted. Generally, the plate current increases with a decrease in negative bias on either or both grids.

Figure 27b shows the set of plate curves for an ideal multiplying tube. In these curves the control-grid voltage lines (e_{g1}) are spaced so that the slope of each line is one unit greater than that of the preceding line. If zero plate current and grid 3 cutoff voltage are placed at the origin, and if the grid 1 curve of zero slope represents cutoff voltage for grid 1, then it can be seen that the plate current will be zero whenever either grid is at "zero" potential. (For simplicity of explanation in connection with this argument, the term "zero" grid voltage will be identical with the value of cutoff voltage for that grid.)

It can also be seen that whenever both grids are above zero, the plate current is equal to the product of the two grid voltages. It should be noted that if the plate current is held at a constant value, the product of the voltages on the two grids must likewise be constant with the result that the two grid voltages are reciprocally related.

Figure 27c again shows the set of pentode curves first shown in Figure 27a. The portions of the curves set off with heavy lines closely approximate straight lines. If these heavy lines are extended (broken lines) until they intercept the $i_b = 0$ axis, they all intersect this axis at approximately the same point, namely $e_{g3} = -10$ volts with respect to cathode. In the region bounded by the values $e_{g3} = -6$ volts to $e_{g3} = -2$ volts and $e_{g1} = -3$ volts to $e_{g1} = -1$ volt, these curves closely approximate straight lines having equally spaced slopes as well as representing equally spaced values of grid 1 voltage. The dynamic zero voltage for grid 3 for these extended curves is about -10 volts. Extrapolating the relationship between the grid 1 curve slopes and voltages to the curve of zero slope, the $i_b = 0$ axis would represent a zero or cutoff voltage of about -4 volts for grid 1. If the tube represented by these curves is operated within the stated boundaries, the plate current will be directly proportional to the product of the grid voltages measured with respect to the respective zero voltage for each grid.

Because of the necessity for extreme stability in the high-gain amplifier, a balanced circuit was used. Figure 28 is a schematic diagram of the frequency-modulation voltage inverter. The voltage from the plate of multiplying tube V55 is dc coupled to one grid of balanced amplifier V57. The fixed voltage ($\alpha x_0 y_0$) is obtained from the plate of V56, another pentode, and is fed to the other grid of balanced amplifier V57. Pentode V56 is used as a fixed voltage source for reasons of circuit stability with respect to plate supply voltage drift. If the plate supply voltage shifts, the voltage difference between the plate of V55 and the plate of V56 ($\alpha x_0 y_0 - \alpha x y$) will remain fairly constant. It is this voltage difference that is to be amplified by the high-gain amplifier. The voltage ($\alpha x_0 y_0$) is adjusted by varying resistor R25 in the cathode of V56 and is set equal to the normal resting voltage of the V55 plate. The balanced output from the plates of V57 is dc coupled to another balanced amplifier stage (V58), the output of which is dc coupled to the y input of the multiplier tube. The voltage y_0 is provided by a 50K potentiometer in the output dc coupling network of V58. The 6AU6 pentode (V59) in the cathode circuit of V58 insures a constant-current cathode supply for balanced amplifier V58 so that variations in the plate voltage supply for V57 will not affect the output voltage of V58. The plate current of V59 is adjusted by a 500-ohm potentiometer in the cathode circuit.

For proper operation of this circuitry in conjunction with the output signal from the amplitude demodulator, two conditions must be met. First, in order to get distortionless output from the voltage-inverter circuit, the input voltage to the multiplying tube must lie at all times within the proper voltage boundaries (Fig. 27c). The second condition requires that the output no-signal voltage level of the amplitude demodulator equal the extrapolated zero voltage for the input grid of the multiplying tube. Because the input-grid extrapolated zero voltage does not equal the actual cutoff voltage, envelopes representing modulations above a certain percentage will not be properly inverted. A frequency-modulated signal having a large modulation factor produces from the output of the amplitude demodulator an envelope whose negative peak voltages approach the resting no-signal voltage level of the amplitude demodulator. Because of the peak-detecting nature of the automatic pulse-amplitude control circuitry of the time-to-amplitude converter, the positive peak voltage of the amplitude-demodulator output envelope will be constant regardless of the percentage of modulation. This peak voltage is placed within the multiplying-tube input grid-voltage boundaries (Fig. 27c). The lower voltage boundary determines the maximum percentage modulation that can be handled by the multiplying tube. This 5915 tube can handle

