# A TRANSISTORIZED DIGITAL COMPUTER WITH BOTH REAL AND STORED TIME ANALOG READOUT OF INFORMATION - FOR USE IN DEEP SPACE INVESTIGATIONS OF MICROMETEOR PHENOMENA

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## PREFACE AND ACKNOWLEDGEMENT

The subject for this paper was taken from work performed at the Oklahoma State University Research Foundation under Contract No. AF 19(604)-5715 during the period 1 May 1959 through 30 April 1960. During this period the author was employed by the Research Foundation as Projects Engineer directly responsible for all work performed under Contract No. AF 19(604)-5715.

Under the above contract, a system was designed and developed, and 24 units produced and delivered to Air Force Cambridge Research Center. Each unit is essentially an electronic system designed for the express purpose of exploiting the data-gathering potential of a specific space vehicle. However, not all sections of the data-gathering system are the original work of the author, some sections having been conceived, developed and carried over from previous contracts. For this reason, only the heart of the system, namely the digital computer section is dealt with in this paper.

In this paper a brief historical background of the development of an "acoustic" micrometeor detection technique is presented, together with details of the specific system evolved for this application. Various design features which include logic, powering, environment and packaging peculiar to this application are presented and discussed. The last chapter evaluates the final production units delivered to AFCRC in terms

of reliability and functional compatibility.

The author wishes to express his most sincere appreciation to Mr. Richard F. Buck for his many valuable comments and suggestions without which the project described in this thesis would have proven an almost insurmountable task. Also special thanks are due Mr. Paul A. McCollum who unselfishly encouraged and aided the author as thesis supervisor.

Mr. Dahl Mitchell, technical assistant to the author, was extremely cooperative in his efforts without personal gain to ensure that the many deadlines and target dates during production of the system were met. The many discussions with Mr. Mitchell concerning problems within the system proved extremely helpful.

Finally, the author wishes to thank Dr. Robert Soberman for permission to use the thesis subject; Mr. Jack Tompkins for lending his personal background of experience in production techniques; Mr. Larry Labarthe for his timely comments concerning saturating core oscillator theory; and Mr. Emerson Oaks for his guidance in the mechanical design of the system package.

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# CHAPTER I

## INTRODUCTION AND HISTORICAL SKETCH

The recent assault by man on the barrier of outer space has brought about a great impetus and stimulus for him to investigate the many phenomena of outer space which heretofore had either been mysterious or entirely unknown. One such phenomenon which lately has drawn more and more attention and investigatory effort is "shooting stars".

Today, we know "shooting stars" to be, in reality, small material bodies that "burned" upon their passage through the atmosphere of the earth. A considerable amount of the early work leading to many (and since verified) deductions was done by Brandes and Benzenberg of Germany near the end of the 18th century. Since this early work, investigations have continued to the present day.

The evolution of "shooting star" or meteor study has resulted in, among other things, a second field of study branching away from the first. This study is concerned with the phenomenon of "micrometeors" rather than meteors as such. The basic differentiation between a meteor and a micrometeor was first made by Whipple in 1950 when he developed a theory of micrometeorites, and in which he worked out the necessary conditions for a meteoric mass to be able to reach the ground without having been completely vaporized. A major point of Whipple's theory is that micrometeors, contrasted to meteors, are sufficiently small to reradiate the heat gained in collisions with air molecules before vaporization of the micrometeor begins to occur. Whipple coined the name "micrometeorite" for meteoric particles that are small enough to reach the surface of earth without having been vaporized in their passage through the atmosphere of earth.[1]

As is true with any experiment, once a need for study was justified, and the experimental data required was established, the next step was to obtain the necessary equipment to measure and record the desired information. One important micrometeor experiment concerns the influx rate and so-called momentum of micrometeors in the upper earth atmosphere, and also outer space. Here, influx is taken to be the number of incident micrometeors per unit of area per unit of time; momentum is the product of the incident particle (micrometeor) mass and an average geocentric velocity.[2]

Contract AF 19(604)-5715 was entered into with Air Force Cambridge Research Center in order to provide the necessary personnel, facilities, and services to accomplish design, development, and construction of special meteoritic microphone detectors for use on satellites and probe vehicles. The detection system was to have a momentum sensitivity in the neighborhood of $10^{-4}$ gram-centimeters per second for impacts on collecting surfaces provided on the vehicle which is used to carry the experiment.

Similar work had been carried on earlier by the Research Foundation, O.S.U., for the purpose of obtaining information by the so-called "acoustic" method on the influx of meteoritic material in the neighborhood of the earth.

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[1]Curtis W. McCracken, "An Analysis of Rocket and Earth Satellite Measurements of Micrometeoric Influx" (unpub. M.S. Thesis, O.S.U., 1959), pp. 1-12.

[2]Scientific Report No. 1, (AF 19(604)-5715, O.S.U., 1959), pp. 1-2.

Under Contract AF 19(604)-1908, a number of instrumentation systems were provided and flown on various types of rockets for this purpose. Under this earlier contract, systems have been developed which are capable of detecting acoustic stimuli created by the impact of small particles on the metallic surface of either the rocket skin or special collecting sections which have been inserted in the rocket skin. In all cases, the electrical impulse provided by the mechanical stimulus has been amplified, detected and (through suitable electronic circuitry), made available as a telemetry input voltage for subsequent radio reception and recording on the ground.

Various designs and models evolved through the period of three years in which this work was conducted. All previous designs have shared one common feature: a real-time recording, event by event, has been obtained. In very early experiments, the direct output of the amplifier was merely detected as an impulse and the wave form of the impulse was telemetered back. Counting of the number of impulses and measurement of the pulse height then afforded a count and a crude indication of the magnitude of the momentum of the impinging particle. Later versions became somewhat more elaborate, providing a number of amplitude-discriminating channels and pulse-standardization circuits through which the telemetered signal consisted of a scaling system, whereby certain discrete pulse-amplitudes and pulse-widths were used to indicate the general momentum range of the individual impinging particle. The time of the occurrence of each event was still indicated as a real-time record, necessitating continuous telemetry from the vehicle to the surface of the earth. Equipment of this type has been flown on Aerobee, Aerobee High, Spaerobee, and Nike-Cajun rockets.

A further refinement of this circuit was evolved for use on the Explorer VI satellite. The instrumentation system for this vehicle still utilized a real-time presentation of the data on the ground, but also had limited storage capabilities, in that the received signal indicated, previous to the time an event occurred, the magnitude of the last preceding event. A d-c analog system was used in which three discrete levels of output signal were possible: a median level indicating a small momentum; an upper level indicating a large momentum; and a lower level indicating a repeat of the preceding event whether large or small in momentum value.

The proposed availability of a special vehicle to be provided by the Lockheed Aircraft Company led to a reconsideration of the experimentation philosophy. A conference was held with representatives of Lockheed in California on 19 and 20 March 1959 to discuss the probable dates of availability and electronics system requirements for compatibility with other instrumentation on the same vehicle. Based upon technical information which was made available at the time of this conference, the Research Foundation proposed an instrumentation system compatible with the requirements as they were then understood. This system was discussed in a following conference held at Air Force Cambridge Research Center 15 and 16 April 1959. As a result of this second conference, a proposal was made and supported under Contract AF 19(604)-5715 to provide a suitable design following the outline of the original proposal.

Further historical understanding of the factors leading to the design described herein can be obtained through study of Quarterly Status Report Numbers 1 and 2 on this contract, covering work done throughout the period 1 June 1959 through 30 November 1959. In addition, earlier work conducted under Contract AF 19(604)-1908 can be evaluated through reference to Quarterly Reports Numbers 1 through 14, which cover the entire program under this contract throughout the period May, 1956 through November, 1959. In particular, Reports Numbers 11, 12, and 13 of this series (covering work done during February through November, 1959) will prove helpful in understanding the background of the original satellite equipment design as well as the general magnitude of the data to be expected, as indicated in Appendix A to Quarterly Report Number 12.

The principal difference between earlier instrumentation systems for acoustic detection of meteoritic material and the present system provided under Contract No. AF 19(604)-5715 lies in the provision of certain internal data storage facilities which permit recording events which may occur during times when the signal is not capable of being received at the ground. A similar function was provided by another organization, (Space Technology Laboratories), for inclusion with the meteor detection system provided for the Explorer VI satellite. In this earlier model, data pulses occuring at the impact of each incoming particle were fed into a binary digital storage system, and held until such time as interrogation from a ground transmitting site triggered a telemetry transmitter within the Explorer VI vehicle.

# CHAPTER II

## BASIC SYSTEM REQUIREMENTS AND DESIGN LIMITATIONS

Certain obvious details affecting both electrical and mechanical design criteria were outlined in the general environmental specifications for the flight vehicle and were published as Lockheed's Specification LMSV-6117A dated 24 June 1959.

Other factors which are not readily apparent from the environmental requirements but which played a significant part in determining design limits and objectives are further discussed in this chapter. Many of these design objectives were worked out during the conferences mentioned in Chapter I.

Before continuing further, it would be well to differentiate between the Lockheed micrometeor detection system in its entirety, which has been the subject of discussion to this point in the paper, and that portion of the system which is the actual basis for this paper, namely the computer portion of the system.

The entire system can be roughly subdivided into three integral interconnected sections. These are the sensor assembly, the computer, and the telemetry sections. The sensor assembly is made up of the detecting plate, crystal microphone, and 100 kc amplifier. One can see then that the sensor assembly is functionally the means by which an incident micrometeor is detected, and a useable pulse-type signal having an amplitude proportional to the momentum of the incident particle is made available at the input of the computer. Thus the sensor section corresponds functionally to tape units, card readers, typewriters, and other input equipment usually associated with more conventional computers.

Similarly the telemetry section of the system functionally performs the job that corresponding output equipment such as tape units, card punches, etc., accomplishes in a conventional digital computer system.

Consider then that the computer section of the system encompasses that portion which receives a pulse-type signal from the input equipment (amplifier), interprets this input signal as useable data, operates on the received information logically, stores the result of this operation until asked for by the output equipment, and finally provides the stored information by means of a suitable signal to the output equipment (telemetry).

## Input

As mentioned above, the input to the computer is the output signal from the sensor. This is typically a burst of 100 kc oscillations extending over a period of 2-7 milliseconds and ranging in amplitude from 1-5 volts about a nominal reference level. Both the period and amplitude parameters of the input burst are functions of the impact momentum of the incident particle, i.e., the detected micrometeor. Ideally, the amplitude parameter reflects a linear relationship with respect to the momentum of the detected particle.

## Logic

Logically, it was desired that the computer perform the following operations with regard to each input signal:

1) The computer should recognize and be able to differentiate between an arbitrary "small" impact and a "large" impact. The nomenclature "A" and "B" impact was coined to identify small and large impacts respectively. An "A" impact was defined as any impact which causes an input signal to appear on the input line, such signal having an arbitrarily selected peak amplitude as a limit. A "B" impact was any impact causing an input signal with a peak amplitude exceeding the arbitrary limit chosen for the "A" signal.

2) It was required that the computer have facilities for counting independently the number of "A" and "B" impacts detected during some arbitrarily chosen period of time. As a first approximation, it was determined that the "A" counter's radix should be a function of both the counting period and the anticipated influx rate of "A" impacts. Further, an arbitrary ratio of 1 to 8 for the "B" radix with respect to the "A" radix was deemed to be sufficient.

3) Since the computer would be programmed by the Lockheed vehicle to have both real and non-real time readout of impact data, it was required that the computer provide storage facilities to retain the received impact data during the non-readout period. Programming is accomplished by means of a mechanically driven commutator, each of the commutator segments corresponding to a

given experiment on the Lockheed vehicle.

4) Both the stored and real time data were required to be converted from digital to analog form prior to being transmitted to the output equipment. The analog form of information is required by the telemetry which is basically a voltage sensitive oscillator.

## Output

As mentioned above, in order for the computer output to be compatible with the input requirements of the telemetry, it was necessary that it be in analog form. The term analog is used here to identify a d-c voltage signal which represents specific information by the value of voltage level with respect to some voltage reference. Certain rather specific restrictions were imposed on the allowable output signal by the Lockheed telemetry system.

1) The output signal was restricted to positive values only with respect to a reference voltage common to both the computer and the telemetry. Further, the peak amplitude of the output signal was specified as equal to or less than 5 volts.

2) The telemetry equipment placed a further limitation on the number of discrete analog voltage levels available for the output signal to represent information. Final recorded data was estimated as resolvable to $\pm$ 2% of full scale; the absolute accuracy was estimated as $\pm$ 5% of full scale. Using the full scale range as 5 volts, this would indicate the minimum detectable step would be of the order of $\pm$ 0.1 volts during real time readout. Likewise, the absolute accuracy estimate required

the nominal voltage levels, used to represent information, to be at least 0.25 volts apart. Note that this last criterion was the limiting requirement insofar as the maximum number of voltage steps available to represent data. Similarly, the resolvability fixed the allowable "drift" which could be tolerated from the computer's output circuit.

## Power

The primary power for the computer, together with the sensor section, was specified as the + 28 volt regulated power supply on the Lockheed vehicle. The entire micrometeor detection system was allocated a maximum of 500 milliwatts from this primary source. Additionally, the following requirements were set forth.

1) All voltages required for the entire detecting system would be stepped down from the primary 28 volt source.

2) The primary power ground should be both a-c and d-c isolated electrically from the system's signal ground.

## Environmental

The environmental cirteria cover essentially two areas, namely thermal and mechanical requirements. The thermal environment was established as encompassing a temperature range bounded by -30° Centigrade and + 85° Centigrade.

The mechanical criteria include the following requirements as set forth by LMSD-6177A:

1) When packaged for shipment, the system shall be capable of withstanding drops to a flat concrete surface in both directions along each of the three major mutually perpendicular axes and each of the principal diagonal axes from a height of 42 inches.

2) The assembled system shall be capable of withstanding vibration at a frequency and amplitude of 3000 cps and 20 g., respectively.

3) The assembled system shall be capable of withstanding an approximately half-sine shock wave with a duration of 6 milliseconds and a peak amplitude of 40 g's.

The above environmental requirements essentially cover the anticipated limits during ascent of the vehicle. Once in orbit, these requirements are considerably reduced such that the thermal limits become $-30^{\circ}$ C to $+60^{\circ}$ C, and the mechanical stresses approach zero. However, once in orbit, the equipment shall be subjected to other environmental stresses. The more significant of these are pressure (vacuum), particle bombardments, and energy radiations.

1) The anticipated pressure is $10^{-8}$ to $10^{-9}$ mm Hg.

2) Several penetrations of the vehicle skin by meteorites per year may be expected. Erosion due to micrometeorites should be negligible for 90-day operation. These effects may be considerably increased if meteoritic showers are encountered.

3) Primary cosmic radiation, that is, the incidence of high velocity nuclei of elements, is anticipated. Because of their extreme velocities, these particles are very penetrative,

and leave an ionized trail through the material, but cause little overall damage due to the low particle flux. Because of the high velocities, shielding weight is prohibitive.

4) At high altitudes there is an intense low energy radiation made up of charged particles trapped in the earth's magnetic field. Its altitude at low latitudes has been reported as above 600 miles. There is reason to believe that the radiation will be encountered at lower altitudes in polar regions. At the present time, sufficient information on the energy distribution of this radiation is not available for use in design considerations.

5) Albedo cosmic rays are secondary cosmic radiation caused by primary cosmic radiation striking the earth's atmosphere and disintegrating the target nuclei into smaller, lower energy particles. The portion of the secondary radiation which is radiated to space is referred to as Albedo cosmic rays. These rays will cause molecular sputtering of the skin and damage to components mounted at the surface. Damage produced is unknown but is expected to be slight for 30-day operation.

6) Ultra-violet, extreme ultra-violet, and X-ray radiation produce damage to non-metallic materials by breaking down the chemical bonds. Thin metallic foils can be employed to shield these materials. This radiation might produce a slight static charge on the vehicle due to photo-electric effect. However, this charge is probably counteracted by other effects.

## Package

Weight and volume requirements for the detecting system were left open, subject to final approval by AFCRC and Lockheed prior to finalization of the system design. This leeway allowed the author considerable freedom to investigate and optimize the package design with regard to other desirable package features.

The principle package design objectives were to: facilitate assembly; ease rework and troubleshooting problems; and provide a relatively compact and rigid framework and case for mounting the system components. Obviously, the final package also had to meet the mechanical environmental requirements outlined above.

# CHAPTER III

## STATEMENT OF DESIGN PROBLEM BASED ON SYSTEM REQUIREMENTS

The basic system requirements and design objectives as outlined in Chapter II directly affect the design of the computer section of the system. Since these requirements were imposed by Lockheed (as the prime contractor), AFCRC (as the experimentor), or the author (as responsible engineer), they could neither be waived nor modified, such modifications of the original specifications having already been included in the subject requirements. Several problems in design of the computer resulted directly from or were implied by these requirements.

The majority of problems were a result of the temperature range over which the computer is required to function. Several typical problems included undesired variations in transistor parameters, solder joints, temperature sensitive insulating tape, metallic part expansion and contraction, and variations to the magnetic parameters of the toroidal core used with the power supply transformer. In fact, one might deduce correctly that but for the thermal environmental specification, the design of a reliable computer would have been greatly simplified.

A second major problem area resulted from the low power allotted the system together with the rather stringent isolation requirements placed on the system power supply. Typical germanium junction transistors available for switching circuit applications are designed for optimum common emitter gain at an emitter current in the order of one milliampere. Because of the limited power available it was necessary to operate the transistors with a nominal saturation collector current of 0.3 milliamperes, considerably below their optimum operating point.

