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FINAL ENGINEERING REPORT - 1962

Project 100-1

Phase I

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Engineering Director

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LIST OF ILLUSTRATIONS

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1.0 WORK STATEMENT

1.1 Conduct design and development program to convert into hardware the fixed base station communications system design and feasibility tested under Task Order No. 1, [redacted] 25X1
[redacted] This task has been modified from 25X1
the original, requiring a more sophisticated package and a more complex coupling technique. For details refer to Sections 2.2.1 and 2.5.

1.2 Fabricate and test one (1) each simplex system consisting of one (1) transmitter and one (1) receiver based on the development program conducted under Item 1.1 above.

1.3 Fabricate and test three (3) each transmitters and four (4) each receivers identical to the model constructed under Item 1.2 above.

1.4 Provide five (5) copies of each monthly letter progress report.

1.5 Provide five (5) non-reproducible and one(1) reproducible copies of engineering drawings describing the equipment fabricated under items 1.2 and 1.3 above.

1.6 Provide ten (10) copies of operation and maintenance handbooks for the equipment fabricated under items 1.2 and 1.3 above.

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2.0 SYSTEM DESIGN

2.1 System Function

The function of System 100-1 is to provide digital communication between two parties utilizing conventional public telephone system as the transmission medium. The message is sent during the course of a normal telephone conversation using the originator's voice [redacted] 25X1 25X1

The system concept is to convert each character of the message into a combination of one to five audio tones. These tone combinations are added to the speaker's voice and transmitted whenever the voice level exceeds a preset threshold. Since tone levels are far below the instantaneous voice power, the voice masks the tones, making them inaudible to anyone listening on the line. All five bits of each character are transmitted in parallel to eliminate synchronization problems, a tone representing a "mark" and the absence of a tone a "space" on the teletype code. Since each tone is at a level considerably below peak speech levels, it is necessary to notch out the speech spectrum around each tone frequency prior to transmission to avoid mistaking voice components for tones at the receiver. In the present system, the five tones are closely spaced requiring one band reject filter to eliminate the undesired spectral components of the voice.

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2.2 Specifications

a. Transmitter

Input: (Telephone Instrument)

DC Source Resistance: approximately 200 ohms
handset off the hook)

AC Source Impedance: 600 ohms (nominal)

Level*: -6 to -26 dbv rms

Frequency Range: 300 to 3400 cps

Output (Telephone Lines)

DC Resistance: Approximately 200 ohms (when
handset is off hook)

AC Impedance: 600 ohms (nominal)

Level*: Voice: -16 dbv $\pm$ 1.5 db

Tone: Adjustable, -40 to -46 dbv

Frequency: Speech: 300 to 3400 cps notched between
1250 and 1950

Tone Characteristics: Tones: 1400, 1500, 1600, 1700,
1800 cps. 40 millisecond
duration, isoleles triangle
envelope

Controls and Indicators

Power: OFF-ON (toggle type circuit braaker)

Standby - Transmit (toggle switch)

Power supply protection (push to reset circuit breaker)

Amber Light: Power indicator

Red Light: End of message indicator

Power requirements

a. 90 to 126 VAC, 45 to 60 cps, 0.6 amperes, nominal, or

b. 180 to 252 VAC, 45 to 60 cps, 0.3 amperes, nominal

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*Level shown is long term average of speech in decibels above 1.0 volts rms.

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Keying Rate: Adjustable from 0 to 100 words/minute during continuous speech (set at factory for 25 wpm, nominal)

Message Tape: 11/16 inch teletype tape (5 level), chadded. Capacity 60 feet (1200 words) maximum

Monitor: A separate earphone is provided to permit local monitoring of the distant speaker's voice.

Mechanical Characteristics

Dimensions (maximum) - 15 inches wide, 9 inches deep, 10 inches high

Weight: - 45 lbs.

Finish: Semi-gloss black enamel

Accessories

220 V motor for tape reader (1 each)

Card Extender (1 each)

Fuses - AGC - 1/4, 3 each

Environment

Temperature 0° to $+50^{\circ}C$

Humidity: 0 to 95% relative humidity

Shock and Vibration: Air transportable

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b. Receiver

Input (Telephone Line)

DC Resistance: Approximately 200 ohms

AC Impedance: 600 ohms (nominal)

Frequency Range: Voice: 300 to 1250 cps and 1950 to 3000 cps
Tones: 1400, 1500, 1600, 1700 and 1800 cps

Level: Voice: -16 to -36 dbv (long-time average of rms voltage)
Tones: -43 to -63 dbv (peak rms voltage)

Tone Characteristics: 40 millisecond duration, isosceles triangle envelope

Output A (to telephone line)

DC Resistance: Approximately 200 ohms

AC Impedance: 600 ohms (nominal)

Frequency Range: Voice: 300 to 1250 cps and 1950 to 3000 cps
(No tones present)

Level: -6 to -26 dbv rms (long-time average of rms voltage)

Output B: Punched paper tape 11/16" wide, 5 level teletype, chadded

Controls and Indicators

Power: OFF-ON (Toggle switch)

Function: Normal - Standby- Receive (rotary switch)

Power supply protection: Push to reset circuit breaker (fuse on power supply for 220 VAC)

Gain Control: Operator sets gain of receiver using VU meter

VU meter: Indicates level of voice signal at input to receiver amplifier

Amber Light: Power "ON" indicator

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Power Requirements

a. 90 to 126 VAC, 45 to 60 cps, 2 amperes nominal, or

b. 180 to 252 VAC, 45 to 60 cps, 1 ampere nominal

Information Rate: Determined by transmitter rate - 25 wpm nominal, 250 wpm maximum

Mechanical Characteristics

Dimensions: Receiver: 21-9/16" wide, 16" deep, 13" high

Perforator: 21-9/16" wide, 17-7/8" deep, 13" high

Weight: Receiver:75 lbs.

Perforator: 79 lbs.

Finish: Hammartone Gray cabinet; clear anodized front panel

Accessories

Power Adapter (1 each)

3 prong to 2 prong power adapter

Card Extender (1 each)

Fuses: 2 amp - 5 each
1/4 amp - 3 each

Microphone: High impedance lapel microphone

Environment

Temperature: 0 to +50°C

Humidity: 0 to 95% relative humidity

Shock and Vibration: Air transportable

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2.3 System Design Considerations

2.3.1 General

An important system consideration is the error and false alarm rates. The error rate is a function of the number of times a transmitted tone, corresponding to a one-bit, is recognized as a zero-bit. This situation arises when sufficient noise signals (undesirable signals) are present in the tone channel and have such a magnitude and phase that when added to the transmitted tone, produce a signal which is less than the established threshold in the receiving equipment. The false alarm rate is concerned with the number of times signals due to noise have sufficient magnitude to exceed the threshold in the receiver.

A second important system consideration is the transmission rate. This parameter is a function of the statistics of speech and the threshold level established in the transmitter relative to the long-time average power level of the speech. The transmission of one bit occurs in 40 ms, corresponding to a maximum rate of 250 wpm. To provide maximum security of the system, transmissions only occur when the instantaneous voice energy exceeds a preselected threshold level. Whenever the speech power exceeds this threshold and the equipment is not in the process of transmitting a bit, transmission of a bit is initiated. At the end of the 40 ms required for this event, transmission of the next bit depends upon the threshold condition above being satisfied. At the time of this writing, statistical information as to the number of times that a given threshold level above the mean speech power is exceeded has not been available. Further, no theoretical means has been found to extract this information from the percentage of time that speech power will exceed a given level referred to the average speech power.

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a. Signal and Noise Levels

This section includes a description of the noise components considered and estimates of their relative power levels. It also includes the various signal levels in the system which are pertinent to determining these noise component levels.

The sources of noise considered are as follows:

A. Distortion components

B. Thermal Noise

C. Speech components which pass through the band reject filter in the transmitter and bandpass tone filters in the receiver.

D. Telephone line crosstalk

Using 50 db as the total useful dynamic range of the telephone line, and 0 db as the average speech level, a maximum level of 12.5 db corresponds to peak limiting speech signal 1% of the time. This amount of peak limiting will not noticeably affect the quality of speech. For the power levels indicated above, line noise for a 3-kc bandwidth is at -38.5 db and line noise for a 100-cycle per second bandwidth is at -52.5 db. A threshold level in the transmission equipment of +10 db relative to the 0 db average speech signal was used during the feasibility experiments and is used in the analysis. This threshold level is the magnitude the speech signal must exceed to initiate tone transmissions. The tone signals are established at 26 db below the transmission threshold, or -16 db relative to the average speech signal. It should be noted that the line noise is -34.5 db below the tone level (using the 100 cps bandwidth value) and has negligible effect on error and false alarm rates.

The attenuation of speech components through the band reject filter in the transmitter and bandpass tone filter in the receiver is shown in Figure 2.3.1.a. Only the output of a tone filter

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adjacent to the edge of the band reject filter can contain signals of any significance. The composite response in Figure 2. 3. 1a shows, however, that these components are attenuated by at least 60 db and need not be further considered.

Cross-talk between telephone pairs is generally maintained at about -30 db. Taking into account the 100-cycle bandwidth of the tone filter, the cross-talk voice energy would be at about -45 db and would not contribute to the error or false alarm rates. On the other hand, the presence of the cross-talk of signals due to tones used by the telephone company can be a serious problem if the tone frequencies fall within the passband of the tone filters in the receiving equipment. This problem can be avoided by careful selection of tone frequencies.

The most significant noise components result from signal distortion produced in amplifiers contained in either the transmitting or receiving equipment or in the telephone line itself. It is assumed that the total distortion will be kept to within 1% corresponding to at least 40 db below the instantaneous speech power.

It thus appears that distortion components are significant in affecting the false alarm and error rates and will now be considered in greater detail. To accomplish this, we will establish the mean power level of these components occurring during the transmission of the tones as well as a statistical model for this source of noise. From this information, we can determine the error and false alarm rates per bit and per character. In addition we can also determine the level of accuracies which must be maintained for a given character error and false alarm rate and thus derive how accurate the AGC's must be.

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b. Bit and Character Error and False Alarm Rates

For 1% total harmonic and intermodulation distortion, the total power in these components is 40 db below the speech power. The mean speech power during transmission of the tones is determined as follows: A transmission is initiated only when the speech power exceeds the +10 db level. However, the speech power will exceed this level 2.8% of the total conversation time. Assuming that 10% of the telephone conversation time can be used in transmitting the digital message, this corresponds to 28% of the transmission time. If we also consider correlation existing between signal samples, the most pessimistic conditions exist if the speech power is always greater than the level exceeded 10% of the time. The speech power distribution thus occurring during ttansmission is considered to vary between 5.6 db and 12.5 db relative to 0 db for the average speech power. Figure 2.3.1b shows the speech power level during tone transmission and the percentage of time it is exceeded varying from 100% at 6.5 db to 0% at 12.5 db. The unprocessed speech signal will actually contain components whose power exceed this 12.5 db level. However, the limiter action in the transmitter has removed these components. Taking the weighted mean of the power as a function of the percent time it is exceeded, an average power level of speech signal during tone transmission of 8.5 db is obtained.

During transmission, the mean energy in the distortion components for 1% total distortion is 40 db below the mean speech power. Although the statistics of the speech is not Gaussian, the statistics of the distortion components can be considered as Gaussian because, in general, the total distortion signal is made up of the sum of many components. The central limit theorem states that the statistics of a signal made up of the sum of many components tends

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toward Gaussian regardless of the statistics of the individual components. The mean power level of the distortion components is 8.5 db -40 db = -31.5 db. The tone levels are at -16 db providing a mean signal-to-noise power ratio of 15.5 db. This is the signal-to-noise ratio used in computing the error and false alarm rates. It will be noted that the reduction in noise level resulting from the use of the 100 cps tone filter is not taken into account, and it is assumed that the distortion components will all lie in the passband of the tone filter. It will be shown that even with this pessimistic model, the error and false alarm rates which can be achieved are practical. The output of the 100 cps bandpass filter is supplied to a detector converting the Gaussian statistics of the signals to a Rayleigh distribution with noise alone and modified Rayleigh with signal and noise. The output of the detector is supplied to a lowpass filter with its cutoff at 50 cps. This filter acts as an integrator producing a signal whose statistics are that which would be attained by summing independent samples at the input to the lowpass filter in groups of two. This results in a distribution as indicated in Figure 2.3.1c where σ is the rms noise at the input to the detector, and where the mean value and standard deviation of the noise alone are 2.5σ and $.928\sigma$ respectively and where the peak signal and standard deviation of the noise distributed about this signal are 17σ and 1.414σ respectively. The magnitude of the signal for three standard deviations (corresponding to the signal level which will be exceeded 0.1% of the time) is 5.28σ and is a case for no tones present. The magnitude of the signal corresponding to the case for tones present and the noise subtracting from the signal and three standard deviations is 12.76σ. A threshold level located anywhere between 7.46σ and 12.28σ will produce a bit error or false alarm rate which is less than 0.1%.

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We now wish to establish the magnitude of the threshold level which will allow maximum gain variations in the system. Letting L = the value of this threshold level, and letting K be the variation in voltage gain, $(5.28\sigma K + L)$ corresponds to the case where the gain is increased to the point where the false alarm rate is 0.1%. On the other hand $\frac{12.76\sigma}{K} = L$ corresponds to the case where the voltage gain has been reduced to the point where the error rate is 0.1%. Solving these two equations for K, a value of 1.55 is obtained corresponding to $\pm$ 3.8 db. This is the voltage gain variation which can be tolerated and still provide a bit error and false alarm rate which is 0.1% or less. Using the value of K indicated above, a threshold level of 8.22σ is obtained, corresponding to a threshold located 6.3 db below the peak tone level. This value of threshold level provides for a maximum variation in both increasing or decreasing the gain from the optimum value.

The character error rate is obtained from the bit error rate assuming a 0.1% error and false alarm rate using the Bernoulli Formula. The character error rate is determined based on having one or more failures or errors in five bita and is given by

$$\sum_{S=1}^{5} {}^{5}C_{s} \quad p^{s} q^{5-s} \qquad = 1 - {}^{5}C_{o} \quad p^{o} q^{5}$$

$$= 1 - q^{5}.$$

$$q = .9987$$

so that the character error rate is 0.64%.

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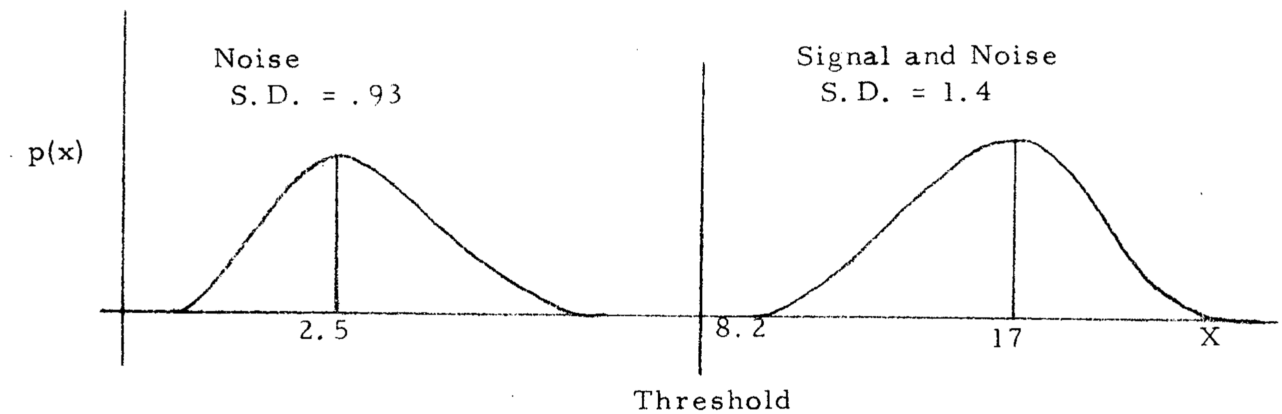

Input S/N to detector of 15.5 db with 100 cps bw detector followed by low pass filter with 50 cps cutoff. Gain variations of $\pm$ 3.8 db of input signal level will result in bit error and false alarm rates of 0.1 percent.

Figure 2.3.1.c

PROBABILITY DENSITY FUNCTION

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2.3.2 Transmitter (See figure 2.3.2)

a. Coupling to Line

The technique of using hybrids to couple the transmitter between telephone and line was extensively explored both in theory and laboratory experiment. The problem of high transmitter gain (to provide good AGC control) was recognized from the beginning of the program. Another problem discovered in laboratory tests is the difficulty of obtaining hybrid balance with the complex telephone instrument impedance. This is further complicated by the time variation in resistance of the telephone (carbon microphone). Even use of another similar telephone as the hybrid dummy load does not yield adequate balance.

An alternate technique using time multiplexing between transmit and receive channels was also tried. This circuit failed when connected to the complex telephone impedance even with substantial guard time intervals.

The final approach adopted for the Phase I equipment is to provide a separate earphone for monitoring the incoming voice signal at the transmitter.

b. Audio AGC

Gain control of the local speaker's voice is needed for two reasons. First, the level of signal fed to the line must be kept below a value that would result in important distortion in the telephone system but high enough to keep the tone levels well above the ambient telephone system noise. Second, a constant output level is desirable from the standpoint of maintaining a uniform information rate for a wide range of speakers and for different telephone operating techniques, i.e. methods of holding the handset. A practical range of levels to be expected for these

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variations is about 20 db.

In the design of the AGC circuit, several factors in addition to dynamic control range are important. The degree of control necessary must be specified since this determines the information rate versus input level characteristic.[1] From the statistics of speech, it may be observed that a very slight change in the speech level into the threshold circuit results in a large change in information rate. In fact, an increase in rms speech power of only 3 db in the vicinity of the 10% point results in a doubling of the information rate.[2] Therefore, it would be desirable to limit the range of inputs to, say, 0.1 db which would cause only a 3% change in information rate. Assuming a linear curve in the vicinity of 10%, however, the loop gain required to maintain the output this closely for a 20 db change in input is quite high resulting in poor transient response and complicating the loop filtering problem.[3] A compromise is therefore necessary in the design of the AGC circuit to achieve a clean output signal with reasonable control. With a loop gain of 60 db, a 20 db change in input results in a 1 db output change. Using LC filters in the loop (high pass and low pass) with 36 db per octave rolloff, distortion can be kept below 40 db. Transient response, however, is still poor due to the delay in AGC control. This results in a very high output for about 100 milliseconds after the start of the input

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[1] Another factor also affecting information rate is the talking rate and articulation of the speaker. Compensation of this factor, however, would require unnecessarily complex circuitry.

[2] See ITT Reference Data for Radio Engineers, P. 874

[3] Speech components above 300 cps must be attenuated sufficiently in the loop control signal before being applied to the gain control element. Otherwise, distortion of the output waveform will result.

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signal before the AGC takes control. The strong signal produces an undesirable "thumping" sound each time the AGC is operated. A technique for avoiding this effect is to employ a clipper circuit. Distortion products in the tone channel bands are removed by the band reject filter. The output of the audio AGC circuit is passed through a highpass filter with a cutoff frequency of 4000 cps to remove any undesired components that might exist outside the useful range of the telephone system.

c. Transmission Rate

It is desirable to predict the transmission rate as a function of threshold setting. As an upper bound, we consider the speech as uniformly distributed; that is, the probability of triggering the transmitter in any 300 microseconds (3 kc) interval is uncorrelated to what has preceeded. Clearly, because of the syllabic character of speech, this assumption is not valid, and the transmission rate we obtain will be an optimistic one serving to show the upper bound. In addition, since there are pauses in the speech, there are intervals of time when no transmissions can occur.

If we assume a trigger rate, λ , uniformly distributed, then the probability of triggering in an interval of time Δt is given by:

$$P_t = \lambda \Delta t \qquad (1)$$

Using (1), the probability of triggering in the kth interval is given by:

$$P_k = \left[(1 - \lambda)\ \Delta t\right]^{k-1} \lambda \Delta t \qquad (2)$$

Letting $t = k\Delta t$ and taking the limit of (2) as $\Delta t \longrightarrow 0$ & $k \longrightarrow \infty$ so that $k\Delta t = t$, we get

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$$P_t = \Delta t \rightarrow 0 \left\{ \lambda \Delta_t \left[(1 - \lambda) \Delta t \right]^{k-1} \right\}$$

$$= \lambda \Delta t \, e^{-\lambda t} \qquad (3)$$

Dividing (3) by Δt yields the probability density function

$$P(t) = \lambda e^{-\lambda t} \qquad (4)$$

The mean value of p(t) is given by

$$\bar{t} = \int_0^\infty t p(t) dt = \int_0^\infty t \lambda e^{-\lambda t} = \frac{1}{\lambda} \qquad (5)$$

We can solve for $\bar{t}$ using (5) λ to obtain

$$\bar{t} = \frac{1}{\lambda} = \frac{\Delta t}{P_t} \qquad (6)$$

$\bar{t}$ is the mean time between the end of one character and the start of the next character. If we denote the basic character length by then the total waveform period, T, is given by:

$$T = \quad + \bar{t} \qquad (7)$$

For our system, τ = 40 ms. We can solve (6) for $\bar{t}$ using the speech statistics of Figure 19 in the FTL handbook, p. 874. P_t is the probability the threshold is exceeded, considering that Δt is 300 microseconds corresponding to a 3 kc speech bandwidth. The results are tabulated in Table 1.