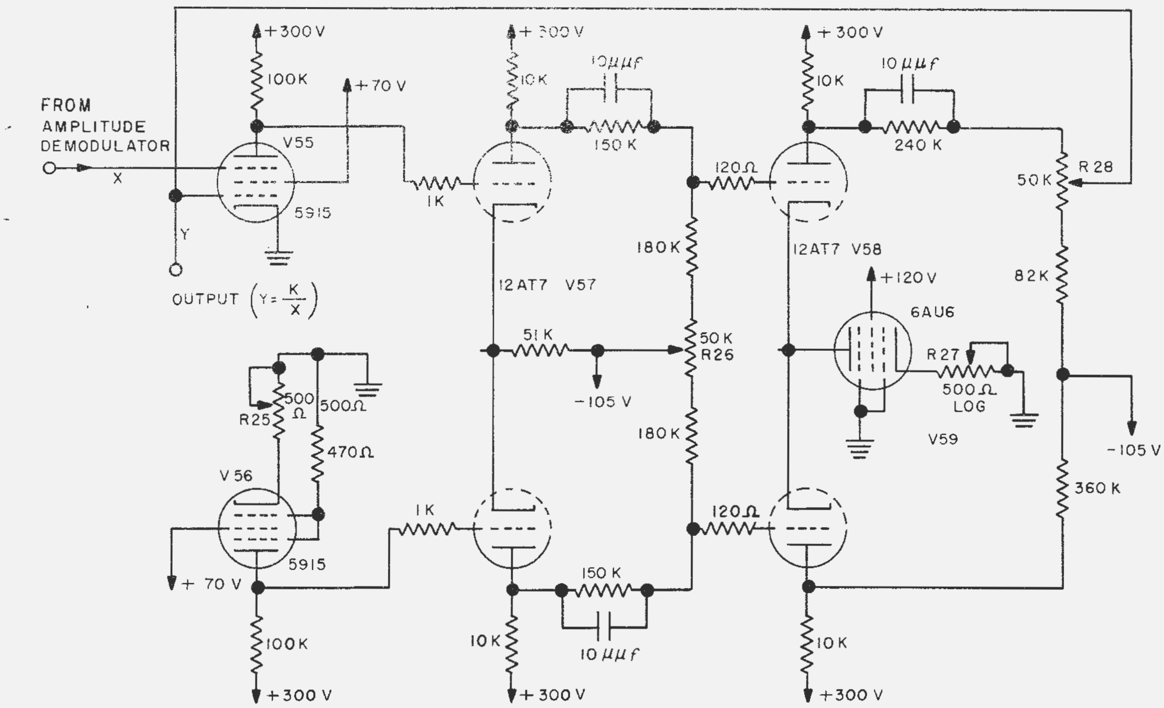

Fig. 28 - Voltage inverter circuit

successfully a maximum-to-minimum voltage ratio of 4 to 1, which allows proper operation with envelopes representing modulation percentages as high as 60% (Appendix B).

To align the voltage inverter, a voltage function of time, $E = k/t$, where $a < t < b$ (Appendix C), is applied to the input grid of the multiplying tube. The maximum-to-minimum ratio of E is determined by the values of a and b. The peak value of E is set equal to the difference between the peak and resting values of the voltage-level output from the amplitude demodulator. The value of E for $t = \infty$ is set equal to the zero voltage for the multiplying-tube input grid. With this hyperbolic waveform applied to the input grid of the voltage inverter, the waveform from the output terminal is observed on a linear time-base crt display. The cathode resistor (R25) of V56 is then varied until a setting is found for which the output waveform becomes a straight line which indicates a linear function of time.

