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Rockwell International

Collins quarterly technical report

Collins Radio Group

Packet Radio Communications

Collins Radio Group

523-0602169-001C3L
1 May 1976

(14) (12) (11)

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DISTRIBUTION/AVAILABILITY CODES		
Dist.	AVAIL. and/or SPECIAL	
A		

(12) 113p.

Rockwell International

(9) Collins quarterly technical report.

(6) Packet Radio Communications.
Collins Radio Group

Principal Investigator:
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ARPA Order ~~No.~~ 2305
Program Code No. ~~P3P10~~
Contractor: Collins Radio Group
(15) *~~Contract No.~~ DAHC15-73-C-0192*
Effective Date: 1 October 1973
Expiration Date: 31 December 1976

Sponsored by:
Advanced Research Projects Agency
Department of Defense

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Government Telecommunications Division
Collins Radio Group
Rockwell International
Dallas, Texas 75207

Printed in United States of America

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SUMMARY

A. TECHNICAL PROBLEM

This project is one part of a larger effort directed at extending packet switching technology into the area of radio communications. The technical investigations are focused on a Packet Radio Network (PRN) to serve fixed and mobile digital terminals in a tactical command and control environment. The objectives of the Collins investigation are:

- Perform research covering the application of radio frequency technology to packet switched communications.
- Participate in the overall ARPA Packet Communications Technology Program under the guidance of the Packet Radio Communications Working Group.
- Develop experimental equipments to support the Packet Radio System/Network experiments.
- Recommend a system architecture to provide terminal-to-terminal security for the packet switched radio network.

B. GENERAL METHODOLOGY

The approach to the packet radio investigation is a combination of theoretical studies, equipment design efforts, and laboratory experiments. Work has been directed in each of the following areas:

- Radio Design and Studies. Studies include development of a design plan for an upgraded packet radio and investigation of critical issues and limitations in realizing a reduced size packet radio.

- Experimental Equipment Development. Circuit design, fabrication, assembly, and testing of three packet radios of upgraded design, including software development. Build and test of nine packet radios of the existing design.

- System Support. Providing support for equipment integration into the experimental test site, including reconfiguration of equipment and software.

C. TECHNICAL RESULTS

Specific results and accomplishments made during the past quarter include the following:

1) The balance of the original packet radios have been completed. Seven PRs have now been transferred to the test bed.

2) A draft of the design plan for the upgraded packet radio has been completed.

3) The revision to the "PR Channel Access Protocol" has been written, documented, and transferred to the test bed. The "Process Control Program" (PCP) has been revised to incorporate SPP and has been documented and transferred.

4) The reduced size PR study has been completed and the final report is included in this quarterly report.

D. PLANS FOR NEXT QUARTER ACTIVITIES

The activity during the next quarter will be focused on the detailed design development of the upgraded packet radio. Development will continue on the build of the nine packet radios of the existing design. Software development will include the incorporation of measurements into the PR protocol.

1.0 INTRODUCTION

In order to realize the full potential of the packet switched radio concept the radio units should be small and light weight. Maximum system flexibility and responsiveness in a tactical environment can be achieved only if the network elements are highly mobile and versatile. A self-contained mobile terminal that is both small and light enough to be hand-carried, while maintaining performance requirements, would allow the highest degree of mobility. To this end, an investigation of size reduction of a packet radio was initiated with two primary objectives in mind. These objectives were: 1) to identify in specific terms the set of constraints to size reduction of a packet radio; and 2) to explore a set of solutions or alternatives to size reduction consistent with performance objectives.

The major effort has been devoted to investigating size reduction in the transceiver section of the packet radio (as distinguished from the digital section). This effort has revealed that the implementation of certain RF functions under performance requirements imposed by other circuit functions and system network requirements present constraints to size reduction. For example, surface acoustic wave matched filters are relatively small in size and consume no power; however, the local oscillator stability necessary for their use causes the local oscillator system to be large and power consuming. As an example of how the system network requirements impact size in packet radio, consider an increase in transmit power requirement. The number of stages in the transmit chain increases, efficiency drops and power dissipation increases, causing the heatsink requirements to increase. More transmit power and reduced efficiency causes increased energy source drain which in turn causes the energy source size to increase. It is these interrelations and their impact size that have been examined and described in this report.

This investigation concludes that the major constraining functional areas to size reduction in packet radio are: 1) RF power amplification, 2) frequency generation, 3) filtering, 4) energy storage and 5) packaging.

It is in these functional areas that significant contributions to size reduction can be made. A broad spectrum of approaches has been considered for achieving size reduction in these areas. This includes relaxing or eliminating some system imposed performance measures, exploring new and more efficient techniques to circuit realizations, and exploiting the advantages of developing technologies. A significant size reduction can be achieved for a reduced power, single channel, single data rate, non-modular packet radio.

Several developing technologies have been examined and those that will have significant impact on packet radio include: 1) the surface acoustic wave oscillator, 2) charge-coupled devide matched filter, and 3) rechargeable Lithium electrochemical cells. If the long term stability of the surface acoustic wave oscillator can be improved, its small size and low power consumption would represent the answer to the frequency generation problem in small packet radios. Charge-coupled device signal processing in packet radio shows considerable promise. These applications include matched filtering, delay, and integration. Charge-coupled device matched filters in packet radio have the potential for very small size, increased flexibility, and increased reliability. Lithium electrochemical cells have the highest energy densities of any available electrochemical system, but they do not currently exist as rechargeable units.

The results of the investigation of size reduction in packet radio are described in three major sections that follow. Section 2.0 addresses those functions in packet radio as subsections that are considered constraints to size reduction. Within these functional areas, basic limitations are highlighted that impact the size of the packet radio. Section 3.0 discusses various alternative solutions and techniques to relieve these constraints. Sections 2.0 and 3.0 are similarly organized to permit ease of correlation between constraints and limitations in a given functional area and potential solutions to size reduction. The last section presents a design plan description of a hand-held packet radio utilizing current technology and applying many of the techniques described in Section 3.0.

2.0 BOTTLENECKS TO SIZE REDUCTION IN PACKET RADIO

2.1 General

The size of packet radio is the result of implementing a specific set of features such as the amount of power to be transmitted, the ability to select one of twenty channels for tranceiver operation, the use of a spread spectrum modulation technique, etc. These features find their origin in the packet radio system specifications and in most cases impose stringent performance requirements on the functions within packet radio. Stringent performance requirements caused size reduction in many functional areas of packet radio to be difficult. Those functional areas considered to offer major resistance to size reduction are RF power amplification, frequency generation, filtering (including matched filtering), energy storage and packaging. The relationship between the functional areas of RF power amplification, frequency generation, and matched filtering and other functions in the radio section of packet radio is diagramed in Figure 2.1.1.

The primary thrust of the investigation into size reduction was concentrated in the functional areas outlined above and as such, was focused on the transceiver section of packet radio. The application of large scale integration to digital functions has been proven. This is not true of the RF analog functions. Monolithic integration does not offer the means to miniaturization for the RF functions as it does for the digital functions.

When a given function in packet radio is implemented, the volume and weight consumed is a result of the interactive combination of packet radio system requirements, requirements imposed by other functions, and limitations that arise from the technology, materials, or techniques employed. The purpose of this section is to identify the set of requirements, regardless of origin, and the set of limitations in each functional area that serve to constrain size reduction. It became apparent during the

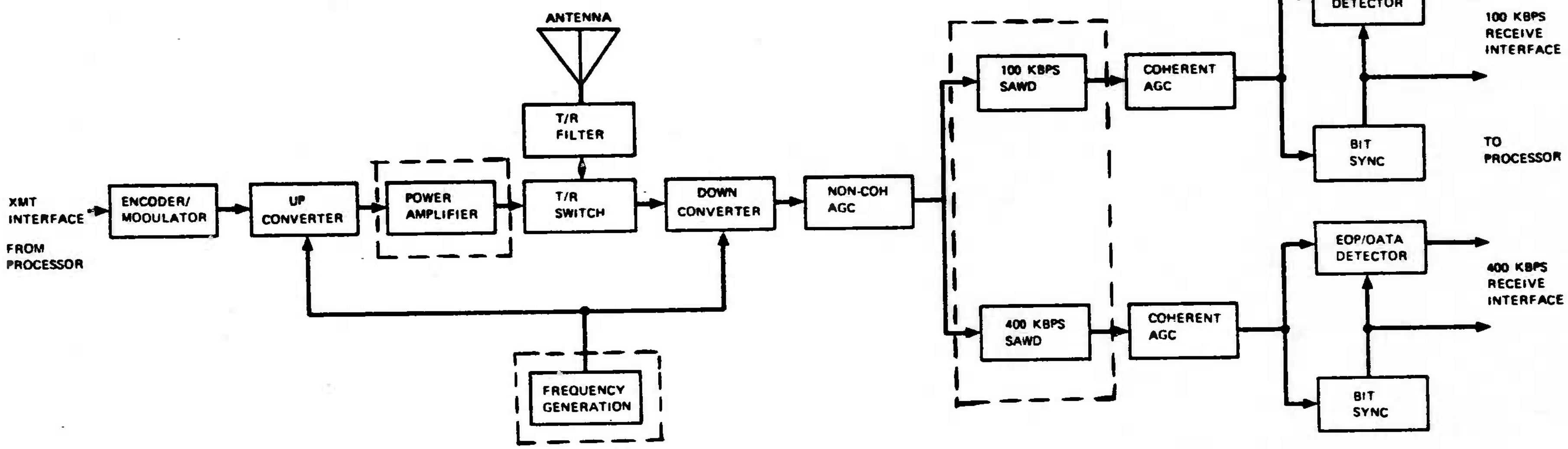

FIGURE 2.1.1 FUNCTIONAL BLOCK DIAGRAM OF RF SECTION IN PACKET RADIO

investigation that in order to identify a meaningful set of requirements and limitations, one must understand the interrelationships existing between various functions in packet radio and the impact on these functions that the packet radio system requirements impose. Therefore, a secondary objective of Section 2.0 is to provide a description of the interrelationships.

