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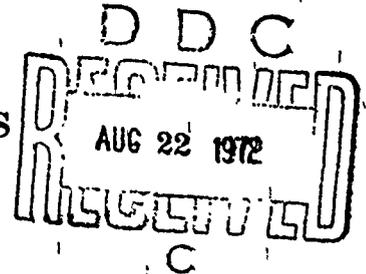
RESEARCH AND DEVELOPMENT  
TECHNICAL REPORT ECOM-0101-F

**ELECTRONICALLY TUNED  
ANTENNA COUPLER**

FINAL REPORT

BY  
C. M. MORRIS

JULY 1972



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This is the Final Report on a program to develop an Exploratory Developmental Model of an electronically tuned, automatic antenna matching unit for use with manpack radio equipment in the 30-to-80 MHz frequency range. This report describes the results of the entire program period, including initial study efforts in the areas of tuning elements, sensors, and networks. The performance of the final deliverable model is discussed, along with suggestions for improvement. Test data on this model, a 10-watt, 30-to-80 MHz coupler using Avco's piezoelectric capacitor, is presented to indicate tuning capability, insertion loss, and tuning speed into the required variable one to four-foot whip antenna. An analysis of the work performed during the study and development phases is shown. Problem areas that appeared during these phases are mentioned, and solutions and design approaches are discussed. Finally, a technical discussion of a 2-to-30 MHz automatic antenna coupler is presented. Major requirements, such as a higher power level (10 watts) and antenna loads (9-foot whip - 2 to 4 MHz, 6-foot whip - 4 to 30 MHz, 10:1 VSWR referenced to 50 ohms) are analyzed. Different approaches to the problem are discussed in detail and trade-offs mentioned. A design example of a completely automatic 2 to 30 MHz antenna coupler compliant with the requirements is given.

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Cincinnati, Ohio

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## 1.0 INTRODUCTION

This program covered the design, development, and fabrication of an Exploratory Development Model of an electronically tuned, automatic antenna matching network for future application in manpack radio sets. This matching network was designed to handle 10 watts of RF power and cover the frequency range of 30 - 80 MHz. It should efficiently match the specified range of manpack antenna impedances to the desired transmitter load impedance, while minimizing control power consumption and tuning time. It was further desired that the network and its component parts be optimized with respect to size, weight, mechanical complexity, and thermal stability. The coupler should feature optimum size for packset compatibility and therefore represent, as closely as possible, a product that can be easily projected for use in future radio equipments.

During the course of the program, techniques were investigated which pertained to electrically tuned matching networks in the 2 - 30 MHz frequency range at power levels up to 40 watts. Component and network specifications necessary for a complete 2 - 30 MHz electronically tuned matching network have been compiled in order to demonstrate feasibility of an HF high power coupler utilizing the most up-to-date concepts developed during the course of the program.

This report covers the complete program period from the initial investigative efforts in the areas of tuning elements, sensors, and networks through the development and final test of the deliverable 30 - 80 MHz Coupler System. Final test data on the 30 - 80 MHz Coupler is shown and discussed, and significant accomplishments and benefits are delineated.

A detailed analysis of the work during the study and development phase is presented. Problem areas that appeared during the development are pointed out, and solutions and design approaches taken are discussed.

Finally a description of a 2 - 30 MHz Coupler is presented. Different approaches to the problem are discussed and trade-offs mentioned. A design example of a 2 - 30 MHz Coupler operating into 6 and 9-foot whip antennas at the 40-watt level is shown.

As a result of this contract Avco Electronics Division has designed and developed an electronically tuned antenna coupler which, with some additional development, can be far superior to any comparable system presently known. This coupler features many revolutionary new techniques and may ultimately provide the Army with the lightest, most compact antenna coupler ever developed for manpack communications equipment. This system was designed specifically with Army user requirements in mind, and offers the following benefits.

- Automatic matching capability for any whip antenna from one to four feet over the 30 - 80 MHz frequency range.
- Compatibility with any type manpack battery. Previous couplers have been limited to use with batteries capable of high current outputs, thus making them unusable with the standard PRC-25 or PRC-7 carbon-zinc, or magnesium batteries.
- Ultra-linearity for compatibility with any desired transmission mode.
- Low insertion loss and increased matching accuracy, allowing greater power delivered to the antenna.

- Operation insensitive to battery voltage variations.
- High power handling capability for a given size.

These benefits are discussed in more detail in a later section. It should also be noted here that the techniques developed by Avco on this program are also applicable to other areas. For example:

- Avco's piezoelectric capacitor provides a versatile electronic tuning element for antenna couplers, wide dynamic range tuners, and pump VFO's anywhere in the frequency range of 2 - 400 MHz.
- Avco's digital VSWR detector replaces four separate sensors in a normal power amplifier-antenna coupler system. It is accurate over a wide range of RF power levels and insensitive to line strays. The reduction in circuit complexity and ease of reproduction provide a sensor system with greater inherent reliability for many automatic tuning applications.
- Avco's new simplified automatic tuning scheme requires only VSWR information to tune. This provides assurance of tuning even extremely badly mismatched loads since high reactance/resistance ratios are no longer a problem.

## 2.0 REQUIREMENTS

The effectiveness of tactical communications can be greatly enhanced by the use of automatic rapid tuning of the antenna matching network. This is especially true for the VHF packset. It has been severely handicapped by lack of automatic antenna matching in the past. Avco has participated in field tests on Army VHF radio equipment which have shown field strength degradation of 6 - 10 db due to changing antenna impedances at a fixed frequency caused by different environmental conditions or operator orientations. The fact that the packset antenna impedance can vary so much at a given frequency that a fixed tuned network permits a degradation of this magnitude is just one of the problems that Avco seeks to correct in its development of an all-electronically tuned antenna coupler system.

An effective and successful development of the electronically tuned coupler will, of necessity, be consistent with the primary objectives of the Army as defined in the development specification.

Emphasis was placed on the development of an exploratory development model unit of 30 - 80 MHz which is compatible with advanced miniaturized packset radio equipment. Techniques discovered during this development that were applicable to a 2 - 30 MHz coupler were noted and analyzed toward their utilization in the extended range system. Most of the requirements for this latter coupler concern parameters that must also be considered for the 30 - 80 MHz system. The major differences, outside of the obvious ones imposed by the extended frequency range, are a greater variety of antennas, linearity, and higher RF power capability. These parameters, as well as those listed below, were investigated in the light of present day component technology and a complete design specification developed for the extended range coupler. The most important of these requirements are:

- Accommodation of whip antenna impedance
- Linearity for SSB operation
- High matching efficiency
- Tuning memory
- Matching speed
- Low power consumption
- Matching accuracy
- Matching sensitivity
- RF power levels
- Bandswitching
- Stability versus temperature and RF power

## 2.1 WHIP ANTENNA IMPEDANCE

The matching network configuration is determined largely by the antenna impedances which must be accommodated over the operating range. These impedances are probably better known by Avco than by others, both because of the company's extensive research and development efforts in antenna design and measurements, and also due to measurement programs conducted at Avco Electronics Division on packset whip antennas. One such program was an experimental investigation carried out under an Army Contract, DAAB07-67-C-0263, to determine the range of packset antenna impedances under actual operating conditions. Techniques were developed to measure the packset whip impedance in its normal field environment and without the influence of external measuring equipment and interconnecting cables. Primary emphasis in the study was on three and four-foot whip antenna characteristics in the 30 - 76 MHz range. A detailed discussion of the results of this investigation, as well as other pertinent antenna characteristics, is given in Appendix A of this report. Figures are shown which illustrate the ranges of antenna resistance and reactance which likely will be exhibited by the specified antennas over their required operating frequency range. The greatest problem of matching these impedances occur at the low end of the 2 - 80 MHz range because the antenna has an extremely high reactance and relatively low resistance (i.e., high Q) in this region. This characteristic is also true for a one-foot antenna in the low end of the VHF range. Plots of the impedance characteristics of the one-foot whip and the high and low Q extremes of the three and four-foot whips are shown in figures 1, 2, and 3. The three and four-foot whip plots show actual field test data and are representative of the high and low Q impedance extremes that these whips present under different environmental conditions and operator orientation.

## 2.2 LINEARITY

Linearity of the antenna coupler is an important consideration, even in couplers intended only for FM radio set usage. Harmonics and spurious products should not be generated by the coupler network; therefore, the tuning components used should possess good linear characteristics. As shown in previous work (Technical Report ECOM-0101-1), the linearity of a saturable reactor at a given power level is directly proportional to the volume of the ferrite used. Therefore, a definite trade-off can be established between coupler size and weight and system linearity in a matching network utilizing saturable reactors. The piezoelectric capacitor, since it is basically a mechanical device, is inherently linear and therefore has a distinct advantage in systems requiring high linearity.

# IMMITTANCE CHART

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ADMITTANCE COORDINATES — 20 MILLIMHO CHARACTERISTIC ADMITTANCE

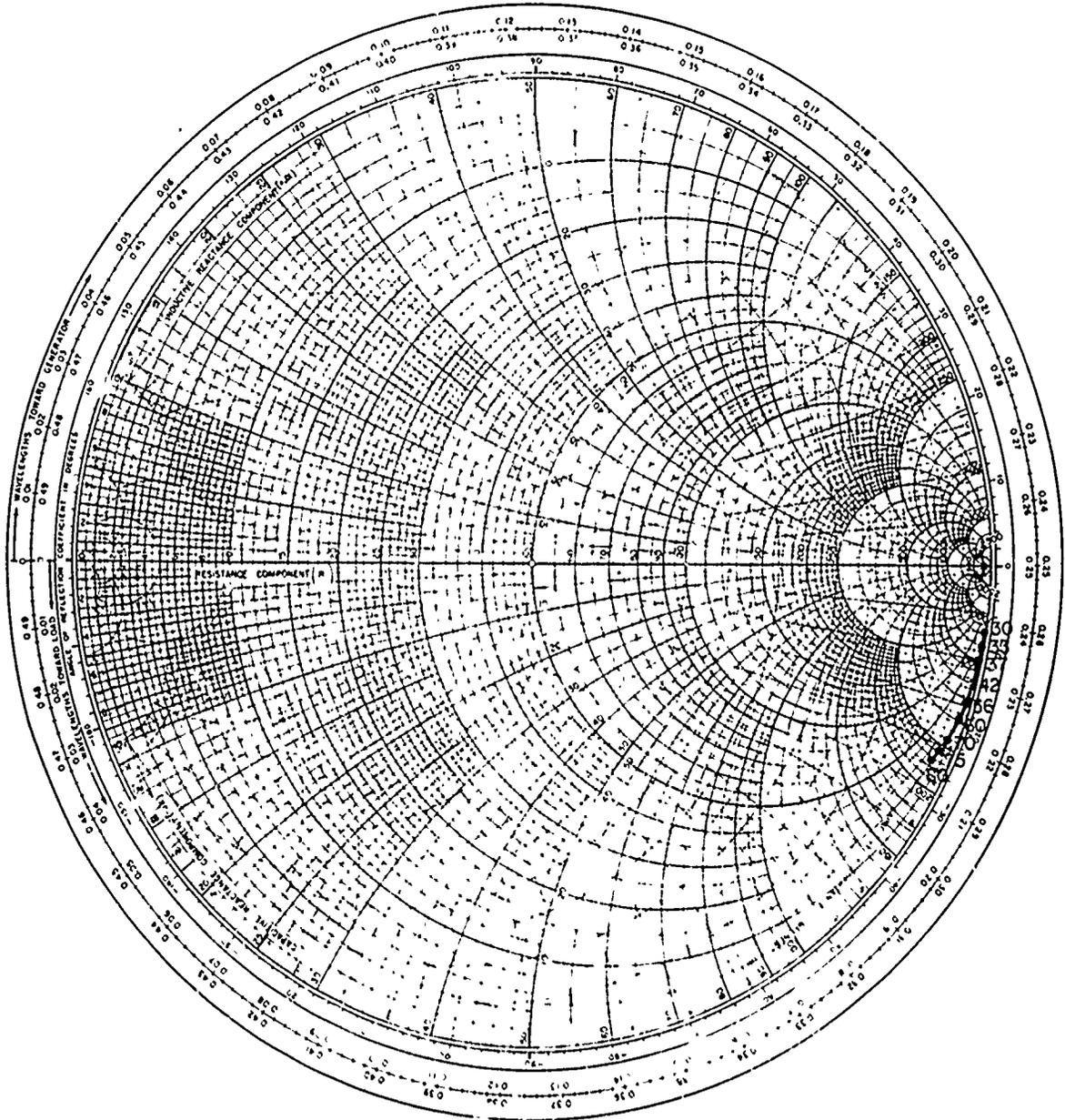


Figure 1. One-Foot Whip Antenna Impedance Plot

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IMPEDANCE COORDINATES — 50 OHM CHARACTERISTIC IMPEDANCE

ADMITTANCE COORDINATES — 20 MILLIMHO CHARACTERISTIC ADMITTANCE

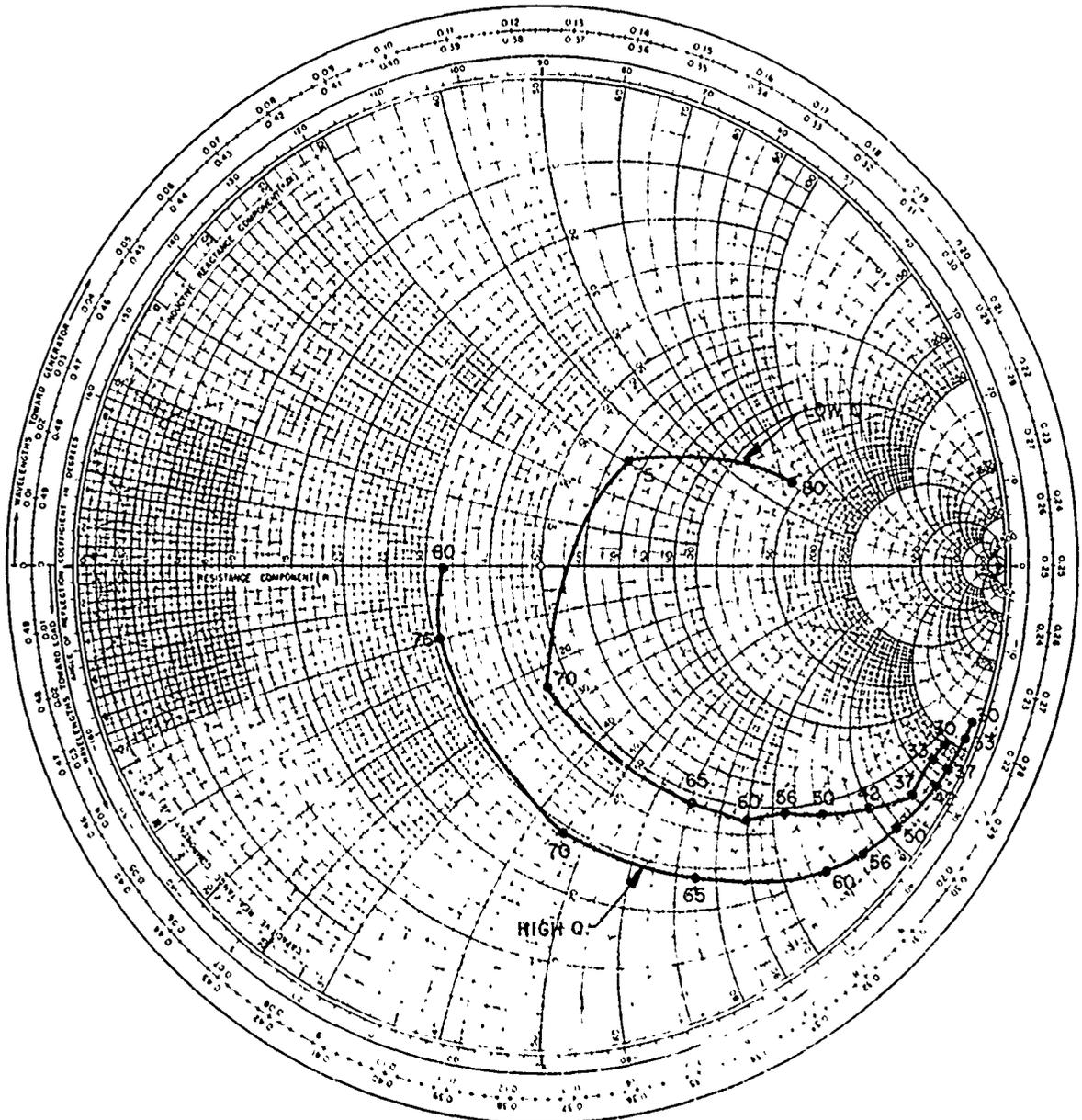


Figure 2. Three-Foot Whip Antenna Impedance Plot



The 2 - 30 MHz Coupler can use linear passive elements such as fixed powdered iron core inductors and capacitors where electronically tunable elements are not available. If saturable reactors are used in each arm of the network, their circulating power-handling requirements must be carefully determined by computer analysis (for a 40-watt system using all the antenna impedances specified in the 2 - 30 MHz region). Using this analysis, the ferrite volume is calculated and the reactors designed with the required linearity specifications in mind. Test results indicate that, as long as the reactors are operated below their calculated power-handling capability, the predicted linearity is achieved. At the low frequencies, where large inductances are needed to resonate the reactances of the very high-Q antennas, the saturable reactors represent only a small percentage of the total inductive reactance; thus, their effect on overall linearity is negligible. At higher frequencies, the antenna impedances are better, less inductance is required, and circulating power-handling requirements are considerably lower. Therefore, smaller volume ferrite cores can be used to achieve the same linear behavior.

### 2.3 MATCHING EFFICIENCY

The network configuration is selected with minimization of component power losses in mind. To this end, computer programs have been set up and run which enable the prediction of insertion loss in each component of the network when matched into the various required load impedances. Network component values have been selected to provide tuning sequences which minimize the inductance required to match given loads. The more efficient a coupler is, the less size and weight it requires, since (especially in the case of ferrite material) both power-handling capability and linearity are proportional to volume of the material.

For the piezoelectric capacitor, power-handling capability is a function of the voltage breakdown. Matching efficiencies can be extremely good because the unloaded Q of this component is very high. The limiting factor here is mechanical mounting configuration and load length strays.

### 2.4 TUNING MEMORY

Tuning memory of at least limited duration is necessary for compatibility with SSB operation and manual or automatic CW operation when the carrier is interrupted. What is required is a memory system for couplers which will provide an indefinite memory capable of maintaining the state of tuning during receive operation, as well as during the above conditions. Once the matching network is tuned, it will remain in this state until a new frequency of operation is selected or the resultant VSWR rises above a predetermined value due to an antenna impedance change. Techniques to accomplish this memory function have been developed by Avco and have been utilized with success in two contracted programs. One method uses a permanent magnet in conjunction with a saturable reactor, the other uses a pilot capacitor control loop together with the piezoelectric capacitor.

### 2.5 MATCHING SPEED

One of the major objectives in the development of an all-electronically tuned antenna coupler is to achieve a tuning speed less than that obtainable with typical analog type couplers with servo-motor tuned inductors and capacitors. These systems should accomplish very rapid tune-ups for the automatic coupler using piezoelectric capacitors, the all-static coupler using saturable reactors, and the combination binary-static coupler which employs a programmer and logic system to switch binary-related coils in addition to saturable reactors.

or piezoelectric capacitors. Tuning speeds for these types of automatic couplers depend greatly on the variety of different antennas and their impedance variations to be matched at any given frequency. This, together with size, weight, and RF power requirements, determines the number (if any) of binary-related coils that must be used to complement the variable tuning elements in the coupler network configuration. Thus, relay switching times and logical tuning sequence, in addition to variable element tuning speed, become important considerations in determining overall coupler speed. Tuning speeds for saturable reactors depend upon the time constant of the control winding and associated circuitry, and the number of tuning pulses necessary. Tuning speeds for the piezoelectric capacitor depend on the digital increments used and the damped time response of the material.

## 2.6 POWER CONSUMPTION

Low power consumption can readily be achieved in antenna matching systems and is a by-product of the techniques used for network tuning. The permanent magnet approach used in the saturable reactor assemblies offers very low DC power consumption (in addition to the previously described memory feature). The reactor control elements are pulsed stepwise to either a higher or lower inductance using a permanent magnet to maintain the desired state bias flux in the reactor. No holding power is required. The relays employed in the 2 - 30 MHz Coupler are the latching type which are momentarily pulsed for operation. Again, no holding power is required to maintain the tuning state. The result of this technique is a network that remains tuned in the "receive" mode without drawing any DC power, and also requires very little actual tuning power because of utilization of the pulsing method. Therefore, a breakdown of coupler operation into receive, transmit, and tune modes shows that no DC power is required in either receive or transmit operation and that, during the short tune mode, power is drawn only in pulsed intervals.

The pilot capacitor control loop approach does require some "holding" power after the tune cycle, but this can be on the order of only 10 - 30 milliwatts. On the other hand, this approach does not require large current pulses during the tune cycle and is therefore compatible with any battery type power source.

## 2.7 MATCHING ACCURACY AND SENSITIVITY

To achieve high tuning accuracy and good sensitivity, careful attention must be paid to the selection and optimization of the RF sensors required for the couplers. The more sensitive a system is, the greater the range of tuning powers it can operate with and the wider variety of load impedances it can react properly to. In other words, in the case of a resistance detector, sensitivity is the ability to provide a recognizable output indicating an  $R_0$ -circle crossing even when the impedance has very high reactance-to-resistance ratios. This has an important bearing on both system tune time and network component selection, since a sensitive detector eliminates the need to avoid certain areas of the impedance plane during the tuning sequence. Accuracy is also an important parameter for two major reasons. First, the greater the accuracy of the detector, the fewer iterative processes will have to be used in the tune cycle, thus shortening the tune time. Secondly, the better the match that is ultimately achieved, the greater the power that will be transmitted.

## 2.8 STABILITY

Since usefulness of the electronic tuning concept depends upon having good stability, careful consideration is given to the effects of temperature and power on component characteristics. The major concern with the piezoelectric capacitor is susceptibility to external vibration and resonance modes inherent in the particular unit. A control loop must be employed to compensate for these problems. The complexity of this loop is proportional to the frequency of the vibration that must be compensated for. Avco's pilot capacitor control method has been shown to be an effective compensation for any temperature change or vibrational perturbation requirement for packset equipments.

Ferrites, on the other hand, are known to have properties which vary with temperature. Thus appreciable effort has been devoted to a study of ferrite properties and methods of stabilization. The two major sources of concern here are:

1. Ambient temperature variations
2. Self-heating of the component due to RF power losses

The ferrite type selected should be one that can maintain its permeability and  $Q$  throughout a reasonable ambient temperature range. A ferrite which suffers a sharp decrease in  $Q$  as temperature rises is not satisfactory since it is susceptible to thermal runaway. The ferrite used in the couplers has been thoroughly tested over the specified temperature range and, while it does exhibit  $Q$  and permeability variations, the change is not radical and operation at any point in this range would therefore be stable.

Self-heating of the ferrite due to RF power losses presents a more serious problem. If the ferrite is not capable of handling the power in a given situation, it will heat up to the point where the inductance of the element will change enough to cause detuning of the coupler. The seriousness of this consideration cannot be over-emphasized. If the matching network is tuned up to the load (especially to a very high- $Q$  antenna), it takes only a very small change in inductance in one of the tuning elements to cause a large variation in input VSWR. Computer analysis has verified this by predicting the change in input VSWR for different  $\Delta L$ 's in the tuning elements for various antenna load impedances. With this information, it is possible to predict the maximum  $\Delta L$  that can be tolerated before more than negligible detuning occurs. This data, together with the ferrite temperature characteristics, makes possible a realistically stable design for the saturable reactor assemblies. RF power levels of 50 watts have been handled readily in several applications with very high- $Q$  loads with no detuning effect.

## 2.9 MECHANICAL

The achievement of small size and weight is a desirable goal and was given considerable attention during the program. Since much present development effort is directed at miniaturized packaging of military equipment, it is not acceptable to utilize an antenna coupler network in a radio that occupies a large percentage of the available volume. The experience with the static coupler has demonstrated the fact that the 30 - 80 MHz Coupler can be very small and easily adapted to light weight and compact sets. Emphasis was on optimizing size and weight for the RF network portion of this coupler.

The logic portion of the coupler easily fits on one card, if hybridized, or it can be made considerably smaller if an LSI approach is used. The CMOS approach to LSI would be very small and also be optimum for power drain considerations.

### 3.0 ACCOMPLISHMENTS

#### 3.1 30 – 80 MHZ AUTOMATIC COUPLER DESCRIPTION

A block diagram of the 30 – 80 MHz Automatic Antenna Coupler developed on this program is shown in figure 4. The deliverable unit is configured in a metal case with the following connections:

- |                          |                           |
|--------------------------|---------------------------|
| ● RF Input               | — Mini-coax connector     |
| ● RF Output              | — BNC connector           |
| ● Power Supplies         | — 9-pin plug-in connector |
| ● Bandswitch, tune lines | — 9-pin plug-in connector |

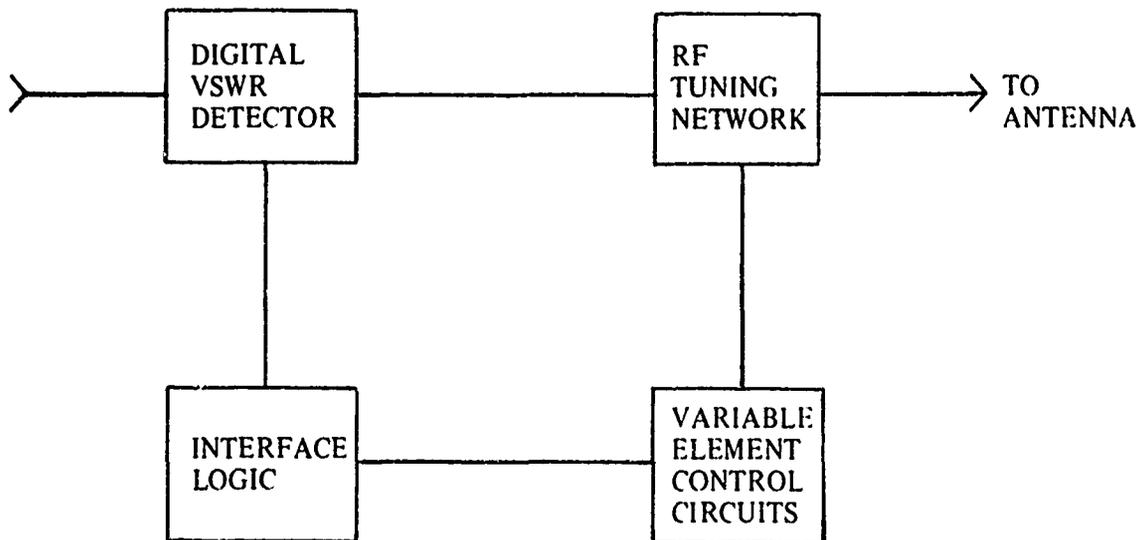


Figure 4. 30 – 80 MHz Coupler Block Diagram

Avco's Automatic Antenna Coupler is a self-contained unit consisting of the following subsystems:

- RF Tuning Network
- Digital VSWR Detector
- Variable Element Control Circuits
- Interface Logic

A brief description of each of these subsystems will best illustrate the principle of operation of the coupler system. A more complete analysis is given in a later section of this report.

##### 3.1.1 RF Tuning Network

The RF Tuning Network consists of two piezoelectric capacitors, the fixed inductors, and the switches necessary for the fixed inductors. The variable capacitors are still the basic cantilever configuration as described in Technical Report ECOM-0101-2. The fixed

inductors are high-Q powdered iron core coils used both in the series mid-section of the  $\pi$ -network and as loading coils on the output. The switches used in the network are TO-5 latching relays. A schematic diagram of this network is shown in figure 5. The network is designed to have the capability to match a variable one to four-foot whip antenna anywhere in the 30 - 80 MHz range. It was tested with simulated antenna loads representative of the impedance values previously shown in figures 1, 2, and 3.

As can be seen from the schematic diagram, the network is basically a  $\pi$ -L type, with the switched inductors in the output arm used as loading coils to resonate the high capacitive reactance associated with the short whips. Each piezoelectric capacitor is really two capacitors mounted back-to-back on the same substrate; this arrangement enables achievement of the required 9:1 tuning range in each of the shunt legs of the network.

The network, as it now stands, is essentially the same as that used to obtain the breadboard data shown in a later section of this report. The only changes made in the network since then are minor mechanical changes made for the sake of mechanical stability and small changes in the fixed coils (inductance changes made by varying the turns on some coils) necessary to eliminate tuning holes which showed up in the breadboard tests.

### 3.1.2 Digital VSWR Detector

The Digital VSWR Detector Module contains the forward and reflected RF power sensors, comparators, and digital encoding circuitry. It provides digital VSWR information to the coupler logic during the tune cycle.

The final module of the Digital VSWR Detector is fabricated on a 3 x 4-inch printed circuit board along with both of the capacitor control loops. The stripline directional coupler is mounted on a separate board directly above it. A block diagram of this detector is shown in figure 6.

The Digital VSWR Detector is the basic control unit in the closed-loop automatic tuning scheme. It samples the forward and reflected RF power at the coupler input, transforms this information into DC signals proportional to forward and reflected power, compares these two signals to get a true indication of the reflection coefficient, and then encodes this in a digital word for processing by the logic system.

### 3.1.3 Variable Element Control Circuits

Two capacitor control loops are utilized as a "memory" for the piezoelectric capacitors and are designed to compensate for capacitor variations due to material drift and environmental conditions. The two loops were fabricated and then aligned with their individual capacitors. These circuits are on the same 3 x 4-inch printed circuit board as the Digital VSWR Detector.

A block diagram of the capacitor control loops is shown in figure 7. The basic concept of operation is that of a closed-loop control system. A small capacitor (pilot capacitor) is mounted on the same plate as the main RF capacitor. This pilot capacitor forms part of a voltage divider which is fed from a low-level 1 MHz oscillator. This "pilot" voltage is detected and used to provide a one-to-one relationship with the main RF capacitor. This value is then used to set the final high-voltage operational amplifier drive level. At the end of the tune cycle this value is fixed proportional to the value of capacitance required to tune. Any deviation due to drift is then automatically compensated for by the generation of an error voltage at the comparator output, which is then used to adjust the main capacitor accordingly.

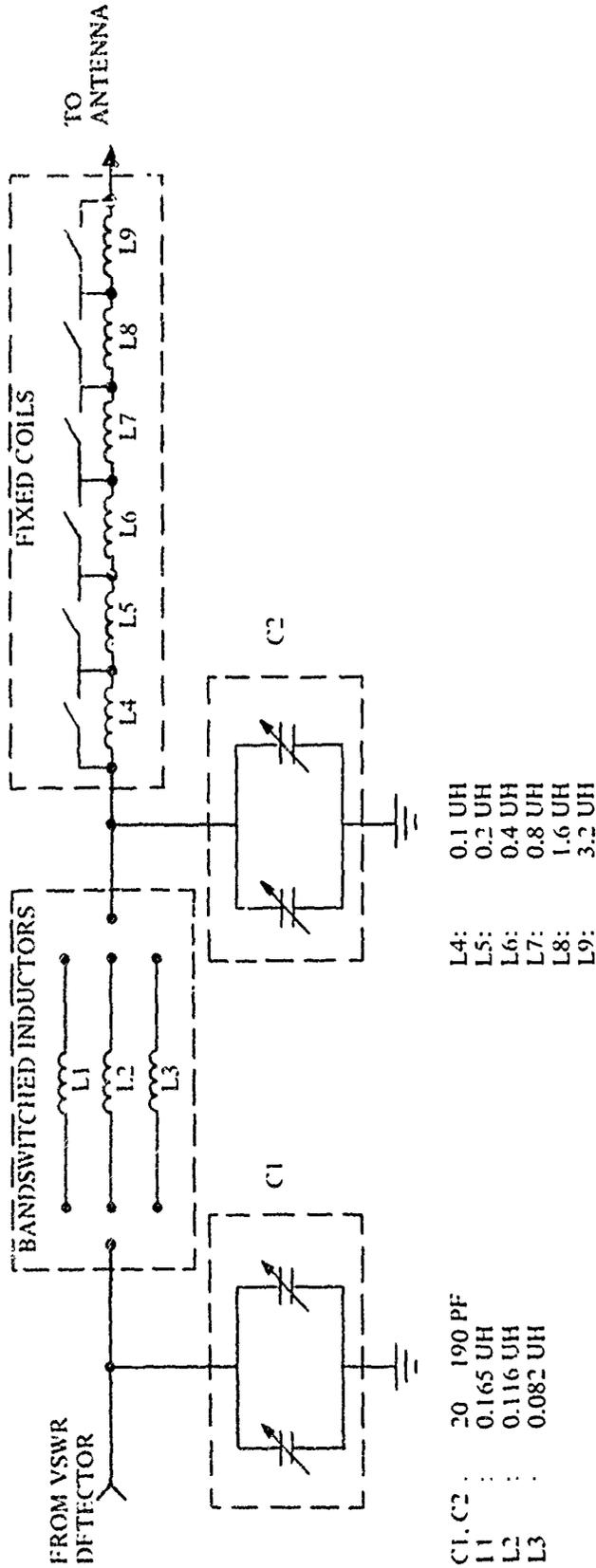


Figure 5. Tuning Network Schematic Diagram

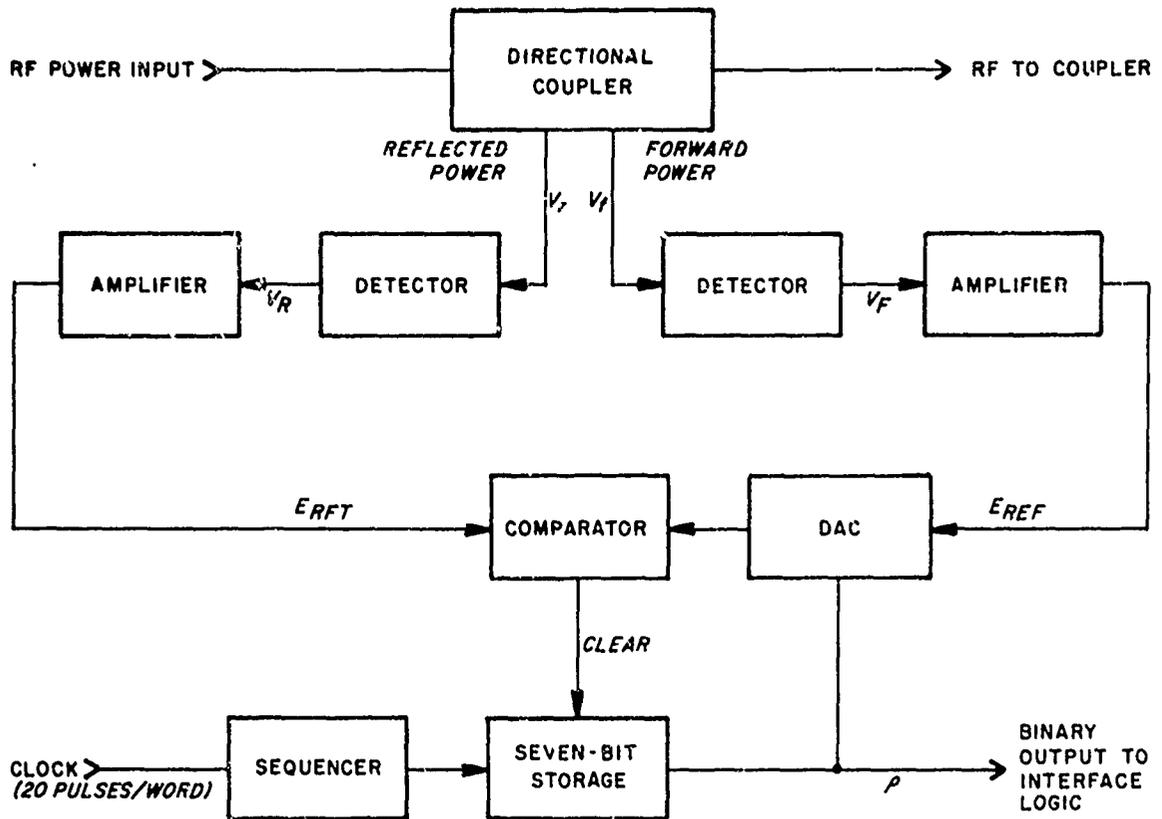


Figure 6. VSWR Detector Block Diagram

### 3.1.4 Interface Logic

The coupler logic system contains control logic for the piezoelectric capacitors, the switched fixed inductors, and interface logic with the outside world. In its breadboard form, the entire logic system was composed of approximately 120 discrete integrated circuits, 17 of which were MSI functions. After consideration of the alternatives, it was decided to fabricate the entire logic system in hybrid form. The discrete IC's were purchased in chip form and mounted on 6 substrates which were, in turn, mounted on one printed circuit board. It was also decided to have the six individual logic substrates and the system printed circuit board laid out by computer by Electronic Graphics, Inc., of Dallas, Texas. The discrete IC chips purchased were beam lead devices wherever possible.

The final logic system then consists of 126 integrated circuit chips (TTL, 5400 series) mounted on 6 substrates. The substrates are in turn mounted on a 3 x 4-inch board in the coupler package. All of the bonding and mounting work was accomplished in Avco's Microcircuit Laboratory.

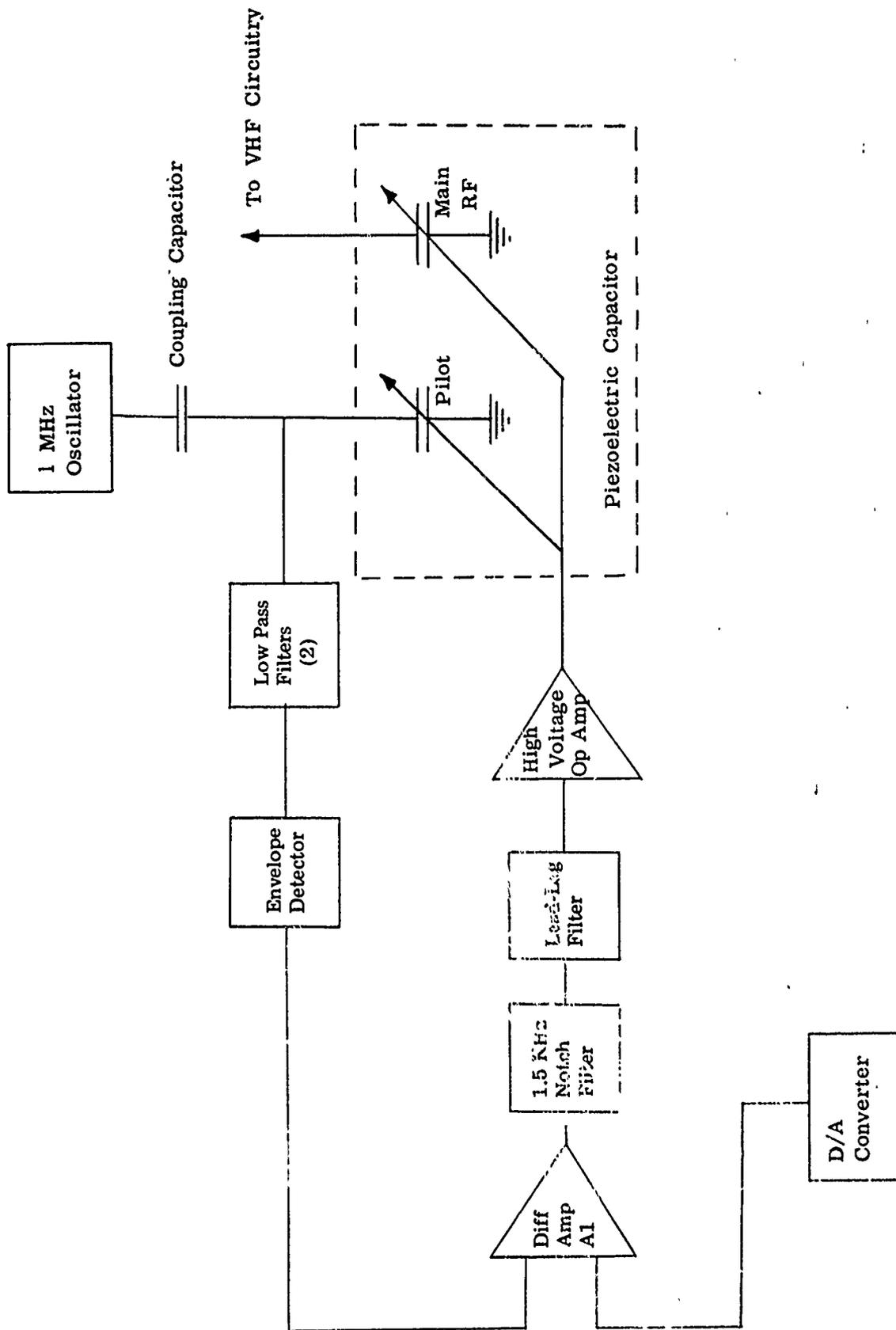


Figure 7. Piezoelectric Capacitor Control Loop Block Diagram

### 3.2 COUPLER SYSTEM MECHANICAL CONFIGURATION

The complete system is composed of three separate 3 x 4-inch boards (RF Network, Detector-Control Loops, and Logic) mounted on top of one another and one smaller (1-1/2 x 3-inch) board containing the directional coupler mounted between the RF Network and Detector-Loop boards. The entire system is enclosed in an aluminum case which has the following connections:

1. The RF input is a miniature coaxial connector mounted directly to the directional coupler board.
2. The RF output is a BNC type coaxial connector mounted on the coupler case and hardwired internally to the RF Network board.
3. Power supply lines, ground lines, and tune cycle information lines are hardwired internally to two 18-pin connectors mounted on the coupler case.