Unfortunately, each 5915 tube has slightly different characteristics, and the alignment procedure must be followed each time this tube is replaced. No tubes were found that were actually incapable of producing a linear waveform from a hyperbolic one by proper adjustment of R25, but the drift in tube characteristics with warm up and the large differences between individual tube characteristics make this type of voltage inverter impractical for field use. For these reasons the voltage inverter has not been permanently incorporated in the present equipment, although other types of inverters are currently under consideration. There is also small likelihood of running into signals having modulations of percentage high enough to necessitate the use of the voltage inverter.

OPERATION

The CMPD is used in conjunction with a slave-sweep display indicator on which the operator can view directly the signal that he wishes to demodulate. The indicator PULSE POLARITY control (Fig. 29) is set to display positive-polarity pulses, and the indicator video GAIN control is set to a predetermined point and locked. The operator first sets the DISPLAY switch to the VIDEO position. The PULSE POLARITY and video GAIN controls are then set so that the amplitude of the raw video signals being analyzed produces a certain upward deflection on the indicator display. When these adjustments have been made, the video pulses are of sufficient amplitude to assure proper operation of all the CMPD circuitry.

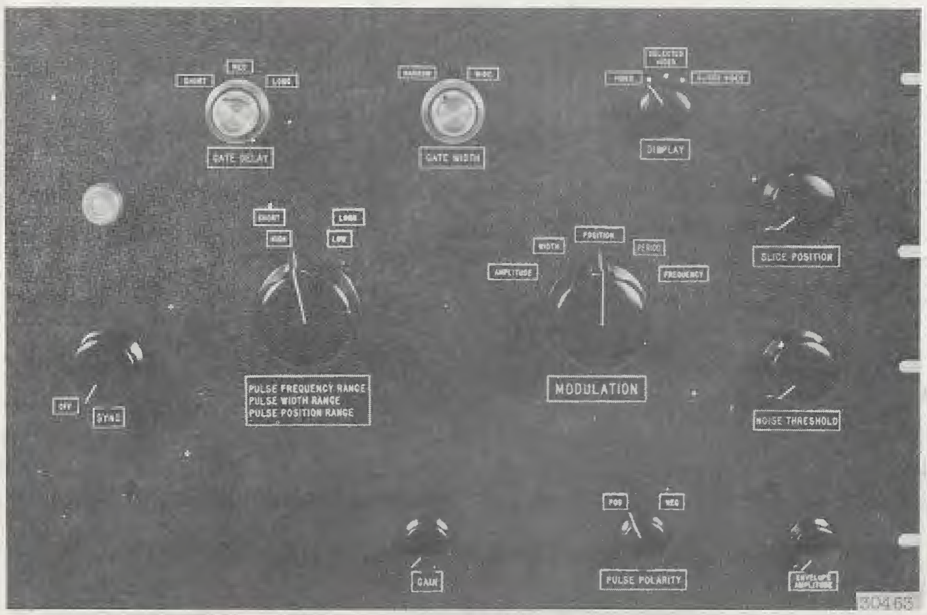

Fig. 29 - Complex-modulated-pulse demodulator control panel

Close examination of Fig. 30, a photograph of a two-channel pulse system with sync pulse as it appears displayed on an AN/SLA-2 analyzer, shows blurred leading and trailing edges on the first information pulse and a blurred trailing edge on the second information pulse. This display indicates pulse position modulation on the first information pulse and pulse width modulation on the second. With this display in view the operator next advances the SYNC control until the pulse selection gate appears on the base line of one of the display sweeps. He then adjusts the GATE DELAY and GATE WIDTH controls until the selection gate includes only the pulse that he wishes to demodulate. Figure 31 shows the display with the selection gate controls set to include the first information pulse alone.