In the discussions that follow, the issues associated with RF power amplification are presented in Subsection 2.2. How limitations on power output, gain, and efficiency of transistor amplifiers affect circuit area required, the size of the heatsink and the size of the power source are analized. Subsection 2.3 points out how stability requirements and operating frequency relate to size of the local oscillator. The problems associated with realizing small low loss filters are identified in Subsection 2.4. The limitations of surface acoustic wave matched filters as they relate to packet radio requirements are also presented. Energy storage is a serious constraint to size reduction. Subsection 2.5 explores the size of the energy source and its relation to fundamental electrochemical principles. Finally, Subsection 2.6 discusses issues associated with package design of the packet radio including trade off between size reduction and modularity.

2.2 RF Power Amplification

The function of RF power amplification contributes to the size of packet radio in three areas: 1) circuit area, 2) heatsink, and 3) power source. The actual circuit area required for generation a given amount of power represents a smaller impact on total volume than either the heatsinking requirements or the size of the power source for maintaining a reasonable duration of operation.

The circuit area required for the RF amplifier chain is determined by the number of stages to achieve the desired power out and the approach to realizing the matching networks. Transistor amplifiers operating at 1800 MHz are capable of a maximum of ten watts output power and power gains of 8-10 dB. To achieve a ten watt output capability, the power amplifier chain must provide approximately 50 dB of gain to the signal emerging from the up-converter. This requires at least four stages preceding the power amplifier. The input and output matching networks of each stage occupy the majority of the circuit space required by the amplifier chain.

The heatsink requirements arise from the limitation on transistor amplifier efficiency and the need to limit the junction temperature of the active devices. The efficiency of the Class C saturated amplifier at 1800 MHz is limited to 50%; the efficiency of amplifiers operated Class A is between 10 and 15%. Since the predrivers are normally operated Class A, the overall efficiency of the power amplifier chain is 30 to 40%. Therefore, 60% or more of the DC input power is dissipated as heat. The problem is to remove the heat with sufficient heatsinking to avoid operating the transistor junctions at elevated temperatures and thus reducing the operating life of the device. The relationship [1] between junction temperature and the operating life of typical microwave power transistor devices is illustrated in Figure 2.2.1.

These curves indicate that no greater than 150°C junction temperature be maintained to insure a mean time to failure of 20 years or better.

The desired output power, the efficiency, and the required supply voltages of the power amplifier chain all influence the size of the self-contained power source. The desired output power and the efficiency interrelate to produce a compound effect on the energy requirements. An increased output power will correspondingly increase energy requirements but also will result in reduced efficiency which increases energy requirements even more. Microwave transistor amplifiers require 24-28 volts for optimum performance, since operating the amplifiers at lower supply voltages reduces power output, gain and efficiency. In order to provide supply voltages in the range of 24-28 volts from an electrochemical system, a DC-DC converter would be needed which further increases the size of the power source and the circuit area.

Packet radio operationally is pulsed system (low duty cycle). Since the packet length (2-20 ms at a maximum duty cycle of 50%) is much longer than the thermal time constant (in the order of a few microseconds) of the semiconductor junctions, operation appear continuous with average power dissipated determined by the duty cycle. Therefore, the fact that semiconductor junctions have finite thermal time constants is not advantageous to minimum heatsink design in packet radio.

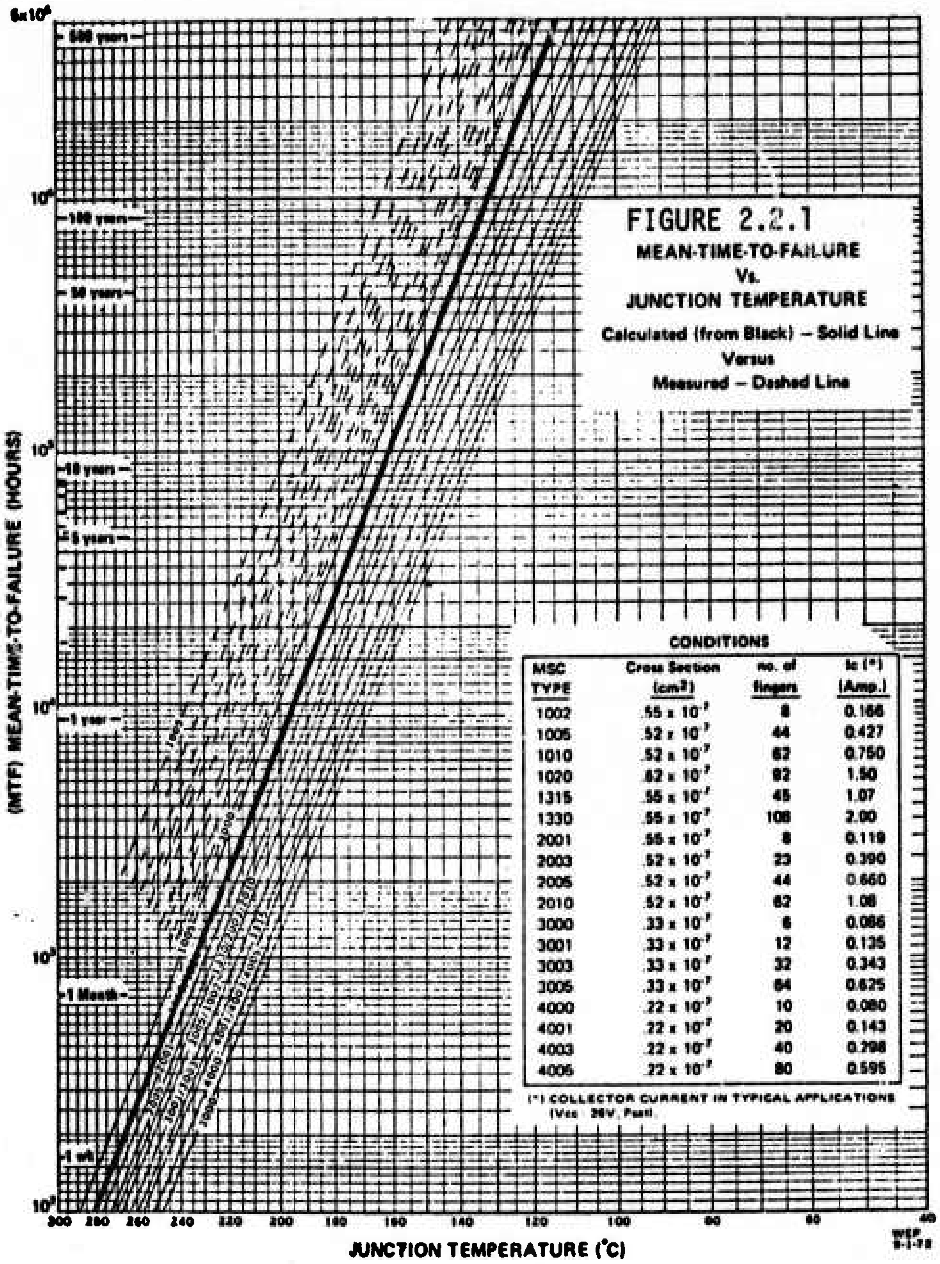

FIGURE 2.2.1

MEAN-TIME-TO-FAILURE
Vs.
JUNCTION TEMPERATURE

Calculated (from Black) – Solid Line
Versus
Measured – Dashed Line

CONDITIONS

MSC TYPE	Cross Section (cm²)	no. of fingers	Ic (*) (Amp.)
1002	$.55 \times 10^{-7}$	8	0.166
1005	$.52 \times 10^{-7}$	44	0.427
1010	$.52 \times 10^{-7}$	62	0.750
1020	$.62 \times 10^{-7}$	82	1.50
1315	$.55 \times 10^{-7}$	45	1.07
1330	$.55 \times 10^{-7}$	108	2.00
2001	$.55 \times 10^{-7}$	8	0.119
2003	$.52 \times 10^{-7}$	23	0.390
2005	$.52 \times 10^{-7}$	44	0.660
2010	$.52 \times 10^{-7}$	62	1.08
3000	$.33 \times 10^{-7}$	6	0.066
3001	$.33 \times 10^{-7}$	12	0.135
3003	$.33 \times 10^{-7}$	32	0.343
3005	$.33 \times 10^{-7}$	64	0.625
4000	$.22 \times 10^{-7}$	10	0.080
4001	$.22 \times 10^{-7}$	20	0.143
4003	$.22 \times 10^{-7}$	40	0.298
4005	$.22 \times 10^{-7}$	80	0.595

(*) COLLECTOR CURRENT IN TYPICAL APPLICATIONS (Vcc = 28V, Pout)

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2.3 Frequency Generation

System decisions, environmental operating conditions related to temperature and ruggedness, and performance objectives for the modulator/demodulator process all contribute to the definition of requirements and constraints for the prime frequency sources in the packet radio. Additionally, for a small packet radio, the size and power consumption levels of the frequency generating circuitry become increasingly important. Packet radio, regardless of size, needs a stable frequency source in the range of 1500 to 1800 MHz. The conventional approach to stable frequency synthesis in the UHF and microwave band employes stable crystal oscillators followed by frequency multipliers and/or phase-locked loops to achieve the required operating frequency. Local oscillators implemented this way assume the stability of the crystal resonator.