The entire Coupler System (including case) is approximately 2 x 3 x 4 inches in size.

### 3.3 COUPLER SYSTEM OPERATION

Coupler operation can be explained in the following fashion:

1. Actuation of the tune switch in the presence of forward RF power initiates the tune cycle. The tune switch was made external to the system for purposes of testing convenience. In an actual system used with a packset radio, tuning can be initiated whenever forward RF power is sensed. It is a simple matter to eliminate the tune switch, since at present it is merely gated with the forward RF power indicator in the logic.
2. The input VSWR (or reflection coefficient) is checked. If this VSWR is high, indicating a non-tuned system, the coupler tune command is given and the logic resets all its storage elements.
3. Tuning proceeds by sensing the input VSWR, comparing it to the previous value, and then taking the appropriate action. The tuning elements are always driven in a direction corresponding to a decreasing VSWR. When a null is reached (indicated by going from decreasing to increasing VSWR), control is transferred to a different variable element. This process is repeated until the VSWR is decreased to a low enough value to be defined as a "tuned condition."
4. The relay-switched output arm loading coils are always sequenced first. Once proper "loading" is achieved, then the piezoelectric capacitors are incremented, starting with the output capacitor, and going back and forth between the output and input capacitors until tuning is achieved. It is never necessary to go back to the output loading coils after their initial sequencing.

### 3.4 COUPLER FEATURES

It was pointed out in the Introduction of this report that there are considerable advantages inherent in Avco's 30 - 80 MHz Automatic Antenna Coupler. A partial listing of these benefits include:

- Automatic matching capability for variable 1 — 4 foot whip
- Manpack battery compatibility
- Linear
- High RF power handling capability
- Insensitivity to battery voltage fluctuations

These benefits are the results of significant technical accomplishments made during the program. These accomplishments are not restricted to systems in the 30 — 80 MHz range, but are applicable with modifications to the HF and UHF range also. The most important program accomplishments include:

- Development of an electronically variable capacitor as a practical tuning element.
- Development of a digital VSWR detector which can replace four separate detectors in a practical amplifier/coupler system.
- Development of a new tuning method which requires only VSWR information to tune.
- Development of a hybrid-package logic system which demonstrates the highest density packaging in the industry.

The discovery and development of the piezoelectric capacitor opened up entirely new vistas for tuning networks. Considerable effort was put into improving and refining the basic cantilever-type capacitor to the point where it could be a realizable tuning element. Although there is still need for improvement, especially in the mechanical mounting and configuration areas, this capacitor offers the following advantages for electronically tuned networks:

- Good linearity
- Low control power
- Small size and light weight
- Good power handling capability
- Low loss

This tuning element together with its control "memory" loop provides increased flexibility in electronic tuning applications. Its small size and ability to handle large powers, low loss characteristics, and ultra-linearity make it ideal for many applications in addition to antenna couplers.

The piezoelectric capacitor is especially well suited for couplers used with portable communications equipment, since very low control currents are needed for tuning, thus making it compatible with any manpack battery. Automatic couplers developed prior to this time all suffered from the problem of DC voltage sensitivity, i.e., successful tuning was dependent on the battery voltage staying at a certain level during operation. This problem was especially acute for operation with the magnesium battery because initial current drains of any magnitude cause a considerable transient drop in DC voltage level. Use of the piezoelectric capacitor eliminates this problem completely, thus providing the Army with a coupler compatible with any battery.

Since it is essentially a mechanical device, the piezoelectric capacitor is extremely linear and independent of power level. Thus, it can be used as an electronic tuning element in any

application requiring wide dynamic range and good linearity. Power handling capability is limited only by the dielectric material; therefore, the capacitor can handle very large RF powers for a small size package.

The digital VSWR detector is the only sensor needed in an amplifier/coupler system. It provides forward and reflected RF power output monitors which can be used in ALC systems or amplifier protective circuits. In addition, the reflection coefficient output is the only coupler tuning sensor required, eliminating the need for detectors such as Resistance, Phase, and Z-Magnitude.

Avco's Digital VSWR Detector features better than 1 percent accuracy over wide RF power levels, is easily reproducible because it can be placed anywhere on the RF line and is not affected by strays, and thus provides a versatile complete sensor system for many automatic tuning applications. Use of this sensor system thus provides a reduction in circuit complexity and therefore a better inherent reliability.

The detector is very accurate, even when sensing VSWR's at high reactance/resistance ratios, since it can always obtain samples of forward and reflected RF power to translate into usable tuning information. Sampling is obtained from a stripline directional coupler, which is easily reproducible, and this information ratioed and transformed in a small, compact digital system to a digital "word" directly proportional to VSWR.

The new tuning scheme featuring the need for only VSWR information to tune provides much greater accuracy and tuning reliability. The variable elements in the network are scanned, searching only for minimum VSWR at the coupler input. Since a positive indication of VSWR is always obtained for any point on the Smith Chart, tuning is assured provided only that the network elements have the necessary matching range. Operators no longer have to worry about an automatic coupler "getting lost" during a tune cycle due to lack of sensor information or other detector inaccuracies.

In addition to the above-mentioned accomplishments relative to the 30 - 80 MHz Coupler, other techniques pertaining to automatic couplers were investigated in the initial phase of the program. Two factors in particular stand out concerning couplers using saturable core reactors as the variable elements. These are:

- Development of a "zone-controlled" iterative tuning scheme for T-network couplers. This technique eliminates the possibility of "getting lost" due to stray input capacity between the network and the detector system. This method is described in detail in Technical Report ECOM-0101-1.
- Development of a current-control method for incrementing the pulses to a saturable core reactor. This is also described in Technical Report ECOM-0101-1 and results in elimination of the supply voltage dependency inherent in the conventional voltage/time control method.

## 2.5 COUPLER PERFORMANCE

Test data was run on the 30 - 80 MHz Coupler with the following results:

- Matching Capability                      1 – 4 foot whip antenna (selected test impedances are shown in figures 1, 2, and 3)
- Matching Accuracy                      See figures 8, 9, 10 and 11
- Insertion Loss                            See figures 12, 13, 14, and 15
- Tuning Time                                260 milliseconds, maximum
- Power Handling                            2 – 10 watts

Figures 1, 2, and 3 show plots of the nominal impedances of the 1, 3, and 4-foot antennas. These impedances were simulated with a variable LC network connected to a 50-ohm resistive load. Input RF power and power into the output 50-ohm load were monitored. Figures 8 through 11 depict the coupler tuning capability by plotting input VSWR after tuning versus frequency for each of the required antennas. Variations in antenna impedances are accounted for by using two values for the 3 and 4-foot whips at each frequency. These two values represent the extremes (high and low Q) of impedances that the antenna should encounter and are based on field test data.

Tuning times were measured by adjusting the external clock frequency (located in the test box) to 2 KHz and timing with a scope how long it took to tune. The clock speed was then adjusted to 20 KHz and tuning capability checked. Theoretical tuning speeds of the coupler are faster than those actually measured; however, due to the fact that the control loop had to be slowed down to compensate for higher-order resonance modes in the capacitors, the speeds were correspondingly slower (this problem is discussed in more detail in a later section).

Reference to the tuning accuracy curves shows that accuracy was not as good as should be expected. This was largely due to three reasons:

1. Reset Problems: A synchronization problem was found in the initial part of the tune cycle. When the tune cycle was initiated, all of the logic systems were not always reset properly. This caused the logic to give an extra pulse or two after the coupler had reached its best tune point, resulting in a final input VSWR which was worse than that actually achievable. Part of this problem seemed to be in obtaining the proper supply voltage turn-on synchronized with the tune button. Another part, however, was due to an IC chip in the logic with a marginal trigger threshold. It was decided not to attempt to attack this problem because of the difficulty of working with the chip-and-wire hybrid logic system.
2. RFI Problems: Undesired RF signals were definitely getting into the capacitor control loops and the VSWR detector, causing erroneous information to be transferred to the logic. This problem was especially severe above 60 MHz. Although considerable decoupling circuitry was added, the problem was never completely solved.
3. VSWR Detector Response: The forward and reflected power detector response curves were not perfectly matched in spots (especially the high band). Unless these response curves are well matched, inaccurate VSWR data will result.

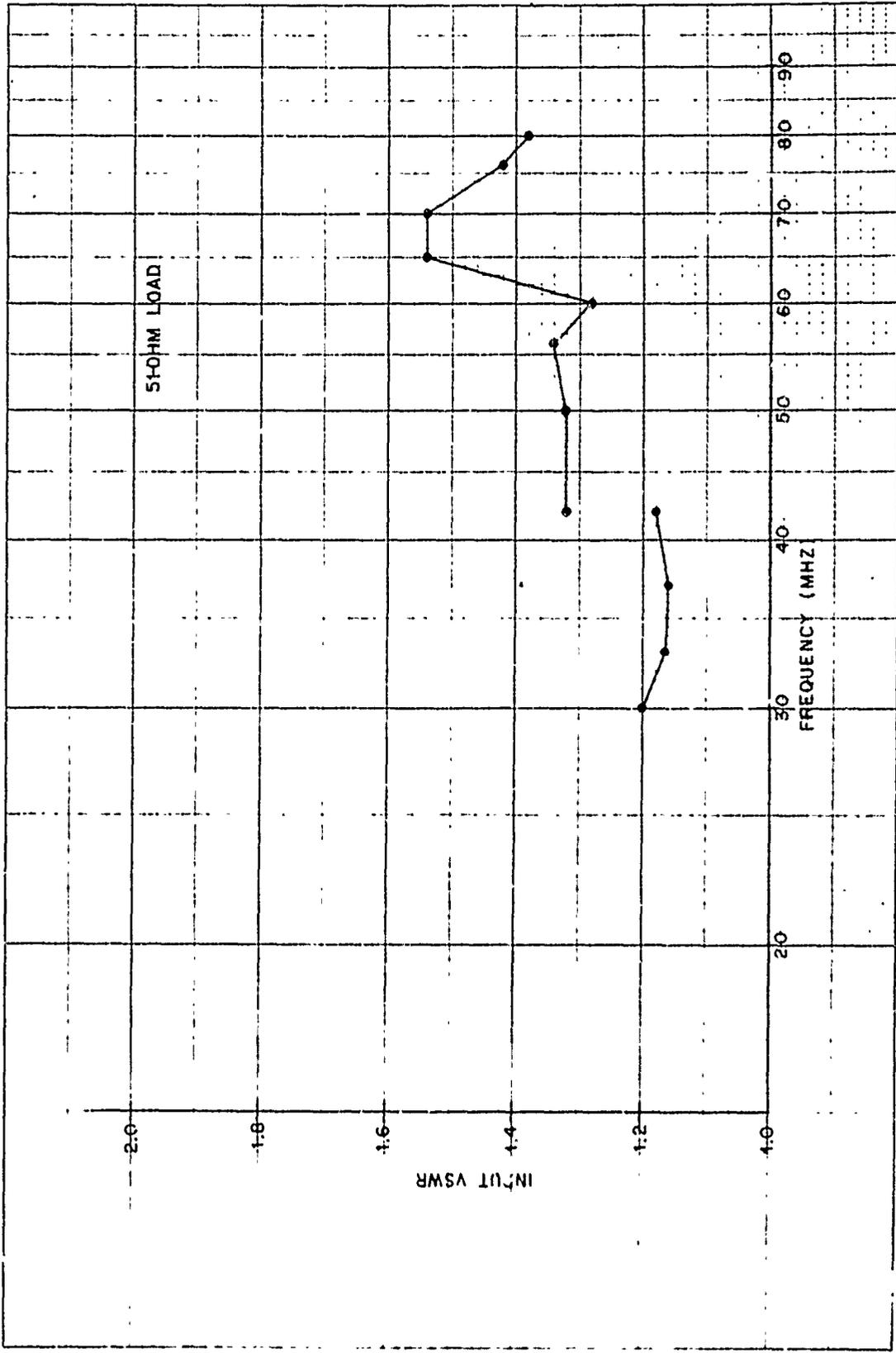


Figure 8. 50-Ohm Matching

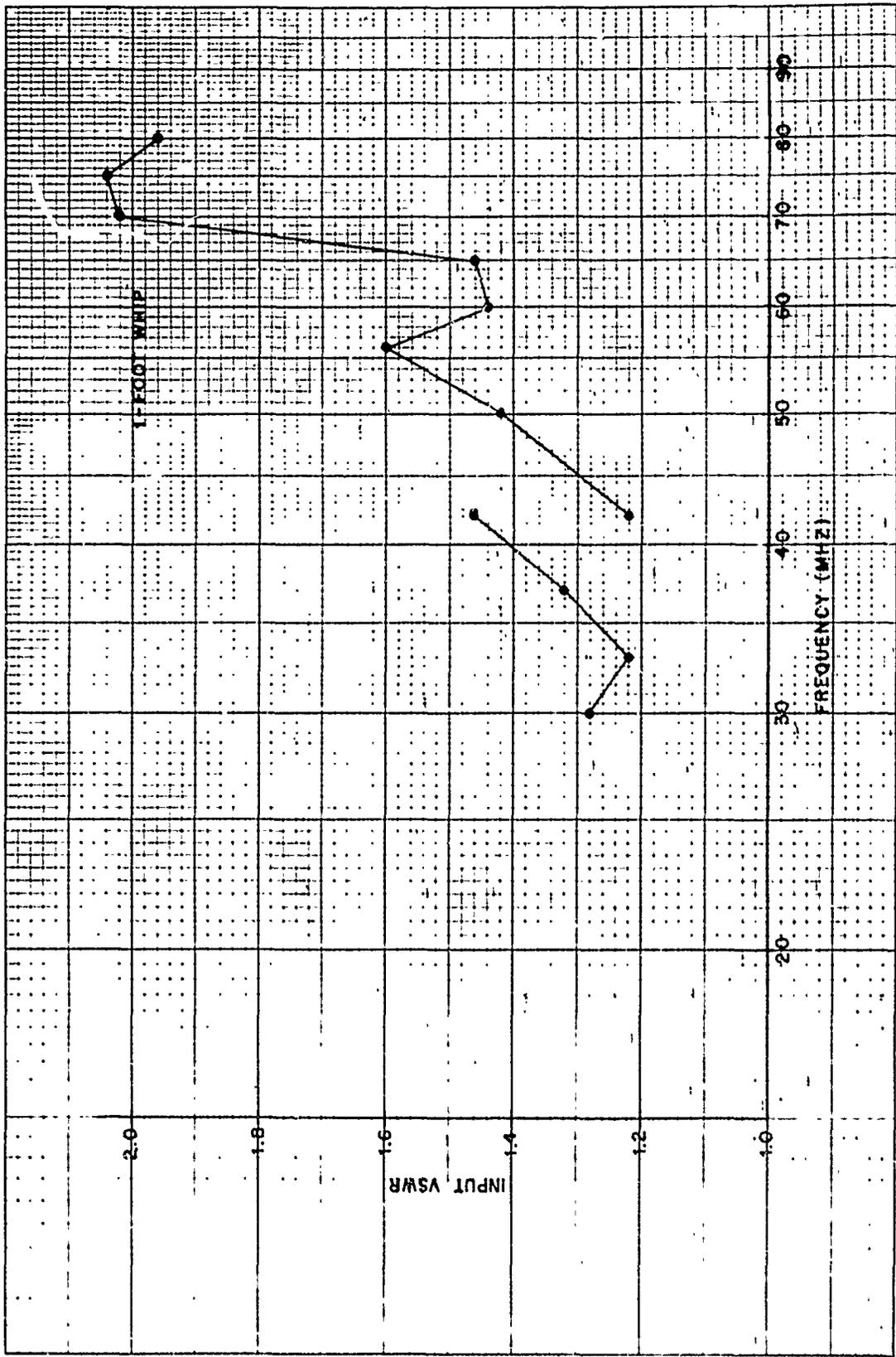


Figure 9. One-Foot Whip Antenna Matching

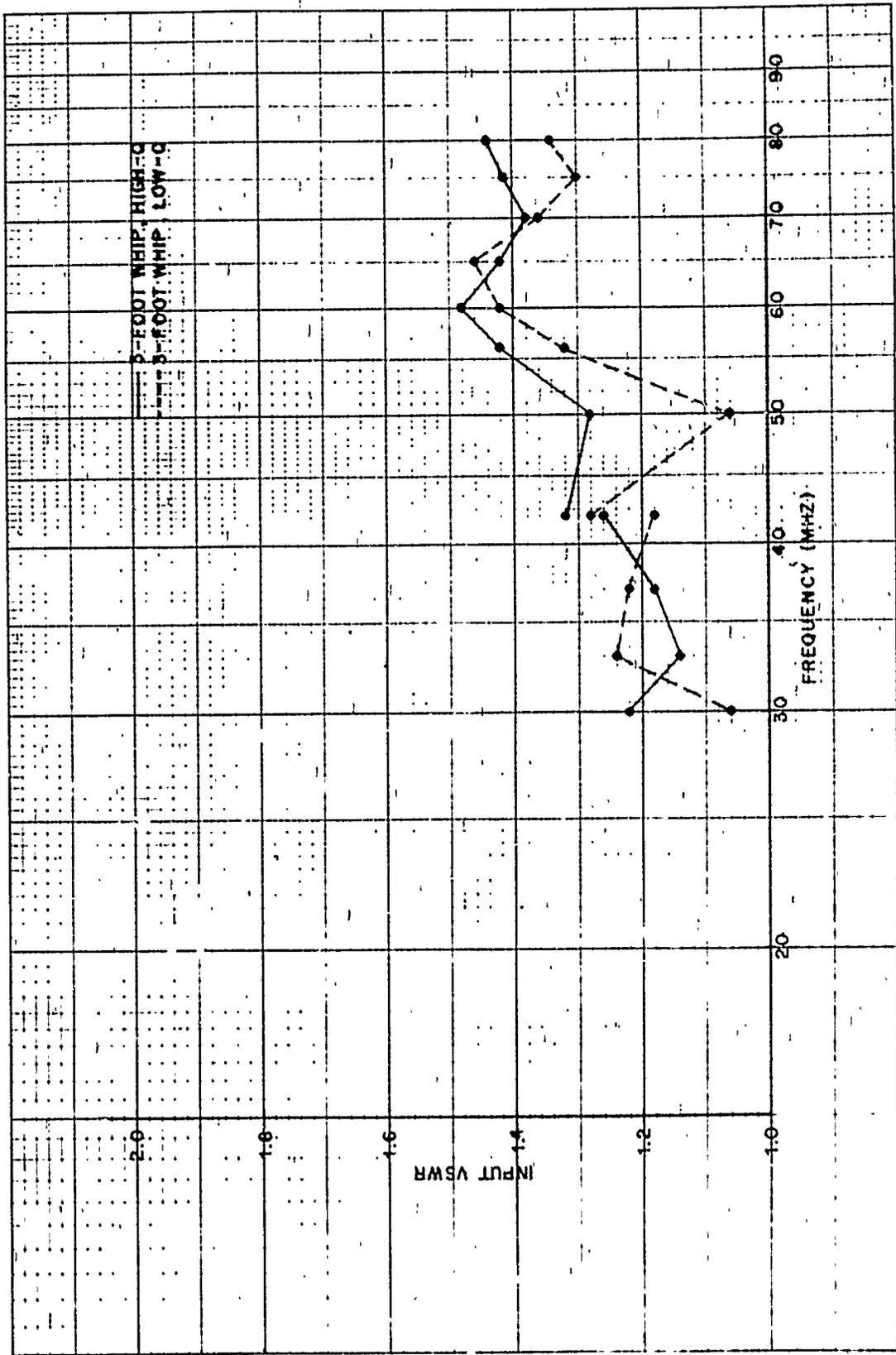


Figure 1C. Three-Foot Whip Antenna Matching

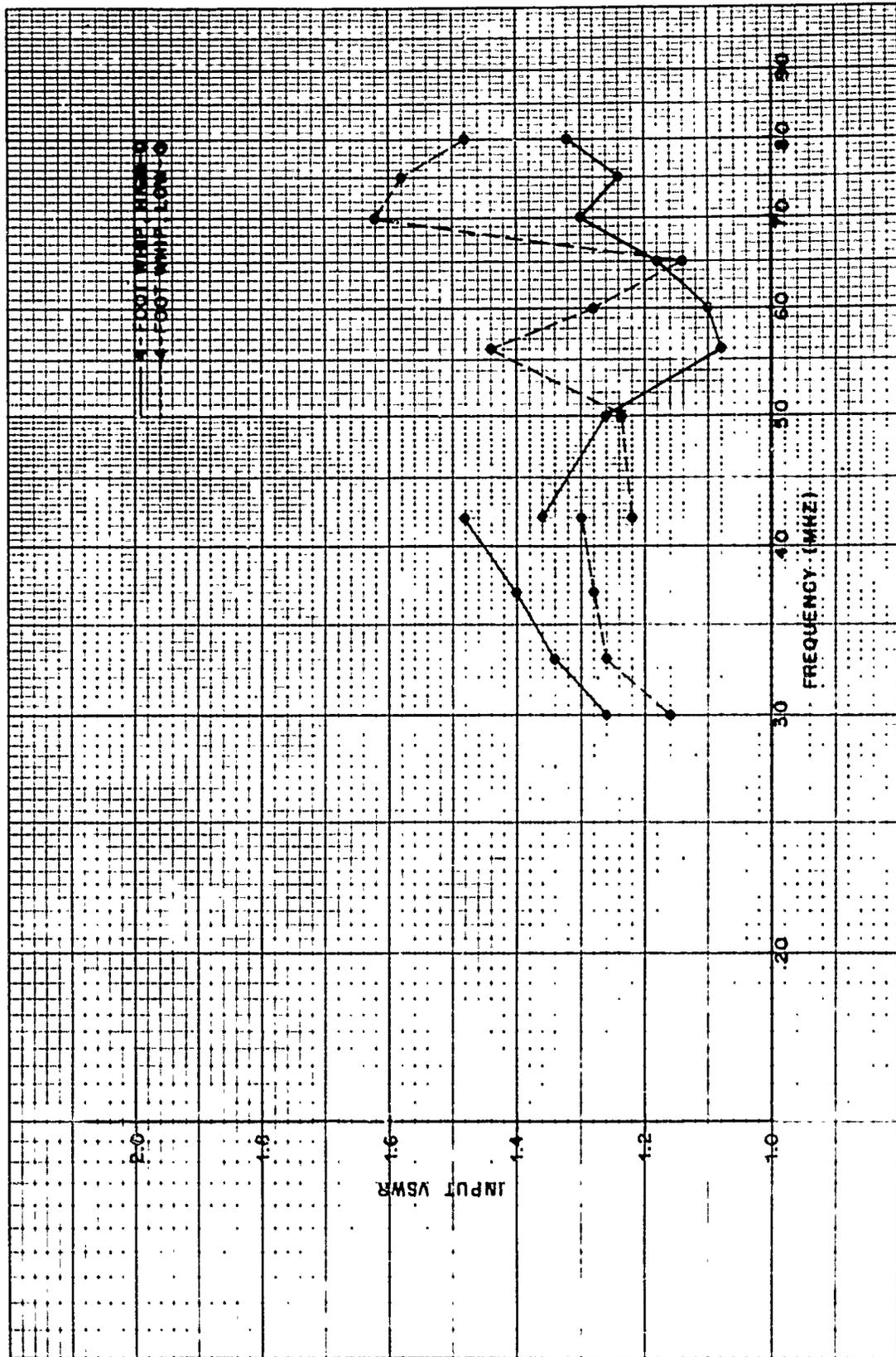


Figure 11. Four-Foot Whip Antenna Matching

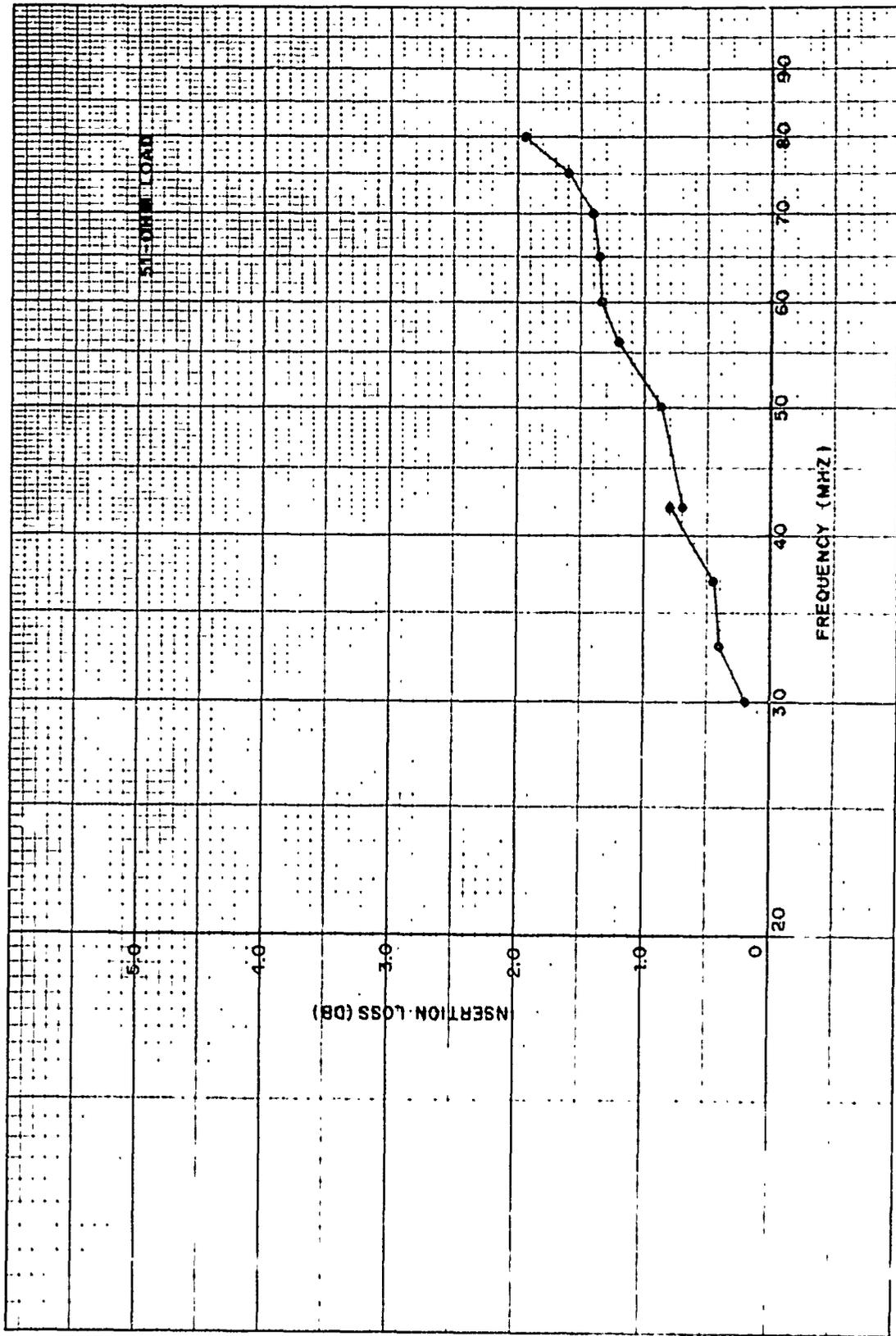


Figure 12. 50-Ohm Insertion Loss

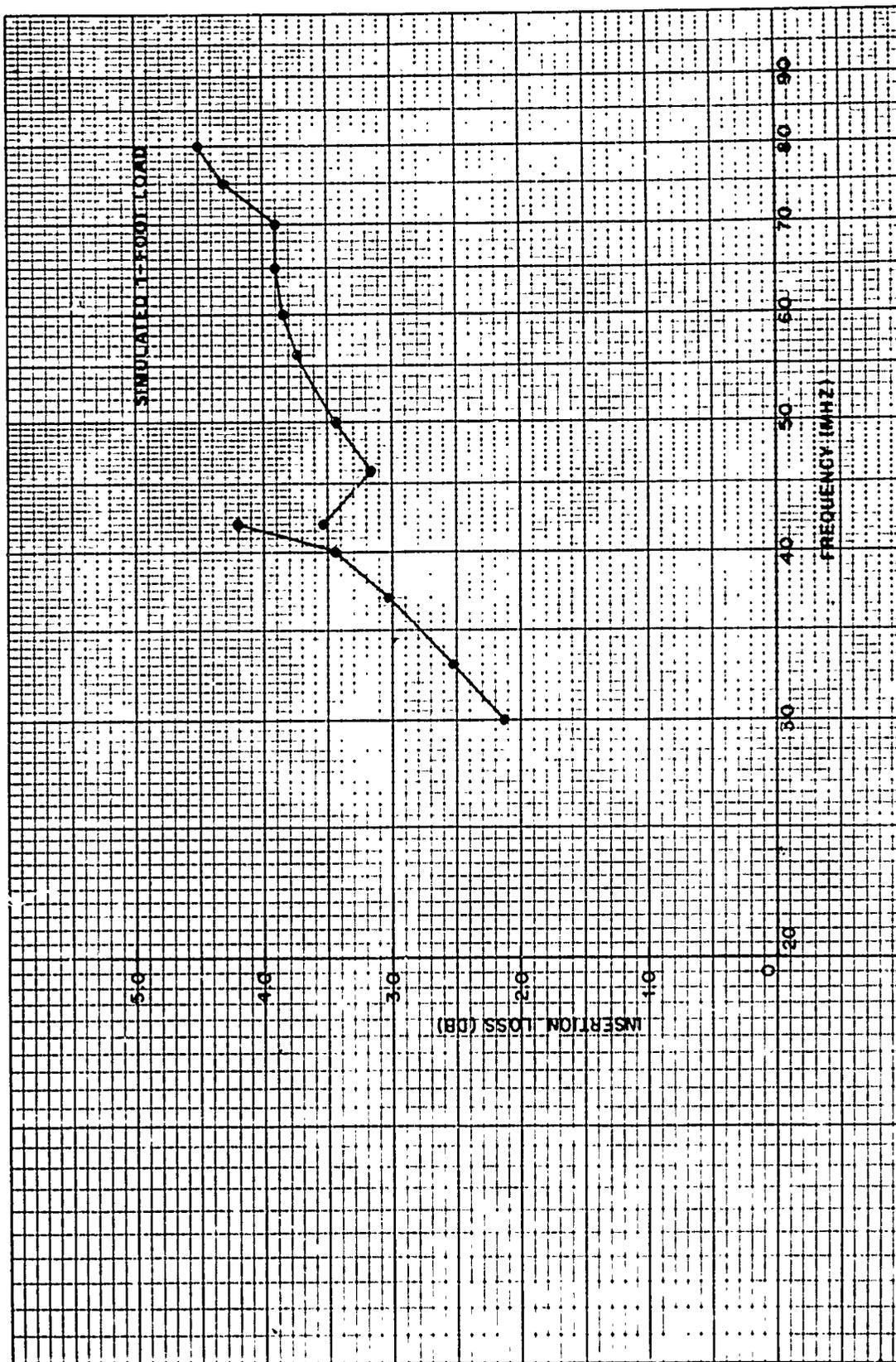


Figure 13. One-Foot Whip Antenna Loss

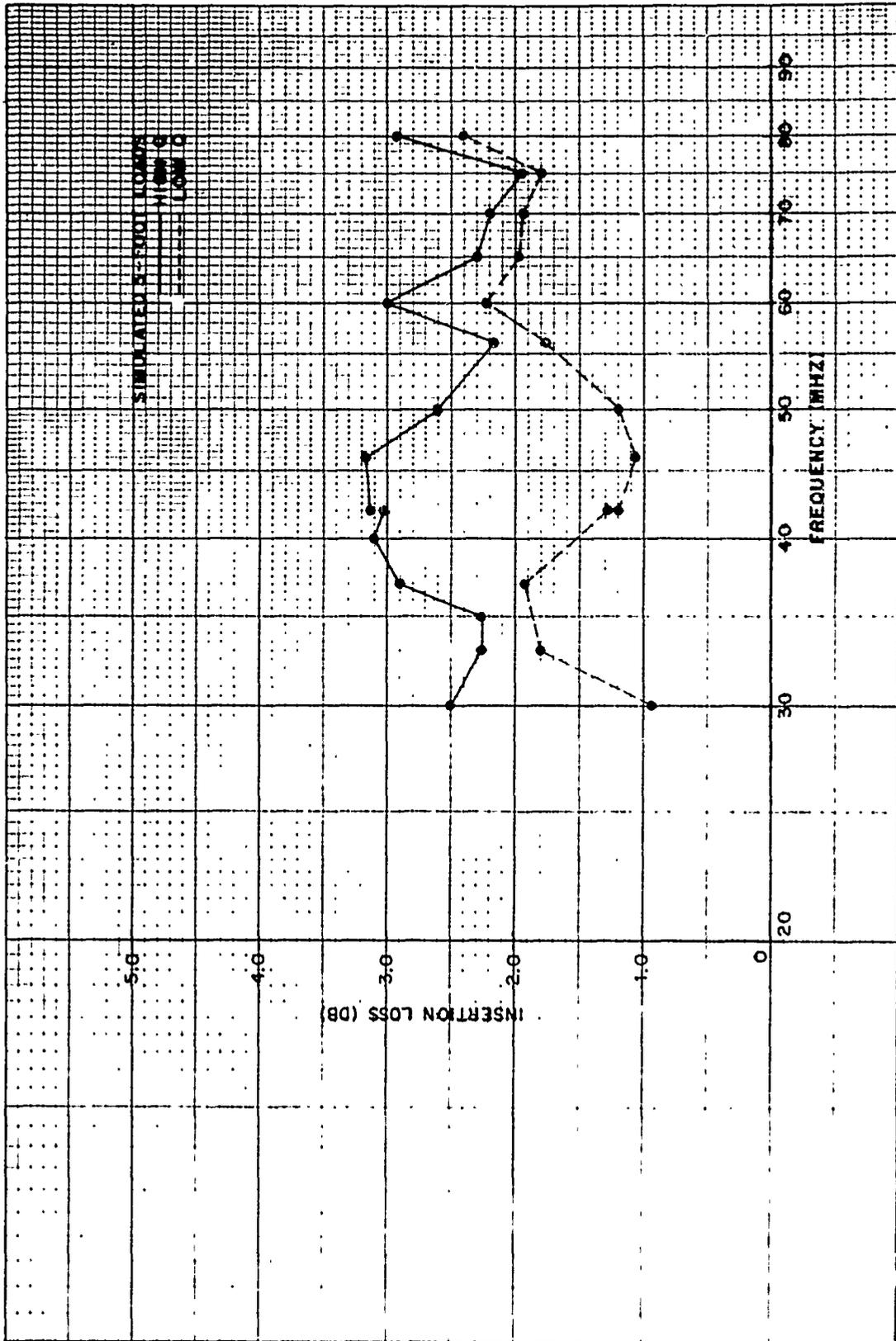


Figure 14. Three-Foot Whip Antenna Loss

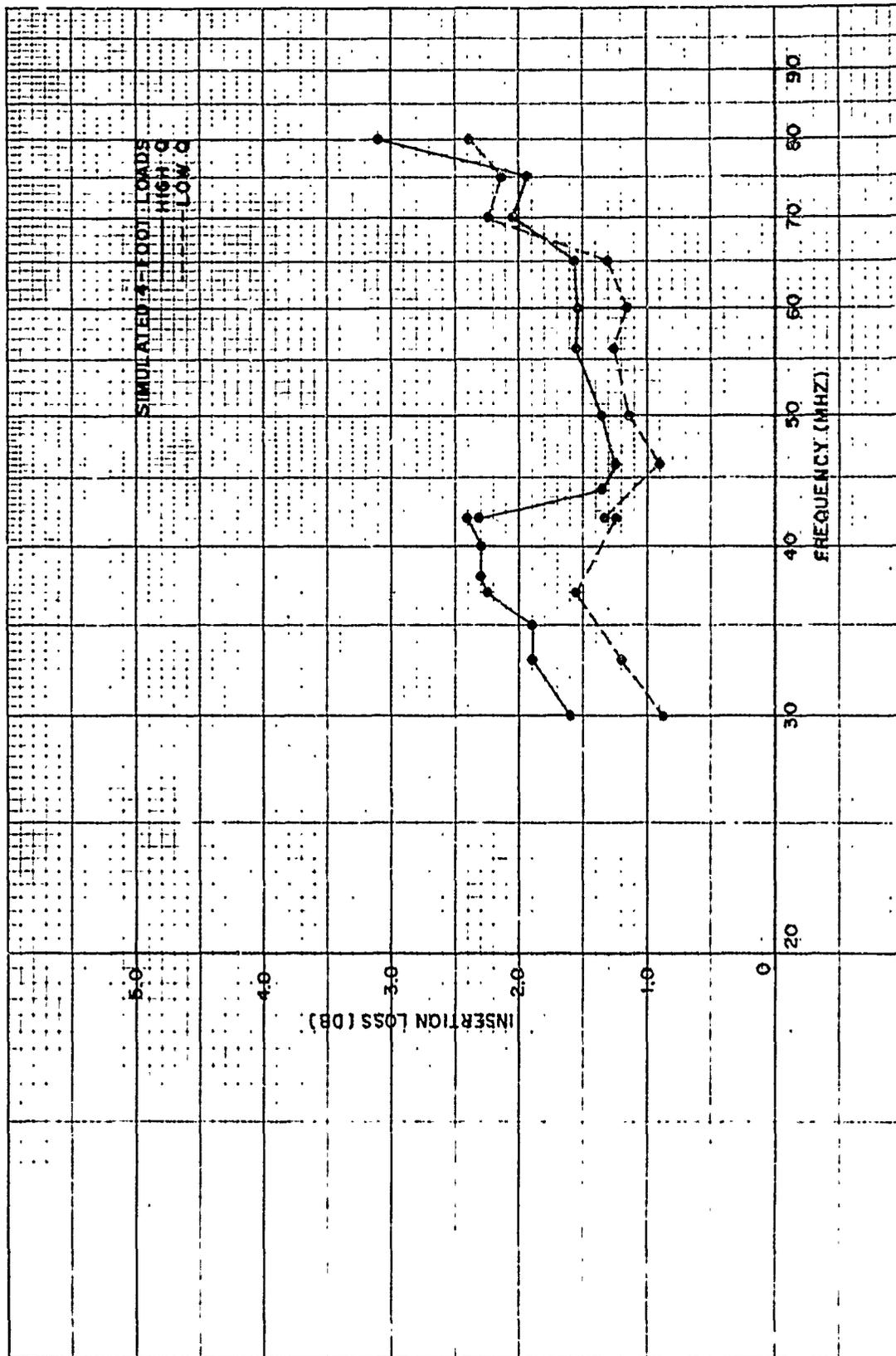


Figure 15. Four-Foot Whip Antenna Loss

Two other areas where improvements can be made in the present system are insertion loss and power consumption. The loss in the high band is higher than expected due to the strays associated with the capacitors. The present system requires approximately 10 watts of control power, most of which is in the logic system. The interface logic power is only necessary during the tune cycle, which is only a small percentage of actual transmit time in a real situation. In this unit, however, an end-of-tune automatic shut-off was not incorporated. In any model to be used with a packset this automatic shut-off would be incorporated, leaving only a "holding" or memory power consumption of about 50 milliwatts. Even this value can be decreased by modification of the control loops.