Words per minute (wpm) is given by the relation:

$$\text{wpm} = \frac{10}{\tau} \qquad (8)$$

Comparing the results from Table I to the empirical data obtained from the feasibility models, it is apparent that Table 2.3.2 data are optimistic by about a factor of 4.

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TABLE 2.3.2
Transmission Rate

Threshold Setting	P_t	$\overline{t}$	T	wpm	wpm re 12 db
9	.041	7.3 ms	47.3	211	+33.6%
10	.028	10.7	50.7	197	+24.6%
11	.019	15.8	55.8	179	+13.4%
12	.013	23.1	63.1	158	0
13	.0078	38.4	78.4	128	-19.0%
14	.0045	66.5	106.5	94	-40.5%
15	.0027	111.	151	66	-58.2%

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d.. Tone Signal Pulse Shaping

Any or all of the five tone generators in the transmitting unit may be operating at any one time. Each time a tone generator is triggered, it emits a signal burst of 40 milliseconds duration. It is of great importance that the energy spectrum of the pulse be controlled and confined to as narrow a region as possible. If energy is emitted at significant levels at frequencies outside of the desired region, such signals could cause errors in the receiving equipment. This problem is especially severe in view of the close proximity of the tone frequencies to each other.

In the case under consideration, the tone frequencies are located at 100 cps intervals between 1400 cps and 1800 cps. Clearly, a minimum of energy should be generated in the spectrum more than 50 cps removed from a given tone frequency; indeed, it is essential that the cross-channel interference level (defined as the unwanted energies in any one tone channel due to tones in all others) be at least 20 db below the wanted tone level. Two approaches exist to achieve the desired result:

1. Narrow band filters can be used to limit the energy spectrum to the required extent, or,

2. The energy spectrum can be limited by shaping of the pulse amplitude.

Approach No. 1 is undesirable for these reasons:

a. It is difficult to maintain center frequency and bandwidth accurately over the temperature range;

b. Filters are large, heavy, and expensive;

c. Five filters are required for the five tones.

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Approach No. 2 makes use of one wave shape generator and five modulators. It has been shown in theory and practice that this approach is practicable.

There exist two types of modulation envelopes for which the energies in the undesired frequency spectrum are particularly low; these are the "cosine-squared" and the "isosceles-triangle" envelopes. Approximate amplitudes of the frequency functions of these pulses can be gleaned from graphs in the I. T. & T. Reference Data for Engineers, page 1013 and page 1014. It is seen that the amplitudes of the two frequency functions are nearly alike. Thus, it is reasonable that a preference for one or the other of the two pulses can be based upon other considerations. A very good reason for selecting the "isosceles-triangle" pulse shape is the fact that such a pulse can be easily generated with good stability of amplitude, symmetry, and period. The total unwanted energy of an isosceles-triangle modulated pulse has been calculated to be 22 db below the proper tone level; and the peak amplitude of an interfering signal in the adjacent channel is calculated to be 28.6 db below the proper level. Measurements taken to date have consistently indicated a maximum interfering signal level in the adjacent channel (due to one oscillator) of 30 db below the desired signal. This result is in good agreement with the theory. It has not been resolved whether the remaining difference of 1.4 db can be ascribed to calibration errors, or whether the slight amount of rounding at the beginning of the triangular ramp is responsible for the improved performance. The interfering signal level in one channel due to tone signals in all others should reduce the margin between the interfering signal and the proper signal to approximately 25 db. This latter figure has been checked experimentally and is in agreement with the theory.

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2.3.3 Receiver

a. Coupling

The receiver is coupled directly to the telephone instrument rather than between telephone and line, and consequently, is somewhat simpler than the transmitter coupling unit. Figure 2.3.3 illustrates the receiver coupling technique. A notch filter identical to the transmitter filter is inserted between the handset microphone and the telephone hybrid. The rejection of the telephone hybrid is rather poor (around 15 db) and cannot be relied upon to isolate the local speaker's voice from the receiver. A better hybrid is furnished with the receiver, but variations in line impedance preclude any possibility of maintaining a high degree of balance. The notch filter assures that the signal leaking past the hybrid does not cause false alarm errors in the received message.

b. Audio Section

Distortion is the most important factor in the design of the audio unit. Distortion is held below 40 db using heavy negative feedback in all amplifiers preceeding the bandpass filters. With the tones 30 db below the transmitter keying threshold, it can be shown that speech distortion becomes an important factor if the dynamic range of input levels is too great (more than $\pm$ 3.8 db). Since the rate of input levels is 20 db or more, it appears desirable to incorporate a coarse manual gain control in the receiver which the operator adjusts with the aid of a meter at the beginning of each call. Figure 2.3.3b is a block diagram of the receiver. The noise plus tone signal from the hybrid is fed to a manual gain control. The control is set on voice peaks with the aid of a VU meter. An emitter follower isolates the bandpass filter from the hybrid (and the variable line impedance). This filter removes the voice from the voice-tone combination and sends the multitone signal to a 60 db amplifier. Separate emitter followers on the output of this amplifier drive the five

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tone channel filters. The filters each have a 3 db bandwidth of 100 cps and center frequencies at 1400, 1500, 1600, 1700, and 1800 cps. The individual tones are delivered to the logic circuit which contains an automatic threshold.

c. Threshold

A threshold must be provided to determine whether a mark or space was sent at the transmitter. The original approach was to provide a closed loop AGC on the tones after speech components are removed by filtering. A loop was designed and built but variable phase shift across the 100 cps bandwidth of the filters proved too severe to allow a reasonable closed loop device.

The final circuit adopted uses a simple open loop servo which rectifies the output of each channel filter and applies a common d-c threshold to each channel detector equal to some percentage of the strongest channel output, e.g. 6 db below the peak of highest level channel. Since this d-c level is always less than the rectified input, highly stable emitter followers may be utilized to provide reliable and precise threshold action.

2.3.4 Packaging

2.3.4.1 Scope

This section outlines the detailed design plan for packaging of the transmitter and receiver units.

2.3.4.2 General Requirements

These units are to withstand a temperature range of $0^{\circ}C$ to $+50^{\circ}C$. The construction is to be of good commercial practice.

2.3.4.3 Transmitter Unit

The following specifications were considered in the packaging of

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this unit. Size: 9" wide by 15" long and 10" high; these are the maximum outside dimensions. Weight: approximately 40 pounds. Finish: Semi-gloss black anamel. A ten-foot external cable is required. Duty Cycle: 1 hour on and 5 hours off.

Transmitter sub-units are as follows: Tally reader, Model 424C modified to ACI drawing 100-1-03-080, card basket and EECO "T" module chassis. ACI 100-1-03-078, filter chassis, ACI 100-1-03-079, and power supply chassis, ACI specification No. 100-1-10-006.

ACI drawing 100-1-03-008 shows the over-all transmitter package. The external dimensions of the box are 9" wide, 15" long by 10" high; weight is 6 pounds.

The case for the transmitter is fabricated from 2 Zero Manufacturing Company .063" thick 9" x 9" x 8" deep, drawn aluminum boxes. One side of each box is removed and the two remaining halves are butt-joined, welded and are reinforced by an open frame member at the joint to form the case. A 1/8" thick aluminum panel is welded to the open end of the case to form a rigid enclosed box. The panel has cutouts into which mount the Tally reader and the card basket chassis. The power supply is mounted to the case below the Tally reader. Two removable plates are provided on the side of the case for access to the connectors located on the filter chassis and for access to the power supply terminals.

The top of the transmitter is enclosed by a cover. The cover is held down to the case panel by screws. A hinged cover over the Tally reader permits access to the reader mechanism for replacing tapes and also provides a storage area for the extension cable. The cover is fabricated from two Zero aluminum boxes .063" thick, 9" x 9" x 2" with one side of each removed. Two handles, one at each end of the case are provided for carrying the unit.

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The blower in the Tally reader is used to exhaust the cooling air from this unit. The cooling air enters through vents at the plug-in chassis side of the transmitter, flows over the EECO "T" modules, cards, filter chassis, power supplies, through the Tally reader and out the opposite end of the case. Excluding the Tally reader, approximately 10 watts of power is dissipated.

The use of four deep-drawn aluminum boxes of the same 9" x 9" size with edge and corner radii gives the transmitter case a clean external contour as well as structural strength. This method was chosen for this design over more costly techniques such as all-welded or one piece deep-drawn construction.

2.4.4 Receiver Units

The following specifications were considered in the packaging of this unit:

Mounting to be compatible with standard 19" relay rack

Duty Cycle: one hour on, five hours off

The two major sub-units of the receiver, the Tally perforator and the receiver control chassis, are packaged separately in standard relay, desk-size steel cabinets approximately 14" wide x 15" deep x 23" high (Bud No. CR1727/Wyco C2119) painted black. The cabinet weight is approximately 40 lbs.

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3.0 THEORY OF OPERATION

3.1 General Design Requirements

No environmental requirements have been specified by the customer. However, to assure reliability, a tentative temperature range of $0^{\circ}C$ to $+50^{\circ}C$ on the equipment is considered reasonable. Allowing for temperature rise inside the package, circuits are designed and tested for a range of $0^{\circ}C$ to $65^{\circ}C$.

Transistors and semiconductor diodes are used in all circuit designs to improve reliability and reduce size. All semiconductors are derated according to their individual specifications for 65° operation. Resistor dissipation is derated by a factor of 5.

The equipment will be connected to a standard telephone system using 600-ohm impedances and levels of approximately -17 dbm.

3.2 Transmitter

3.2.1 General

The theory of operation of the transmitter is explained by pursuing the flow of a signal through the system and observing the operation of the individual circuits.

Input and output circuits, as well as the operation of the power supply, are discussed first.

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3.2.2 Detailed Theory

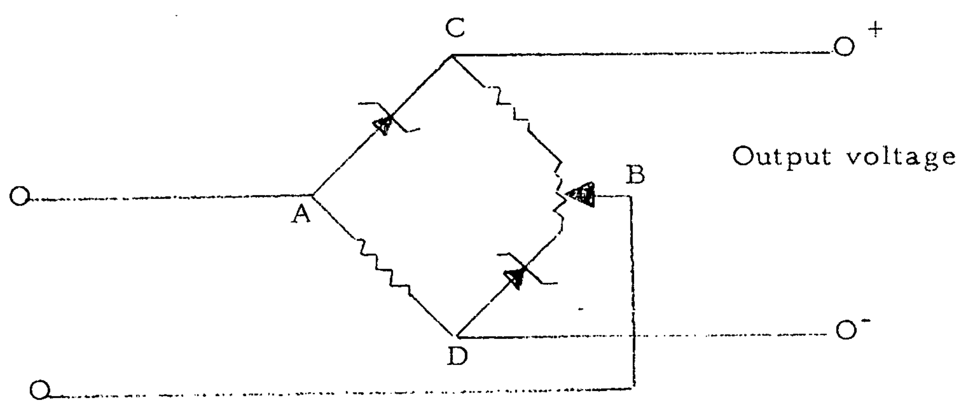
3.2.2.1 Power Supply

The transmitter power supply is an integral unit which operates off 115 V of 230 V line voltage.

Theory of Operation of Power Supply - The EM1140B Power Supply converts 115 volt, 60 cycle power to plus and minus 12 volts well regulated DC power and -24 volts unregulated DC power. An electrical schematic of the circuit is shown in the attached drawing. One power transformer is used for all outputs to minimize weight and size.

Since the two regulated sections are identical, only one will be considered for this discussion. Input power is fed to the transformer where the line voltage is stepped down. The reduced voltage is rectified in a full wave bridge circuit and filtered by a choke input filter.

The regulator section is made up of a comparison circuit, error signal amplifier, bias supply and series regulating transistor. The comparison circuit is a bridge consisting of two resistors, two zener diodes and a potentiometer as shown below.



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The output voltage is applied across the arms CAD and CBD to generate an error signal. The potential of point A with respect to D is the output voltage minus the fixed zener voltage, and is a potential that varies considerably for changes in output voltage. The potential of point B with respect to D is approximately equal to the fixed zener voltage and is fairly constant. Therefore, the error signal which is the difference of potential between points A and B will vary directly with output voltage. If the output voltage increases, the error signal will increase and vice versa.

Reference is now made to the power supply schematic, Figure 3.2.2. The error signal is fed to a transistor amplifier which drives the series regulating transistor. In order to operate properly, the base of the series regulating transistor must be biased negatively with respect to the emitter. The bias is obtained from the -12 volt supply for the +12 volt regulator and from a separate bias supply for the -12 volt regulator. When the output voltage increases, the error signal changes so that the base current of the series regulating transistor decreases. When the base current decreases, the collector current decreases and the voltage across the transistor increases. This decreases the output voltage and compensates for the original rise in output voltage.

The third output is unregulated and consists of a bridge rectifier and filter section to provide a low ripple DC voltage. The primary of the transformer is made up of two sections, which are wired in parallel for 115 volt operation and in series for 230 volt operation. Overload protection is provided by a 0.5 ampere circuit breaker when the power supply is wired for 115 V AC operation; when it is wired for 230 V AC operation, a fuze inside the power supply provides the protection.

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3.2.3 Input and Output Circuits

In operation, the transmitter is inserted in the telephone line between the transmitter telephone and the line.

3.2.3.1 Input Circuits

The input signal to the transmitter is obtained from the mouthpiece of the transmitter telephone whenever (a) operating power is applied to the transmitter, and (b) the transmitter telephone is "off the hook."

The operation of this circuit is as follows (Refer to Transmitter Schematic Diagram 100-1-03-064):

Assume that operating power has been applied to the transmitter. Lifting the phone off the hook completes the biasing circuit which enables the microphone in the mouthpiece of the transmitter telephone. The biasing circuit consists of resistors R_1 and R_4; capacitors C_4, C_9 and C_{10}; inductor L_1; relay coil K_1; and the carbon microphone of the telephone.

With the exception of the carbon microphone (which is outside of the transmitter) and the inductor L_1 (mounted on the filter chassis), all parts of the biasing network are mounted on TB_1.

Resistors R_1 and R_4 determine the microphone bias current of approximately 50 ma. Capacitors C_9 and C_{10} in conjunction with the resistors R_1 and R_4 form a filter network for the bias current. Inductor L_1 represents the load impedance for the transmitter telephone. The signal return path is provided by C_4. Bridged across Load L_1 is the first amplifier stage of the transmitter (Q_2, Card No. 10).

3.2.3.2 Output Circuits

After the input signal has been processed by the transmitter, it appears as the output signal of transformer T_1, and is impressed through capacitors C_2 and C_3 across L_2 and R_2. When the phone is off

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the hook (and K_1 thus is closed) L_2 and R_2 also serve to terminate the incoming telephone line. It is seen that under this condition any signal present across L_2 and R_2 is transmitted.

Transformer T_1 and inductor L_2 are mounted on the filter chassis. R_2, C_2, and C_3 are located on TB_1.

3.2.3.3 Audio and AGC Circuits

(a) First Audio Amplifier (Part of Card No. 10) - The input signal from the transmitter telephone appears at connector P101 pin 20 after having passed through the input circuit (see 3.2.3.1). From here, it is routed to P10 - 17 on audio card No. 10.

P10 - 17 is the input terminal of the first audio amplifier. This circuit consists of transistor Q_2 connected in common emitter configuration. The gain of this stage is 10 db V, and the maximum peak-to-peak input signal occurring at the base of Q_2 is 4.0 volts. The output signal of the amplifier is capacitively coupled via P10 - 20 to P9 - 22 on the AGC card (card No. 9).

(b) AGC Loop (Card No. 9 and Part of Card 8) - The AGC card contains most of the circuits that form the AGC loop. Two amplifiers, which are also a part of the AGC loop, are located on the audio card (Card No. 8).

The AGC loop is designed to maintain the average voice power per syllable at a relatively constant level, regardless of fluctuation of the incoming signal level. This effect is achieved by providing a "fast-attack-slow release" gain control characteristic in the loop. The loop gain is sufficient to maintain the output signal level within $\pm$ 1 db for input signal level fluctuations of $\pm$ 10 db.

Gain Control is achieved by means of a variable voltage divider

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consisting of R_1 and R_{20}. R_{20} is a "Varistor," i.e., a current-sensitive resistor. Its characteristic is such that an increase in the varistor bias current causes the varistor AC resistance to decrease. It is apparent that, if the varistor bias current is varied as a function of the voice power, gain control can be exercised over the incoming signal.

To explain the operation of the AGC loop, assume initially that no signal is present at P9-22. In this condition Q_1 is reverse-biased and only a small current flows through the resistor network of R_2, R_{20} and R_3. This current causes the varistor R_{20} to assume a better defined value of resistance than it can without a bias current. Capacitor C_2 acts as a filter to isolate the varistor from the line, and simultaneously, provides the signal ground.

Assume now that a signal is introduced at P9 - 22, small enough not to exceed the AGC threshold. This signal appears across the varistor R_{20} which is typically 1 K ohm under the given minimum bias condition. Transformer T_1 is bridged across R_{20}. It serves two purposes: (a) it provides a 2:1 step-up of the signal voltage, and (b) it allows the introduction of the DC bias current to Q_2 without loading effects or I_{co} problems.

Q_2 operates as an emitter follower amplifier providing a high termination impedance for transformer T_1, and a low driving impedance for the following filter. FL_2 is a high-pass filter having a cutoff frequency of 300 cps; its input and output impedances are 600 ohms; and its function is to prevent low frequency oscillations within the AGC loop, as well as to reject undesired low frequency signals which are not part of speech.

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Q_3 is an amplifier which is biased to achieve a gain of 26 db with a large dynamic operating range. The output of Q_3 is transformer coupled to the clipper circuit and to emitter follower Q_5. (Both circuits are on Card No. 8). The secondary winding of T_2 is returned to ground in order to obtain a signal swing which is symmetrical with respect to ground. This characteristic is necessary for the operation of the clipper circuit.

The signal in the AGC loop next flows into amplifiers Q_5 and Q_6. Q_5 is an emitter follower which performs two functions; (a) it provides the high impedance necessary to prevent loading of T_2, and (b) it provides isolation between the clippers and amplifier Q_6. This isolation is required to prevent signal distortion at T_2 prior to AGC attack when very large signal voltages (up to 20 volts peak-to-peak) can occur at T_2. Q_6 provides the final voltage amplification of the AGC signal prior to its application to the detector circuits.

Transformer T_1 (on card No. 8) couples the AGC signal to the detector driver Q_4 (on the AGC card). This transistor is biased so that its current output capability is limited only by its own DC current amplification factor and the drive available at the base. Since the voltage applied to the base of Q_4 can become much larger than $\pm$ 12 V with respect to ground, diode CR_2 is installed to limit the base voltage to +12 volts. The negative swing is limited to -12 volts by Q_4, since its collector-base junction becomes forward biased at larger voltages. The output of Q_4 is direct-coupled to detector diode CR_4.

The entire detector circuit, including its output current amplifiers Q_5 and Q_6, is DC coupled. This approach eliminates the time-constant problem inherent in AC coupled detectors. To achieve good AGC

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threshold stability over the temperature range from $0^{\circ}C$ to $65^{\circ}C$, compensating techniques were applied to the circuits. For instance, to insure that the AGC attack and release time constants are stable with temperature, transistor Q_4 obtains its bias voltage from Diode CR_3. Since diode CR_3 and the emitter-base diode of Q_4 are similar, their temperature characteristics are similar, and the DC bias voltage applied to CR_4 is very close to zero. This action insures that the quiescent DC voltage on the memory capacitor C_{11} is always near zero volts.

The DC signal present on C_{11} is applied to two complementary emitter followers in tandem. The use of complementary PNP-NPN followers reduces the DC shift through the amplifier pair to a very small amount. Transistor Q_5 is biased to present a very high impedance to the time constant network C_{11}, R_{16}, thus making the time constant essentially independent of the load.

At this point it should be noted that values of R_{15}, R_{16}, and C_{11} have been determined through listening tests. R_{15} and C_{11} control the AGC attack time and the network C_{11}, R_{16} controls the release time. Either time constant can readily be changed by varying R_{15} or R_{16}.

The output voltage of Q_6 is approximately zero volts in the quiescent state. In order to establish a definite AGC threshold, a 5.0 volt zener diode is inserted in the output of Q_6. The diode is held in the break-down condition by a current through R_{19}. Thus, in the quiescent condition a voltage of +5.0 V is applied via low-pass filter FL 1 to the base of the varistor control transistor Q_1, holding it off. The selection of a 5-volt zener diode as the AGC threshold is prompted by two considerations: (a) a 5-V zener diode has an

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extremely low temperature coefficient, and (b) operation with high level signal makes any changes with temperature in the junction voltage of Q_1 insignificant as far as threshold stability is concerned.

The low-pass filter FL 1 has a 200 cps cut-off frequency. Its purpose is to insure stable operation of the AGC loop.

The AGC loop is designed to reduce fluctuations of $\pm$ 10 db in the average voice level to less than $\pm$ 1.0 db. The average RMS Voice level at P 9-22 is 0.5 volts. The AGC threshold is exceeded with 0.1 volts average RMS.

As mentioned before, the attack and release time constants of the AGC system were determined in listening tests. The results indicated that an attack time constant in the order of 200 milliseconds is desirable. Because of this relatively slow attack characteristic, it is clear that initial bursts of voice energy far in excess of the desirable levels are applied to Card No. 8. In order to avoid overloading of the succeeding stages prior to AGC attack, amplitude limiters are provided on Card No. 8. The release time constant of the AGC is several seconds.