Quartz is usually chosen for the crystal resonator because of its temperature characteristics. For high frequency operation, face shear mode vibrations in quartz plates are employed in which the resonant frequency is determined by the thickness of the plate. The usual fabrication processes limit the minimum thickness of the quartz plates to about 100 μm. The fundamental mode of the oscillator occurs when the plate thickness equals one acoustic wavelength. This establishes the maximum fundamental frequency at about 25 MHz. Resonances will also occur when integral numbers of acoustic wavelengths fit in the quartz plate thickness. Overtone operation to a maximum of about 150 MHz is possible. Overtone operation is not always desirable since overtone resonators can exhibit spurious responses at unwanted frequencies in addition to the set of overtone responses. Unless the crystal oscillator circuits used have relatively high selectivity, and consequently good stability, oscillation at spurious frequencies can occur. [2]

The stability of a local oscillator is characterized in three time frames: 1) short, 2) medium, and 3) long term. Short term implies stability over a time interval of about a microsecond to a second, medium term covers a time frame on the order of one hour to a day, and long term implies a time scale of a month to a year. Short term stability, when viewed in the frequency domain, is a measure of the single sideband phase noise about the carrier. The short term stability of a local oscillator is a function of both the crystal resonator Q and the approach to synthesis. Figure 2.3.1 shows how the phase noise can vary in the band about the carrier for the direct approach (multiplication) and the indirect approach (phase-lock system) to local oscillator synthesis. Neither method offers any clear advantage at small frequency offsets from the carrier, but the indirect method gains the advantage as the offset is increased. It is the crystal resonator Q that is most important in determining the short term stability. Crystal resonators operating in the fundamental mode generally have Q's in the range of 10,000 to 100,000. Short term stabilities of a few parts in 10^{10} can readily be achieved with quartz crystal oscillators.

Medium term instability of the local oscillator is caused principally by temperature variations. The frequency-temperature characteristics of quartz crystal resonators depend on how the crystal is cut with respect to the crystal lattice planes. The frequency variation as a function of temperature for several cuts is illustrated in Figure 2.3.2. Since the AT-cut exhibits the smallest variation in frequency with temperature, it is most often used for temperature stable sources. The AT-cut frequency-temperature characteristics are magnified in Figure 2.3.3 to illustrate the limits of crystal stability. The cut, $\theta = 35°10'$, exhibits the smallest frequency change across the temperature range. For a temperature change from 0 to 60 degrees, ± 5 parts per million (ppm) is achieved. If, as in the packet radio, the stability requirements on the local oscillator are more stringent than ± 5 ppm, temperature control or temperature compensation of the crystal oscillator is required.

Present temperature control techniques utilize a heating element, usually deposited on the quartz for maximum efficiency. The quartz is

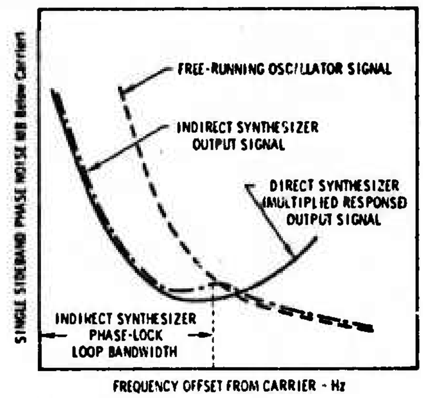


FIGURE 2.3.1 LOCAL OSCILLATOR SHORT TERM STABILITY

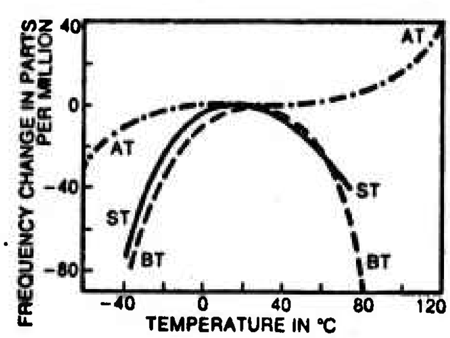


FIGURE 2.3.2 FREQUENCY-TEMPERATURE CHARACTERISTICS OF VARIOUS QUARTZ CRYSTAL RESONATORS

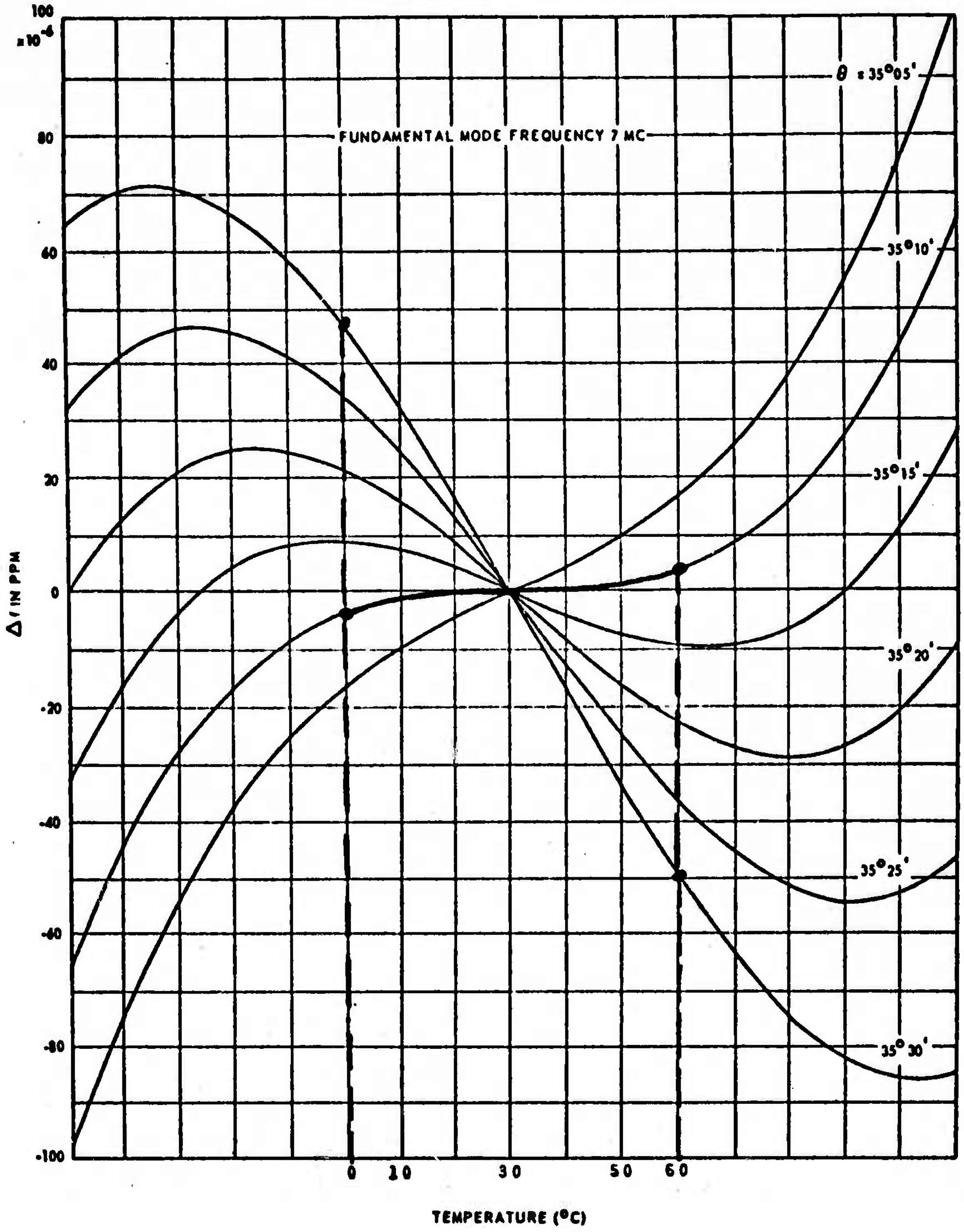


FIGURE 2.3.3 FREQUENCY - TEMPERATURE - ANGLE CHARACTERISTICS OF PLATED AT-TYPE NATURAL QUARTZ CRYSTAL RESONATORS.

then maintained at an elevated temperature during transceiver operation. Although the temperature controlled oscillator can be implemented in small size, its major disadvantages are the input power requirement for the heating element and warm-up time of several minutes. The power consumption can be as much as 500 mw at 0°C.

The temperature compensated oscillator, on the other hand, requires low input power and no warm-up time but suffers from the large size generally required for the compensation network. Temperature compensation networks make use of the non-linear reactances (obtained via thermistors and varactors) to compensate for the frequency versus temperature character of the crystal. Conventional temperature-compensated crystal oscillators with ± 1 ppm stability can occupy as much as 3 cubic inches and consume 50 mw of power.

