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DESIGN STUDIES FOR A MULTIFILTER HYDROGEN LINE RADIOMETER

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Introduction

At the NRAO an autocorrelation type hydrogen line radiometer has successfully been used with the 300-foot telescope during the past year. With the 140-foot telescope nearing its completion (early 1965) it was felt that an independent hydrogen line receiver should be available for use with the new dish. The question of which receiver goes to which telescope was left open, and thus the new receiver should also be able to work at the 300-foot telescope, with its stricter requirements concerning integration time. An arrangement with a moving feed to increase the possible integration time was not considered.

At the time when the need for a new hydrogen line receiver was anticipated, no final decision on the receiver type was made. This report will give the results of some preliminary studies of a filter type receiver. It should perhaps be remarked that at the time of this writing, about three quarters of a year after this work was initiated, the case for an autocorrelation receiver is very strong. This is largely due to the greater familiarity with this receiver type gained during the intervening period, and its considerable versatility. The main advantages of a filter receiver are its somewhat lower cost, and the 30% better theoretical sensitivity.

The bandwidth and number of channels were not specified exactly at the outset. It was understood, though, that the bandwidth should be about 10 kHz (or perhaps smaller). The total band covered by the channels should be close to 2 MHz. In the hydrogen line profile catalog published by Muller and Westerhout [1], the widest profile is about 1.9 MHz (the corresponding point on the sky cannot be reached with the 300-foot telescope, but well with the 140-foot). The final number of channels will have to be settled when (and if) it is decided to build a filter type receiver.

It has been assumed that by the time the present receiver would be ready for operation, a parametric preamplifier giving a system noise temperature of 100 °K would be available. This may not become true, but the nonlinearity effects arising from such a low system noise temperature have been taken into account in the present design. Assuming a system temperature of 100 °K, a bandwidth of 10 kHz, and an integration time of 10 seconds, the theoretical sensitivity is about 0.5 °K rms.

The experimental work done so far has almost exclusively dealt with the detailed design of the individual channels. In this report, though, consideration will be given to all the major parts of the receiver, mainly in the form of proposals or remarks, and often without the backing of experimental data.

The construction philosophy adopted is that after the design has been made at the NRAO, the actual fabrication of units be given out to a company on a competitive basis. Printed circuitry should be used as much as possible.

It was specified from the beginning that the output data be recorded on magnetic tape.

Block Diagram

The gross design of the receiver is conventional (fig. 1), employing the well-known frequency switching technique (rate 400 Hz).

The signals received by the antenna are amplified in the low-noise parametric preamplifier. They are then combined alternately in a crystal mixer with two LO frequencies, one fixed at 1393 MHz, and one that is slightly variable around 1390 MHz (but fixed at any given time).

The resulting 30 MHz IF is further amplified, passed through a gain modulator and converted to the final IF through mixing with a second LO. After amplification at the final IF the signal is distributed to the various channels. Two versions of the channel IF sections are considered, one employing standard crystal filters, and the other using a contiguous comb filter set.

A square law detector extracts the 400 Hz square wave modulation containing the signal information. This is passed through a narrow band selective amplifier centered at 400 Hz, and is finally converted to a DC voltage in a synchronous detector.

The DC voltage is smoothed in an RC network with a time constant of 10 seconds. The outputs of all the channels are sampled at 10-second intervals and fed to the digital output system and recorded on magnetic tape.

Two levels of calibration signal should be provided, say 20 °K and 100 °K. This is most conveniently achieved by two switched neon bulbs inserted through the bottom plate of the feed horn.

An isolator between the parametric amplifier prevents the LO from leaking into the amplifier and overloading it.

The system indicated in the block diagram for obtaining the square wave voltage for the frequency switch, calibration lamps, gain modulator and the synchronous detectors has two advantages over a circuit starting with a 400 Hz oscillator. First, letting 800 Hz pulses trigger a bistable multivibrator, perfect symmetry of the generated 400 Hz square wave can be ascertained. Second, it may turn out that the 30 MHz IF needs to be blanked out during the switching transients. This can very conveniently be done by driving the suppression unit directly from the pulse generator, and having the pulses triggering the square wave generator delayed in a separate delay circuit (monostable multivibrator) preceding it.

The outputs from the square wave generator feeding the frequency switch, calibration lamps, and the gain modulator should give a voltage with phase 0° as well as phase 180° with respect to the reference voltage for the synchronous detector. This will assure easy changes of the output DC voltage polarity, and of the calibration signal polarity.

Also, a small phase variation around the nominal values for the three outputs mentioned above (common to all of them), will provide correction for a possible uniform phase shift in the channel audio sections.

Other parts of the receiver will be discussed later.

Selecting Frequencies

Great care must be taken when local oscillator and filter frequencies are selected, in order to avoid self-generated interference in the receiver. There are, of course, in general many combinations of frequencies that give a system free from self-interference, but as soon as a couple of frequencies have been chosen, the choice of the remaining ones is very restricted.

The first IF is chosen as 30 MHz in order to adhere to the standard adopted at the NRAO. This immediately gives a first LO frequency of about 1390 MHz (or 1450 MHz), which is obtained by the multiplication of an original, lower frequency. A multiplication factor of 100 was selected, mainly in order to simplify the setting of the first LO. This will increase the price of the multiplier by about $1000 (at a level of $5000), since it is more difficult to make multipliers involving factors of 5 ($100 = 2^2 \cdot 5^2$) than such with factors of 2 or 3. A starting frequency of about 13.9 MHz is obtained with the factor chosen.

The second LO and filter frequencies now have to be selected in such a way that they themselves, or their harmonics, or combinations of harmonics, fall outside a 2 MHz wide band around 30 MHz (the first IF) and outside a similar band at the final IF. The final IF depends on the second LO frequency, of course. It turns out that two bands, at about 7 and 13 MHz, can be used. The lower band is then preferred.

When checking the possible interfering frequency combinations it is most convenient to do it systematically by means of a table. In a horizontal row the first LO frequency and its harmonics are written, and in a vertical column the corresponding quantities for the second LO frequency. The interfering frequencies can immediately be written down as the sum and difference of all the possible combinations. In appendix 1 this is done for 5 different second LO frequencies. Only the differences are in the table, since the sums obviously are much higher than 30 MHz.

The final choice is a second LO of 23.2 MHz and a final IF band between 5.8 and 7.8 MHz.

Signal Levels in the Channels

The analog-to-digital converter to be used in the output system has a full scale capability of ± 10 V. To use its full range means an inconveniently high output signal from the synchronous detector. An operational amplifier before the A/D converter brings the required output signal down to, say, ± 0.5 V which is easier to manage. The maximum hydrogen signal from the galaxy, about 120 °K, then should not give a DC output larger than 0.5 V. Of course, the synchronous detector must be capable of delivering a higher output voltage so that there is no risk of overloading.

It is in general desirable to keep the audio (400 Hz) amplifier gain low, in order to reduce the chance for pick-up from the reference voltage for the synchronous detector. On the other hand, the audio input signal available is commonly limited by the second detector. The reason is that if a square law detector is required, a low detector output level, less than about 10 mV, is necessary. Even so, the dynamic range over which the square law is valid is fairly narrow.

Previous experiments by Claude Bare [2] with a detector circuit employing two diodes had indicated the feasibility of a square law detector operating at a high output level with a great dynamic range. It was decided, therefore, to assume an operating point of 100 mV DC.

With an IF bandwidth of 10 kHz and an output time constant of 10 seconds, the relative rms noise fluctuations will be

$$\frac{\Delta T}{T} \approx \frac{1}{\sqrt{10^4 \cdot 10}} \approx 0.003,$$

corresponding to a modulation at the input of the audio amplifier of $0.003 \cdot 100 = 0.3$ mV. The maximum modulation signal, assuming a system noise temperature of 100 °K, and a maximum signal of 120 °K, will be 120 mV peak-to-peak. If the system noise is higher, say 200 °K, the maximum modulation will be lower, in this case 60 °K (still assuming an operating point of 100 mV). Thus the audio gain must be high enough to allow for the least favorable case.

Tests on the detector circuit showed that for an operating level of 100 mV DC, about 400 mV rms are needed at the detector input. Allowing a factor of five to avoid clipping of the noise peaks, the channel must be capable of delivering $0.4 \times 5 = 2$ V rms into the detector.

Audio Circuit

The design of the audio section for the channels was approached in two different ways. With the required performance specified, some companies were asked to make a complete audio unit, comprising selective 400 Hz amplifier, gain control, synchronous detector, and signal and reference voltage transformers. An audio unit was also designed at the NRAO with commercially available parts. Full-wave type synchronous detectors were used in all cases.

Of the companies asked, two, Rawco Instruments, Inc. (Fort Worth, Texas), and Natel Engineering Co., Inc. (Van Nuys, California), submitted designs, and also prototypes for evaluation.