The ground isolation requirement coupled with the obvious need to reduce losses between the primary source power and the system supply voltages dictated use of a transformer coupled d-c-d-c converter. Temperature effects on the transformer have already been mentioned as a major problem. In addition, the rather violent switching transients associated with the primary oscillator caused considerable energy radiation and resulted in undesired noise on the computer's power and signal lines. (Actually the major effect of the radiated energy was apparent in the amplifier section. However, this particular aspect of the system design will not be dealt with further on the grounds that this problem area is not an integral part of the computer.)

Of course, the major overall problem involved the outright and complete design of the computer itself. Since the voltages used, together with their regulation, play such an integral part in not only the individual circuit designs, but also in the logical philosophy of the computer, the power supply has been included as a unique section of the computer. Thus the overall problem might be stated as the design of a logical system taking into account all, or as many as possible, specific problems outlined or implied above, and meeting all the general requirements and criteria specified in Chapter II.

# CHAPTER IV

## LOGICAL APPROACH TO THE COMPUTER DESIGN

In the design of any system, whether it be electrical, mechanical, structural, or other functional type, one should first set down in some systematic sequence the following bases for design:

1) Design objectives - this tabulation should answer as completely as possible the question, "what function (s) must the system perform?".

2) Intersystem compatibility - in the probable event that the subject system being designed is in reality a subsystem of some larger, more complex network of other systems, then the interrelationship between the subject system and adjacent systems must be examined, and considerations which must be taken, listed to insure compatibility between interconnected systems.

3) Design limitations - lastly, all system requirements, specifications, limitations, in short, all known factors which could conceivable influence or restrict the systems design should be studied thoroughly by the designer prior to proceeding forward to the next step in the design.

The author has attempted to fulfill the major steps listed above throughout the first three chapters. At this point the reader should have gained an insight into the peculiar problems which confronted the author in the design of the subject computer.

## Selection of Basic Building Blocks

The next step in the design of any system is an "art" rather than a "science". The scientific aspect of the design will come later. The artistic step referred to is that of establishing a system philosophy. In general, any system organization will reflect the designer's personality and/or background and experience. Where one designer might use a so-called brute force approach, a second designer might use finesse. The end result in either case would be an acceptable design provided the system requirements were satisfied.

Reviewing the computer requirements as outlined in Chapter II, it was decided to subdivide the computer into the following functional sections:

1) Input pulse amplitude differentiator and shaper
2) Adder
3) Storage register
4) Digital to analog converter
5) Output buffer
6) Power supply.

Based on previous experience, it was felt that sections (2) and (3) above could be combined into a single function block where the two described operations are performed simultaneously. Similarly, sections (4) and (5) above were combined into a single design problem. Hence the logical organization of the computer contains the following named sections:

1) Power Supply - takes the primary power from the + 28 volt regulated supply on the vehicle and supplies the well regulated multiplicity of voltages required by the individual computer circuits.

2) Discriminator - accepts an input signal, differentiates between signals by signal amplitude to determine whether the input signal represented an "A" or a "B" impact, shapes the signal into a pulse shape suitable for use with digital pulse techniques, and transmits the resultant pulse on the "A" or "B" output lines from the discriminator.

3) A and B Registers - accept any pulse appearing on their respective input line from the discriminator, count the number of pulses received independently, remember or store the total pulses received, and provide suitable outputs representative of the sum stored.

4) A-B Mixer - provides suitable means to receive the information readout from the A and B registers, identify and transform this assumed data from digital to an analog type signal, and with suitable powering provide a useable signal to the telemetry or output equipment.

The above outline represents the approach toward organization of the computer employed by the author. This is not to say that the above represents the only solution to the problem (s). Nor is there any known mathematical equation which will verify that this approach is even the optimum one. At most, one might comment that the above approach does represent a very simple straightforward solution to the problem.

Notice that in the above scheme an essentially analog signal is converted to digital, and then reconverted back to analog prior to readout. An alternate scheme might have been to leave the input signal in analog form throughout the required sequence of operations. This approach was purposely avoided because of the complexity of operations and equipment usually required for storing analog data. A rather obvious scheme for handling analog storage would be through the use of a core with a reasonably linear B-H characteristic over some range. Hence by powering the input signal the volt-second energy under the input signal wave shape could be made to drive the core to some point on the B-H curve where B, the magnet flux in gauss, would be proportional to the lines of flux switched and hence representative of the area under the input wave shape. The drawbacks to this scheme are obvious with regard to the requirements and limitations already set forth. Principal among these are the relatively large amount of powering required for the core driver circuits, the need for timing, gating and other rather complex control circuits during any readout cycle, not to mention the radiated electrical noise induced into the computer system during switching of the core. Note that the amplifier, used in conjunction with the computer and packaged in close proximity to the computer, will saturate on an input of 20 microvolts from the microphone.

For much the same reasons as above, a switch core array using a row and column technique to correspond respectively to input signal amplitude and number of impacts received was also considered and discarded as being an impractical approach.

Capacitor storage was also considered and discarded since timing and gating circuits would be required. Additionally, the as yet undetermined period of storage coupled with the unknown effect of anticipated static charge collecting on the vehicle made the need for a recharging cycle a very definite possibility. The author did not feel the end result of such an approach warranted the obvious design effort entailed.

An obvious approach in the register section, and the one which was used in the final design, involves the use of a bistable or two-state circuit. Any number of these two-state circuits can be arranged by one of many commonly used techniques into a stepping or counting ring. Further, by using this approach, each circuit or building block not only functions as a counter, that is, flipping or changing from one stable state to another with the incidence of an input signal, but each block also remembers it received an input signal inasmuch as each of the two states are stable. Three such building blocks readily come to mind:

1) Latching relays
2) Cryotron circuit
3) Eccles-Jordan trigger.

Again from a practical standpoint, only the Eccles-Jordan circuit provides a realistic answer. Relays are inherently bulky, power consuming, and unreliable beyond several thousand operations. Although a very definite possibility for the future, cryotron circuits are not yet beyond the development state of the art. Also, for this application, the coolant required for the necessary near absolute zero thermal parameter would impose insurmountable problems with regard to the restrictions already imposed on the system.

The final basic consideration which had a significant bearing on the computer design philosophy regarded the selection of the basic amplifier device. After reviewing the foregoing discussion, the choice lay between vacuum tubes and transistors. The vacuum tube is inherently a high impedance device which is in itself a desirable characteristic insofar as powering is concerned. However, the heater power required more than offsets this savings in signal power. Additionally, the vacuum tube has an inherently short life, its operating theory being based on self-destruction, and is relatively bulky and fragile when compared to a typical junction transistor. The transistor, on the other hand, is theoretically ideal for switching circuit applications, closely approaching an electrically controlled mechanical switch in its characteristics. Small, rugged, requiring no heater power, and having a theoretically infinite life expectancy, the transistor is particularly suited to an application such as the subject computer system.

Today there are available two similar types of transistors which inherently differ in certain of their characteristic parameters. These two types are commonly differentiated by the type of semiconductor material used in their construction, namely, germanium and silicon. The principal differences include price, base to emitter gain, thermal characteristics, bias levels, internal resistivities, and frequency response. In all cases mentioned above except thermal characteristics, germanium units still enjoy an advantage over silicon devices in low voltage switching applications although the gap is slowly narrowing. In the case of thermal considerations, germanium semiconductor devices are typically rated to operate between junction temperature limits of $-55^\circ$ C to $+85^\circ$ C. Notice that this allows a safety factor of $20^\circ$ C at each extreme of the thermal range over which the computer must operate. It is granted that by using germanium units at the exclusion of silicon devices the author has exposed himself to a considerable variation in transistor parameters throughout the thermal environment range. However, although imposing limits on the individual circuit designs, none of the associated problems proved impossible.

## Computer Design by Section

Again, experience and common sense are the principal tools which can be realistically employed during this phase of system design. This is not to say the designer can let his imagination have full sway. Certain physical and scientific laws must always be respected. Nevertheless, considerable freedom in design is still allowed even at this late stage in the problem.

Sectional design of this system was started at the A-B mixer section. As described in Chapter II, the Lockheed telemetry system requires that data be represented by zero to +5 volts, and resolvable to $\pm$ 2% of full scale. The accuracy estimated by Lockheed is $\pm$ 5% of full scale. Based on this, it was ascertained that eight discrete levels would be available to represent data reliably for transmission to monitoring stations. A survey of existing components indicated that the choice of 0.6 volts per output step would be a reasonable choice. This value was selected as a basis for design of the A-B mixer. Two different types of A-B mixers were tentatively examined in the original approach. These were designated as "voltage mixer" and the "current mixer".

The "voltage mixer" uses resistor voltage divider networks working into a high impedance load. A schematic of a typical "voltage mixer" as bench tested during the early development of the prototype system, is shown in Figure 4-1. This scheme offered several advantages over the "current mixer" approach. However, it also had a very basic disadvantage which far outweighed the advantages. The primary advantage of the "voltage mixer" circuit was that only a single level of logic was required throughout the logic section of the system. Thus, triggers operating between $\pm$ 6.8 volts and ground could work into this circuit and the digital information could be converted to analog form falling within the required 0 to +5 volt range. The major disadvantage observed was very significant voltage variation with respect to thermal environmental change for the common collector transistor circuit used in the "voltage mixer". Bench testing of this "voltage mixer" indicated an output voltage variation at each level of typically 1.2 volts throughout the thermal environmental range of $-30^{\circ}$ C to $+60^{\circ}$ C. Since during the early stages of development of the system it had not been definitely specified whether the temperature data of the system-mounting location on the Lockheed vehicle would be available, it was felt that this output variation, with respect to temperature, could not be tolerated in the system. For this reason, the "current mixer" was chosen over the "voltage mixer".

The "current mixer" does not exhibit any significant output voltage variation with respect to temperature change. However, since it utilizes grounded base transistor stages within the circuit to develop discrete amounts of current (and hence voltages) across the bias resistor ($R_6$) shown in Figure 4-2, the use of a second signal-level within the logic section of the system is necessitated. If appropriate values of emitter resistors $R_1$, $R_2$, and $R_3$ in the "current mixer" are chosen, then

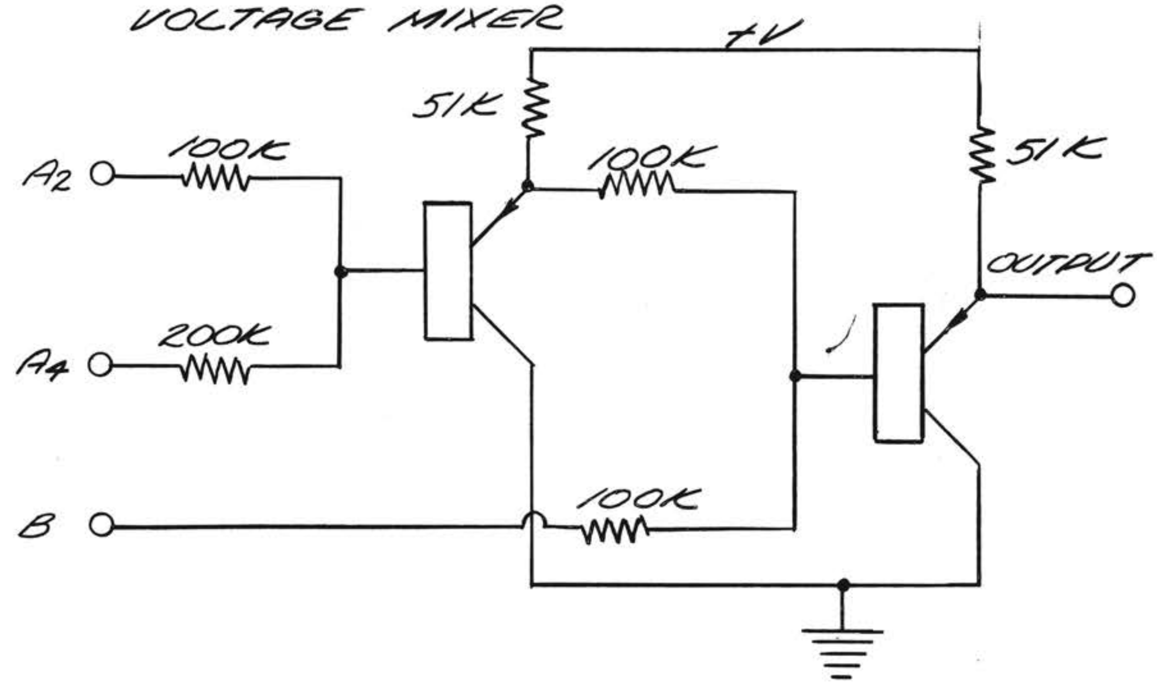
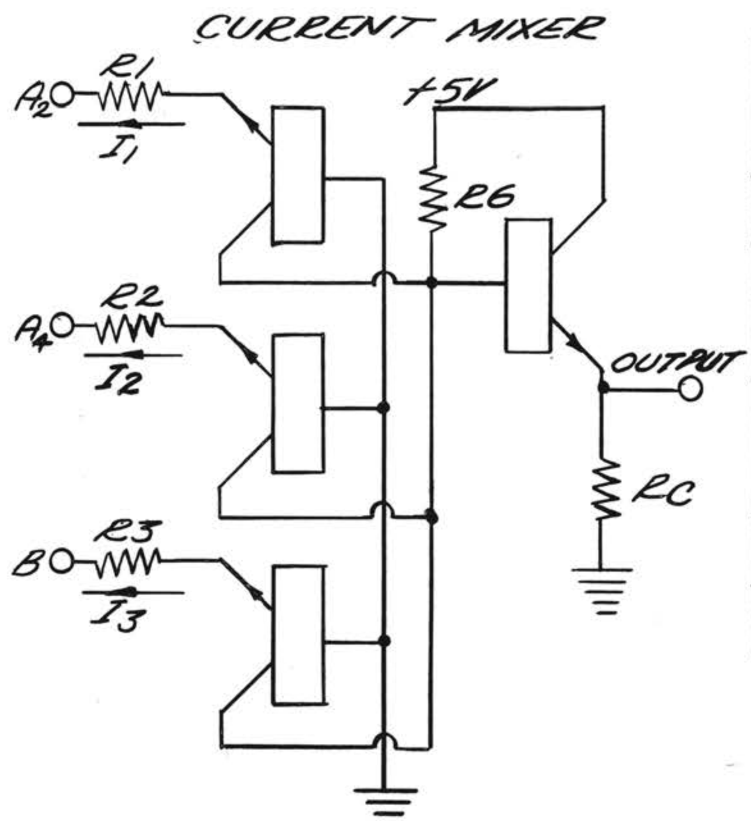



FIGURE 4-1




| $I_1$ | $I_2$ | $I_3$ | $E_o$ |
|---|---|---|---|
| 0 | 0 | 0 | 4.8V |
| 1 | 0 | 0 | 4.1 |
| 0 | 1 | 0 | 3.4 |
| 1 | 1 | 0 | 2.7 |
| 0 | 0 | 1 | 2.0 |
| 1 | 0 | 1 | 1.3 |
| 0 | 1 | 1 | 0.7 |
| 1 | 1 | 1 | 0.1 |

FIGURE 4-2

with the outputs from the A and B registers at one or the other of two discrete levels ( say 0 and -5 volts), an exact value of emitter current can be caused to flow through the transistors for the one level (-5 volts) and essentially no current for the other level (0 volts). Since the attenuation of this current through the transistor is of the order of only one to ten parts per thousand, essentially all emitter current through each of the transistors is caused to flow into the collector circuit. Consider each of the currents $I_1$, $I_2$ and $I_3$ as a unique binary position. This is illustrated by the table in Figure 4-2. By choosing an appropriate value of bias resistance returned to + 5 volts as shown in Figure 4-2, the voltage appearing at the base of the output emitter-follower stage transistor will have one of eight discrete values, depending upon the unique binary combinations of the input currents.

From the discussion above, it is apparent the values of the emitter resistors must be chosen such that the currents in each of the respective input legs are proportional to 1, 2, and 4, reflecting the value of each of the binary positions. The voltage parameters of this circuit are such that with all three of the common base stages conducting, none of the three are saturated. The reason for using the common base stage is to provide a suitable impedance transfer between the input resistors and the bias resistor such that the change of voltage across the bias resistor does not reflect back across the emitter resistors. Therefore, regardless of the voltage drop across the bias resistor, the voltage across each of the emitter resistors (and hence $I_1$, $I_2$, and $I_3$) is dependent only on the input signal levels.

Several other significant features of the A-B mixer should be pointed out. Assuming input levels of -5 and zero volts, observe that the -5 volt level must have a low impedance looking back into the source. This is due to the fact that the common base stage is conducting its maximum current at this time, and any impedance looking back into the driving circuit will cause a reduction in the five volts across the emitter resistors. These emitter resistors are carefully bridged and chosen to an exact value so as to minimize possible error and insure a smooth progression of analog output voltage steps.

Observe from Figure 4-2 that only three inputs to the A-B mixer, labeled $A_2$, $A_4$, and B, are shown. As mentioned above, 0.6 volt per output step was chosen as the standard analog voltage increment. This limited the number of discrete voltage levels available to represent the contents of the A and B registers to eight.

The implications of this are that the contents of only three of the register triggers may be readout. The bases used to choose the triggers indicated for readout are presented in a subsequent section of this paper.