### 3.6 IMPROVEMENTS

Avco feels that the concept of a realizable capacitively tuned coupler has been proven on this program. Faults still exist in the present Demonstration Model; however, these are just faults with this particular model and not with the coupler concept as a whole. A description of the areas that need improvement and what can be done is the purpose of this section.

#### 3.6.1 Power Consumption

The logic system should be done in CMOS. This will reduce the logic power drain from the present 10 watts to approximately 5 milliwatts. Under these circumstances it can either be left "on" all the time, or still automatically shut off at end-of-tune cycle as previously mentioned. Power for the capacitor control loops then becomes the largest item. This can be reduced considerably by making a very slow loop (just enough to compensate for capacitor changes due to temperature, manpack vibration requirements, and material creep) and switching in this loop only after the tuning cycle is nearing completion. In this way, it is conceivably possible to reduce the loop memory power to around 15 milliwatts.

#### 3.6.2 Tuning Speed

As mentioned above, if the capacitors are run open-loop during tuning, they can then run close to their theoretical maximum (a clock rate of 200 KHz). A memory loop is really not necessary during the tune cycle - only afterwards. The only closed loop needed during tuning is the loop containing the VSWR information - capacitor drive voltage.

#### 3.6.3 Tuning Accuracy

The greatest inhibiting factor here is RF interference. The other problems mentioned above (reset and VSWR response matching) can be solved by proper circuit adjustments. The coupler, utilizing the VSWR tuning scheme, should tune as accurately as the network allows. There are two ways to implement an end-of-tune. The first is to set a VSWR threshold (1.3:1, for example) and then stop the tune sequence as soon as the sensor sees that the input impedance gets within the 1.3:1 circle. The second approach is to let the coupler tune until the capacitors run completely through the smallest increments. This will always result in the most accurate "tune point", although the tuning speed will be slightly longer.

#### 3.6.4 Insertion Loss

A different mechanical mounting configuration should be made for the capacitors. The Q of the capacitor by itself is quite high, however, strays associated with mounting lower the overall Q considerably. The capacitor leads were shortened as much as possible in the

present model (from the breadboard) but, since the same network was used in both the breadboard and the final unit, no major mechanical changes could be made.

### 3.6.5 Size and Weight

A number of things can be done to reduce the size, and therefore the weight, of the present coupler model. Among these are:

1. Fabricate the entire logic system in MOS LSI. This would completely eliminate one of the three circuit boards in the present system.
2. Simplify the capacitor control loop as mentioned before. In this fashion, the logic wafers, VSWR detector and control loops could be placed on one board.
3. Remount the capacitors. Presently, the two capacitors are mounted on a foam rubber type pad which adds considerably to the overall height of the system. This was done to provide some isolation (to decrease susceptibility to resonant vibrations) between the capacitors and the circuit board.

In addition to these, when the coupler is used in conjunction with a radio set, a complete case such as presently used will not be necessary. Provision should be made for a shield and that's all.

The present case also is about 1/4 inch longer than necessary in order to accommodate the connectors used with it. This also would not be necessary when the coupler is used together with a complete radio system.

## 4.0 DETAILED DISCUSSION

A detailed discussion of both the investigative work on the program and the development effort on the 30 - 80 MHz Coupler can be found in the earlier reports (ECOM-0101-1, 0101-2, and 0101-3). This information will not be repeated here; however, a brief summary of the topics covered, together with some discussion of the more significant results, will be given.

### 4.1 PROGRAM EFFORT

#### 4.1.1 Initial Program Goals

Primary emphasis on the program was put on the development of a 30 - 80 MHz Antenna Coupler. This unit would incorporate an LSI package for the logic and would physically represent, as close as possible, a final mechanical design compatible with a small packetset radio.

#### 4.1.2 Initial Program Approach

Original program approach centered around constructing the 30 - 80 MHz Coupler using saturable reactor technology. At that time, permeability tuning was thought to be the optimum electronic tuning technique for this type application. Since the components for this technique were well defined (and also being concurrently studied under a materials analysis contract at Avco), only minimal effort was to be devoted to component

development on this program. Major emphasis would be placed on developing improved techniques in the associated areas of networks, detectors, and logic systems.

#### 4.1.3 Initial Program Effort

Initial effort on the program was concentrated in three areas:

1. Analysis of ferrite temperature characteristics in an effort to determine an optimum thermal saturable reactor configuration: Heat rise and inductance drift of the saturable core reactors were plotted against RF power dissipation for different yoke assembly heat sinks. Computer analysis was used to determine worst-case dissipations in any network inductor for a given antenna. From this analysis and the thermal data, the optimum ferrite size and yoke configuration were determined for the coupler saturable reactors.
2. Analysis of RF network configurations in an effort to determine the minimum element saturable reactor network: Extensive computer effort was made to determine element values, insertion losses, node voltages, and branch currents for each antenna over the frequency range.
3. Development of a "current-control" technique for the saturable reactor pulses: A method was developed which enabled more accurate control of the DC pulses used to vary the saturable reactor's inductance in an attempt to remove some of the dependence of the saturable reactor system tuning on input DC supply voltage level.

A more extensive discussion of these efforts is described in Technical Report ECOM-0101-1.

#### 4.1.4 Piezoelectric Capacitor Initial Effort

Discovery of the piezoelectric capacitor early in the program opened up a completely new area for investigation. As a potential tuning element, this capacitor offered many advantages as previously outlined in this report. Because of the promise this component had for this particular application, it was discussed with ECOM and a common conclusion was reached to investigate it further. Since this represented a whole new area of component technology, an extensive effort had to be planned in order to incorporate the capacitor into a realizable antenna coupler. This investigative effort had to be, by necessity, much longer than the minimal component study effort originally planned.

Effort in the other program areas of networks, detectors, and logic systems, since these are all dependent on the variable tuning element, had to be delayed until the characteristics of the piezoelectric capacitor were better defined. Therefore, the entire study phase of the program was extended in order to accommodate the additional effort required on the piezoelectric capacitor.

#### 4.1.5 Study Phase Effort

During the remainder of the study phase, primary emphasis was placed on an analysis of the piezoelectric capacitor from the standpoint of its applicability to the Electronically Tuned Antenna Coupler System. Capacitance range, Q-factor, and dielectric materials were investigated for different configurations. Different biasing methods were also studied. A closed-loop control system was devised to compensate for environmental effects and to provide a system memory.

The original work and plans in the associated areas of networks, detectors, and logic, since it was based entirely on a saturable reactor type system, had to be either considerably modified or changed completely. The results of the new work in these areas can be summarized as follows:

- A new network was designed and all the significant parameters established through computer analysis.
- A new tuning method was established, which depended only on VSWR information in order to execute a successful tune cycle.
- A digital VSWR detector was developed to be compatible with a digital tuning logic.
- A control loop was developed for the capacitor which would compensate for all disturbances.

At this point, preliminary design work had been accomplished in all the above-mentioned areas. The new tuning method has been computer simulated and looked good. The test capacitors on hand had exhibited a low enough mechanical Q to assume compatibility with the control loop. It was therefore decided that, although a schedule slippage was unavoidable, the program could be completed without any increase in funds provided no problems of any magnitude developed during the initial development phase.

#### 4.1.6 Initial Development Effort

The digital VSWR detector was breadboarded and tested. During this effort two problems were encountered:

1. Since the Coupler had to tune over a 2 – 10 watt range, the correct ratio of forward to reflected power had to be maintained exactly for similar VSWR's independent of the RF power level. In order to assure this, operation had to be in the same region of the peak detector diodes (either the linear or the square-law region) over this power range. A redesign effort was necessitated to determine the optimum coupling level between detector and RF line; also, a diode investigation was undertaken to find units having a wide "predictable" region. The original idea of using a pad in the forward power line during a tune cycle was discarded because of system complexity and because it restricted the versatility of the detector. Therefore, the peak detector, directional coupler, and D/A converter interface circuitry had to be modified before the problem could be resolved.
2. A special D/A converter had to be found that would handle a variable reference voltage (in this case, the reference is the forward RF power sample which varies over a 5:1 range).

The initial effort on the capacitor control loop emphasized loop speed in an effort to develop as much compensation for environmental effects (especially vibration). A loop was developed having a 1 KHz unity gain frequency which worked very well with the test capacitors. This loop was then redesigned for minimum power consumption and performance was verified with the test capacitor.

Two new capacitors were then made for the RF network breadboard. Initial tests on these models disclosed a much higher mechanical Q than possessed by the test models. This led to the following problem:

The capacitor control loop would not lock-up since higher order resonance modes of considerable magnitude were discovered in the capacitor (they were unnoticed previously because of the lower Q models). These new resonance modes appeared in the 1 KHz - 2 KHz frequency range, or right in the area of the loop unity gain point.

The only immediate solution to this problem involved lowering the loop unity gain point to reduce its speed; however, this not only reduced compensation needed for external vibration but also led to longer tuning times. In order to really solve the problem, a two-pronged approach had to be undertaken:

1. Investigate the piezoelectric material to determine how to fabricate capacitors having lower mechanical Q. In this line, different types of damping material, different material sizes and thicknesses, and different mounting configurations were studied.
2. Study the effects of compensating networks, e.g., active-element notch filters, on the capacitor control loop response.

Because of this control loop problem, a complete automatic system could not be assembled. However, the network was built up and manually tested with the VSWR detector. In this fashion, tuning range and other static system characteristics could be analyzed. Initial tests proved the accuracy of the new tuning approach, but pointed out, however, the following two problems:

1. The physical placement of the capacitors on the RF network board led to excess stray capacitance, which restricted the tuning range because the necessary minimum capacitance could not be achieved.
2. The piezoelectric relays developed by Avco and used as the switches in the RF network required readjustment after initial use. Further study of the force - control voltage characteristic for these relays was necessary to assure reliability. Also to be resolved was an apparent contradiction with the manufacturer's data on the material's deflection versus applied voltage characteristic. In the mean time, it was decided to use small commercial latching relays while the above investigations were going on.

These problems necessitated not only a new RF network layout and fabrication, but also a study of pertinent relay parameters (especially loss, contact capacitance, and switching time) of commercial latching relays.

#### 4.2 BREADBOARD ANALYSIS

After fabrication of a new network using TO-5 latching relays, the control loop's speed was reduced considerably, the VSWR Detector and interface logic system were tested, and the complete coupler breadboard was assembled and tested. Tuning data into the simulated antenna loads is shown in figures 16, 17, 18, and 19. This data reflects the complete breadboard, with a discrete IC logic system, operating in an automatic mode.

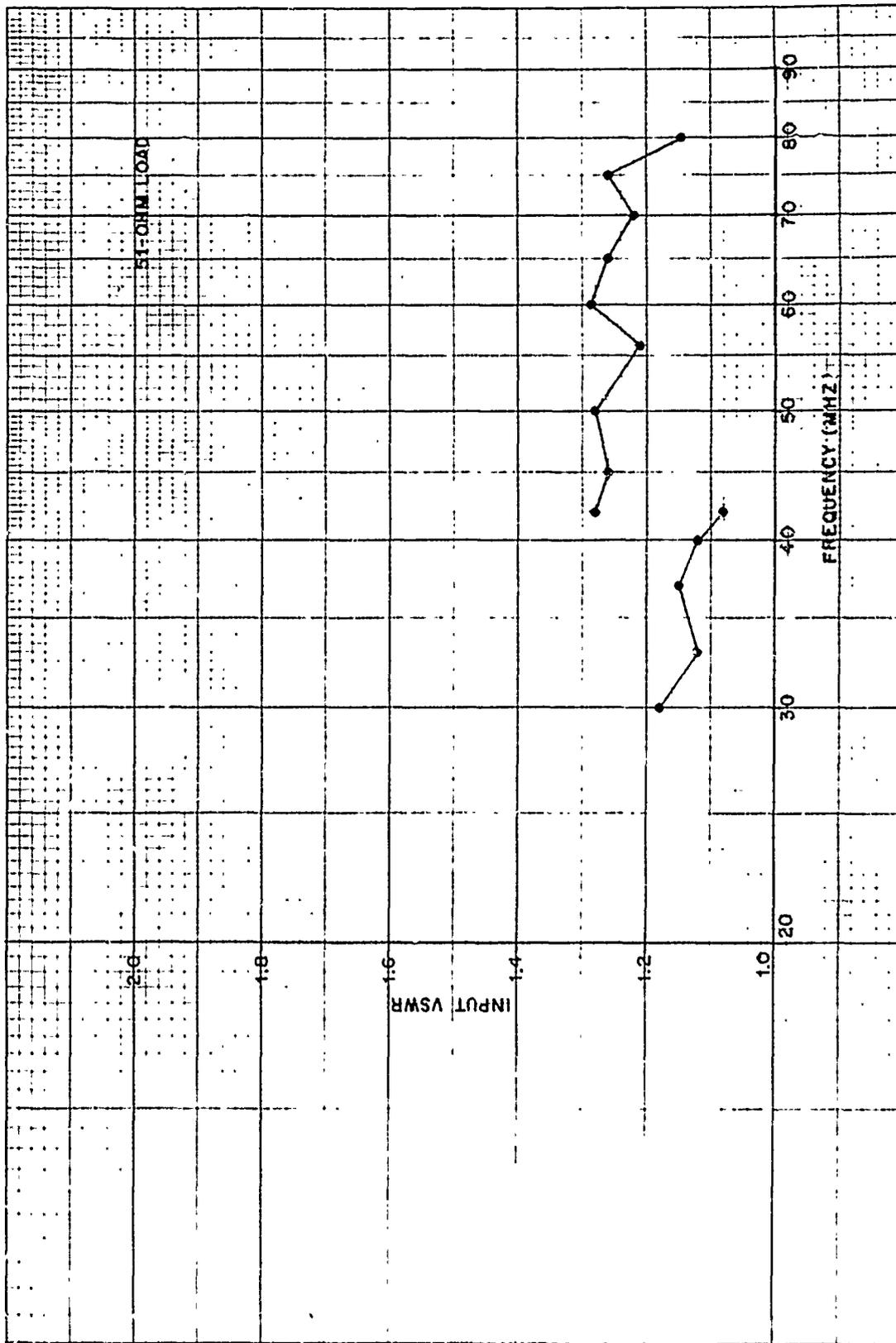


Figure 16. Breadboard 50-Ohm Matching

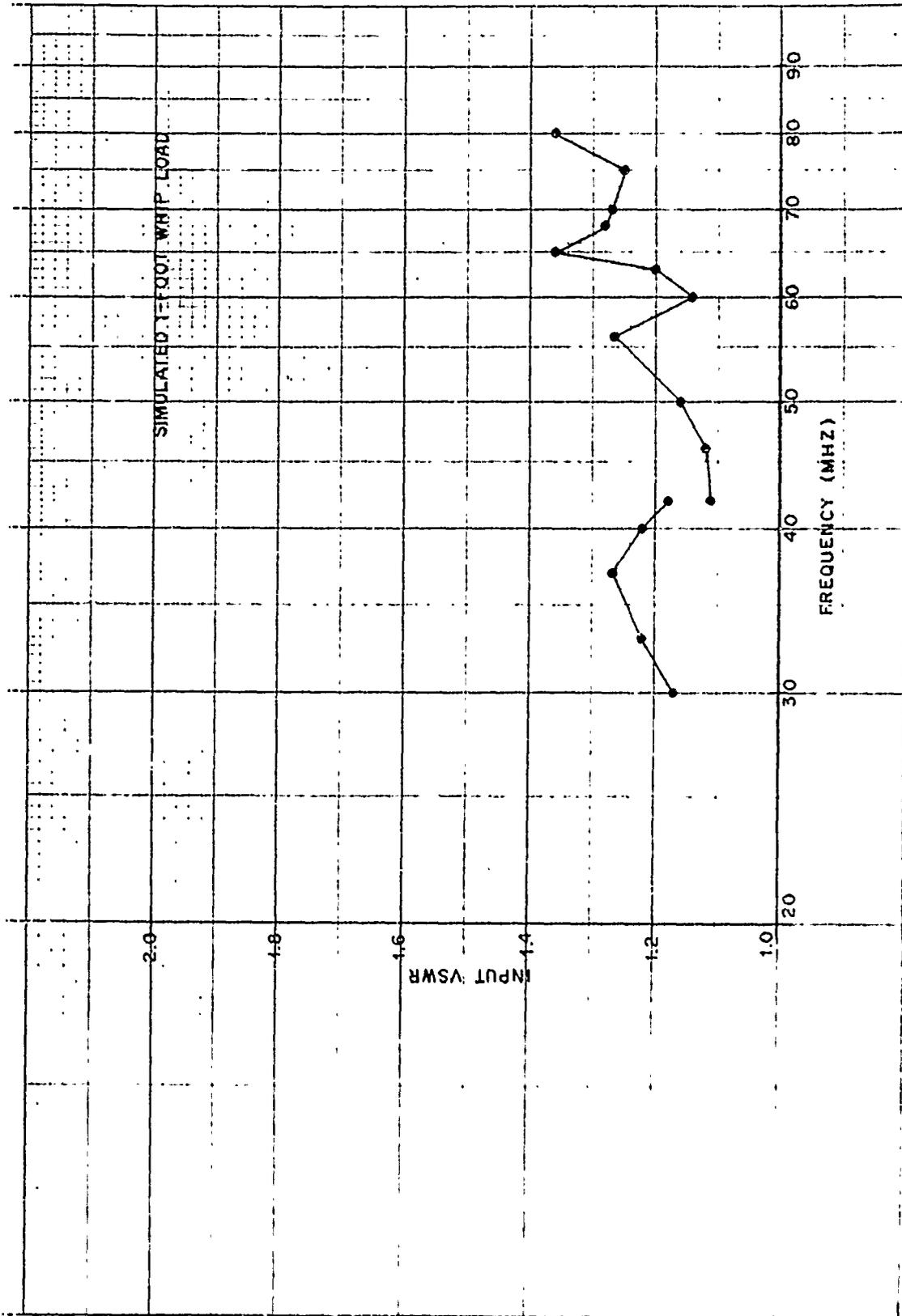


Figure 17. Breadboard One-Foot Whip Antenna Matching

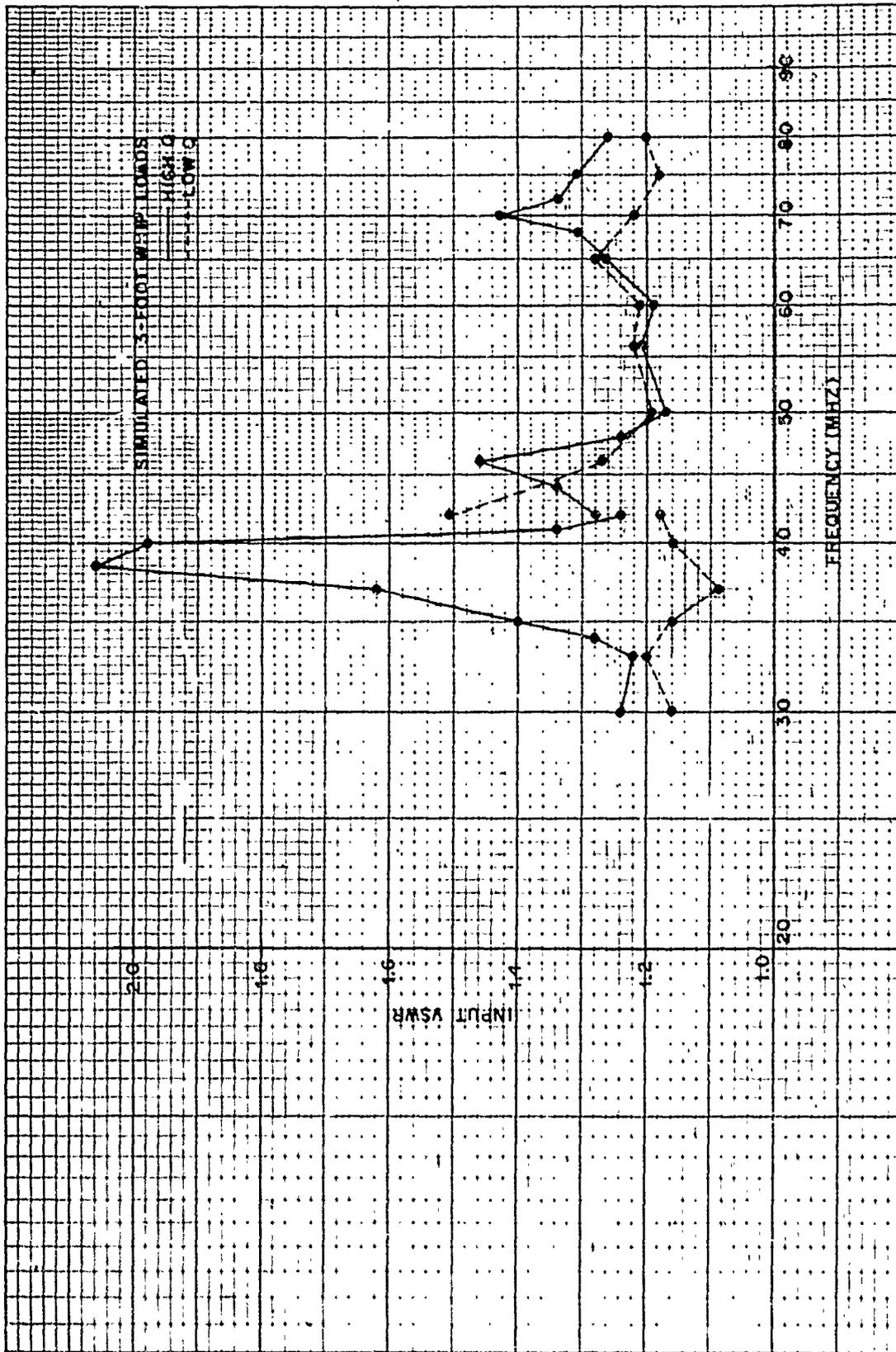


Figure 18. Breadboard Three-Foot Whip Antenna Matching

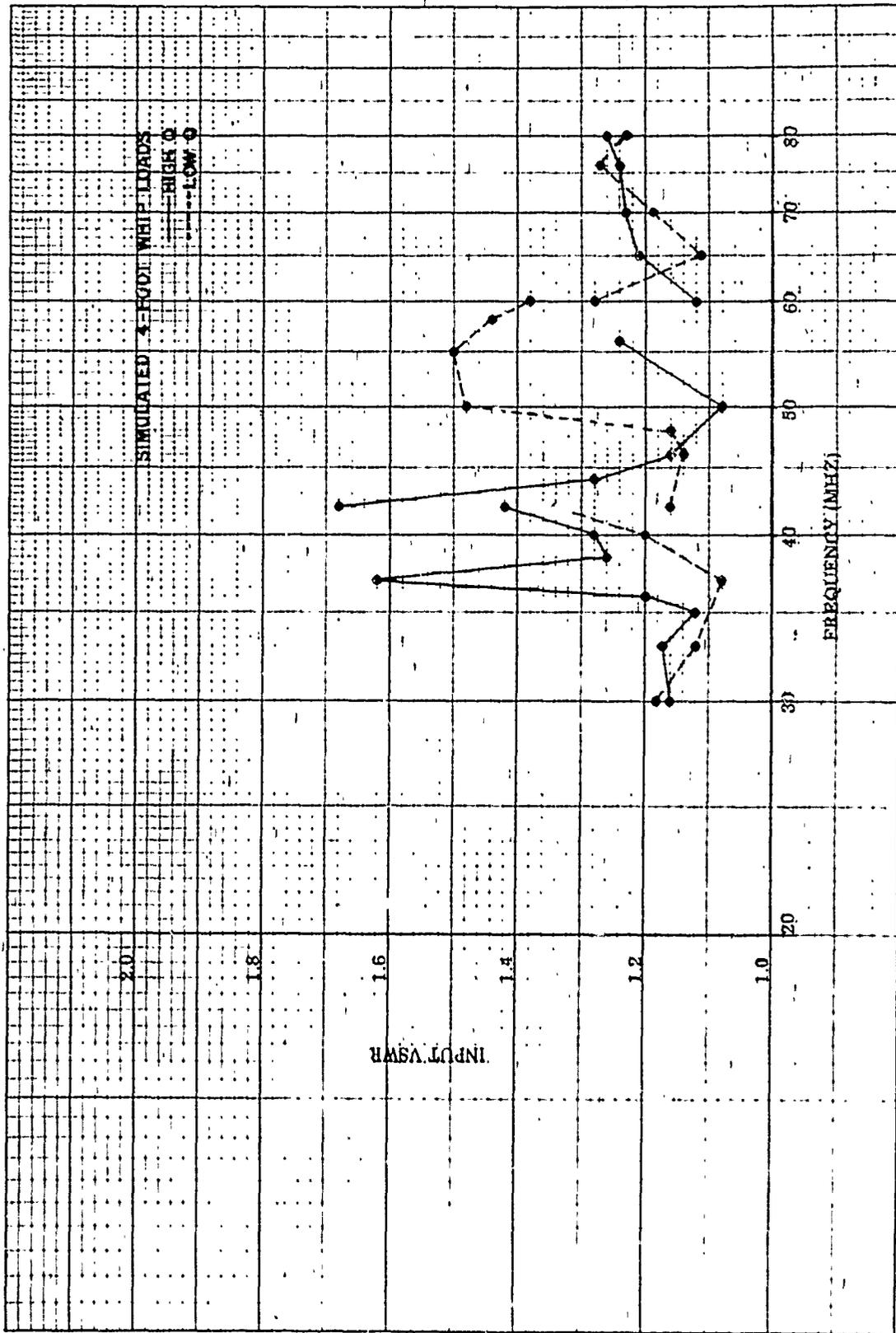


Figure 19. Breadboard Four-Foot Whip Antenna Matching

### 4.3 REDESIGN EFFORT

A redesign effort was initiated to consider the problems brought to light during the breadboard tests. A summary of the work in this phase is as follows:

- **Tuning Capability:** The VSWR Detector forward and reflected power curves must be closely matched (versus frequency) to insure accurate tuning. This was done in a separate adjustment. It was also important to isolate actual "tuning" problems from RFI induced problems. Therefore, a thorough analysis of the decoupling and shielding requirements was made. The values of the mid-section series inductors were changed to increase the network matching range and eliminate the "tuning holes" that showed up in the breadboard tests.
- **Insertion Loss:** The piezoelectric capacitors were remounted and leads shortened as much as possible to help alleviate the low-Q problem at the high frequencies.
- **Tuning Speed:** Due to the higher-order resonance modes (discussed in Technical Report ECOM-0101-3) of the present configuration piezoelectric capacitor, the control loop speed had to be cut back considerably. This was the only way, other than an entirely new capacitor configuration, to guarantee closed-loop stability during tuning. Therefore, the clock frequency was reduced from 200 KHz to 20 KHz. Lowering the loop speed has another undesirable feature, namely, attenuation to external vibrations is reduced. However, it was found that the capacitor resonance modes vary greatly from unit to unit and with mounting configurations. Therefore, the control loops could not be standardized with the "fix" and still maintain high speed. The only real solution was an entirely new capacitor type.

The other factor affecting tuning speed is the setting of the VSWR "stop". The system can be adjusted so that tuning will automatically stop whenever the input impedance gets within a predetermined VSWR. Tuning time will vary, depending on the input accuracy required, i.e., a poorer match requires less time than searching for the "optimum" match (which the system will do if allowed to run without a "stop").

From the above it is obvious that the major decisions involved the piezoelectric capacitor. The analysis of the capacitor from both a packaging and performance viewpoint assumed that:

1. The entire unit (capacitor plus mounting brackets and leads) must not resonate in the band of interest.
2. The unit must be isolated from the network board to maintain control loop stability.
3. The unit must fit into the original package area-wise (3 x 4 inches).

It was decided to use the original capacitors in the final model (building a "new configuration unit" had a severe time risk, and there was no guarantee of improved performance) and to find a mounting arrangement that minimized insertion loss while still maintaining stability. Therefore, the final configuration is a compromise in that a foam rubber isolator is used with each capacitor, which still required the RF leads to be slightly

longer than desired due to the elevated mounting. However, as was previously mentioned, the control loops are very dependent on the capacitor mounting configuration, especially if the capacitors are not sufficiently isolated from the board.

After ascertaining that the breadboard logic system was functioning correctly, the system drawings and specifications were updated prior to building the final model. Original plans were to put the logic into an LSI configuration. However, after discussion with different LSI vendors (Collins, Motorola, General Instrument, and Solid State Scientific), the average quote for development of monolithic LSI came to two or three wafers with the total cost being too great for the program. Bipolar LSI was also considered. Texas Instruments was contacted concerning the applicability of their DRA (discretionary routing approach). However, their approach also required three packages to ensure doing the job.

Since the cost for either of the above approaches exceeded the budget, another had to be found. The major objectives here were to maintain the original concept of a small package and still ensure a timely delivery schedule within the budget.

It was then decided to utilize the hybrid approach, whereby the individual IC chips would be purchased and used to build up the logic system in hybrid packages. This approach would not only enable meeting the budget requirements, but would also still enable utilization of a small overall system package since it was determined that all of the chips could be mounted on one 3 x 4-inch board. The chips would be purchased and mounted by Avco's Microelectronics Laboratory into a small number of individual substrate packages. These packages would then be mounted on a master printed circuit board used for the complete logic wiring.

#### 4.4 FINAL DEVELOPMENT PHASE

The work effort on the Exploratory Development Model was broken down into the following phases:

- Layout Phase
- Fabrication Phase
- Module Test Phase
- System Test Phase

##### 4.4.1 Layout Phase

This phase included mechanical layout and release of all the modules and the overall system assembly. Layouts for the VSWR detector, control loops, and system were completed in-house. It was decided, however, to have the layout for the entire logic system (including the VSWR detector logic) done by an outside source whose major field was computer layout techniques. Because of the circuit complexity, it was felt that a considerable savings could be achieved by this means. Electronic Graphics of Dallas, Texas, was selected as the vendor for the layout work. This effort was to include not only the layouts for the individual substrates (containing the IC chips), but also the printed circuit board for the overall system.

Therefore, effort during this phase was mainly directed towards partitioning of the logic, finalizing the substrate package sizes, and determining test procedures for the individual packages. It was finally decided that six individual packages were required for all the IC

chips in the system. After interfacing with the vendor, it was also determined that both sides of the logic board would have to be used, since, in addition to the six substrate packages, three D/A converters and the clock circuitry had to be mounted on this board.

Initial contact with Electronic Graphics was made during the last part of May 1971. The finished partitioned logic drawings were delivered during June 1971. Promised delivery dates for all layout items were six weeks to two months.

#### 4.4.2 Fabrication Phase

The VSWR detector and capacitor control loops were fabricated without any difficulty. Work on the logic system was proceeding slowly due to a delivery delay in receiving layouts from Electronic Graphics. The individual substrate layouts were being sent in one at a time, the first one being received at Avco the first week of August 1971.

In the meantime, it was decided to utilize beam lead type IC chips in the logic wherever possible. A beam lead is simply a chip with pre-attached wires, or beams, connected to the active elements. Ultimately, these wires are bonded to the passive hybrid components. The conventional assembly steps, such as wire bonding, probing, and die bonding are virtually eliminated, as the attaching process is completed in one step. The result of using beam lead chips is higher reliability and much easier assembly and repairability.

Some of the IC chips, however, were not available in beam lead form. The majority of them were available and they were ordered for use wherever possible. Motorola is the vendor supplying the IC chips, beam lead and conventional IC.

The total logic network contains 126 (5400 series) IC chips mounted on the 6 individual substrates. These substrates are made up of 4-conductor and 3-dielectric layers, averaging approximately 25 chips per 1-inch by 3-inch area of substrate. This represents probably the highest density of hybrid circuitry achieved in the industry. The entire system, mounted on the 3 x 4-inch board, represents a size reduction of about 6 to 1 from discrete to hybrid.

#### 4.4.3 Module Test Phase

This phase included test and alignment of the Capacitor Control Loops and the VSWR Detector. After fabrication of the logic wafers, they were DC tested, but they were not completely tested for timing errors.

After assembly of the six wafers on the parent board, the complete logic was tested and debugged. The test setup included a bank of lights to monitor the digital "word" from the VSWR Detector, a bank of lights to monitor the output fixed coil register, and a test fixture which allowed pulsing the logic slowly, one at a time, with provision to stop anywhere in the sequence to accurately check logic sequencing and aid in troubleshooting.

One problem that arose during the testing of the logic was a heat problem which prohibited testing for long durations of time. About 10 watts was drawn by the total logic system, and this was all concentrated on the one small board. No method of heat sinking or heat removal was found practical. Even forced air cooling was ruled out since this caused condensation to form on the glass cover enclosing the logic. Therefore, extreme care had to be exercised during the troubleshooting stage in order not to overheat the logic and cause any permanent damage to the chips. Many chips had to be replaced as it was, either because they were defective to begin with or because they were damaged in probing.

#### 4.4.4 System Test Phase

The final program effort was a system test phase, the results of which have already been described in the first section of this report.

### 5.0 CONCLUSIONS AND RECOMMENDATIONS

The concept of a realizable capacitively tuned antenna coupler has been proven on this program. The piezoelectric capacitor has successfully demonstrated its usefulness as an electronically controllable tuning element. The test results on the deliverable coupler system have illustrated the successful integration of many new concepts, among them being:

- The piezoelectric capacitor
- The digital VSWR detector
- The "one-detector" tuning scheme
- High-density hybrid logic packaging

The overall design approach for this coupler had a number of difficult system problems (which were not resolved in previous couplers) to consider. Among these were:

1. Matching a variable one to four-foot whip with its many impedance variations due to environment and operator orientation.
2. Compatibility with all types of manpack batteries. This included the problems of DC voltage fluctuation, voltage transients, and power consumption.
3. Ability to tune over a varying RF power range.
4. Linear operation.
5. Small size and light weight.
6. Fast tune time.

These were the major requirements on the coupler system -- no system in the past had been developed to handle satisfactorily all of these design considerations. Avco's deliverable model demonstrated the feasibility of a coupler to meet all of these requirements. Because of the newness of many of the design concepts, though, there is room for improvement in the demonstration model as it now stands. Refinements have been made in some techniques in order to correct problems in the first model. These suggested improvements include.

1. Power Consumption: Total power drain could be reduced to about 20 milliwatts by making a CMOS logic and a slow memory loop (paragraph 3.6.1).
2. Tuning Speed: The capacitors could run open-loop during tuning, with the slow loop switched in as a memory after tuning is completed (paragraph 3.6.2).
3. Size and Weight: With a CMOS logic and a simplified control loop the logic, loops, and detector can be on one board (paragraph 3.6.5).

## APPENDIX A

### ANTENNA CHARACTERISTICS

#### A-1. TYPES OF MANPACK ANTENNAS

The types of whip antennas specified in ECOM Development Description EL-CP0000-0047B, paragraph 3-2, are typical of those used with manpack radios. The 9-foot and 6-foot whips are electrically short monopoles at the lower frequencies — from 2 to 4 MHz for the 9-foot whip and from 4 to about 15 MHz for the 6-foot whip. By commonly accepted definition an electrically short antenna is one whose length is less than  $\lambda/10$ . Such whips exhibit very high capacitive reactance, as discussed below, which is aggravated by the need to keep the whip diameter very small to achieve flexibility and light weight.

#### A-2. WHIP REACTANCE

The typical packset whip consists of a No. 22 copper conductor or six No. 30 conductors embedded in a fiberglass envelope which supports and protects the conductors. The small effective diameter of the antenna conductor results in a very large ratio of L/D, length to diameter, which controls the characteristic impedance and reactance of the radiator. Equations A-1 and A-2 give the value of input base reactance and characteristic impedance for a short whip.

$$Z_0 = 60 \left[ \ln \left( \frac{4L}{D} \right) - 1 \right] \text{ ohms} \quad (\text{A-1})$$

$$X = Z_0 \cot (2\pi L/\lambda) \text{ ohms} \quad (\text{A-2})$$

where:

L = antenna length  
D = antenna diameter  
 $\lambda$  = wavelength

As an example, the  $Z_0$  and input reactance of the 9-foot whip at 2 MHz, assuming an effective conductor diameter of 1/16 inch are:

$$\begin{array}{ll} Z_0 & = 470 \text{ ohms} & \text{from equation A-1} \\ X & = -4060 \text{ ohms} & \text{from equation A-2} \end{array}$$

In the practical situation, the base reactance of a short whip is somewhat less than the value calculated from equation A-2 because of a shunting capacity at the base. This is present because the insulator which supports the whip introduces an additional capacity of several picofarads. The effect of this shunting capacity is to reduce the apparent input reactance and resistance of the antenna and can readily be calculated for a given antenna.

### A-3. WHIP RESISTANCE

The antenna resistive component is made up of the radiation resistance which produces useful radiated power and the loss resistance which is made up of ohmic losses and ground losses which dissipate RF power. The radiation resistance is related to antenna electrical length by equation A-3 which holds for a short whip radiator.

$$R_r = 400 \left( \frac{L}{\lambda} \right)^2 \text{ ohms} \quad (\text{A-3})$$

At 2 MHz, the 9-foot whip has a radiation resistance.

$$R_r = 0.13 \text{ ohm}$$

which is appreciably less than the loss resistance which exists at this frequency in a manpack whip.

### A-4. IMPEDANCE RANGE UNDER OPERATING CONDITIONS

The expected range of antenna resistance and reactance for the specified antennas over the frequency range is shown in figures A-1 and A-2. These values are the result of data measured by Avco Electronics Division on a number of similar packset whips. The effect of a shunt capacity (about 5 pf) at the base insulator is included. The range of values at a specific frequency is primarily the result of variations which occur as the operator moves, bends, stoops, or lies down. Variations can also result from bodily differences from operator to operator and different conductivity of the ground over which the operator is moving. The coupler networks were to be designed to accommodate the entire range of resistance and reactance variation, shown in figures A-1 and A-2, which is considerable.

Referring to figure A-1, the wide variation in antenna resistance may, at first, seem unexpected. However, the typical packset whip exhibits such a large and variable input resistance because it lacks a sufficient counterpoise system for efficient operation. The only counterpoise permitted with the tactical manpack antenna is the radio set enclosure, which is too small to give good efficiency or stability. Some insight into this problem can be obtained by use of figure A-3 which shows the equivalent circuit of a manpack radio, antenna, and operator. RF currents in the antenna and radio set enclosure couple into the operator and ground in a rather complex fashion. Large variations in antenna impedance can occur because the circuit impedances of figure A-3 are highly variable.

A graph of radiation resistance versus frequency for the specified whip antennas is given in figure A-4, which also shows the typical loss resistance expected with a packset whip operated on a man's back over average soil. It is evident that the radiation resistance is much less than the loss resistance at frequencies up to about 20 MHz and thus the radiation efficiency of the antenna is quite low until the whip length is greater than about  $\lambda/8$ .