The synchronous detector used in the Rawco unit is a transistor chopper. Tests made showed a good linearity up to about 2 V DC output, where distortion occurred. With maximum gain the undistorted output was lower. When the input was shorted, a full-wave rectified sine voltage appeared at the output, giving a DC offset of the order of 10 mV. The price asked for the unit was \$220 for quantities greater than 100.

The Natel design employs a diode synchronous detector in a common bridge circuit. The bridge must be carefully balanced with resistors to prevent the reference voltage from appearing at the output. The prototype supplied by Natel was found to give a maximum DC output of about 2.5 V, with good linearity. A rather large, unsymmetrical (with respect to ground) square wave was present at the output with short-circuited input, giving rise to a DC offset that increased with the square of the amplitude of the reference voltage up to over 20 mV. The unit was sent back to the factory and readjusted. After this the DC offset was less than 1 mV. One month later, though, the offset had increased to 7 mV, possibly due to aging of components.

The price of $130 was asked for the unit for quantities greater than 100.

On both the Rawco and the Natel units the reference voltage input had to be loaded by 100 Ω in order to avoid a distorted square wave. The drive current thereby increased to about 50 mA, which is too large. A sine wave reference voltage did not give the same problem, and reduced the necessary drive current to 5-10 mA, without much affecting the output DC voltage.

The NRAO design is shown in figure 2. The selective amplifier is a White Instruments active filter type 252 (reassembled for printed circuit board mounting and called type 2290). It has been used extensively at the NRAO and works very well. Its 3 dB bandwidth is 20 Hz and the voltage gain 20 at the nominal load impedance of 100 kΩ. In the present application the load impedance is very much lower, about 1 kΩ, reducing the voltage gain by a factor of 2. This loss is picked up in the drive transformer for the synchronous detector, which has a 1:4.5 voltage ratio (impedance 1.2 kΩ:25 kΩ), giving a net voltage gain of about 2.3. The transformer is a Microtran type MMT 18-M. The capacitor across its secondary corrects for the phase shift that occurs. Some other transformers with higher voltage ratios were also tried, but their low primary impedance caused distortion.

Two transistor choppers were tested as synchronous detectors, Airpax type 6025-8 (type 6025 reassembled for printed circuit board mounting), and Plug-In Instruments type C-24. Both types worked very well, and with a DC offset around 1 mV. The Airpax chopper was chosen, finally, being directly available in a printed circuit board version.

In the final prototype unit the DC offset increased to 2 mV due to the more compact construction. DC output versus square wave input voltage is depicted in figure 3 for the final unit. The dependence of the DC output on the reference voltage level was investigated. A selected output voltage of 0.50 V stayed constant for reference voltage variations between 30 and 0.3 V rms. Appendix 2 gives the cost estimate for the unit.

The NRAO design was finally adopted, both from performance and cost considerations.

Detector

It is desired to have the receiver output device deflection due to a source of radiation translated into the equivalent change in antenna temperature, ΔT, at the receiver input. In general a correction dependent on the detector law and the ratio $\Delta T/T_N$, where T_N is the system noise temperature, is required in order to find ΔT from the deflection ΔV.

The relation between the equivalent input noise temperature T_{IN} and detector output voltage can be written [3]

$$T_{IN} = C \cdot V_{OUT}^{\alpha},$$

where C is a constant. For a linear and a square law detector α is equal to 2 and 1, respectively. In [3] the correction factor for obtaining the true change in input temperature due to a source is derived. As an example, an α of 1.33 and $\Delta T/T_N = 1$ gives a 10% correction. α must be < 1.03 to give a correction less than 1% for $\Delta T/T_N = 1$.

For a square law detector, $\alpha = 1$, the correction factor becomes unity for all values of $\Delta T/T_N$. It is therefore desirable to have a square law characteristic over as wide a dynamic range as possible. The experiments by C. Bare [2] mentioned before showed promise, but α was not constant (and equal to unity) over a wide enough range. The suggestion in his report of using more diodes was followed up, and new tests made with the detector circuit shown in figure 4. The detector must be driven from a low impedance source (in this case 25 Ω). The values of the resistors R_1, R_2 and R_L were found by experiment, and this was done within an hour or two. Figure 5 shows a series of curves plotted during the alignment procedure (frequency 7 MHz).

In a mass production of channels it is, of course, very inconvenient to have to adjust each of the detector assemblies individually. Therefore it is desirable to use diodes with as similar properties as possible. These can certainly be obtained at a moderately increased price from a diode manufacturer. In order to test the spread in the detector characteristics with ordinary diodes, three measurements were made with the finally selected resistor values on independent sets of diodes. The result is shown in figure 6, indicating a very reassuring similarity between the three curves.

The behavior of the detector circuit at various temperatures was also investigated. During the measurements the detector assembly and the driving transistor amplifier were enclosed in an oven. By mistake the measurements were not made on the finally adopted circuit, but on one with the values $R_1 = R_2 = 1.0$ kΩ, $R_L = 2.2$ kΩ. However, the temperature dependence can be appreciated also in this version, and is depicted in figure 7.

At the normal operating temperature the deviation from a square law is very small over an output range between 10 and 400 mV, as seen in figure 6. The selected operating point of 100 mV then allows a $\Delta T/T_N = 4$, which will not be reached for normal observation with the proposed system.

A large signal will be observed, though, during possible absorption measurements on Cas A. With the 300-foot telescope this source gives an antenna temperature of about 2000 °K. On the source, at the signal frequency, the receiver noise plus hydrogen line will amount to about 200 °K. At the comparison frequency the source antenna temperature plus the receiver noise gives 2100 °K. The modulation signal will be 1900 °K, giving a square wave of 1900 mV p-p at the normal operating point of 100 mV (if the detector were square law up to 2100 mV). In this case an IF attenuation of 10 dB will cause the signal levels to vary between 20 and 210 mV, still within the square law range.

Channel IF Section

a) Conventional design.

In a multifilter receiver it is a great convenience to have only the filter itself as the frequency determining element. Therefore, it was decided to build an untuned IF amplifier, identical for all channels.

It was stated earlier that an output capability of 2.5 V rms into the detector circuit is necessary. Tests with the detector circuit just described also showed that the output impedance should be low, pointing to an emitter follower output stage. In order to estimate the gain required in the channel, it should be realized at first that a strong bandwidth compression of the noise signal takes place during the passage of the filter. From 2 MHz to 10 kHz, the signal level is reduced by 23 dB. The input level should be fairly low, say below 100 mV, to ease the driving requirement for all the parallel channels. This leads to a minimum gain of 51 dB, neglecting the filter insertion loss. On the other hand, the desirability to keep each channel as simple as possible calls for a low-gain design, since any increase in the cost of a channel will be multiplied by a large number in a multichannel system. It turned out, however, that the cost argument did not prove as serious as anticipated.

A number of designs, essentially by C. Cooper, were tried. The one finally adopted is shown in figure 8. It consists of two identical halves, each comprising a grounded base input stage, a common emitter stage, and an emitter follower. The filter is placed between the two subunits. The output impedance is 25 Ω, the input impedance for the circuit in figure 8 is about 45 Ω for each of the two halves. The input impedance is determined by R_1, and to some extent by C_2.

The frequency response is determined mainly by the decoupling capacitors C_1 and C_3, which determine the negative feedback. The desired response in this case is obtained by suppressing the low frequencies by a suitable amount, a method which of course wastes the high inherent gain at low frequencies. The response curve with the filter replaced by a 3.5 dB pad is shown in figure 9. The gain was found to be 60 dB.

The position of the filter after a three-stage amplifier gives a very good isolation between adjacent channels.

The transistor 2N 706 has a permissible free air power dissipation of 300 mW. With a heat sink the rating increases to 2 W. The emitter follower transistor dissipates 200 mW DC, which is within the power rating, although a heat sink would provide additional security. The emitter follower stage of the first amplifier half does not need the output capability of the last part. It can be rebiased for lower power consumption, although it is not done here.

Figure 10 shows the input impedance vs. frequency for the circuit in figure 8, and in figure 11 the temperature dependence of the gain is given. In appendix 3 the cost for each channel is estimated.

Since a fairly long signal cable between the driver unit and the various channels is anticipated, it is suggested that the driving unit contain one emitter follower for each channel. The emitter followers can all be driven from a single source (also an emitter follower). The power requirements for this source will be very moderate. Assuming an input impedance of 200 Ω for each of the emitter followers, and a required signal level of 100 mV, 100 channels, say, will need $100 \cdot \frac{0.1^2}{200} \cdot 1000 = 5$ mW.

Several companies were asked to provide filters. The lowest quotation was given by McCoy Electronics Company, with a price of $35 per filter in quantities over one hundred. Previous experience with this company has been positive.

b) Contiguous comb crystal filter.

Another possibility considered for the channel IF section is a contiguous comb filter set, proposed by Damon Engineering. All the filters are immediately adjacent to each other without any resistive isolation in between, and are driven from a common source through a reactive correction network. Virtually all power delivered by the source is dissipated in the filter loads, the main loss being the filter insertion loss.