The 0.6 volt step was predicated on the basis of anticipated drift of component parameters over the full range of temperature anticipated. The base to emitter forward bias voltage will increase with decreasing temperature at an essentially constant rate of 2 millivolts per degree Centigrade. Thus, over the complete range of $-30^{\circ}$ C to $+60^{\circ}$ C, the $V_{BE}$ of a typical germanium transistor will vary by approximately 0.2 volt with respect to the $V_{BE}$ at one of the limiting temperature end points. Consider the effect this will have on a particular analog output level, say with all three of the common base mixer transistors conducting. Assuming all other parameters constant, a 0.2 volt shift across each of the emitter resistors causes a 0.1 volt shift across the common base bias resistor across which is being developed the final output voltage drive. In addition, if all variances in other parameters are considered, the net effect (which has been typically observed) is a shift of 0.2 volts in the output signal. Adding this possible drift to the 0.25 volt requirement imposed by the telemetry, note that the 0.6 volt step chosen allows a 0.15 volt margin for safety, or approximately 25% per step. The foregoing discussion was the primary basis for choosing the eight 0.6 volt output steps rather than some other combination of steps and incremental voltages.

The input impedance of the telemetry is approximately 250 kilohms. Therefore, from an output drive consideration there was little or no problem. The common collector configuration was used as the final stage for several reasons. There is no signal inversion, voltage gain is essentially unity, high current gain is possible with a relatively high input and low output impedance, and very little possibility of saturation exists. The common collector configuration is conventionally referred to as a line driver because of its capacitive drive capabilities. This dovetails rather well with the as yet undertermined length of cable interconnecting the computer output and telemetry input.

Once the input requirements of the A-B mixer were fixed, the A and B registers could be considered. In a previous discussion above, the use of a basic Eccles-Jordan circuit, henceforth referred to as a trigger, as the basic building block, was justified. Since the trigger is a bistable circuit, that is, with two stable states, each trigger can count to two at most. This implies that any counting system used must have a basic radix of two. However, recalling that only eight output steps are available for the readout of information, notice that this corresponds to the three inputs envisioned for the A-B mixer. Consider that at least one input line must be assigned for the readout of B inpact data, leaving only two lines available for the digital readout of A impact data. Thus, if straight binary readout were used in the A register, only four A impacts could be counted and stored prior to either reset, overflow, or saturation of the register depending on the register's design.

Previous data available from past micrometeor experiments indicates that an A register having a capacity of at least eight is essential for satisfactory functioning of the computer as regards the experiment with which it will be used. Therefore, in order to obtain the desired register capacity of eight, three triggers with binary inputs are serially connected. However, only the last two (the 2 and 4 position triggers) are connected for readout to the A-B mixer. Hence the counting and storing of A impact data is by straight binary, but the readout is by biquaternary. A single trigger with a binary input and its output connected to a corresponding input on the A-B mixer comprises the B register.

Consider now the full significance of the design just discussed. The A register has a counting and storing capacity of eight A impacts as required. The B register has a capacity of two which exceeds the 1 to 8, B to A, ratio requirement. At first glance it appears the biquaternary readout of the A register introduces a considerable error inasmuch as the register can always contain one more than appears on the readout lines. For instance, three A impacts will be readout as two, an error of 33%. However, consider also that as now designed, the A register is never reset, except when the register is filled after seven impacts, the eighth A impact automatically recycling the ring to zero. Nor is there any need to reset the register when only partially filled. Therefore, the "odd" one is never lost, nor does the probable error ever become greater than plus one. Consequently after several hundred A impacts, the relative error becomes insignificant.

From the foregoing discussion one might correctly deduce that final data reduction of the telemetered information is required. A more detailed discussion of this procedure is included at the end of this chapter.

It was desired that the discriminator perform two basic functions, that is, differentiation between input signals by amplitude (whence came the name discriminator), and pulse shaping. In addition, due to the nature of the input signal described in Chapter II, some means was required to insure only a single digital pulse appeared on the appropriate A or B output line. After considerable study the following circuit arrangement shown in Figure 4-3 was evolved.

The discriminating function is accomplished by a standard threshold gate using a capacitor, diode and two resistors. This circuit differs from a conventional gate in that the gating bias is provided by a d-c bias voltage (set by the two resistors $R_1$ and $R_2$), rather than a coincident pulse. Thus any time a signal appears on the input line with an amplitude sufficiently large to exceed the threshold bias, the difference between the input signal and the bias is superimposed on the d-c bias at the diode anode. The capacitor $C_1$ differentiates the negative portion of the input so that a sharp trigger pulse appears at the collector of $T_1$. Observe that the diode serves as a half-wave rectifier blocking the positive half of the input signal burst. $R_1$ and $R_2$ are of the order of 400-500 kilohms to minimize the loading of the input driver.

DISCRIMINATOR

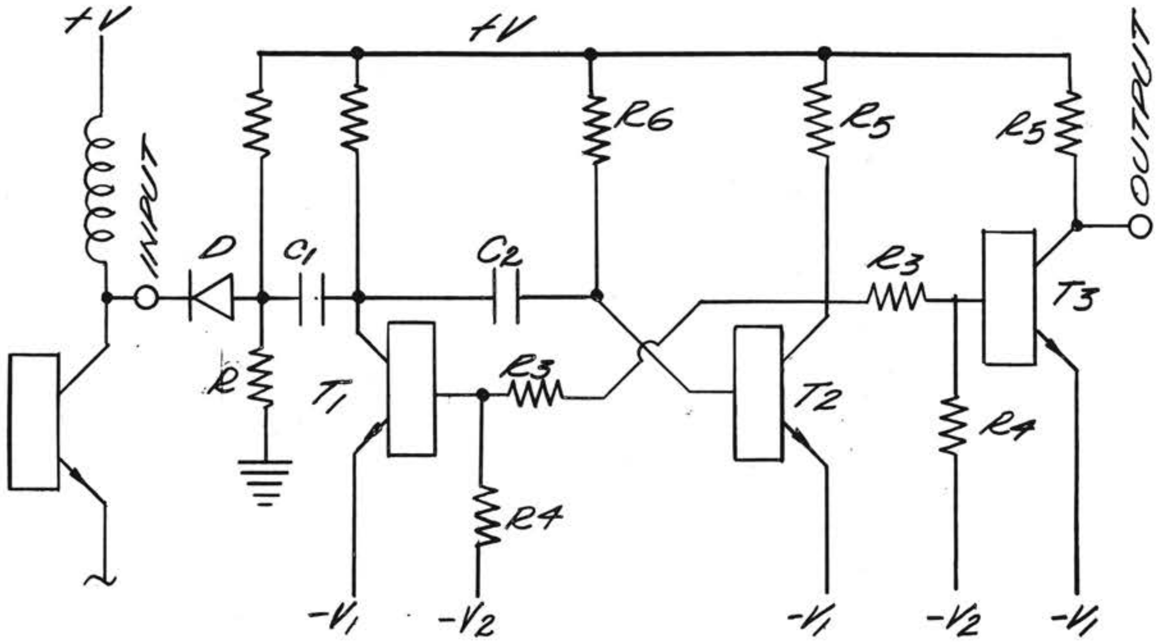



FIGURE 4-3

TYPICAL INPUT SIGNAL

VERTICAL - 1 VOLT PER DIVISION

HORIZONTAL - 1 MILLI-SECOND PER DIVISION

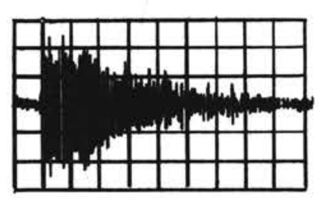


FIGURE 4-4

A similar discriminator circuit with the diode and $C_1$ interchanged was considered. This would pass a positive rather than a negative signal. However, this circuit was not suitable for this application because of the 100 kc characteristics of the input signal. Observe that an input with a peak amplitude less than the bias voltage would not allow sufficient time between input signal peaks for the capacitor to complete discharge except by making $R_1$ and $R_2$ relatively small, and hence loading down the input circuit. The result is that the charge build-up across the capacitor will eventually overide the diode bias. A typical input signal waveshape is depicted in Figure 4-4.

The threshold differentiating gate is used to trigger a conventional monostable multivibrator or single shot. This is a common practice generally used in pulse shaping circuits to blank out any undesirable transients on the signal line. A negative shift at the collector of normally "off" $T_1$ is coupled into the base of normally "on" $T_2$ by $C_2$, which causes $T_2$ to turn off. Transistor $T_2$ turning off allows the potential at its collector to go positive which turns $T_1$ on. At this time the collector of $T_1$ is effectively shorted to $-V_1$ so that after the triggering transient has initiated "flipping" of the single shot, the input is shorted out and the timing components $R_6$ and $C_2$ assume control. Thus the amplitude of the resulting square wave pulse will be determined by the voltage returns of the circuit, while the pulse length is fixed by the values of $C_2$ and $R_6$. Note that the single shot pulse length should completely bound the 100 kc burst of input signal to prevent multiple triggering on a single impact.

The final stage of the discriminator is a simple common emitter configuration necessary to provide powering on the A and B pulse lines, and insure a sharp leading edge on the output pulse. The output pulse might have been taken from the collector of either $T_1$ or $T_2$. However, for reasons to be discussed in Chapter V, the extra stage was added and the circuit configuration was finalized as shown.

Observe that by means of the circuit just described, input pulses with amplitudes less than some arbitrary value of voltage can be rejected. Thus it is ideally suited to B impact discrimination with respect to A impacts. However, since the bias on the A impact discriminator is set only high enough to reject the nominally 0.2 to 0.4 volts of noise typically appearing on the input line, all B impacts cause a pulse to appear on the A register input line also. It would have been possible to design an additional feature into the discriminator whereby the A pulse line would be shunted whenever the B single shot flipped. Had this been done, the total number of impacts over a given period of time would have been the sum total counted by both the A and B registers. With the present design, the total number of impacts regardless of magnitude is readout by the A register, the B impacts by the B register, and the total number of A impacts is the difference between the A and B registers. In each case, one arithmetic operation is required to deduce the identical information and hence there is no logical advantage in one scheme over the other. However, from a circuit design standpoint, the scheme chosen is much the simpler.

The power supply section was not a logical design problem, but rather required the more specific considerations generally taken in a basic circuit design. Therefore, this section will be covered in more detail in Chapter V. In brief, it might be mentioned that a basic transistor-saturable reactor using a magnetically coupled astable multivibrator circuit to excite the transformer primary was chosen as the fundamental power supply circuit. Basis for this selection was the power input limitation implying a high voltage-transformation efficiency requirement, the isolation of signal and power ground requirement, and a size and reliability consideration.

Figure 4-5 is a block diagram of the computer which has just been evolved. Before summarizing this chapter, there are a few loose ends which should be tied up. Referring to Figure 4-5, observe that data is read into and out of the computer on a single line, but that at one point, namely between the A and B registers and the A-B mixer, data flow is parallel on three information lines. Thus, one might note that both serial and parallel logic are uniquely combined in a successful effort to optimize the design from the standpoint of component count, timing and control circuit reduction, coupled with a reliably simple mode of data readout. Furthermore, it should be obvious to the experienced engineer that with a few timing and control circuits, this computer system would serve as the basic building block of a much more complex and powerful computing system.

Several automatic self-checking features are an inherent outgrowth of the logical design of the computer, and should be mentioned. The reader might question the practicality of any need for checking against any component or circuit failure on the grounds that with the computer orbiting in deep space, nothing can be done to remedy a system malfunction. Actually, the inaccessibility of the system while operating is one of the principal basis for requiring checking features. For example, in the event of failure to the B register trigger, this can be recognized and allowed for in the data reduction of the readout information, and the

ELECTRONIC COMPUTOR FOR THE LOCKHEED MICROMETEOR SYSTEM

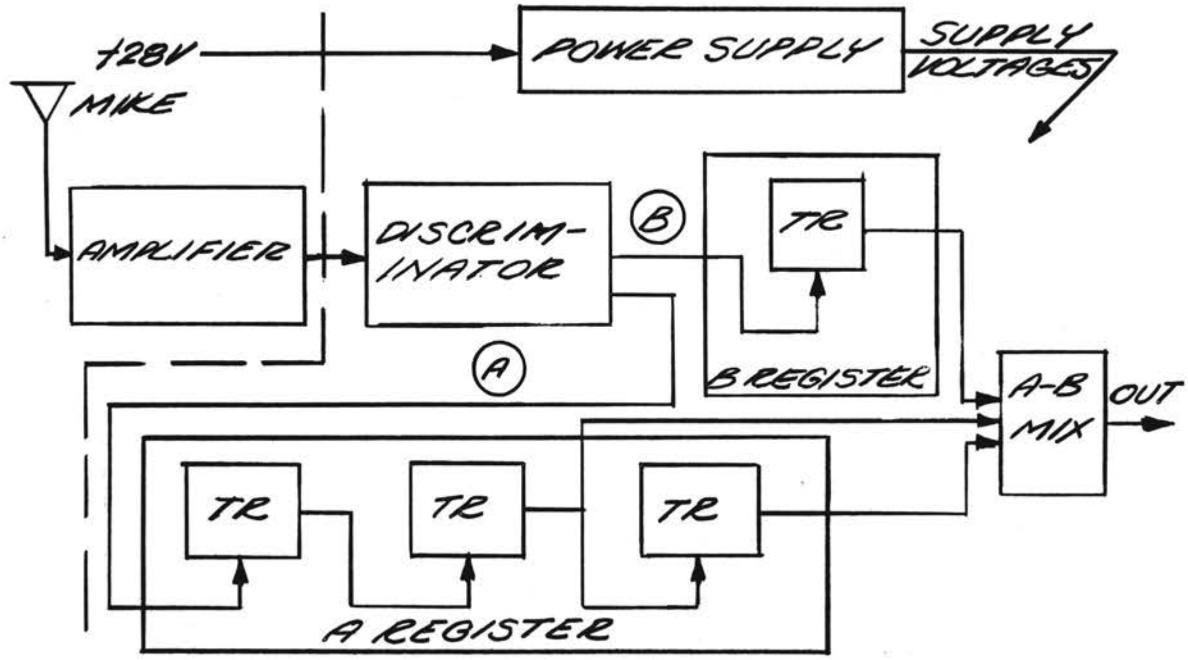



FIGURE 4-5

TYPICAL READOUT RECORD

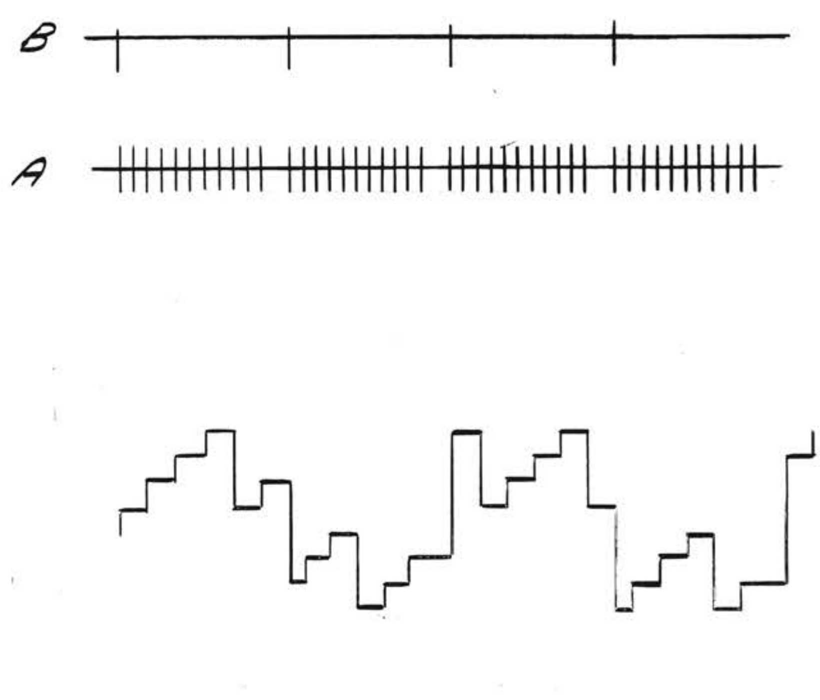



FIGURE 4-6

total impacts still be reported with complete assurance of validity.

All checking is accomplished by observing the real time readout signal. Since all counting is sequential, over a given period of time a definite pattern is established which the experienced observer can distinguish. If all component parts of the computer are functioning properly, then over a sufficiently long period of time there is a 100% probability that eight discrete voltage levels in the output signal will be observed. Referring to the table in Figure 4-2, if any of the triggers which provide the readout fail, a comparison of output signal levels against the chart will quickly indicate the malfunctioning block. For instance, consider the $A_4$ trigger latching up in the "on" position. Then over a sufficiently long period, the output signal would contain only the 0.1, 0.7, 2.7, and 3.4 volt levels.

Normally, for a B impact, the output signal will, for example, change from 4.8 volts to either 2.0 volts or 1.3 volts depending on whether or not the A register contains the "odd" one mentioned previously. Thus, in the event an output signal was observed changing from 4.8 to 3.4, multiple triggering would be indicated and all data would become suspect. However, in this case the suspected data should not be discarded, since subsequent observations might indicate the particular malfunction. For example, a change in output signal from 4.8 volts to 0.7 volts would pin down the malfunction as being in the A register, tending to verify the B impact data. Similarly a jump from 4.8 to 2.0, and then immediately back to 4.8 volts would definitely restrict the multitriggering malfunction to the B register.

Other checks include a periodic stepping of the output signal indicative of an unstable, oscillating component or block, and repeated output signal steps corresponding to simultaneous A and B pulses indicating malfunction of the B impact discriminating circuit.