The long term stability of a crystal resonator indicates how its frequency changes with age. Long term stability is basically a function of the cleanliness of the manufacturing process. High stability crystals require manufacture in clean rooms with purity to the molecular level. Long term stabilities of 1 part in 10^8 per week is generally the industry standard for quartz crystal resonators. [3]

Because multiple temperature-compensated or temperature-controlled crystal oscillators would be required to achieve programmability of a direct synthesis approach to a local oscillator, a phase-lock system employing a voltage-controlled oscillator (VCO) which assumes the medium and long term stability of a single stable crystal oscillator is generally the most practical route. A block diagram of a programmable phase-lock system approach to frequency generation is presented in Figure 2.3.4. The phase detection circuitry compares the phase of the frequency standard to the divided output of the VCO and if the two signals differ in phase, an error voltage is generated and applied to the VCO, causing it to correct in the direction required for decreasing the difference. The correction procedure continues until lock is achieved, after which the VCO will continue to track the incoming signal. The low pass filter filters the error voltage

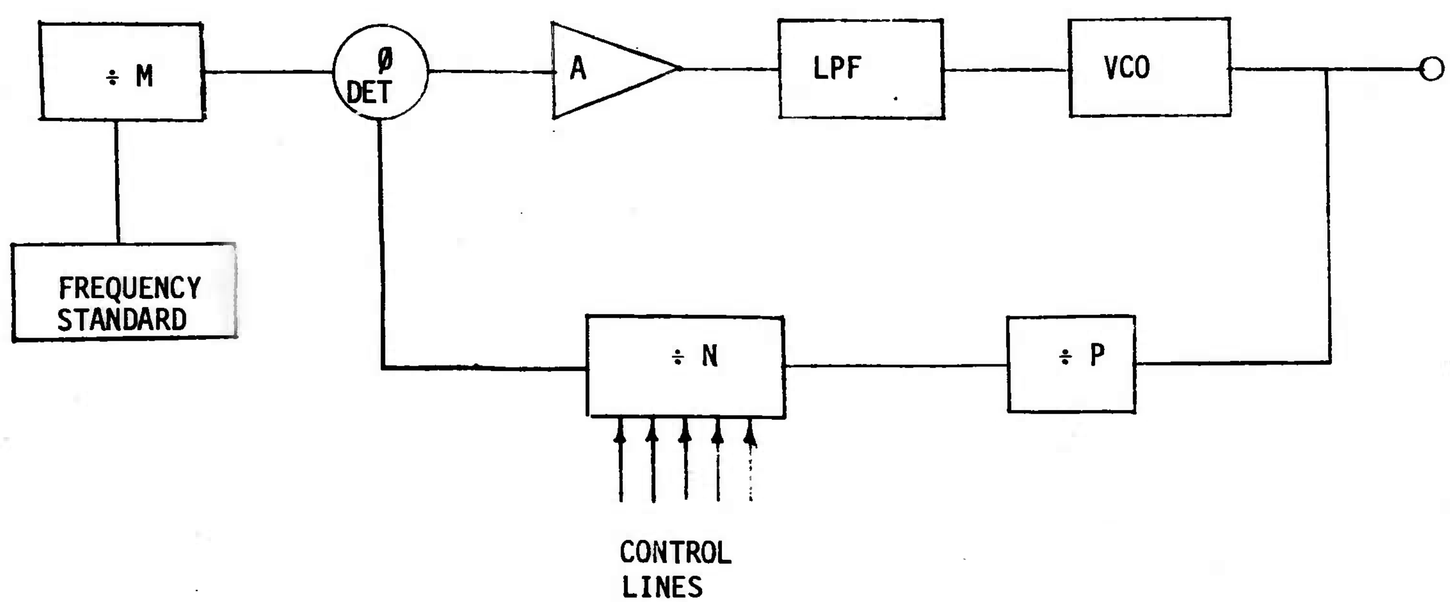


FIGURE 2.3.4 PROGRAMMABLE FREQUENCY SYNTHESIZER

and establishes the loop bandwidth. The bandwidth of the loop filter along with other loop parameters such as loop gain, and VCO gain constant determine the lock-up time, pull in range, and noise performance. The integer N is determined by the logic levels on the control lines and since the loop constrains the output frequency of the countdown circuitry to be the same as the standard, the VCO frequency must change for each new value of N.

In order to assure fast switching speed for the programmable frequency synthesizer, the countdown circuitry has to be fast and the loop lock-up time has to be short. Depending on the divide ratio and the required switching speed, the countdown circuitry is power consuming. The countdown logic necessary to convert an 800 MHz signal to 400 KHz consistent with 1 ms switching speeds will consume between 1 to 2 watts of power. The VCO with supply voltage regulation will consume more than one watt. Total power budget of the synthesizer approaches 3 watts. Since the frequency generation circuitry consumes power during all modes of transceiver operation, standby, receive, and transmit, 3 watts of continuous power consumption represents a severe constraint to size reduction.

2.4 Filter Functions

Filters in communications systems are required to discriminate and separate signals based on some signal parameter such as frequency, phase, amplitude, or combinations of these. In this context, both frequency filters and matched filters are discussed in this section as they relate to the packet radio.

Frequency Filters

The creation of unwanted frequency components is unavoidable when generating a signal suitable for communications purposes or when processing a received signal, frequency discriminating filters are required. Filter realizations can be quite varied depending on the particular application. For example, a low pass audio frequency filter can be realized adequately with discrete inductors and capacitors, whereas in the higher frequency ranges distributed element and cavity filters become more attractive.

The important factors that characterize a filter are center frequency, bandwidth, skirt selectivity, passband insertion loss, inband ripple, and ultimate stop band rejection. The main elements of a filter are reactances-- (inductive and capacitive) achieved either through lumped or distributed components. It is the losses in these inductive and capacitive components that limit many filter performance measures. Losses tend to decrease the rate of attenuation rolloff (poorer skirt selectivity), increase the attenuation within the passband and in certain cases, prohibit the realization of narrow bandpass filters [8]. A conventional measure of the quality of any reactance is the quality factor Q, which describes how many times the reactance of an inductor or capacitor is greater than its resistance. The quality factor Q equivalently expressed for a resonant combination of inductors and capacitors is

$$Q = 2\pi f \frac{\text{Energy stored}}{\text{Power dissipated}}$$

As shown in Figure 2.4.1, for a given relative bandwidth and a given number of resonant sections, there is a minimum quality factor necessary to realize the filter.

In the search for high Q resonators for effective filter design, the piezoelectric crystal exhibits very high Q (>10,000) and small size. The crystal resonator is very good for narrow bandwidth (less than a few percent) filter applications below 100 MHz. It is practically impossible to construct crystals whose fundamental frequency is above approximately 25 MHz since their physical dimensions become so small that good quality crystals cannot be produced. Harmonic operation is possible above 30 MHz, but unpredictable spurious responses discourage the use of crystal filters except for very low percentage bandwidths.

Microwave filters are designed using distributed reactances. Filter elements may be realized from sections of transmission line. The impedance of a section of transmission line can behave as a capacitance, inductance, series resonator, or parallel resonator depending on the relationship between the length of the line and the wavelength. Sections of transmission line can be made to have relative high Q and therefore are used extensively in microwave filters. Since parallel and series resonant circuits are realized with shorted and opened quarter-wave length transmission lines, the size of the filter depends on the center frequency and dielectric constant of the transmission line dielectric. Figure 2.4.2 shows how a quarter-wave section of transmission line changes with dielectric constant.

Cavity filters are another category of distributed element filters whose high Qs make them attractive at microwave frequency. Cavity resonators usually constructed with highly conductive walls can have very low losses and, therefore, very high Qs. Characterization of cavity resonators may best be done using the principle of electromagnetic theory; however, their behavior at frequencies on or near resonance is similar to lumped circuits.