(c) Audio Card (Card No. 8) - The input signal is applied to Card No. 8 at P8 - 1. R_1, a high-resistance in the order of 20K ohms, feeds the signal to the clipper diodes CR_1 and CR_2. The clipping level is adjustable by means of potentiometers R_2 and R_4, with C_1 and C_2 providing the signal paths to ground.

The value of R_1 is selected in test; this is a convenient means of adjusting the gain of the system between the signal input and the clipper diodes to the desired value.

The output of the amplitude limiter circuit is applied to an emitter follower which in turn supplies the signal to the 4000 cps low-pass filter FL 1. This filter operates between 600 ohm impedances and

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limits the signal bandwidth to 4000 cps.

Having been passed through the filter, the signal is amplified by transistor Q_2. This stage has a gain of 20 db and can handle a large signal swing.

Emitter follower Q_3 is used to supply the impedance match and power required to drive the band reject filter FL 3.

FL 3 operates between 600 ohm impedances; its reject band lies between 1350 and 1850 cps; and its insertion loss is in the order of 7 db.

FL 3 is capacitively coupled to the base of emitter follower Q_7. This stage provides the driving power for output amplifier Q_4 as well as for low pass filter FL 4. The tone bursts are combined with the transmitted voice power at the base of Q_4. Bias network CR 4, R_{26}, and C_{14} establishes the operating of Q_4 in the most linear region. Linear operation of Q_4 is necessary in order to avoid the generation of frequencies between 1350 and 1850 cps which have just been removed from the signal spectrum by the preceding filter FL 3. Signals in the reject band, generated by non-linearities of Q_4, are approximately 70 db below the operating signal level. Aside from providing gain, Q_4 establishes a constant, resistive system output impedance of 600 ohms.

The output signal from Q_4 is applied to hybrid transformer T_1 on the filter chassis. Most of the signal appears on the output winding of T_1 and is impressed across L_2, R_2 of the output circuit as described in 6.2.2. A small amount of the output signal appears across the third winding of the hybrid transformer and is available at J 10 - 21. Also present at J 10 - 21 are signals which may be

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arriving from the telephone line. In order to monitor these incoming signals, an amplifier is provided on Card No. 10 which drives an earphone.

It is obviously desirable to balance hybrid transformer T_1 sufficiently well so that the level of the signal transmitted from Q_4 and appearing at J 10 - 21 is small compared with the signals received from the telephone line. Unfortunately, it is not generally possible to achieve a good degree of balance with the hybrid transformer. This is due to the fact that the hybrid transformer balance is a function of the impedance presented by the telephone line to the transformer. Since nearly every telephone line presents a different impedance, no single balance network will yield satisfactory results. In the given circuit, a balance of approximately 12 db is obtained by means of balancing potentiometer R_1. The amplifiers on Card No. 10 are designed to prevent "blasting" the transmitter operator's ear with his own voice. (d) Audio Card No. 10 - Card No. 10 contains the audio amplifiers which are required to enable the transmitter operator to listen to the incoming call.

The incoming telephone signal (along with the leakage of the outgoing signal) appears at P 10 - 21 of Card No. 10. Audio amplifier Q_3 is isolated from the hybrid transformer by emitter follower Q_1. This insures that no distortion will be introduced into the line regardless of the drive level applied to the amplifier.

A set of amplitude limiter diodes is connected across the output of the emitter follower. This technique insures that the signal level applied to Q_3 will never exceed 300 mv peak-to-peak

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and, consequently, Q_3 cannot be blocked by excessive incoming signal levels. The gain of Q_3, when the amplifier is loaded with a 1000-ohm impedance earpiece, is approximately 15 dbv. This gain is sufficient to raise the incoming signal to a comfortable listening level.

It was indicated previously that under most operating conditions, a significant level of the transmitter operator's voice can be present at P10-21. To limit the signal level applied to the earpiece under this condition, limiter diodes are connected across the earpiece. The distortion introduced by the amplitude limiter diodes is not objectionable as was borne out by listening tests.

3.2.3.4 Logic and Tone Generator Circuits

(a) Detector Card (Card No. 7) - The detector card contains circuits which monitor the instantaneous signal level of the transmitted voice present at J8-22. After amplification and fullwave rectification, the signal is applied to an adjustable threshold circuit which generates an output pulse every time the signal exceeds the threshold. In flowing from P8-22 to the detector card, the signal passes through filter F1-4 mounted on TB1 which is a part of the filter chassis. F14 is a low-pass filter with a cutoff frequency of 750 cps. Its purpose is twofold: First, it prevents noise spikes from triggering a tone burst when there is no voice energy present to mask the tone; and secondly, it increases the probability of masking a tone by triggering only on low-frequency

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components of the voice which exceed the threshold. This latter fact is based on evidence that the rate of decay of a syllable composed of high frequencies is much faster than the rate of decay of a low-frequency syllable. The output of the 750 cps low-pass filter is applied to P7-1. Amplifier Q_6 accepts this signal and provides a voltage gain of approximately 28 db. Emitter follower Q_7 is direct-coupled to the collector of Q_6, and provides the power gain required to drive step-up transformer T_1. The center tapped secondary winding of T_1 feeds two diodes connected as a fullwave rectifier. In order to achieve good rectification efficiency, the diodes are held in the conducting state during quiescent signal conditions by a small current caused by a voltage developed across R_1 and CR_3. This current of approximately 20 micro-amperes per diode is relatively independent of temperature. This effect is achieved by deriving the detector biasing current from a voltage source (CR_3 and R_1) which exhibits a temperature characteristic similar to that of detector diodes CR_1 and CR_2.

Potentiometer R_3 is required to permit balancing of the two detector legs which otherwise would differ in their outputs because of inherent differences in diode and winding impedances.

The detected output is developed across R_4.

It is in order here to consider the reason why a fullwave rectifier was inserted instead of applying the AC signal directly to the threshold.

Fullwave rectification of the audio signal provides two significant improvements over the use of unprocessed audio: (a) it causes a higher

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tone transmission rate, and (b) it increases the probability of masking the tone by voice energy.

To explain the mechanism of achieving higher tone transmission rates by using fullwave rectification, note that through fullwave rectification, use is made of the positive and the negative signal excursions; and that twice as many signals are sampled by the threshold.

The use of positive and negative signal excursions speeds up the transmission rate by providing the possibility of exceedin g the threshold several milliseconds sooner than in case of no rectification. Thus, there is an increased probability that two, three, or even more tone bursts can be transmitted per syllable, simply because of the fact that threshold conditions are detected sooner, tone transmission is initiated sooner, and less dead time occurs between tone bursts.

The fact that threshold conditions are detected without unnecessary delay permits the initiation of the tone burst at the earliest possible moment, thus increasing the probability of masking the tone burst even in the case of very short syllables.

Returning to the description of the operation of the detector circuits, it is remembered that the signal had been traced to R_4. Emitter follower Q_1 employs a type 2N1420 transistor which features a very high current amplification factor. The intention is to prevent any undue loading of the detector circuit by developing a very high input impedance to the base of Q_1.

E mitter follower Q_1 does not operate as a linear amplifier over

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the full dynamic range of the input signal. To understand its operation, it is necessary to first explore the operating characteristic of the threshold stage Q_2.

The emitter voltage applied to Q_2 is derived from voltage divider R_{12} R_{13} and is adjustable between +12 and +6 volts. Under quiescent conditions and until the signal reaches threshold level, Q_2 is held in saturation by a voltage negative with respect to the emitter. This negative voltage is supplied through R_9.

It becomes obvious at this point that follower Q_1 need not operate at all until the base voltage at Q_1 approaches +6 volts. During quiescent conditions when the base voltage of Q_1 is near zero volts, R_9 will continue to supply base current to Q_2 as if Q_1 were non-existent. Indeed, under such conditions, Q_1 is reverse-biased by the following mechanism. Assume that R_{12} is adjusted to +8 volts. Q_2 is now saturated, and the base current of Q_2 is approximately 1.7 ma. Because of voltage divider action, the emitter of Q_1 will be at approximately +7 volts, indicating that Q_1 is reverse-biased. Certainly, Q_1 could be designed to operate as a linear amplifier by decreasing R_9 considerably in value. However, this approach generates two problems: (a) it tends to reduce the input impedance to Q_1, thus loading the detector, and (b) it causes excessive currents to flow through the base of Q_2, thus abusing Q_2 and also causing unnecessary loading of the divider R_{12} - R_{13}. It is seen, therefore, that non-linear operation of Q_1 is desirable.

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Suppose, now, that the input signal at the base of Q_1 exceeds +8 volts. Now Q_8 goes into the conducting state and its emitter will follow the base. It is apparent that any further positive increase in signal voltage at the base of Q_1 will cause Q_2 to be turned off, and a negative-going pulse to be generated at the collector of Q_2. Its duration is equal to the length of time that the threshold is exceeded.

A threshold stability of better than $\pm$ 1 db is achieved through the use of complementary transistors, i.e., Q_1 is an NPN, and Q_2 is a PNP silicon transistor.

The normal operating level (long-time average of RMS voice voltage) at the base of Q_6 is approximately 50 mv rms. The wiper of R_{12} is typically set to +8 VDC.

(b) EECO Rack - The EECO Rack contains most of the digital logic elements of the transmitter unit, namely, a Schmitt Squaring Amplifier and three One-Shot Multivibrators.

The input to the EECO rack consists of a negative pulse which is derived from Q_2 on the detector card and which is available at P7-8. The pulse is capacitively coupled to the squaring amplifier Z_1. Its output is applied to a gate circuit consisting of R_1, R_2, and CR_1. If no pulse has been applied to this gate during the preceeding 40 milliseconds, the gate is enabled, and the output pulse from Z_1 is passed to one-shot Z_2. This circuit provides a 20-millisecond time delay. The normal (positive-going) output pulse of Z_2 is passed through an emitter follower (part of Z_3) and is fed to P6-1 to be used in the control of the ramp generator. The trailing edge of this pulse

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serves as the trigger for the second 20 millisecond one-shot Z_4.

The inverted (negative-going) output of Z_2 is applied to Pg-21 to drive emitter follower Q_1 on Card No. 6.

At the moment when the timing interval of Z_2 ends, Z_4 is fired which also operates as a 20 millisecond delay circuit. Z_4 performs several functions. Immediately upon firing it delivers a negative pulse to trigger the 4 m sec one-shot Z_5. Simultaneously, the normal (positive-going) pulse is applied., via an emitter follower (part of Z_3) to the gate which controls Z_2. This positive pulse inhibits the gate and prevents retriggering of Z_2 until Z_4 has timed out. The very same pulse also clamps (through CR_2) to zero the trailing edge of the negative pulse of Z_2. This technique yields a fast fall time for Z_2.

The inverted (negative-going) output of Z_4 is applied to P6-9 to be used for control of the tone generators. At the end of the 20 millisecond interval of Z_4, the input to the gate is removed and Z_2 can be retriggered.

(c) Ramp Generator - The ramp generator comprises transistors Q_2 and Q_3. The control signal for the generator appears at P6-1. It consists of a positive pulse (-11 V to -3 V) of 20 millisecond duration. In order to achieve good linearity of the trangular waveform, the circuit operates between -12 V and +12 V.

The ramp generator operates as follows:

The -11 to -3 volt signal at P6-1 is shifted by a 15 volt zener diode (CR_1) to obtain a +4 to +12 volt swing at the base of Q_2. In the quiescent condition, the base of Q_2 is at +4 volts. In this condition, Q_2 is saturated, its collector is at +4 volts, and Q_3 is cut off since its emitter is clamped to ground by diode CR_2.

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During this period, there is a constant +200 mv on the emitter capacitor of Q_3.

Assume, now, that a pulse arrives which elevates the base of Q_2 to +11 volts. Q_2 now turns off, since its emitter is returned to +10 volts. With Q_2 off, the base of Q_3 is returned to -12 volts which causes Q_3 to go into saturation. The emitter of Q_3, which initially was clamped to ground, now begins to shift negatively, trying to follow the base toward -12 volts. The time constant is dimensioned so that the emitter capacitor of Q_3 changes to - 1.2 volts in 20 milliseconds. Good linearity of the ramp voltage is achieved by limiting the emitter swing to -1.2 volts.

After 20 milliseconds of charging, the base of Q_2 returns to +4 volts. At this moment Q_2 again goes into saturation and Q_3 is cut off. The emitter capacitor of Q_3 now charges from -1.3 volts toward +12 volts through the emitter resistor of Q_3. This charge time is so dimensioned that after 20 milliseconds, the voltage on the capacitor is +200 millivolts. At this moment, diode CR_2 conducts, and the ramp is completed.

The output of the ramp generator is amplified by emitter follower Q_5 and applied to all five of the tone frequency modulators. Note that the ramp signal will be fed to the tone modulators every time that the logic circuits are triggered. This is true regardless of the tape reader input. The ramp signal appears at P6-8.

(d) Tone Oscillator and Modulator (Card No. 1)

There are five tone oscillators and five modulators in the transmitter. The modulators are identical, and the tone oscillators

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differ only in the inductance values used in the frequency determining networks. Consequently, only one oscillator and one modulator will be discussed.

The tone oscillator is a conventional two-transistor type with the frequency determining elements connected between the emitters of the transistors. This design makes the oscillator frequency essentially independent of the load, provides a relatively low output impedance, a large voltage swing, and good starting characteristics.

The oscillator is controlled by a transistor switch (Q_3) which normally clamps the collector of Q_4 to ground, inhibiting operation. To start the oscillator, Q_3 is reverse-biased. This allows the collector of Q_4 to swing positive and oscillation begins. The tone oscillator signal is coupled to the modulator and impressed across diodes CR_2 and CR_3. The tone modulator is a very simple circuit. It operates as follows:

Assume that initially no input is present at the audio channel. Thus, the ramp generator is in its quiescent state. Remembering that the quiescent voltage at J6-8 is slightly positive, it is seen that the emitter of Q_1 tends to be above ground by about 400 mv due to the junction drop of emitter follower Q_1. This voltage insures that diodes CR_2 and CR_3 of the modulator are conducting, and that no output can exist from the modulator.

Now assume that at t=0, a signal is received which calls for the generation of a ramp, and that, simultaneously, the input from the tape reader requires the generation of a certain tone.

The following events now take place:

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(1) The proper tone oscillator is turned on and its output is fed into the modulator; (2) the emitter of Q_1 follows the negative-going ramp, and an increasing reverse bias is established across diode CR_2; the oscillator signal is therefore always clipped in its positive swing, but it will build up to the equivalent of the instantaneous reverse bias of diode CR_2 in its negative swing. Signals above this threshold are short-circuited by the large emitter capacitor of Q_1. Since the amplitude of the tone is larger than that of the ramp, a 100% modulated wave results.

It is desired to multiplex the outputs of the five modulators. This cannot be accomplished passively for reasons of isolation and impedance matching. Amplifier Q_5 is therefore inserted after the modulator. This stage is AC coupled and has unity voltage gain. In order not to load the modulator, a high input impedance must be achieved for the amplifier. The ensuing requirement for a high impedance bias circuitry calls for the use of a silicon transistor as the amplifier. There is one additional consideration which requires attention; namely, the fact that the modulator output consists of a negative-going pulse only. Thus, the average output of the modulator is a DC voltage, the magnitude of which is a function of the repetition rate of the pulse. If the amplifier were to follow the average output to any degree, the linearity of the amplification would be lost, and the output wave shape would no longer be triangular. It is necessary, therefore, that the time constant of the coupling network be of the order of one cycle of the tone oscillator. The implication is that now the waveform of the carrier tone suffers some distortion. However, since the unwanted frequencies are far removed from the desired spectrum, it is easy

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to reject them. Low-pass filter C_7, C_8, L_1 on Card No. 6, having a cutoff frequency of 2000 cps, serves this purpose.

(e) Tone Control Circuits -

The tone control circuits are located on Cards No. 1 and 6. In conjunction with the tape reader, these circuits initiate and terminate the tone bursts at the proper instants. Throughout the following paragraphs, reference will be made to the timing diagram of Figure 3. 2. 2. 2. 4.

To explain the operation of the tone control circuits, assume first that no audio signal is present at the transmitter input. Consequently, the system is in its quiescent condition, and the following DC voltages are present at the control circuits: Transistor Q_3 on Card No. 1 is in saturation, causing the tone generator to be "OFF". Q_3 is held in the "ON" condition by a voltage of +4 volts which is derived through a zener diode from emitter follower Q_6 on the Ramp Generator Card No. 6. If the tape reader senses a hole in the tape, the +4 volt potential is present also at diode CR_1. This diode is "ON" because of the return of R_2 to +5 volts. Clearly, with +4 volts applied to the base circuit of Q_3, the transistor will be in saturation.

Consider the second possibility, namely, that the tape reader does not sense a hole in the tape. In this case, no DC continuity exists between Q_6 on Card No. 6 and diode CR_1 on Card No. 1. However, Q_3 on Card No. 1 is "ON" whenever there is no DC continuity through the tape reader, i. e., whenever the reader does not sense a hole in the tape.

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Consider, now, the final possibility, namely, that an audio signal has exceeded the threshold, and that the first 20 millisecond one-shot Z_2 has fired. Call this time t_o.

At time t_o, the emitter of Q_6 on Card No. 6 swings from - 3 V to -11 V; and, due to zener diode CR_3, the voltage at P 6-22 swings from +4 V to -4V. Assuming that the tape reader senses a hole, DC continuity exists between P 6-22 and P 1-1. Since Q_6 on Card No. 6 represents a very low source impedance for negative-going signals, it is capable of discharging capacitor C_1 on Card No. 1 very quickly, thus reverse biasing Q_3 and permitting the tone generator to start. The above voltage conditions persist until Z_2 times out after 20 milliseconds. Let this time be called t_1. At t_1, the voltage at P 1-1 swings to +4 volts, reverse biasing CR_1.

Remembering that a tone burst lasts for 40 milliseconds, it is apparent that the tone generator must continue to run for 20 milliseconds past time t_1. This operation is effected by dimensioning the time constant $R_2 \times C_1 >$ 20 milliseconds. Thus, the tone generator tends to run for longer than t_1 + 20 milliseconds. The exact end of the 40 millisecond tone burst at time t_2 is effected through transistor Q_2. This circuit operates as follows: Let t_2 be the instant when the second 20 millisecond one-shot times out. At time t_2, the voltage at P6-9 swings from -11 V to -3 V and transistor Q_4 is turned "ON" through the resistive voltage shift network R_2 and R_{20}. Switching of Q_4 causes a sharp negative pulse to appear at the base of Q_2 on Card No. 1. The effect is that Q_2 conducts hard for a brief instant. Its current gain is sufficient to remove quickly the residual negative charge from C_1 (which held Q_2 "OFF"

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and the tone generator "ON"), and to replace it by a positive charge, which causes Q_3 to conduct and squelch the output of the tone generator. Thereafter, Q_3 is held "ON" through R_2, and Q_2 is held "OFF" through R_{14} on Card No. 6.

Since the tape reader participates in the tone control operation only during the first 20 milliseconds, (from t_o to t_1), the succeeding 20 millisecond interval (t_1 to t_2) can be used to advance the paper tape reader by one step. Thus, ample time is available for the advance of the mechanism, and for the settling of the tape reader contacts.

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3.3 Receiver

3.3.1 General

An over-all block diagram of the receiver is shown in Figure 3.2 and a detailed schematic diagram in Figure 100-1-04-034. Input signals consisting of notched speech plus information tones are applied to the receiver via a public telephone system. The message information detected by the receiver is contained within the 1300 to 1900 cps frequency band. Message signals are transmitted to the receiver at low amplitude, about 27 db below the rms level of the 300 to 3000 cps voice signal.

The received message consists of alpha-numerical characters in standard five-baud teletype code. The five bauds comprising each character are transmitted simultaneously in the form of 40-millisecond pul ses having a triangular shape. A separate carrier frequency is used for each of the five pulses. The five carrier frequencies are:

1400 cps
1500 cps
1600 cps
1700 cps
1800 cps

Bandpass filters are used in separate amplifier channels to separate the individual code pulses from the composite signal. The detected and amplified pulses are supplied to logic circuits which drive a Tally Paper Tape Perforator. The receiver is capable of processing input code groups at any rate up to 25

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characters per second, or approximately 250 words per minute. The receiver output is a punched paper tape suitable for use on any standard teletype printer.

The receiver utilizes a special sliding-threshold agc circuit which provides a constant output from each of the five detector channels over a dynamic range of 10 db. A manual gain control circuit and VU meter provide for adjustment of the input signal amplitude to compensate for the wide variations between local and long-distance transmissions.

An Engineered Magnetics Model EM 1140B regulated power supply provides operating voltages for the receiver.

3.3.2 Detailed Description

3.3.2.1 Input-Output Circuit

The input-output circuit of the receiver consists of the output amplifier and band-rejection filter which notches the voice spectrum of the receiving station operator to prevent errors in the received message; the input circuit for the receiver; and coupling circuits which permit the use of a single-pair transmission line for the two-way conversation.

The voice signal from the receiver telephone is applied to pin A of J3 in the receiver. This signal is generated in a high-impedance crystal microphone which develops a peak-to-peak voltage of about 200 millivolts across a load impedance of 75 K ohms presented by R9 and the input impedance of Q1, both of which are located on TB2. The conventional carbon microphone normally used in the telephone is replaced to minimize distortion and to reduce the possibility of false alarm errors in the received message.