A typical readout record is depicted in Figure 4-6. Also indicated on a common time scale are the simulated A and B impact pulses. Once a standard procedure has been established for data reduction, the process becomes reasonably simple and straight forward. Consider the case of real time readout such as shown in Figure 4-6. The following steps are required.

1) With an ordinary straight edge, draw a horizontal "mean axis" midway between the fourth and fifth voltage levels.

2) Count all legitimate ($\geqq 0.6 \pm 0.2$ volt changes) output signal steps including those that pass through the "mean axis," and multiply by two. The product will give the total number of impacts during the period of time represented by the record with a maximum possible error of one.

3) Count all signal steps that cross the "mean axis". This represents the total number of B impacts.

4) Subtract number of B impacts recorded from the total number of impacts recorded. This difference is the total number of A impacts.

5) Make any of the reliability checks described above necessary to verify validity of recorded data.

The procedure for reducing stored time readout is somewhat similar to the five steps outlined above except no significant reliability checks can be made, and a table corresponding to the one in Figure 4-2 becomes necessary. In this case, only two voltage levels of the output signal are recorded, namely the one corresponding to the last previous interrogation at time $T_1$, and the signal just received, say at time $T_2$. Then the period of time corresponding to the number of impacts stored is $T_2$ minus $T_1$.

1) A zero voltage reference should be established and the ground recording equipment accurately calibrated to record the output voltage signal with respect to the voltage reference. Actually any means of calibration is satisfactory as long as an index is established on the record corresponding to each of the eight output signal levels.

2) From a table similar to that shown in Figure 4-2, determine the number of A and B impacts contained in storage at times $T_1$ and $T_2$ respectively. From this deduce the number of A and B impact pulses stored during the storage period.

3) Observe subsequent readout signal to correlate stored inpact flux rate with actual flux rate. This should indicate whether either register recycled during the stored time period with a reasonable degree of accuracy. Tentative programming presently calls for a store time to real time ratio of 97/3. Present micrometeor influx data to date indicates that, except in the case of a so-called meteoric shower, there is little likelihood of either register recycling during the store time period.

In summarizing Chapter IV, the author wishes to point out that the logical design and organization of the complete computer has been discussed in detail. An attempt has been made to justify in the reader's mind every phase of the design, and to present a firm basis for each major design decision. A discussion of the computer's sequential operation has been included along with significant diagrams and sketches.

# CHAPTER V

## DESIGN OF THE BASIC CIRCUIT

Again, correlating the thesis subject to a generalized system design effort, one should consider that there are at least two generally accepted and legitimate approaches in the design of any system. One requires a standard array of functionally individual building blocks to be designed by the engineer, analyzed and described or represented by some mathematical equation (generally LaPlace for a servo-type system, Boolean or switching algebra for a logical or decision making system, etc.), and these subsystem blocks made available to the system design engineer. In the second approach, the system designer works out a specific system design using interconnected black boxes or building blocks, if you will, which the circuit designer subsequently must fill with suitable components and devices that will functionally satisfy the system requirements. Common characteristics of both approaches are that the so-called building blocks must be compatible one to the other (so that when assembled there will not be any abrupt functional discontinuities apparent in the system), and the fact that normally the production engineer is not consulted until too late to make any changes in design which might facilitate manufacture.

Fortunately in this specific case the author was responsible for all three phases mentioned above in the evolution of this particular system. Thus it was not necessary to adhere to either philosophy of system design since all problems of coordination and communication were essentially eliminated. A typical result of this arrangement was, for example, that the registers were specifically designed to use an existing trigger while, on the other hand, the A-B mixer and power supply were singularly designed to complement the system requirements. The fastidious reader who views, for example, the one-to-one correspondence between the three binary inputs to the mixer and the magic figure of eight sufficient and necessary readout voltage steps, with suspicion should bear this in mind.

The foregoing point was brought out since if the reader assumes the subject system developed sequentially as described in this thesis, then certain aspects of the system's design (to be covered in this and the following chapter) which are facilitated by design decisions discussed in previous chapters, might appear to be a fortunate coincidence. The author wishes to assure the reader that such is not the case, and that with but few exceptions many of the apparently small details which "just happened to" dovetail were the result of considerable study and speculation.

## Standard Components

The circuits design aspect of the problem was greatly simplified by certain characteristics of the system. Several of these already described in Chapter IV were the basis for choosing a germanium transistor as the basic component. Further the author reasoned that by restricting the types of transistors used to one standard type plus, at most, two special types for specific applications, the tasks of procuring, stocking and testing would be greatly simplified. The overall system was examined with regard to the use of a common type transistor throughout, and the implied effect this would have on logical signal levels and power supply requirements. This investigation indicated that either a PNP or an NPN type transistor could be used for the standard. Thyratron type transistors for possible use in the trigger circuits were purposely disregarded on the grounds that their "latch-up" current requirement was prohibitively large.

Based on past experience, it was anticipated that a critical problem in the design of all circuits would be transistor characteristic drift due to changes in the environmental temperature. Further, transistor theory predicted that, of those transistor parameters of concern to the switching circuit designer, $I_{CBO}$ and $h_{FE}$ (the collector cutoff current and common emitter d-c current gain respectively) would be the most adversely affected by themperature changes. $I_{CO}$, the thermal leakage current, is considered a component of $I_{CBO}$ above. As a general rule of thumb, it was predicted that for the junction transistors considered, the $I_{CBO}$ specified at $25^{\circ}$ C would double for each $10^{\circ}$ C increase in ambient temperature, and the $h_{FE}$ specified at $25^{\circ}$ C would halve at $-30^{\circ}$ C. The drifts of other transistor parameters were then taken into consideration in each circuit design, but were not felt to be significant enough to be used as a basis for selection of the standard transistor type.

Arbitrary limits of $I_{CBO}$ at $+ 60^{\circ}$ C and $h_{FE}$ at $-30^{\circ}$ C were set at less than 80 microamperes and greater than 20 respectively. Using the rules of thumb mentioned above, these limits established acceptable values for these parameters at $+25^{\circ}$ C of $I_{CBO}$ less than 7 microamperes with $V_{CB}$ equal to 15 volts, and $h_{FE}$ greater than 40 with $I_C$ equal to or less than 500 microamperes. The $V_{CB}$ and $I_C$ conditions specified above are additional parameter conditions arbitrarily chosen by the author based on his "feel" for the problem. Subsequent design calculations will verify these "ballpark" figures.

Other considerations taken in the selection of a standard transistor were the package size, power rating, cost, availability, number of equivalent types, and reliability based on the manufacturer's reputation. Since power requirements are negligible, and power dissipation primarily a function of the package, any transistor in the less than 150 mw class in general satisfied these requirements. Similarly, any units which met the $I_{CBO}$ and $h_{FE}$ specifications were satisfactory in all other respects. Based on a survey of transistor types available from local jobbers, a 2N78 was chosen as the standard unit with a 2N254 specified as the backup unit. Both are NPN grown junction germanium transistors; the 2N78 manufactured by General Electric, the 2N254 manufactured by Texas Instruments.

Based on much the same reasoning which substantiated the decision to base essentially all circuit designs around a standard transistor, it was also decided to select a standard type diode, resistor and capacitor for the basic switching circuit designs. The few special components used will be mentioned and their use justified where applicable.

Both silicon and germanium point contact and junction diodes were considered as the standard diode. Silicon and point contact germanium units typically exhibit a "forward" drop of 0.7-1.0 volts with a forward current equal to 1.0 milliamperes. Junction germanium diodes, on the other hand, are available which exhibit a $V_F$ of 0.25 volts at $I_F$ equal to 1.0 milliampere. Even though the forward drop might not appear a significant parameter, consider for example the full bridge rectifier circuit which would be the logical choice to use with the power supply. Assume silicon diodes which typically exhibit a $V_F$ of 1.0 volts at $I_F$ equal to 10.0 milliamperes were used and a load current of 10.0 milliamperes, d-c. The power lost in rectification would then be approximately 20 milliwatts. This represents 4% of the total 500 milliwatts allotted the system from the primary source.

A second consideration was the nearly 0.5 volt difference in input signal necessary to trigger the discriminator single shot. Thus many impacts would be rejected by a low conductance diode that would otherwise be registered by the high conductance unit.

A IN305 diode was selected as a standard component. This unit is a gold-bonded junction type diode manufactured by Raytheon.

The choosing of a standard resistor type was greatly simplified by a size consideration and the availability requirement. Ohmite 1/10 watt Little Devil Resistors are available in all ± 10% tolerance RETMA values and have a good history of reliability.

Many capacitor types were considered before choosing a standard. The final selection came from a physical size versus capacity relation. The type selected was the $V_K$ series manufactured by Vitramon, Inc. Capacitors in this series are available from 47 mmf to 10,000 mmf with a ± 20% tolerance. Although there are six different package sizes in the series, lead spacing is standard for all packages. The maximum package size in the series is 0.300 by 0.300 by 0.100 inches. Capacitance change due to temperature is specified by Vitramon as less that ± 7.0% the 25° C value, over a temperature range of -30° C to +60° C.

Figure 5-1 depicts the packages of the four standard components. Observe that with the exception of the resistor, all components shown are

STANDARD COMPONENT PACKAGES

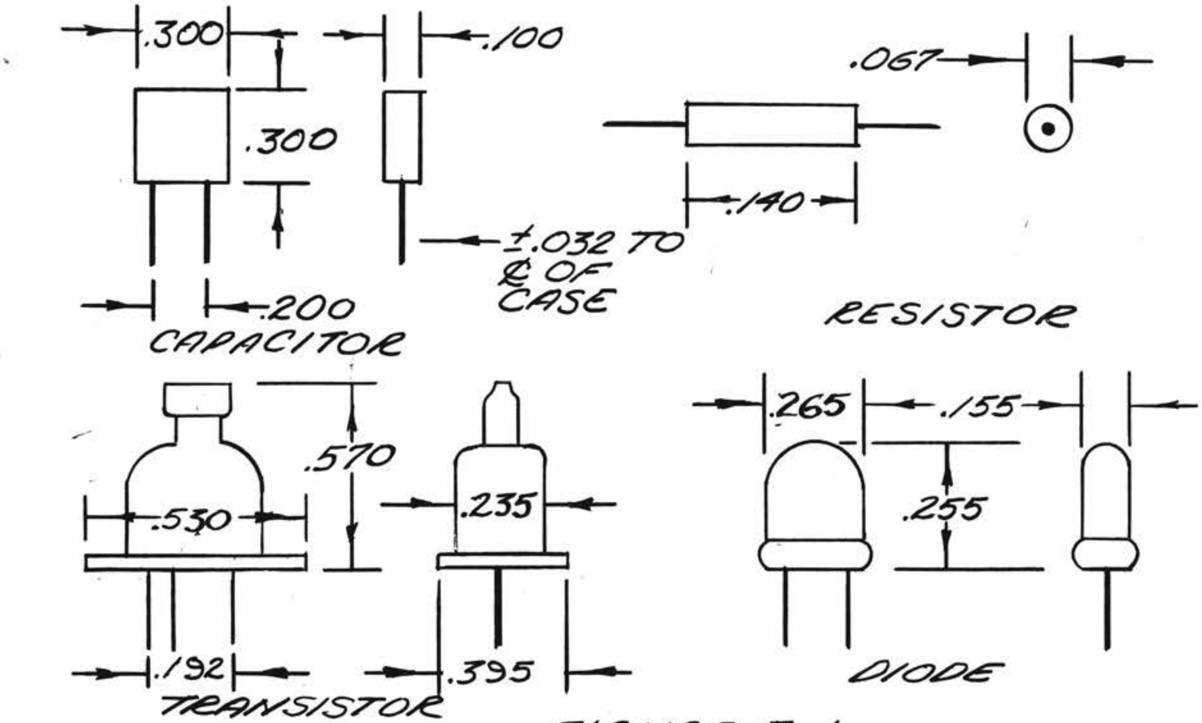
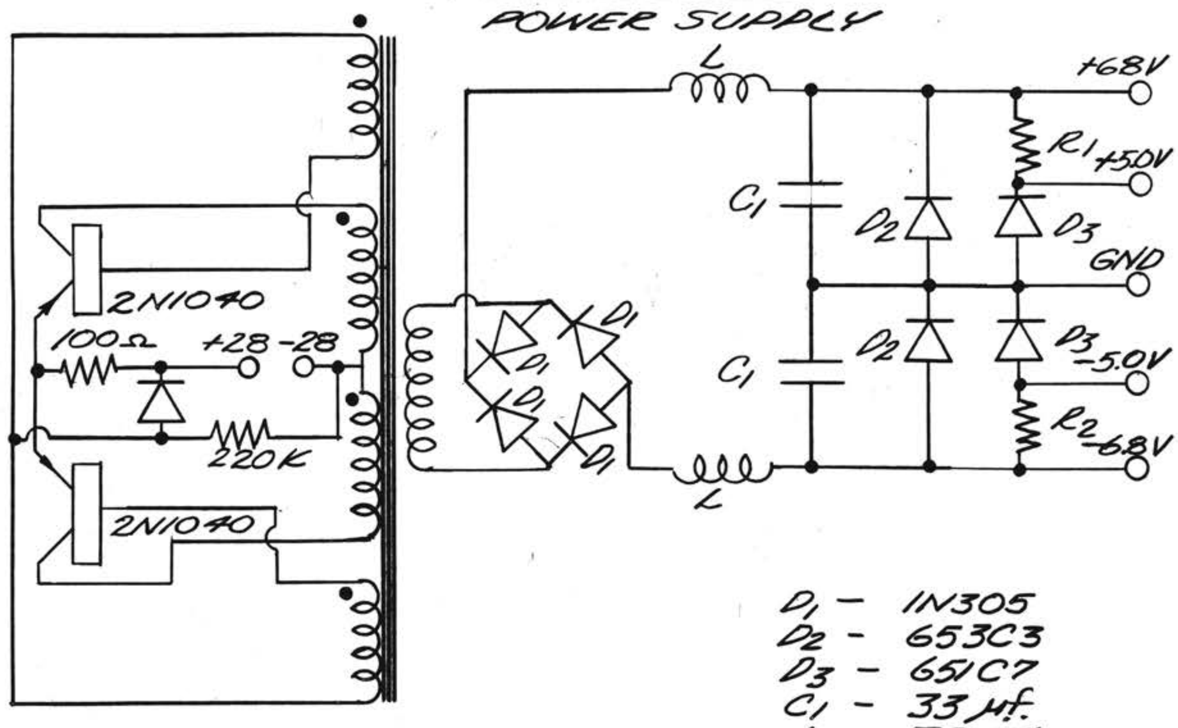



FIGURE 5-1

POWER SUPPLY




$D_1$ - 1N305
$D_2$ - 653C3
$D_3$ - 651C7
$C_1$ - 33 μf.
L - 72 MH.
$R_1$ - 820 Ω
$R_2$ - 390 Ω

FIGURE 5-2

mounted upright. This is not only a space saving feature, but also greatly facilitates circuit assembly where etched circuit boards are used. In the cases of the diode and capacitor, the mode of mounting had a significant bearing on their selection as a standard type.

## Power Supply

As mentioned in Chapter IV, it was decided to use a transformer coupled TSR (transistor saturable reactor) circuit for the power supply. The power supply schematic is shown in Figure 5-2. The theoretical calculations necessary for the design are shown in Appendix A as are the calculations for all the circuits used.

The first step in the design of any circuit is specifying the load. In the case of the power supply, this includes specification of the supply voltages for the rest of the system. From the system requirements discussed in Chapter II, one supply voltage has already been fixed at +5 volts with respect to signal ground. Also discussed in Chapter IV was the need of a low impedance negative signal voltage at the inputs to the A-B mixer. In order to minimize unbalance in the secondary of the power supply transformer, -5 volts with respect to signal ground was chosen as a second supply voltage.

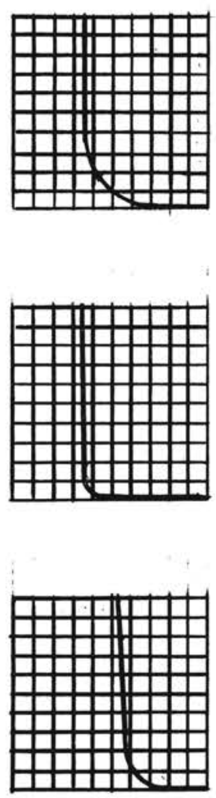
No mention has been made as yet of the voltages required by the amplifier circuit. Although not an integral part of the computer, the amplifier does share the common power supply with the computer, and therefore must be considered in the power supply design. The prototype and first production models of the micrometeor detection system were designed to use an amplifier developed during the period 1956-59 under Contract No. AF 19(604)-1908. This circuit requires one positive supply voltage in addition to signal ground. Because of the extreme sensitivity of this circuit, it was essential that it be returned to a very low impedance, well regulated and filtered supply. As shown in Figure 5-2 the author proposed to use zener diodes to provide regulation. A survey of available zeners indicated that of those types intended for low power applications, a zener with a 6.8 volts rating offered the lowest dynamic internal impedance.

This internal impedance measurement, $Z_z$, also is a commonly used index to indicate the amount of current required to bias the zener diode around the knee of its current-voltage curve. Thus a zener with low $Z_z$ will require less power to regulate than one with a relatively higher $Z_z$. This is apparent from Figure 5-3.

Based on the above discussion, coupled with the fact that the complement of +6.8 volts, namely -6.8 volts, could be utilized as the off bias return for NPN switching circuits referenced to -5 volts, a third and fourth output voltage requirement (i.e., plus and minus 6.8 volts) was added to the supply.