The absence of the isolation problem, together with the very favorable drive situation, indicated the possibility of eliminating the individual channel IF amplifiers completely. This would increase the reliability of the system by getting rid of an appreciable number of transistors and components. The only difficult point seems to be the direct matching of the filter to the detector circuit. The output impedance of the filter must be low (less than 100 Ω); otherwise it is not possible to obtain a square law detector characteristic; on the other hand the input impedance of the detector is about 1000 Ω. The attractive feature of a transistorless channel IF amplifier had to be abandoned, and a circuit containing an emitter follower output stage (see fig. 12) was adopted. Damon Engineering proposed to build the filter bank, output stage and detector circuit for $246 per channel (they bid on a 127-channel system). They would also supply the driver unit and the correction network.

The cost of a system using the contiguous comb filter design will be about $136 per channel higher than for the conventional design described earlier.

Digital Output System

The digital output system will be provided by A. Shalloway and his group. The block diagram is shown in figure 13. The DC output signal from each channel is fed to a wetted mercury relay tree, which performs the scanning by means of suitable control pulses at a rate of about 3 milliseconds per channel. The operational amplifier increases the level to one appropriate for the analog-to-digital converter, and also provides a very high input impedance. In the DIT (data input transfer) unit additional information (time, position, frequency, etc.) is supplied to the data, and is recorded along with it on magnetic tape via the scanner. The last three units comprise the NRAO standard telescope output system.

The remaining features of figure 13 will be discussed in a little while.

Monitoring of Receiver Output

In addition to the recording of the output data on magnetic tape a visual monitoring system is necessary for checking the performance of the receiver. It is also desirable to have a permanent record of the observations in analog form. With a very large number of channels it is clearly impractical to use an analog multipen recorder (or recorders), nor is it practical (or possible, in view of the high scan rate) to record the channel outputs sequentially on a single recorder.

A simple way of achieving an analog display is to feed the output of the operational amplifier to the y-plates of an oscilloscope with a long persistance screen. In synchronization with the scanning of the channels the electron beam is displaced in the x-direction, see fig. 13. In this way the instantaneous power spectrum observed is made visible, and can be photographed whenever a permanent record is wanted. Time and date can easily be included. By having the x-sweep increase linearly with time each channel will make a short horizontal line. A spot representation of each channel is obtained by letting the x-voltage increase in steps. Magnification of a certain part of the spectrum is easily done, too. Provision should be made for starting the scanning of the channels at arbitrary times in order to facilitate testing. Negative marker pulses on every ten channels will provide an easy means for finding a particular channel on the oscilloscope screen.

The details of the system just described will not be elaborated upon here.

The detector voltages in the individual channels should also be monitored, and a jack be provided in each channel for this purpose. Variations in detector output voltage between the channels will take place for the following reasons: 1) RF + main IF bandpass not flat, 2) individual channel amplifiers not exactly identical (the conventional design), 3) filter insertion loss variable, 4) detector efficiency variable. With the square law detector described above it is desirable to keep the operating levels as equal as possible in order to achieve a uniform performance.

While it is possible to get an accurate reading on a meter of the various detector levels, it is obviously impossible to get an overall picture of all the channels by plugging into a jack one at a time, or by throwing an n-pole switch, or by reading n simultaneous meters (n is the number of channels). By providing a separate set of mercury relays (see fig. 13) all the detector voltages can be scanned and displayed on an oscilloscope in a similar manner as the signal output described before. A time constant of about 1 sec is used to smooth the detector voltage. It will not be necessary to keep a continuous record of the detector voltages, so the same oscilloscope as for the hydrogen line display can be used, by just throwing a switch temporarily.

There is one obvious difficulty connected with the detector levels in this receiver. (as in all low-noise receivers). During the half cycle of the switching when the comparison frequency band is received the 100 °K system noise will give a detector voltage of 100 mV (the chosen operating point). At the maximum hydrogen line frequency an additional 120 °K will be received and the mean detector voltage will increase to 160 mV. It is recognized that this in effect constitutes one possible alternative to the line display, described earlier, although of inferior quality. At any rate, it shows that the detector voltages should be adjusted to their operating levels with the antenna pointing towards a cold part of the sky.

The arrangement of scanning the detector voltages has an additional advantage. By connecting the detector outputs (via an operational amplifier) to the A/D-converter, the voltages can be registered on magnetic tape. Changing the noise level in the main IF amplifier chain with an attenuator, and recording the outputs, will give all the detector characteristics at once. The results can easily be processed on the computer.

In this way the exponent α for each detector circuit can conveniently be checked as often as desired, and possible corrections to the observations of the hydrogen line be applied (if α is too far from unity). It is not necessary to worry about the sky background, since the front-end can be disconnected and replaced by a termination or a booster amplifier (to keep the noise level high enough).

The positioning of the time constant nets for the signal and the detector outputs indicated in figure 13 is in order to decrease hum pick-up.

A rough cost estimate for a complete 127-channel receiver is given in appendix 4.

References

[1] Muller, C. A., and Westerhout, G., "A Catalogue of 21 cm Profiles", B.A.N., 13, 151-195 (No. 475), 1957.

[2] Bare, C., "Power Detector", NRAO Electronics Division Internal Report No. 24, February 1964.

[3] Hvatum, H., "Detector Law", NRAO Electronics Division Internal Report No. 6, December 1962.

APPENDIX 1

Table of possible interfering frequencies generated in the receiver for different 2nd LO frequencies. Combination frequencies falling within 2 MHz of 1st or 2nd IF band center are underlined.

2nd IF Band	1st LO / 2nd LO	13.9	27.8	41.7	55.6	69.5	83.4	97.3
7.0 - 9.0	22	8.1	5.8	19.7	33.6			
6.5 - 8.5	22.5	8.6	5.3	19.2	33.1			
5.8 - 7.8	23.2	9.3	4.6	18.5	32.4			
5.5 - 7.5	23.5	9.6	4.3	18.2	32.1			
5.0 - 7.0	24	10.1	3.8	17.7	31.6			
	44	30.1	16.2	2.3	11.6	25.5	39.4	
	45	31.1	17.2	3.3	10.6	24.5	38.4	
	46.4	32.5	18.6	4.7	9.2	23.1	37.0	
	47	33.1	19.2	5.3	8.6	22.5	36.4	
	48	34.1	20.2	6.3	7.6	21.5	35.4	
	66		38.2	24.3	10.4	3.5	17.4	31.3
	67.5		39.7	25.8	11.9	2.0	15.9	29.8
	69.6		41.8	27.9	14.0	0.1	13.8	27.7
	70.5		42.7	28.8	14.9	1.0	12.9	26.8
	72		44.2	30.3	16.4	2.5	11.4	25.3
	88				32.4	18.5	4.6	9.3
	90				34.4	20.5	6.6	7.3
	92.8				37.2	23.3	9.4	4.5
	94				38.4	24.5	10.6	3.3
	96				40.4	26.5	12.6	1.3
	110						26.6	12.7
	112.5						29.1	15.2
	116.0						32.6	18.7
	117.5						34.1	20.2
	120						36.6	22.7
	132							34.7
	135							37.7
	139.2							41.9
	141							43.7
	144							46.7

APPENDIX 2

Cost estimate for audio circuit (in quantities larger than 100).

White active filter ------------------------	$ 32.00
Signal transformer ------------------------	6.50
Chopper --------------------------------	25.00
Potentiometer ----------------------------	1.00
4 Capacitors ------------------------------	.50
Connectors ------------------------------	10.00*
Printed board plus mounting of parts ---------	25.00*
	$100.00

* Estimated cost.

APPENDIX 3

Cost estimate for the conventional filter channel (in quantities larger than 100).

26 Resistors --------------------------------	$ 1.00
17 Capacitors -------------------------------	.75
6 Transistors (2N 706) ----------------------	7.25
Connectors, jack --------------------------	10.00*
Potentiometer ------------------------------	1.00
Filter --	35.00
Detector circuit ----------------------------	5.00
Printed board plus mounting of parts ----------	50.00*
	$110.00

* Estimated cost.

APPENDIX 4

Cost estimate for a complete 127-channel receiver.
Numbers within parentheses are cost for the Damon design.