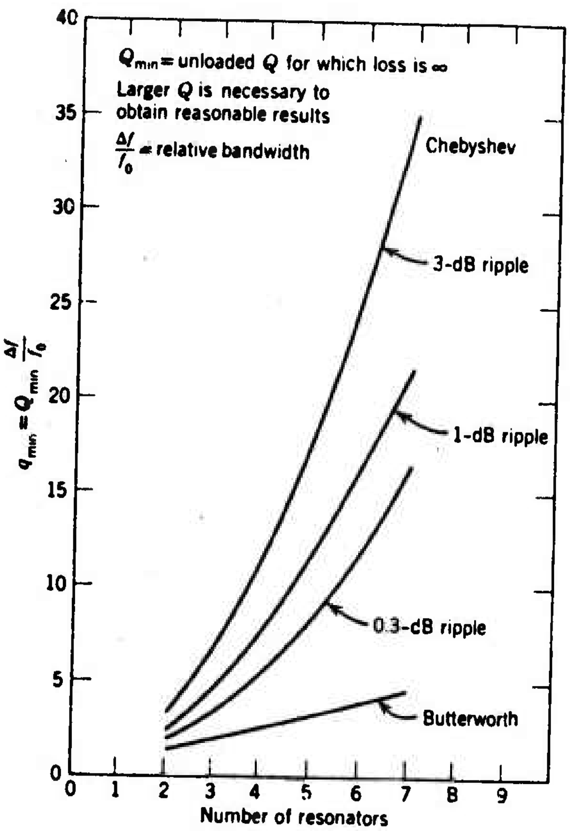
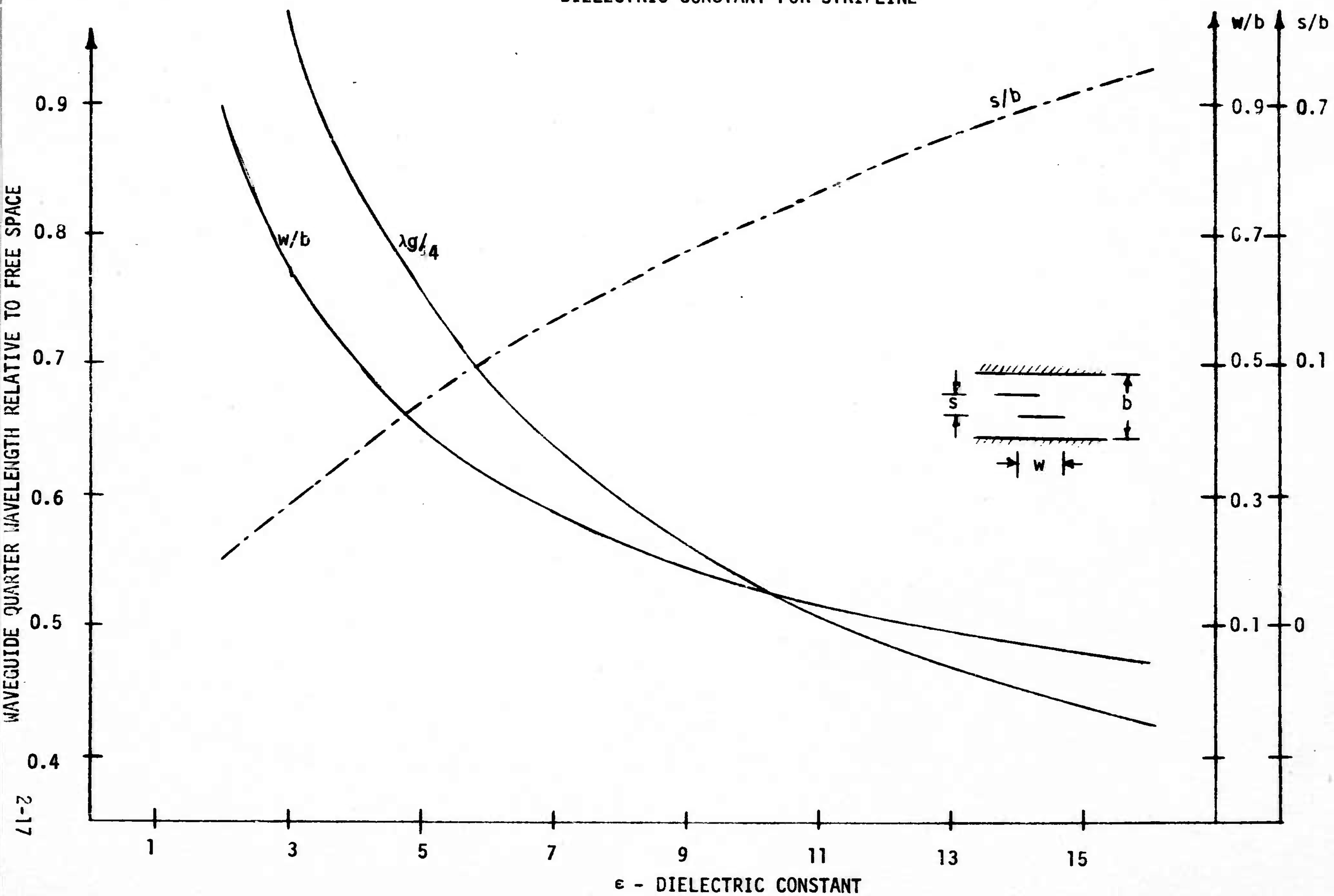


FIGURE 2.4.1 RELATIVE MINIMUM UNLOADED Q FOR BUTTERWORTH AND CHEBYSHEV FILTERS

FIGURE 2.4.2 WAVELENGTH AND ASPECT RATIOS VERSUS DIELECTRIC CONSTANT FOR STRIPLINE



A given cavity can have many possible modes, and for each mode, the resonant frequency is determined by the mode, the cavity dimensions, and the constants of the dielectric filling the cavity. For simple cavities the cavity dimensions must be on the order of a half-wave length to satisfy the electric and magnetic field boundary conditions at the frequency of interest [9]. To summarize, high Q cavity resonators can produce very low loss filters, but their size and weight make them unattractive for small radio applications.

Matched Filters

Surface acoustic wave devices (SAWD's) are used in the packet radio as matched filters in the modulation/demodulation of minimum-shift-keyed (MSK) signaling waveforms. SAW devices offer several advantages over conventional signal generators and matched filters in communications systems applications. In comparison, SAW devices are small, simple, and offer efficient and economical solutions to the synchronization of spread spectrum sequences [10].

The capabilities and limitations of SAWD's as matched filters the packet radio define the performance criteria for many functional blocks. Most performance limitations of SAWD's are direct functions of the physical properties associated with the propagation of acoustic waves of the crystal substrate material. Surface wave propagation constants of various common substrate materials are listed in Table 2.4.1. Lithium Niobate ($LiNbO_3$) has the highest piezoelectric coupling coefficient. ST quartz has the smallest temperature coefficient of delay. Bismuth Germanium Oxide ($Bi_{12}GeO_{20}$) exhibits the lowest acoustic velocity. These properties imply that devices constructed using $LiNbO_3$ have the lowest insertion loss; those built on ST quartz have the most stable temperature characteristics, and $Bi_{12}GeO_{20}$ provides the longest delay time for a given substrate size.

The insertion loss of SAWD matched filters is defined as the ratio of the power in the correlation peak to the power in the input waveform. There are several loss mechanisms responsible for the high insertion loss of SAWD

Material	Cut	Propagation Direction	Coupling Coefficient k^2(%)	Temperature Coefficient (ppm/°C)	Velocity (mm/μsec)
Quartz (HC)	-20° rotated Y	X	0.25	-32	3.209
Quartz (ST)	+42.75° rotated Y	X	0.16	0	3.157
$LiNbO_3$	Y	Z	4.5	-90	3.488
$Bi_{12}GeO_{20}$	110	001	0.85	-140	1.62

TABLE 2.4.1 SURFACE WAVE CHARACTERISTICS OF VARIOUS SUBSTRATES

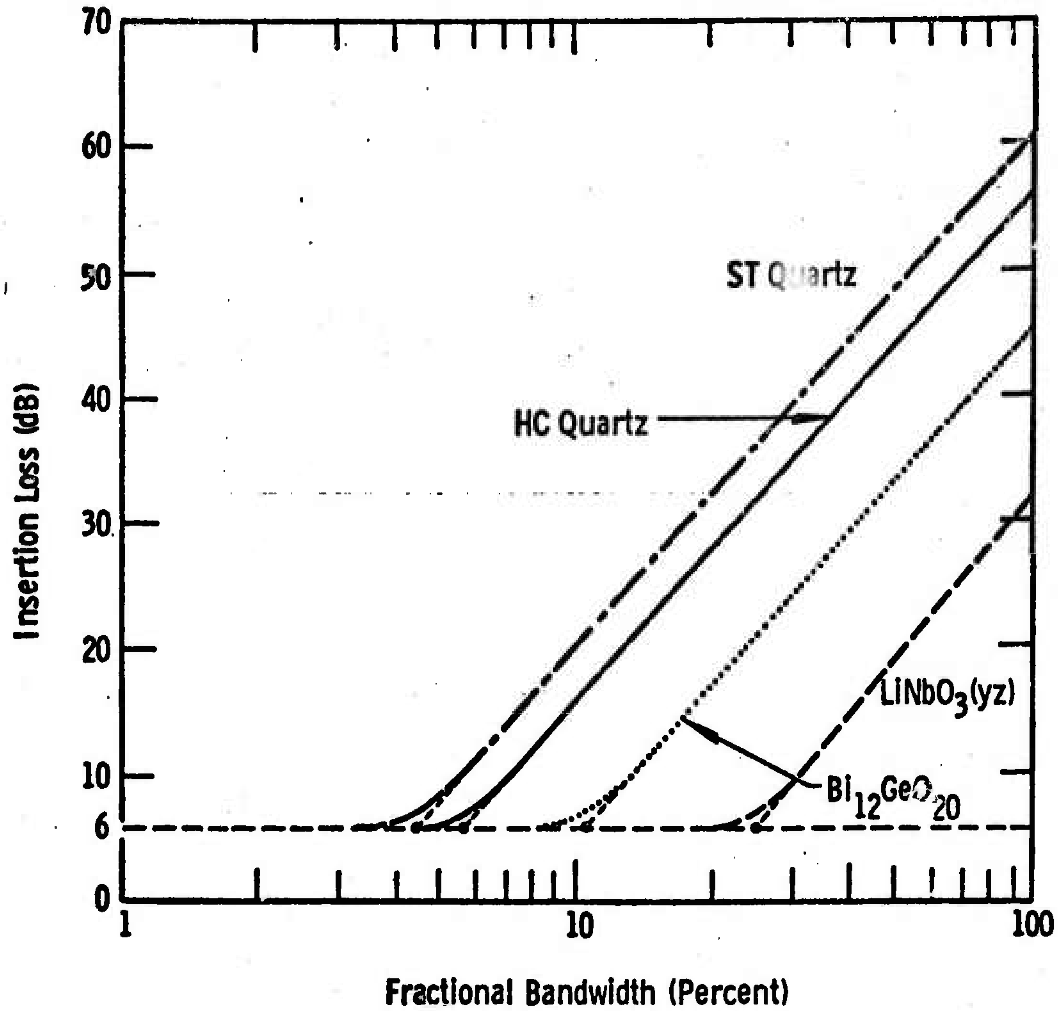


FIGURE 2.4.3 MINIMUM ACHIEVABLE INSERTION LOSS FOR TWO TRANSDUCERS ON VARIOUS SUBSTRATES

matched filters. An approximation to the insertion loss due to the transducers can be made based upon the fractional bandwidth and Q of the device. Hartmann et al [11] have developed graphs of insertion loss for two transducers on various substrates (see Figure 2.4.3). For bandwidths greater than about 5% on quartz the device Q is quite large. Thus, if the device is simply conjugately matched the overall response will be too narrow for the signal of interest. Hartmann's curves assume a matching resistance is placed across the device terminals to achieve the required circuit Q. These curves do not include parasitic effects which can be substantial. For broadband transducers, parasitic capacitance typically accounts for 3-7 dB of loss while the parasitic resistance and capacitance of the coded transducer contributes 3-5 dB. SAWD's used in packet radio have center frequencies of 67 MHz; the input transducer is required to have a fractional bandwidth of approximately 30%; the coded transducer fractional bandwidth requirement is about 20%. When the transducer losses for ST quartz obtained from Figure 2.4.3 are combined with parasitic losses, the losses can be greater than 50 dB. However, when the correlation or processing gain (21 dB for 128 chips/bit) is taken in consideration, the insertion loss is about 30 dB. This kind of insertion loss means that signals emerging from SAWD matched filters must be amplified to re-establish reasonable levels for further processing.