The loss resistance plotted in figure A-4 includes ohmic losses in the whip conductor, loss due to the imperfectly conducting ground, and loss due to currents which flow in the operator's body. The conductor loss is generally much smaller than the other two loss components for manpack antennas. This has been determined by a series of extensive measurements conducted by Avco Electronics Division, starting in 1958 and extending over a period of several years during which a number of different manpack and portable antennas

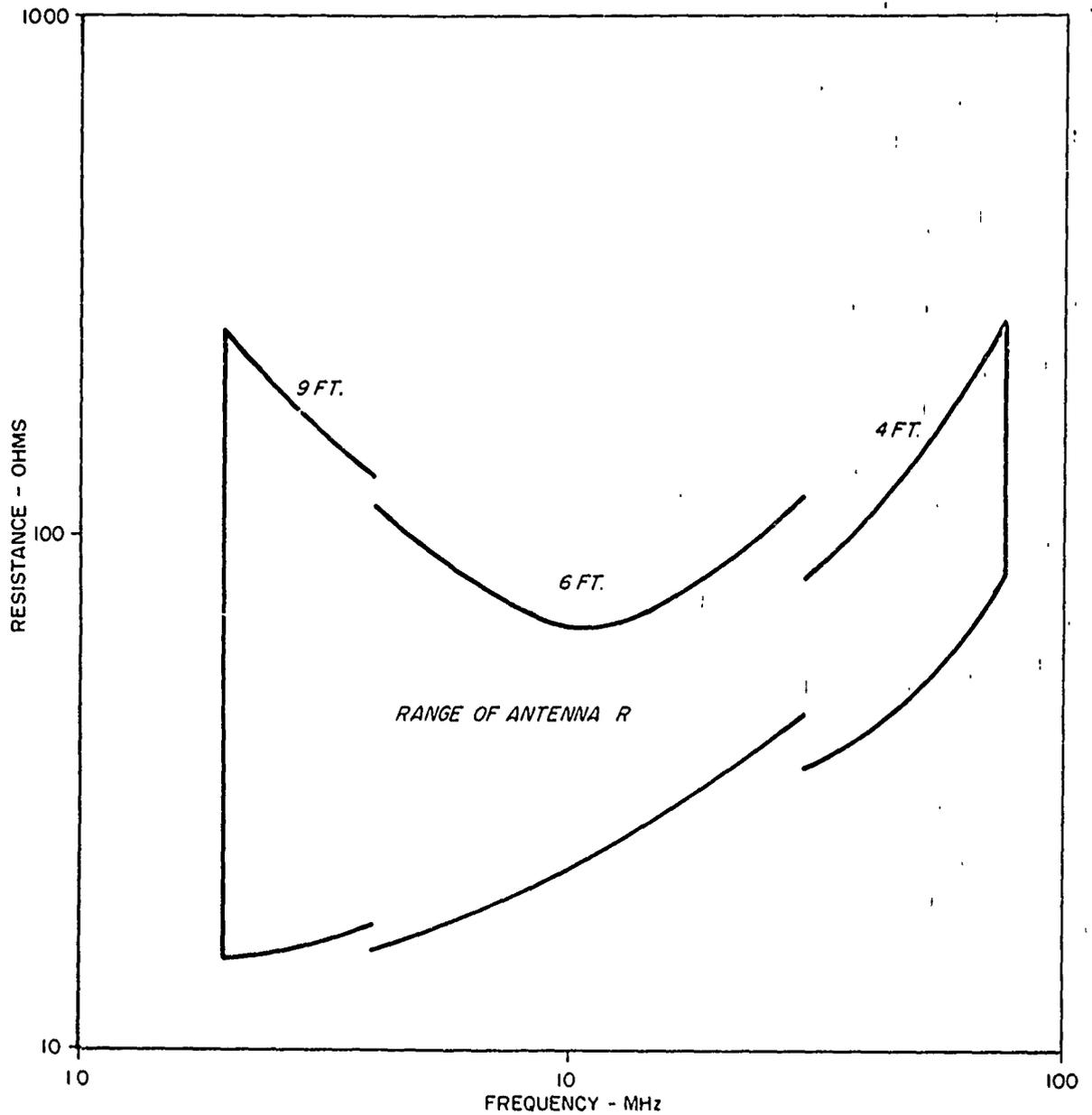


Figure A-1. Packset Whip Antenna Resistance

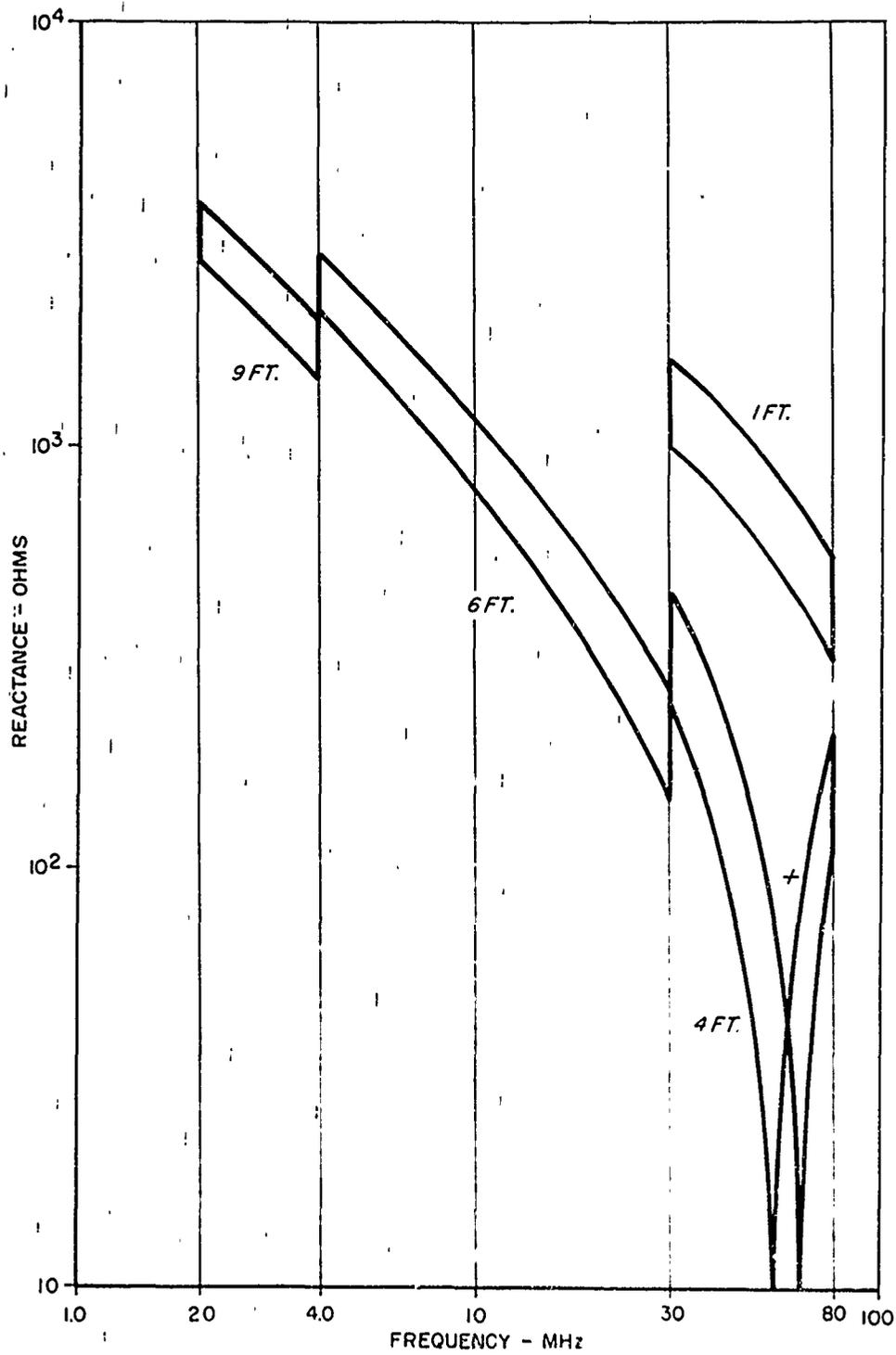
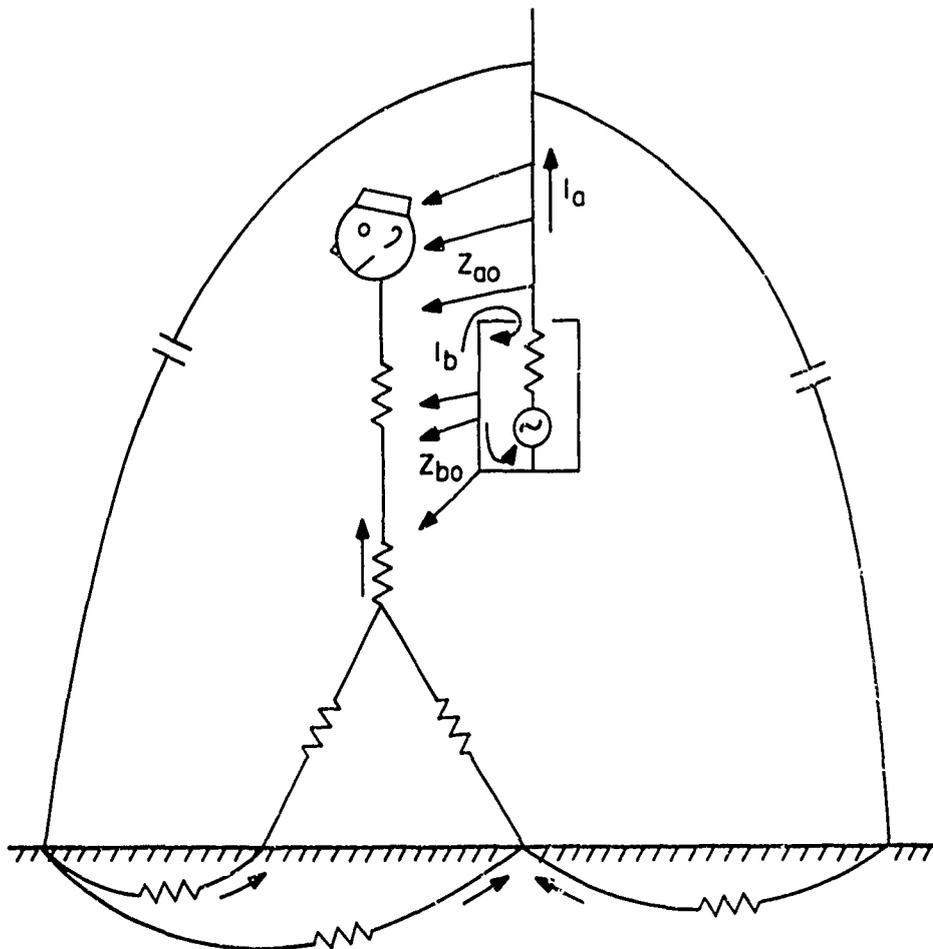
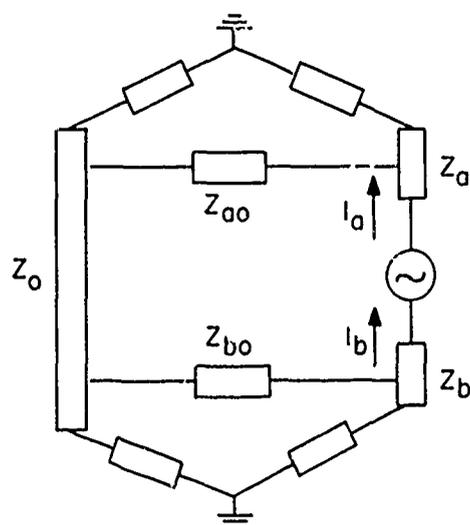


Figure A-2. Packset Whip Antenna Reactance



A. Pictorial Equivalent Diagram



B. Equivalent Circuit Diagram

IMPEDANCES:  
 $Z_o$  - OPERATOR  
 $Z_a$  - ANTENNA  
 $Z_b$  - RADIO BOX

Figure A-3. Packset Operator Circuits

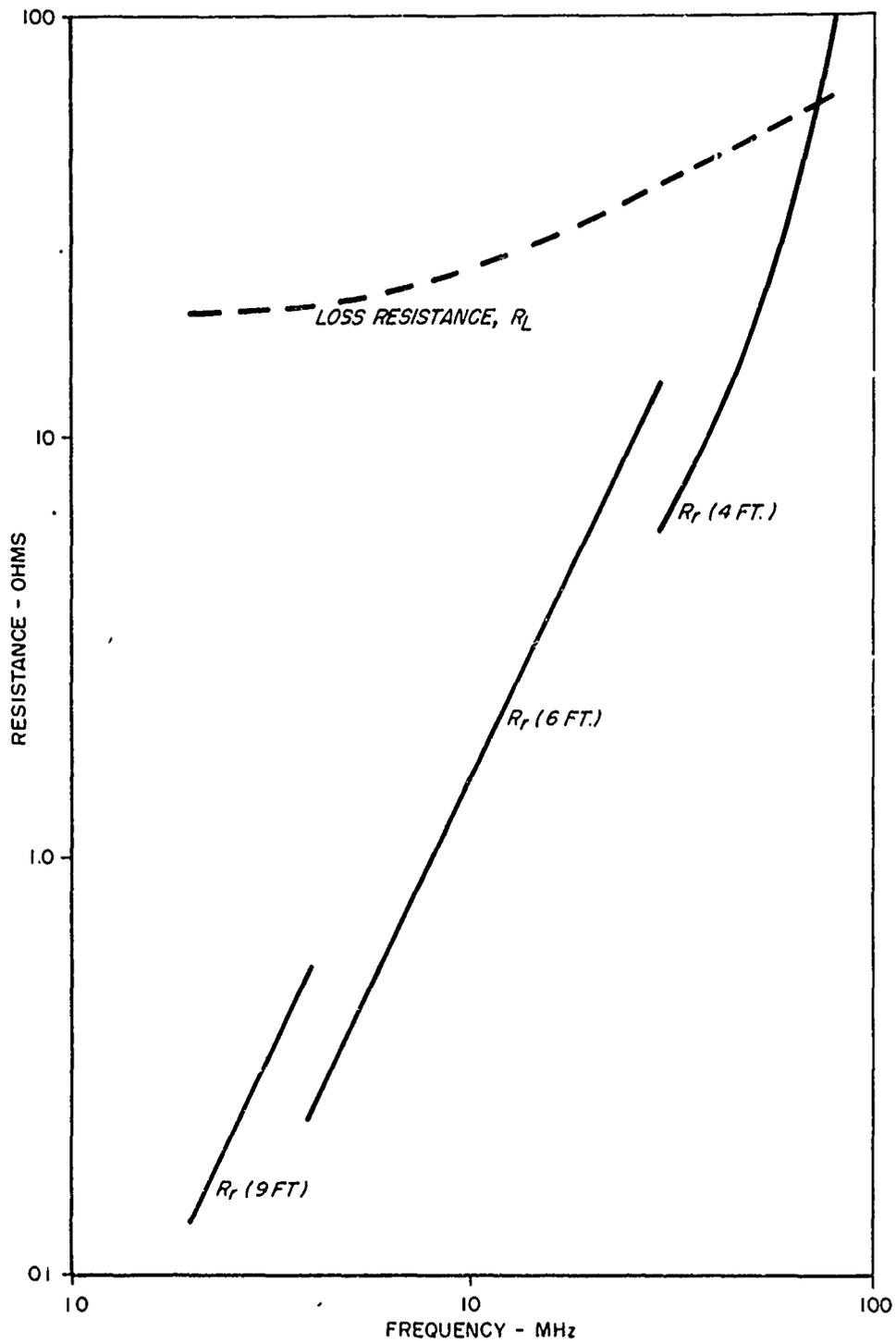


Figure A-4. Radiation Resistance and Loss Versus Frequency for Packset Whip Antennas

have been measured. These measurements were carried out in conjunction with development programs for the AN/PRC-42 (2 - 12 MHz), PT-5 (28 - 68 MHz), AN/PRC-70 (2 - 76 MHz), AN/TLQ-17, and Independent Research and Development Programs (IRAD) on electrically small antennas.

A previous packset antenna study was undertaken to determine the range of packset antenna impedances under actual operational conditions. This information was necessary in order to design the automatic antenna matching network that was developed for a new VHF radio. The measurement techniques developed for this task allow the antenna impedance to be determined in the normal field environment and without the influence of external measuring equipment and cables. All of the RF bridge circuitry, signal source, and DC battery supply was located within the packset housing. Measurements were made over the frequency range of 30 to 76 MHz and pertinent results for a 3-foot and 4-foot whip are given in table A-1.

#### A-5. EFFICIENCY AND RADIATION PATTERNS

An estimate of radiation efficiency, assuming the transmitter is matched to the antenna impedance, can be made by use of figure A-4, which shows radiation resistance and typical loss resistance. The estimated radiation efficiency ranges from about 1 percent at the lower frequencies to more than 50 percent above 30 MHz. It is important to note that even above 30 MHz the efficiency may be very low (i.e., 5 to 10 percent) if the antenna is coupled heavily to the operator's body. In this situation, the operator's body may be dissipating most of the RF power and communication range will be appreciably less than expected. Also, some hazard may exist to the operator if the transmitter power level is more than a few watts.

The rather heavy coupling between radio antenna and operator will also result in some radiation pattern distortion. The human body has a resonance in the range of 60 to 80 MHz and acts as an antenna director below this frequency. That is, the radiation pattern will be strongest in front of the operator if the packset is carried on the operator's back. The front-to back ratio may be as high as several db at frequencies around 40 to 60 MHz due to this effect. The magnitude of the effect depends upon the operator size and antenna-to-operator coupling. In general, it is far better to reduce this coupling by raising the whip or using a higher frequency where the coupling is less and the overall radiation efficiency will be greater.

Table A-1. Mea

TEST CONDITIONS	FREQUENCY (MHZ)	NO. 1		NO. 2		NO. 3		NO. 4		NO. 5		NO. 6		M C S 3 4 S E A C
		OPERATOR STANDING ON DAMP GROUND WITH HELMET ON (OPEN FIELD)		OPERATOR STANDING ON DAMP GROUND WITH HELMET OFF (OPEN FIELD)		OPERATOR STANDING ON DRY GROUND WITH HELMET ON (OPEN FIELD)		OPERATOR STANDING ON DRY GROUND WITH HELMET OFF (OPEN FIELD)		OPERATOR STANDING WITH BACK 2 FT FROM VEHICLE, DRY GROUND AND HELMET ON		OPERATOR STANDING 3 FT FROM 40 FT POLE STRUCTURE, DAMP GROUND AND HELMET ON		
		R	X	R	X	R	X	R	X	R	X	R	X	
MEASURED IMPEDANCE OF WHIP ANTENNA 3/32 IN. DIA x 3 FT LONG	30	49	-397	51	-417	50	-407			64	-455	54	-437	
	33	28	-319	31	-359	24	-279			28	-319	21	-299	
	37	23	-264	21	-244	24	-274			25	-284	28	-268	
	42	118	-279	93.0	-286			81.1	-237	121	-186	72.7	-260	7
	50	113	-195	113	-196	118	-204	114	-206	108	-187	116	-193	1
	56	29.8	-119	36.0	-118	35.8	-125	33.6	-110	33.6	-110	29.5	-118	3
	60	16.4	-104	18.4	-104	14.3	-102	23.4	-101	21.8	-103	36.2	-135	1
	65	12.1	-86.2	13.5	84.9	12.4	-88.1	14.9	-84.7	17.9	-84.1	17.4	-89.3	1
	70	8.97	-46.1	12.6	-50.5	9.4	-53.2	8.8	-55.3	12.8	-51.4	17.9	-52.0	1
	76	24.8	-3.5	18.4	-4.6	20.9	-1.5	23.6	-4.2	25.7	-4.1	19.8	-3.1	2
MEASURED IMPEDANCE OF WHIP ANTENNA 3/32 IN. DIA x 4 FT LONG	30	26	-299	29	-329	29	-329			38	-308	33	-318	
	33	21	-239	35	-252	18	-254			30	-248	31	-253	
	37	23	-189	20	-189	30	-213			21	-204	20	-229	
	42	57.8	-204	54.1	-203			30.3	-179	37.6	-164	37.5	-198	3
	50	69.8	-92.6	73.0	-90.1	71.6	-91.4	72.4	-89.4	83.4	-86.3	70.5	-97.1	7
	56	66.2	-30.9	66.2	-30.9	67.7	-34.5	63.0	-34.9	85.3	-34.5	64.5	-52.2	6
	60	32.8	-37.7	33.1	-32.0	39.6	-41.0	38.2	-41.0	41.3	-36.1	36.7	-35.4	4
	65	40.1	+36.1	36.7	+42.3	46.3	-34.9	53.0	+28.2	54.6	+31.5	53.9	+28.6	4
	70	72.3	+148	65.1	+146	73.1	+143	71.5	+146	95.3	+147	75.7	+126	8
	76	373	+302	336	+314	309	+299	347	+302	316	+229	320	+288	3

NOTE: All values

48.1

48.2

Table A-1. Measured Terminal Impedances of Small-Diameter Whip Antenna

FREQ (MHz)	NO. 4		NO. 5		NO. 6		NO. 7		NO. 8		NO. 9		NO. 10		NO. 11	
	OPERATOR STANDING ON DRY GROUND WITH HELMET OFF (OPEN FIELD)		OPERATOR STANDING WITH BACK 2 FT FROM VEHICLE, DRY GROUND AND HELMET ON		OPERATOR STANDING 3 FT FROM 40 FT POLE STRUCTURE, DAMP GROUND AND HELMET ON		OPERATOR STANDING 2 FT FROM 40 FT POLE STRUCTURE, DAMP GROUND AND HELMET OFF		OPERATOR PRONE ON DRY GROUND WITH HELMET ON, ANTENNA VERTICAL		OPERATOR PRONE ON DRY GROUND WITH HELMET ON, ANTENNA HORIZONTAL		OPERATOR KNEELING ON DAMP GROUND (WITH HELMET ON, ANTENNA ABOUT 60° ANGLE WITH GROUND)		RADIO SET SLITTING ON DAMP GROUND	
	R	X	R	X	R	X	R	X	R	X	R	X	R	X	R	X
707			64	-455	54	-437			32	-308	28	-319	67	-344	44	-357
79			28	-319	21	-299			19	-269	24	-279	28	-319	40	-378
74			25	-284	28	-268			23	-269	20	-229	23	-259	22	-249
	81.1	-237	121	-186	72.7	-260	72.9	-251	59.8	-224	66.1	-220	77.5	-220	12.0	-238
74	114	-206	108	-187	116	-193	121	-201	103	-171	122	-196	122	-174	93.0	-175
75	33.6	-110	33.6	-110	29.5	-118	36.5	-110	18.2	-103	37.7	-116	43.0	-112	31.7	-111
702	23.4	-101	21.8	-103	36.5	-135	18.4	-104	14.2	-89.9	20.2	-87.7	29.2	-102	33.9	-118
75.1	14.9	-84.7	17.9	-84.1	17.4	-89.3	18.7	-88.0	25.4	-66.3	27.7	-68.6	29.4	-76.6	29.3	-69.0
72	8.8	-55.3	12.8	-51.4	17.9	-52.0	17.6	-51.1	36.2	-28.3	22.9	-31.6	33.4	-39.8	40.6	-27.4
75	23.6	-4.2	25.7	-4.1	19.8	-3.1	22.4	+1.6	51.2	+22.8	45.7	+20.3	52	0	73.7	+35.9
79			38	-308	33	-318			29	-218	23	-259	26	-299	61	-314
74			30	-248	31	-253			19	-220	15	-214	18	-259	34	-328
73			21	-204	20	-229			13	-184	34	-177	37	-191	32	-182
	30.3	-179	37.6	-164	37.5	-198	33.1	-202	45.7	-150	17.9	-14.5	39.1	-157	23.4	-132
74	72.4	-89.4	83.4	-86.3	70.5	-97.1	76.4	-105	85.2	-59.7	93.3	-58.3	98.1	-85.3	86.7	-67.7
74.5	63.0	-34.9	85.3	-34.5	64.5	-52.2	63.9	-48.1	56.8	-35.5	97.3	-11.9	110	-21.4	56.7	-44.3
71.0	38.2	-41.0	41.3	-36.1	36.7	-35.4	43.8	-39.5	79.0	0	75.0	-1.3	69.5	-28.1	70.8	-21.6
74.9	53.0	+28.2	54.6	+31.5	53.9	+28.6	43.7	+24.2	84.3	+70.7	62.3	+71.7	82.7	+64.6	59.8	+73.8
74.8	71.5	+146	95.3	+147	75.7	+126	82.5	+143	221	+167	151	+208	135	+122	171	+154
79.9	347	+302	316	+229	320	+288	307	+329	530	+103	512	+272	404	+147	537	+56

NOTE: All values in ohms

48.2

48.3

Table A-1. Measured Terminal Impedances of Small-Diameter Whip Antenna

TEST CONDITIONS	NO 1		NO 2		NO 3		NO 4		NO 5		NO 6		NO 7		NO 8		NO 9		NO 10		NO 11		
	R	X	R	X	R	X	R	X	R	X	R	X	R	X	R	X	R	X	R	X	R	X	
MEASURED IMPEDANCE OF WHIP ANTENNA 3/32 IN DIA x 3 FT LONG	30	49	397	51	-417	50	107		61	-455	54	-437			32	-308	28	319	67	311	43	157	
	33	28	319	31	359	24	279		28	319	91	299			19	269	24	279	28	319	40	158	
	37	23	264	21	244	24	274		25	244	28	268			23	269	20	229	21	249	22	219	
	42	118	279	93	246			81	237	121	186	72	260	72	251	59	221	66	220	77	220	12	218
	50	113	195	113	-196	118	204	111	206	108	187	116	193	121	201	107	171	122	196	122	171	91	177
	56	298	-119	160	118	358	125	336	110	316	110	295	118	365	-110	103	103	37	116	130	112	31	111
	60	164	-101	181	104	143	102	231	101	218	103	362	135	184	101	142	309	26	277	212	162	139	118
	65	121	862	135	819	121	881	149	847	179	841	174	893	187	880	254	863	277	846	294	766	291	196
	70	897	-461	126	505	94	532	148	553	128	514	179	520	176	511	362	543	229	316	111	108	106	274
	76	248	35	181	16	209	15	216	42	257	41	198	-31	224	116	312	128	457	203	52	0	777	159
MEASURED IMPEDANCE OF WHIP ANTENNA 3/32 IN DIA x 4 FT LONG	30	26	299	29	-320	29	329		38	308	33	318			29	278	23	239	26	291	61	311	
	33	21	239	35	252	18	254		30	248	31	253			19	220	15	211	18	229	31	328	
	37	23	189	20	189	30	213		21	201	20	229			13	184	31	172	37	191	32	187	
	42	578	204	541	-203			30	179	37	164	37	198	33	202	45	150	179	115	79	157	23	172
	50	698	926	730	901	716	914	724	894	834	863	705	871	764	105	852	597	933	583	981	853	867	677
	56	662	-309	662	309	677	315	630	349	853	345	645	522	639	481	568	355	973	119	110	214	567	113
	60	328	-377	331	-320	396	-410	382	-410	113	-361	367	-354	438	-395	790	0	750	13	695	281	708	216
	65	401	361	367	423	463	-349	530	282	516	315	539	286	437	242	843	707	623	271	827	616	598	278
	70	723	148	651	146	731	143	715	146	953	147	757	126	825	143	221	167	151	208	135	122	171	151
	76	373	302	336	314	309	299	347	302	316	229	320	288	307	329	530	103	512	272	104	147	537	156

NOTE: All values in ohms

48-1

48-2

48-3

## APPENDIX B

### DESCRIPTION OF 30 - 80 MHZ COUPLER CIRCUITS

#### B-1 INTRODUCTION

A detailed description of operation of the circuits which comprise the 30 - 80 MHz Coupler can be found in Technical Report ECOM-0101-3. It is summarized here for purposes of easy reference.

#### B-2 TUNING NETWORK

A schematic diagram of the tuning network is shown in figure B-1. C1 and C2 are piezoelectric capacitors, while the inductors are either air core or powdered iron core types. A complete description of how the network values were determined is given in Technical Report ECOM-0101-3.

#### B-3 CAPACITOR CONTROL LOOP

##### B-3.1 STATIC CAPACITANCE VARIATIONS

The control loop has the capability of correcting for the "static" drift of the capacitor. The magnitude of the correction is determined by the open-loop DC gain. It was considered quite satisfactory for the electronically tuned coupler to have a correction factor of 50. Therefore, if the capacitor tends to drift 5 percent, the loop would correct in such a manner that the resulting capacitance variations would be only 0.1 percent. A correction factor of 50 is obtained by allowing the static (DC) open loop voltage gain to be 50.

##### B-3.2 DYNAMIC CAPACITANCE VARIATIONS

Exposing the piezoelectric capacitor to acceleration forces with sufficiently low frequency content to excite the fundamental mode of vibration of the Bimorph plates, which used as electrodes in the capacitor, results in a corresponding capacitance modulation. The natural frequency for the Bimorph elements used in the piezoelectric capacitor is approximately 250 Hz. Therefore, if the capacitor is vibrated at a frequency of 250 Hz and the mechanical Q of the Bimorph plates is assumed to be 10, then the peak deflection around the quiescent position is 10 times larger for the 250 Hz acceleration than for a static acceleration of the same size. For this reason, it is desirable to reduce the effect of mechanical vibrations for frequencies to the order of 500 Hz. This can be achieved by the capacitor control loop having an open-loop unity gain frequency above 500 Hz. The original control loop had a unity gain frequency of 1 KHz. The resulting correction to an excitation at 250 Hz was  $5 \cdot Q = 50$ , if a mechanical Q-value of 10 was assumed.

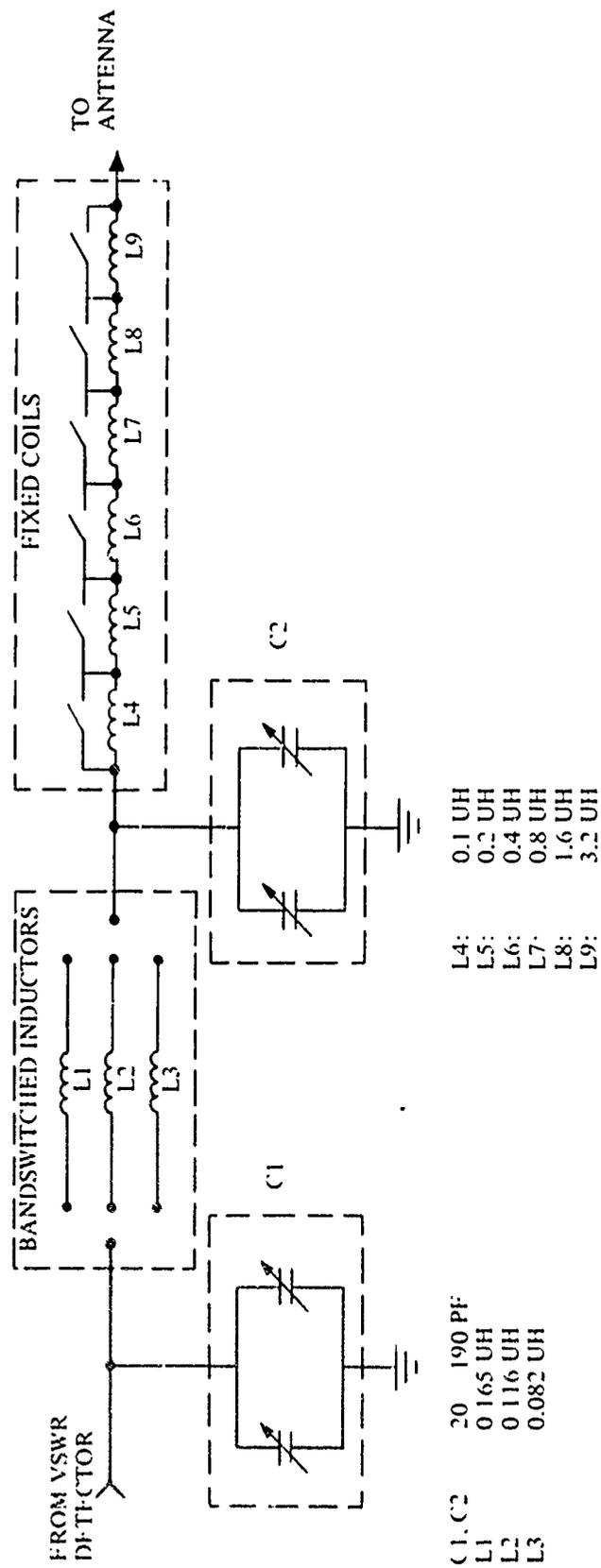


Figure B-1. Tuning Network Schematic Diagram

### B-3.3 TIME RESPONSE

The deflection of the Bimorph element and thus the capacitance value is dependent upon the voltage applied to the brass vane, which is physically located between the piezoelectric ceramic plates constituting the Bimorph element. The piezoelectric capacitor transfer function, capacitance versus input control voltage, has approximately a second-order low pass characteristic. The time response for such a system to a step change in control voltage is well known and can be obtained from most standard text books in control systems.

Time and frequency response measurements with various piezoelectric capacitors have verified the low pass transfer characteristics with reasonable accuracy. The simple theory given here is slightly complicated by the presence of higher-order vibrational modes, which do not to any great extent modify the step response of the piezoelectric capacitor.

Piezoelectric capacitors with various mechanical configurations have been tested. The mechanical  $Q$  value of the Bimorph elements has been found to vary from 2.5 up to 30. The fundamental self-resonant frequency for all capacitors measured has been approximately 250 Hz, which is determined by the length dimension (1.25 inch) of the Bimorph element. The variations in  $Q$  value will give rise to significant deviations in step response for the various capacitor configurations.

The time response to step input can be written:

$$\Delta C = \left[ 1 - \frac{c \cdot \omega_n t / 2Q}{\sqrt{1 - 1/4Q^2}} \sin \left[ \omega_n \sqrt{1 - 1/4Q^2} t + \cos^{-1} \frac{1}{2Q} \right] \right] \quad (\text{B-1})$$

where:

- $\omega_n$  = natural frequency,
- $Q$  = mechanical  $Q$  value
- $\Delta C$  = static capacitance change corresponding to the control voltage change.

The second term in the expression for the step response is the error term. The settling time for 19 percent error is roughly, for large  $Q$ ,  $t \approx \frac{4Q}{\omega_n}$ , which for  $Q = 30$  and  $\omega_n = 500$  radians/sec gives  $t = 75$  msec.

For  $Q = 2.5$ , the settling time can be estimated as 6 msec.

A control loop with a unity gain frequency sufficiently higher than 250 Hz and a phase margin in the order of  $45^\circ$  will substantially reduce the settling time. The original loop had a unity gain frequency of 1 KHz with a phase margin near  $15^\circ$ . The resulting settling time for 10 percent error was less than 1 msec when using a capacitor with an open-loop settling time of about 10 msec. It was desirable from a system standpoint to reduce the response time as defined here to at least 2 msec.

It should be noted that the loop design reducing dynamic capacitance variations discussed previously is in agreement with the design criteria for fast response. However, an additional design constraint, namely large phase margin, is required for fast settling time.

## B-4 BASIC OPERATION

The pilot capacitor provides the means for regulating the main capacitor without interfering with the RF path through the main capacitor. A block diagram of the closed loop is shown in figure B-2. Here a certain DC voltage from the D/A converter corresponds to a certain pilot (and main) capacitor.

The frequency used to measure the pilot capacitor was chosen to be 1 MHz. This frequency was selected because of certain undesirable low frequency coupling effects caused by stray capacitances between brass vane and pilot capacitor. The 1 MHz bandpass filter reduces the undesired low frequency content of the 1 MHz signal before entering the envelope detector. Also, VHF effects present at the pilot capacitor due to leakage from the main capacitor are attenuated by the band pass filter.

The lead-lag filter shown in figure B-2 advances the phase at the unity gain frequency of the loop. This increases the stability of the loop and also reduces the ringing. A Bode plot of the open loop gain characteristics is shown in figure B-3. Here an idealized capacitor transfer is assumed. The mechanical Q of the capacitor is assumed to be equal to one. The 12 db/octave slope above the natural frequency  $\omega_n$  is motivated by the assumption of a second-order capacitor transfer function.

The lead-lag filter has a 20 db attenuation depth and the phase advance at the unity gain frequency,  $4 \omega_n$ , is approximately  $50^\circ$ . The capacitor transfer phase at this frequency is very close to  $-180^\circ$ . Consequently, the loop phase margin is in the order of  $50^\circ$ , which guarantees excellent transient behavior and thus short settling time.

## B-5 LOW POWER DESIGN

Figure B-4 shows the block diagram of the low power control loop. The following characteristics of this loop apply:

- Total power consumption is less than 40 mW.
- The Avco developed output stage is capable of  $\pm 50V$  output voltage swing to cover the full range of the piezoelectric capacitor. It should be noted that the transistors are biased such that the stage responds also to very small AC signals.
- Only two low-pass filter sections following the envelope detector are required.
- A 1.5 KHz notch filter inserted to eliminate the effects of higher-order modes of Bimorph vibration on the stability of the loop is shown schematically immediately following the Fairchild micro-power operational amplifier  $\mu A 735$ .
- The noise voltage generated in the control loop and fed to the brass vane (and thus modulating the capacitor) is reduced in two ways. First, the filtering for frequencies above 1 KHz is increased. Secondly, the total amplifier gain has been reduced, which has been made possible by a corresponding increase in envelope detector sensitivity.