Front End:	Parametric amplifier	k$15		k$	
	Mixer-preamplifier	2			
	LO (Manson)	8			
	Multiplier	5			
	Other	3		33	
IF and Channels:	IF amplifiers, etc.	3			
	127 channels x 210 (346)	27	(44)	30	
	(Driver for Damon filters)		(3)		(50)
Switching:	Pulse generator, square wave generator, etc.	3		3	
Output System:	Scanner	2			
	A/D converter	2			
	Logic	3			
	Multiplexer	2			
	Memoscope	3			
	Camera	1			
	Other	2		15	
Power Supplies:		4		4	
				k$85	(105)
			+10%	10	10
			k$	95	115

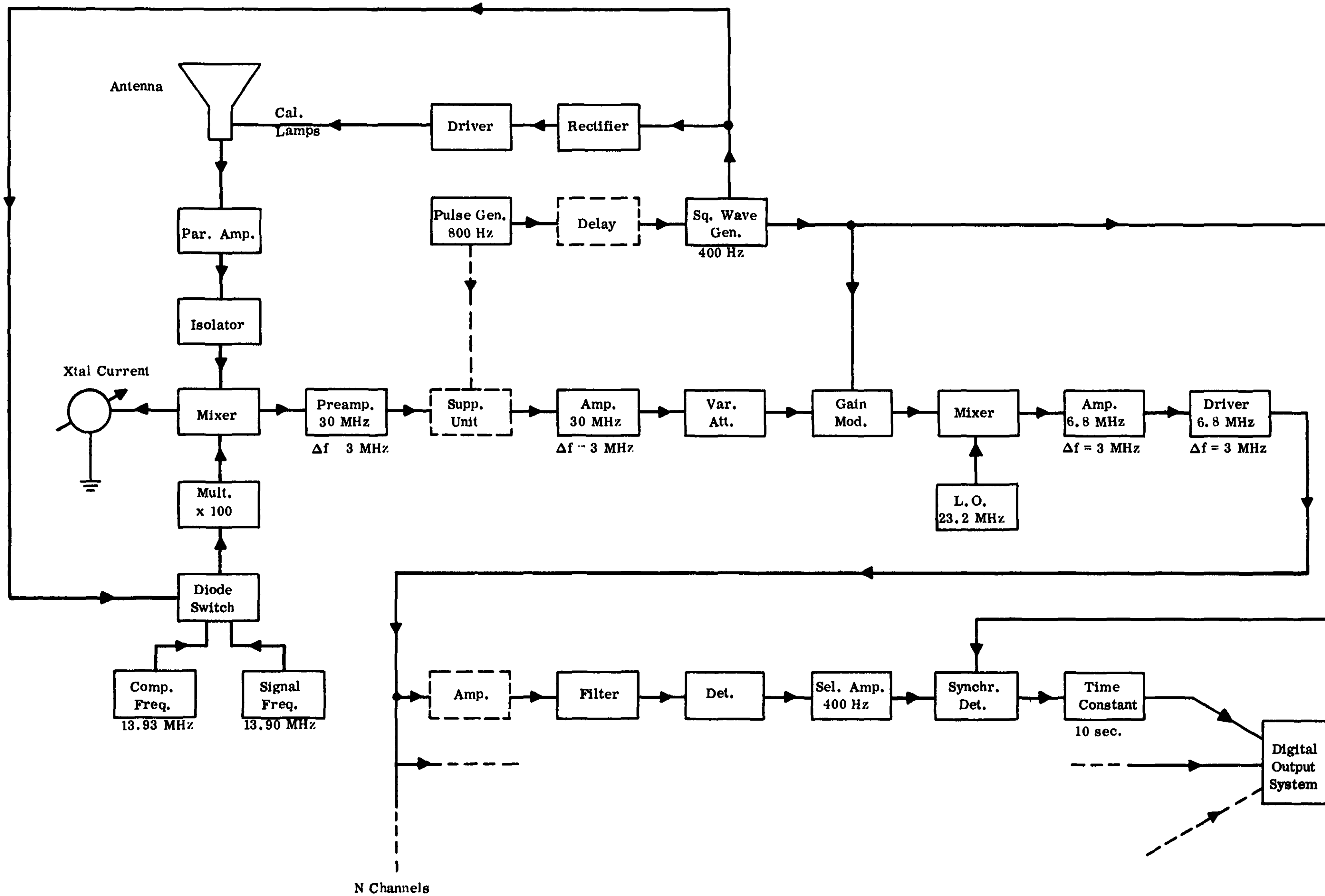


Figure 1

Block diagram of multifilter receiver.

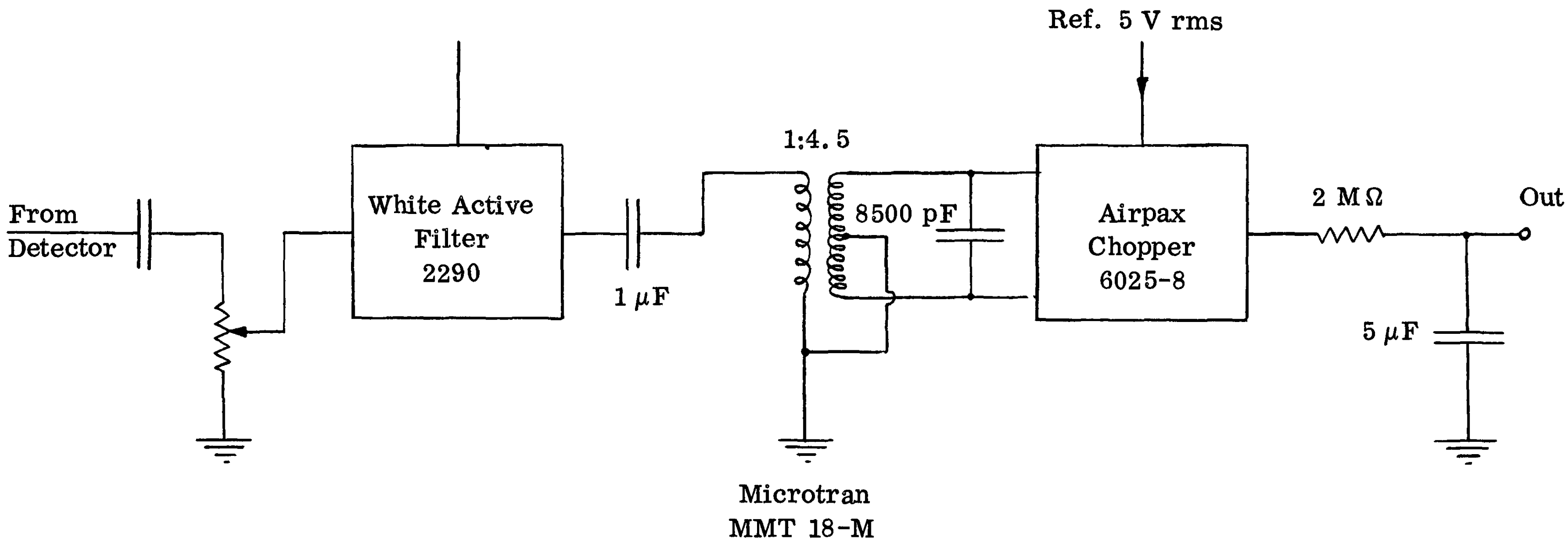
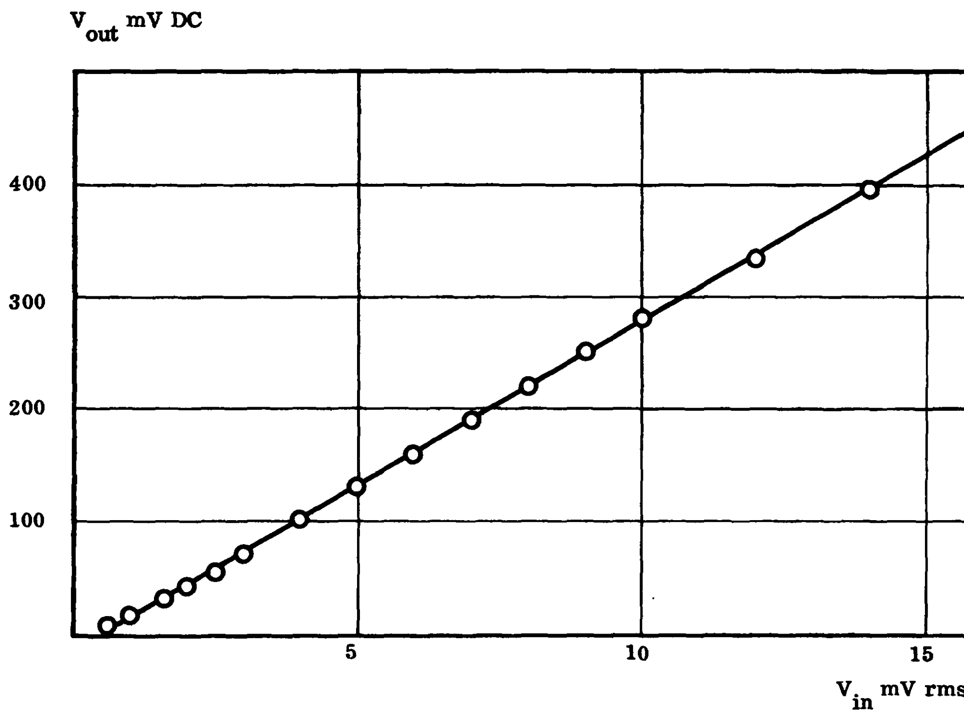