The zener diodes chosen to regulate the supply voltages at the required values were obtained from the T.I. 650 series. The T.I. IN746-IN759 series, a family of zeners having voltage ratings comparable to the 650 series, but available only in an axial type package, were also investigated. It was found that the IN7xx types have obviously softer knees relative to the 650 types with corresponding voltage ratings. This is shown in Figure 5-3 where the dynamic characteristics of a type IN756A and type 653C3, both rated at 6.8 ± 5% volts, may be compared. These curves are representative of those units tested. An additional factor leading to the specifying of the 650 zener series was

TYPICAL ZENER DIODE CHARACTERISTICS



THESE CHARACTERISTICS WERE TAKEN WITH A TEKTRONIX TYPE 575 TRANSISTOR CURVE TRACER. THE VERTICAL SCALE WAS 0.1 MILLIAMPERE PER DIVISION, THE HORIZONTAL SCALE 1 VOLT PER DIVISION. READING FROM THE TOP CURVE DOWN, THE DIODE TYPES SHOWN ARE 1N756A, 653C3, AND 651C7, ALL MANUFACTURED BY T.I.

FIGURE 5-3

its package compatibility to the standard diode depicted in Figure 5-1.

As indicated above, the supply voltages, in addition to the regulating requirements, were also required to be well filtered. A conventional L-section filter was selected as being optimum in this application on the grounds that ripple could always be further reduced by the rather simple expediency of adding more sections. Initially a search was made for the largest valued capacitor available with voltage and physical size parameters compatible with system requirements. Because of the special nature of this application, the standard family of capacitors were bypassed in search of a polarized tantalum unit in the 25-50 microfarad range having a working voltage equal to 150% of 6.8 volts, or approximately 10 volts. A Mallory unit (TAM 33-10) was found which was satisfactory, and specified for use in the filter. This unit has a capacitance of 33 microfarads $\pm$ 20%, a working voltage of 10 volts, and a rectangular package with parallel lead orientation similar to the standard capacitor package.

Using a standard equation to find the critical inductance necessary, $L_C$ was found to be 75 millihenrys. Referring to catalogs on powder cores, a molybdenum permalloy (A.E. Co. Part No. A-050056-2) toroidal core was selected as being the smallest size available which would give the desired inductance. Using appropriate charts and tables, an inductor design was worked out resulting in a coil having 75 millihenries inductance and a nominal 44 ohms d-c resistance. Again using a standard equation the ripple factor was calculated and found to be 0.0012. Using 6.8 volts as the basic d-c voltage, the a-c ripple checks out to be approximately 8.0 millivolts, which is satisfactory.

The rectifier circuit used is a standard full wave bridge type which uses four standard diodes. The basis for using this type diode has been covered previously, and need not be repeated here.

The transformer design turned out to be one of the more challenging and perplexing problems of all the circuits. As indicated in Appendix A, the design of the regulator and filter circuits are relatively simple and straightforward insofar as determination of component values is concerned. However, Appendix A does not reveal the complete story regarding the design of the transformer coupled TSR portion of the circuit, since how does one equate engineering judgement to an equation?

After considerable study of the problem, the author set down three basic steps to be determined prior to initiating design of the TSR. These were:

1) Selecting suitable transistors with which to excite the transformer primary.

2) Selecting a suitable core size and material for the transformer.

3) Selecting preliminary operating parameters around which to design the circuit.

Steps (2) and (3) above are interrelated inasmuch as the optimum operating parameters are a direct function of the core material and size used.

Bearing in mind the desirability of low losses, (i.e., high efficiency) in the power supply, germanium transistors were considered before silicon on the basis that the greatest losses directly attributible to the transistors would be due to high saturation resistance and/or low $h_{FE}$. Obviously, the standard transistor type would not suffice since a maximum collector swing of greater than 56 volts, or twice the primary supply voltage, was anticipated. The standard transistor has a 15 volt collector to emitter rating.

Of all the units considered, only one type meets all the requirements deemed necessary by the author. This is a type 2N1040 manufactured by Texas Instruments. Its package size is comparable to and, in fact, smaller than the standard unit. Cost wise it compares very favorably with any silicon unit which could fill this application. The collector voltage rating is 80 volts which allows sufficient margin for safety. The forward current transfer ratio, $h_{FE}$, is 70 at 25° C in the collector current region (4 to 20 milliamperes) where it will be operated. The common emitter saturation resistance is low (0.2 ohms), and the power rating (5-7 watts at $75^\circ$ C) more than ample.

As mentioned above, steps (2) and (3) are interdependent. Consider the well-known transformer equation:

$$E = 4.44\ B_m AfN \times 10^{-8} \tag{5.1}$$

where: $E$ = Supply voltage

$B_m$ = Flux density

$A$ = Cross section area of core

$f$ = Frequency

$N$ = Primary turns.

Rearranging the above equation and combining constants to facilitate discussion;

$$K = \frac{E}{B_m AfN} \tag{5.2}$$

Note that for this particular design E, the primary supply voltage from the Lockheed vehicle, is also a constant. Therefore;

$$B_m AfN = k' \text{ (constant).} \tag{5.3}$$

Normally the minimum area of the core is limited by the transformer power requirement. However in this application, area is only significant in its related effect on the other three variables above. From a size and weight consideration it is desirable, where these parameters are factors in the design, to strive for a minimum area. Also, since core losses are partially a function of core weight, being generally specified for a given core material in watts per pound at a reference frequency, the desirability of keeping A small is further indicated.

If $B_m$ and N are assumed fixed, momentarily, then reducing A causes the frequency to increase. Core losses can be theoretically predicted for a given core material by the equation:

$$\frac{R}{\mu L} = K_1 f + K_2 f + K_3 f^2 \quad . \tag{5.4}$$

Notice that core losses increase by a power of two as frequency increases. Therefore one might conclude that frequency might be minimized at the expense of core area with favorable results.

The cross sectional area of the core and the maximum number of turns are related in the following manner. If the overall dimensions allowable for the core are fixed, then as area increases, the diameter of the eye must decrease. Hence the number of turns which can be realistically threaded through the eye of the core must also decrease.

$B_m$, the saturation flux density, also differs with the type of core material. Because of the extremely low power application intended for this transformer, only tape wound toroidal cores were investigated. This was done on the basis that the tape thickness could be made very small, and thus equivalent to thin lamination type construction. Thin laminations tend to decrease eddy currents within the core, a principal cause of core losses. Secondly, since there are no air gaps to contend with in the toroid configuration, the maximum effective working permeability for a given core material is obtained, while flux leakage is reduced to a minimum.

Three different core materials, all exhibiting a square loop magnetic characteristic and available in the tape core package were investigated. These were Supermendur, Deltamax, and Supermalloy type material. The names used here are trade names coined by Arnold Engineering Company, one of the principal suppliers of magnetic materials in this country. Reference is made to A. E. Co. literature for the type and percentage of metal making up each alloy.

Table I shows typical magnetic properties of these core materials. The three types listed cover a wide range of core materials in the very high permeability group of alloys. Readily apparent from the table is the wide range of peak saturation flux density. Also the typical trend of all other parameters with respect to $B_m$ can be seen. For example, consider Supermendur with $B_m$ equal 22,000 gausses.

TABLE I

TYPICAL PROPERTIES OF CORE MATERIALS

Frequency = 400 cps

| Parameter | Supermalloy | Deltamax | Supermendur |
|---|---|---|---|
| Specific Gravity (gm/cc) | 8.77 | 8.25 | 8.15 |
| Electrical Resistivity ($\mu$ohm-cm) | 65 | 45 | 26 |
| Peak Induction, $B_m$ gauss | 7,500 | 15,000 | 22,000 |
| Residual Induction, $B_R$ gauss | 5,000 | 14,000 | 21,000 |
| $B_R/B_m$ ratio | 0.7 | 0.92 | 0.97 |
| Max. Permeability ($B/H_C$) | 100,000 | 40,000 | 19,000 |
| $H_C$, oersteds ($\Delta B = B_m$) | 0.1 | 0.5 | 9.76 |

One might correctly assume that for a fixed core size with a maximum number of turns also fixed, the natural frequency of a Supermendur transformer would be one-third that of a similarly built Supermalloy transformer, $B_m$ for Supermalloy being approximately one-third that of Supermendur. From our previous discussion one might incorrectly assume that as a result core losses in the Supermendur transformer would also be less, due to the $K_3f^2$ term of equation 5.4. Not mentioned was the fact that $K_3$, the eddy current constant, is a function of the resistivity of the core material. Since power is proportional to $I^2$, it should be deduced from the table that the $K_3f^2$ portion of core loss for Supermendur at a given frequency and core size is comparable to that of a Supermalloy core of the same size, but three times the frequency.

Another consideration to be taken in predicting toroidal core performance is the loss commonly termed hysteresis loss. This is a function of the core's B-H characteristics, and is proportional to the area enclosed by the individual core's B-H curve or hysteresis loop. When inspecting different core materials with regard to choosing one for a low power application, a good first approximation of the relative hysteresis component of core loss is the product of $H_C$ and $B_m$ and the $B_R/B_m$ ratio for each core.

So-called tape cores are available in a variety of tape thicknesses. The more common are 1, 2, 4 and 12 mil tapes. By using a thinner tape core, the effect is similar to using thinner laminations in that eddy currents are effectively reduced. This effect is counterbalanced by the so-called stacking factor which varies fr
about 0.7 for 1 mil tapes to 0.95 for 12 mil tapes. The produ

the stacking factor and the gross cross sectional area of the core used gives the effective area to be used in equation 5.1. A good rule of thumb to use in selecting the tape thickness is: use 12 mil for frequencies less than 60 cycles, 4 mil up to 400 cycles, 2 mil between 400 and 1000 cycles, and 1 mil above 1000 cycles. These are not rigid requirements, but rather suggested first approximations.

The core finally chosen for the power supply transformer was a composite made up of two 6T8043-S1 cores taped together. The next size larger core was too large in its outside diameter to be compatible with packaging requirements. A single 6T8043 core size was found by trial and error to have too small a cross sectional area for the transformer to be designed for 1000 cps operation. The rather unique solution of paralleling the two magnetic core circuits effectively doubled the core area with no adverse effects on the other transformer parameters.

The reader should correctly conclude from the above discussion, which only briefly touched on one phase of transformer design (i.e., a saturating transformer using a square loop magnetic core material) that there are no hard and rigid rules and/or procedures set forth in the literature which must be followed. The author's experience indicated a major problem encountered in this particular design was the selection of a suitable core size and material based on the requirements and limitations to be met. Once this is accomplished, the procedure outlined in Appendix A will prove extremely accurate in completing the transformer design.

To the author's knowledge, the particular TSR circuit, that is, the transistor multivibrator with transformer coupled feedback, depicted in Figure 5-2 is not a commonly used configuration. This particular circuit evolved from the author's efforts to neutralize certain adverse circuit characteristics observed at the high and low temperature extremes specified for the unit. At both temperature extremes the net effects to the TSR were similar; the dropping out of the drive transistors from bottoming, an apparent overheating of the transistor junctions, and a final thermal instability or runaway condition in the circuit taking place. The primary causes at the temperature extremes were theorized to be entirely separate. This theory was substantiated partially by the fact that the "fix" in the circuit configuration, (i.e., the sensistor added between the common emitters of the transistors and the positive return of the primary supply) based on the theory did alleviate the problem.

Temperature evaluation of the circuit isolated the problem to the transistor oscillator circuit. Using this as a starting point, it was proposed that, with increasing temperature to $+60^{\circ}$ C, $h_{FE}$ approximately doubled the $+25^{\circ}$ C value for this particular transistor operated in this collector current range. This coupled with a slight but significant decrease in $V_{BE}$ resulted in a substantial increase in $I_p$, the transformer primary winding excitation current which could be supported by a constant base drive. The implications of this, requiring saturating transformer operational theory, are discussed below.

Consider a single transistor of the driving pair. When driven from "off" to "on", initially the collector sees a large inductive impedance, and $I_p$ or $I_c$, which is considered to be the sum of two current components (a constant resistive $I_R$ component and an exponentially increasing inductive $I_L$ component) equals $I_R$. Since the base drive is a square wave, the transistor is driven deep into saturation. As $I_L$ increases, $I_p$ increases until $I_B$, the base drive, can no longer support the load current. This value of $I_p$ is essentially $I_B$ times $h_{FE}$. The exponential increase in $I_L$ is a result of fewer lines of flux being switched in the core as saturation of the core is approached. This is shown in Figure 5-4.

The circled numbers on the two curves are corresponding points on the flux and current waveshapes with respect to time. In the region of transistor switchover depicted by the shaded area of curve (b), the power dissipated across the transistor junction is a maximum. From points 3 to 4 the change in flux is very small, hence the feedback drive holding the opposite transistor off during the period 1-4 is decreased. When this can no longer supply the required $I_{CBO}$, the second transistor starts to conduct, the base drive polarity is reversed to the first transistor, and it is driven "off to point five on Figure 5-4(b).

Consider now that as temperature is increased, $I_p$ max. essentially doubles which results in an increased storage time for the transistor being switched "off". Also the transistor being turned "on" responds slower due to the greater $I_{CBO}$ which must be overcome and the increase in core characteristics bounded by points 3 and 4. Hence the time base of the shaded area on curve (b) is also increased until a temperature is reached where the increase of the shaded area causes a corresponding increase in power dissipation across the transistor junction such that the increase of junction temperature above the ambient causes transistor thermal runaway.

At the low temperature extreme, the $h_{FE}$ transistor parameter decreases by approximately 50% with respect to the 25° C value, while the $V_{BE}$ parameter increases. The net result is that the $H_c$ (proportional

TSR PRIMARY CIRCUIT OPERATING CHARACTERISTICS

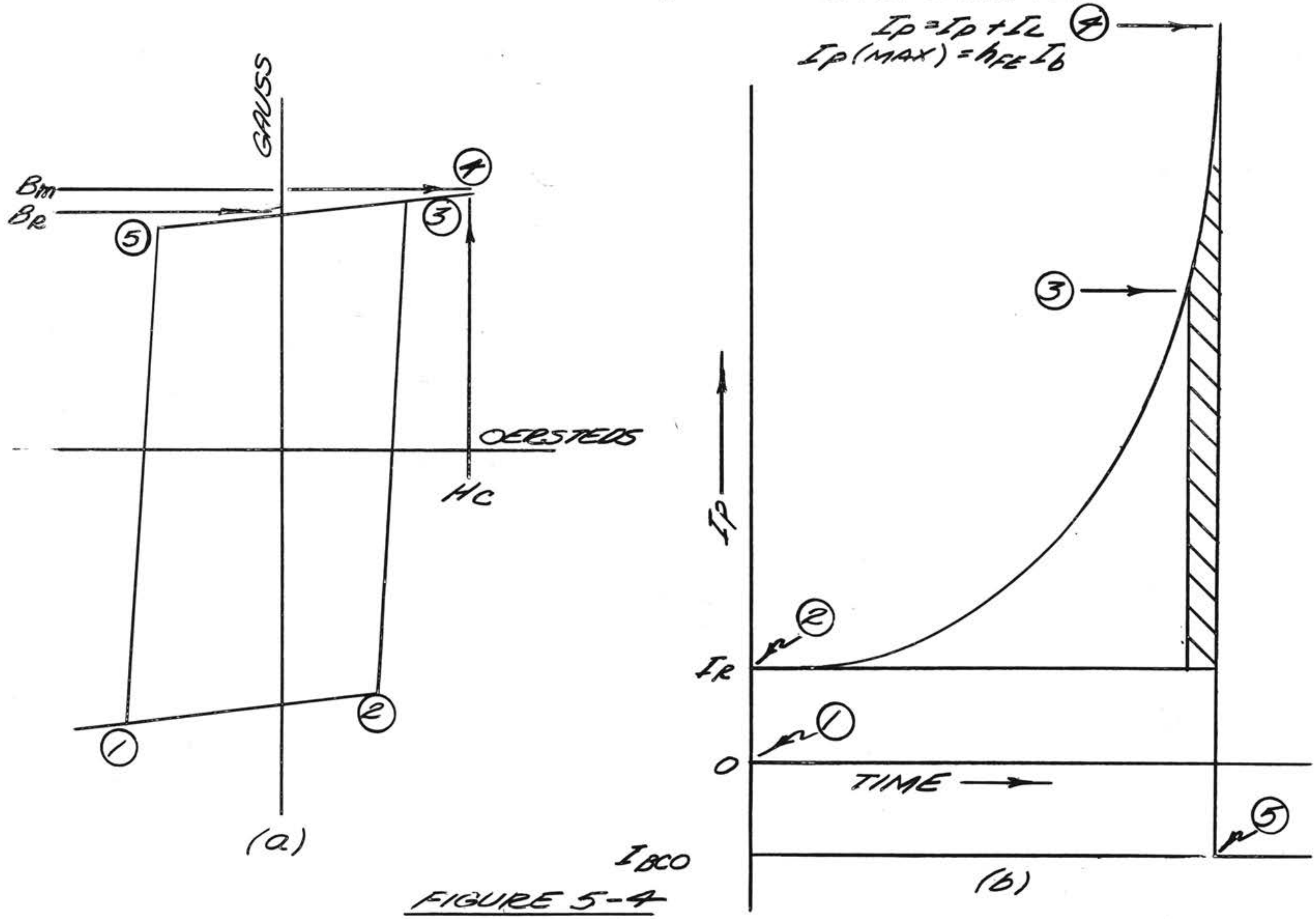



FIGURE 5-4

to the product of $I_p$ and the number of primary turns on the transformer) corresponding to point 3 on curve (a) is not reached, the transformer no longer saturates, but rather follows minor hysteresis loops, $B_m$ of equation 5.1 decreases thus increasing frequency, and core losses increase correspondingly. Since $I_R$ component of $I_p$ represents losses, $I_R$ also increases and less $I_L$ is available to switch flux, thus resulting in less drive, ad infinitum, until oscillations cease.