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The need for low distortion in this application is obtained from the following considerations. The spectrum of normal speech is highly peaked in the region below 700 cps and decreases rapidly at frequencies above 1000 cps. When the speech is distorted (distortion in a carbon microphone is as much as 30%), however, the spectrum tends to flatten and additional energy appears above 1000 cps. Since the voice signal, due to the local speaker, is already 20 db stronger than that of the distant speaker, and since the notch filter (FL1) has finite attenuation in the message bandwidth, distortion in the local speaker's voice can cause a significant number of errors to occur whenever the receiving operator speaks. The use of a linear crystal microphone minimizes this effect.

Additional isolation of the local speaker's voice from the receiver input is provided by the hybrid transformer T2 which is described in detail in the later part of this section.

Returning to the input circuit, the voice signal from the microphone passes through emitter follower Q1 on TB2. This signal is applied to the base of Q2 where it is amplified to a level of about 850 millivolts peak-to-peak. After further amplification in Q1 (audio card No. 1), the speech signal passes through the band reject filter FL1 which attenuates speech energy in the 1300 to 1900 cps message bandwidth when the function switch is in the RECEIVE OR STANDBY position. In the normal position the un-notched signal at the base of Q1 by-passes the notch filter via S1-A1 and is supplied directly to Q2 base on audio card No. 1. Q2 is an emitter follower that isolates the band-reject filter from line impedance variations. The output of this stage is fed through a 330-ohm resistor (R9) to hybrid transformer T2. R9 plus the

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source impedance of emitter follower Q2 add to yield a resistance of about 350 ohms, which is just equal to the resistance of gain control R1 (front panel). Hybrid transformer T2 is exactly balanced when R2 (chassis, Figure 8.1) plus R6 (TB2) just equals the line resistance. When this occurs, the voltage drop across R6 - R2 is exactly equal to and 180° out-of-phase with the voltage across pins 7 and 9 of T2. Consequently, no current flows in R1 (gain control) due to a signal applied to pin 4 of T2. (Since the line impedance varies for different telephone circuits, the balance is not perfect and some feed-through occurs.) Instead, the signal is coupled through T2 to T1, where it is delivered to the telephone line for transmission to the transmitter station.

The incoming voice signal arriving from the transmitter station is applied to the telephone earpiece through terminal D of J3. In the "NORM" mode, this voice signal is applied directly from terminal 9 of the transformer T2 to J3-D, through contacts of S1-A2 and S1-A5. In the "STANDBY" and "RECEIVE" modes, emitter followers Q1 and Q2 on terminal board TB1 are placed in series with the voice signal output at T2-9. In these modes, the received signal is also applied to GAIN CONTROL R1 in the receiver input channel. The emitter followers provide isolation between the receiver input channel and the earpiece to minimize distortion and voice feed-through.[1]

Note 1: The earpiece in the receiver telephone acts like a low-efficiency microphone without isolation; unnotched speech can appear at the input to the information channel and can cause false alarm errors.

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In the "STANDBY" and "RECEIVE" modes, voice input signals received over the external telephone transmission lines at J3-E and -F are coupled through transformers T1 and T2 to the emitter followers and the receiver GAIN CONTROL R1.

3.3.2.2 Receiver Input Channel

The receiver input channel consists of a gain control, a signal-level monitoring circuit, and a preamplifier. GAIN CONTROL R1 provides a means of adjusting the input signal level to compensate for the variation between local and remote transmissions. For local signals, the maximum peak-to-peak amplitude of voice input signals may be as high as 3 volts; for input signals from remote transmitters, the input may be as much as 20 db below this level, or about 300 millivolts peak-to-peak.

To permit proper adjustment of GAIN CONTROL R1, a signal-level monitoring circuit is provided. The monitoring circuit consists of emitter follower Q5, amplifiers Q6 and Q4, a peak clipper circuit, and VU meter M1. The transistor stages are mounted on audio card No. 1, the clipper and meter circuits on terminal board TB1. The voice input signal at the arm of GAIN CONTROL R1 is applied to the monitoring circuit through emitter follower Q3, which also drives the preamplifier circuit.

Emitter follower Q5 at the input to the monitoring circuit drives a two-stage cascade amplifier, Q4 and Q6, the output of which is applied to METER ADJUST potentiometer R14. This potentiometer provides a means of calibrating the VU meter. The amplified output at the arm of R14 is coupled by C6 and C5 to VU meter M1.

Positive and negative peak clippers are connected across the

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output to protect the meter against high-amplitude peak signals. The clipper diodes, CR1 and CR2 on terminal board TB1, are back-biased by voltage divider resistors R4 through R7, so that only high amplitude peaks of the signal cause the diodes to conduct.

The dial of VU meter M1 is marked to indicate the optimum signal level, and GAIN CONTROL R1 is adjusted so that the meter reading coincides with this mark. Since the ratio between average level of the voice signal and the low-level message signal is nearly constant, this adjustment brings the input signal amplitude within the limits of the dynamic AGC range of the receiver.

The preamplifier circuit consists of bandpass filter FL2, and a three-stage amplifier and emitter follower located on audio card No. 2. The bandpass filter passes the low-level message pulses within the 1300 to 1900 cps band, and rejects all components of the voice signal.

The three-stage preamplifier, consisting of transistors Q1, Q2 and Q3 on audio card No. 2, provides approximately 60 db of gain for the 1300 to 1900 cps signal components. An over-all gain of approximately 100 db is required for the receiver; however, the preamplifier gain is limited to 60 db in order to maintain linearity and to prevent intermodulation distortion. Intermodulation distortion might occur if two tones were received simultaneously, or if an exceptionally strong signal outside the message band were not attenuated sufficiently by the bandpass filter. This could cause a false output from one or more of the five detector channels, and in turn cause an incorrect character to be punched by the Tally perforator.

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The input signals to the preamplifier consist of triangular pulses of 40-milliseconds duration, at five different carrier frequencies, spaced 100 cps apart, from 1400 to 1800 cps. Transformer coupling is used between the three amplifier stages and between the third stage and output emitter follower Q4. Thermistor RT1 in the emitter of Q3 provides compensation for gain variations with temperature changes. The preamplifier is disabled by the operating mode selector switch, except in the "RECEIVE" mode. When S1 is set to either "NORM" or "STANDBY", the emitter of amplifier Q4 is shorted to ground through contacts of S1-B5. In "RECEIVE" mode, the short is removed allowing Q4 to conduct.

The output of the preamplifier circuit is applied from the emitter of Q4, through J6-22 to the inputs of the five audio detector channels.

3.3.2.3 Audio Detector Channels

The five audio detector channels provide further amplification of the message pulses. All five detector cards are identical except for the bandpass filter at the input of each card which selects only one of the five modulated frequencies. Each bandpass filter has a 3-db bandwidth of 100cps. Since the five detector cards are identical, only detector card No. 1 will be described.

The input signal to the detector card consists of the amplified message signal from emitter follower Q4 in audio card No. 2. This signal is applied through J7-1 to emitter follower Q1, which provides an impedance match between the preamplifier output and bandpass filter FL3, and isolates the input circuits of the five detector cards.

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The bandpass filter FL3 has a bandwidth of 100 cps, centered at 1400 cps, so that only the 1400-cps message pulses are coupled to the first amplifier stage Q8. Amplifiers Q8, Q2, and Q3 in cascade provide 40 to 50 db of amplification of the 1400-cps signal. Potentiometer R3 at the input to transistor Q2 provides a gain adjustment for the amplifier circuit. The narrow bandwidth of the amplifier eliminates all intermodulation distortion, since no two signals within the 100-cps band can generate a cross-product that might be interpreted as a valid signal.

The amplified 1400-cps signal at the collector of Q3 is applied in parallel to emitter followers Q4 and Q7. Emitter follower Q4 drives the diode detector, and Q7 supplies one of the five input signals to the AGC circuit. (See paragraph 3. 3. 2. 5) Emitter follower Q4 and transformer T1 provide a low-impedance charging source for the RC network at the output of the detector. Detector diode CR2 is back-biased by the AGC circuit, through the secondary winding of T1. Capacitor C1 in AGC card No. 2 provides an AC ground return for the secondary of T1. The AGC circuit establishes both a minimum and a dynamic threshold bias level on the detector.

Capacitors C4 and C7, connected across the detector output, are charged very rapidly to the peak value of the input signal, through the low-impedance source of T1 and Q4. A delay is introduced, however, by the RC time constant of resistor R18 and series capacitors C5 and C8. This delay provides sufficient time for the AGC circuit to respond to the input signal, and also prevents any stray noise pulses of short duration from exceeding the trigger threshold. Thermistor RT1 forms a part of the voltage divider which establishes the bias level for diode detector CR2 and switching

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transistor Q5. The thermistor acts to maintain this bias at a constant level over a wide range of temperature variation.

Switching transistor Q5 is normally conducting. When an incoming signal pulse exceeds the AGC threshold, CR2 conducts and a positive-going signal is coupled to the base of Q5, cutting off the transistor. This produces a negative-going pulse at the collector of Q5 and at the output of emitter follower Q6. The negative-going output of Q6 is coupled via P103-22 to a Schmitt trigger at the input to the control logic circuits.

During any 40-millisecond interval, a negative-going output pulse from any of the five detector cards represents a binary 1, or true signal; conversely, the absence of a pulse represents a binary 0, or false signal. The combination of the five outputs represents the teletype code for a particular alphabetical or numerical character in the incoming message.

3.3.2.4 Control Logic Circuits

The control logic circuits respond to the outputs of the five detector channels to provide simultaneous 1-millisecond output pulses to the perforator. The logic circuits include a 10-millisecond delay which allows sufficient time for detection of signals in all five channels, and a 1-millisecond one-shot multivibrator which synchronizes the outputs of the five channels.

Assuming that a signal above threshold level is detected in channel 1, a negative-going output signal will be coupled from the emitter of Q6 in detector card No. 1, through P103-22 to the channel-1 Schmitt trigger Z1. Because of the delay in the detector circuit, the transition from the false to true state is relatively

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slow: the Schmitt trigger, however, provides an output pulse with a very fast rise time. This output pulse is coupled to the channel-1 20-millisecond one-shot multivibrator, Z6. Identical Schmitt trigger circuits, Z2 through Z5, and 20-millisecond one-shots, Z7 through Z10, are provided for the other four detector channels.

The output of the channel-1 20-millisecond one-shot, designated Z6, is combined in an AND gate with the Z14 term from a timing circuit. The Z6 term becomes true when the 20-millisecond one-shot fires; the Z14 term becomes true 10 milliseconds later, for a period of 1 millisecond. During the 1-millisecond period when both Z6 and Z14 are true, an output pulse is coupled to the perforator by emitter follower Z11-A. At the end of the 20-millisecond period of Z6, the circuit returns to the quiescent, or off, state, ready to respond to the next incoming signal. In the same manner, 1-millisecond output pulses are delivered simultaneously to the perforator from all other channels where a signal was detected.

The timing circuit, which synchronizes the five channels, consists of an OR gate, a 10-millisecond delay one-shot, a 1-millisecond one-shot, and associated emitter followers. The output signals from the 20-millisecond one-shots in the five signal channels are combined in the OR gate, comprising CR1 through CR5 in the logic card. If any one of the inputs to the OR gate is true, the 10-millisecond one-shot Z13 is triggered. This one-shot provides a 10-millisecond delay to allow time for all five detector channels to respond to an input signal. Normally, all five bets for each character in the message code are transmitted simultaneously; however, the transmission time for each of the five carrier frequencies may vary several milliseconds when the signals are transmitted over long distances.

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At the end of the 10-millisecond delay, the trailing edge of the one-shot output triggers the 1-millisecond one-shot Z14. The 1-millisecond output pulse from Z14 is applied to emitter followers Q1 and Q2 in cascade, on the logic card. Emitter follower Q2 provides the D1, D2, D3, and D4 and D5 terms for the output AND gate in each signal channel, as described above. The output of Q2 is also coupled by capacitor C1, through P110-8 and J1-40, to the Tally perforator. This signal is applied through a delay circuit to the sprocket drive in the perforator to advance the tape one step after the appropriate code has been punched.

3.3.2.5 AGC Circuit

The AGC circuit in the receiver provides a sliding-threshold voltage for each detector channel, to provide positive triggering of the control logic circuits over a 10-db dynamic range of input signal variations.

As described in paragraph 3.3.2.3, the amplified signal at the collector of Q3 in each detector card is applied to emitter follower Q7. The output of Q7 is coupled to the AGC circuit through transformer T2 and diode CR3, which comprises one element of an OR gate. The combined signals from all five detector circuits are applied through the OR gate to the base of emitter follower Q1 on AGC card No. 1. The level of AGC voltage developed by the circuit is determined by the input signal having the highest amplitude. Two emitter followers, Q1 and Q2 on AGC card No. 1, are connected in cascade to provide isolation of the input circuit and a low-impedance source for the AGC detector. The response of the AGC circuit is sufficiently rapid to follow the linear rise of the

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triangular modulation envelope of the input signal.

Diodes CR1, CR2 and CR3 provide temperature compensation for emitter followers Q1 and Q2, and AGC detector diode CR4. Diode CR4 rectifies the input signal, and the negative AGC output voltage is developed across a filter network comprising C3 and R6. Two additional emitter followers, Q3 and Q4, provide sufficient power to drive the five parallel output stages. Emitter followers Q5 through Q9 provide a separate AGC output circuit for each of the five signal channels. A potentiometer in the emitter circuit of each output stage permits separate adjustment of the AGC voltage supplied to each channel, to provide compensation for variations in gain between channels, and for differences in input signal levels at the different modulation frequencies. Diodes CR5 and CR6 in series between Q3 and Q4 provide temperature compensation for the output stages.

The five outputs from AGC card No. 1 are coupled to AGC card No. 2, where each is combined with a fixed bias circuit. The fixed bias circuit establishes a minimum threshold voltage for the associated detector channel. For detector channel 1, the minimum threshold is established by the setting of potentiometer R1, which forms part of a voltage divider connected across the -12 volt supply. The minimum threshold is adjusted to optimum level and applied through isolation diode CR2 to transformer T1 on detector card No. 1. When the AGC voltage from Q9 and potentiometer R19 in AGC card No. 1 is less than the fixed bias level, the AGC volt age has no effect. However, when the AGC voltage exceeds the fixed bias or minimum threshold, CR2 is biased off and the AGC voltage is applied to

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the channel 1 detector. Diode CR1 provides temperature compensation for CR2, and acts to maintain a constant level of fixed bias.

The bias circuits for channels 2 through 4 are identical to the channel 1 circuit.

The action of the AGC circuit is such that signals having a peak amplitude less than the minimum threshold established by the fixed bias circuit cannot produce a detector output signal. Signals having a peak amplitude greater than the minimum threshold produce a detector output signal which is held nearly constant over an input range of 10 db.

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4.0 PACKAGING

4.1 Design of Transmitter Sub-Units

4.1.1 Reader

The Tally reader purchased per ACI specification drawing No. 100-1-10-018 is 7-7/8" x 8-1/2" x 6-1/4" high; weight, approximately 13 lbs. This unit is flange-mounted and secured to the case panel by 6 No. 8-32 screws. The reader unit panel and cover are fabricated from rigidamp material supplied by Barry Controls of Glendale, California to quiet the operation of this unit. To complete this unit, a control box, consisting of a toggle switch (standby and transmit positions), two indicator lights, one for power, the other for end of message, and two circuit breakers are mounted on the Tally reader panel. A small aluminum box is used, only requiring a hole through the Tally panel for the cables and for the mounting screws, with all the control components mounting in the box.

A supply bin for 30' of 11/16" teletype paper tape is mounted to the read head mechanism to handle the pre-punched tape.

4.1.2 Card Basket

Card Basket and EECO "T" Module chassis size is approximately 8-1/2" x 6-1/2" x 6"; weight, approximately 6 pounds. This chassis is fabricated from aluminum in the shape of an open box with 10 taper pin Kennedy printed circuit connectors mounted at the bottom. Ten card guide grooves are machined in the sides for the cards. Along one side of the chassis is mounted a bracket holding the Seven EECO "T" modules (2 spares). Flanges are located at each end for mounting this chassis to the case panel by 4 screws. A cable and plug connect

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this unit to the filter chassis. The plug-in cards have components mounted between terminals and use point-to-point wiring. EECO "T" modules are purchased items. Refer to ACI drawing 100-1-03-078 for details.

4.1.3 Filter Chassis

This chassis is frabricated from aluminum plate. The transformer and filters are mounted to this chassis by a hold-down plate. A small component board and three connectors are also mounted on this chassis. The connectors form the junction box interconnecting the sub-chassis. The approximate size of this unit is 6-1/2" x 8-1/2" x 4"; weight, 9 pounds. Refer to ACI drawing 100-1-03-079 for details.

4.1.4 Power Supply

The power supply is purchased per ACI specification 100-1-10-006. For details refer to this specification.

4.2 Design of Receiver Units

4.2.1 Tally Perforator Unit

The Tally perforator 420 and drive package 1424 is packaged per ACI Drawing 100-1-04-074 on a 10-1/2" x 19" panel. The perforator unit is enclosed in a sound-proof box with a hinged cover over the punch head and chad box for access to the tape supply. The tape supply reel is mounted to the right of the perforator on the front of the panel. For details, refer to the ACI drawing. The perforator and drive package unit is ordered from Tally Register Corporation per ACI Specification 100-1-10-019. Approximate weight of this unit is 35 pounds.

4.2.2 Receiver Control Unit

The receiver control is packaged in a 10-1/2" x 19" panel x 14" deep chassis and mounted in a standard 10-1/2" cabinet modified to

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ACI drawing 100-1-04-056.

Mounted on the chassis are one card basket for 10 plug-in cards, one power supply (ACI 100-1-10-006), 2 component boards, the EECO plug-in modules, 7 filters, 2 transformers, connectors, and the necessary inner component cabling.

Switches, meters and controls are mounted on the front panel of the unit. Refer to ACI drawing 100-1-04-056 for details.

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5.0 PERFORMANCE DATA

Several tests were conducted on the breadboard and model to determine performance of the system. No precise performance specifications have been established by the customer but the following basic requirements appear reasonable from a systems viewpoint based on results obtained in the feasibility program:

1. Information rate: 25 to 50 words per minute
2. Voice quality: telephone toll quality
3. Tone level: set below threshold of audibility
4. Error rate: less than 1 percent
5. Ambient temperature: 0 to 50^{o}C

Some of these parameters are quite subjective and difficult to measure. For example, voice quality and tone audibility are very hard to measure in quantitative terms. The error rate is dependent on a number of factors - tone level, line distortion, pulses and crosstalk on the line, etc. The approach followed was to first determine a reasonable tone level based on listening tests by several observers and then to make error rate measurements at and around this tone level for different values of simulated line distortion. In addition, tests were made over several local telephone circuits with measured performance characteristics.

5.1 Breadboard Tests

5.1.1 Output Level

This test was performed to determine the AGC control characteristics using CW input. Figure 5.1.1 shows that the dynamic range of the AGC is 20 db and the variation in output level for this range of inputs is 1.0 db.

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5.1.2 Keying Rate for Various Speakers

Keying rate as a function of input level was measured using several different speakers. In each test, the input level was varied to determine the center of the AGC control range. This data is plotted in Figure 5.1.2.

5.1.3 Tone Audibility

In this test, each aural observer listens to the transmitter output (using headphones) which contains both a recorded voice and the information tones. The highest tone level at which each observer can no longer hear the tones is noted. Each observer listens to several different types of voices including both male and female. The same number of tones is keyed each time the threshold is exceeded, a much more severe condition than actually exists in normal operation. Table 5.1.3 below lists the audibility thresholds for the observers used in the test. It is believed that slightly higher tone levels than those indicated would be safe because the observers had a priori information on the characteristics of the tones.

TABLE 5.1.3

Tone Audibility

Threshold of Audibility -db below rms speech power

Observer	One Tone	Two Tones	Three Tones
	-30	-30	-27
	-30	-33	-36
	not audible	-27	not audible
	-27	-27	-33
	-27	-30	-30
	not audible	-27	-30
	-27	-24	-27

25X1

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5.1.4 Error Rate

A test was performed to determine the effect of line distortion on system error rate for several tone levels around -27 db. Four one-minute recordings of different speakers were used to make the statistics invariant. Line distortion was simulated using an amplifier whose peak distortion could be set using a switch. Figure 5.1.4 summarizes the data obtained in this test. The error rate for the breadboard is slightly poorer than that obtained using the model. This is probably due to the improper rejection characteristic of the filter used in the transmitter breadboard.

5.2 Model Tests

5.2.1 Keying Rate vs. Line Voltage

The effect of variations in line voltage on information rate was measured using the transmitter model and a tape recorded input. This data is tabulated below:

TABLE 5.2.1

Keying Rate vs. Line Voltage

Keying Rate - Char./Minute

Speaker	117 VAC	105 VAC	128 VAC
	290	287	303
	210	212	232
	200	220	225
	240	235	245

25X1

5.2.2 Keying Rate Vs. Temperature

Information rate was checked for the transmitter model over a temperature range of 0 to +50°C. Table 5.2.2 summarized the

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results obtained. A tape recorded input signal was used.