Figure 2. Audio Section Circuit



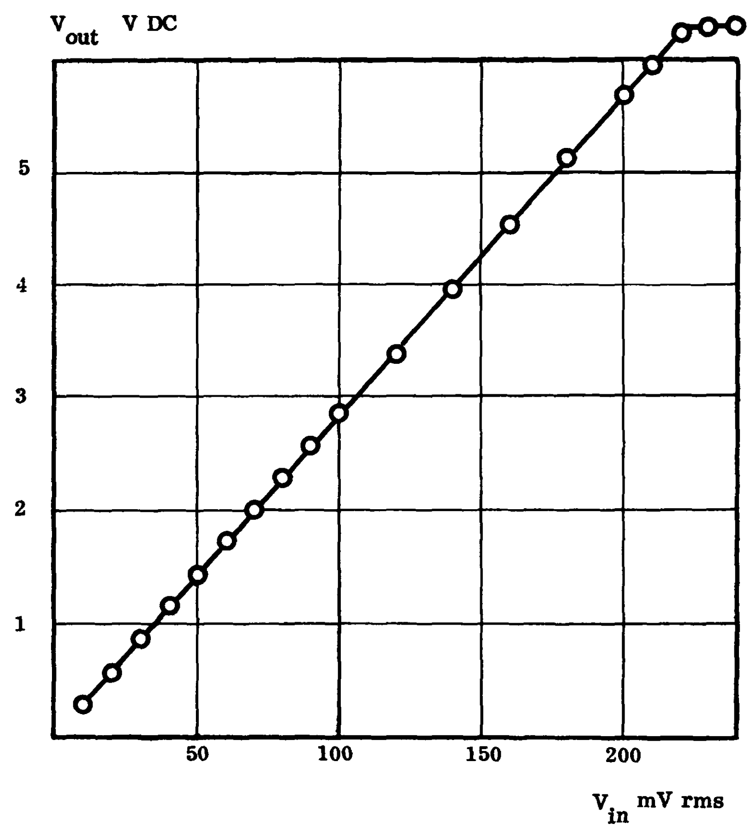
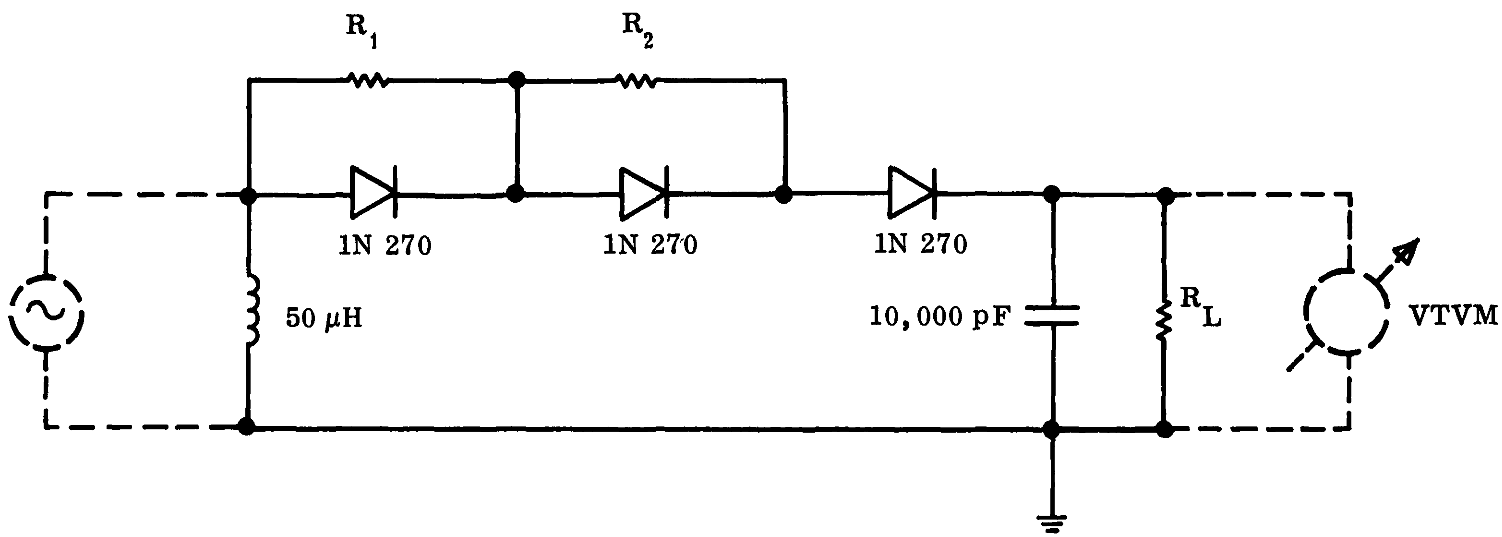


Figure 3

Input-output characteristics of the audio unit in figure 2 at small input levels (top) and large input levels (bottom).



Final Values:

$R_1 = 2.2\ k\Omega$

$R_2 = R_L = 1.0\ k\Omega$

Figure 4

Detector circuit used.

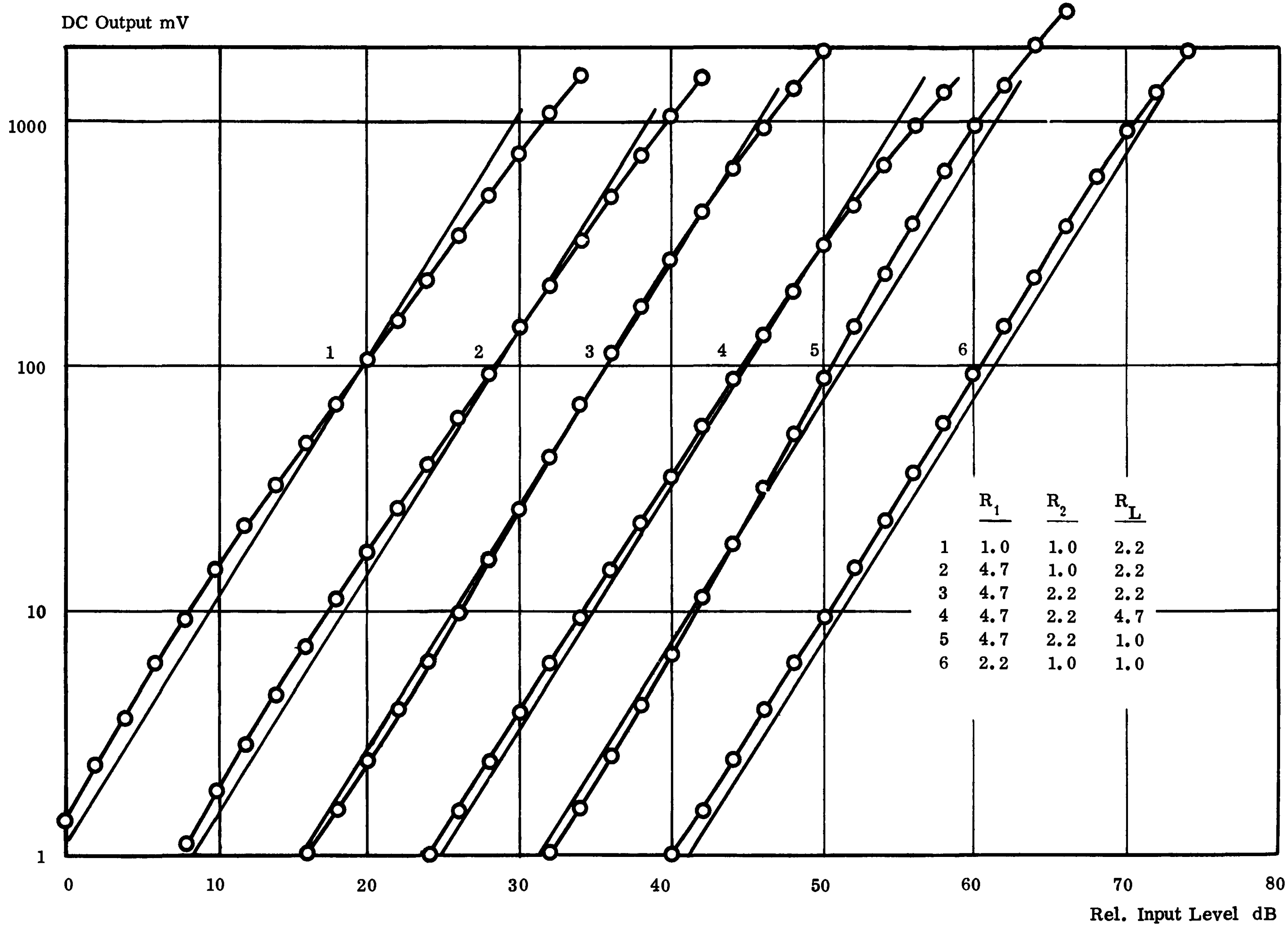


Figure 5

Detector law plots for various circuit parameters of the detector in figure 4. The curves are displaced horizontally for clarity. The slope of straight lines added for reference corresponds to a perfect square law.