To check for proper operation of the selection gate circuitry the operator then switches the DISPLAY switch to the SELECTED VIDEO position. Only the selected pulse and sync pulse should now appear on the display. The DISPLAY switch is then set to the SLICED VIDEO position, and the SLICE POSITION control is advanced until a stable pulse appears

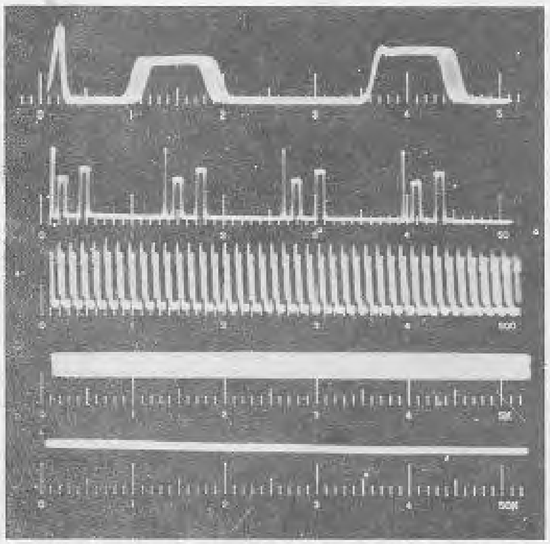

Fig. 30 - Two-channel pulse-system display with sync pulse, indicating pulse-position modulation on the first information pulse and pulse-width modulation on the second

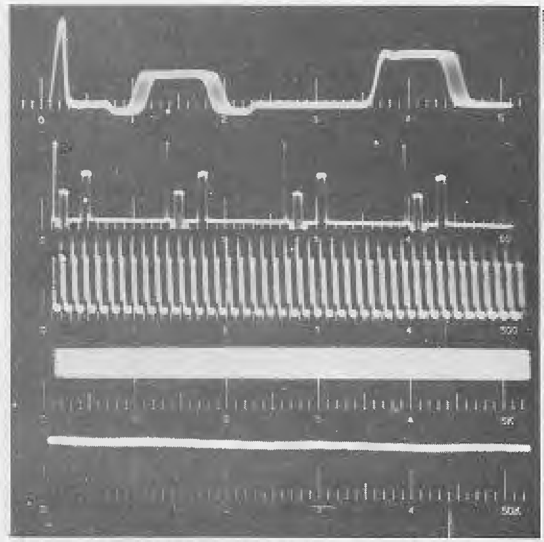

Fig. 31 - Two-channel pulse-system display showing negative-polarity pulse selection gate bracketting the first information pulse

on the display. (In the case of amplitude-modulated pulses the pulses are not sliced but instead are sent directly to the amplitude demodulator.) If the incoming video pulse is noisy, the SLICE POSITION control is adjusted so that the sliced pulse appearing on the display has a minimum of width or position jitter.

In the case of a pulse system where the sync pulse is absent, the SYNC control is turned OFF, allowing the signal to pass through the pulse selector without a selection gate being present.

In the case of pulse width demodulation (second information pulse, Fig. 32) the RANGE switch is set to either the LONG or the SHORT range depending on whether the pulse is longer or shorter than 250 μsec. If the operator wishes to demodulate the position parameter of a pulse, the setting of the RANGE switch would depend on whether the time interval between the sync pulse and information pulse were longer or shorter than 250 μsec. For pulse period or pulse frequency demodulation the RANGE switch would be set to either the HIGH or the LOW position depending on whether the pulse repetition rate was above or below 4 kc.

The NOISE THRESHOLD control is then advanced all the way clockwise, and the MODULATION switch is set to demodulate the proper parameter. The modulation envelope is now available at the output terminals of the amplitude demodulator. For aural reception of information carried by voice-modulated communication channels, a one-tube audio amplifier has been temporarily incorporated in the CMPD to amplify the voltage output from the amplitude demodulator to a sufficiently high level to drive headphones. The ENVELOPE AMPLITUDE control determines the gain of this amplifier. In the case of a noisy voice-modulated signal the operator may have to back off the NOISE THRESHOLD adjustment until the signal-to-noise ratio becomes a maximum. For other types of modulation this