In both cases the primary current increases beyond a safe limit with respect to the transistors, and the transformer secondary voltage drops to zero. The first case is described as a positive feedback loop around the transistor base-emitter circuit which approaches, and eventually exceeds unity. The second is a result of a decrease in loop gain around the transistor base-collector circuit which approaches and becomes less than unity.

By utilizing a sensistor, (i.e., a temperature sensitive resistor with a large (0.7%/$^\circ$ C) positive temperature coefficient) in the transistor emitter circuit, the feedback is reduced around the base-emitter circuit at high temperatures, while the base drive, and hence overall gain around the base-collector circuit, is increased at low temperatures. As a result the RMS value of $I_p$ is stabilized within a range of 6-8 milliamperes over a thermal test environment of -30$^\circ$ C to +70$^\circ$ C.

## Switching Circuits

Again based on the author's previous experience, it was decided to supplement the standard transistor decision with a standard switching circuit building block to be used extensively in the other, more complex switching circuits making up the discriminator and register sections.

The circuit chosen was a saturating, so-called voltage mode, inverted amplifier circuit using the common emitter configuration. This type circuit is commonly used in switching applications where the pulse repetition rate is in the 10 to 100 kc range. The C.E. configuration provides both current and voltage gain, the current gain for the overall circuit theoretically approaching the $h_{FE}$ of the transistor as a limit.

The supply voltages for the basic inverter block are +6.8 volts for the collector return, -5.0 volts for the emitter reference, and -6.8 volts for the base "off" bias return. (As mentioned previously, the +5.0 volt return is reserved for the output driver.)

As indicated in Appendix A, a modified "worst-case" design based on the author's experience was employed throughout. Because of the care taken to regulate and filter the supply voltages, it was not felt necessary to take any tolerance on them, as is normally done in applying a "worst-case" design philosophy. To compensate for this, the resistors were segregated in stock in values of 0 to + 10% tolerance and 0 to -10% tolerance. All the resistors for a single production unit were then taken from the same tolerance group.

The trigger configuration is simply two basic blocks cross-coupled from collector to base. The "worst-case" on consideration for the basic block load current is the total current required to drive its own collector to saturation ($11.8^V/R_c$), the input current to another block which is "off", and the input current to the lowest impedance mixer input (400 microamperes). $R_c$ must be small enough so that $6.8^V/R_c$ is sufficient current to drive one other "on" stage plus the total $I_{CBO}$ of itself and the $I_{EBO}$ of the common base mixer stage.

The binary coupling input circuit for the trigger was designed for a negative shift input to take advantage of the low saturation resistance of the NPN transistors. Base triggering was used with the trigger circuits rather than collector pullover to reduce the trigger voltage shift required. Emitter triggering was not applicable due to the low impedance separate return used.

The design equation used to calculate the capacitor values for the binary coupler and the base input network evolved from the author's own experience. A subsequent search of the literature has revealed at least one reference (General Electric Transistor Manual) which gives comparable design formulas for this application. In both cases the results indicate capacitor values in the same order of magnitude.

Figure 5-5 shows a typical binary trigger circuit completely assembled Again note the use of the standard diode as the steering diodes in the binary coupling circuit.

The design of the single-shot circuit utilized with the discriminator used one of the standard inverters as the normally "off" side of the multivibrator. However, the normally biased "on" stage had to be specifically designed for this application. A general discussion of this circuit is included in Chapter IV. A schematic is depicted in Figure 4-3.

Capacitors $C_1$ and $C_2$ were found from the calculations in Appendix A to require values beyond the value of the standard capacitor series. However, the desired values were readily available in a tantalum type. The insensitive period for the circuit was arbitrarily selected as five to ten milliseconds. This was based on previous experience with the amplifier circuit which would be used to trigger the monostable multivibrator. The basic electrical impulse from the microphone-amplifier system consists of a sharp leading edge, followed by a relatively long wave train of approximately 100 kilocycle fundamental ringing. The amplitude envelope of this wave-train is of random and unpredictable shape; low velocity stimulation at the momentum values anticipated indicates that the duration is roughly proportional to the momentum stimulus. For an impinging particle with momentum of the order of $10^{-2}$ gm cm/sec, the train is essentially decayed to zero in about 6 milliseconds. Use of recovery times of this magnitude insures that double triggering of the logic will not occur, and results in a maximum utilizable duty cycle for succeeding stimuli.

BINARY CIRCUIT

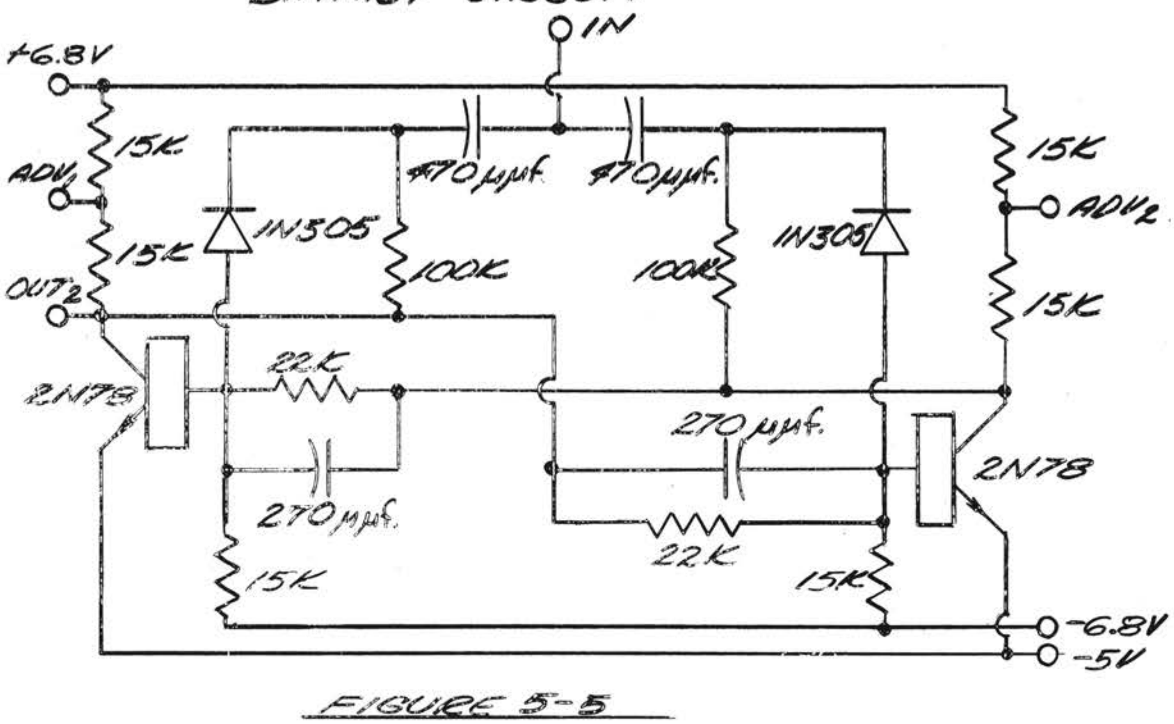



FIGURE 5-5

A final output stage, again a standard inverter block, was added to the discriminator circuit to power the A and/or B trigger pulse lines. Actually this pulse doesn't require much power. However, since it is working into a capacity load, and a sharp leading edge is required, a very low impedance source was desired. This is available from the standard inverter block turning "on". The desired pulse might have been taken from the collector of $T_1$ in Figure 4-3, however, it proved necessary to couple the input signal into the single shot stage at this point to obtain the desired desensitizing of the input signal modulations. Also it was felt that any possible source of line noise should be isolated, since a 0.2 volt pulse on the binary input line is sufficient to flip the triggers. A positive trigger pulse was not used since for it to be a low impedance pulse with a fast rise time implies the utilization of a PNP transistor, which is a choice contrary to the design philosophy adopted.

Briefly summarizing Chapter V, and covering any minor details that might have been implied but not spelled out previously, the bases for and the details of the design of all circuitry used in the computer have been discussed with a sincere effort to reveal the author's thinking in this area. Observe that in considering the logic used, a lower signal voltage should be considered a logical "1", while a higher signal level is a "0". Although contrary to conventional practice, this logic is not unique in that many large computer systems not only use negative logic, but use both negative and positive logic coupled with dual logic levels interchangeably. A final point involves temperature indication. To provide a key to output signal drift relative to temperature, a Ruge Associates type BN1200 temperature gauge is mounted adjacent to the A-B mixer section. One side of this gauge is connected to signal ground, the other side of the element being brought out to a separate commutator bar on the programmer. This permits actual in-flight measurements of the ambient system temperature within that portion of the computer most sensitive to temperature variations.

## CHAPTER VI

## PACKAGE DESIGN

Several different ways of packaging this system for mounting on the Lockheed vehicle were considered. These included both the internal package of the electronic components themselves and the external dimensions of the package. Preliminary discussions with Lockheed indicated that size and weight of the package were not critical as compared to the forerunner of this system. The size of the detecting plate was tentatively specified as approximately 11-1/2 inches by 6-1/2 inches. The original specification was to include the microphone and electronics all in an integral package about 1-3/4 to 2 inches deep. Therefore, it was agreed that a circuit-package having external dimensions of approximately 11-1/2 by 3 by 2 inches would be satisfactory. Figure 6-1 depicts the over-all package adopted for the electronics package.

Based on experience encountered with previous meteor-detecting systems, it was believed that the building block philosophy of internal packaging would prove very satisfactory insofar as testing, trouble shooting, and replacing any failures detected within the system prior to shipment were concerned. Thus, (for example) in the event a single trigger within the system failed prior to shipment, it would be possible to remove that particular circuit from the system and replace it with a comparable component or circuit. Note that it would not be necessary

EXTERNAL PACKAGE ASSEMBLY

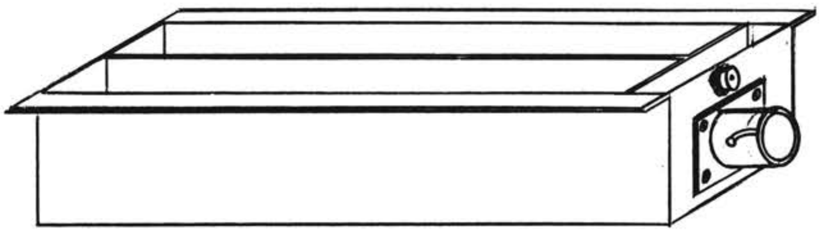


FIGURE 6-1

INTERNAL PACKAGE ASSEMBLY

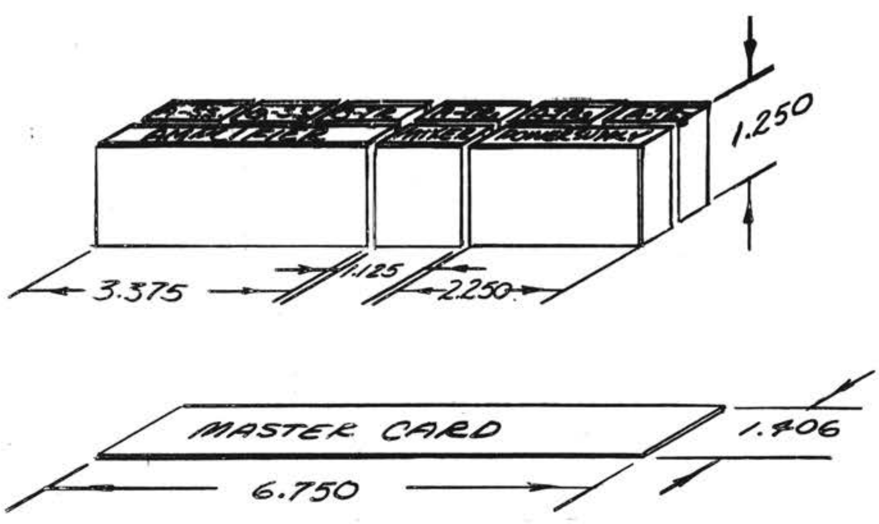



FIGURE 6-2

to rebuild the entire system since individual spare circuits or building blocks could be maintained in stock for replacement purposes.

Using this as the basic internal packaging-criteria, suitable means of interconnecting the various building blocks were investigated and considered. More to facilitate standardization and ease of construction of the various building blocks than anything else, it was decided to utilize etched-circuit boards on which to mount the individual circuit components. By bringing suitable leads out of the etched-circuit boards, it was then possible to interconnect them by means of a master etched-circuit board. This could then be completely assembled into a single package and mounted within a metal container of some type.

A rectangular basic etched-circuit board was arbitrarily selected, having dimensions of 1-1/8 by 1-1/4 inches. Tentative circuit layouts indicated that all circuits used in the system could be mounted within this space. Also, in order to meet environmental shock and vibration requirements tentatively specified by Lockheed, it was determined that after mounting the components making up a circuit on a particular etched-circuit board, these cards and components would then be placed in a mold and Eccofoam used to encapsulate the building block. A suitable mold was machined and prototype modules were cast and evaluated. Very little difficulty was encountered in reproducing a uniform size module for use throughout the system.

It is apparent that by using this philosophy of packaging, the ratio of volume utilization by components to the volume available is quite small. Actually, the packing factor is less than 20%. However, as indicated, weight and volume were not prime considerations in the packaging, while ease of construction, assembly, testing, replacing failures, and actually revising the entire system design were. These are greatly facilitated by this type of packaging.

Following the building and testing of the prototype model, the above described package was revised to take advantage of the experience gained in design and development of the prototype.

Initially five microdot connectors were used to interconnect the system power and signal lines to the Lockheed vehicle system. Connecting and disconnecting all five microdots during testing proved a time consuming and tedious effort. In addition there was a tendency to connect cables to the wrong receptacle. To alleviate these difficulties, four of the microdots were replaced by a single Amphenol (P/N 67-02E12-7) 7 pin quick release receptacle with one microdot connector being retained for the microphone input. Figure 6-1 depicts the final external package used with the system.

Another revision in the initial package philosophy is reflected in Figure 6-2. It was implied above that all modules were constructed to a common size. This was true for the prototype model. However, subsequent testing and evaluation indicated that the modular assembly shown in Figure 6-2 is a more satisfactory arrangement. All modules except the amplifier and power supply assemblies retained the standard module size. The power supply assembly layout was changed from three single modules to a single "equivalent-two" module, with the amplifier requirements being reduced from four singles to an "equivalent-three" block.

A final consideration in packaging was the shielding of energy radiation from the power supply. Since the power supply was designed to oscillate at 1 kc with a square wave as the fundamental wave shape, considerable noise resulting in "false" pulsing was picked up by the by the amplifier and other signal lines. To reduce this radiated energy, the entire power supply module was electrically and magnetically shielded in the final design. This was greatly facilitated with the single "equivalent-two" module as compared to the three single modules.

Since weight was not a critical parameter, brass was used in the external package assembly rather than a lighter metal like aluminum. The reason for this was that brass is much more workable than the lighter metals which tend to flow and melt when being drilled and/or welded. Except for the end plate to which the Amphenol connector is attached, all other joints are welded for added strength.

## CHAPTER VII

## SUMMARY AND CONCLUSIONS

The primary object of this thesis was to present the philosophy, procedures, and methods used by the author to solve a unique problem in system design. The subject was treated in a logical manner, in that the development of the system was described sequentially, chapter by chapter, as it evolved from an idea in March, 1959 to hardware six months later.

The prototype model has a record of more than 1000 hours testing and evaluation to the Lockheed specification described in Chapter II. Once the initial "bugs" and malfunctions were worked out of the system, the unit has enjoyed a good reputation for reliability. Of the eleven systems delivered to AFCRC to date, only one legitimate malfunction has been recorded. This was due to the catastropic failure of a sensistor in the power supply section during a 100 g load shock test performed by AFCRC on unit Serial No. 1 in their test facilities. The O.S.U. Research Foundation does not have the necessary equipment to perform this test, hence the faulty mounting technique leading to this failure was not discovered prior to shipment. Subsequent production units were reworked to eliminate this fault.

The reliability record described above tends to verify the approaches taken and decisions made by the author. This system was unique only in the function performed and the design limitations placed on it. It is believed by the author that, in general, the systematic procedures outlined and used to develop this system are applicable to a wide range of systems. The systems referred to here are meant to include either a simple, single stage circuit or an assembly of subsystems each composed of thousands of individual circuits.

Based on experience, both prior to and as a result of compiling the material for this thesis, the author has concluded that an engineer's basic training assists him in the logical organization, study and solution of a great many different types of problems. However, several of the problems encountered in the development of the subject system leads the author to conclude that the organizing and study must be supplemented by common sense, unbiased thought, experience, and the ability to differentiate between sound judgement and guess work.