TABLE 5.2.2

Keying Rate Vs. Temperature

Speaker	$+25^{\circ}C$	$+55^{\circ}C$	$0^{\circ}C$	Max. Change From $25^{\circ}C$ Value
	251	242	257	3.6%
	195	183	193	6.2%
	162	146	153	9.9%
	222	197	220	2.3%

25X1

5.2.3 Error Rate Vs. Distortion

The error rate tests performed on the breadboard (5.1.4) were repeated using the model and a tape recorder. With the tones set 27 db below rms speech power and a peak line distortion of 1/2% (-46 db), the error rate is 0.3%. Data for other distortion values is listed below in table 5.2.3.

TABLE 5.2.3

Model Error Rate Measurements

Peak Distortion (db below peak speech)	Total No. of Characters Transmitted	Total No. Of Errors	Character Error Rate - %
-62	1143	1[1]	0.1
-54	1162	0	0
-40	1203	30	2.5

Note 1: This error occurred at the end of one message and may be due to transient from switching to "Standby."

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5.2.4 Local Line Tests

Error rate was also measured with the model transmitter and receiver connected through the telephone exchange. Two different configurations were tried [redacted] 25X1

1. Diamond-to-Diamond and
2. Diamond-to-Triangle

Before these tests were run, frequency response, distortion and insertion loss were measured for both lines. Frequency response data is plotted in Figure 5.2.4. Neither line exhibited any measurable distortion (output distortion was identical to input distortion which was less than 0.1% or -60 db.) Insertion loss for Di-Di is approximately 10 db while the Di-Tri is 20 db. Each line was dialed five times and a five-minute passage transmitted each time. The results are tabulated below.

TABLE 5.2.4

Error Rate over Local Circuits

Line	Total No. of Characters Sent	Total No. of Received Errors	Error Rate
Diamond/Diamond	860	0	0%
Diamond/Triangle	1092	57	5.2%

Errors in the Di-Tri test were apparently due to clicks from dialing at other locations which are either inductively or capacitively coupled to the line used in the test. Error-free copy was obtained on several occasions when the line was quiet.

5.2.5 Error Rate Vs. Receiver Gain Control Setting

In all of the previous error rate tests, the receiver gain control was accurately adjusted to establish proper tone levels at the input to

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the detectors. Since this control is set on voice peaks using the VU meter, it is subject to some error which depends on the care exercised and skill of the receiving operator. This test was made to establish both the resettability error using several operators and the variation in error rate due to improper setting of the gain control. Table 5. 2. 5a lists the reset errors for five operators using two different recorded voices.

TABLE 5. 2. 5a

Deviation in Gain Control from Optimum Setting Using Inexperienced Operators

Operator	Average Gain Error (Three measurements on each voice) Voice A	Voice B	
	-1.2 db	-1.8 db	25X1
	+0.8 db	+0.4 db	
	-1.0 db	-0.1 db	
	+1.8 db	+2.0 db	
	+0.6 db	-1.1 db	

With a keying threshold 6 db below the tone peak and a minimum keying threshold 8 db below nominal, the receiver AGC range is $\pm$ 4 db.

In every instance, operators were able to set the gain well within these limits as Table 5. 2. 5a shows. However, as a check on Receiver AGC performance, error rate was measured at these extreme ends of AGC control. Table 5. 2. 5b is a compilation of error rate vs. gain control setting with a simulated line distortion of 1%.

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TABLE 5.2.5b

Error Rate vs. Receiver Gain Control Setting
(1% simulated line distortion)

Gain Setting	No. of Characters Sent	No. of Errors Received	Error Rate - %
Optimum	900	21	2.3
4 db high	828	32	3.8
4 db low	860	13	1.5

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6.0 CONCLUSIONS

All tests conducted on the transistorized equipment developed under this contract confirm the feasibility, demonstrated previously with a vacuum tube version, of using audio tones to carry covert information via standard public telephone installations. A significant reduction in the transmitter size was accomplished by packaging and by grouping the five tones in a single wide slot between 1300 and 1900 cps as opposed to the five individual slots spread between 700 and 2400 cps employed in the feasibility model. In addition to size reduction, improved quality in the notched speech was obtained.

As a consequence of the extreme importance of system security, tone levels approximately 10 db lower than those used in the feasibility transmitter are presently being used. For this reason, receiver gain is 10 db higher making it more susceptable to noise and line distortion.

A method for obtaining the same degree of security at higher tone levels (6 to 8 db) through use of a magnetic tape delay technique (ACI proposal No. P-027) has been proposed. It is believed that this modification is essential to assure reliable long-range operation.

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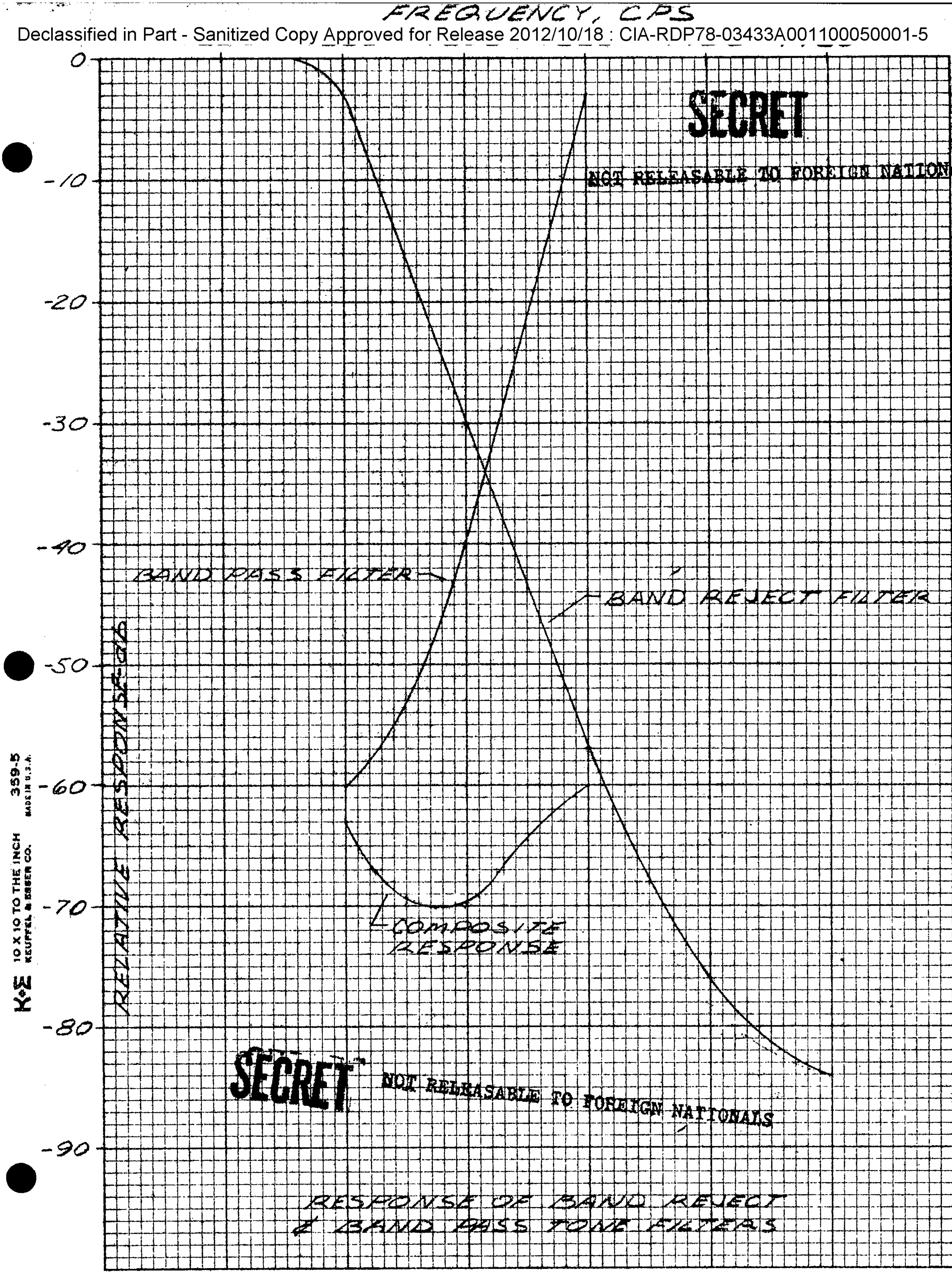

FIG. 2.3.1 a

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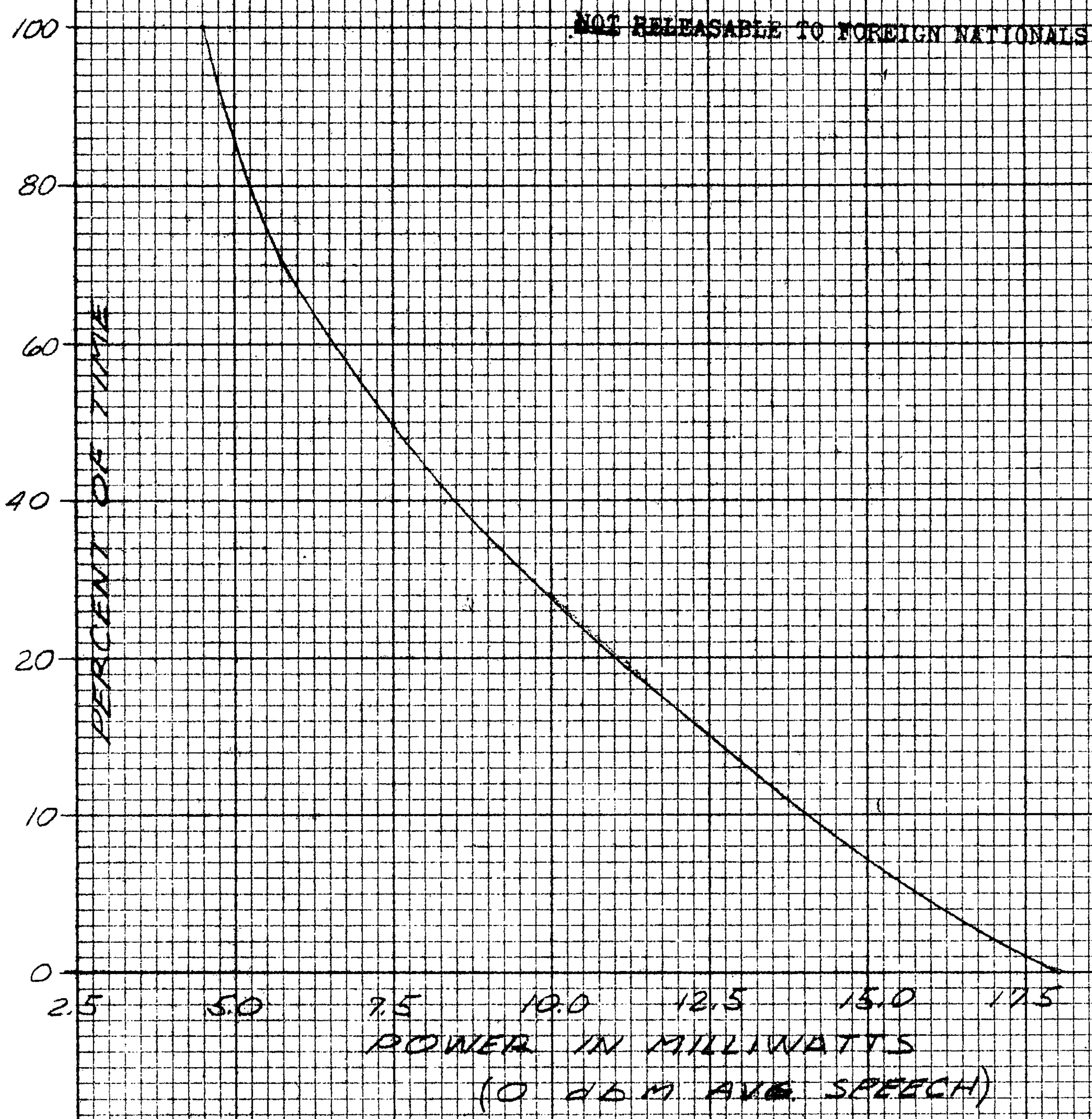

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SPEECH POWER DISTRIBUTION DURING TRANSMISSION

FIG. 2.3.1 b

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KEUFFEL & ESSER CO. MADE IN U.S.A.

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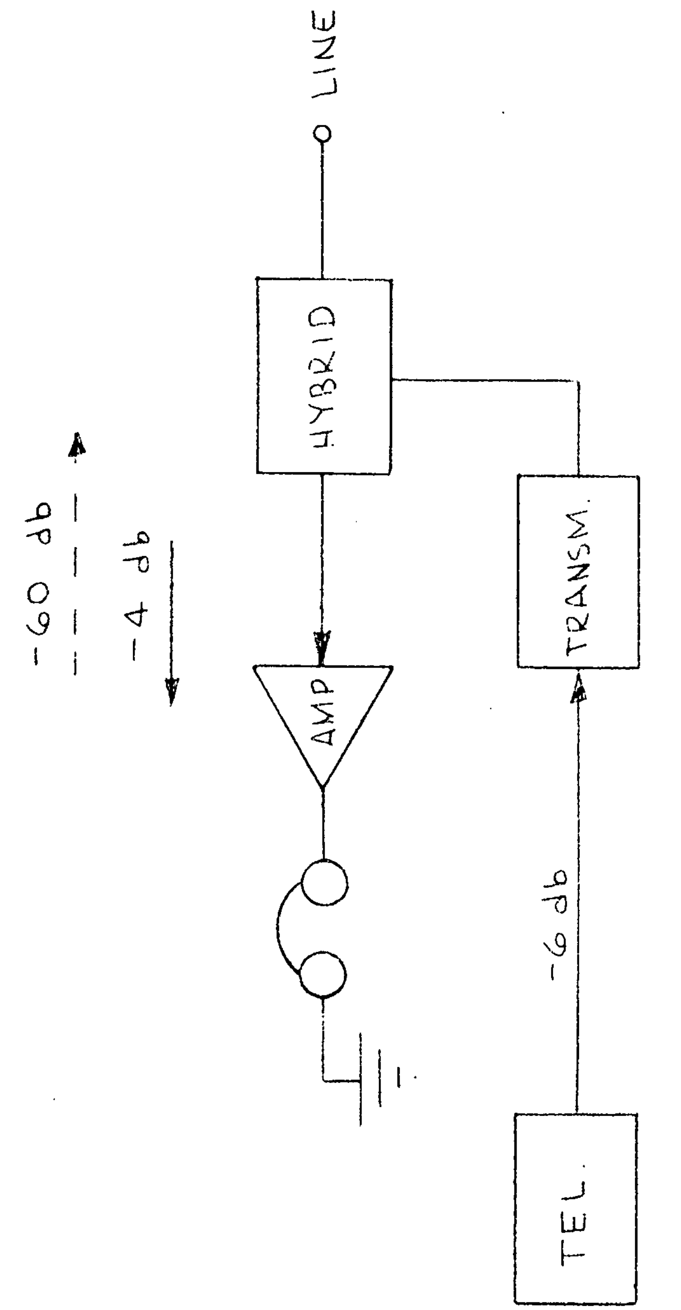

FIGURE 2.3.2 TRANSMITTER COUPLING

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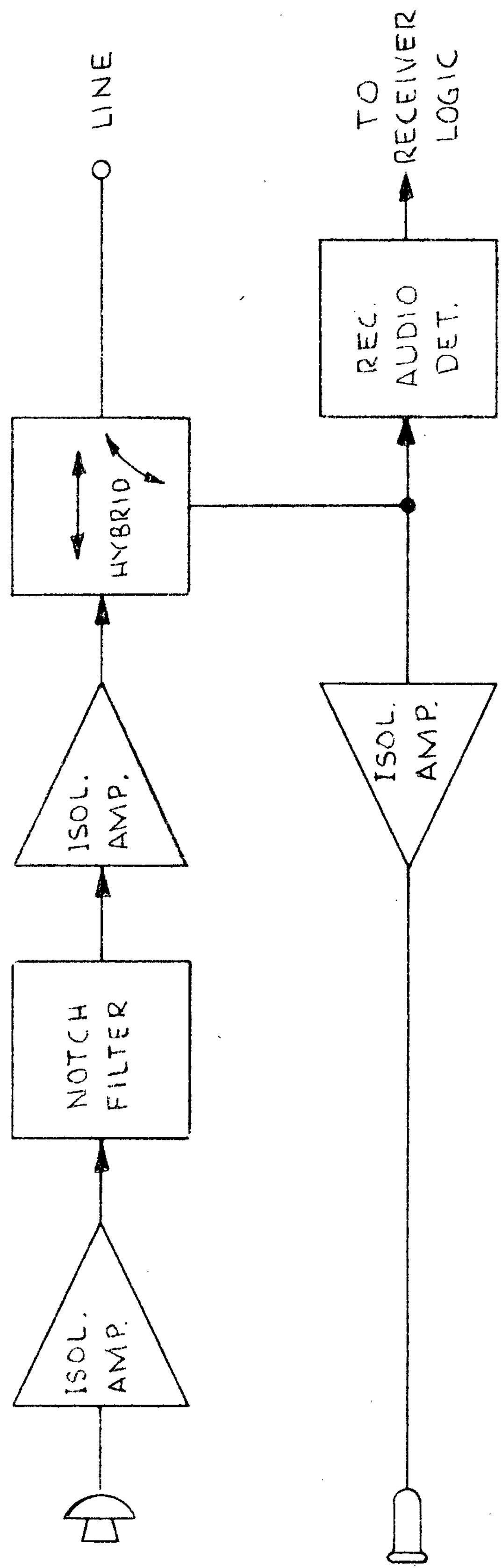

FIGURE 2.3.3 RECEIVER COUPLING TECHNIQUE

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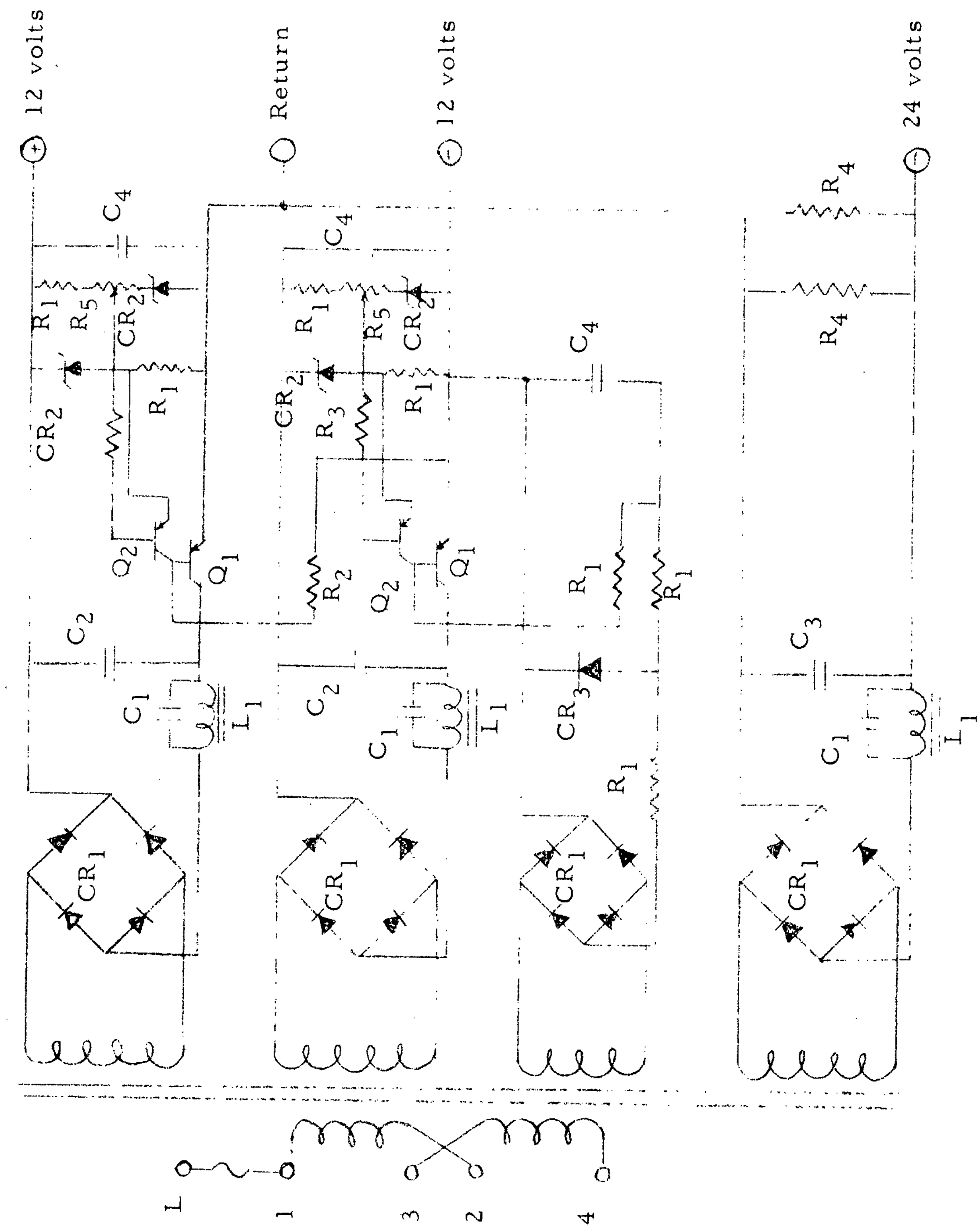