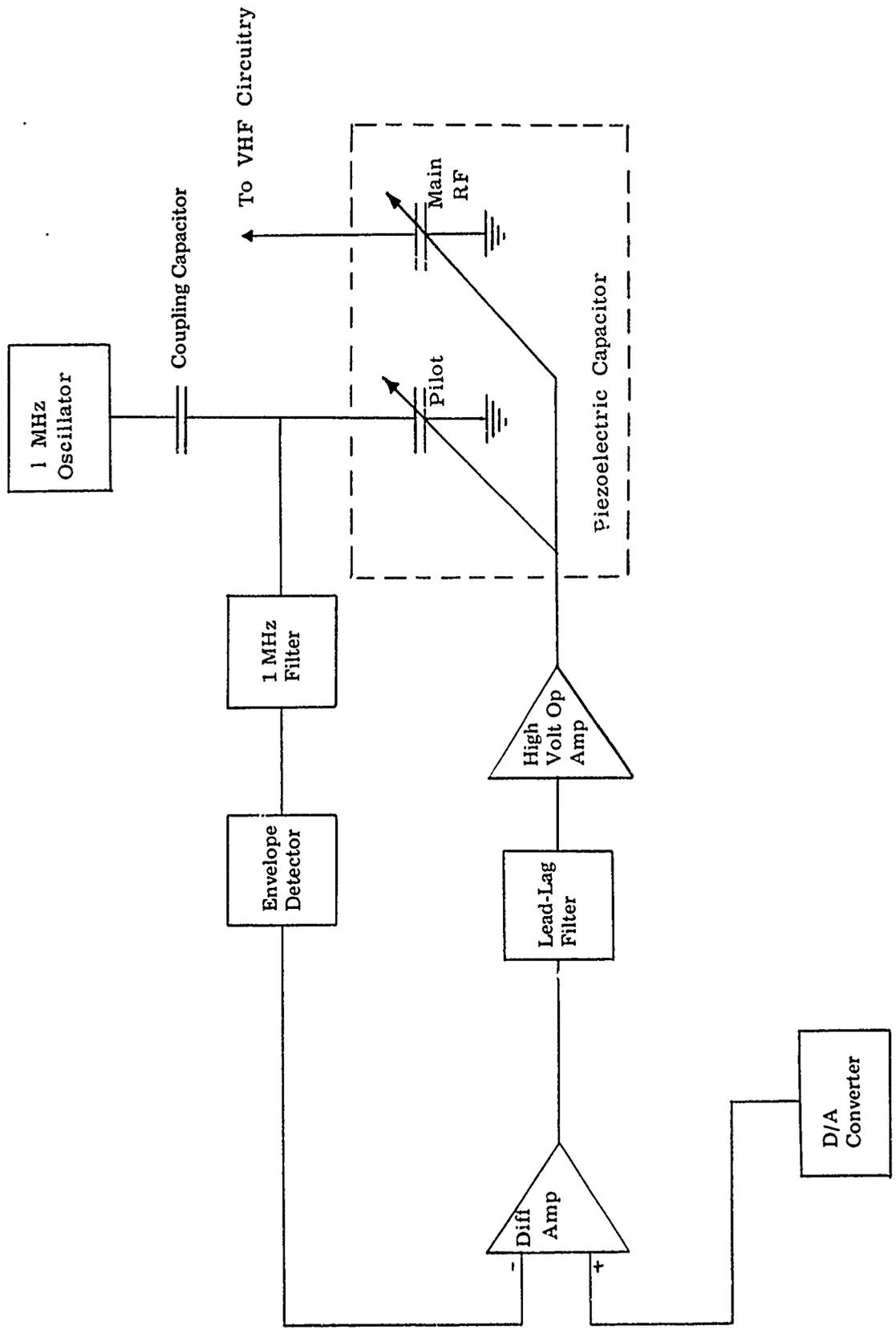
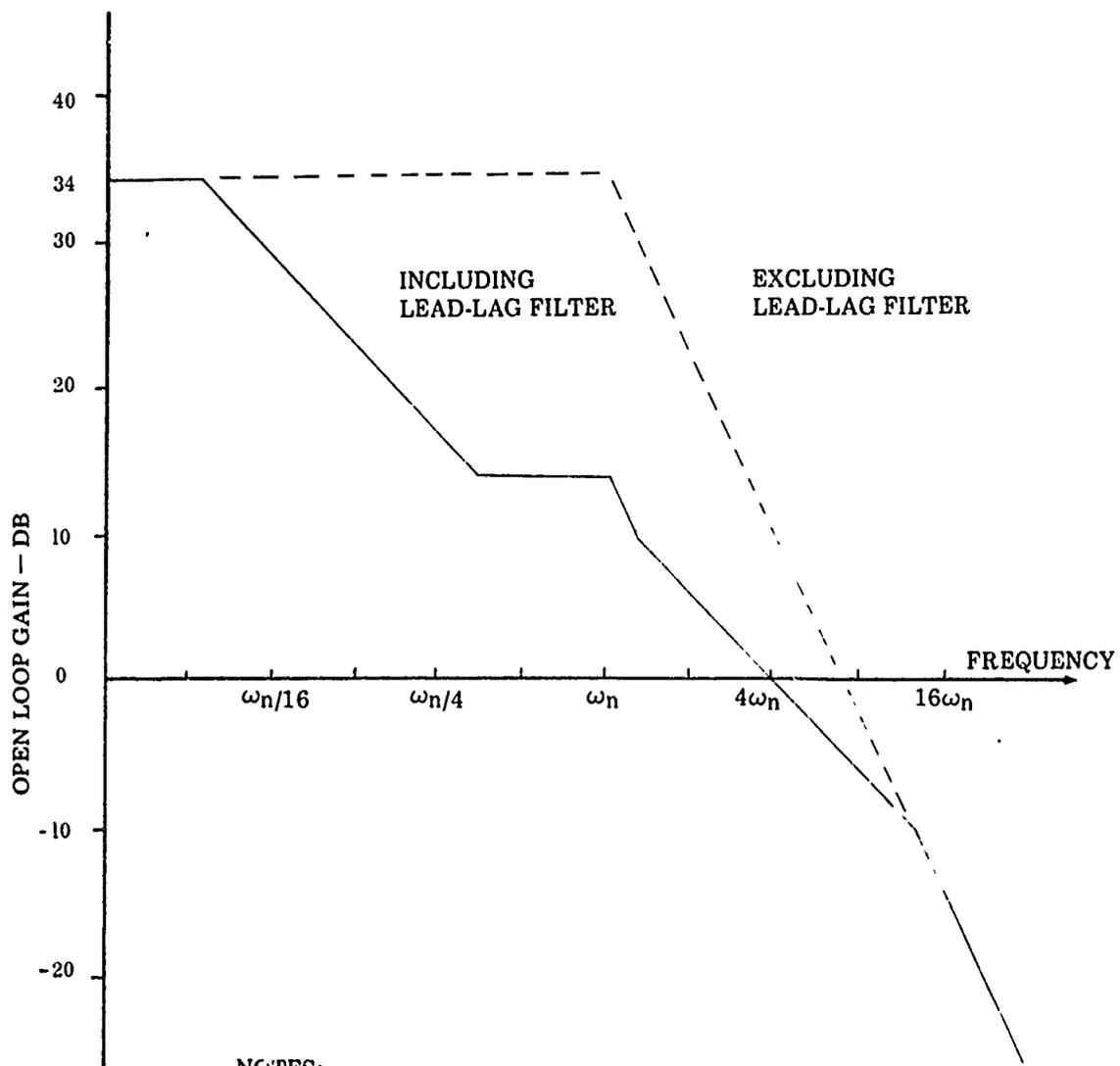


Figure B-2. Piezoelectric Capacitor Control Loop Block Diagram



**NOTES:**

1. Assuming idealized capacitor with mechanical Q equal to 1.
2. In practice  $\omega_n$  corresponds to 250 Hz.
3. Theoretical phase margin:  
     50 with lead-lag filter  
     0 without lead-lag filter

Figure B-3. Bode Plot of Open-Loop Gain

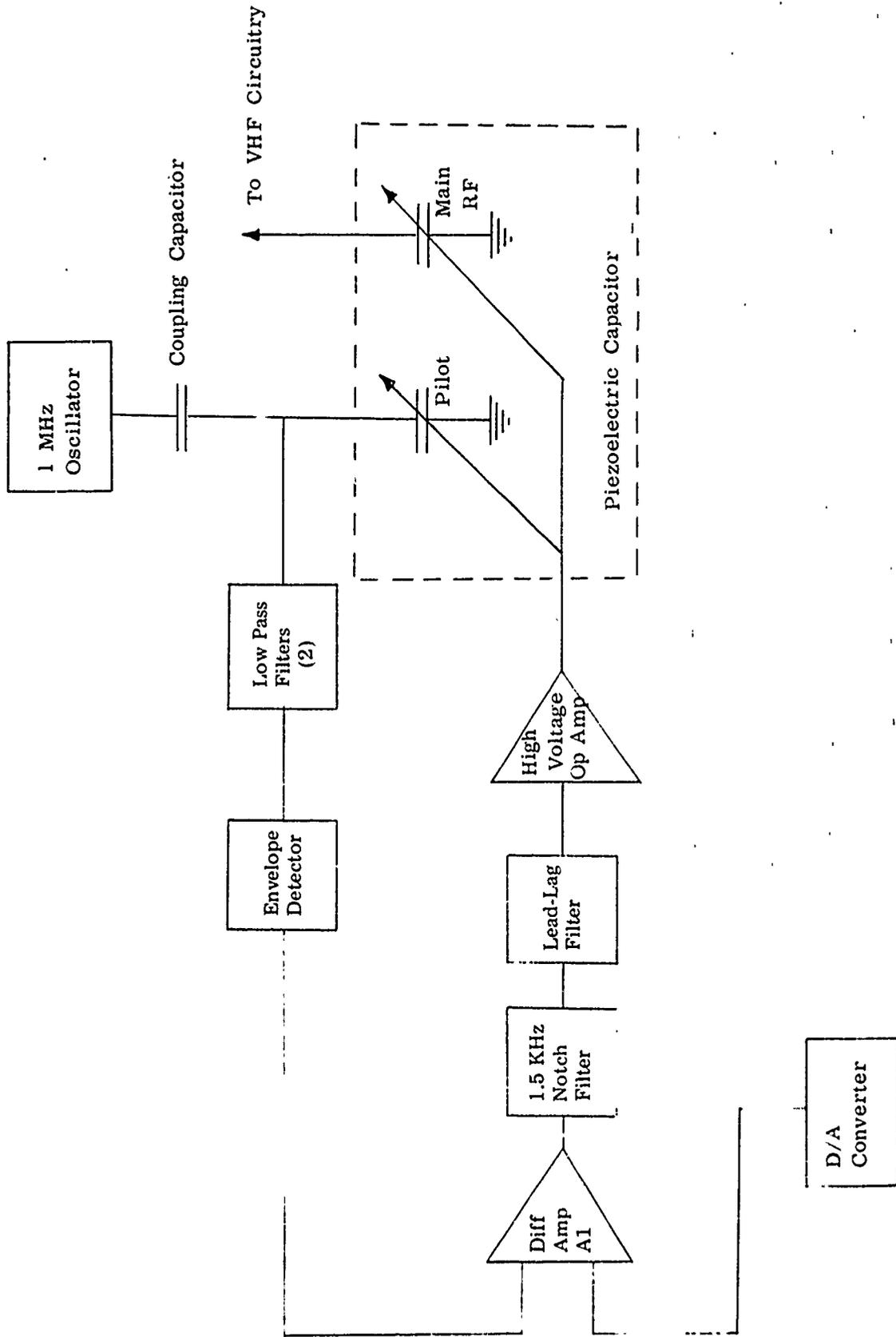


Figure B-4. Low Power Control Loop Block Diagram

## B-6 DIGITAL VSWR DETECTOR

### B-6.1 GENERAL DESCRIPTION

The basic purpose of the VSWR detector is to provide information concerning the VSWR of the antenna coupler. Tuning is accomplished by changing an element value. Then the VSWR is compared to the previous value to see if the VSWR has increased, decreased, or remained the same. A decision is made in the interface logic, an element is changed, and again the VSWR is compared. This information is provided to the interface logic in binary form. There are seven-bits of binary information.

The basic method used to measure the VSWR is as follows: It is known that the directional coupler, when properly terminated, has the capability of extracting information concerning the forward and reflected power or voltage. This is due to its directional properties which can be derived mathematically. The coupled ports will provide the following:

$$\begin{aligned} V_2 &: kV_F \\ V_3 &: kV_R \end{aligned}$$

If a means is provided to ratio these two quantities, the following relationship will be true:

$$\frac{V_3}{V_2} = \frac{kV_R}{kV_F} = \frac{V_R}{V_F} = \rho V \quad (\text{B-2})$$

This ratio is known as the reflection coefficient and can be related mathematically to VSWR by the following:

$$\text{VSWR} = \frac{1 + \rho V}{1 - \rho V} \quad (\text{B-3})$$

The means utilized to make the division is a dividing analog-to-digital converter. This also provides a digital output which is compatible with the interface logic.

Figure B-5 is a block diagram of the VSWR detector. The directional coupler provides signals proportional to the forward and reflected power to the load. These signals are detected via a peak detector which also contains the proper time constants for high-speed operation. These signals are amplified to the proper level for use with the digital-to-analog converter (DAC) by means of the amplifier blocks. The DAC provides an analog signal at its output which is proportional to the reference signal and the digital binary input signals. This DAC is an eight-bit unit with only seven bits in use. This enables the analog output to provide N increments of the analog output, N being defined as:

$$N = 2^{\# \text{bits}} = 2^7 = 128 \text{ increments.} \quad (\text{B-4})$$

The storage provides temporary storage for each bit of the VSWR word. The sequencer provides the proper sequence of inputs to the storage. The comparator compares the analog output of the DAC with the processed reflected power signal. The comparator output controls the storage or clearing of each bit of information depending on the input of the sequencer.

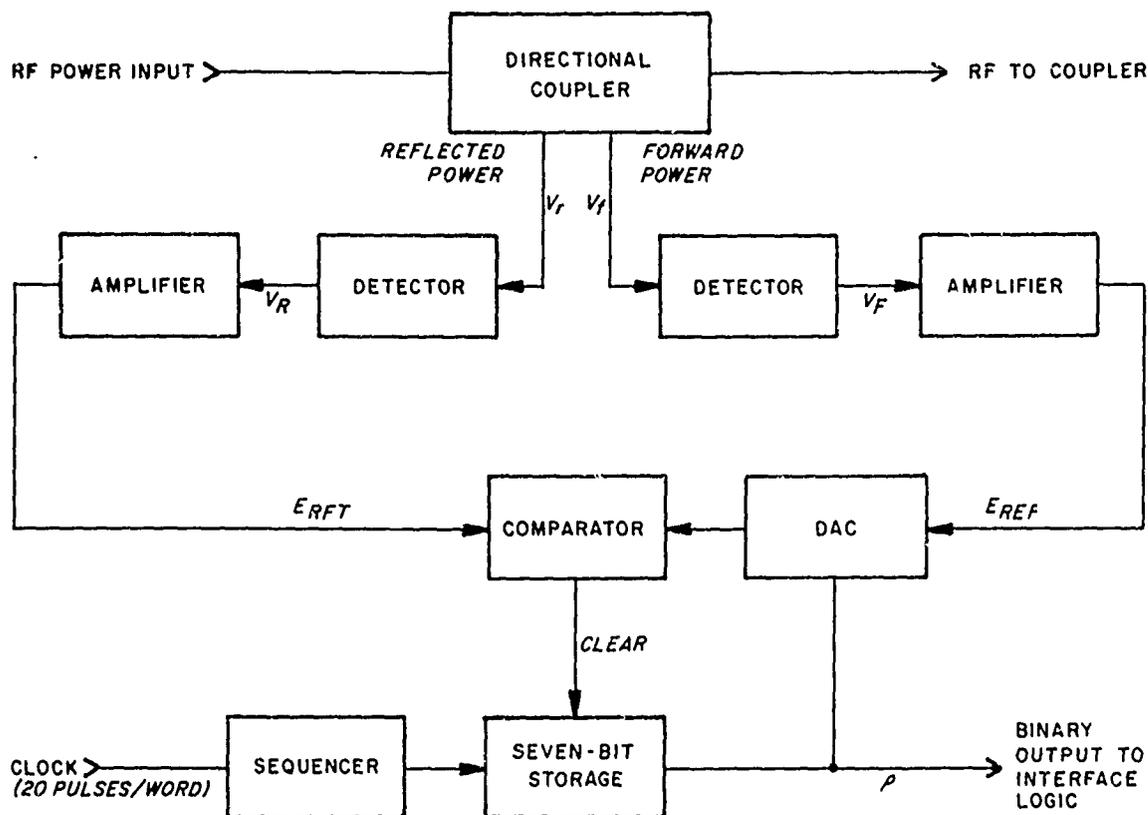


Figure B-5. VSWR Detector Block Diagram

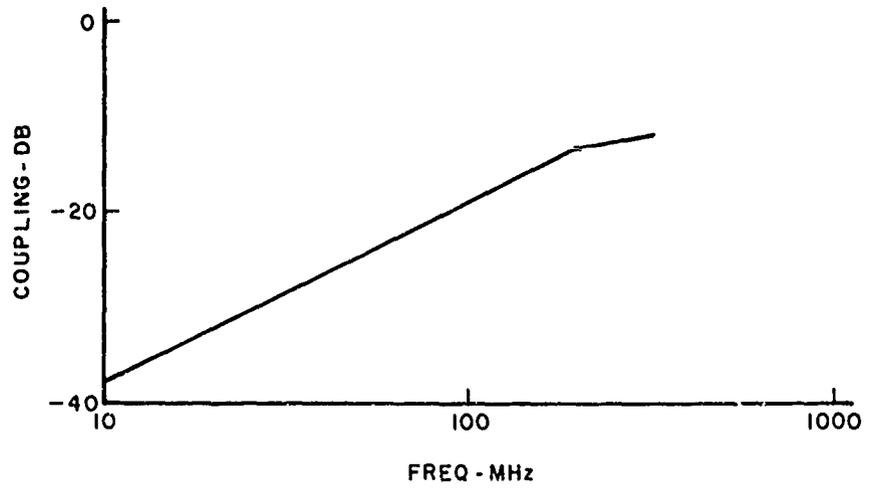
### B-6.2 DIRECTIONAL COUPLER

Stripline techniques were utilized in the design of the directional coupler. This technique provides:

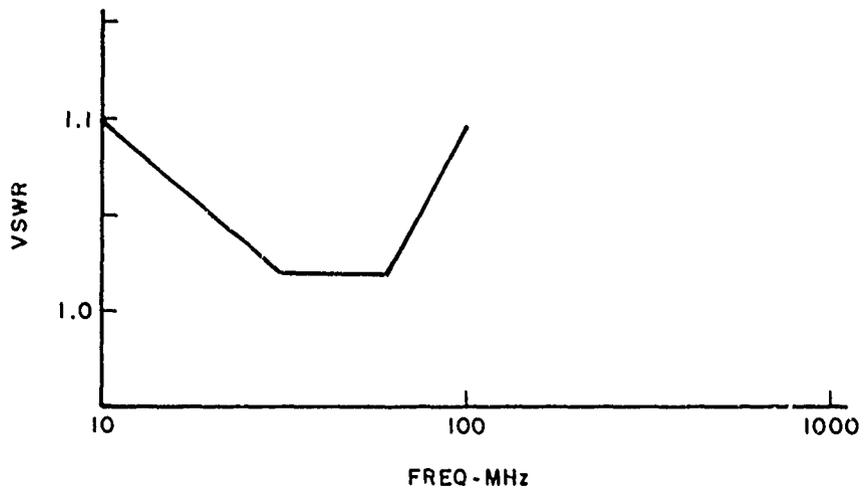
1. Reproducibility
2. Accuracy
3. Shielding from extraneous signals
4. Ease of construction

The physical directional coupler is 6 inches long, one inch wide, and approximately 1/4 inch thick. It is constructed from special printed circuit board. The dielectric constant and dimensions of this board are very closely controlled. Conventional printed circuit techniques are used to construct this board which is designed for a coupling factor  $K = 10$  db at 360 MHz, where its electrical length is  $90^\circ$  or 1/4 wavelength. Coupling at 80 MHz is -20 db and decreases to -28 db at 30 MHz. This gives very little insertion loss due to insertion of the directional coupler between the generator and the antenna matching network. Directivity is greater than -20 db. Figure B-6 is a graph of the critical parameters versus frequency. Coupling, VSWR, and isolation are displayed for each coupled line. An independent coupled line is used for the forward monitor and reflected monitor.

a. COUPLING



b. VSWR OF INPUT



c. ISOLATION

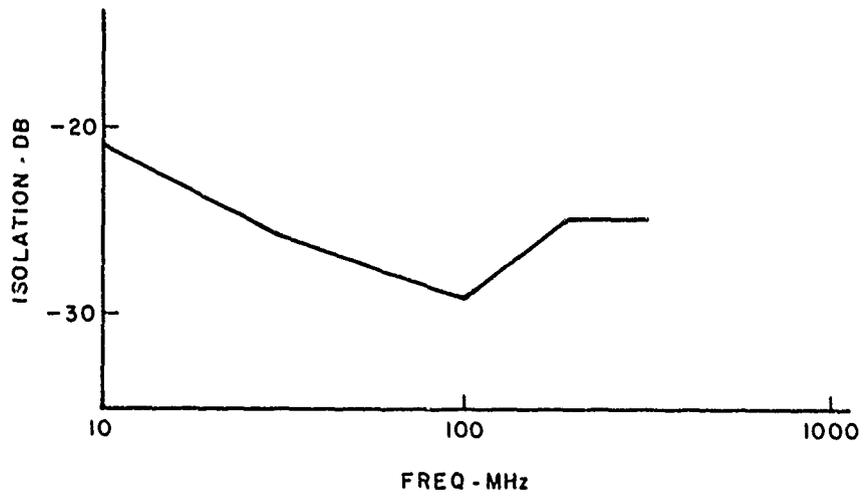


Figure B-6. Directional Coupler Parameters

### B-6.3 DETECTOR

Figure B-7 is a simplified version of the detector circuitry. This is a very simple peak detector circuit. The AC from the RF outputs is supplied to the diode. RF is peak detected. The decay time constant is  $t_d = R_d C_1$  and the attack time constant is  $t_a = r_d C_1$ , where  $r_d > R_d$  and can be neglected in its effect on  $t_a$ . The greatest linearity problem occurs in this portion of the design. Since  $\rho$  is a function of the ratio:

$$\rho V = \sqrt{\frac{P_r}{P_f}} = \frac{V_r}{V_f} \quad (\text{B-5})$$

where  $\rho V$  is only the absolute value of the reflection coefficient. The following must be true in order to detect a decrease or increase in  $\rho$  for all values of  $P_f$  and  $P_r$ .

$$\rho : \left( \frac{E_{RFT}}{E_{REF}} \right)^N$$

where  $E_{RFT}$  is the amplified and detected value of  $P_r$  and  $E_{REF}$  is the amplified and detected value of  $P_f$ . It can be proven that detection of changes in  $\rho$  and direction of change can be determined for  $N > 0$ . Therefore, the following is true:

$$V_r : (E_{RFT})^N$$

$$V_f : (E_{REF})^N$$

or

$$E_{RFT} = K_1 \sqrt{V_r} \quad (\text{B-6})$$

$$E_{REF} = K_2 \sqrt{V_f} \quad (\text{B-7})$$

where  $K_1 = K_2$  and are gain constants. It can be shown that the diode transfer curve must be:

$$V_F : (V_f)^n$$

where  $V_f$  is the RF input and  $V_F$  is the DC output. For a typical diode, it is possible to operate in a region where  $n = 1$  and where  $n = 2$ . The  $n = 2$  region was chosen for its greater operating region and its lower operating level. Figure B-3 is a typical transfer characteristic of an HP 5082-2824 diode. It is seen that the square law region ends at -20 dbm and the linear region begins. Therefore, the maximum input power to the diode is -20 dbm in order to remain in the  $n = 2$  range. This will occur at the forward power level of 10 watts.

### B-6.4 AMPLIFIER

Figure B-9 is a schematic diagram of the amplifier section of the design. The forward amplifier has a buffer stage since it has a very low output impedance requirement in order to interface with the reference line of the DAC. The gains of both amplifiers must be such as to meet the following requirements.

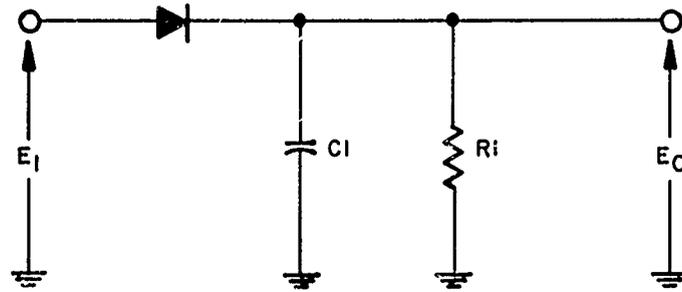


Figure B-7. Detector Circuitry

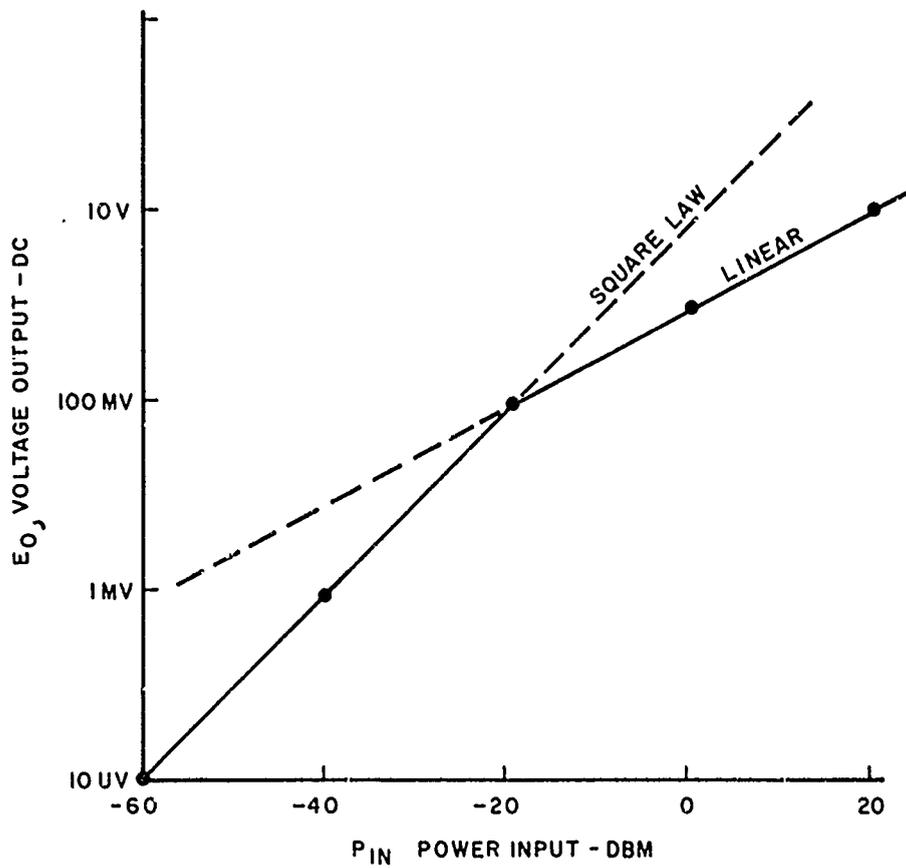
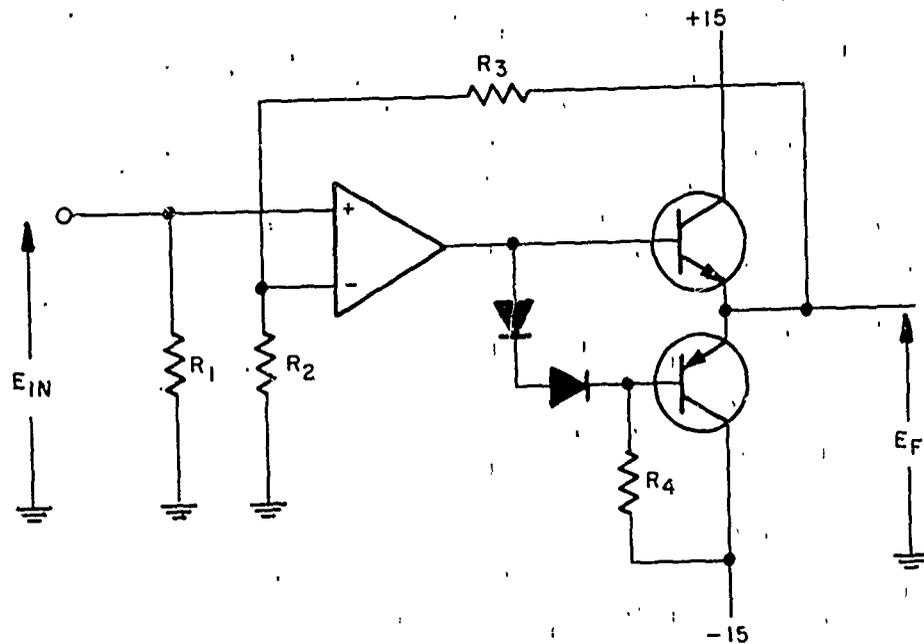
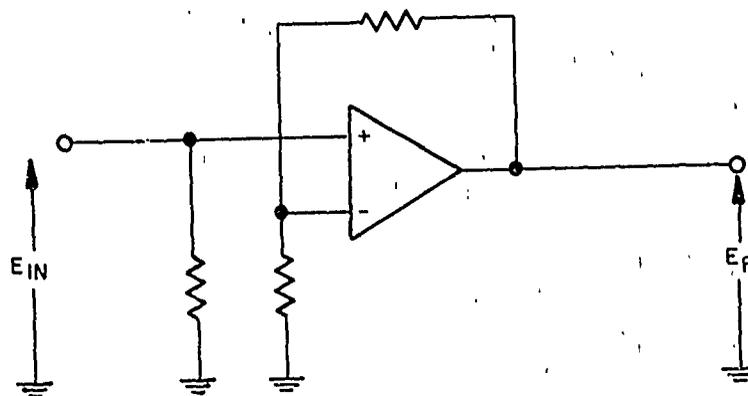


Figure B-8. Transfer Characteristic of HP 5082-2824 Diode



a. FORWARD AMPLIFIER



b. REFLECTED AMPLIFIER

Figure B-9. DC Amplifier Schematic Diagram

$$E_{REF} = E_{RFT}$$

when

$$\rho = 1.0 \text{ or } VSWR = \infty$$

and also

$$E_{REF} = 9.0 \text{ volts DC}$$

when

$$P_F \text{ (to antenna coupler) } = 10.0 \text{ watts.}$$

The output impedance of the  $E_{REF}$  amplifier must be less than five ohms. The slew rate of both amplifiers are compatible with the speed of operation.

### B-6.5, D/A CONVERTER

Three types of digital-to-analog converters were investigated during this program. They are the Precision Monolithic DAC-01 and the Micro-Networks MN302 and MN829 versions. Table B-1 is a comparison of critical parameters of the three units.

Table B-1. Comparison of Digital-To-Analog Converters

UNIT	BITS	SETTLING TIME	POWER CONSUMP.	ACCURACY	REFERENCE
MN302	8	3 $\mu$ sec	400 mW	$\pm 1/2$ LSB	External
MN829	8	3 $\mu$ sec	400 mW	$\pm 1/2$ LSB	External
DAC-01	7	3 $\mu$ sec	250 mW	$\pm 1/2$ LSB	Trim

The DAC-01 was found to be unsuitable for variable reference usage because elaborate interface circuits are required for linear operation. The MN302 was found to be very suitable operating over a variable reference range of 4 to 9 volts although the MN302 is designed to be operated as a fixed reference device.

The MN829 is a modification of the MN302. Its primary deviation from the MN302 is that it is specifically designed to operate over a reference variation of 4 to 9 volts. Satisfactory results were derived from the MN302.

### B-6.6 A/D TECHNIQUES

There are many analog-to-digital conversion techniques available for use, some of which are.

1. Tracking method
2. Ramp method
3. Successive approximation method

The successive approximation method was utilized in this design. Although it requires more parts than the other methods, it is the fastest conversion method, Figure B-10 is a simplified version of this method.

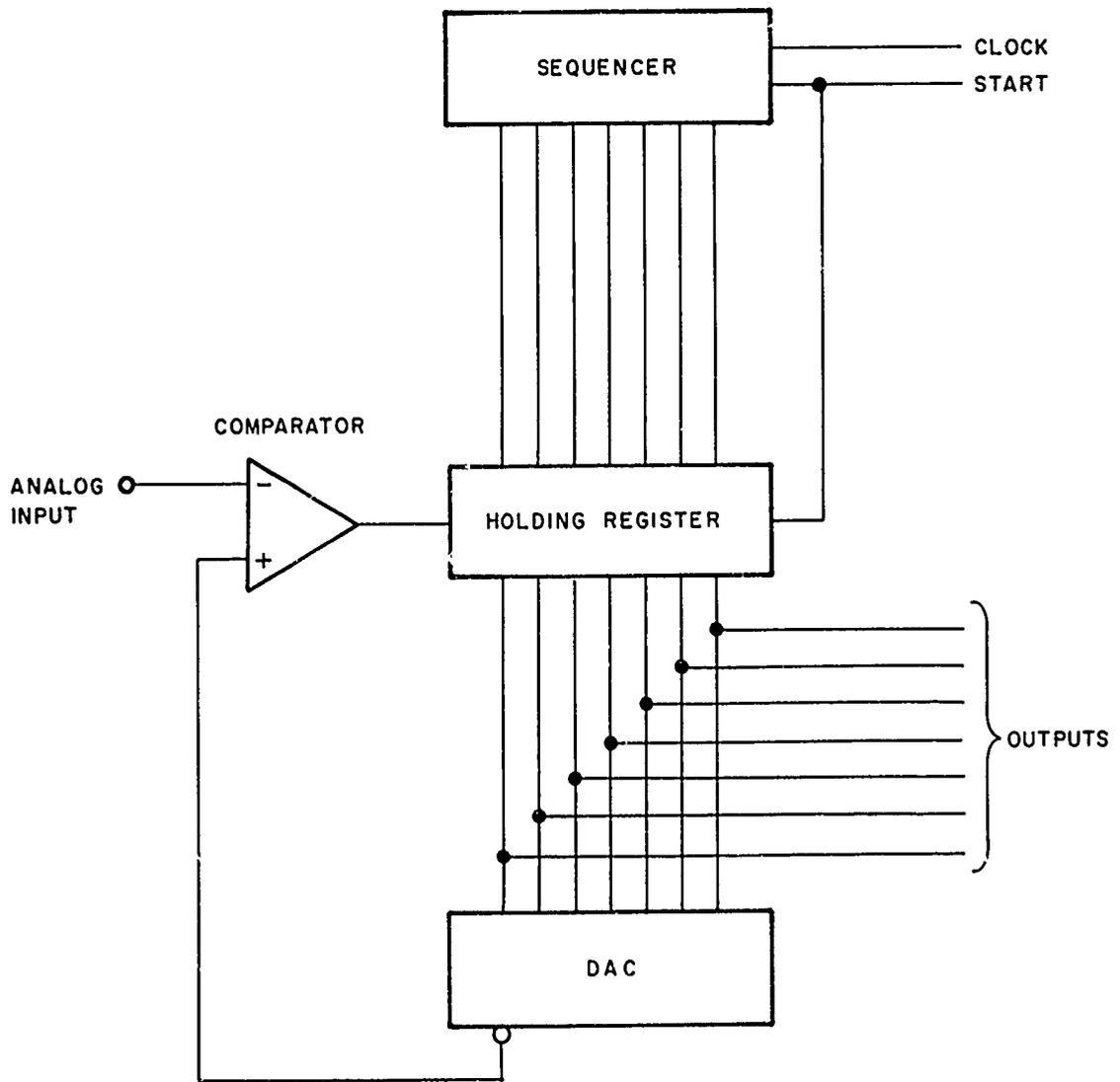


Figure B-10. A/D Conversion Technique

## B-6.7 STORAGE

Bit storage is accomplished by means of a data flip-flop. There is one storage register for each bit of information. Figure B-11 is a typical storage register with associated circuitry.

The data flip-flop used in the breadboard is a Motorola MC 7479. The flip-flop triggers on the positive edge of the clock pulse. During the clock transition, the state of the input, D, is transferred to the Q output, which is pin 13. The set input is used to clear the "VSWR word" before approximating a new input to the VSWR detector. The "clock in" line receives pulses from the interface logic. The sequence bit input is derived from the sequencer. This input is "high" only when the storage register in question is being worked. The comparator input is either "high" or "low". A "high" comparator input indicates that the analog output of the DAC is lower than the signal being approximated. The "low" input is the opposite. The MC 1810 is used to gate the clock pulses to the flip-flop only at the proper time interval. The sequencer input is double the period of the clock pulses. The MC 1810 is a Quad-Two input "Nor" gate. The truth table for the MC 1810 is as follows:

INPUTS		OUTPUT
9	10	8
L	L	H
L	H	L
H	L	L
H	H	L

A typical operational cycle of a single register would be as follows. The timing diagram for this cycle is included in figure B-12. At  $t_1$  the sequencer input goes "low". This indicates that this register is to be worked. Also, the "clock input" goes "low." Therefore, the "clock" goes high, causing data on line D to be transferred through the flip-flop. This bit is now present on one of the binary inputs of the DAC, causing an increase in the analog output of the DAC. Information as to whether the bit is too large or too small is transferred back via line D. If line D is "high" at  $t_3$ , when the data is transferred through the flip-flop, the bit will remain. This is true of this example. At time  $t_3$  the sequencer works on the next register in the sequence. The sequencer line to this flip-flop is "high" and clock pulses no longer reach the data flip-flop. The data remains in the register until the "clear word" line goes "low" and resets the register. If the comparator input would have gone "low," then at  $t_3$  the low would have been transferred through the register.

## B-6.8 SEQUENCER

The purpose of the sequencer is to apply pulses to the storage register in the proper time sequence. In this manner the "VSWR word" can be approximated by the successive approximation method. This method requires the successive placement of each bit into each storage register, beginning with the most significant bit (MSB) and ending with the least significant bit (LSB). At the time each bit is placed into the storage register, a decision is made whether to keep the bit or remove it.

The sequencer is constructed of an AN-5490 and a MC 8301. The former is a decade counter and the latter is a BCD-to-decimal decoder. The decimal output of the MC 8301 is applied to each storage register. The output of the decade counter is applied to the BCD input of the MC 8301.

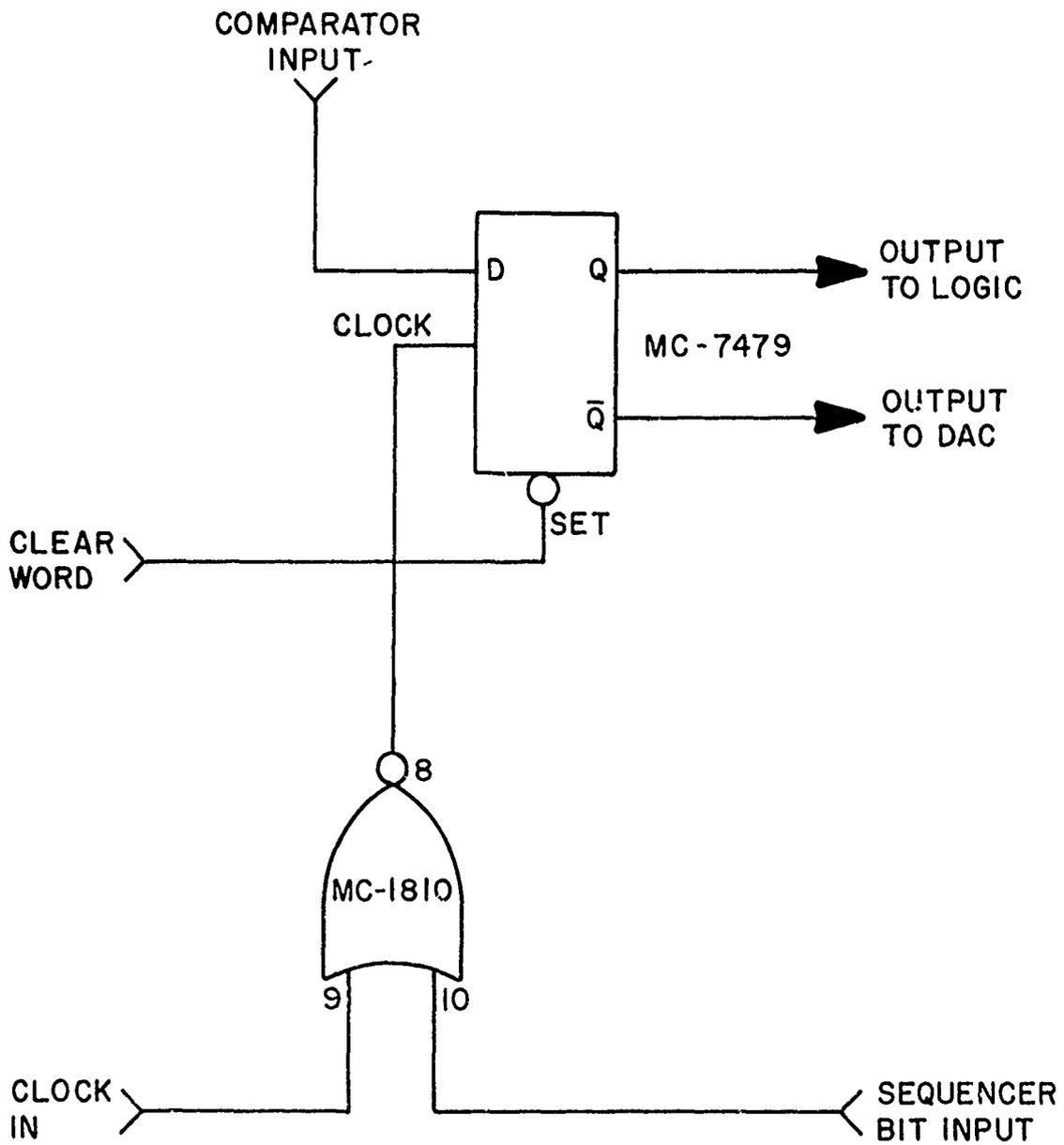


Figure B-11. Typical Storage Element

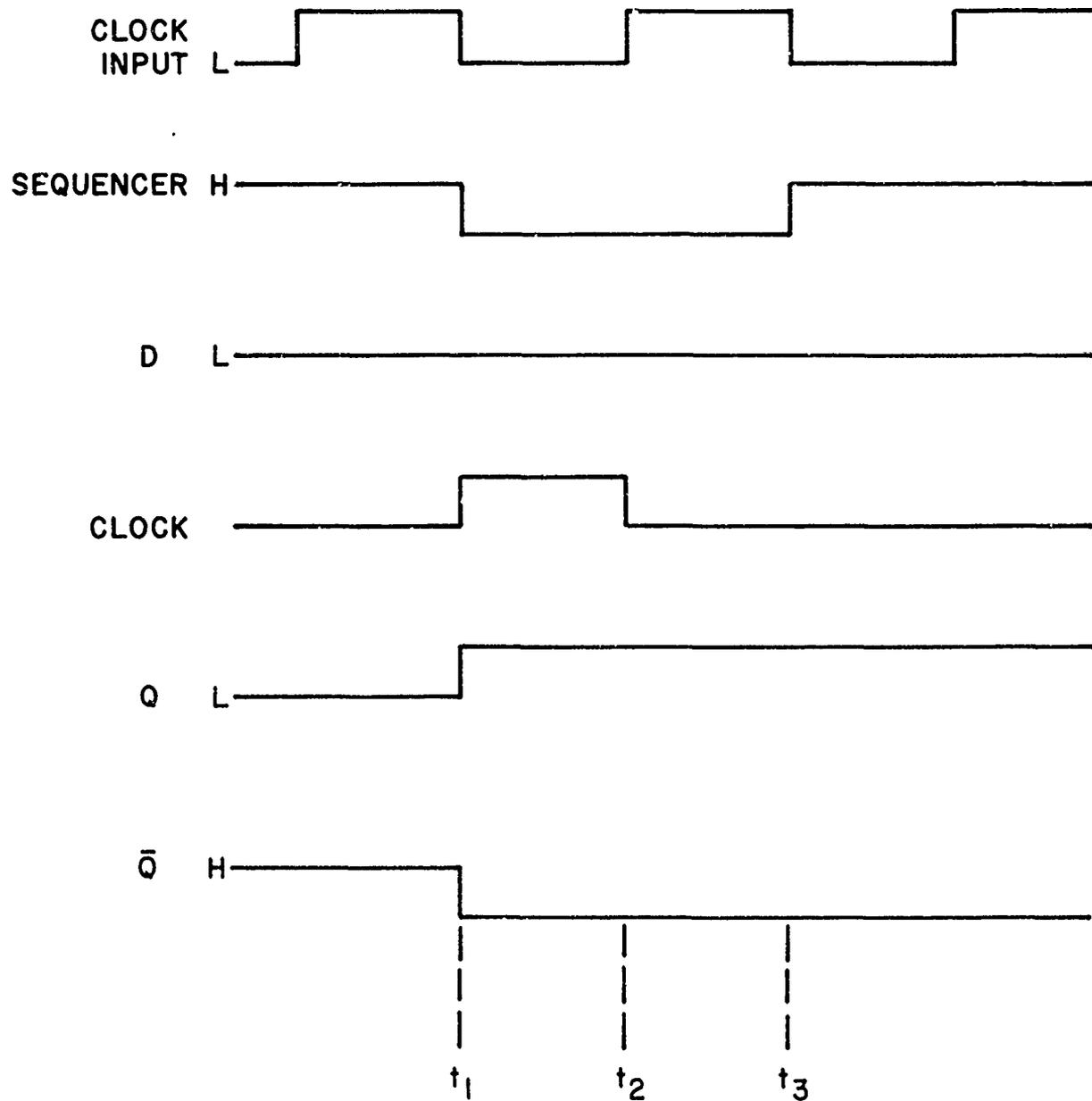


Figure B-12. Timing Diagram for Typical Storage Register

## B-6.9 COMPARATOR

The comparator consists of a simple operational amplifier. A type  $\mu A741$  was used in place of the conventional comparator because of interface problems. The inputs of the comparator were required to handle 10 volts at times. To protect the logic, the output of the comparator was zenered at 5 volts; this also protects the logic from negative inputs.