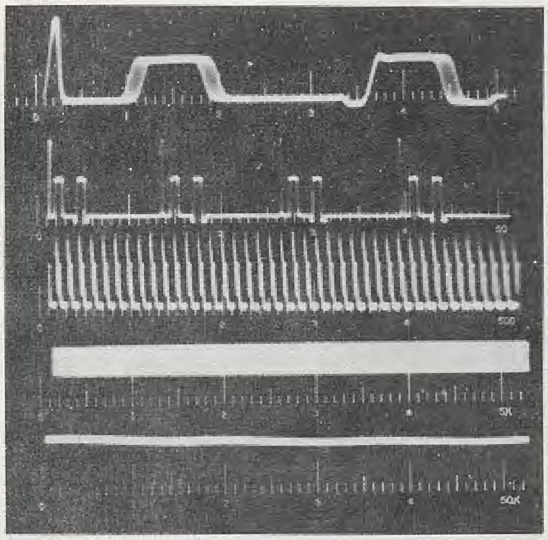

Fig. 32 - Two-channel pulse-system display showing negative-polarity pulse selection gate bracketting the second information pulse

adjustment must be made visually on a slave-sweep type analyzer. Techniques for making this adjustment are still under investigation.

If the operator discovers a pulse signal having more than one of its parameters simultaneously modulated, each parameter may be demodulated separately by setting the MODULATION switch to the position corresponding to the type of demodulation desired. Measurements have been made regarding crosstalk between two modulations on the same pulse. Table 1 lists the relative crosstalk between channels. A 1.2-μsec pulse at a repetition rate of 60 kc was amplitude modulated with a sine wave. The output from the demodulator was recorded for all five positions of the RANGE switch with zero db as the output level when the RANGE switch is in the amplitude demodulation position. This procedure was followed with respect to each parameter of the pulse.

TABLE 1
Crosstalk Between Channels

Type of Modulation on Pulse	Output Signal (db)				
	Output Channel Selected				
	Amplitude	Width	Position	Period	Frequency
Amplitude	0	-26	-32	-40	-40
Width	-29	0	-40	-40	-40
Position	-26	-40	0	-40	-40
Period	-25	-32	-32	0	0
Frequency	-25	-32	-32	0	0

A photograph of the output signal from the boxcar generator for an amplitude-modulated input signal having a pulse repetition rate of 75 kc and a modulation frequency of 6.2 kc is shown in Fig. 33. The same output signal with the NOISE THRESHOLD control adjusted so that the lower amplitude input pulses do not affect the output of the boxcar generator is shown in Fig. 34. This photograph might represent a noisy-signal situation where it is undesirable to have the low-amplitude noise pulses trigger the dunking circuit in the boxcar generator.

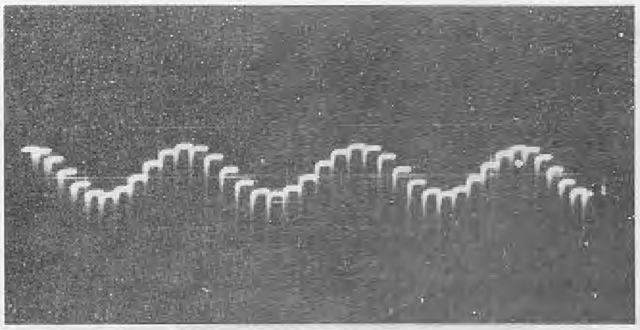

Fig. 33 - Boxcar-generator output signal for an amplitude-modulated input signal having a pulse repetition rate of 75 kc with a modulation frequency of 6.2 kc

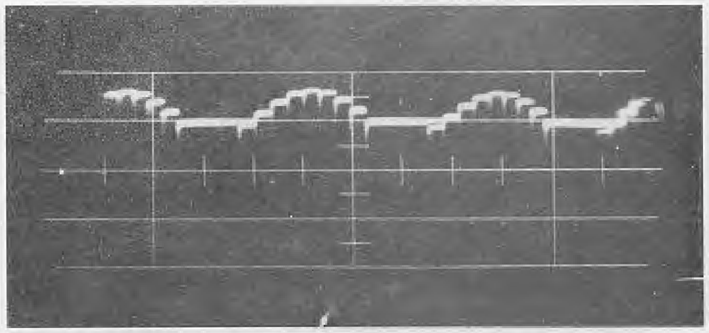

Fig. 34 - Boxcar-generator output signal with NOISE THRESHOLD control set for noisy signal, thus eliminating the contribution to the envelope from the lower amplitude pulses