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## APPENDIX A

### Switching Circuits

Reference: Pulse and Digital Circuits by Millman and Taub

Basic Inverter Block Design:

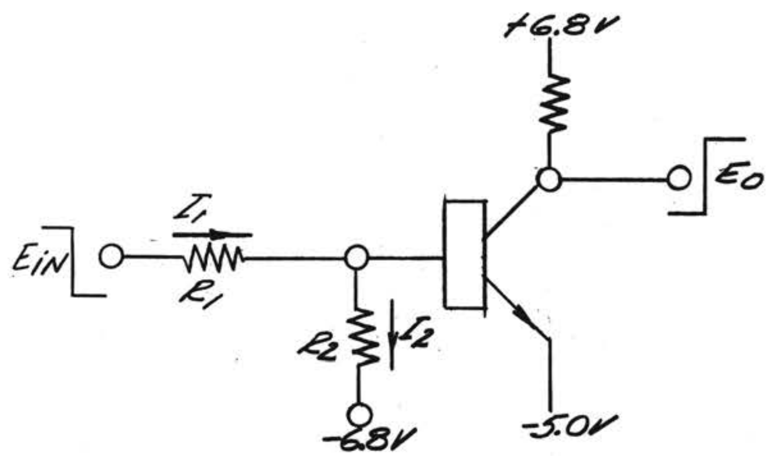



Fig. A-1. Basic Inverter Block Configuration

Design Criteria:

1. Resistors: $\pm$ 10%, 1/10 watt (Ohmite)
2. Transistors: $\beta \geqq 20$ at -30° C.

   $I_{CBO} \leqq 80$ μa at +60° C.

   $$V_{BE} = \begin{cases} +0.3^v \text{ (on)} \\ -0.2^v \text{ (off)} \end{cases}$$

$$I_b = \frac{I_c \text{ (max)}}{\beta \text{ (min)}} = I_1 - I_2 \text{ (on case)}$$

Assume: $I_c$ (max) $\leqq$ 1.2 ma.

$E_{in} \geqq +0.5^v$

Whence (assuming +10% change in $R_1$ and -10% change in $R_2$):

$$\frac{1.2}{20} = \frac{5.2^v}{(1.1)(R_1)} - \frac{2.1^v}{(.9)(R_2)}$$

$$.06 = \frac{4.72}{R_1} - \frac{2.33}{R_2} \qquad (A.1)$$

$I_{CBO} = -I_1 + I_2$ (off case)

Assume: $E_{in} \leqq -4.9^v$

Whence: $I_{CBO} = \dfrac{-0.3}{(.9)(R_1)} + \dfrac{1.6}{(1.1)(R_2)}$

$$.08 = \frac{-.33}{R_1} + \frac{1.45}{R_2} \qquad (A.2)$$

Solving equations (1) and (2) simultaneously (Determinants)

$$\Delta = (4.72)(1.45) - (.33)(2.33) = 6.08$$

$$R_1 = \frac{\Delta}{(.06)(1.45) + (.08)(2.33)} = 22.25^k$$

$$R_2 = \frac{\Delta}{(4.72)(.08) + (.33)(.06)} = 15.3^k$$

Choosing next smallest RETMA size resistors:

$R_1 = 22^k$

$R_2 = 15^k$

Checking selected size resistors:

$$I_b = \frac{4.72}{22} - \frac{2.33}{15} = 59.2\ \mu a$$

$$I_{CBO} = -\frac{.33}{22} + \frac{1.45}{15} = 82\ \mu a$$

Verifying $I_c$ (max): (Note: These calculations based on two stages cross coupled for the trigger configuration, and driving lowest impedance mixer input leg as depicted in Figure A-2 below.)

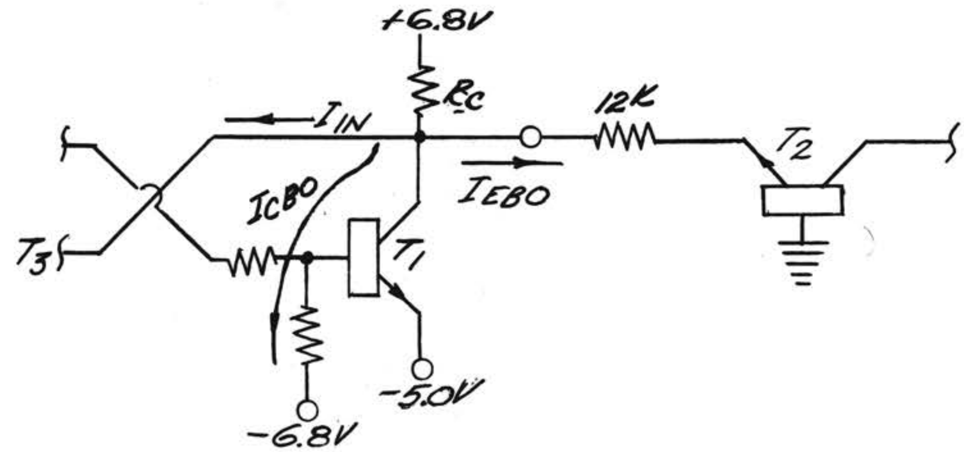



Fig. A-2. Maximum Load Configuration

$$I_c\ (T_1\ \text{off}) = I_{in}\ (T_3) + I_{CBO}\ (T_1) + I_{EBO}\ (T_2)$$

$$\frac{6.8 - 0.5}{(1.1)(R_c)} \approx \frac{4.72}{22}\ ;\ R_c = \frac{6.3}{(1.1)(.214)} = 27^k$$

Using calculated value of $R_c$:

$$I_c\ (T_1\ \text{on}) = I_{in}\ (T_3) + I_c + I_1$$

$$I_c\ (\text{max}) = \frac{2.3}{(.9)(15)} + \frac{11.8}{(.9)(27)} + \frac{5.0}{(.9)(12)}$$

$$= 0.17 + 0.485 + 0.463$$

$$= 1.118\ \text{ma.}$$

$$\therefore\ I_b\ (\text{max})\ \text{Required} = \frac{1.118}{20} = 56\ \mu a$$

Thus the basic design of the inverter block is satisfactory.

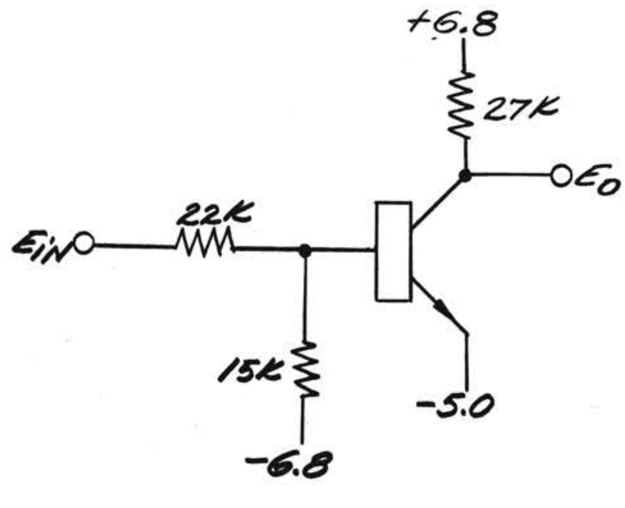



Fig. A-3. Final Inverter Block Circuit with Component Values

Monostable Multivibrator Design:

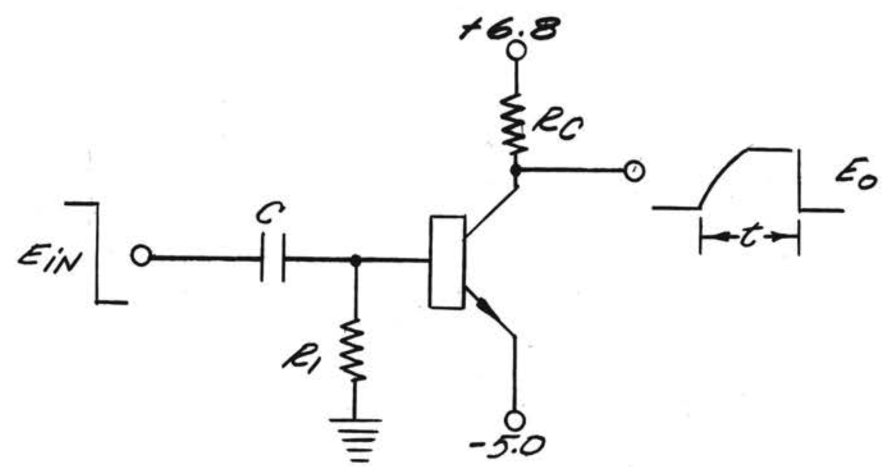



Fig. A-4. "On" Stage Configuration

Design Criteria:

1. $\Delta E_{in} \leqq 4.0$ volts

2. $t \geqq 7.5$ ms.

From physical size considerations, choose C = 0.22 μfd.

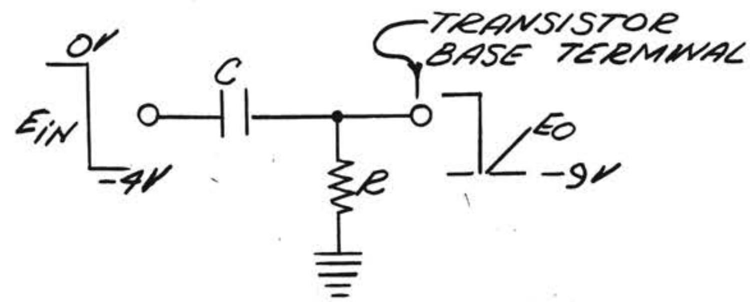



Fig. A-5. Differentiating Network for Base Input Signal

$$E_o = -9\epsilon^{\frac{-t}{RC}}$$

Letting:

$$t = 7.5 \times 10^{-3}$$

$$C = 0.22 \times 10^{-6} \text{ f.}$$

$$E_o = -5$$

Then:

$$\frac{5}{9} = \epsilon^{-\frac{7.5 \times 10^3}{122 R}}$$

$$\ln \frac{5}{9} = -0.59 = \frac{-34 \times 10^3}{R}$$

$$R = \frac{34^k}{.59} = 57.6^k$$

Choosing nearest RETMA size resistor:

$$R = 56^k$$

$$I_b = \frac{5}{(1.1)(56)} = 81 \ \mu a$$

$$\therefore I_c \text{ (max)} = (20)(81) = 1.6 \text{ ma.}$$

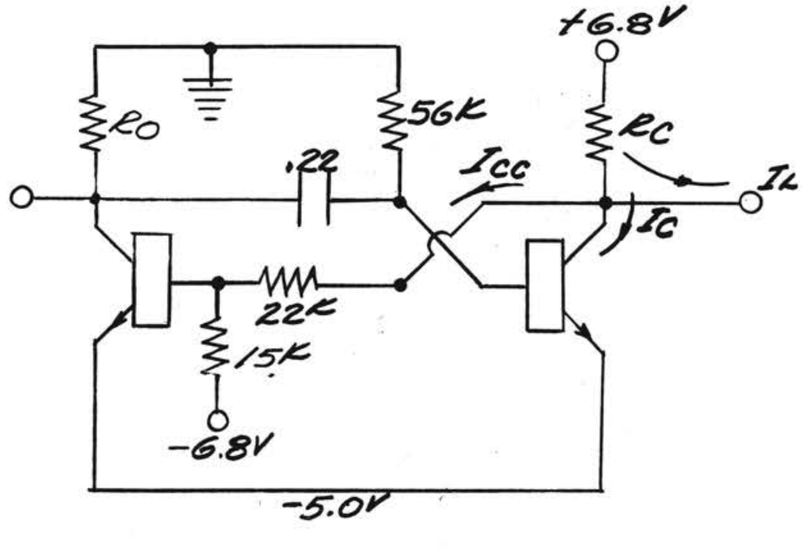



Fig. A-6. Basic Inverter Block Interconnected to "On" Stage

For $R_c$:

$$\frac{6.3}{1.1\ R_c} \geq \frac{4.72}{22}$$

$$R_c \leq \frac{(6.3)(22)}{(1.1)(4.72)} = 26.7^k$$

Choose $R_c = 22^k$ (next smallest RETMA value)

For $R_o$:

$$\frac{5.0}{.9R_o} \leq 1.2$$

$$R_o \geq \frac{5.0}{(.9)(1.2)} = 4.6^k$$

$\therefore$ Choose $R_o = 16^k$

## Coupling Capacitor Calcuations

Reference: <u>Transients In Electrical Circuits</u> by Lago and Waidelich

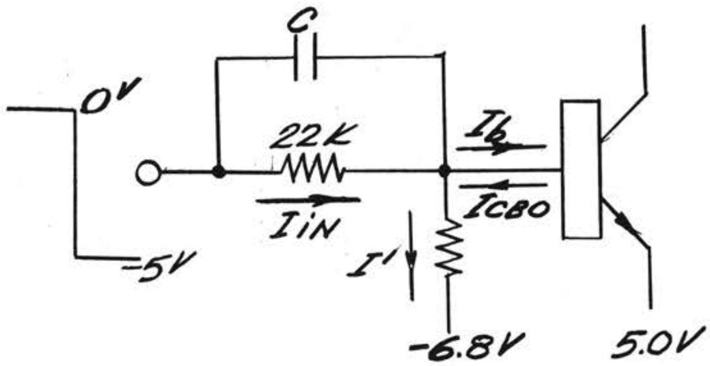



Fig. A-7. Base Speedup Capacitor

Assume: $e_{in}$ = Step Function

$I_b = 60\ \mu a$

$I_{CBO} = 80\ \mu a$

$I'$ = Constant

$\therefore\ \Delta I\ (\text{base}) = 60 + 80 = 140\ \mu a$

$Q_o = CE = 5C$

Choose: $I_{cap.} = (10)(\Delta I_b)$ for 1.0 μ sec.

$\therefore\ Q_o = 5C = (10)(\Delta I_b)(10^{-6})$

$C = (2)(140)(10^{-12})$

$= 280\ \mu\mu f.$

Choose: $C = 270\ \mu\mu f$.

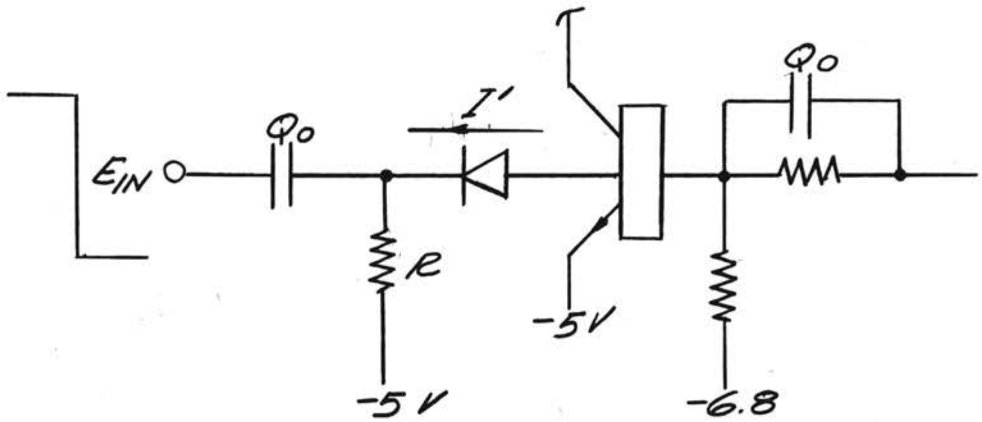



Fig. A-8. Binary Coupling Capacitor

Assume: $\Delta e_{in} = 3.0^{V}$

$I'$ (req'd) = $(10)(140)(10^{-6})$ for 1 μs.

$\therefore Q_0' = 3C = (10)(140)(10^{-12})$

$C = 470\ \mu\mu f$

Checking Recovery Time:

"On" to "Off"

Allow $1 \times 10^{-6}$ seconds for trigger to flip.

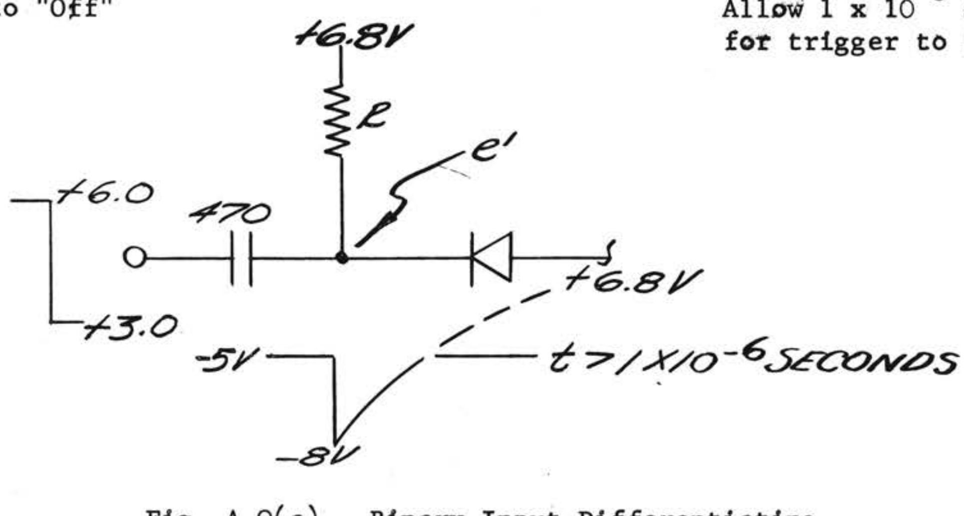



Fig. A-9(a). Binary Input Differentiating Circuit

$$\Delta e' = 14.8\ (1 - e^{-t/RC})$$

$$\frac{3}{14.8} = 1 - e^{-t/RC}$$

$$e^{-t/RC} = 1 - 0.2 = 0.8$$

$$e^{t/RC} = 1.25$$

$$t/RC = \ln 1.25 = 0.223$$

$$\therefore\ R \geq \frac{t}{0.223C} \geq \frac{10^{-6}}{(.223)(470)(10^{-12})}$$

$$R \geq \frac{10^{6}}{105} \cong 10K$$

"Off"to "On"

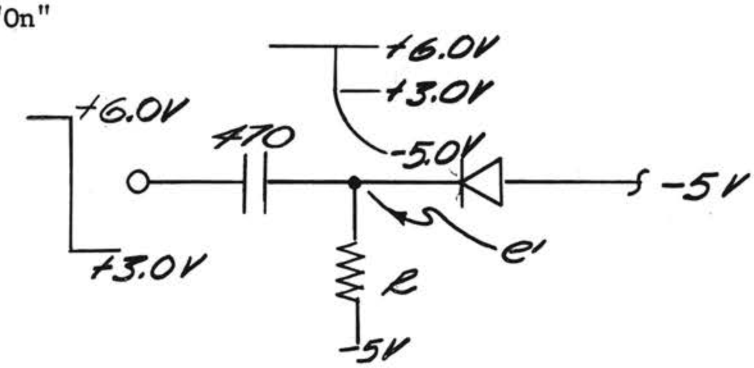



Fig. A-9(b). Binary Input Differentiating Circuit

Require $e' = -5^{V}$ at $t = 10$ ms.

$$3RC \leq 10 \times 10^{-3}$$

$$R \leq \frac{10 \times 10^{-3}}{(5)(470)(10^{-12})}$$

$$R \leqq \frac{(10)(10^6)}{1.4} \leqq 7 \text{ megohms}$$

(Note that this is not the "worst-case". The "worst-case" occurs when single shot recovers.)