## B-6.10 OVERALL PERFORMANCE

The overall design goal of the VSWR detector, qualitatively speaking, is the following:

1. A seven-bit binary output is related to the VSWR of the load as reflected through the coupler RF tuning network.
2. A VSWR of 1:1 (a reflection coefficient of zero) corresponds to all bits being "low."
3. A VSWR of infinity (a reflection coefficient of unity) corresponds to all bits being "high."
4. For a constant VSWR, the binary output should not change over the operating range of forward power.

As a means of evaluating the results, a value is given to each output line. These are as follows:

LINE	VALUE
1 (MSB)	64
2	32
3	16
4	8
5	4
6	2
7 (LSB)	1

The term "binary weighting value" is found as follows. The output lines which are "high" are recorded. Their respective value is found from the above chart. These values are summed. Their sum is called the binary weighting value. A binary weighting value (BWV) of 127 corresponds to a reflection coefficient of 1.0. A BWV of zero corresponds to a reflection coefficient of 0.0. There is a gain adjustment available to assure that both amplifier outputs are equal when a short circuit is placed on the output of the VSWR detector. This assures a BWV of 127 for a short circuit and is the only adjustment necessary.

Figures B-13 and B-14 are performance curves at the two extremes of forward power levels.

## B-7 INTERFACE LOGIC

The digital system necessary for the antenna coupler is divided into three subsystems. These are:

1. Switch control logic
2. Detector interface logic
3. Main tuning logic

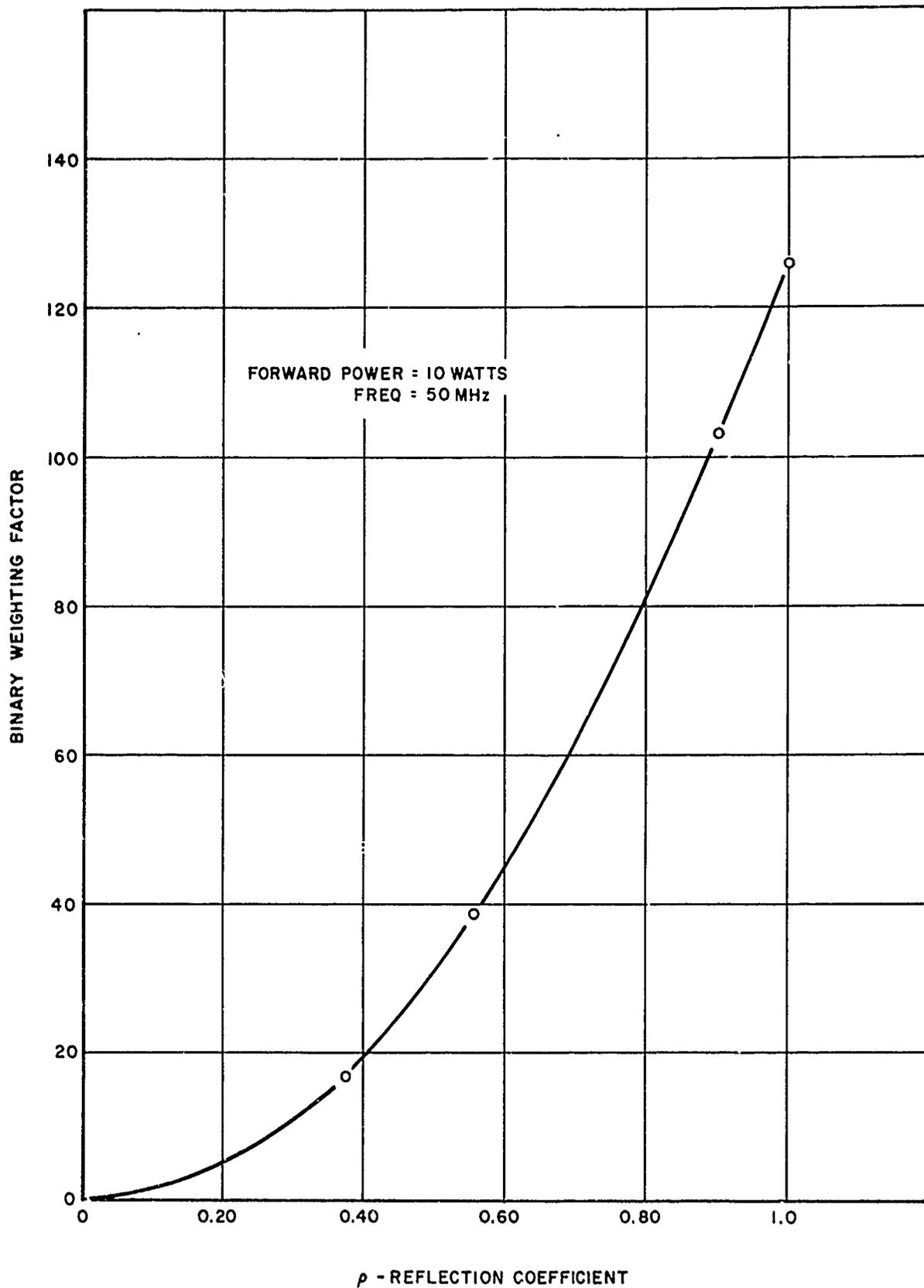


Figure B-13. Binary Output Versus Reflection Coefficient for VSWR Detector  
(Forward Power of 10 Watts)

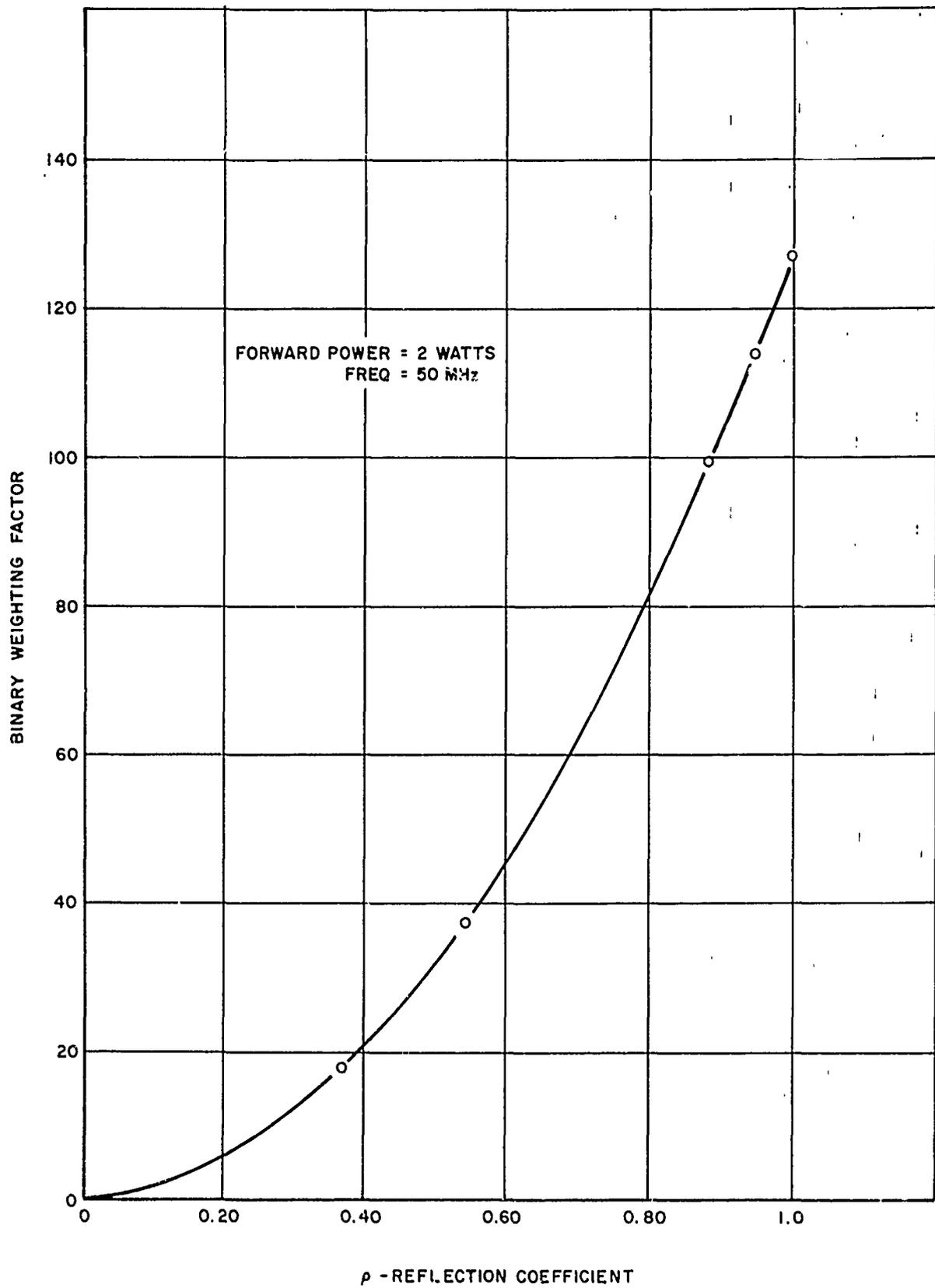


Figure B-14. Binary Output Versus Reflection Coefficient for VSWR Detector (Forward Power of 2 Watts)

Operation of the antenna coupler logic has been described previously in Technical Report ECOM-0101-2. The basic tuning sequence was outlined in that report and a complete flow diagram shown. Minor modifications made to the logic scheme since then include the addition of circuitry to pre-program the output fixed coil band so that those coils which would never be used in a particular band are automatically locked out. The tuning sequence has been completely computer simulated using the program described in Technical Report ECOM-0101-2. The results of this simulation served to verify the validity of the tuning scheme using the reflection coefficient as the only sense information.

## APPENDIX C

### 2 - 30 MHZ AUTOMATIC ANTENNA COUPLER

#### C-1 OVERALL REQUIREMENTS

Any consideration for an automatic antenna coupler in the HF range must take into account all of the parameters already discussed in section 2.1. Of special importance in the HF region is the antenna impedances, since normal whips used with radio sets in this region have very high Q's with a large capacitance reactance that must be resonated. The complete requirements discussed for this design include:

- Frequency 2 - 30 MHz
- Matching Capability 9-foot whip, 2 - 4 MHz  
6-foot whip, 4 - 30 MHz  
10:1 VSWR circle, referenced to 50 ohms
- Matching Accuracy 1.3:1
- Power Handling 40 watts
- Insertion Loss 1 db maximum for VSWR's less than 10:1
- Tune Time 1 second maximum
- Tune Power No power in transmit and receive, pulsed during tune cycle
- Harmonics Self-generated harmonics down 80 db
- Memory Yes

#### C-2 DESIGN CONSIDERATIONS

Consideration must be given to why an automatic antenna coupler is necessary. Some of the major benefits of this approach are:

- Maximum power transfer to the antenna under all operating conditions
- Consistently high amplifier efficiency
- Predictable filter characteristics

Since a good VSWR (1.3:1) is always maintained, a reliable design for harmonic attenuation is assured.

- Reliable power amplifier performance

The possibility of excessive voltages and currents is minimized due to the matched conditions; therefore, protective circuitry is made less complex.

- Automatic compensation for impedance variations of manpack antenna in field
- Automatic maintenance of tuned condition in receive mode

The primary function of the antenna coupler is to efficiently match all of the required antennas to enable maximum transfer of power from the power amplifier to the antenna. Any proposed matching network configuration is determined largely by the antenna impedances which must be accommodated over the operating frequency range. The greatest impedance matching problem occurs at the low end of the 2 — 30 MHz range because of the tremendously high Q's exhibited by the antennas in this region. At the high end of the band, the usual problem of not being able to achieve a low enough minimum capacitance has been alleviated by using small fixed capacitors instead of one large variable unit. This is also one of the major disadvantages in using a binary-type shunt capacitor bank, since the relay contact capacitances in parallel would be appreciable at 30 MHz. Another point to consider is that the antenna impedances shown in table C-1 are merely nominal values and are subject to variation. The range of values at a specific frequency is, primarily, the result of variations which occur as the operator moves about. Variations can also result from bodily differences from operator to operator and from the different conductivity of the ground over which the operator is transmitting. This 2 — 30 MHz coupler is designed to accommodate the entire range of antenna resistance and reactance variations.

Therefore, the major item that must be resolved before going any further is the choice of network configuration to be used. An assumption is made in this design discussion that the following basic antenna coupler components are available:

- Tuning Components
  1. Saturable core reactor
  2. Piezoelectric capacitor
  3. Switched fixed elements, either inductors or capacitors
- Sensors
  1. Resistance detector
  2. Phase detector
  3. Conductance detector
  4. VSWR detector
  5.  $|Z|$  detector

It is also assumed that the reader has some familiarity with the above-mentioned components and circuits, since a theoretical description of them does not form part of this report.

Table C-1. Impedances for 6-Foot and 9-Foot Whip Antennas

FREQUENCY (MHZ)	6-FOOT WHIP IMPEDANCE (OHMS)		9-FOOT WHIP IMPEDANCE (OHMS)	
	R	X	R	X
2	—	—	19	-2600
3	—	—	17	-1900
4	14	-1800	14	-1500
5	12	-1400		
6	11	-1200		
7	10	-1000		
8	10	- 870		
9	10	- 760		
10	10	- 680		
12	11	- 580		
14	12	- 450		
16	13	- 390		
18	15	- 340		
20	18	- 280		
22	20	- 240		
25	27	- 200		
30	38	- 120		

### C-3 NETWORK CONFIGURATION

#### C-3.1 NETWORK TYPES

It is known that any load can be tuned by only two (reactive) elements in the right arrangement and having the appropriate values. Figures C-1 through C-4 outline the different possible combinations. The shaded areas define the range of impedances that can be tuned if both RF elements are assumed to be variable. The arrows indicate a possible path followed by the image of the load impedance transformation if only one tuning element is changed at a time.

From these plots it can be seen that the output section of the matching network must have either L-match configurations as shown in either figures C-2 or C-3. Figure C-2, however, shows an output shunt inductor which is not a desirable situation because of stray capacitances to ground. Therefore, the output section of the 2 - 30 MHz matching network must have a series inductive arm as shown in figure C-3. This is, of course, necessary because in the low frequencies the whip antenna impedances all fall in region 6 of the Smith Chart (figure C-5). However, as indicated in table C-1, over the complete 2 - 30 MHz frequency range the antenna impedances vary considerably and do not remain in just one region. Therefore, more than a two-element network is needed to provide the necessary matching capability.

In addition to an output inductive series arm consideration must be given to what other elements are necessary. Two network types come immediately to mind:

- T-network
- $\pi$ -L network

# IMMITTANCE CHART

IMPEDANCE COORDINATES — 50 OHM CHARACTERISTIC IMPEDANCE  
ADMITTANCE COORDINATES — 20 MILLIMHO CHARACTERISTIC ADMITTANCE

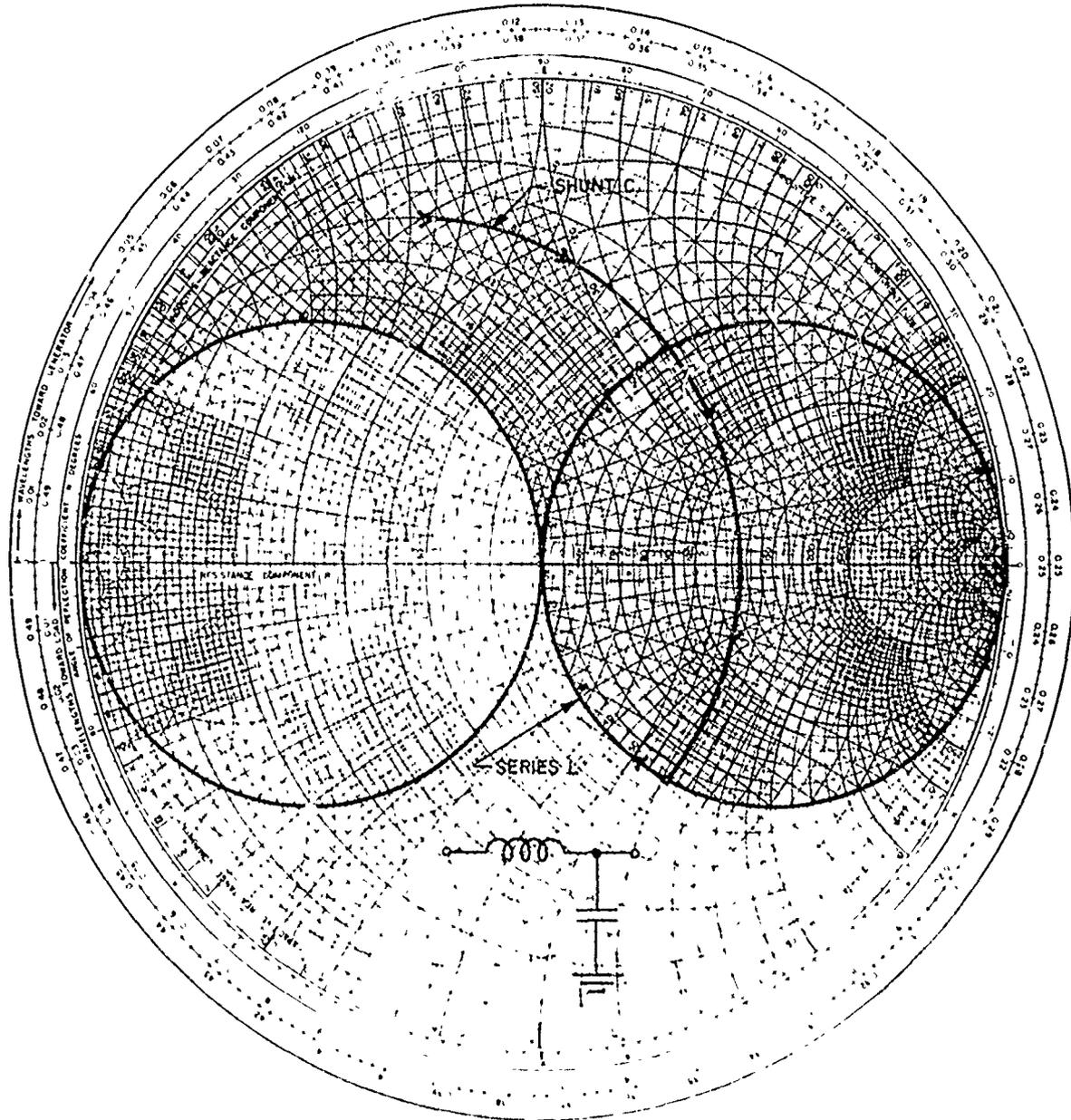


Figure C-1 Tuning Diagram for Series Inductor and Shunt Capacitor



### IMMITTANCE CHART

IMPEDANCE COORDINATES 50 OHM CHARACTERISTIC IMPEDANCE  
ADMITTANCE COORDINATES 20 MILLIMHO CHARACTERISTIC ADMITTANCE

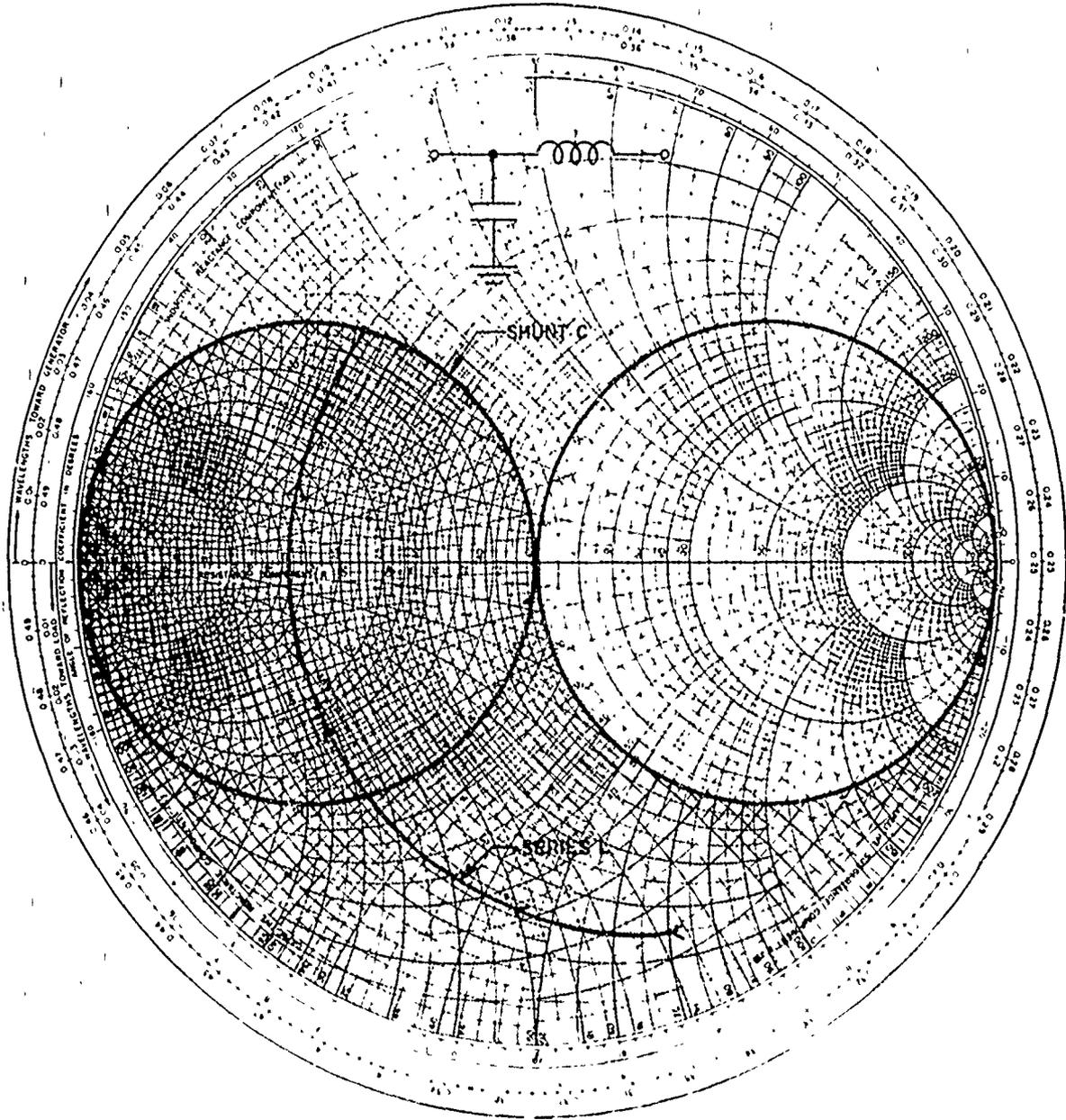


Figure C-3. Tuning Diagram for Shunt Capacitor and Series Inductor

# IMMITTANCE CHART

IMPEDANCE COORDINATES — 50 OHM CHARACTERISTIC IMPEDANCE  
ADMITTANCE COORDINATES — 20 MILLIMHO CHARACTERISTIC ADMITTANCE

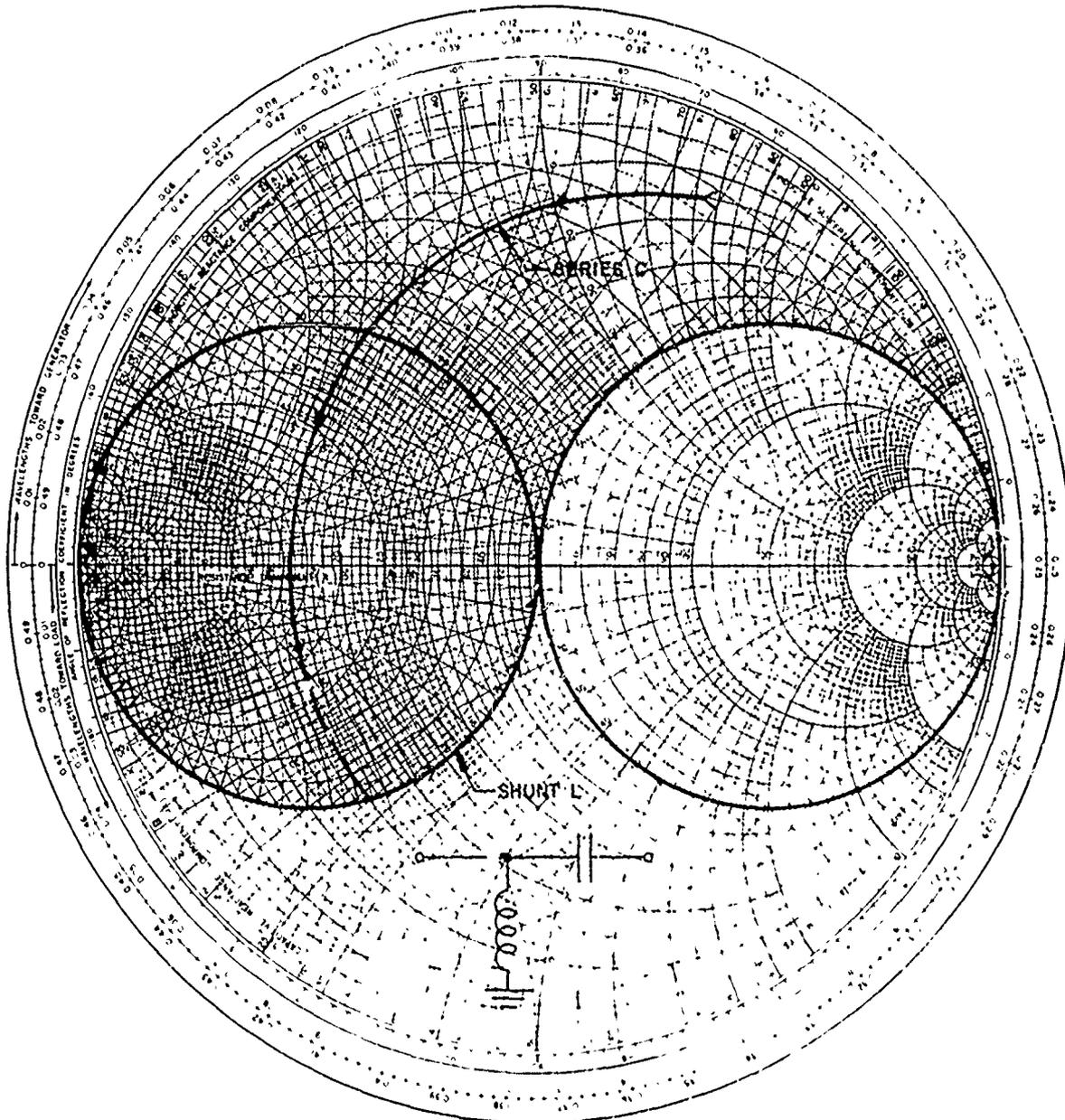


Figure C-4. Tuning Diagram for Shunt Inductor and Series Capacitor

# IMMITTANCE CHART

IMPEDANCE COORDINATES — 50 OHM CHARACTERISTIC IMPEDANCE  
ADMITTANCE COORDINATES — 20 MILLIMHO CHARACTERISTIC ADMITTANCE

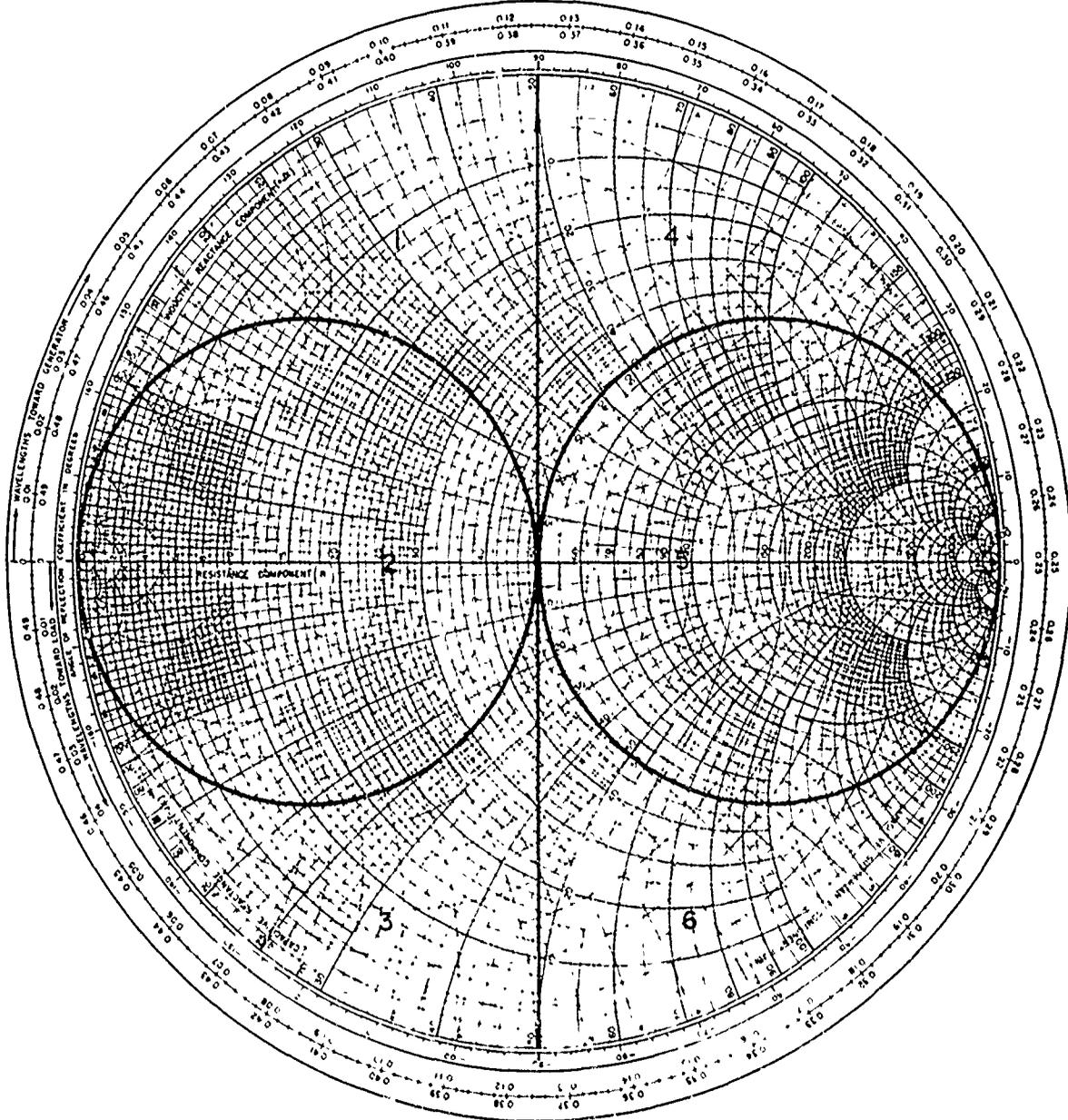


Figure C-5. Impedance Regions

Of these two, the T-network has the advantage of requiring one less branch than the  $\pi$ -L type. However, this brings up the question of the type of variable tuning element to be used. To provide the matching range of interest, two variable elements must be used. The following applies:

- If both variable elements are inductive, the T-network should be used.
- If both variable elements are capacitive, the  $\pi$ -L network should be used.
- If a variable capacitor and inductor are used, much more flexibility is obtained. A T-network should, however, still be given primary consideration.

Also to be considered is whether the variable tuning element is a continuous-tuning type or a switched-bank of fixed components. This will be considered in a later section.

Initial consideration will be given to a T-network, with inductive horizontal series arms and a capacitive vertical shunt leg. This network is a combination of figures C-1 and C-3.

### C-3.2 TUNING METHODS

Tuning methods include the following:

1. Tune L1 for R and L2 for phase and bandswitch C1
2. Tune C1 for R and L2 for phase and bandswitch L1
3. Tune both L1 and C1 for R and L2 for phase.

These different tuning methods will be considered in this section. Reference is made to figure C-6, which shows the basic T-network. The input and output arm inductors have a resistance associated with them which is indicated by  $R = X_L/Q$ . It is assumed that sensing information is available in all cases. This information can take the form of:

1. Resistance Detector:  $R \{Z_L\}$
2. Phase Detector:  $\phi \{Z_L\}$
3. Conductance Detector:  $R \{Y_L\}$
4. Impedance Detector:  $|Z_L|$
5. VSWR Detector:  $\rho = \frac{P_{\text{reflected}}}{P_{\text{incident}}}$

The type of sensor ultimately required depends on the tuning method adopted. All of the above detector systems have been built and used in practical radios, therefore, the type of network or tuning method is not restricted by the sensor system.

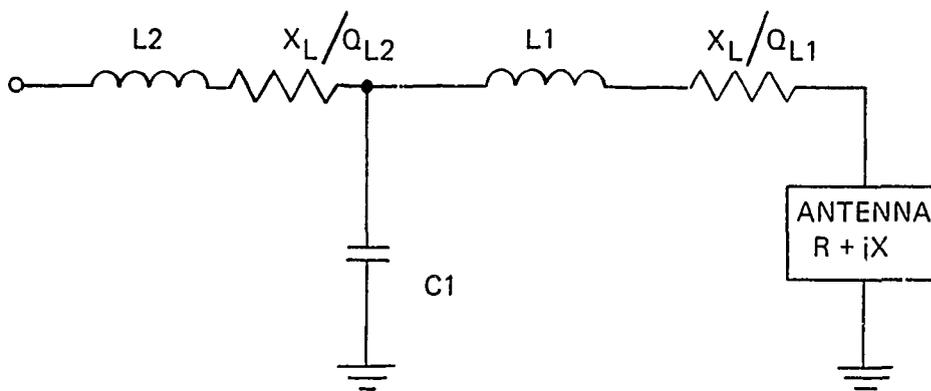


Figure C-6. Basic T-Network

### C-3.2.1 Case No. 1 (Tune L1 for R and L2 for Phase)

If L1 and L2 are tuned, L1 has to be tuned for  $R = 50$  ohms. If L2 is momentarily short-circuited, then the L1-C1 combination tunes as shown in figure C-7. Here tuning of a load  $25 + jX$  is assumed. Such a load will be located on circle I, and will move around the circle in the direction of the arrow if L1 is made smaller. Because of the shunting capacitor C1, the coupler output impedance lies somewhere on circle II (C1 here assumed to be 20 millimhos). The intersection of circle II with the 50-ohm circle (A) is the tuning point.

As can be seen from this diagram, with decreasing L1 value the real part of the coupler output becomes successively smaller than 50 ohms, equal to 50 ohms (tune point), larger than 50 ohms, again equal to 50 ohms, and finally again smaller than 50 ohms. There is no essential difference between points  $<50$  ohms for large L or  $<50$  ohms for small L. This means that tuning is only possible if a point between A and B can be found. Since, in this case A equals 12 millimhos and B equals 20 millimhos, the distance of A and B equals  $83 - 50$ , or 33 ohms. This means that the  $\Delta L$  steps to be made by L1 must be such that  $\Delta\omega L < 33$  ohms. This is only true for a 25-ohm load and a 20-millimho parallel capacitor; for other values of load on C1,  $\Delta L$  might be smaller or larger.

In a computer program in which such a tuning system was simulated, it turned out that a system using steps of 2 mah was only useful up to about 6 or 7 MHz with the 9-foot whip antenna. This means that, if binary tuning is used, a "smallest possible step scanning" has to be used. If binary tuning is impossible (because too many binary steps are needed), it may be possible to use a combination of binary and saturable reactor tuning. However, the requirement of having to be able to make very small steps requires a cooperation between switch and saturable reactor logic. This amounts to a trade-off between the following possibilities:

1. Make a binary step on: apply all available pulses to the saturable reactor. When there are 126 binary steps (as required) +40 pulses, the total tuning and retuning time becomes:

$$126 \times 5 + 126 \times 40 = 5670 \text{ milliseconds}$$

# IMMITTANCE CHART

IMPEDANCE COORDINATES - 50 OHM CHARACTERISTIC IMPEDANCE

ADMITTANCE COORDINATES - 20 MILLIMHO CHARACTERISTIC ADMITTANCE

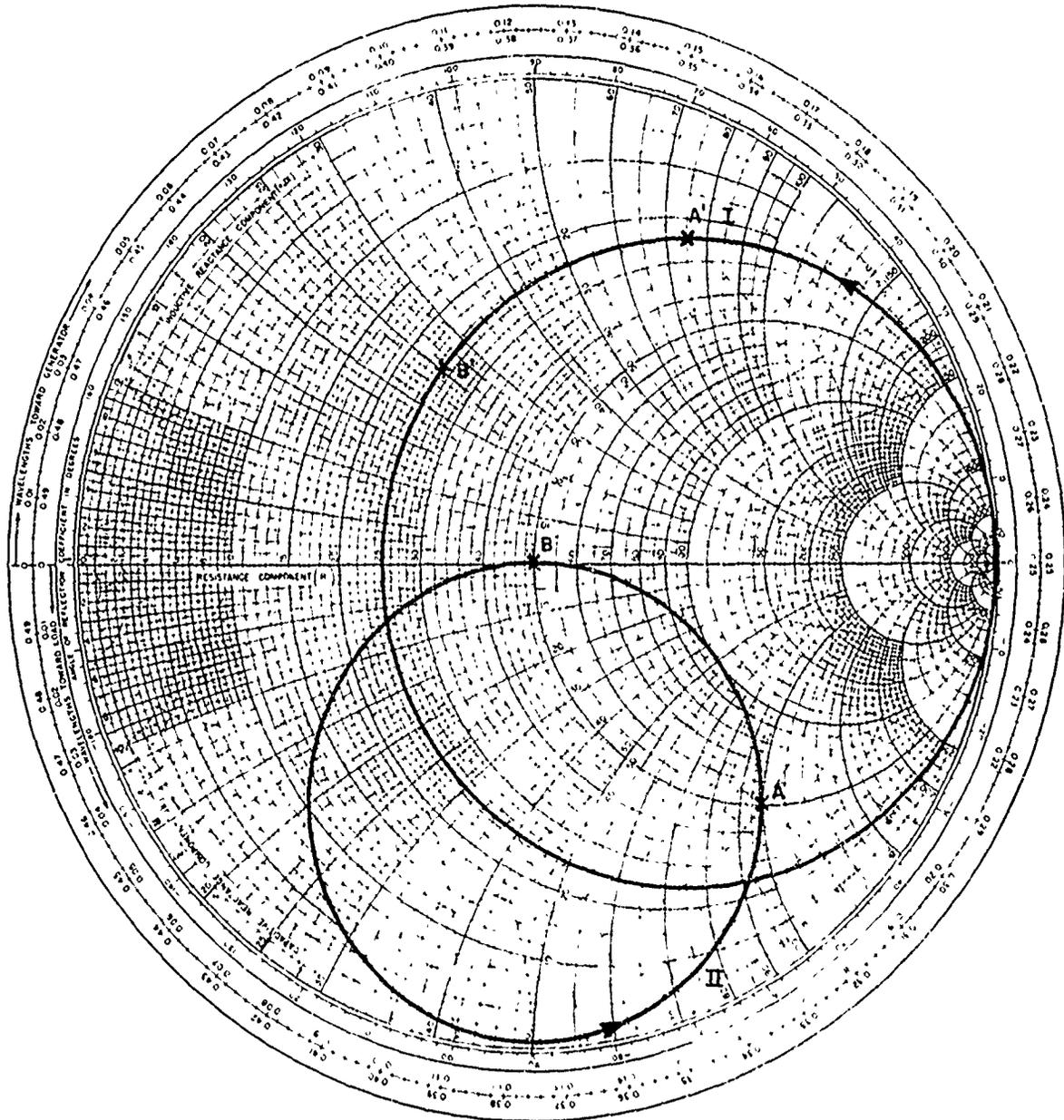


Figure C 7. Tuning Diagram for Case No. 1

2. Make binary steps and apply only enough pulses to the saturable core as necessary to obtain the required resolution. Make the binary steps add or subtract as required (up -down counter). This requires extensive logic hardware. The first estimate would be somewhere around 150 to 200 IC's. (This also requires a very high supply current.)