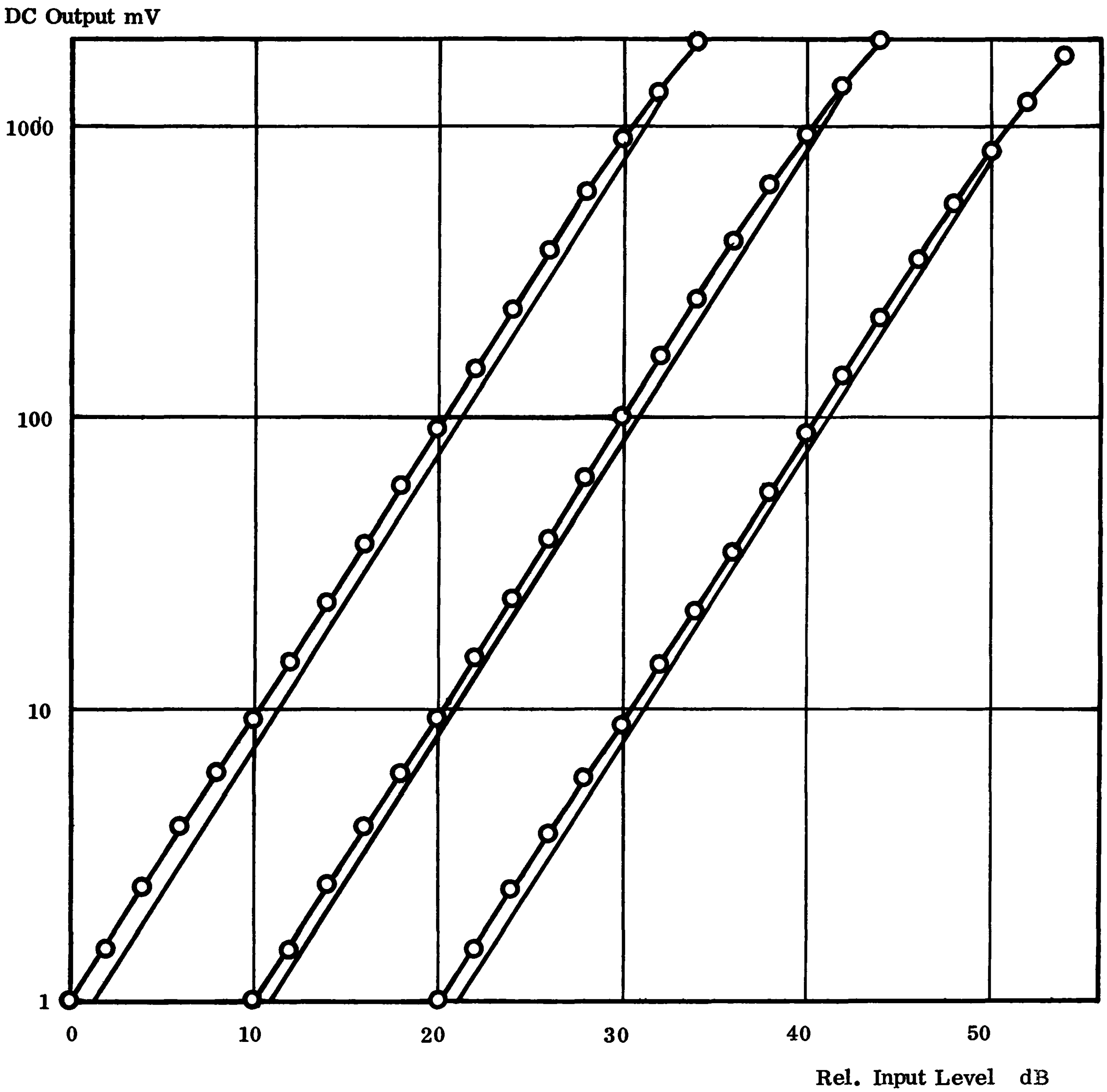


Figure 6

Detector law plots obtained with constant circuit parameters for three different sets of diodes.

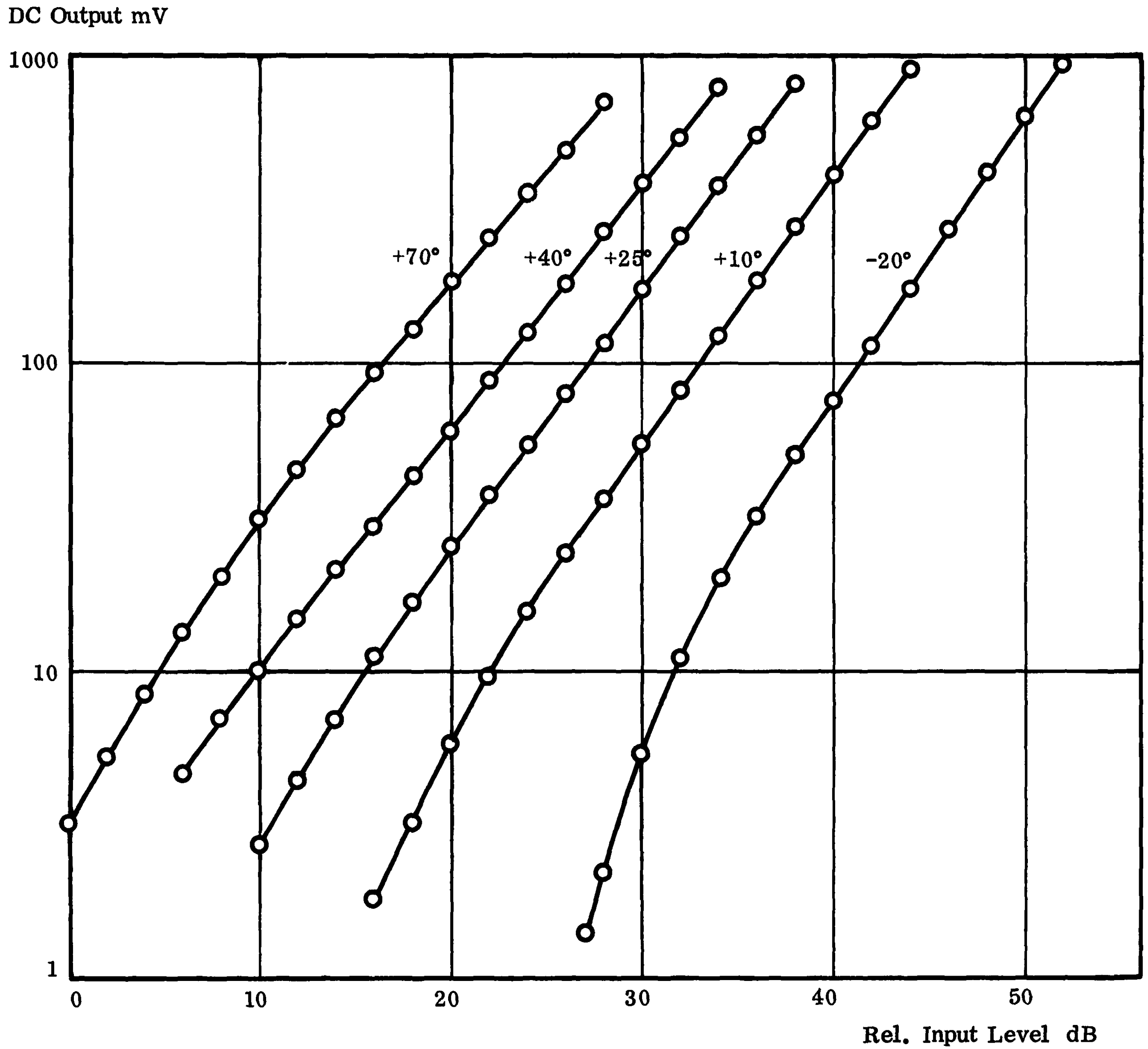


Figure 7

Dependance of detector law on temperature.