CONCLUSIONS

The Complex-Modulated-Pulse Demodulator has proven to be a most versatile tool for the reproduction of modulation envelopes from complex-modulated pulse signals. It performs this function quickly and accurately for pulse signals modulated with respect to pulse amplitude, pulse width, pulse position, pulse period, and pulse frequency. With a minimum of manual tuning controls the operator can select any desired pulse from a pulse train and even demodulate separately different kinds of modulation on the same pulse with little crosstalk between modulation functions.

This equipment has also proven that pulse signals which have no apparent modulation when viewed on current slave-sweep analyzers can actually carry low-amplitude modulating envelopes that, when reproduced by the CMPD, have a very favorable signal-to-noise ratio. The CMPD has thus become a device necessary not only for the detection of information but also for the simple determination of the general type of modulation present.

[redacted]

It is anticipated that in the future the CMPD, or modifications thereof, will be the major tool used for the demodulation of complex-modulated-pulse signals such as those used in secured guided missile and communications systems.

The equipment utilizes standard electronic components, and no special techniques are required.

ACKNOWLEDGMENTS

The author wishes to express his deep appreciation for the wealth of invaluable advice received from John H. Markell and Henry K. Weidemann.

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APPENDIX A
Calculation of Distortion of Frequency-Modulated Signal Using Period-Modulation Detection

In the case of frequency-modulated pulse signals, the signal emerging from the amplitude demodulator will be of the form $E = k/(1 + \beta \cos \omega t)$ where β is the modulation factor and ω is the modulation frequency. If the above expression is expanded and x is substituted for $\beta \cos \omega t$, the series obtained is $E = k(1 - x + x^2 - x^3 + x^4 - x^5 + \ldots (-1)^n x^n + R_n)$ where R_n is the remainder after n terms and is equal to $(-x)^{n+1}/(1+x)$. Expanding x, x^2, x^3, and x^4 we get:

$$x = \beta \cos \omega t = \beta \cos \omega t$$

$$x^2 = \beta^2 \cos^2 \omega t = \frac{1}{2}\beta^2 + \frac{1}{2}\beta^2 \cos 2\omega t$$

$$x^3 = \beta^3 \cos^3 \omega t = \frac{3}{4}\beta^3 \cos \omega t + \frac{1}{4}\beta^3 \cos 3\omega t$$

$$x^4 = \beta^4 \cos^4 \omega t = \frac{3}{8}\beta^4 + \frac{1}{2}\beta^4 \cos 2\omega t + \frac{1}{8}\beta^4 \cos 4\omega t.$$

The amplitude contributions of these terms to ω, 2ω, 3ω, and 4ω are:

ω	$\beta + \frac{3}{4}\beta^3$
2ω	$\frac{1}{2}\beta^2 + \frac{1}{2}\beta^4$
3ω	$\frac{1}{4}\beta^3$
4ω	$\frac{1}{8}\beta^4$

Adding the 2ω, 3ω, and 4ω amplitudes we get:

Total amplitude for $2\omega + 3\omega + 4\omega = \sqrt{\left(\frac{1}{2}\beta^2 + \frac{1}{2}\beta^4\right)^2 + \frac{1}{16}\beta^6 + \frac{1}{64}\beta^8} = \frac{\beta^2}{8}\sqrt{16 + 36\beta^2 + 17\beta^4}$

as compared with $\beta + \frac{3}{4}\beta^3$ for the amplitude of the ω component. The ratio