FIGURE 3.2.2 Power Supply, Schematic

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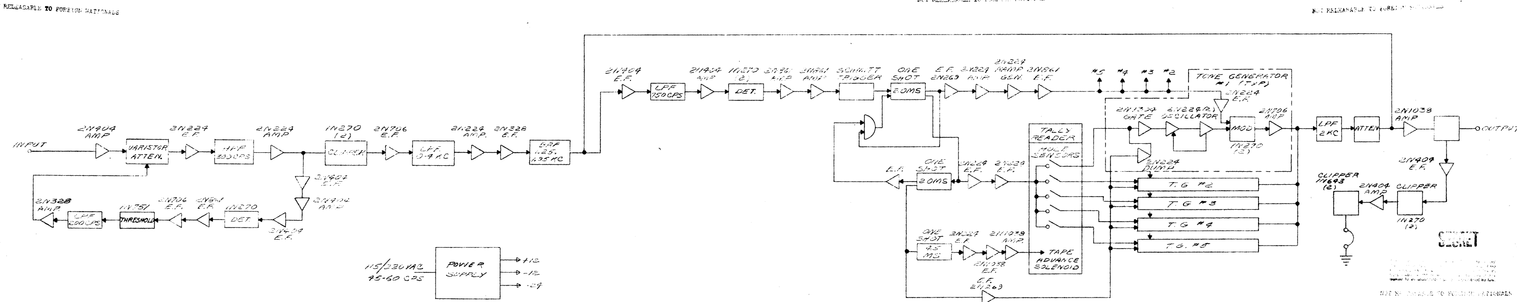

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PROJECT 100-1, PHASE I Copy # 11

FIGURE 6.0A
100-1-10-035 A

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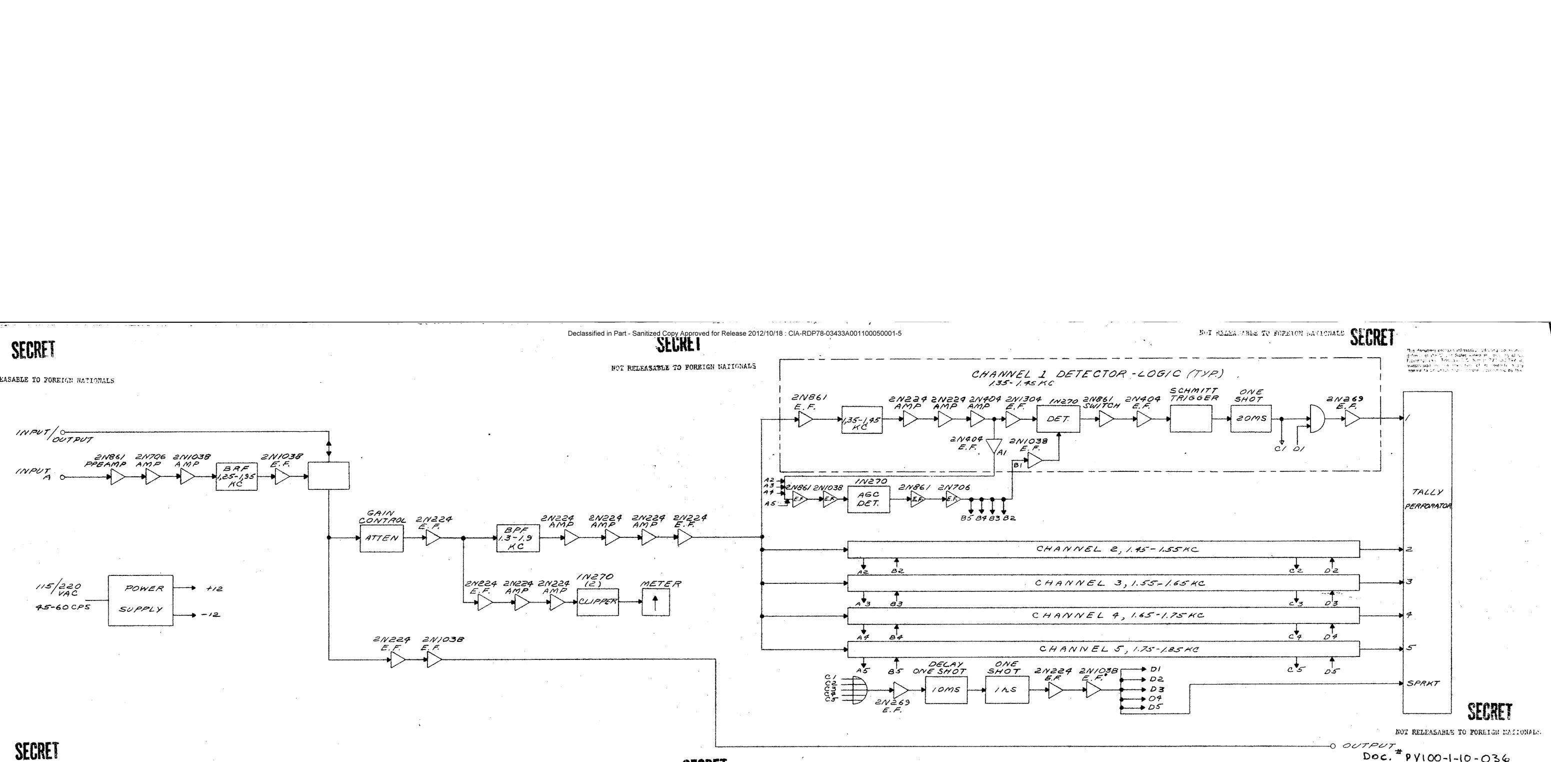

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RECEIVER BLOCK DIAGRAM
PROJECT 100-1 PHASE I

FIGURE 6.0A
100-1-10-036 A

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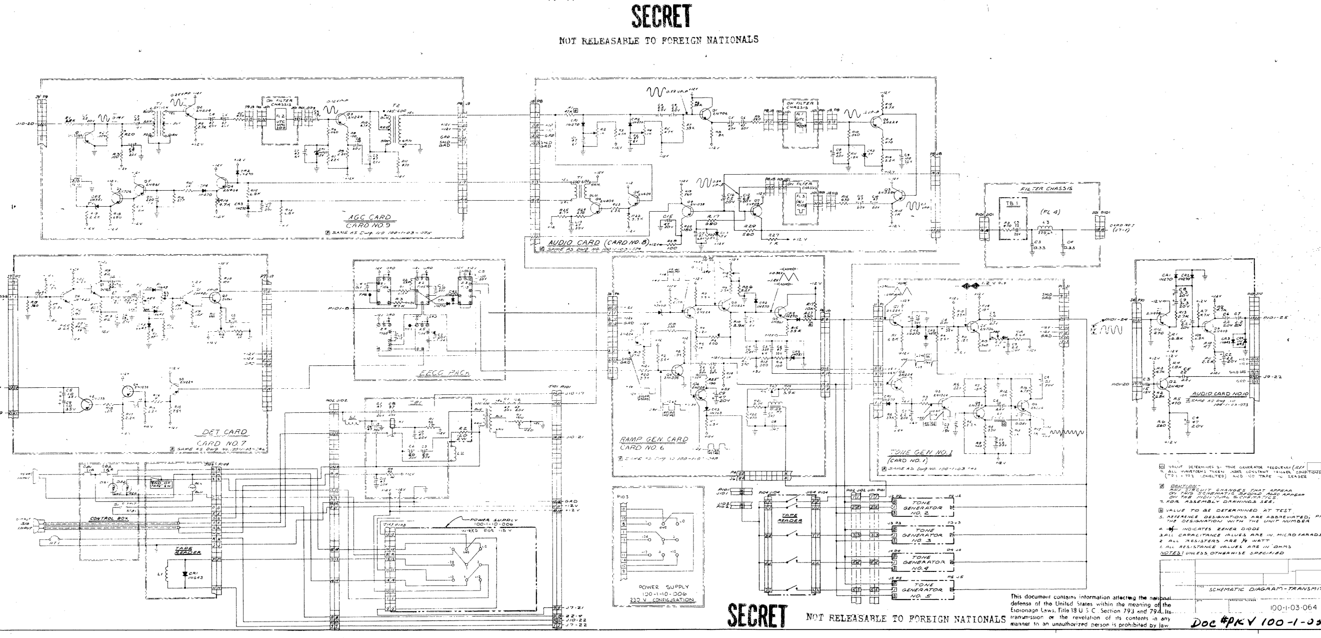

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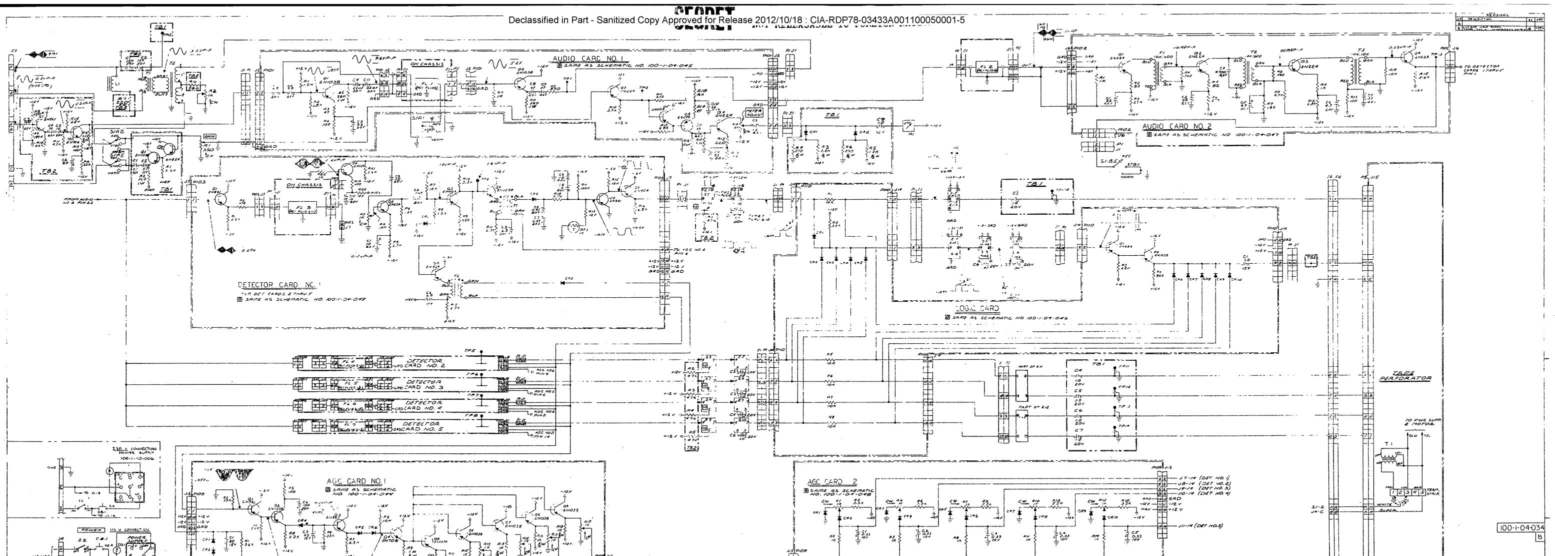

10. EECO T-106 MODIFIED ACCORDING TO DRAWING NO. 100-1-10-024
9. SIGNAL LEVEL MEASURED WITH GAIN CONTROL FULL CW, FUNCTION SWITCH REC POS., TRIANGULAR INPUT AT J3 ANY ONE OF TONE FREQ'S.
8. ANY CHANGES THAT ARE MADE ON THIS MUST ALSO BE MADE ON THE INDIVIDUAL CARD SCHEMATICS
7. ALL COMPONENTS NOT SPECIFICALLY LOCATED ARE EITHER ON CHASSIS OR FRONT PANEL
6. ALL DIODES ARE 1N270.
5. REFERENCE DESIGNATIONS ARE ABBREVIATED. PREFIX THE DESIGNATION WITH THE UNIT NUMBER.
4. INDICATES ZENER DIODE
3. FOR ASSEMBLY DRAWING SEE
2. ALL CAPACITANCE VALUES ARE IN MICROFARADS
1. ALL RESISTANCE VALUES ARE IN OHMS, 1/4 WATT

NOTES: UNLESS OTHERWISE SPECIFIED-

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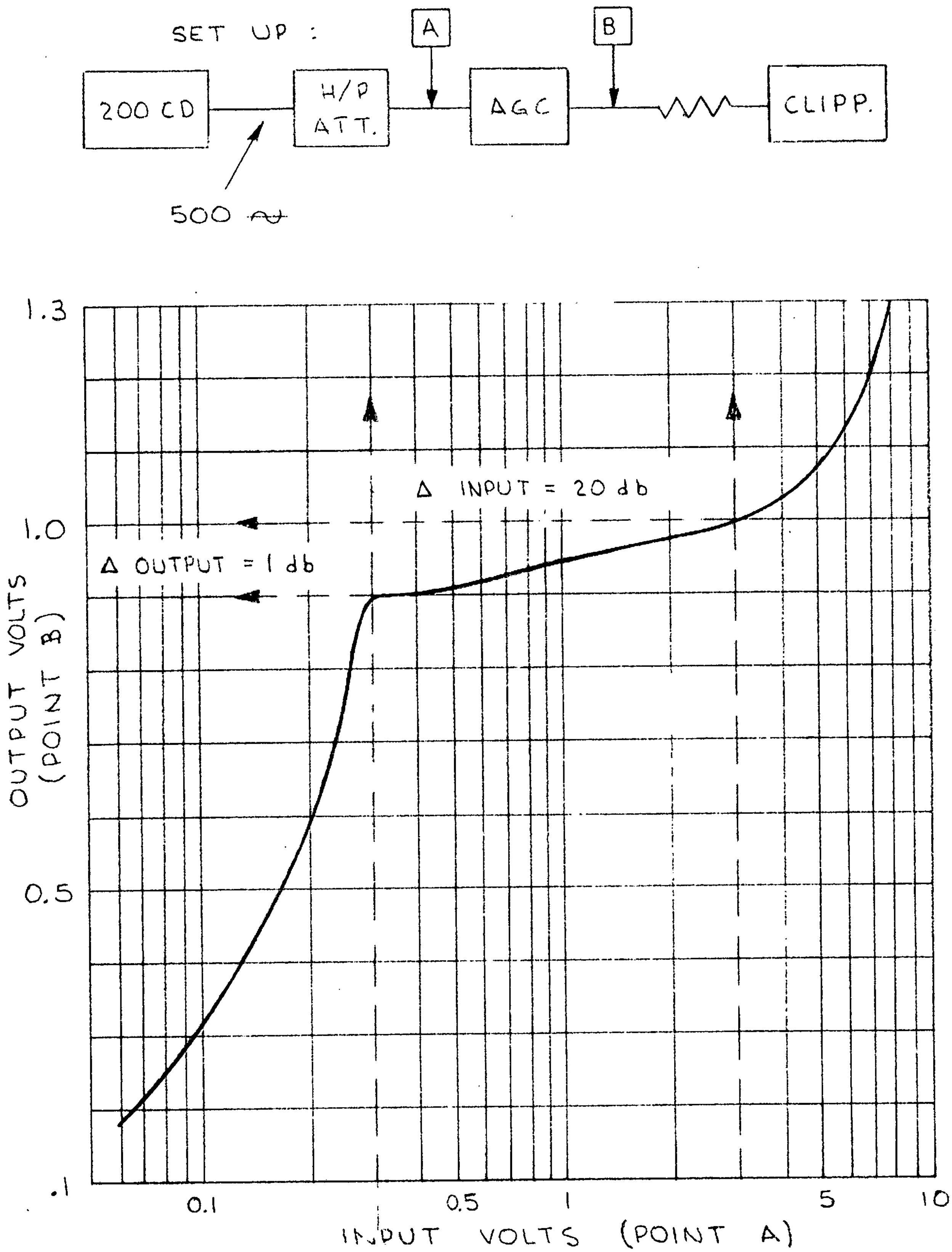

FIGURE 5.1.1 A.G.C. RESPONSE

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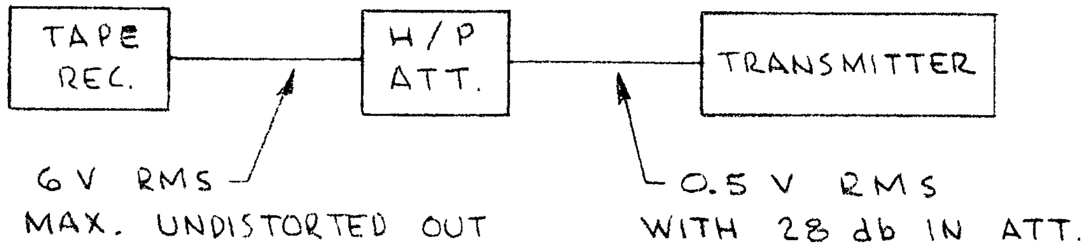
TEST SETUP:



CLIPPERS AT 12 db
THRESHOLD AT 145 MV RMS +20 db - 1.4 SV P-P

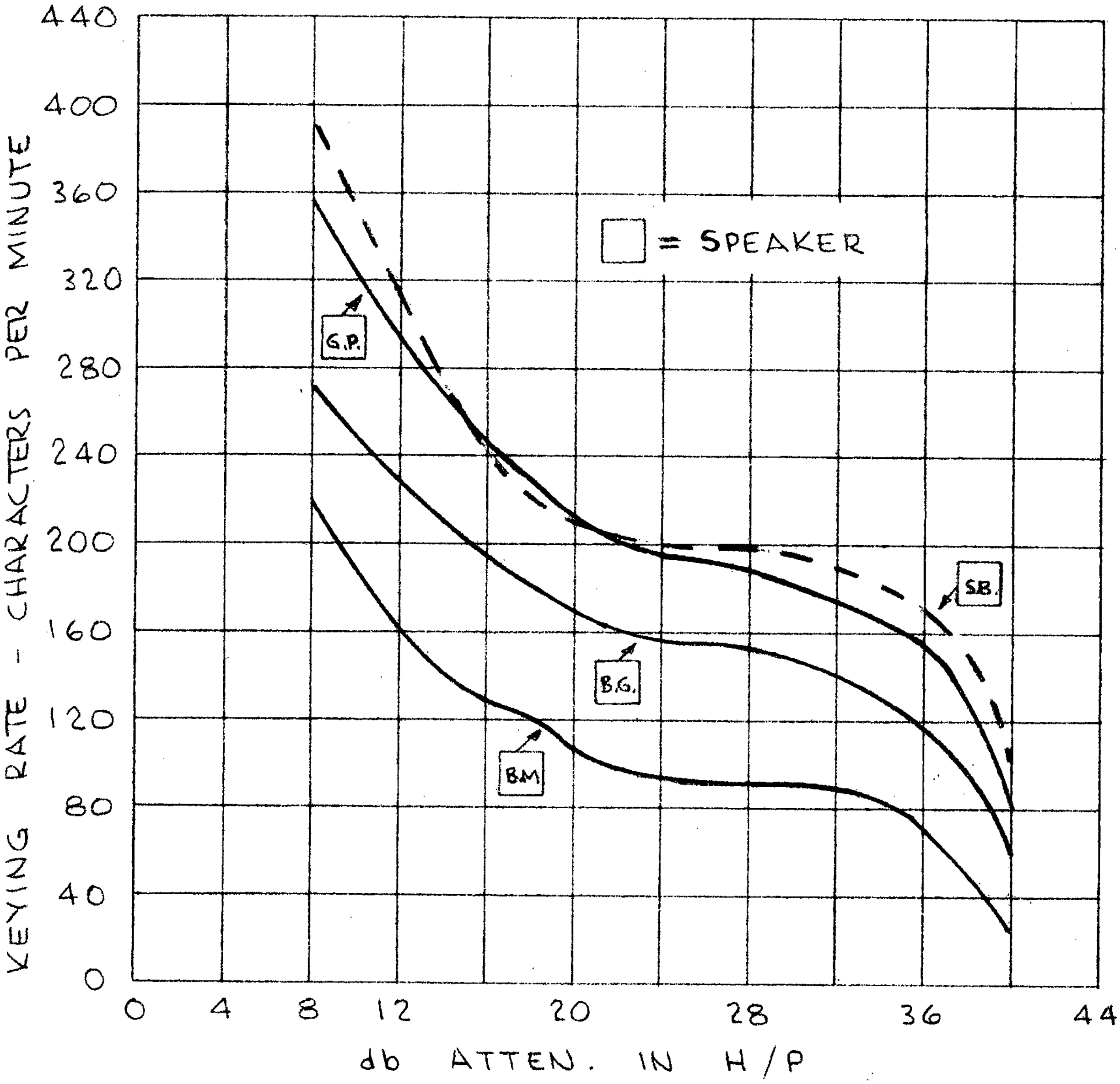


FIGURE 5.1.2 AGC CONTROL

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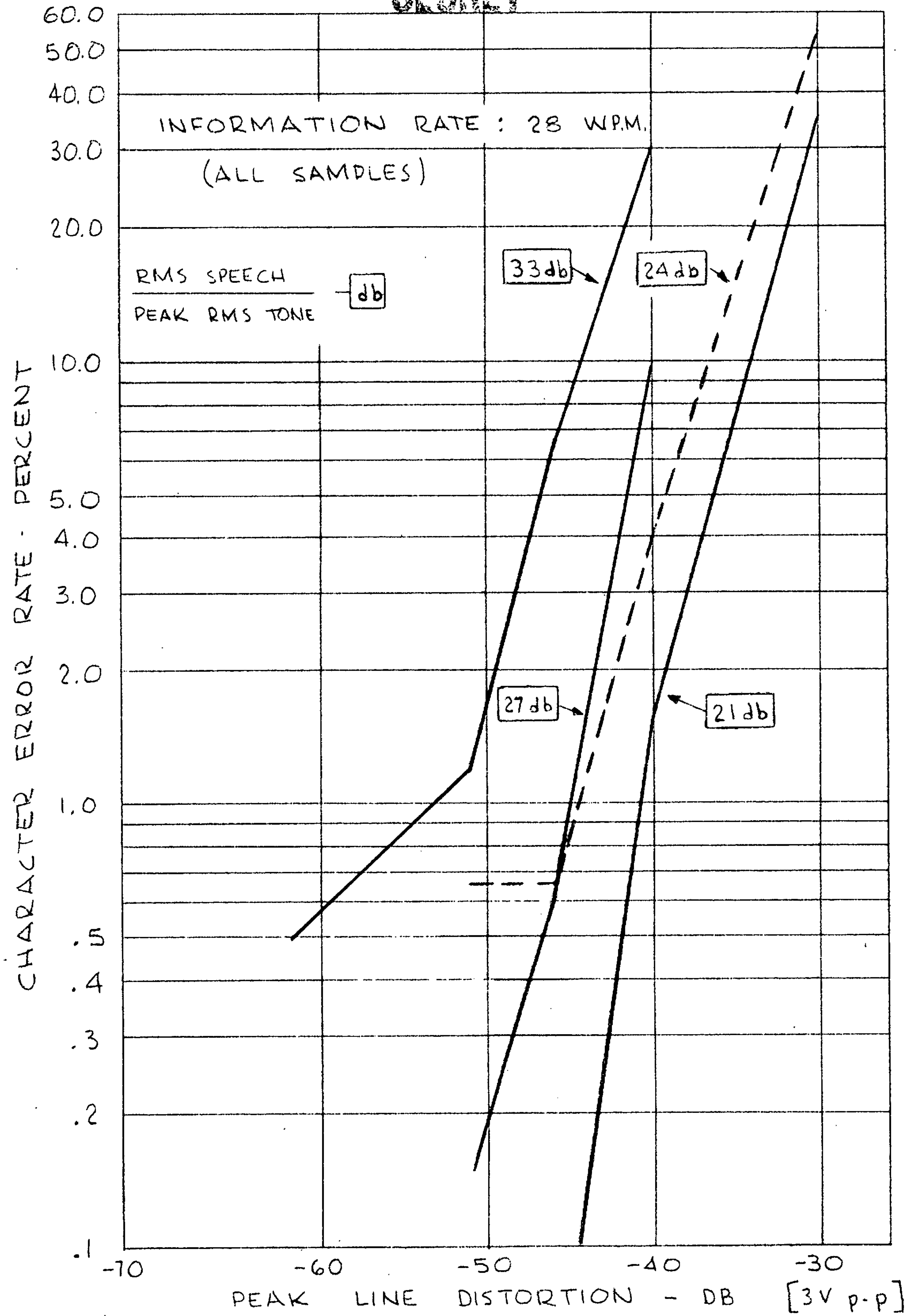

FIGURE 5.1.4 ERROR RATE VS LINE DISTORTION

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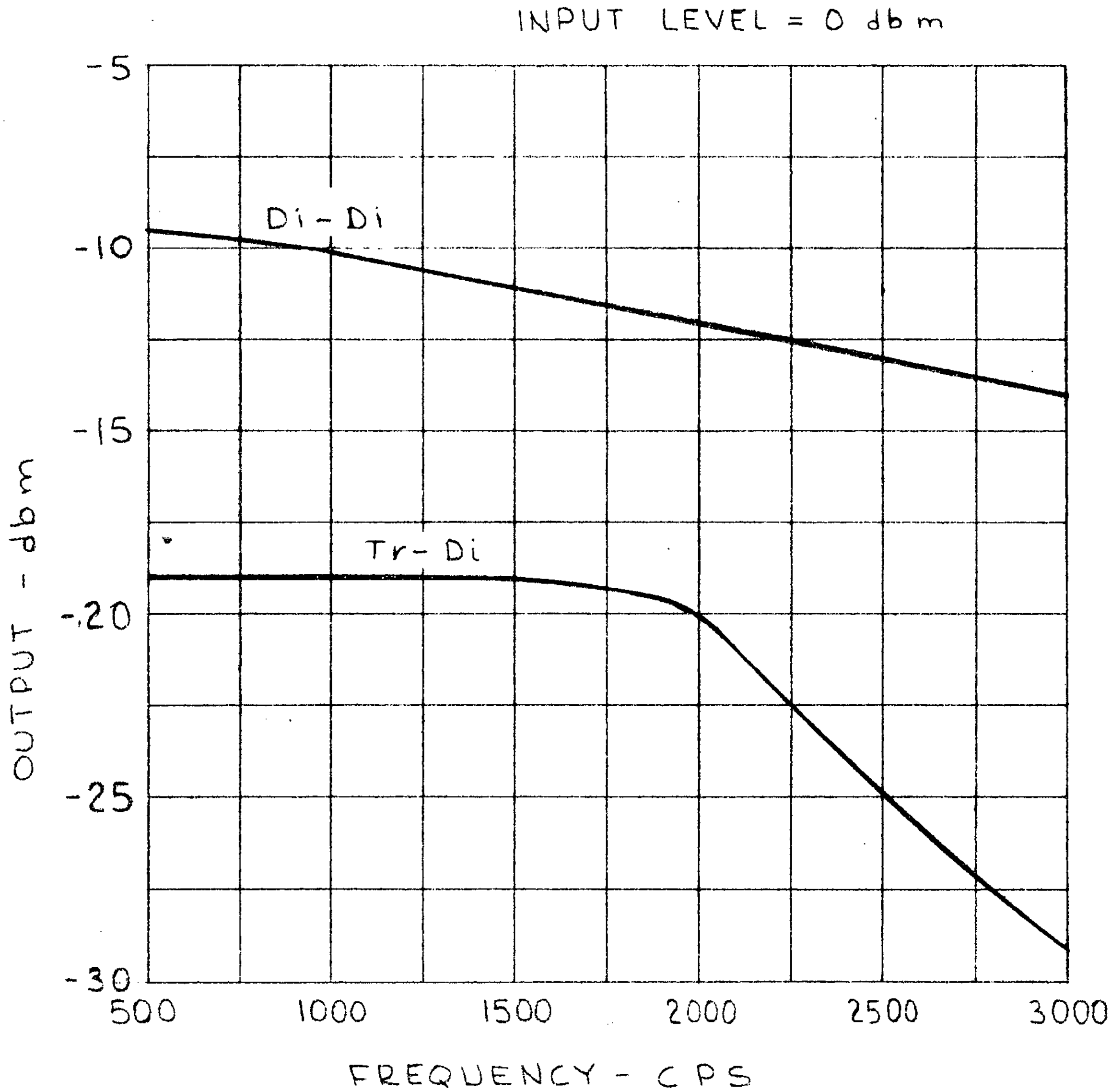

FIGURE 5.2.4 LINE FREQUENCY RESPONSE

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APPENDIX

TONE AUDIBILITY TESTS

I. Introduction

This report details the results of two experiments conducted in the interval 2 December 1961 to 5 December 1961 to determine certain characteristics about the audibility of tones masked by noise, and the statistics of speech after various kinds of filtering.

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II. THRESHOLD OF AUDIBILITY VS. REPETITION RATE

Figure 1 shows the test instrumentation for this experiment. Random noise (band-limited to 300 to 3 KC by the low-pass filter in the summing circuit and the series high pass filter) was added to pulse-modulated tones. The pulse shape was triangular with a 40 ms duration. An attenuator in series with the tones was used to adjust the levels so the threshold point could be determined.

With no attenuation in the tone channel, the relative levels of the tone and noise were as follows:

E_{2_n} = 32 mv rms

E_{2_t} = 10.5 mv rms (for 100% duty cycle)

This corresponds to a tone/noise ratio of -10 db

Table I contains the actual audibility data taken on four observers. To determine the threshold point, the observer would listen to the output and the operator would key a few tone-bursts through, and adjust the attenuator setting until the observer "heard" the bursts about 50% of the time. It was noted that the threshold was quite sharp, 1 or 2 db, making the difference between 5% audibility and 95% audibility.

a. Analysis of Results

The data showed striking uniformity for the observers used. For single pulses, the average tone/noise level for audibility was -13 db. As the repetition rate increased the audibility level was lowered until at 100% duty cycle, the average audibility level was -18 db, an improvement of 5 db.

As it relates to the system under consideration, the most significant quantity would be the change in audibility in going from a 100% duty cycle to, say a 40% duty cycle. The data indicate there is 3 db difference between these two cases.

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b. Conclusions

Reducing the repetition rate can improve detectability IF the limiting factor is bursts of tones, staccato effect.

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TABLE I

(see Figure I)

Initial Conditions:

E_{1_t} = 1.2 v p-p or 262 mv rms (40 ms pulses)

E_{2_t} = 10.5 mv rms (0 db attenuation)

E_{2_n} = 32 mv rms

Pulse Repetition Interval	Threshold - DB Below Noise Observer 1	2	3	4	Ave.
40 ms (continuous)	-17	-18	-19	-19	18.3
100 ms	-14	-15	-16	-14	15.0
200 ms	-15	-15	-15	-14	14.8
Single Pulse	-13	-12	-14	-12	12.8

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III. DE-SENSITIZATION OF EAR

It is a known phenomena that the ear tends to become de-sensitized to a weak signal if it is also receiving a strong signal simultaneously. No data were readily available as to the recovery time of the ear, and this experiment was performed to collect that basic data.

FigureII depicts the instrumentation used to perform this experiment. The output from the random-noise source was gated to a summing circuit and then through an amplifier to the headset. The gate circuit was also used to gate a single tone burst to this summing circuit, and the time position of this tone could be varied from being completely covered by the noise to being "in-the-clear" by a substantial time interval.

Table II contains the actual data taken during this test on three observers. The threshold was determined by having the observer listen for the pulse and adjust the attenuator until it was "barely heard." It was again noted that the threshold was quite sharp.

With the attenuator set at -10 db, the readings at the summing circuit were:

$E_{1_n} = 27$ mv rms

$E_{1_t} = 27$ mv rms

Therefore, with no additional attenuation, the tone/noise ratio was 0 db. Note: Figures in Table II were attenuator settings, and 10 db should be subtracted from them to arrive at the changes.

a. Analysis of Data

It became very evident as the data were being taken that the ear has a very quick recovery time. When the pulse was completely inside the noise, the average threshold was a tone/noise

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ratio of -13 db. (This compares very favorably with the previous test, even though two of the observers were different.) When the pulse was half in the clear (20 ms covered, 20 ms clear), the audibility changed to an average of -23 db, or 10 db different. And when the pulse occurred right after the noise ($\pm$2 ms) the average audibility level was -34 db or 21 db lower than for the completely covered case. Going from zero delay for the pulse to 50 ms delay (i.e. tone occurs 50 ms after noise ends), the audibility threshold average was -41 db, a net change of 24 db from being completely covered, but only a 7 db change from the zero-delay case. (Note: At the 50 ms delay the spread in thresholds has become much larger, as the limiting factor becomes the individual observer's hearing sensitivity.)

b. Conclusions

1. The ear recovers the majority of its sensitivity within a few milliseconds after the cessation of a louder signal.

2. Audibility of a pulse transmitted essentially "in the clear" is a function primarily of the observer's basic hearing sensitivity.

3. The system MUST be designed so as to minimize the number of pulses transmitted "in-the-clear."

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TABLE II
(See Figure II)

Initial Conditions:

E_{IT} = 27 mv rms, triangular wave shape

E_{IN} = 27 mv rms

Tone/Signal Ratio (Nominal) - 10 db

Attenuator Setting -10 db

Threshold - DB down from Noise

Time Delay - τ ms	Observer 1	2	3	Ave.
-40 (covered)	-24	-26	-20	23.3
-20 (half covered)	-33	-34	-33	33.3
2	-47	-45	-41	44.3
10	-48	-48	-43	46.7
20	-48	-51	-43	--
50	-51	-57	-45	--

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IV SPEECH STATISTICS OF FILTERED SPEECH

Figure III depicts the setup used for this experiment. Taped passages from four speakers were used as the input signal and manually adjusted to about the same rms level using the Ballantine True RMS Voltmeter with an additional 14-second integration time. This signal was then fed through an AGC circuit and a band reject filter which eliminated energy in the frequency band of 1300 - 1900 cps. This signal was then fed through an LPF to the threshold and transmitter element. The threshold was adjustable by attenuator settings.

Table III contains the data for RMS voltage readings taken at the four points of interest in the system which show what is happening to the speech power and gives some "feel" for the spectral distribution. An amplifier gain was adjusted between the output of the LPF and the input to the transmitter so it would see essentially a constant signal.

At each of the LPF settings (four in all) the data were taken as to the number of characters transmitted per second for different threshold levels. The basic raw data are contained in [redacted] notebook, and are summarized in Figures 4 - 7 for LPF bandwidths of 500 cps, 750 cps, 1000 cps and 3000 cps respectively. Figure 8, then, is a composite of the "average" curves taken from Figures IV & VII. In the case of the "average" curve for 750 cps LPF, the rms deviation from the curve for the data was computed and was 0.71 characters per second. 25X1

Visicorder graphs were made during the tests which showed the band reject filtered speech and the transmitted pulses. From these graphs it was possible to determine the number of pulses that were transmitted "in-the-clear" or partially so. This data

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is in the notebook. It is too difficult to present in this paper. Details of these experiments are contained in [redacted] notebook. 25X1

a. Analysis of Results

As would be expected, the presence of an LPF reduces the transmission rate in a rather direct relation to the filter cutoff frequency. There is also an essentially linear relation between keying rate and threshold setting, at least for high thresholds as we were using. Finally, "in-the-clear" transmissions are minimized as the LPF cutoff frequency is lowered and the threshold is raised.

The AGC works extremely well regardless of the speaker's characteristics and holds the output quite constant. As is evident from the voltage readings, the filters do have different effects on different speakers. It was not as pronounced a difference between male and female as might have been expected.

b. Conclusions

1. An LPF feeding the transmitter is highly desirable to reduce "in-the-clear" transmissions. A cutoff frequency of 750 cps seems appropriate.

2. The threshold setting is somewhat critical as it effects both transmission rate and "in-the-clear" transmissions. A threshold of 14 db above the rms power that exists after the AGC circuitry seems appropriate.

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TABLE III

Signal Levels for Filtered Speech (Refer to Figure 3)

Speech Level - RMS volts after 14 seconds integration

E_4 for LPF cutoff frequency

Speaker	E_1	E_2	E_3	500 cps	750 cps	100 cps	3000 cps
1	.450 0db	-135 0 db	.150 0 db	1.27 0 db	1.27 0 db	1.29 0db	1.25 0 db
2	.410 -.8	.138 +.2	.150 0	1.38 +.7	1.30 +.2	1.37 +.5	1.28 +.2
3	.465 +.3	.135 0	.122 -1.8	1.20 -.5	1.05 -1.5	1.12 -1.2	1.08 -1.2
4	.535 +1.5	.132 -.2	.140 -.6	1.48 +1.3	1.35 +.5	1.30 0	1.22 -.2

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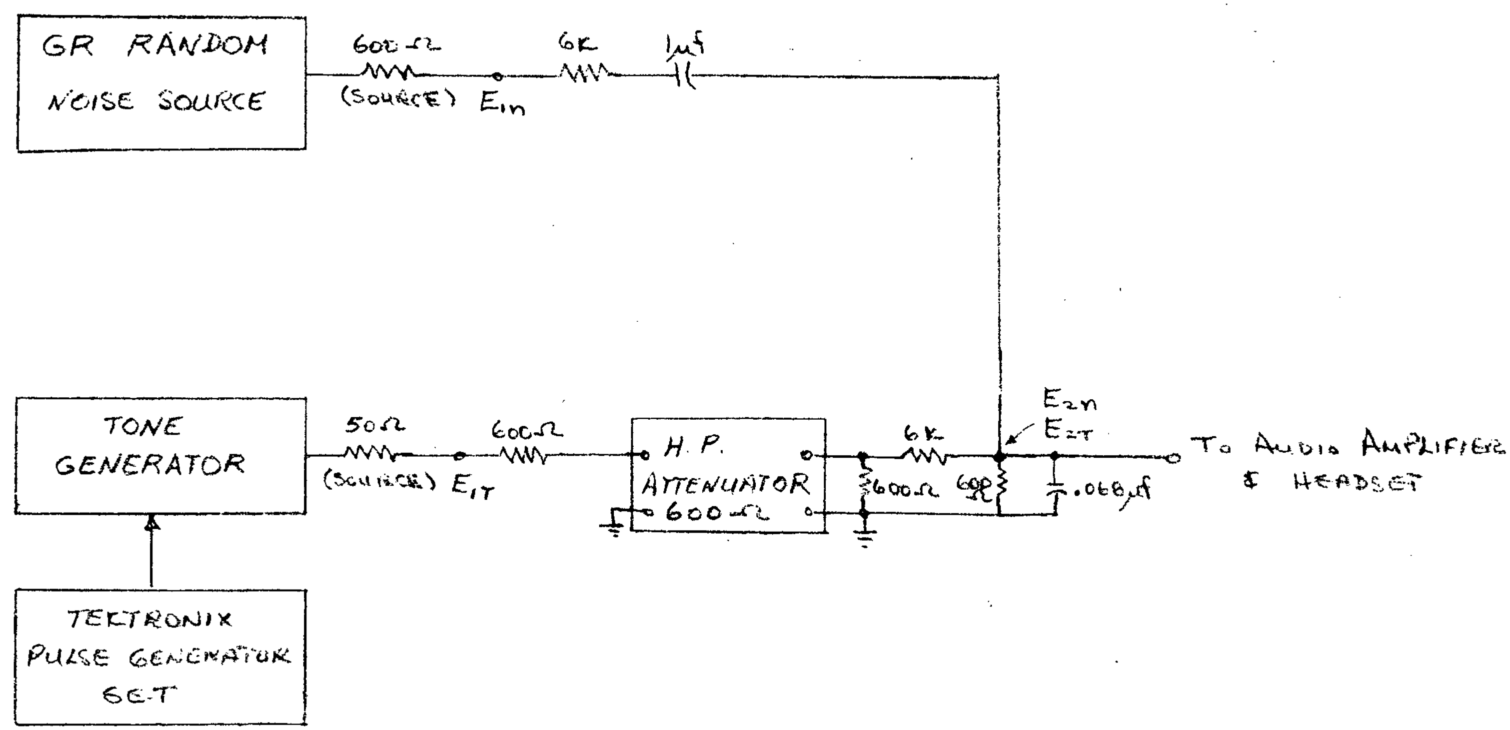

FIGURE I. — BLOCK DIAGRAM, AUDIBILITY THRESHOLD TESTS

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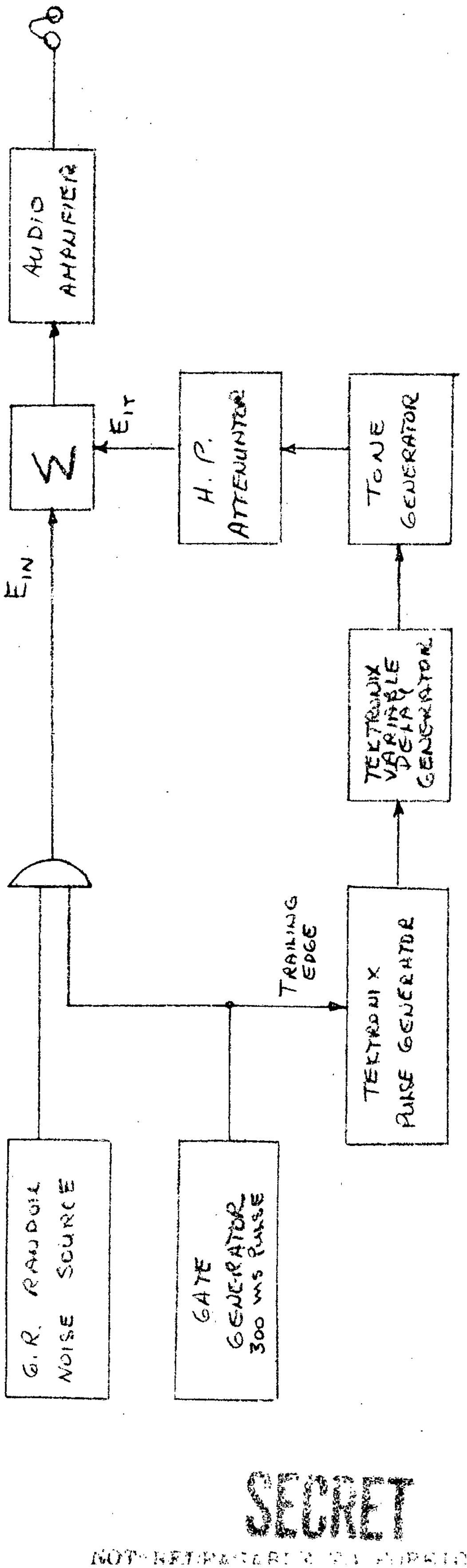