The above stated dilemma might be avoided by using any one of the other tuning methods.

It might be suggested that the problem of finding the 50-ohm tuning point in this configuration can be avoided by a pretuning procedure. The idea is to tune the L1 - load combination for some given condition. Possible conditions to pretune are:

1.  $\phi = 0^\circ$
2.  $|Z| = 50$  ohms
3.  $G = 20$  millimhos

Conditions 2 and 3 only will work for antennas where  $R < 50$  ohms. Very many frequencies apply to antennas where  $R < 50$  ohms. Since, for those antennas the problem of finding proper R- tuning is just as severe as everywhere else, conditions 2 and 3 offer no real solution.

However, neither is it possible to tune, for every frequency, each antenna such that  $\phi = 0^\circ$ , because this is only possible if the reactive part of the antenna at that frequency is capacitive. If an antenna is given by  $R + jX$  with  $X > 0$  and  $R < 50$  ohms, there is no way of pretuning the L1-plus-load combination, because changing the value of L1 will never bring the L1-plus-load combination in another section of the immittance chart.

The word "section" in the above context means an open region in the immittance chart bounded by and not containing one of the lines:

$$\begin{aligned} \phi &= 0^\circ, G = 20 \text{ millimhos} \\ R &= 50 \text{ ohms, } Z = 60 \text{ ohms} \end{aligned}$$

### C-3.2.2 Case No. 2 (Tune C1 for R and L2 for Phase)

If only C1 and L2 are considered, the network falls into the category of figure C-2. Although characteristics of some antennas lie partly within the shaded tuning range, there is no band where the characteristics of all the antennas lie within this range. This means that a series L is needed to change the antenna impedance in such a way that tuning will be possible. This again means that L1 will be a bandswitched inductor. The values for L1 are established by considering all antennas at the low end of the band, where  $\omega L$  has its smallest value, and computing which value of L is needed in series with the antenna impedance to make it positive and larger than 50 ohms.

A computer program was written for this circuit and the resulting data shown in table C-2 was obtained. The columns in the table represent the following:

BAND	—	Band of interest
MHZ	--	Frequency of operation
UH	—	Value of L2 in uh
PF	—	Value of C1 in picofarads
POUT	—	Output power to antenna in watts

PLOSS — Loss of power in L1 in watts  
 PCIR — Circulating power in L2 in watts  
           = | voltage across L2| x |current through L2|  
 VSWR MAX — The value of C1 is changed by 0.1 pf. The  
               figure in this column represents the VSWR  
               after such a change.

Table C-2. Results of Computer Program Analysis of 9-Foot and 6-Foot Whip Antennas

9-FOOT WHIP							
BAND	MHZ	UH	PF	POUT	PLOSS	PCIR	VSWR MAX
1	2	23.94	683.5	16	33.8	300	1.00
1	3	122.46	52.4	12	37.6	2440	1.29
2	3	11.60	527.1	27	22.4	218	1.00
2	4	62.46	56.7	21	28.4	1569	1.14
6-FOOT WHIP							
3	4	4.42	724.2	26	23.6	111	1.00
3	5	40.87	58.0	21	28.3	1283	1.12
3	6	55.86	28.8	18	31.5	2105	1.45
4	6	4.92	349.3	27	22.2	185	1.00
4	7	19.24	68.8	24	25.3	846	1.00
4	8	26.82	37.0	23	26.9	1348	1.24
4	9	31.64	24.3	21	28.4	1789	1.53
4	10	34.42	17.8	20	29.7	2162	1.96
5	10	5.60	123.5	31	18.4	351	1.02
5	12	8.27	56.3	30	19.5	623	1.07
5	14	13.96	23.7	29	20.3	1228	1.38
6	14	1.51	215.3	38	11.3	133	1.00
6	16	3.61	73.2	38	11.8	363	1.03
6	18	4.77	42.2	38	11.5	540	1.08
6	20	5.55	28.1	39	10.9	697	1.15
7	20	1.16	127.2	43	6.1	146	1.00
7	22	1.85	67.9	43	6.0	256	1.02
7	25	2.21	41.1	44	5.2	347	1.04
7	30	2.58	22.5	45	4.5	486	1.09

**Circulating Power.** The circulating power is important because it establishes the size of the ferrite used as the saturable core. It was found that, for a 20 by 20 mm cube of 4E2 material, the maximum allowable circulating power at 2 MHz was 330 watts. However, this was established for a -30 db IM figure and on the condition that only third-order products were present. The feeling is that a safe limit has to be 5 times lower, or 60 watts at 2 MHz. For higher frequencies the power capability is higher in direct proportion, so that 30 watts/MHz is allowable.

From table C-2, it can be found that the largest circulating power is about 800 watts/MHz (at 3 MHz for 9-foot whip). This means that the 4E2 reactor should only contribute 30/800 of the total L2 inductance or approximately 5 uh.

If restricted to a 5 uh saturable core reactor, there should be no problems from overloading the reactor. There is a second reason to restrict this inductance. As can be seen from the computer output data, at high frequencies only a small value of L2 is needed. Because the expected range of 4E2 material under a varying polarizing field is only 1:4, taps will have to be provided on the saturable core reactor to obtain the smaller values of inductance. The self-resonant frequency of the coil must be kept above the frequency range where it will be used (2 - 30 MHz). This restricts the maximum inductance to a value that is expected to be 4 to 5 uh.

**VSWR.** From the computer data a more severe fact is evident. A very small change in C1 (by 0.1 pf) is enough to cause large amounts of detuning. A typical example is the whip antenna for low R values or band 4 at 10 MHz. Here the VSWR becomes 5.49 when C1 is changed from 12.4 to 12.5 pf. To find the reason for this, note that L1 in band 4 has a value of 35 uh. For 10 MHz this is an impedance of 2.2 kilohms. The whip antenna has an impedance of 2 -j 190 ohms. So the total impedance of L1 plus whip antenna becomes:

$$2 -j190 + j2200 = \dots + j2010$$

This represents a Q of about 1000. Obviously L1 is much too large; however, it is necessary to tune the 6-foot whip antenna in the same band at 6 MHz. The whip antenna impedance at this frequency is 11 -j1200 ohms. If the result is checked at 6 MHz, band 4, for the 6-foot whip antenna, it is seen that, after a 0.1 pf change, the VSWR is still 1.00. So, it is advantageous to try to obtain a small resultant reactance for the combination of L1 and the load.

This tuning system requires a saturable reactor in series with C1 in order to obtain enough resolution. Based on the data gathered, it appears that the LC series circuit gives rise to high circulating power in the saturable core reactor. Furthermore, a change of 0.1 pf out of 12.4 pf represents less than 1 percent, and also detuning as a result of temperature change may be anticipated. For these reasons, the third tuning system is favored.

### C-3.2.3 Case No. 3 (Tune Both L1 and C1 for R and L2 for Phase)

The general idea for this is as follows:

Tune L1 so that the resultant impedance of L1 plus load is tunable, with the reactive part of this resultant impedance as small as possible. This can be done by adding inductance to all loads except those fulfilling these conditions:  $R > 50$

or

$$|Z| > 50 \text{ and } \phi = +$$

If it is decided that a load falls outside this area, inductance should be added until

$$|Z| > 50 \text{ and } \phi = +$$

It is important to be aware of the fact that this tuning can be done by the "50 percent adding" technique. This means that 50 percent of the maximum available inductance is added. A "too small" or "too large" decision is then made, and this 50 percent inductance is kept in or removed from the circuit while another 50 percent of the then available inductance is added, and so on. This means that only  $n$  steps are needed for  $n^2 - 1$  possible values.

After this, C1 is tuned. Because the L1-plus-load combination by now has a much lower Q, C1 does not need to have too large a resolution. At this instant of study C1 is expected to need a 1 pf resolution. Then L2 is tuned binarily as mentioned above.

It has to be investigated if only one saturable core reactor as part of L2 is sufficient. Otherwise a second reactor must be installed in the L1 arm, thus avoiding a CL series circuit. It is anticipated that 7 switches for L1, 14 switches for C1, and 4 switches for L2 plus saturable core reactor would be required.

The pulse logic is not much different from the one implemented for exclusive saturable reactor tuning. The switching logic, however, is much larger because of the large amount of switches. If very fast retuning after a small change of load or frequency is desirable, the counters for the switches should be of the up/down type, which requires about two times as many logic elements.

### C-3.3 CONCLUSIONS

When restricted to an L-C-L T-network, the 3-element tuning described under Case No. 3 seems to be the best. However, there is still a number of problems associated with this approach. These include:

1. Size and amount of logic circuitry required
2. Required resolution for C1
3. Problems associated with small values of C1 because of the relay stray capacitance.
4. Influence of the Q of L2 on "R-shift". (If L2 is switched into the circuit and tuned for phase =  $0^\circ$ , then a resistance  $\omega L_2/Q$  is added. This changes the R = 50 ohms tuning, so that retuning of R must take place. This can be solved by employing an "iterative" type tuning approach; however, this is difficult to implement with a binary-switched system

#### C-3.4 MODIFICATIONS TO CASE NO. 3

An optimum solution, it would seem, would be to implement Case No. 3 as described above with a piezoelectric capacitor as the shunt element. In the 2 – 30 MHz frequency range, however, due to the large values of shunt C required at the low end of the range, the piezoelectric capacitor would have to be paralleled with switched fixed capacitors. With a present range of approximately 20 – 200 pf in the piezoelectric capacitor, two or possibly three fixed capacitors would have to be used in an arrangement as shown in figure C-8. In this configuration a continuously variable capacitance of 20 – 560 pf can be obtained (the minimum value is, of course, greater because of the stray capacity to ground).

Another modification to the basic tuning method (Case No. 3) is to use a saturable core reactor as the input L2 exclusively. Three cores would have to be used, since the required input L range is approximately 0.3 – 25  $\mu$ h. These three cores, or inductors, can be all put in the same reactor assembly as shown in figure C-9.

The output L1 can still be a binary-switched arrangement providing the tuning method assures that a spot within the region corresponding to  $|Z| > 50$  ohms and  $\phi = +$  can be accurately determined. This would be no problem as long as the output L1 is stepped in an inverse binary fashion, i.e., smallest step first and proceeding to the largest in steps equal to the smallest. The smallest step would have to be chosen with resolution sufficiently small not to "skip" over the above-mentioned region. Proceeding in a regular binary fashion would not necessarily assure this unless elaborate precautions are taken. This can take the form of additional sensors, for example.

The obvious disadvantage to this approach is, then, tuning time. Seven binary steps in the output L1 mean that:

$$2^7 = 128 \text{ steps could be used.}$$

Assuming 5 milliseconds per relay switch and sample period, this is:

$$5 \times 128 = 640 \text{ milliseconds.}$$

Capacitor tuning would be as follows, assuming the smallest capacitive increments must be used:

Clock speed for capacitor	=	20 KHz
Increments	=	128
Capacitive sweep	=	$50 \times 128 \times 10^{-6} = 64$ milliseconds

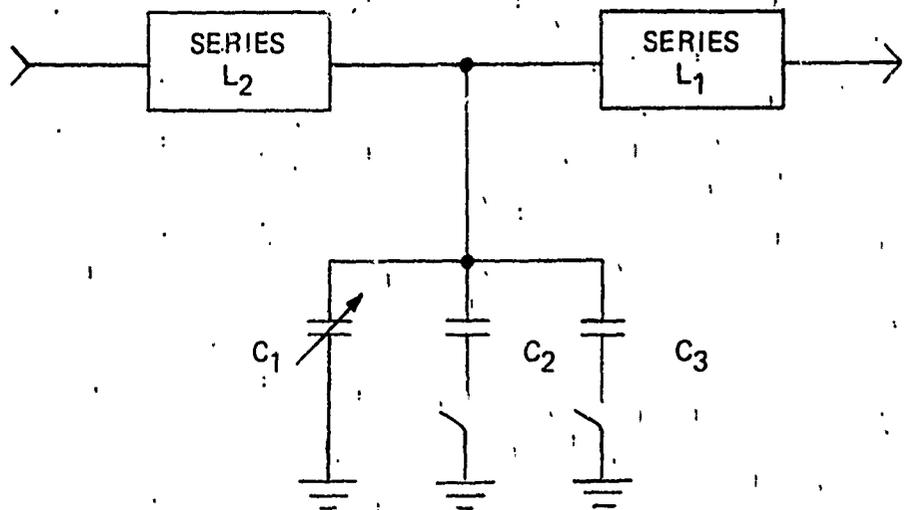
If two additional fixed capacitors are needed, then:

$$\text{Total C time} = 192 \text{ milliseconds}$$

Input reactor time, at a maximum, assuming 16 pulses at 10 milliseconds a pulse and 3 reactors, is:

$$3 \times 16 \times 10 = 480 \text{ milliseconds}$$

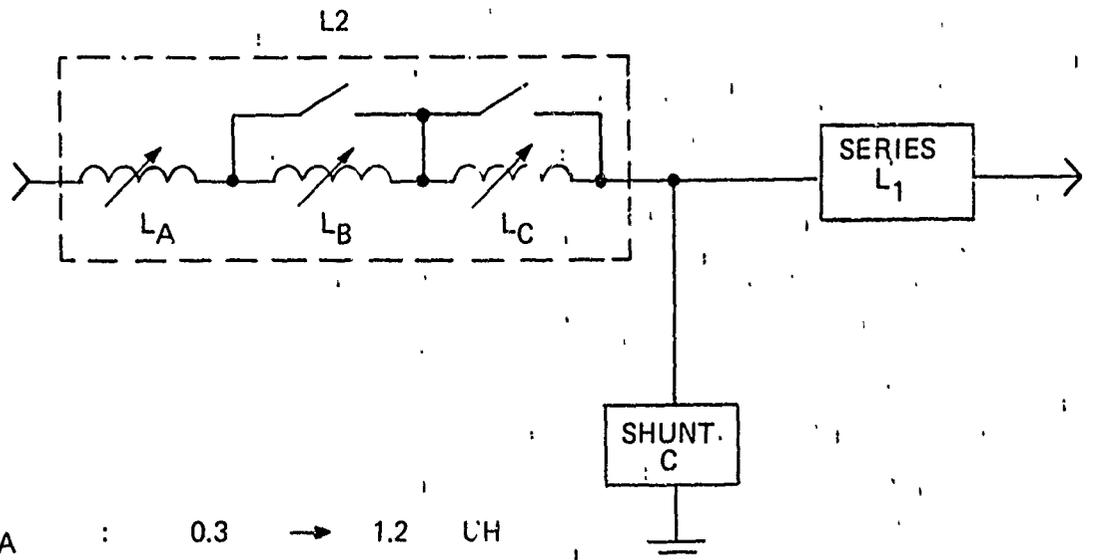
Therefore, a once-through time of  $640 + 192 + 480 = 1.312$  seconds could be achieved.



$C_1$  : 20 - 200 PF, PIEZOELECTRIC CAPACITOR

$C_2, C_3$  : 180 PF FIXED CAPACITORS

Figure C-8. T-Network with Shunt C Leg Using Piezoelectric Capacitor



$L_A$  : 0.3 → 1.2 UH

$L_B$  : 1.0 → 5.0 UH

$L_C$  : 4.5 → 22 UH

Figure C-9. T-Network with Saturable Reactor Input L Arm

This time, however, does not take into account any iterations which are necessary due to the stray reactances and finite Q's of the inductors.

To summarize, then, the obvious advantages to this type approach are:

- Flexibility — no bandswitching required
- Wide Tuning Range — variable L and C give large matching capability at any frequency

Disadvantages of this approach are:

- Two separate variable elements necessitate two separate control circuit types
- Complex logic
- Need for R,  $\phi$ , and  $|Z|$  detectors
- Possibly long tuning times

#### C-3.5 ALTERNATE T-NETWORK APPROACH

One other possible T-network method should be considered. This is depicted in figure C-10, which shows use of variable C1 and L1 components. L2 would then be bandswitched. This network has a considerable matching flexibility and could be implemented in the following way. The output arm, L1, would still have to be a combination of switched fixed coils and a saturable core reactor. The shunt leg, C1, would be a piezoelectric capacitor paralleled by switched fixed capacitors.

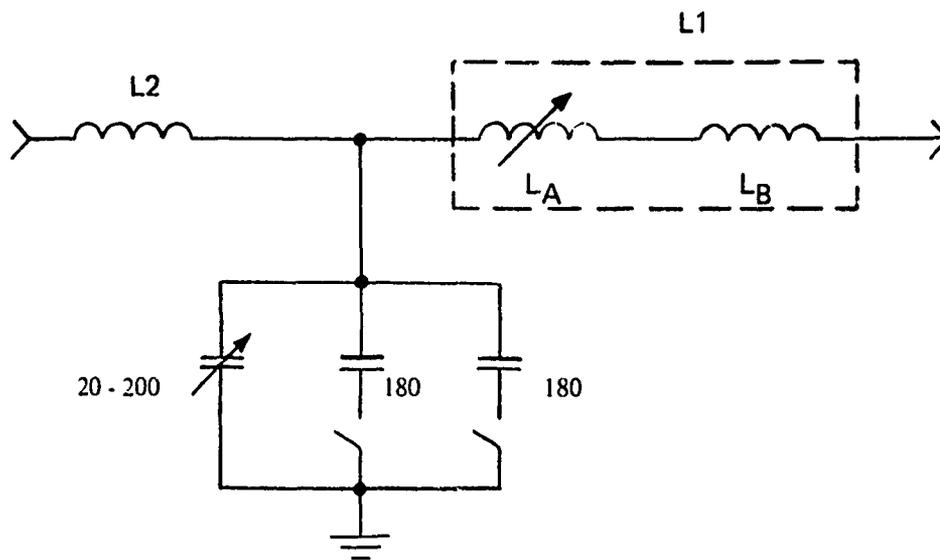
Since the input inductance is fixed for any given band, the output combination of C1 and L1 is tuned for a point which is equal to  $50 \text{ ohms} + X_{L2}$  as shown in figure C-11. The way this must be done is to vary L1 in increments and, between each incremental step, to vary C1 through its full range at the same time checking to see whether the tuning point is reached. In other words, L1 is first stepped, then C1 is swept, and VSWR is monitored. If no tuning point is reached, then L1 is stepped again, C1 is swept, and VSWR is monitored. This process is continued until tuning is achieved.

Advantages to this approach are:

- A one-step tuning procedure
- Only a VSWR detector needed (possibly only a reflected power detector)

Disadvantages are:

- Two separate logic and control systems required
- Tuning time could be long



$L_A$	:	0.3	→	6	UH	(SATURABLE CORE REACTORS)
$L_B$	:	5	→	150	UH	(BINARY RELATED SWITCHED FIXED INDUCTORS)

Figure C-10. Alternate T-Network

An example of the tuning times in the present system is:

Number of reactor pulses = 16  
 Number of reactor cores = 2  
 Number of fixed coils = 6

Therefore, there are 32 steps per coil and  $2^6$  or 64, possible coil switches. Then the number of inductor steps becomes:

$$32 \times 64 = 2048$$

If the total capacitor sweep time (including switching in the fixed C's) is 192 milliseconds as before, then the total tuning time (maximum) is:

$$192 \times 2048 \approx 400 \text{ seconds}$$

If the capacitor only had to be swept after a fixed coil change, then maximum tuning time would be approximately 12 seconds.

### IMMITTANCE CHART

IMPEDANCE COORDINATES — 50 OHM CHARACTERISTIC IMPEDANCE  
ADMITTANCE COORDINATES — 20 MILLIHM CHARACTERISTIC ADMITTANCE

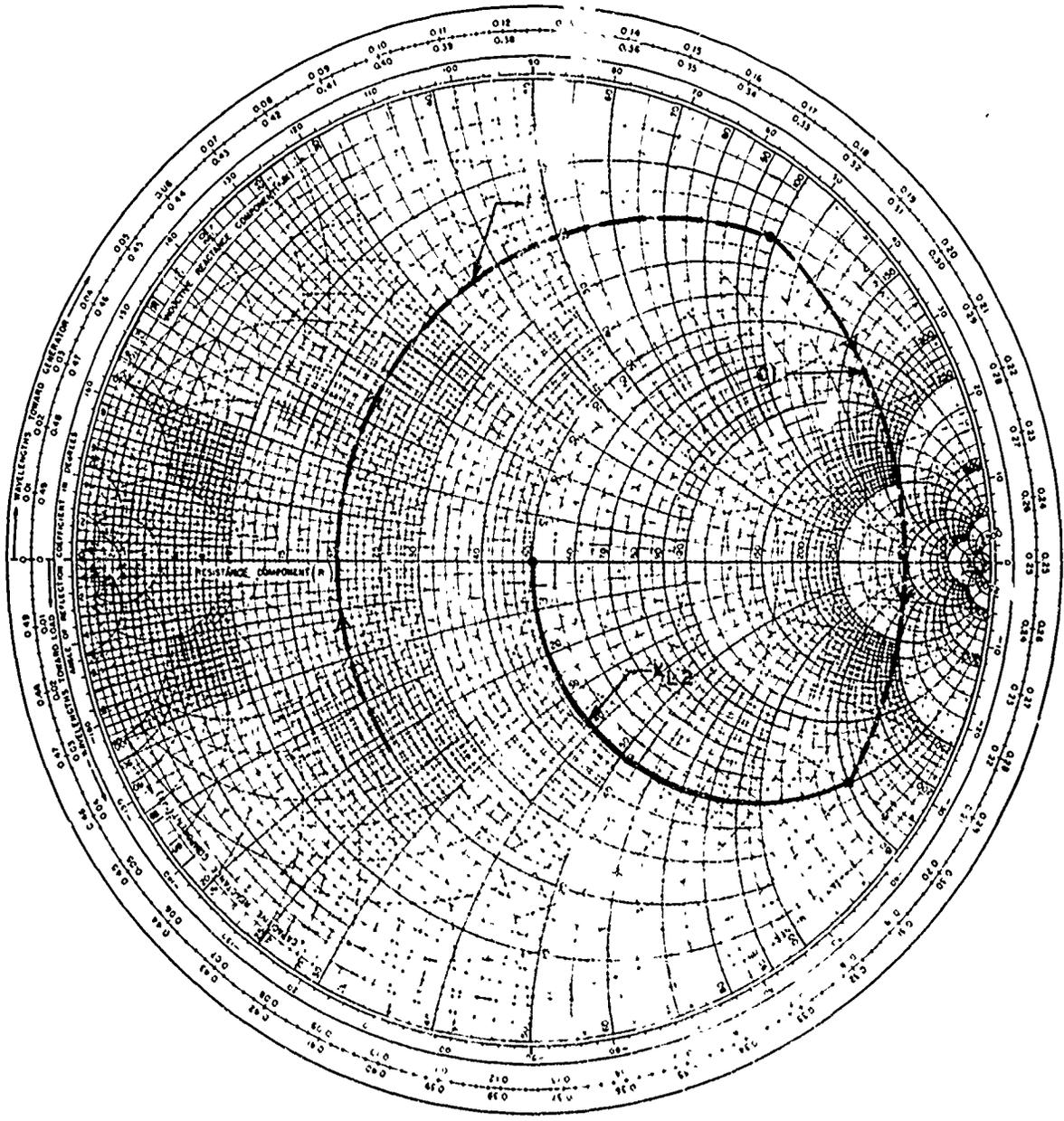


Figure C-11. Tuning Diagram for Alternate T-Network

Another possible means of implementing this approach is to sweep both the inductor and the capacitor simultaneously. It is theoretically possible to tune this way if the capacitor could be swept through its complete range about 40 times, while the inductor is swept through once. During this time, VSWR is continuously monitored and tuning achieved when a VSWR null is detected. Considerable implementation problems exist with this technique, but its tuning simplicity and accuracy still make it an attractive idea.

### C-3.6 $\pi$ -L NETWORK

A  $\pi$ -L network in the 2 – 30 MHz frequency range can be implemented in a method similar to the 30 – 80 MHz Coupler previously described. Output loading coils with values equal to those used in the T-network would be used, and the series mid-section inductor arm would still have to be bandswitched. The VSWR tuning method can be used, offering a large advantage in overall system simplicity.

The main disadvantage to using this type approach in the 2 – 30 MHz range is the large value of capacitance necessary in the shunt legs. As an example, the input capacitor in the network shown in figure C-12 should have a 20 – 500 millimho range. This means that the minimum C must be:

$$C_{\min} = \frac{20 \times 10^{-3}}{2\pi \times 30 \times 10^6} \approx 100 \text{ pf}$$

and the maximum C must be:

$$C_{\max} = \frac{500 \times 10^{-3}}{2\pi \times 2 \times 10^6} \approx 4000 \text{ pf}$$

Since the present piezoelectric capacitor maximum is only 200 pf, this means that at least five fixed switched capacitors must be added in parallel. This would be a difficult network to realize in practice, and additional study would have to be made before recommending its usage.

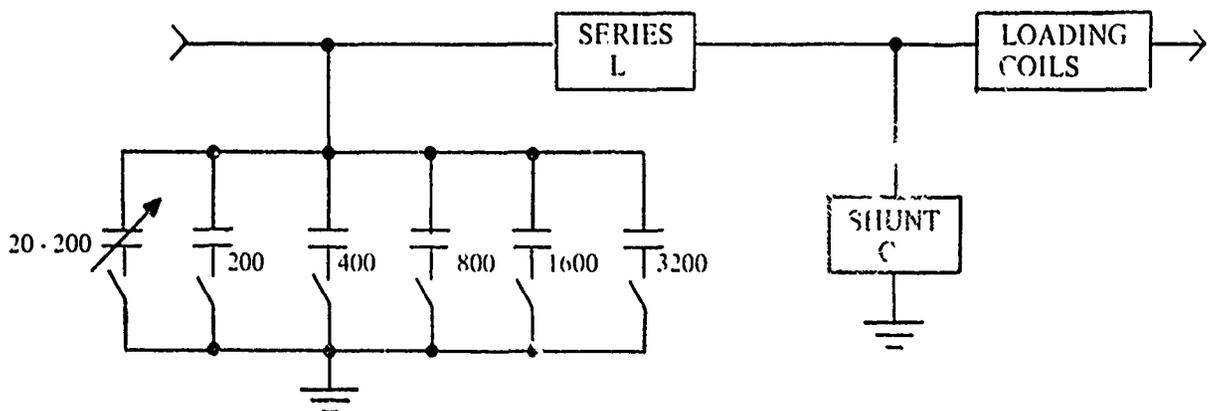


Figure C-12.  $\pi$ -L Network

## C-4 COUPLER DESIGN

### C-4.1 CONCEPTS

At present, the most practical approach to an automatic 2 — 30 MHz antenna coupler is to use a T-network with saturable core reactors in the input and output arms and a bandswitched shunt capacitor leg. Switched fixed loading coils will still have to be used in the output arm. The reasons for this approach are as follows:

1. A minimum number of network elements is preferable. (The T-network using a binary/static approach is a minimum element automatic coupler.)
2. Stray capacitance associated with the shunt leg can be considered part of the bandswitched capacitor and will have no adverse effects.
3. Impedances which cannot be matched by this network have a value larger than a certain  $|Z|$ , and this  $|Z|$  can be made as large as desired by proper selection of the shunt leg capacitor.
4. The wide matching requirements (whips plus 10:1 VSWR circle) mean that no pre-programming is possible. Under these circumstances, this type network gives, at present in a practical sense, the best tuning times with the least overall system complexity.

Other concepts important in an antenna coupler are:

- Memory
- Power Consumption
- Linearity

Tuning memory in the coupler is necessary for compatibility with SSB operation and manual or automatic CW operation when the carrier is interrupted. In the coupler, once the matching network is tuned, it will remain in this state until a new frequency of operation is selected or the system VSWR rises above a predetermined value due to an antenna impedance change. Since the latter changes are quite common, due to the different orientation and environmental effects described previously, the rapid retuning feature incorporated in the coupler is a distinct advantage. In fact, since retuning speed is so great (100 ms worst case), the coupler could be programmed to automatically retune in voice modes whenever the handset is keyed. This would insure the operator of a matched condition whenever he transmits, since any VSWR variation due to orientation or environment would automatically be compensated for. Coupler memory is accomplished by using latching relays for the coarse tuning elements and permanent magnet type saturable reactors for the fine tuning elements. A more complete description of this technique is given in later sections of this appendix.

Lower power consumption is readily achieved in the coupler and is a by-product of the techniques used for network tuning. The saturable reactor tuning elements will be pulsed stepwise to a higher or lower inductance using a permanent magnet to maintain the desired bias flux in the reactor. No holding power is required. Since the relays employed are of the latching type which are pulsed momentarily for operation, no holding power is required.

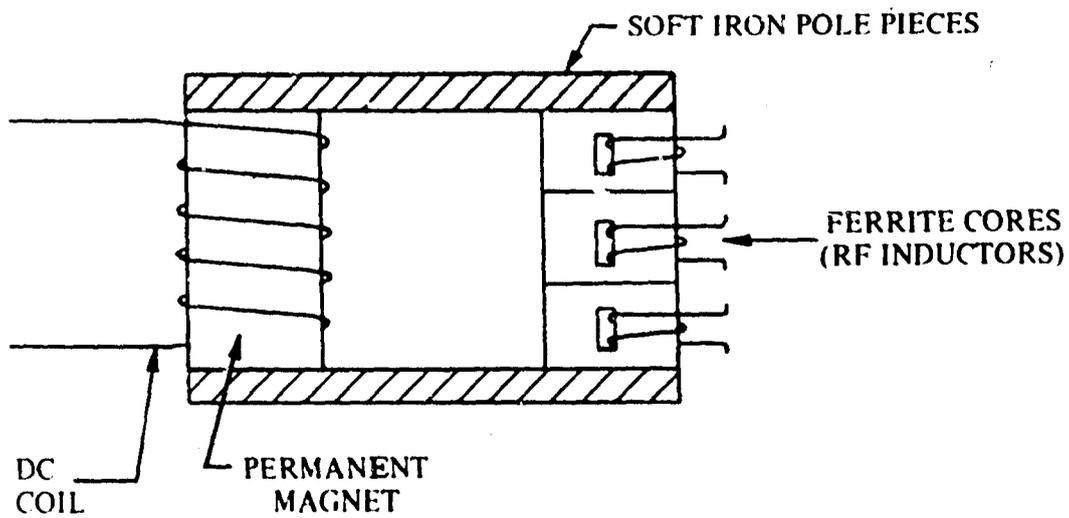
The coupler uses linear passive elements such as fixed inductors and capacitors in addition to the vernier saturable reactor tuning. As will be shown later, the linearity of the reactors is directly proportional to the ferrite volume for given power levels. The hybrid network approach offers the optimum combination of ferrite volume and number of fixed coils to achieve the required linearity. All the network operations were computer simulated in order to cover the diverse antenna conditions, and then the results were verified by tests on actual networks in the laboratory and in the field.

This network configuration has been utilized in previous Avco equipments, and it has been found to provide realizable performance. The system discussed here, however, features several design improvements. Among these are:

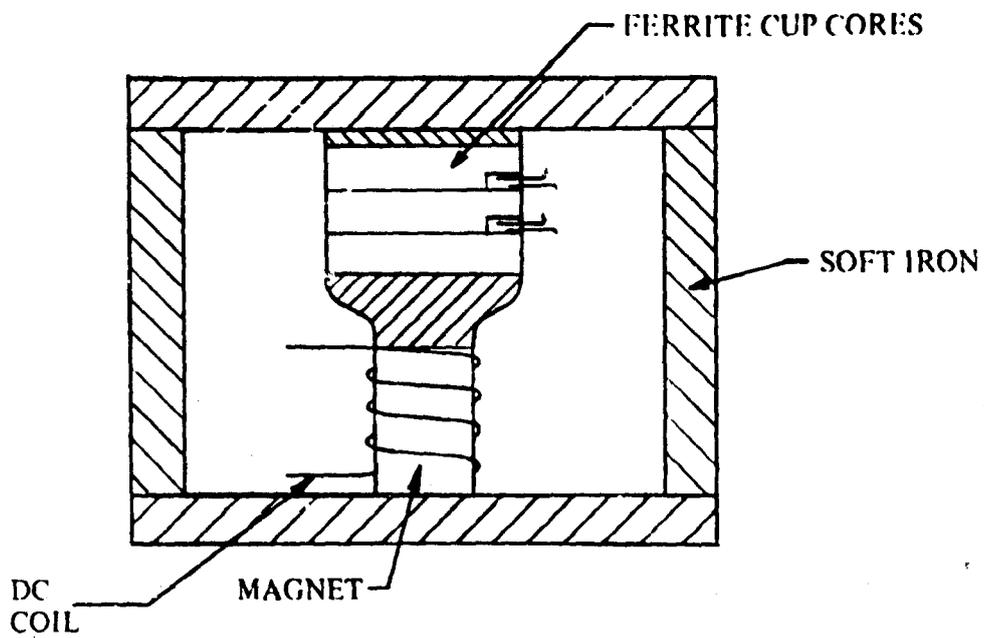
- New configuration saturable reactor yoke assembly. Cup-cores are used in an optimum magnetic flux circuit which provides greater inductance range and power-handling capability for a given size ferrite. The present technique also allows for the best heat removal from the ferrite. A comparison between the previous and present reactor systems is shown in figure C-13.

This figure shows the magnet in the same leg of the assembly as the ferrite, an optimum condition for saturation with a given magnetic field. Care must be taken so that the side pole pieces are not too close to the center. When the pole pieces are too close to the center, the flux may not follow the iron path, but instead will "jump" across the air gap and, therefore, not go through the ferrite gap.

- Current control of the reactor pulses. This technique eliminates the dependency of the energy pulses on supply voltage level. A simplified technique for this was described in Technical Report ECOM-0101-1. The circuit necessary here would feature temperature and overshoot compensation as well. Basic operation is the same, i.e., the incremental pulses given the reactor to change its value are set by a prescribed current level. The energy pulse into the DC coil then cuts off at this level. In this case, 16 pulses would probably be used to cover the full reactor inductance range.
- Use of the iterative "zone-controlled" tuning method prevents the logic from "getting lost" by automatically compensating for any stray reactances and finite element Q's. This technique is also described in Technical Report ECOM-0101-1.
- Use of a CMOS logic system. Tuning power, size, and weight are all considerably reduced by making the logic system out of CMOS logic. Ultimately, of course, such a design lends itself very well to a CMOS LSI system, which in this case would probably require two or three LSI chips.
- Hybrid packaging of reactor drive circuitry. The reactor drive circuitry lends itself to a transistor matrix which is shown in figure C-14. A Darlington-type arrangement is used. In the figure, terminal 1 is connected to the reactor DC coil, while terminal 5 is connected to the current-control sampling resistor. One of these matrices is used with each reactor, and the entire circuit can be put in a one-inch square hybrid package.



A. Previous Reactor Configuration



B. Present Reactor Configuration

Figure C-13. Reactor Configurations

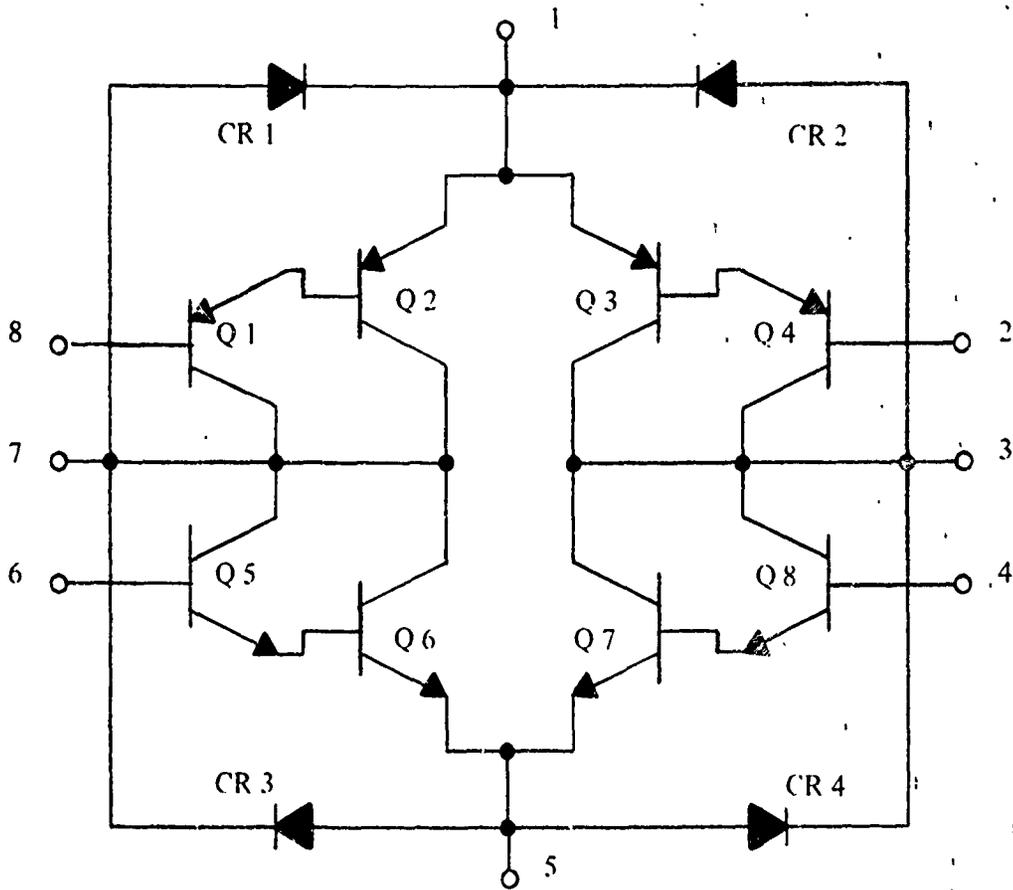


Figure C-14. Reactor Drive Circuitry

- Use of "fast scan mode" to improve tuning times. This means that, during the initial part of the tuning process when the first crossing of the 50-ohm circle is sought, the reactor will be swept and monitored by the R - detector to see if a crossing has occurred. The previous method was to pulse it incrementally whenever it was used. Since the tuning method requires going through the reactor each time a new fixed coil combination is inserted, this represents a considerable time savings. If there are 64 possible coil combinations and the reactor is pulsed incrementally and then sampled as before, the time for this could be:

Pulse time = 4 milliseconds  
 Sample time = 4 milliseconds  
 Number of pulses = 16

Then

$$16 \times (4+4) \times 64 = 8.2 \text{ seconds}$$

Using the new approach gives:

Swept pulse time = 12 milliseconds

Thus

$12 \times 64 = 768$  milliseconds

In this new approach, the R - detector is monitored during the sweep pulse and, if a crossing is detected, the reactor is swept through in a normal fashion. Some overshoot will occur in the inductance (in reactor L). However, this does not present a problem, since, if a crossing is detected in the overshoot portion — and therefore not detected during a normal incremental reactor sweep — the system will automatically switch to the next coil combination, thus automatically compensating for the apparent overshoot.