$$\frac{\frac{\beta}{2}\sqrt{16 + 36\beta^2 + 17\beta^4}}{4 + 3\beta^2}$$

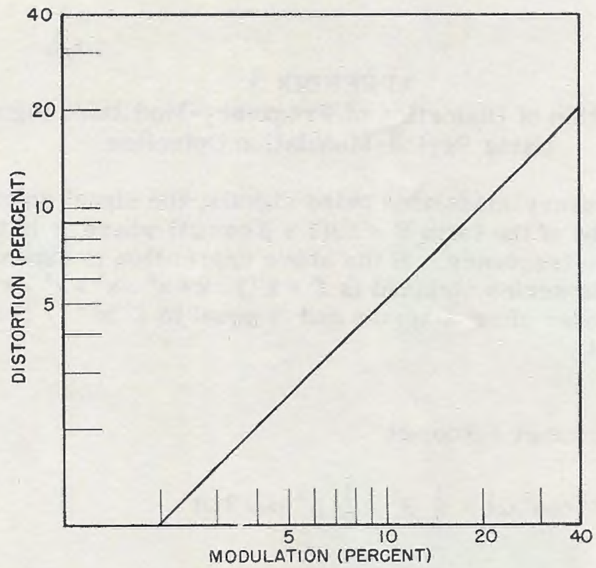

Fig. A1 - Percentage distortion vs. percentage modulation curve

is the relative distortion in the amplitude demodulator output envelope for a signal having a frequency modulation factor of β. Figure A1 is a plot of the percentage distortion against the percentage modulation.

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APPENDIX B
Percentage Frequency Modulation as a Function of
Multiplying-Tube Input-Voltage Ratio

A frequency-modulated pulse signal having a maximum-to-minimum pulse-repetition-rate ratio of x produces from the output of the amplitude demodulator a pulse whose amplitude varies over a voltage ratio of x. The maximum voltage ratio that the multiplying tube can handle is seen to be also equal to the maximum permissible frequency ratio. If x is designated as the upper frequency limit, then 1 will be the lower. The center frequency is $(x+1)/2$. The frequency deviation is $(x-1)/2$. The frequency modulation factor is the quotient of the frequency deviation and the center frequency and is equal to $(x-1)/(x+1)$. Example: For a voltage ratio of 4 the maximum percentage frequency modulation allowable would be one hundred times the modulation factor $= (100)(4-1)/(4+1) = 60\%$.

* * *

APPENDIX C
Arbitrary Function Generator

To expedite the development of the frequency-modulation voltage inverter an arbitrary function generator capable of producing hyperbolic voltage-vs.-time waveforms was developed to eliminate the painful task of plotting dc-wise the transfer characteristics of the inverter. By using this instrument the frequency-modulation inverter may be lined up in a matter of seconds rather than days.

In Fig. C1 an opaque paper mask with the upper edge cut in the form of an hyperbola is placed over the face of a short-persistence-screen cathode-ray tube. In front of the tube, shielded from all external sources of light, is placed a sensitive photomultiplier tube. The output of this tube is connected to the input of a high-gain dc amplifier, the balanced output of which drives the vertical deflection plates of the cathode-ray tube. The dc amplifier output is so connected to the crt deflection plates that light from the crt spot falling on the cathode of the photomultiplier tube will drive the upper deflection plate negative with respect to the lower one. The spot is thus driven down to the edge of the mask and is held there. If a linear sweep voltage is applied to the horizontal deflection plates, the spot will follow the edge of the mask with each sweep to produce at the vertical deflection plates a voltage waveform (Fig. C2) that when plotted against time is a very close copy of the paper mask.