FIGURE II.- BLOCK DIAGRAM, EAR DE-SENSITIZATION EXPERIMENT

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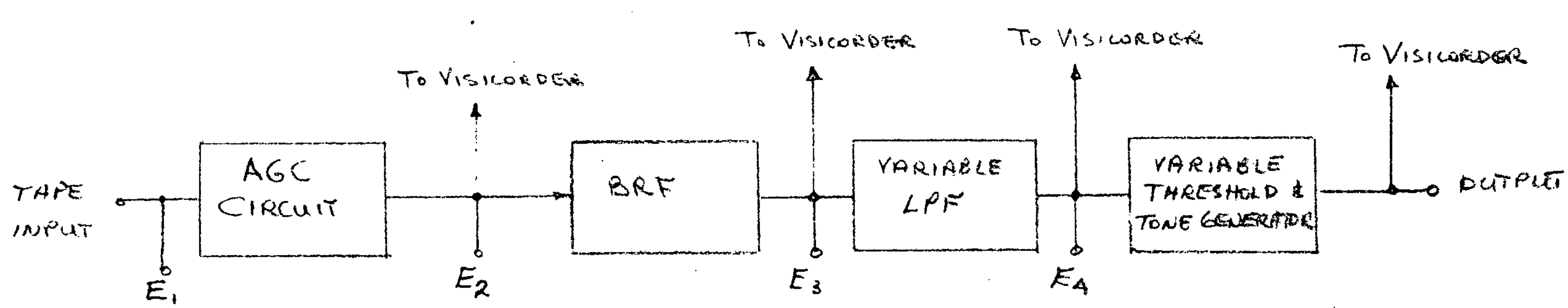

FIGURE III. - BLOCK DIAGRAM, FILTERED SPEECH STATISTICS EXPERIMENT

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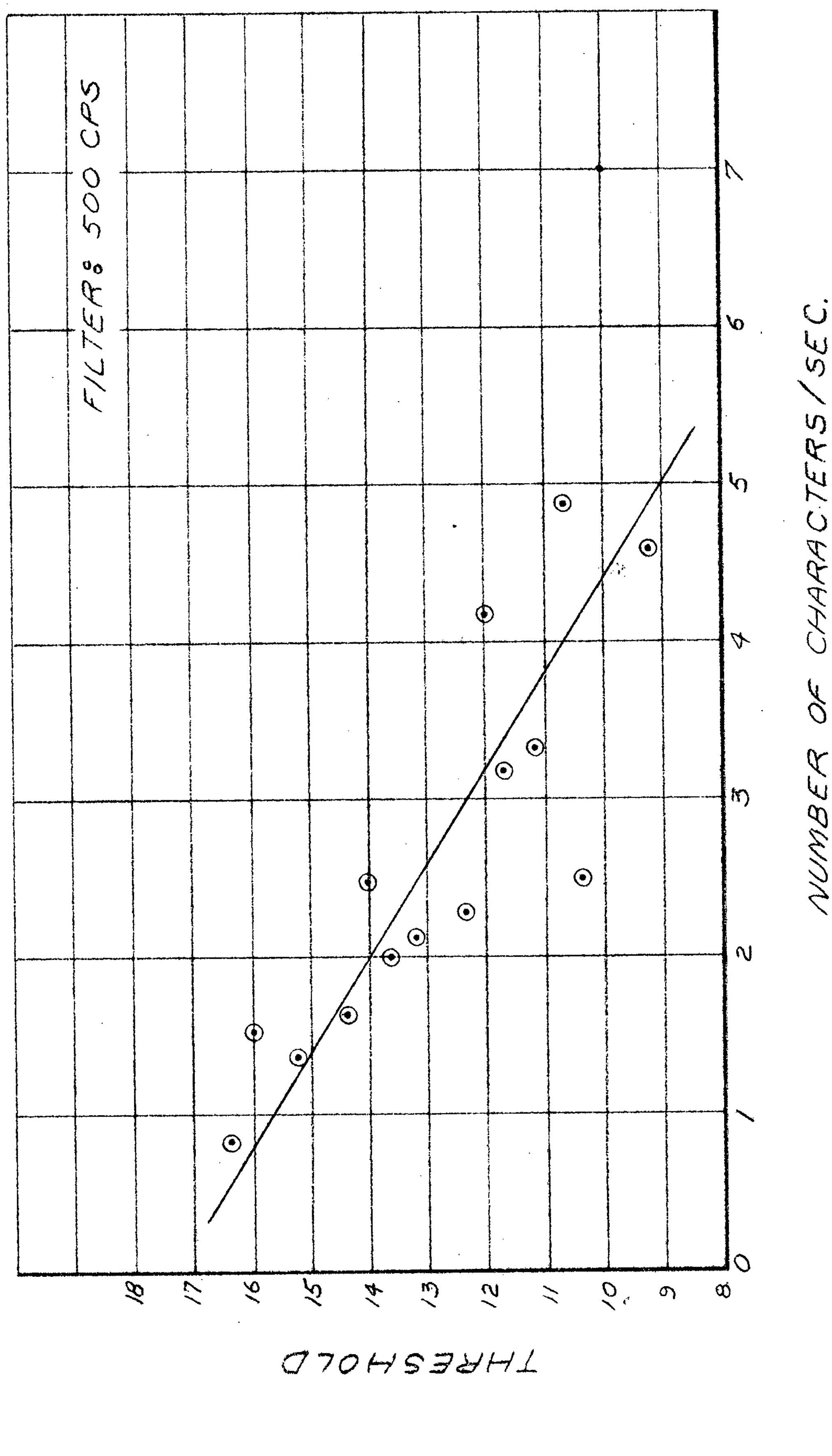

FIGURE IV - TRANSMISSION RATE VS THRESHOLD LEVEL, LOW PASS FILTER CUTOFF OF 500 CPS.

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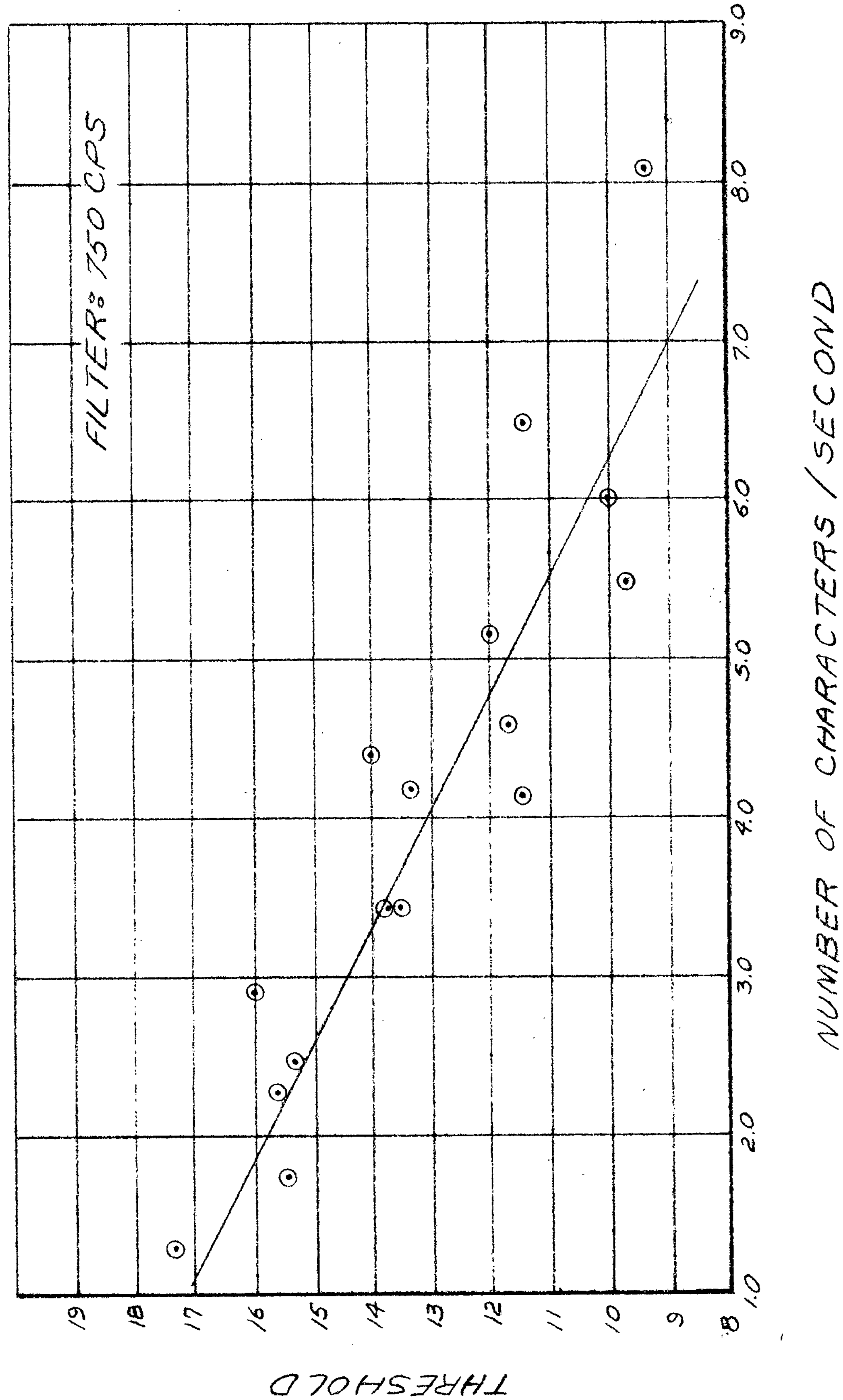

FIGURE II - TRANSMISSION RATE VS THRESHOLD LEVEL, LOW PASS FILTER CUTOFF OF 750 CPS

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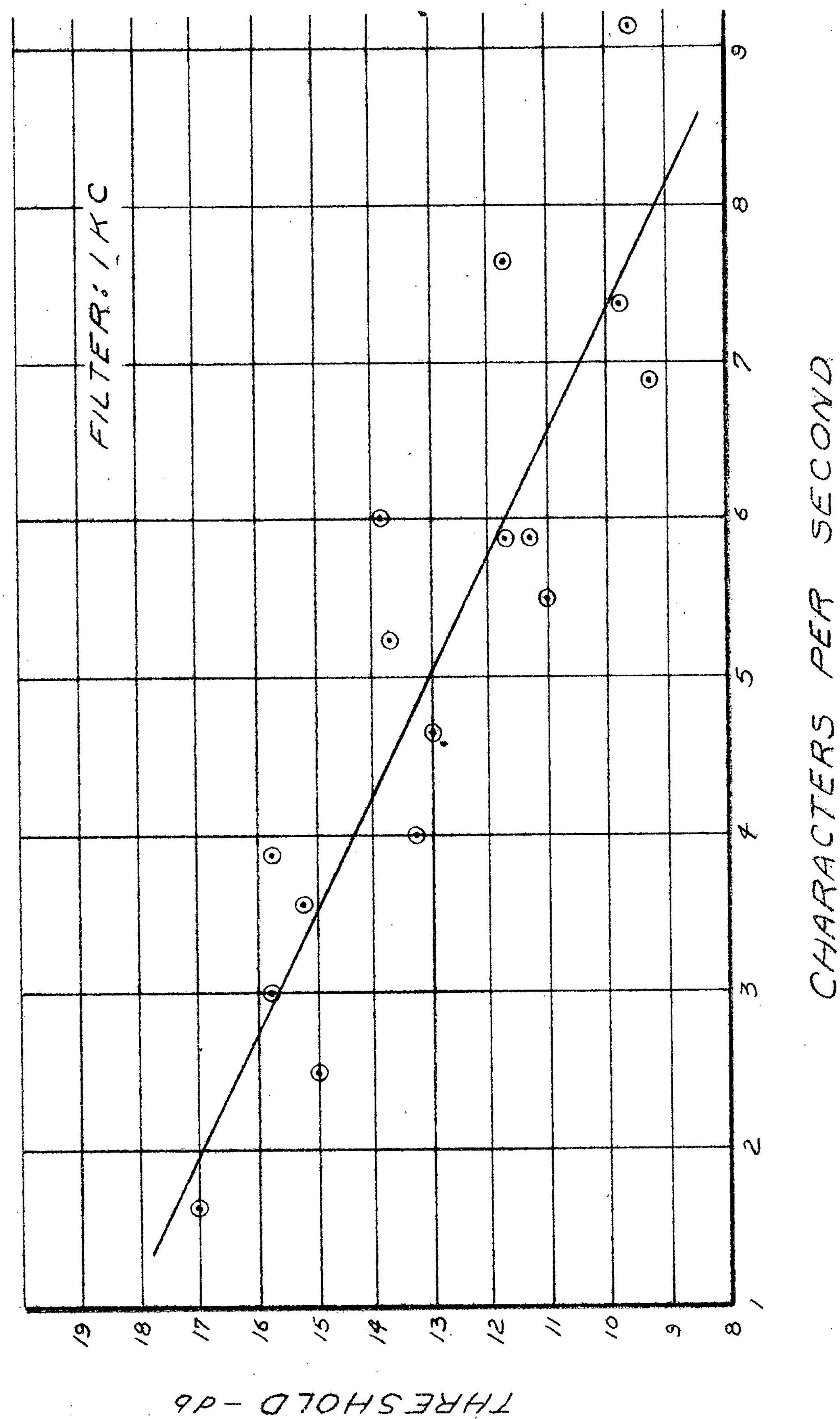

FIGURE III - TRANSMISSION RATE VS THRESHOLD LEVEL,
LOW PASS FILTER CUTOFF OF 100 CPS

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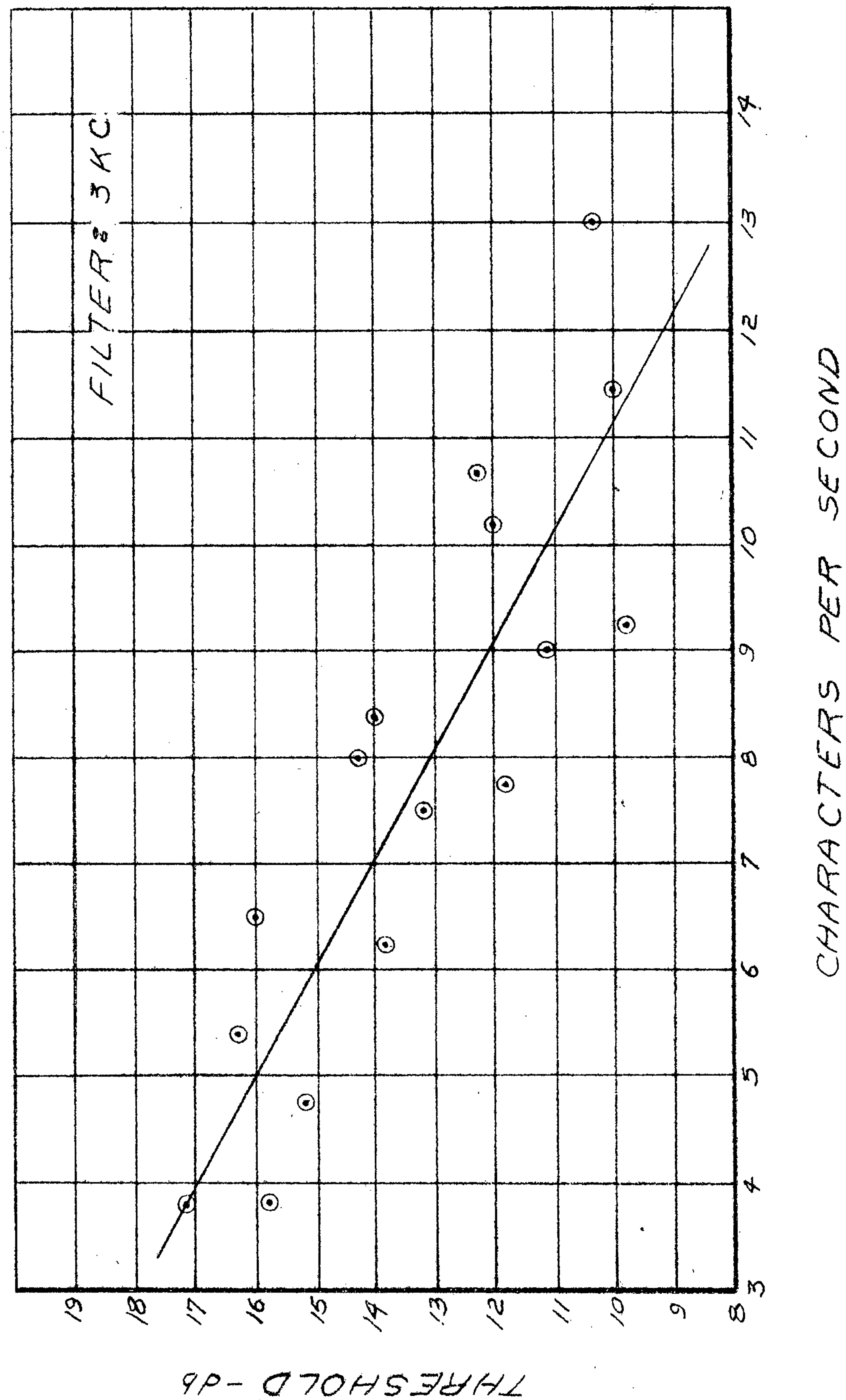

FIGURE VII— TRANSMISSION RATE VS THRESHOLD LEVEL, LOW PASS FILTER CUTOFF OF 3000 CPS

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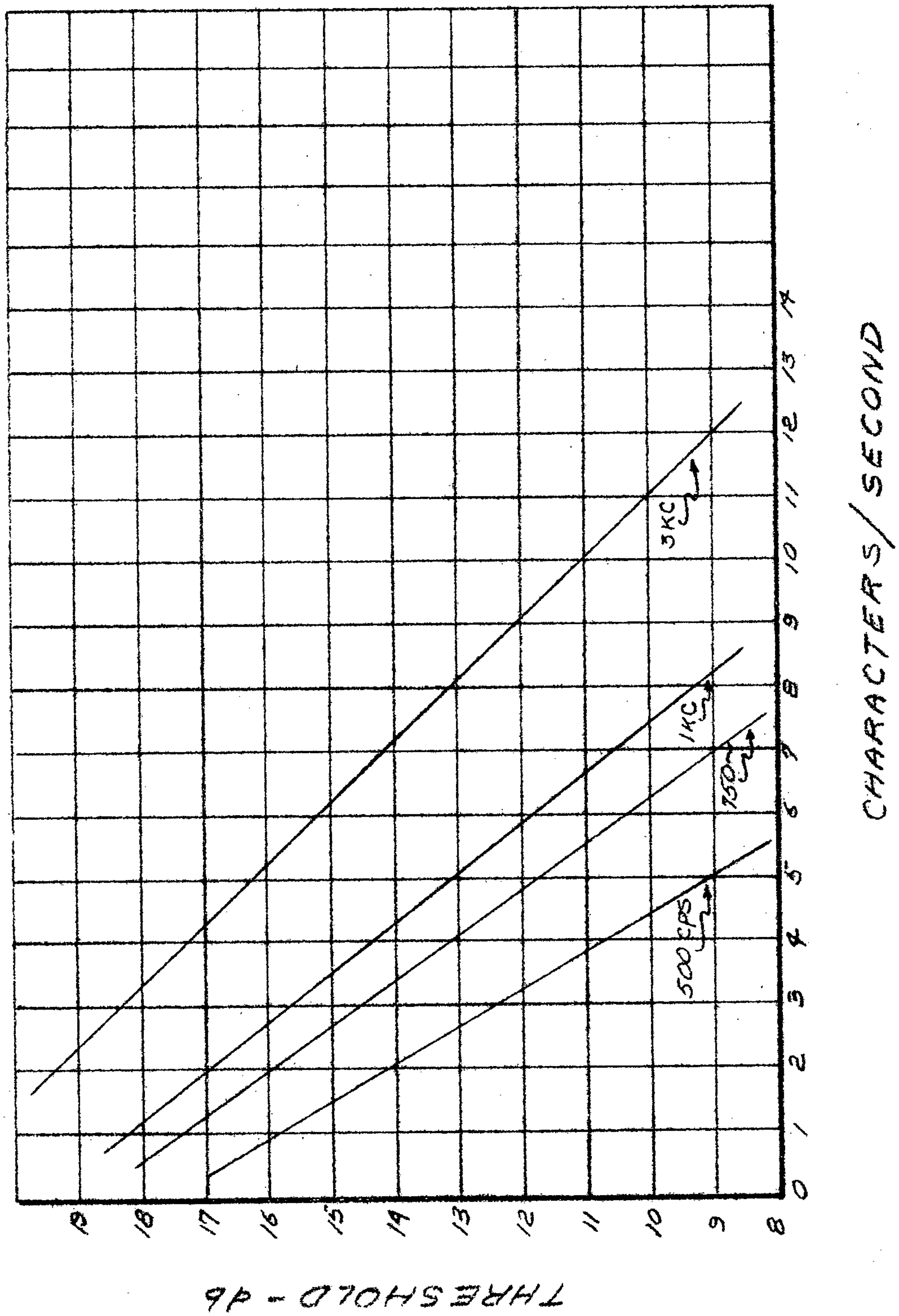

FIGURE VIII - NOMINAL TRANSMISSION RATE VS THRESHOLD LEVEL FOR VARIOUS LOW PASS FILTERS

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