#### C-4.2 BLOCK DIAGRAM DISCUSSION

Figure C-15 shows a block diagram of the 2 — 30 MHz Antenna Coupler system. Reference to the block diagram of the antenna coupler shows that the major components of this system are:

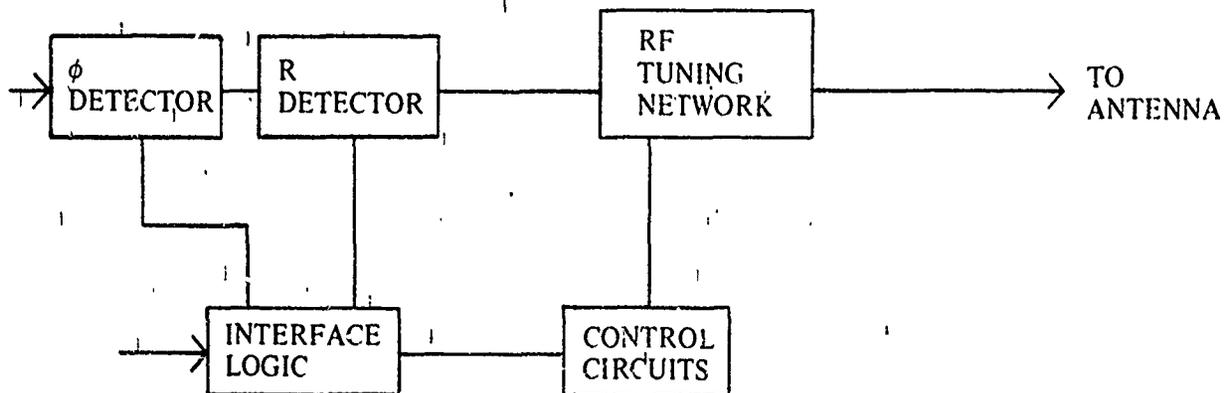


Figure C-15. 2 — 30 MHz Coupler Block Diagram

- **RF Tuning Network.** This network is a T-type network composed of a saturable reactor in each series arm and a capacitive shunt leg. Additional fixed coils are used in the output arm to obtain the necessary range at low frequencies. The mid-section capacitive leg has seven capacitors that are automatically preset according to band information, while the loading inductive arm's coils are logically switched prior to the actual tuning sequence.
- **Control Circuit.** The control circuit contains the control networks for the variable inductors, memory circuits, and control circuitry for the inductor relays.

- **Interface Logic.** The logic circuit utilizes the information from the detectors and system interface lines to control the operation of the antenna coupler.
- **Sensor System.** Sensing information about coupler input phase and resistance is provided for the coupler tuning sequence by the  $\phi$  and R detectors.

### C-4.3 RF TUNING NETWORK

The network for matching the required load impedances is shown in figure C-16. Note that this circuit is a single T network using variable inductors in the two series arms and bandswitched fixed capacitors in the shunt leg. A saturable reactor is combined with binary switched coils in the output arm, while the input arm has just a saturable reactor.

Three separate ferrite core inductors are used in one yoke assembly in both the input and output saturable reactor. These cores are electrically shielded from each other to minimize any unwanted coupling effects. The values of each inductor, along with the total range of variable L covered by each reactor assembly, are shown in the schematic diagram (figure C-16). Values of the fixed coils are also given. It is noted that, while these coils are switched in a binary fashion, their initial values are not exactly binary related. This is because the apparent inductance will vary with frequency due to both winding stray capacity and relay capacity effects which appear in parallel across each coil. Therefore, the initial design values must contain some overlap to compensate for this. Care must be exercised in the selection of these coils to insure that each coil is used well below its self-resonant frequency point. The best switching relays available at this time are small vacuum latching relays made by Jennings which feature, besides high voltage capability and memory, a very low contact capacity. Tests on several of these relays show a contact capacity of only 1.5 pf as compared with the standard half crystal can type which measured 4 pf. These are specially made RF relays with a 3 kv breakdown and 200 watts power-handling capability, well above the levels required.

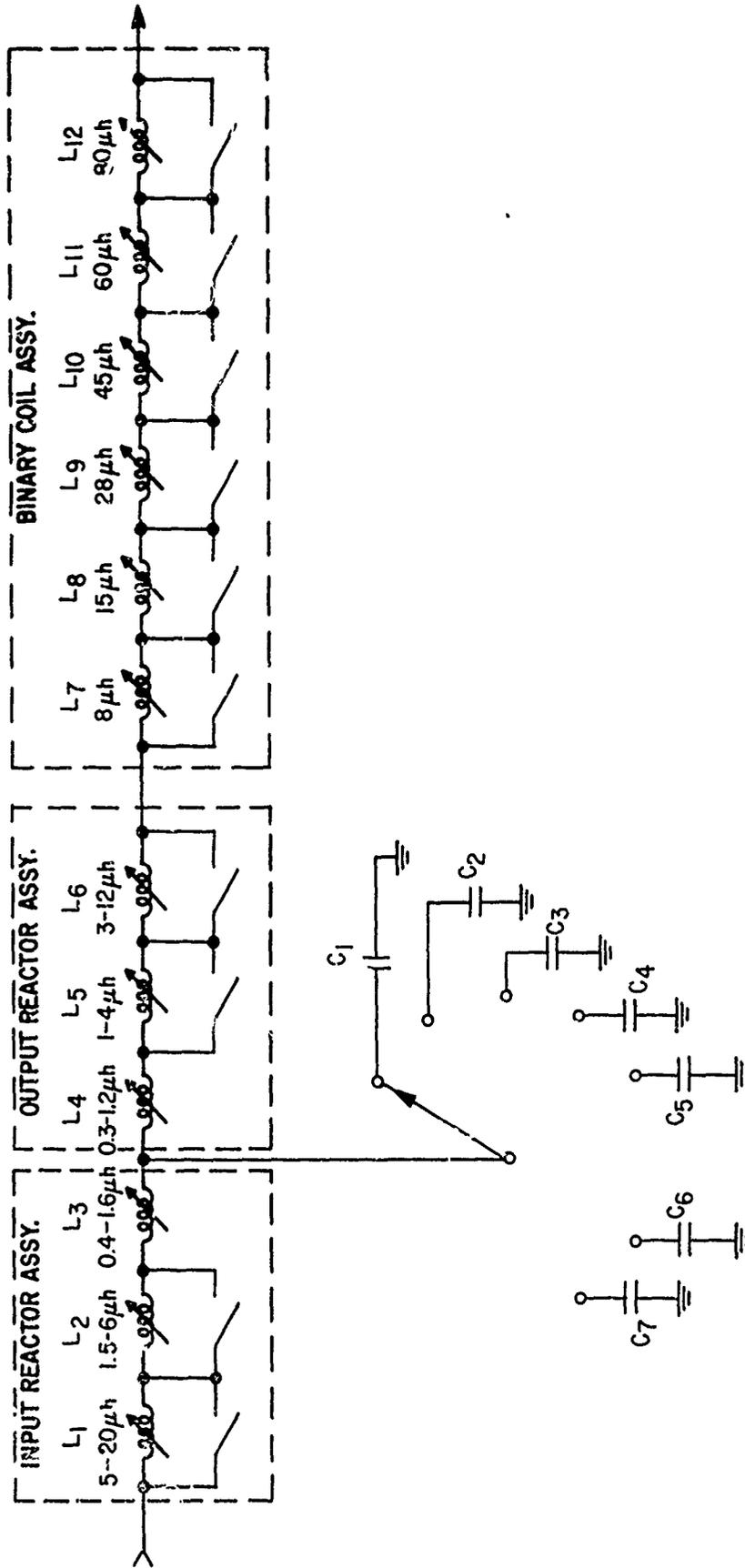
#### C-4.3.1 RF Tuning Element

The basic saturable core reactor as developed and refined by Avco has been explained in many reports. The theory is quite well developed and will not be discussed here except to summarize its usage in the present coupler system.

The basic idea of the antenna matching network described here depends on the concept that the relative permeability ( $\mu_r$ ) of a core material and the inductance of a coil wound on this core material can be changed by applying an external magnetic field.

The second important concept is the use of a permanent magnet to supply the external field and the changing of the strength of this magnet by current pulses through a coil surrounding it. Changing the field strength of the magnet will cause changes in  $\mu_r$  of the core material. In this way a range of inductance values is available from the RF inductor, each corresponding to a certain field strength of the permanent magnet. This is true if the magnet changes from zero field to its maximum value. In the case where the magnet is allowed to change from maximum field via zero field to maximum reversed polarity field, there will be two values of magnetic field strength corresponding to each inductance value.

The basic construction of the saturable reactor assembly was indicated in figure C-13.



BANDSWITCHED CAPACITOR ASSEMBLY

Figure C-16. Tuning Network Schematic Diagram

- Permanent Magnet Material. Given the requirement of small weight and volume, the magnetic material to be selected should be the one with the highest possible energy product (BH). Large energy products are obtainable in ceramic type magnets. However, the coercive forces in magnets of this type are very high, and so it is virtually impossible to simply change the magnetic field strength with a pulsed coil. Thus a metallic magnet material was selected, exhibiting a very high energy product. The particular magnet chosen is an Indian General Alnico V-7 with  $(BH) = 5.9 \times 10^4$  joule/m<sup>3</sup> (or  $7.5 \times 10^6$  gauss-ersted).
- Ferrite Material. The choice of ferrite material was based on the following considerations:
  1. Useful frequency range — The material should exhibit a reasonable Q from 2 — 30 MHz.
  2. Rate of change — The material should exhibit as large a change of  $\mu_r$  as possible under influence of an external field.
  3. Power capability — This requirement relates to the property of the ferrite to withstand high RF power levels without introducing considerable distortion.

The rate of change as defined under consideration 2 above will depend on the  $\mu_r$  without magnetic field. The higher the initial  $\mu_r$ , the larger the change. However, a high  $\mu_r$  is contradictory to the requirement of a large useful frequency range. Therefore, a compromise has to be effected. The best ferrite material found that gives good characteristics over a wide frequency range is Ferroxcube's 4E2.

It should be noted that two different methods of applying the magnetic bias field can be used. The parallel bias will sustain a field mainly parallel with the alternating field generated by the RF coil. The orthogonal bias consists of a field mainly orthogonal to the alternating RF field (see figure C-17). The advantage of the orthogonal field is the increase in Q that can be achieved. In brief, the losses in ferrite at high frequencies are due to two phenomena: domain wall motion and electron spin resonance. The first one appears at lower frequencies and the second one appears at higher frequencies although the two peaks may almost coincide. By applying an orthogonal field of large enough magnitude, the domain walls will disappear and the electron spin will precess about the direction of the external field, thereby strongly reducing the losses. However, one should realize that a very strong field is necessary to achieve this. This means that the available variation of inductance is limited, because the external field may only be allowed to change about the higher values of field strength.

The selected approach to the problem has been to use a ferrite material with a high enough initial Q at the high frequencies without any external field applied. Then a parallel external field is applied to change the  $\mu_r$  of the material. As can be seen from figure C-18, a parallel field will produce a larger change in inductance than an orthogonal field of the same strength. A preference for a

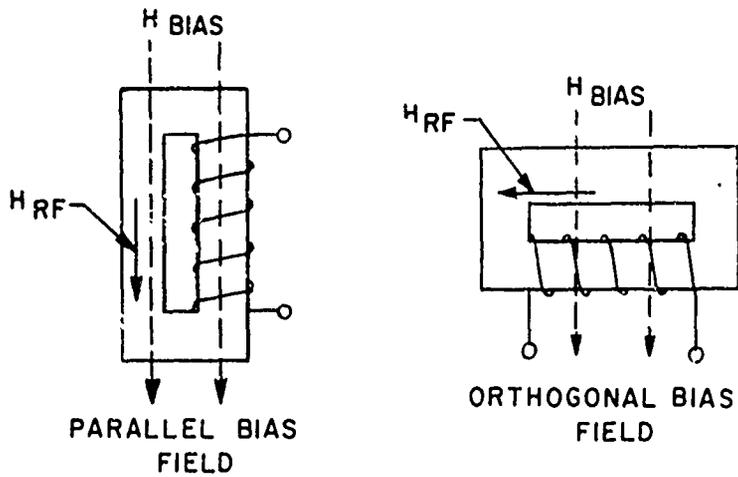


Figure C-17. Bias Methods

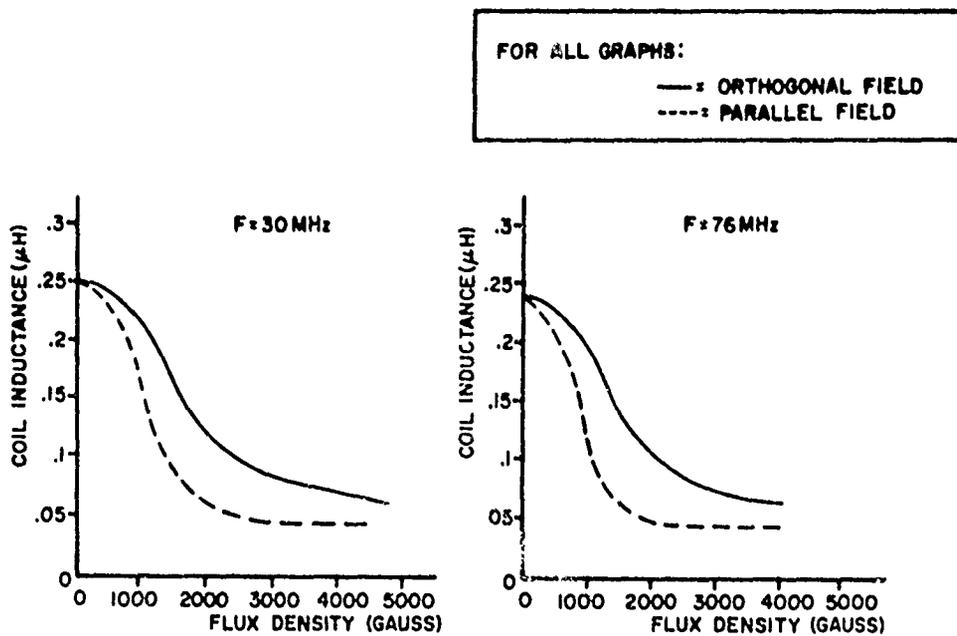


Figure C-18. Inductance of Coil as a Function of Flux Density

parallel field leads to still another design concept. When the ferrite cube in figure C-19 is considered, it can be noted that the DC field (dotted lines) is only partially parallel to the RF field (solid lines). At the end pieces (shaded) the RF and DC fields are almost perpendicular to each other. Because parallel fields are favored for the reasons previously explained, the ratio  $L/d$  should be made as large as possible so that the relative size of the end pieces, where the fields are not parallel, is decreased.

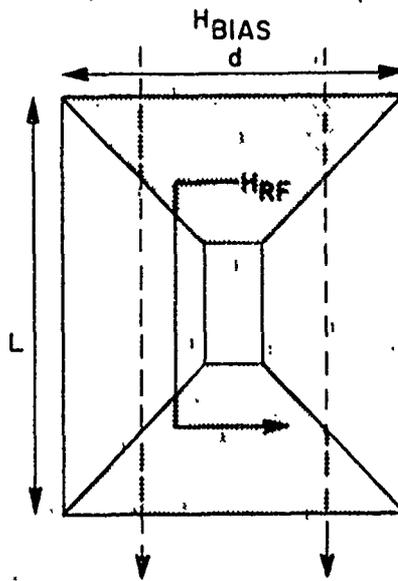


Figure C-19. Parallel and Orthogonal Bias

Given a parallel field, it is easy to see why an inductance change will take place (see figure C-20). The initial  $\mu_r$  of the ferrite, given by the slope of the BH curve of the material, changes when the bias point is moved along the BH curve, as a result of an externally applied magnetic field. Although, theoretically,  $\mu_r$  can be reduced to zero with a high enough external field the field strength required to do this is very high. In this application, fields are used with a strength which reduces  $\mu_r$  to about 4, from an initial value of 16.

Another consideration becomes evident from looking at figure C-20. The RF field has to be much smaller than the DC bias field. The actual limit here is imposed by the distortion that can be tolerated on the RF waveform as a result of the nonlinearity of the ferrite.

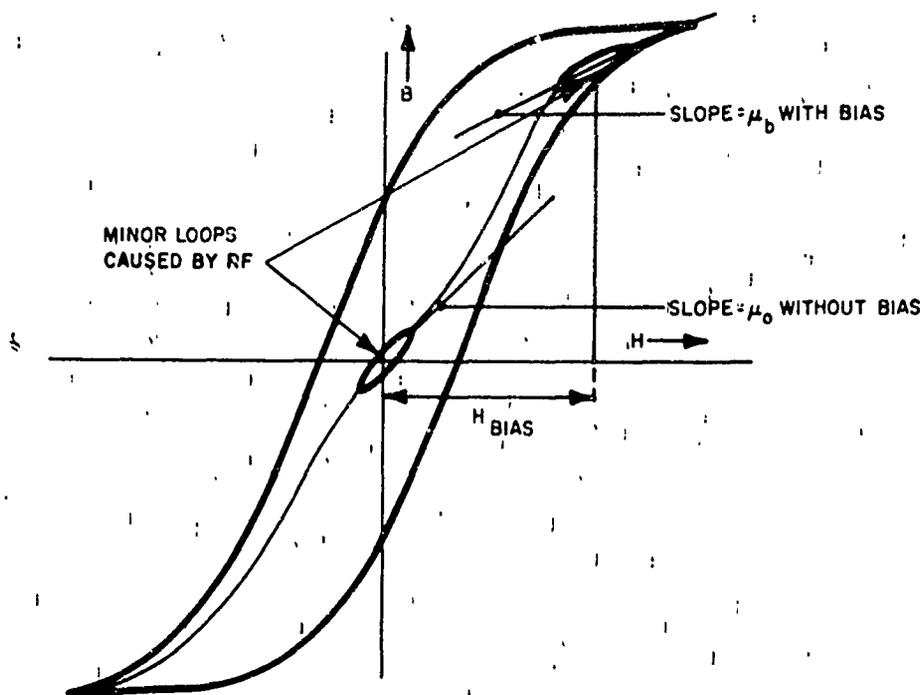


Figure C-20. Hysteresis Loops

#### C-4.3.2 Tuning Concept

The basic network tuning method proceeds in four steps. These are:

1. Rapid Scan Mode
2. Normal Scan Mode
3. Coarse Tune Mode
4. Fine Tune Mode

A simplified version of the overall tuning philosophy can be described in the following manner (see figure C-21).

Assume that  $L_2$  has zero value (in reality it is minimum). By varying  $L_1$  over the range  $0 \rightarrow \infty$ , the antenna plus  $L_1$  impedance will be varied over the range

$$(R_A + j X_A) \rightarrow (R_A + j \infty)$$

In a Smith diagram (see figure C-22) this constitutes part of a reactance circle for  $R =$  constant. In the given example in figure C-22 where

$$R_A + j X_A = 100 - j100$$

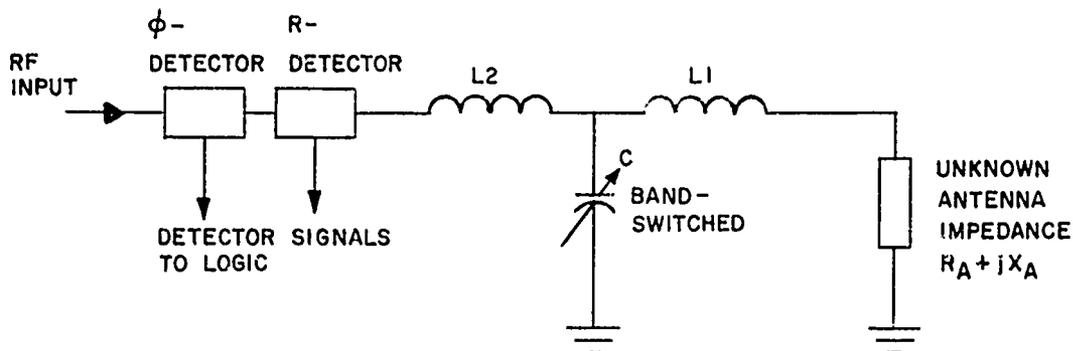


Figure C-21. Network Example for Tuning

the circle I is the image of all impedances that can be acquired by varying the value of L1. Now by considering the parallel capacitor C with reactance  $X_C$ , the input impedance of the coupler network shifts in the negative reactance direction along the constant conductance circles. If it is assumed that  $X_C = -10$  millimhos, circle II is the result in the example of figure C-22. Two intersections exist with the circle  $R = 50$  ohms. At those intersections the impedance (in ohms) is given by

$$\bar{Z} = 50 + jY$$

If the inductance L2 is added to the network and a value of L2 selected in such a way that

$$X_{L2} = -Y$$

where  $X_{L2}$  is the reactance of L1, then the final network will exhibit an input impedance of  $50 + j0$  ohms; i.e., the network is tuned for 50 ohms.

It should be pointed out that, in general, it is not necessary for circle II to intersect with the  $R = 50$ -ohms circle. If no such intersection exists, the desired tuning is not possible. By choosing  $X_C$  large (i.e., having a small capacity) an intersection for circle II with the  $R = 50$ -ohms circle can always be found. However, for very large values of  $X_C$  this intersection will move in the direction

$$R = 50, |Z| \rightarrow \infty$$

So very large values of  $X_{L2}$  will be necessary to achieve tuning. Furthermore, intersection A might be located on the positive reactance side of the 50-ohm circle (see circle III). This means that tuning for  $\phi = 0^\circ$  at A' is only possible if  $X_{L2}$  would be negative; so L2 then has to be capacitive. Because of the exclusion use of saturable core reactors as variable elements, points like A' are excluded and for tuning only the intersections of the kind at B' are to be used.



Now, the following tuning procedure can be established:

1. Make  $X_{L2}$  as small as possible by minimizing  $L2$ .
2. Select the right capacitor for  $C$  (its value depends on the band corresponding to the selected frequency).
3. Vary  $L1$  until  $\text{Re}(\bar{Z}) = 50$  ohms.
4. Vary  $L2$  until  $\text{Im}(\bar{Z}) = 0$  ohm.

In order to make the decision

$$\begin{aligned} &\text{Re}(\bar{Z}) < 50 \text{ ohms or} \\ &\text{Re}(\bar{Z}) > 50 \text{ ohms or} \\ &\text{Re}(\bar{Z}) = 50 \text{ ohms} \end{aligned}$$

a resistance detector is used. A phase detector is used for the logical decision:

$$\begin{aligned} &\text{Im}(\bar{Z}) < 0 \text{ or} \\ &\text{Im}(\bar{Z}) > 0 \text{ or} \\ &\text{Im}(\bar{Z}) = 0 \end{aligned}$$

The description of these detectors is given in previous reports.

If  $L1$  is varied as indicated in step 3 above, the relationships indicated in figure C-23 exist. Figure C-23C is constructed from figure C-23B, assuming the following logic relationship:

If output R detector is +5 volts, increase  $L1$   
If output R detector is -5 volts, decrease  $L1$

From figure C-23B it is now evident that "B" is a stable tuning point, whereas point "A" is unstable. This is very suitable, because as was stated before, point A does not necessarily enable the phase coil ( $L2$ ) to tune for positive values of  $X_{L2}$ . However, if initially  $X_{L1}$  is smaller than the value of  $X_{L1}$  at "A", no tuning is possible, because the indicated logic relationship would lead to a continuously decreasing  $X_{L1}$ . This problem can be solved by an initial inversion of the logic relationship.

If the input impedance  $\bar{Z}$  has a negative phase angle before  $X_{L2}$  is varied from minimum and after  $X_{L1}$  is tuned for  $\text{Re}(\bar{Z}) = 50$  ohms, tuning of  $X_{L2}$  (after it has been switched in) is accomplished in a simple way by relating the phase detector output polarity to the increase or decrease of  $X_{L2}$ . At this point it is noticed that it is not necessary to make use of the conditions

$$\begin{aligned} &\text{Re}(\bar{Z}) = 50 \text{ ohms and} \\ &\text{Im}(\bar{Z}) = 0 \text{ ohm} \end{aligned}$$

for which the output of the R and phase detectors would be zero. The only information needed consists of:

$$\begin{aligned} &\text{Re}(\bar{Z}) < 50 \text{ ohms or } > 50 \text{ ohms and} \\ &\text{Im}(\bar{Z}) < 0 \text{ ohm or } > 0 \text{ ohm} \end{aligned}$$

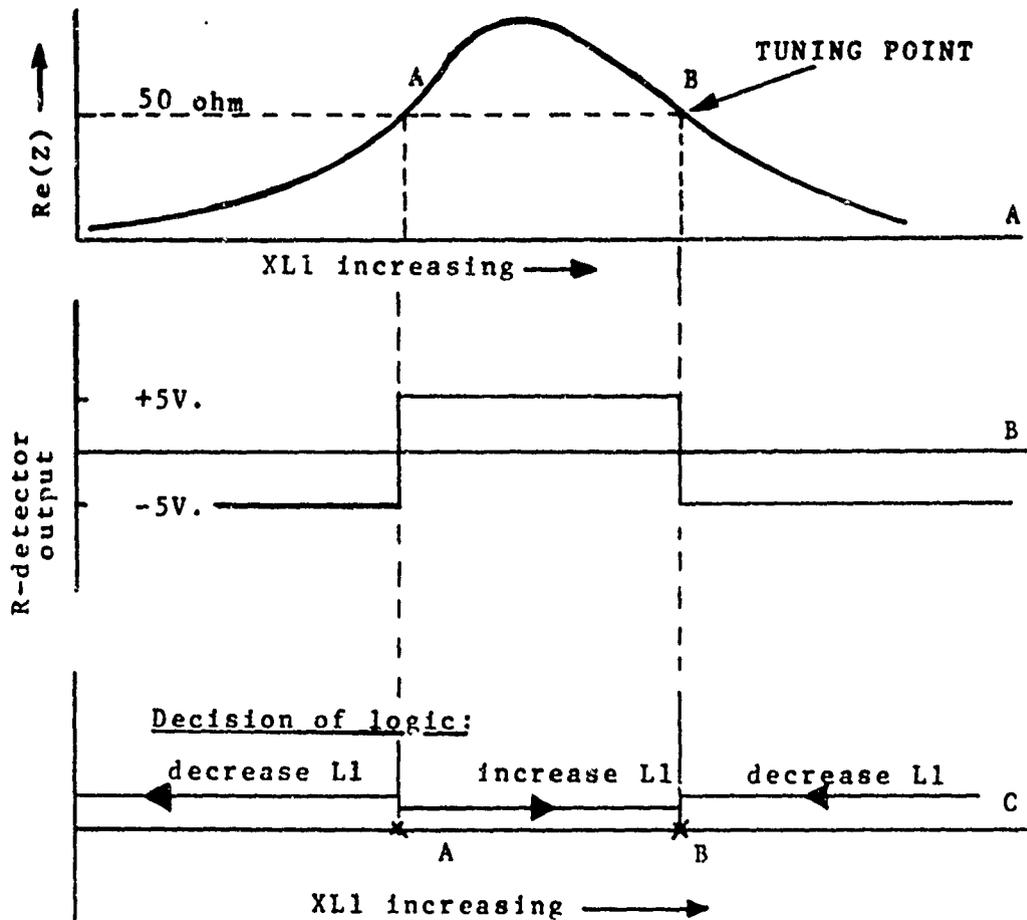


Figure C-23. Relationships in Output Arm of Tuning Network

This means that the system only recognizes the network's "absolute position" in the tuning diagram so it can make the decision in which direction to move; however, the input signals from the R and  $\phi$  detectors do not contain sufficient information to decide on "quality" of tuning. Nevertheless, with this information the system will be able to move into the direction of a tuned condition...

#### C-4.3.3 Practical Tuning Method

The conceptual tuning method described in the last section must be modified in reality for several reasons. Among these reasons are:

1. Faster tuning times required
2. Stray reactances
3. Finite element Q's
4. Detector inaccuracies

Therefore, the proposed method in four parts as mentioned previously functions as described below:

- Initial Conditions. Two sets of initial conditions can be discussed: a "home" condition and the "present" state.

1. Home Condition. This condition has both L1 and L2 at minimum and the shunt C in the proper band. The coupler should automatically go to a "home" condition under any of these occurrences:

- a. Radio Silence Mode
- b. Frequency Band Change
- c. Power Switch Turned On
- d. Whenever a "fine tune" retune attempt fails

2. Present State. Whatever elemental condition the coupler was last in.

- Rapid Scan Mode. This mode, described briefly before, operates in the following manner:

Starting from a "home" condition, the output reactor is swept with a scan pulse through its complete  $L_{\min} \rightarrow L_{\max}$  range while monitoring the R-detector at the same time to see whether a valid crossing of the 50-ohm circle is achieved. If no crossing is indicated, the smallest fixed inductor is switched in and the reactor is reset and swept again through its range. This process is repeated until a valid 50-ohm crossing is obtained. Then logic control goes to Normal Scan Mode.

- Normal Scan Mode. Following the initial R-crossing indication, the reactor is reset and incrementally pulsed through its range pulse-by-pulse in order to detect this R-crossing with a smaller L increment. As was previously mentioned, it may be necessary to switch to the next coil combination to compensate for the  $X_L$  overshoot in the rapid scan pulse. Once the first R-crossing is again detected in the Normal Scan Mode, the logic control goes to Coarse Tune Mode.

- Coarse Tune Mode. The procedure followed in this mode is well described in Technical Report ECOM-0101-1. It is an iterative type procedure where the logic control is determined by which zone of the Smith Chart the input impedance is in at any given sample period. The ultimate aim of this mode is to achieve the first zero-phase crossing. Once this is achieved, logic control goes to Fine Tune Mode.

- Fine Tune Mode. This mode is again zone-controlled, the ultimate aim being to achieve four consecutive crossings of the  $R = 50$  ohms, and  $\varphi = 0^\circ$  lines. This is also described in Technical Report ECOM-0101-1.

#### C-4.3.4 Calculation of Parallel Capacitor Values

The value of the capacitor to be used in each frequency band depends on the following considerations and restrictions (see figure C-21):

1. C should be as large as possible (to keep L2 as small as possible).
2. For each antenna impedance ( $R_A + jX_A$ ) which has to be tuned, the combination of C, L1, and  $R_A + jX_A$  should exhibit at least one value of L1 for which the substitute impedance of C, L1, and  $R_A + jX_A$  can be expressed as

$$\bar{Z} = 50 + jX \quad (C-1)$$

with X having a negative value.

The substitute impedance Z, for C parallel with L1 and  $R_A + jX_A$  in series, can be expressed as:

$$Z = \frac{(R_A + jX_A + jX_{L1})(jX_C)}{R_A + jX_A + jX_{L1} + jX_C} = \frac{-X_A X_C - X_C X_{L1} + jR_A X_C}{R_A + j(X_A + X_{L1} + X_C)} \quad (C-2)$$

and

$$\text{Re } \bar{Z} = \frac{-X_A X_C R_A - R_A X_C X_{L1} + R_A X_A X_C + R_A X_{L1} X_C + R_A X_C^2}{R_A^2 + (X_A + X_{L1} + X_C)^2} \quad (C-3)$$

$$\text{Re } \bar{Z} = \frac{R_Z X_C^2}{R_A^2 + X_A^2 + X_{L1}^2 + X_C^2 + 2X_A X_{L1} + 2X_A X_C + 2X_{L1} X_C} \quad (C-4)$$

Because of condition 2 above, a solution has to be found for  $X_{L1}$  from the equation

$$\text{Re } (\bar{Z}) = 50 \text{ ohms} \quad (C-5)$$

This solution represents the value of  $X_{L1}$  for which tuning is achieved. From equations C-4 and C-5:

$$50R_A^2 + 50X_A^2 + 50X_{L1}^2 + 50X_C^2 + 100X_A X_{L1} + 100X_A X_C + 100X_{L1} X_C = R_A X_C^2 \quad (C-6)$$

This is a second-order equation for  $X_{L1}$ , which might have none, two identical, or two different solutions, depending on the value of the discriminant. Two identical solutions mean (using the Smith diagram) that the circle traced by varying the value of  $X_{L1}$  is tangent to the  $R=50$ -ohms circle (figure C-24). So two identical solutions for  $X_{L1}$  will indicate the largest useful C-value. Thus the discriminant of the quadratic equation should be zero:

$$B^2 - 4AC = 0 \quad (C-7)$$

where  $B = 100 X_A + 100 X_C$

$$A = 50$$

$$C = 50R_A^2 + 50X_A^2 + 50X_C^2 + 100X_A X_C - R_A X_C^2$$

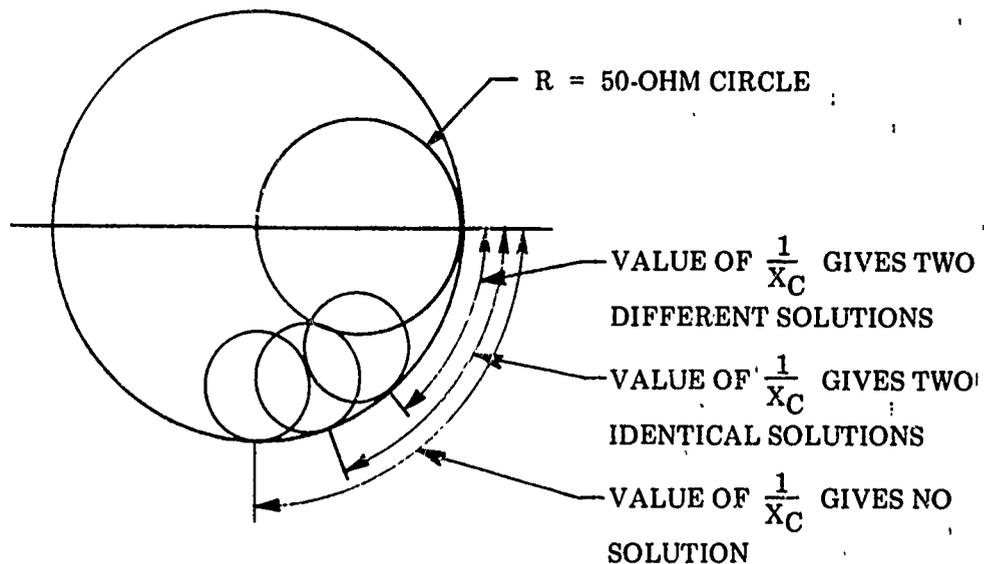


Figure C-24. Parallel C Determination

or

$$10,000 (X_A + X_C)^2 - 200 [50R_A^2 - R_A X_C^2 + 50(X_A + X_C)^2] = 0 \quad (C-8)$$

$$50 R_A^2 = R_A X_C^2$$

$$X_C = \phi \sqrt{50 R_A} \quad (C-9)$$

Because  $X_C$  represents a capacitive reactance, the negative solution for  $X_C$  is selected. This expression (equation C-9) presents a simple formula for the upper limit of the parallel capacitor. It also enables one to decide that for a given capacitor,  $X_C = X_{C0}$ , the impedance values that cannot be tuned are contained inside the circle  $R = R_{A0}$  in the Smith diagram, where

$$R_{A0} = -X_{C0}/50 \quad (C-10)$$

However, there is another class of non tunable impedances. These are the values of  $R_A$  and  $X_A$  for which solutions of equation C-6 exist, but for which both of these solutions have negative values. If the solutions are written as:

$$X_{L1}, X_{L2} = \frac{-B \pm \sqrt{B^2 - 4AC}}{2A} \quad (C-11)$$

It is noticed that  $A = 50$  and so the denominator is always positive.

For both solutions to be negative

$$B > B^2 - 4AC$$

$$100X_A + 100X_C > \sqrt{200R_A X_C^2 - 10,000R_A^2} \quad (C-12)$$

So tuning is possible if:

$$100X_A < \sqrt{200R_A X_C^2 - 10,000R_A^2} - 100X_C \quad (C-13)$$

It is noticed that this inequality can always be satisfied by choosing  $X_A$  small enough. Here  $X_C$  has a negative value and so the term  $(-100X_C)$  gets a large positive value for small capacitors.

**C-4.3.5 Conclusion.** Even if  $X_C$  is selected according to equation C-9, tuning might still not be possible if  $X_A$  is very large (i.e., the antenna has a high inductive component) (see figure C-25). Furthermore, if tuning is not possible, the possibility exists of choosing a smaller capacitor value for  $C$  and thus raising the value of the right hand side of expression C-13 until the inequality is satisfied.

### IMMITTANCE CHART

IMPEDANCE COORDINATES — 50 OHM CHARACTERISTIC IMPEDANCE

ADMITTANCE COORDINATES — 20 MILLIMHO CHARACTERISTIC ADMITTANCE

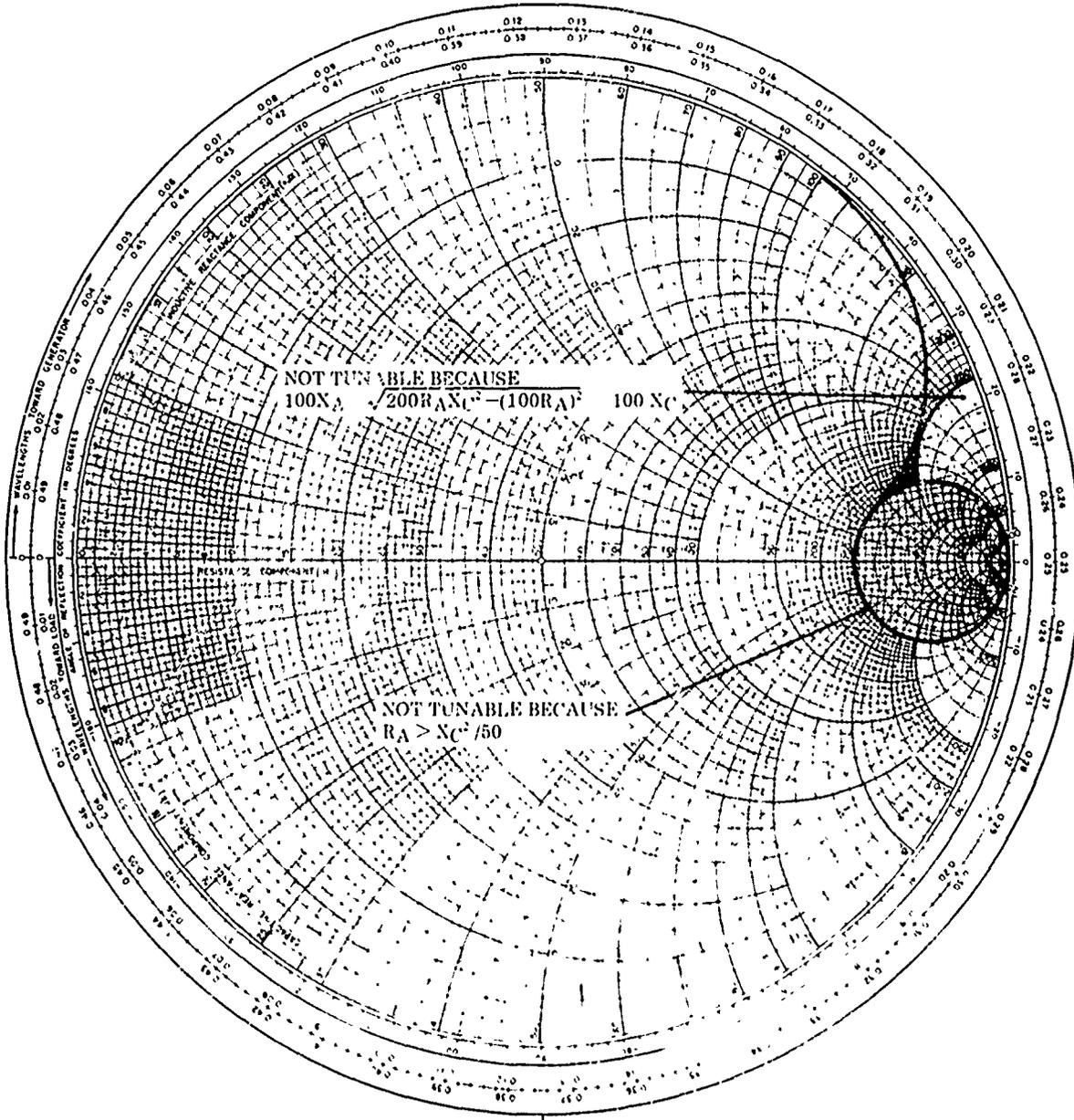


Figure C-25. Non-Tunable Areas