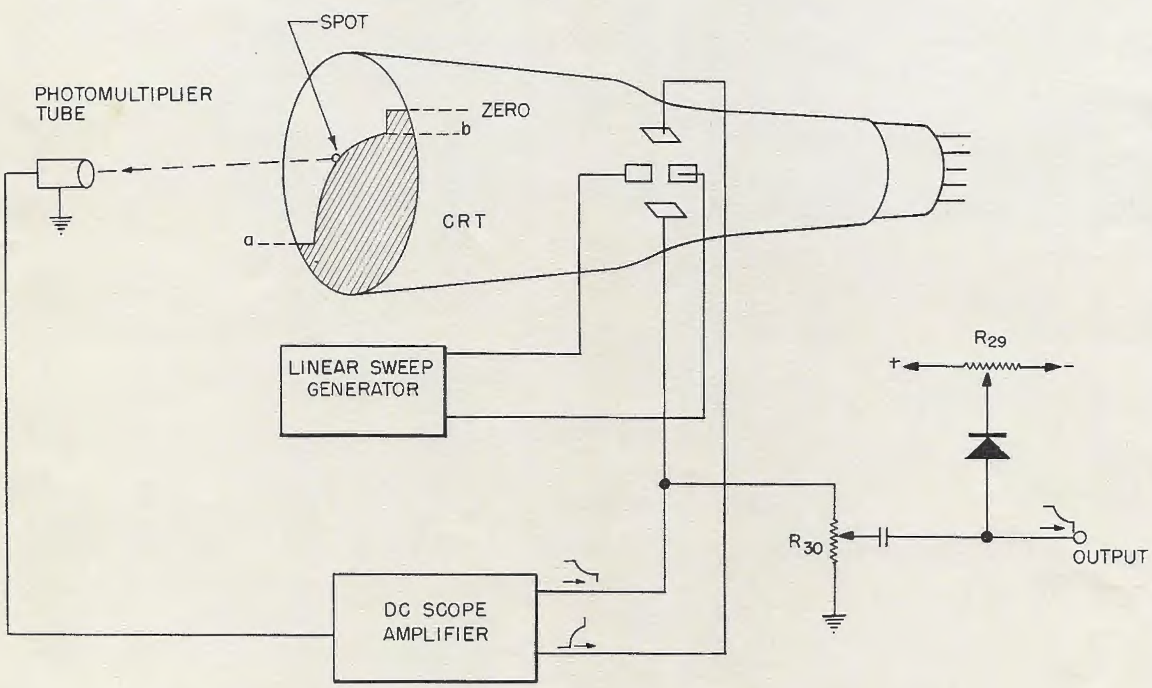

Fig. C1 - Diagram of arbitrary function generator setup

The shape of the mask determines the maximum-to-minimum voltage ratio (a/b) available (voltages are measured with respect to the voltage at $t = \infty$, i.e., the t-axis). Potentiometer R29 (Fig. C1) in conjunction with the crystal diode determines the absolute value of the positive peak voltage of the waveform. Potentiometer R30 controls the amplitude of the waveform, thereby setting the absolute value of the zero (t-axis) voltage.

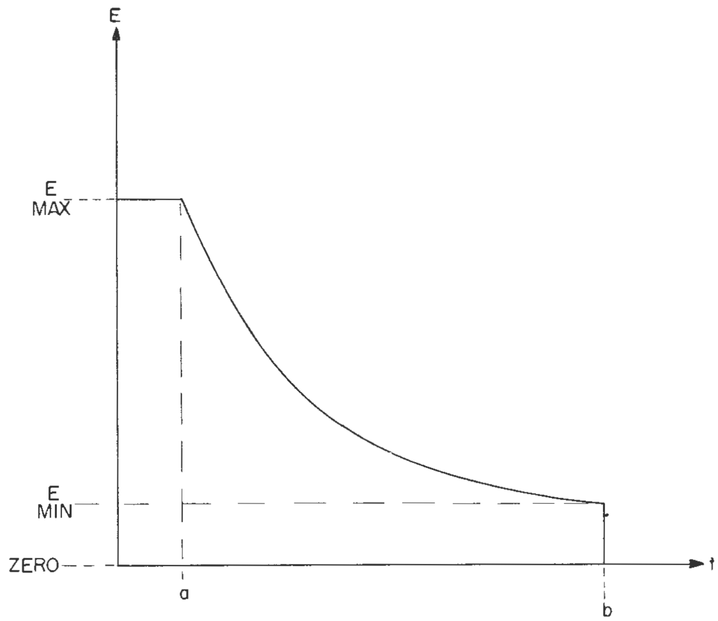

Fig. C2 - Hyperbolic output waveform

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