The center frequency or frequency of maximum response of a SAWD transducer is determined by the finger spacing

$$f_0 = v/d$$

where v is the acoustic wave propagation velocity and $d = \lambda_0$ is the acoustic wavelength at frequency f_0. For low frequencies, below 30 MHz, transducers become inconveniently large. Standard photolithography can produce finger widths down to about 1.5 μm or a f_0 of 500 MHz. For narrower fingers, electron beam techniques must be used to make the original mask although reproduction from the mask can be achieved using optical techniques. This method of fabrication and the propagation losses encountered at frequencies above 200 MHz restrict SAWD applications to below a few gigahertz [12].

The center frequency of the SAWD matched filter in packet radio establishes the IF frequency. Flexibility of the IF frequency could allow an optimum choice to be made based upon the many functional requirements of packet radio. From a fabrication standpoint, flexibility in the center frequency of a SAWD is easily obtained, at least between 30 and 500 MHz. From an applications standpoint, SAWD matched filters are used in packet radio for the differentially coherent detection of 10 μs MSK waveforms. A functional block diagram of differentially coherent MSK detection incorporating a surface acoustic wave device is presented in Figure 2.4.4. The SAWD matched filter is designed for a given center frequency of operation. It can be shown that any offset between the effective center frequency of the device and the IF frequency causes a degradation in the performance of the detection process. A given frequency offset corresponds to a certain phase error across the length of the device. Figure 2.4.5 shows how the signal-to-noise ratio E_b/N_o and the hybrid output channels ratio, Σ/Δ, degrades as a function of phase error and the amplitude imbalance, ϵ, that could exist between signals A and B at the hybrid input ports. The Σ/Δ is the quantity upon which the bit decision is made and theoretically is very large for an ideal device. If acceptable performance can be established at $\geq$ 14 dB Σ/Δ, for a 10 μs device, this corresponds to a maximum phase of error of 0.4 radians or a maximum frequency offset of 6.4 KHz.

Contributions to the total frequency offset arises from several sources. Local oscillator instability of both the transmitter and receiver can cause the IF frequency to deviate from f_o. A local oscillator at 1500 MHz with a stability of $\pm$1 ppm will cause a maximum deviation of $\pm$1500 Hz. This maximum increases to $\pm$3 KHz when both the transmit and receive LO instabilities are taken in consideration. As discussed earlier, the propagation velocity of the acoustic waves for a given finger spacing determines the center frequency. The propagation velocity is a function of temperature; therefore, the SAWD center frequency becomes temperature sensitive. ST quartz has a zero first order temperature coefficient of delay, but the second order coefficient is not zero. It is parabolic as shown in Figure 2.4.6 and for a temperature variation of $\pm$35°C, the frequency is subject to change -38 ppm. (Notice that the frequency