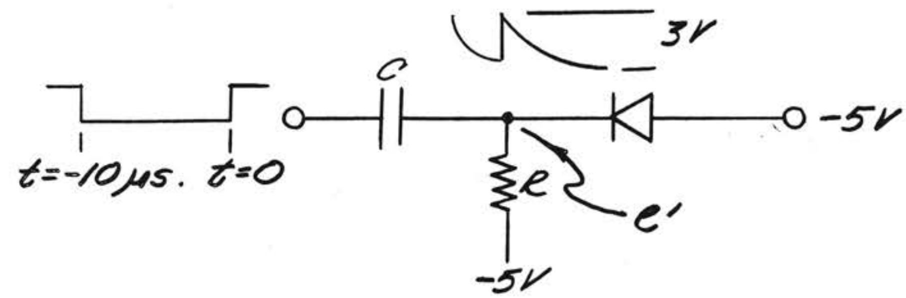



Fig. A-9(c). Binary Input Differentiating Circuit

Require $e' = -5^V$ at $t = 1.0$ ms.

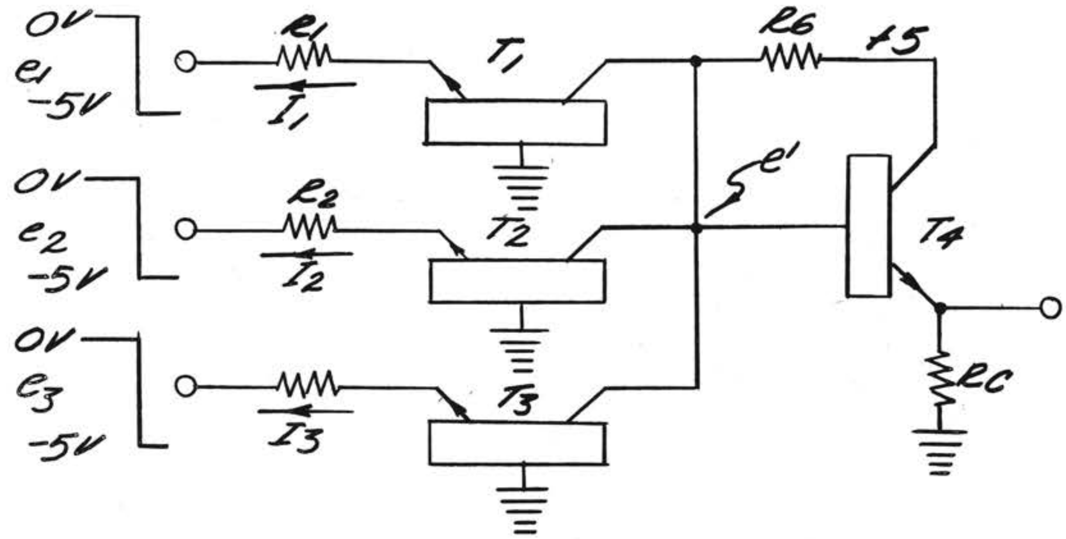
$$3 \text{ RC} \leqq 10^{-3} \text{ sec.}$$

$$R \leqq 700 \text{ kilohms}$$

Choose: R = 100 kilohms

A-B Mixer

Reference: <u>Switching Circuits and Logical Design</u> by Caldwell




$$e' \cong (I_1 + I_2 + I_3)R_6$$

Fig. A-10. A-B Mixer

For e' to represent binary inputs $e_1$, $e_2$ and $e_3$, then $I_1$, $I_2$ and $I_3$ must have weighted values.

$$\therefore \; I_3 = 2I_2 = 4I_1$$

$$I\,(\max) = I_3 = 400\ \mu a$$

$$\left.\begin{aligned} \therefore \; R_3 &= \frac{5}{400} \times 10^6 = 12.5^k \\ R_2 &= \frac{5}{200} \times 10^6 = 25^k \\ R_1 &= \frac{5}{100} \times 10^6 = 50^k \end{aligned}\right\}$$

Select
Exact Values
From Stock

To insure $T_1$, $T_2$, and $T_3$ never saturate:

Let e' (min) = 0.1 volt

$$I_{max}\,(R_6) = 400 + 200 + 100 + I_b\,(T_4)$$

$$I_b = \frac{I_c}{h_{FE}} \quad (T_4)$$

Note that when I ($R_6$) is max.

$I_c$ ($T_4$) is negligible:

$$\therefore I_{max} = 700 \ \mu a$$

Let e' (max) = 4.8 volts

$$I_{min} (R_6) = I_b (T_4)$$

Critical case is for I ($R_6$) = max.:

$$\therefore R_6 = \frac{(5.0 - 0.1)}{700} \times 10^6$$

$$= \frac{4.9}{0.7} \times 10^3 = 7.0 \text{ kilohms}$$

Choose: $R_6$ = 6.8 kilohms

Check for $I(R_6)$ = min.:

$$I_b \ (max) = \frac{(5.0 - 4.8 - V_{BE})}{6.8 \times 10^3}$$

$$I_b \ (max) = \frac{0.05}{6.8} \times 10^{-3} = 9 \ \mu a.$$

$$I_c \ (max) = (h_{FE})(9) = 180 \ \mu a.$$

$$\therefore R_c \geqq \frac{4.8}{180} \times 10^6 \geqq 27 \text{ kilohms}$$

Choose $R_c$ = 56 kilohms

# Power Supply

References:


1. Electron-Tube Circuits by Seeley, pp. 215-240.
2. Junction Transistor Electronics by Hurley, pp. 444-453.
3. Bulletin PC-203 by Magnetics, Inc.
4. Designing DC-DC Converters by Mangetics, Inc.
5. Bulletin TC-101A by Arnold Engineering Company.


Regulator:

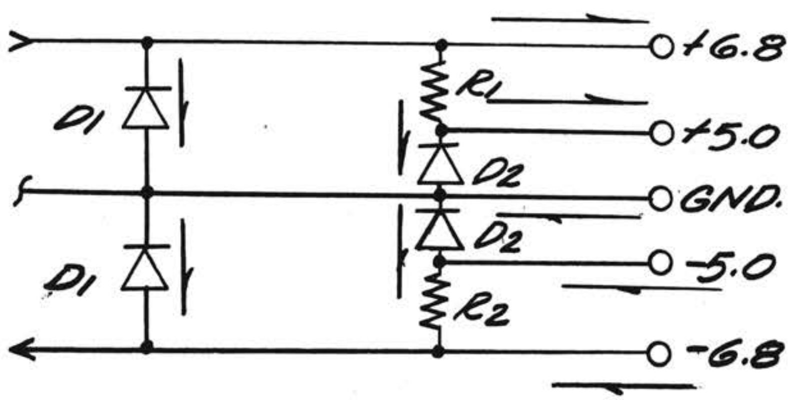



Fig. A-11. Zener Regulator Circuit

D-C Current Requirements:

$I(+6.8) = 7.0$ ma.

$I(+5.0) = 1.0$ ma.

$I(-5.0) = 3.0$ ma.

$I(-6.8) = 1.0$ ma.

$I_z(\text{min.}) = 1.0$ per zener

$\therefore I\ (\text{Total}) = (7.0 + 1.0 + 2.0)$ max.

$I_{dc}\ (\text{max.}) = 10.0$ ma.

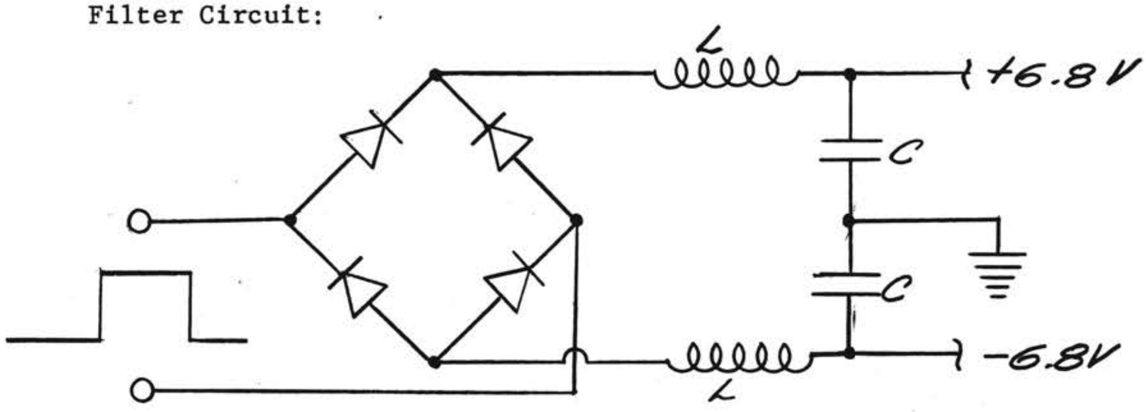
$$R_1 = \frac{(6.8 - 5.0)^{v}}{2\ \text{ma.}} \cong 820 \text{ ohms}$$

$$R_2 = \frac{(6.8 - 5.0)^v}{4 \text{ ma.}} \cong 390 \text{ ohms}$$

Choose: $D_1$ - 653C3

$D_2$ - 651C7

Filter Circuit:




f = 1000 cycles/sec

Fig. A-12. Filter Circuit

Ref. (1), p. 226:

$$L_c = \frac{2R_L}{3\omega} = \frac{R_L}{(3)(\pi)(1000)}$$

$$R_L = \frac{6.8^v}{10 \text{ ma.}} = 680 \text{ ohms}$$

$\therefore L_c$ = 72 mh. (design for 75 mh.)

Ref. (3), p. 2:

Choose: core size - 55050

Ref. (3), p. 33:

N(75 mh.) = 1200 Turns

Ref. (3), p. 32:

Wire size = AWG #38

Ref. (3), pp. 12-13:

$$R_{dc} = \frac{K_1 \times N \times K_2}{1000} \longrightarrow K_1 = 0.057;\ K_2 - 648.2$$

$$R_{dc} = \frac{(0.057)(1200)(648.2)}{1000} = 44 \text{ ohms}$$

Choose: C = 33 microfarads

Ref. (1), p. 224:

$$\text{Ripple Factor} = r = \frac{e}{E_{dc}}$$

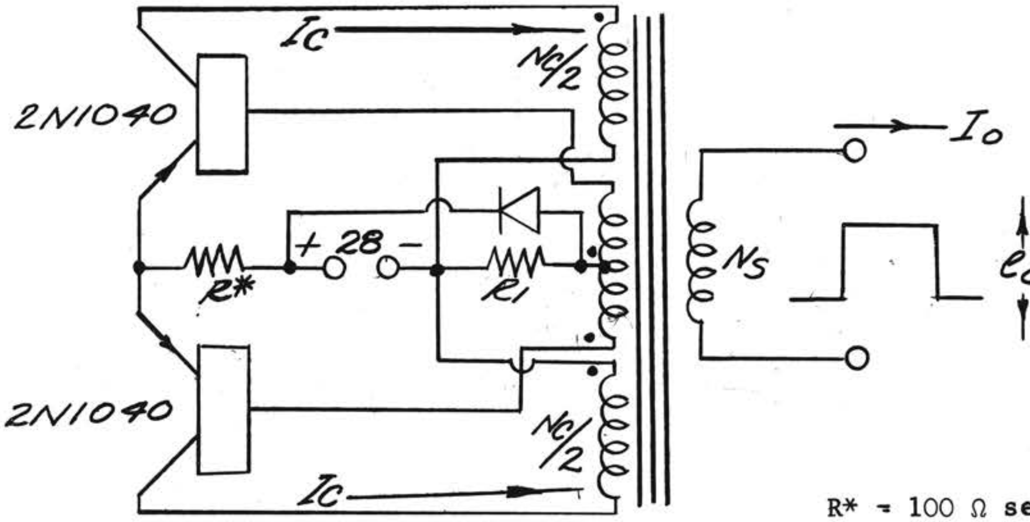
$$r = \frac{\sqrt{2}}{12\omega^2 LC}$$

$$r = \frac{(1.414)}{(12)(4)(\pi)^2\ (10^6)(.075)(33 \times 10^{-6})}$$

$$r = \frac{1.414}{1175} = 0.0012$$

$$\therefore\ e = (0.0012)(6.8) = 8.0 \text{ mv.}$$

Transformer Design for TSR:




$R^* = 100\ \Omega$ sensitor
$R_1 = 220$ kilohms

Fig. A-13. Transistor Saturable Reactor

Load Requirements:

$I_o = 10$ ma.

$e_o = E_{rect.} + E_{filt.} + E_{load} + 5\%$

$e_o = (2)(0.3) + (2)(10)(44)(10^{-3}) + 13.6 + 5\%$

$e_o = 15.85$ volts

Choose: Freq. - 1000 cps.

Two 6T8043-51 Cores
stacked together

Ref. (5):

$B_s = 6{,}500$ Gauss

$A = 0.0177\ \text{in.}^2$

From System Specifications:

$$E = 28 \text{ volts}$$

Ref. (4):

$$N_c = \frac{E \times 10^8}{12.9 \times f \times A \times B_s}$$

$$N_c = \frac{(28)(10^8)}{(12.9)(1000)(0.0117)(6500)}$$

$$= 2860 \text{ Turns}$$

$$N_c/2 = 1430 \text{ Turns}$$

$$N_b = \frac{N_c \times 2V_b}{2E}$$

$$I_c = \frac{I_o \times e_o \times 1.15}{E}$$

$$I_c = \frac{(10)(15.25)(1.15)}{28} = 6.25 \text{ ma.}$$

$$\therefore I_b = \frac{6.25}{20} = 0.312 \text{ ma. for } h_{FE} = 20$$

$$V_{BE} = 0.3 \text{ volts for } 0 < I_b < 10 \text{ ma.}$$

$$V_f = (\text{diode}) = 0.3^V$$

$$V\ (R^* \text{ at } 6 \text{ ma.}) = 0.6^V$$

$$\therefore V_b = (0.3) - (0.3) + (0.6) = 0.6 \text{ volts}$$

$$N_b = \frac{N_c \times 2V_b}{2E} = \frac{(2860)(1.2)}{56}$$

$$N_b = 62 \text{ Turns}$$

$$N_s = \frac{N_c \times e_o}{2E}$$

$$N_s = \frac{(2860)(15.8)}{(2)(28)}$$

$$N_s = 810 \text{ Turns}$$

Power Supply Evaluation:

Measured Parameters (sec. connected)

| | |
|---|---|
| $R_s = 93$ ohms | $I_{in} = 7.0$ ma. |
| $R_c = 180$ ohms | $I_{out} = 9.5$ ma. |
| $R_{choke} = 52$ ohms | $I_{bias} = 0.12$ ma. |
| $E_{in} = 28$ volts | $\Delta e_o = 0.005$ volts |
| $E_{out} = 13.9$ volts | freq. = 1 kc. |

(Secondary open)

| | |
|---|---|
| $E_{in} = 28$ volts | $I_{bias} = 0.84$ ma. |
| $I_{in} = 1.0$ ma. | |

Losses

| | |
|---|---|
| *Core = (28)(1) | = 28.0 mw |
| Sensistor = $(7.0)^2(100)$ | = 4.9 mw |
| Pri.Cop. = $(7.0)^2(180)$ | = 9.0 mw |
| Sec.Cop. = $(9.5)^2(93)$ | = 8.5 mw |
| Rect. = (9.5)(0.3) | = 2.85 mw |
| Choke = $(9.5)^2(52)$ | = 4.70 mw |
| Bias = (0.12)(28) | = 3.35 mw |
| Total | = 61.3 mw |

*Includes switching and base drive.

| | |
|---|---|
| $P_{in}$ = (28)(7.0) | = 196 mw |
| $P_{out}$ = (13.9)(9.5) | = 132 mw |
| Difference | 64 mw |

Overall efficiency = 132/196 = 67.5%

$$r = \frac{\Delta e_o}{E_o} = .005/6.8 \times 100 = 0.075\%$$

VITA

Dave C. Mueller

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Master of Science

Thesis: A TRANSISTORIZED DIGITAL COMPUTER WITH BOTH REAL AND STORED TIME ANALOG READOUT OF INFORMATION - FOR USE IN DEEP SPACE INVESTIGATIONS OF MICROMETEOR PHENOMENA

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