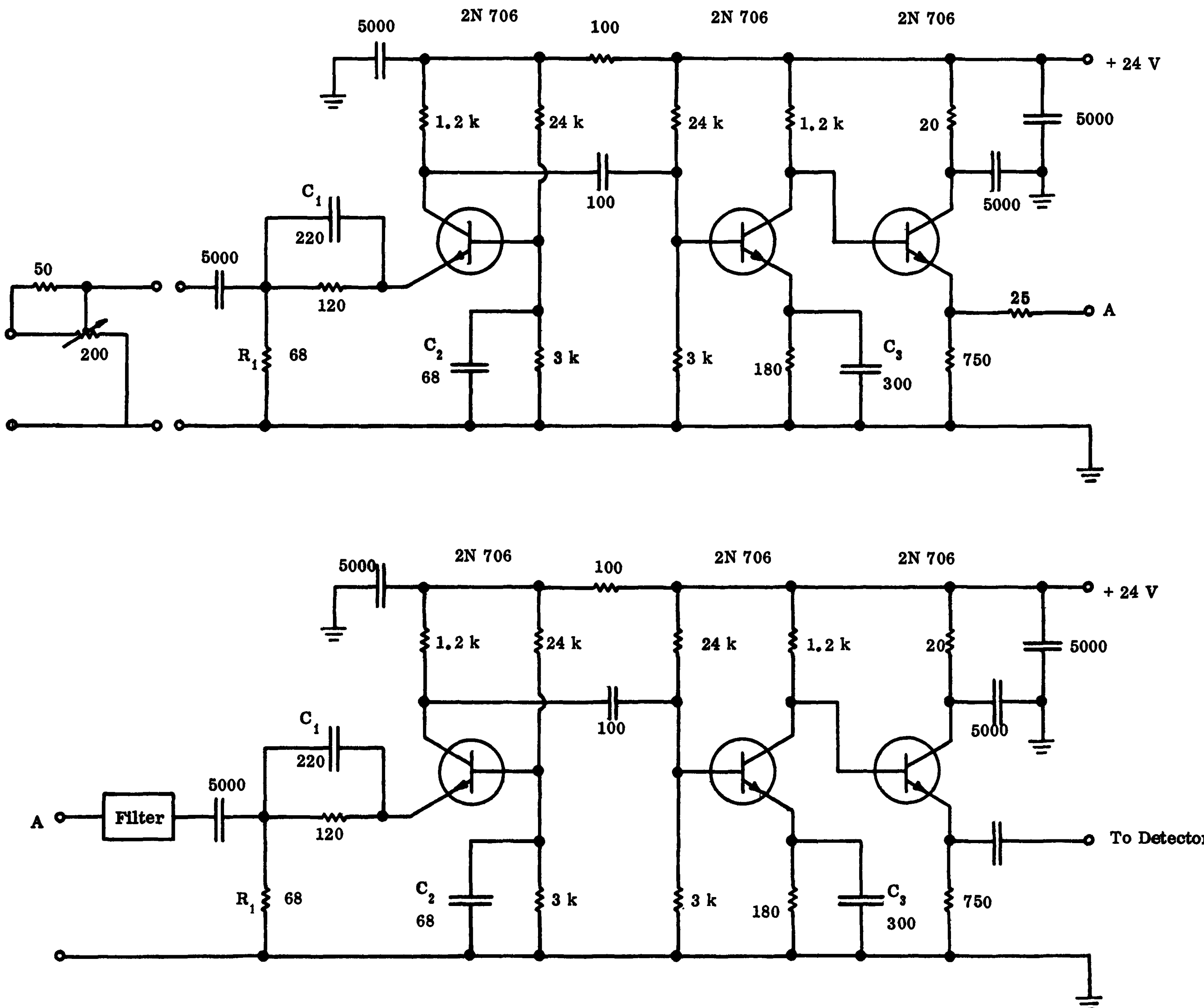
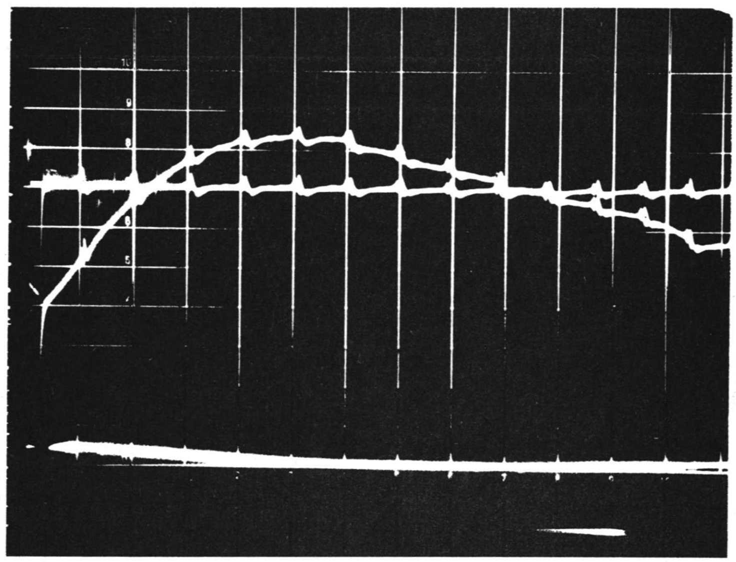


Figure 8

Channel IF amplifier circuit.

7 MHz



- 1 dB

Figure 9

Frequency response of channel IF amplifier without filter. Distance between frequency markers 1 MHz.

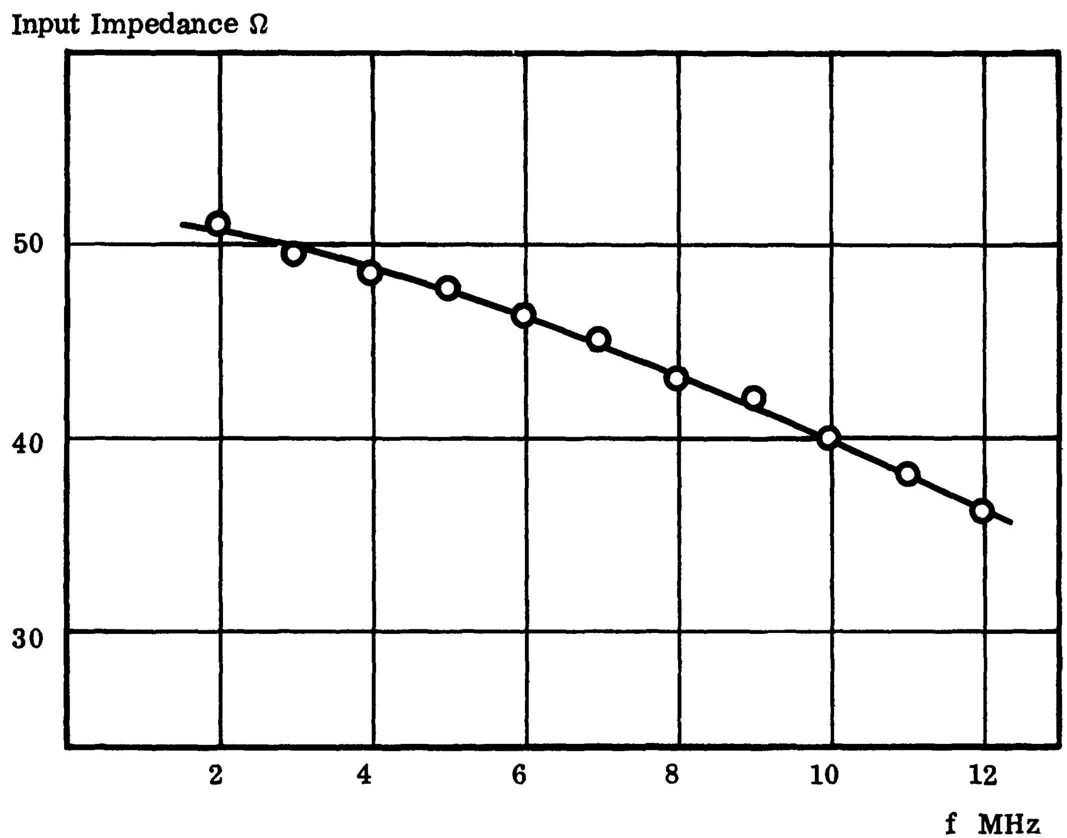


Figure 10

Input impedance vs. frequency for channel IF amplifier.