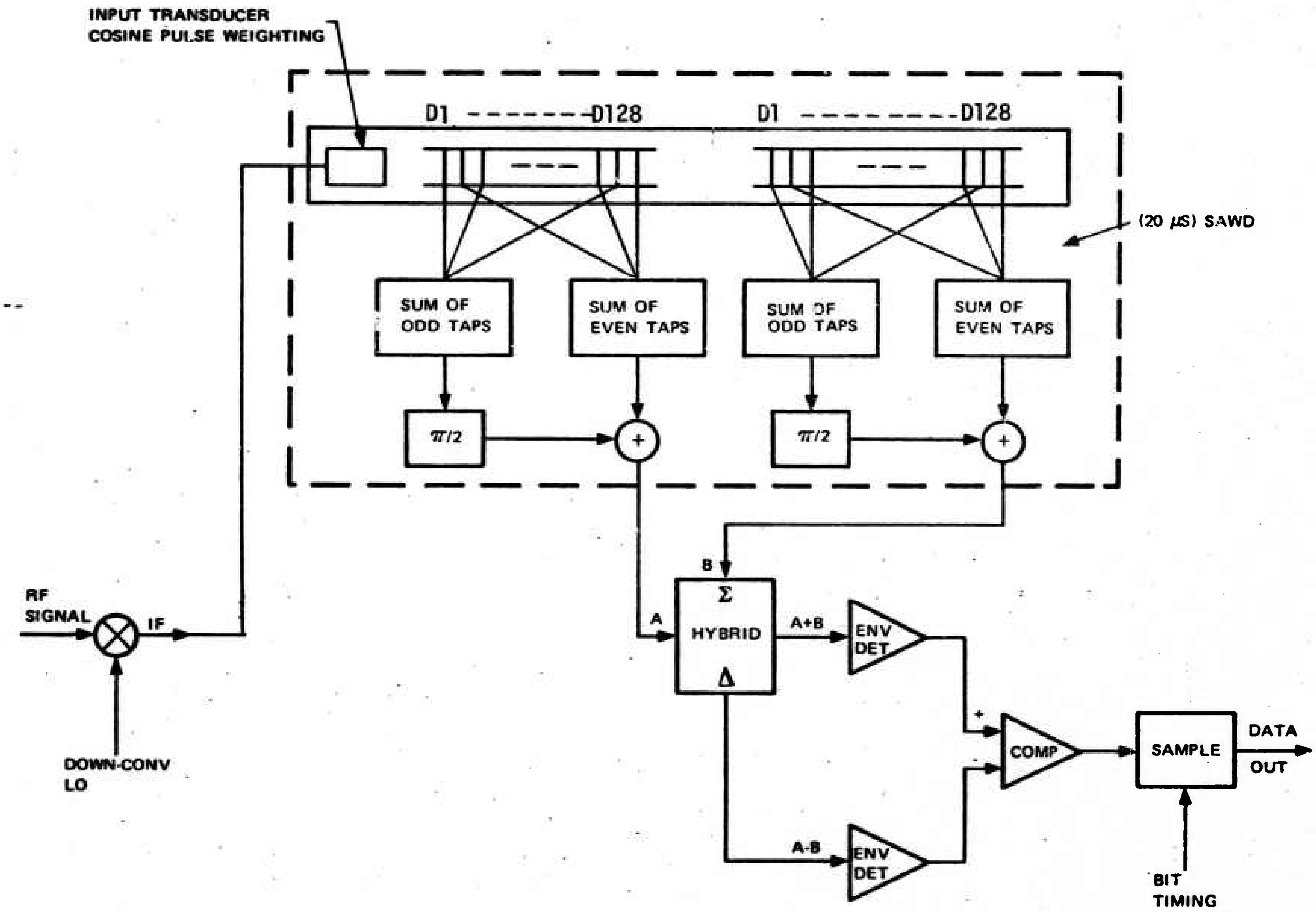


FIGURE 2.4.4 DIFFERENTIALLY COHERENT MSK DEMODULATION USING SURFACE ACOUSTIC WAVE DEVICE

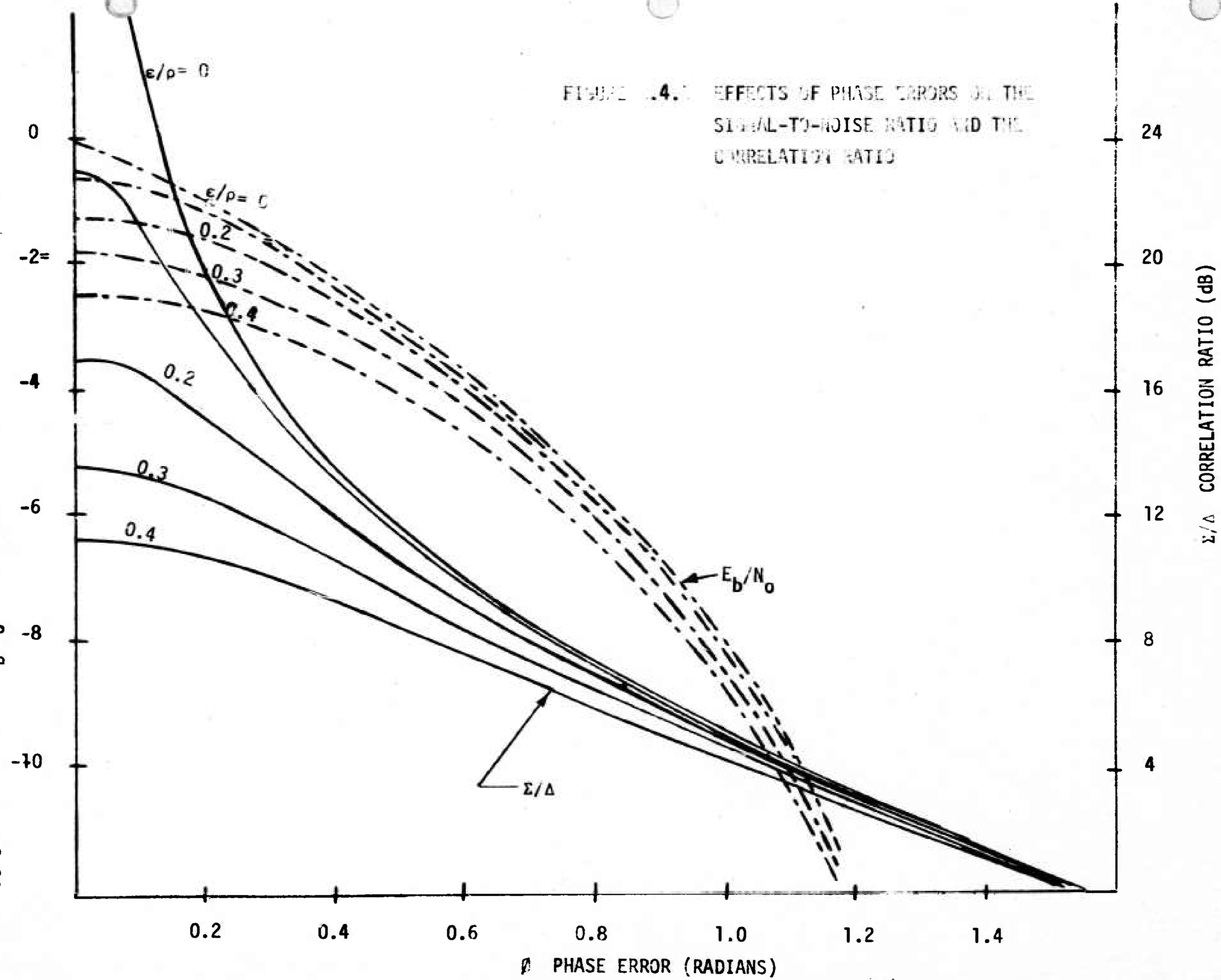


FIGURE .4. EFFECTS OF PHASE ERRORS ON THE SIGNAL-TO-NOISE RATIO AND THE CORRELATION RATIO

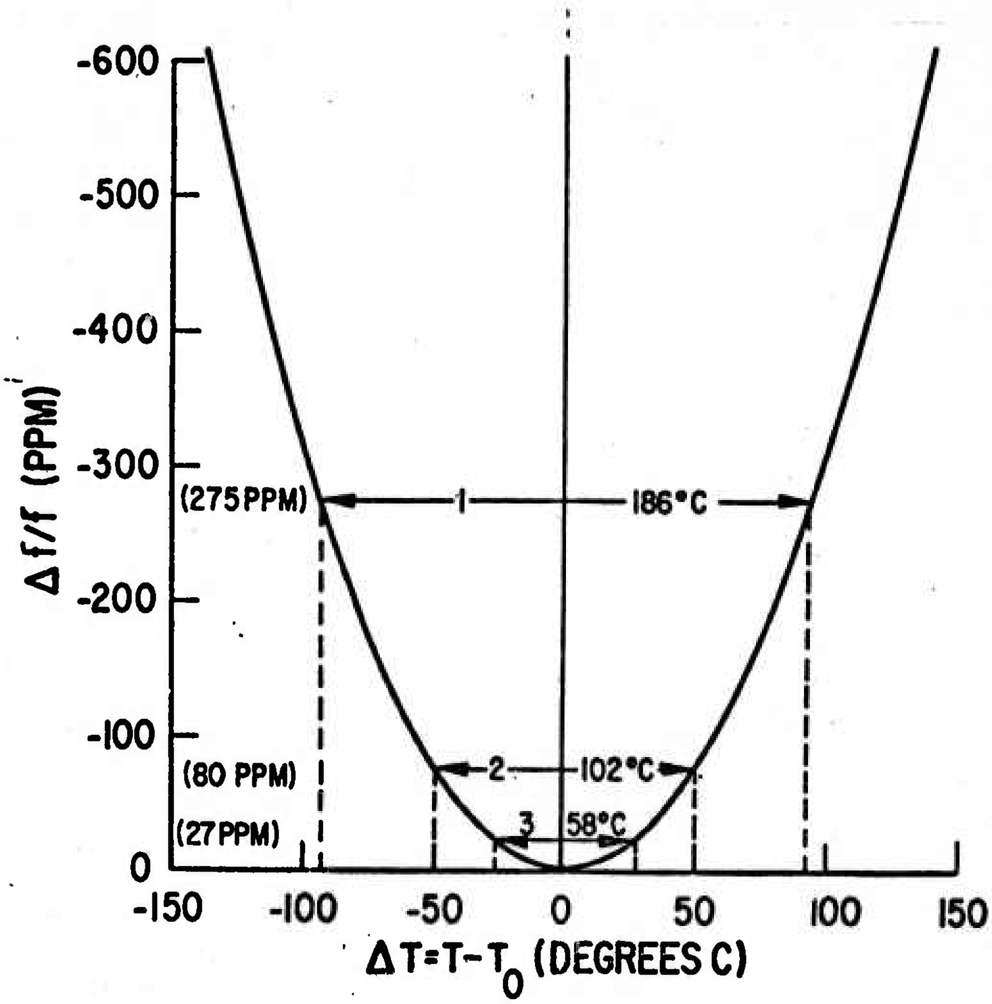


FIGURE 2.4.6 FREQUENCY VERSUS TEMPERATURE CHARACTERISTICS FOR ST QUARTZ

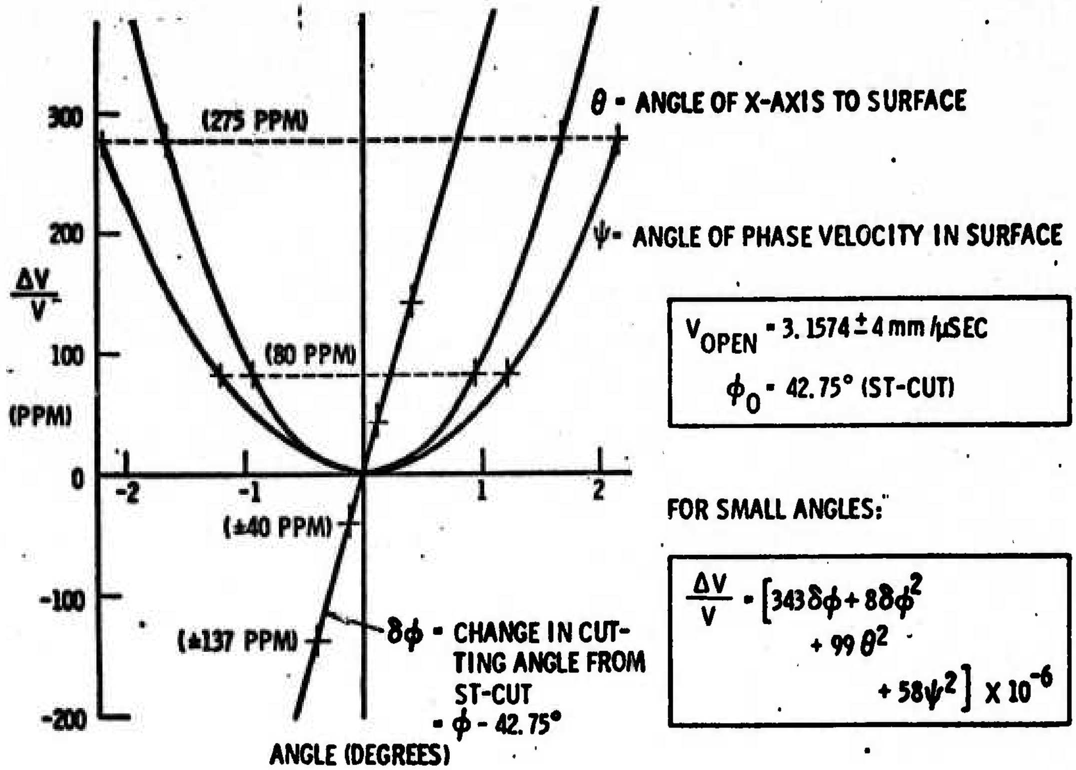


FIGURE 2.4.7(a) THE EFFECTS CRYSTAL ORIENTATION ERRORS ON ACOUSTIC WAVE VELOCITY

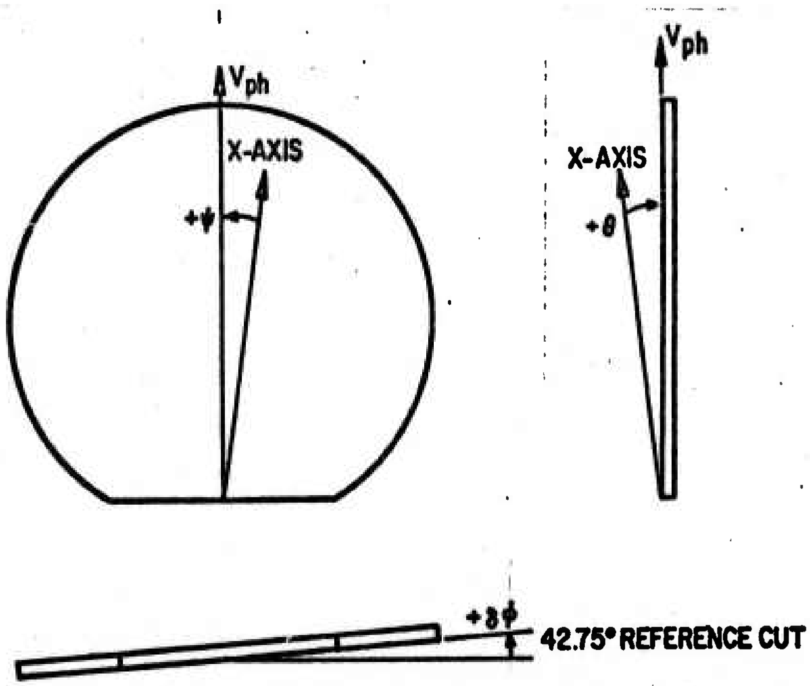


FIGURE 2.4.7(b) DEFINITION OF ANGLES FOR VELOCITY CALCULATIONS ON ST QUARTZ

change is always negative if the nominal operating temperature is chosen to be T_0 which is about 20°C for ST quartz). At 70 and 300 MHz, expected frequency offset could be as large as -2.66 KHz and -11.4 KHz respectively. A third major source of frequency offset error is crystal orientation errors. ST quartz crystals are cut to an accuracy of $\pm$15 minutes of the ST angle (ϕ = 42.75°). From Figure 2.4.7(a) the maximum frequency change can be determined if the crystal orientation errors are known. Since the contributions to frequency change are small for small values of θ and ψ, these angles are assumed zero. Therefore, for ϕ of $\pm$15 minutes, the frequency offset can be as much as $\pm$85 ppm. This is $\pm$6 KHz at 70 MHz and $\pm$25.5 KHz at 300 MHz. A summary of the worst case frequency error contributions are presented in Table 2.4.2. Recalling that ± 6.4 KHz maximum total frequency offset will maintain a Σ/Δ of dB, it is clear from Table 2.4.2 that designing for worst case situations is difficult if not impossible to do, since Σ/Δ degradation is severe at both center frequencies shown. However, it is also clear from Table 2.4.1 that using compensation techniques one can come much closer to achieving a maximum total frequency offset of ± 6.4 KHz more readily at 70 MHz than 300 MHz. Therefore, one concludes that there is not much flexibility in the frequency of operation of SAWD matched filters for packet radio applications unless measures are taken to reduce error contributions from the primary error sources listed in Table 2.4.2.

The propagation velocity of the acoustic wave determines for the most part the size of the SAWD matched filter. For differentially coherent detection of MSK in packet radio, a 20 μs waveform is required in the device. For a propagation velocity in quartz of about 7.5 us per inch, the SAWD is at least 2.5 inches long and is actually about 3 inches including input transducer. The actual crystal dimensions are 0.3" x 3" x .025". Since the propagation velocity is independent of frequency, the transducer represents the only change in crystal dimensions for a change in operating frequency. The size of the SAWD matched filter is further increased by the need for matching networks at its input and output ports. These networks in discrete form can consume considerable space in comparison to the crystal itself. The overall package dimensions of the 20 μs SAWD matched filter is 1.6" x 4" x 0.42".

The greatest impact on the size of the packet radio due to the SAWD matched filter is indirect through its center frequency limits (need for low IF), frequency stability (increased stability of the frequency sources) and insertion loss (need for additional amplification).

ERROR SOURCE	70 MHZ	300 MHZ
	Δf	Δf
LO (± 1 PPM)	± 3 KHZ	± 3 KHZ
TEMP. (± 35°C)	-2.66 KHZ	-11.4 KHZ
ORIENTATION (±15')	± 6 KHZ	± 25.5 KHZ

TABLE 2.4.2 WORST CASE FREQUENCY ERRORS FOR SAWD MATCHED FILTER AT 70 MHz AND 300 MHz

2.5 Energy Storage

Although much activity in recent years has been directed at the development of small high density power sources for mobile systems, this activity has not kept pace with miniaturization in electronics and packaging. Therefore, the power source in the packet radio represents one of the most challenging constraints to size reduction.

One of the most common and practical approaches to energy storage is the electrochemical system. The electrochemical system is one where the energy is stored in the form of chemical energy; i.e., active materials that take part in chemical reactions to provide an electric current. The characteristics of an electrochemical system (cell or battery) are largely determined by the properties of the active materials involved in the chemical reactions. As many combinatios of active materials are possible, the number of cells available with differing characteristics are many.