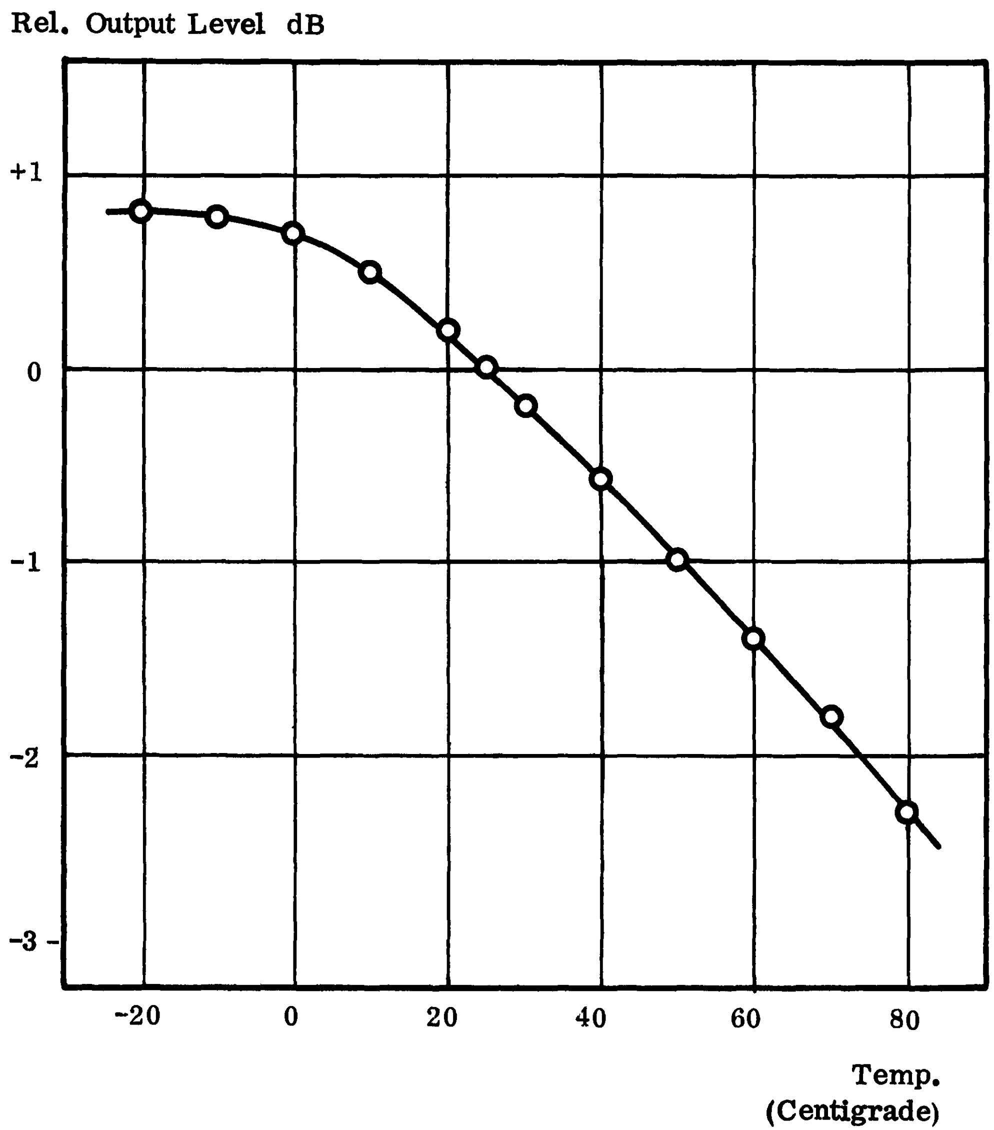


Figure 11

Temperature dependance of channel IF amplifier gain.

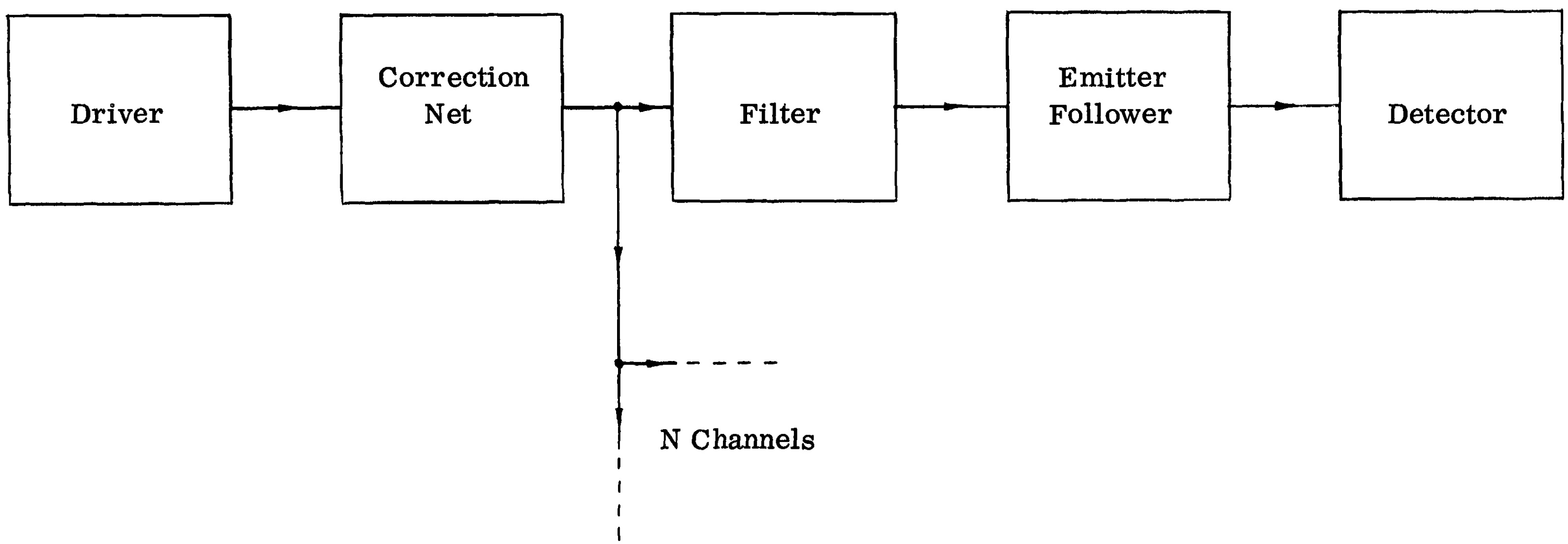


Figure 12. Block diagram of channel IF section with contiguous comb filter.

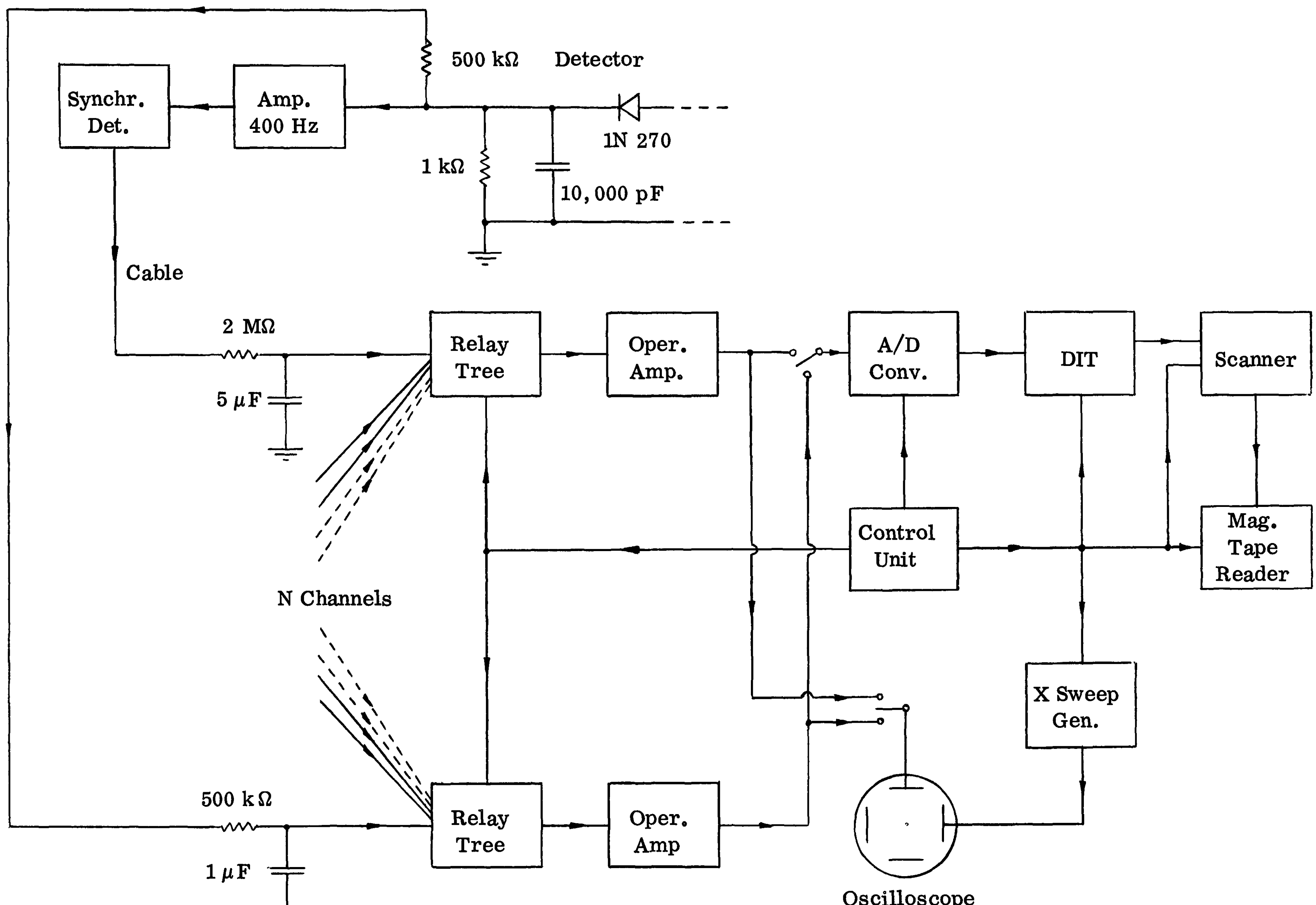


Figure 13. Digital output system and monitoring circuits