In 1834 Faraday formulated his fundamental law of the relation between the amount of electrochemical action and the quantity of electricity passing through a cell. The first part of Faraday's law states that the chemical effects produced at the electrodes by the passage of an electric current are in direct proportion to the magnitude of the current and the time that it flows. This law establishes the relationship between the output of battery and the amount of active materials that it contains. The second part of Faraday's law states that quantities of various substances, which are liberated or consumed by the action of a given quantity of electricity, are in all cases proportional to the equivalent weights. That is, the quantity of electricity (96,500 coulumbs or 28.80 ampere-hours) that liberates one gram--equivalent weight is the same for all substances. This quantity has been named the "faraday" and it is a fundamental constant of electrochemistry. [15]

One consequence of Faraday's law is that is allows the calculation of the theoretical maximum energy density of a given electrochemical system.

Since a gram-equivalent weight of a substance is determined from its atomic weight and valence, it follows that the lightest oxidizing and reducing agents available will result in the highest energy density systems. Table 2.5.1 [16] provides a comparison of the chemical properties of common types. Notice that lithium is the best reducing agent since it has the highest oxidation potentital. It is the lightest of all reducing agents and has the highest ampere-hour capacity per gram. Among the best light-weight oxidizing agents are oxygen and air, the halogens (flourine, chlorine, bromine, etcl), sulphur, and metallic compounds of these. [17]

Practical limitations in battery technology cause the achievable energy densities to be much less than the theoretical maximum that may be determined with Faraday's law. One reason is that the active material taking part in the reaction is seldom more than 25% of the total active material present [15]. Table 2.5.2 provides electrical characteristics of some non-reversible (primary) electrochemical cells; Table 2.5.3 presents the relevant electrical characteristics for reversible (secondary) cells. [18] The secondary cells tend to have lower energy densities on the average as compared to primary cells. However, primary cells have limited power density (power density is the rate of energy output). The silver-zinc secondary cell exhibits the highest energy density of the common secondary cells--50 watt-hours per pound. While 50 watt-hours per pound is hardly enough for many applications requiring small high density power sources, it does compare favorably with many of the most common secondary systems. For reasons presented earlier, lithium provides the highest energy density of all the cells discussed.

The power consumption requirements for a small packet radio could be 5 to 10 watts continuous average power. If it is desired that the duration of operation between battery change or recharge is 10 hours, the energy storage requirement is 50 to 100 watt-hours. From Table 2.5.3 the worst case situation would require 2-1/2 pounds silver-zinc battery occupying more than 40 cubic inches. The same worst-case condition can be met with a one-pound 12 cubic inch lithium battery. Clearly, the inability to provide very small lightweight energy sources represents one of the major bottlenecks to size reduction in packet radio.

All current electrochemical systems are far from realizing their maximum energy densities and improvements can be expected. New combinations of materials are being explored. However, those materials believed to have the ability to provide the highest energy density electrochemical systems are currently under investigation. Some lithium cells currently under laboratory development have demonstrated energy densities greater than 200 watt-hours per pound. [19] It is energy densities of this magnitude and greater where the energy source would cease to be a constraint to size reduction in packet radio.

TABLE 2.5.1 CHEMICAL PROPERTIES OF COMMON REDUCING AGENTS IN ELECTROCHEMICAL SYSTEMS

REDUCING AGENT	ATOMIC WEIGHT	OXIDATION POTENTIAL	AMPERE-HOUR PER GRAM
CADMIUM	112.40	+0.403	0.477
COPPER	63.54	-0.337	0.840
IRON	55.85	+0.440	1.440
LEAD	207.21	+0.126	0.517
LITHIUM	6.94	+3.045	3.862
MAGNETIUM	24.32	+2.370	2.204
NICKEL	58.71	+0.250	0.913
SILVER	107.87	-0.799	0.248
SODIUM	22.49	+2.714	1.166
ZINC	65.37	+0.763	0.820

TABLE 2.5.2 PRIMARY ELECTROCHEMICAL CELLS

	Leclanche	Zinc-Chloride	Alkaline	Magnesium	Mercury-Oxide	Silver-Oxide	Divalent Silver-Oxide	Lithium
1. Energy output Watt-hours per lb Watt-hours per in.^{3}	20 2	44 3	20 to 35 2 to 3.5	40 4	46 6	50 8	70 14	100 to 150 8 to 15
2. Nominal cell voltage	1.5	1.5	1.5	2.0	1.35 or 1.4	1.50	1.5	2.8
3. Practical drain rates High (50 mA) Low (50 mA) Pulse	100 mA/in.^{2} Yes Yes	150 mA/in.^{2} Yes Yes	200 mA/in.^{2} Yes Yes	200 to 300 mA/in.^{2} Yes No	No Yes Yes	No Yes Yes	No Yes Yes	No Yes Yes
4. Impedance Z_i	Low	Low	Very low	Low (delay on start up)	Low	Low	Low	Less than 1 ohm
5. Temperature range Storage Operating	-40 to 120°F 20 to 130°F	-40 to 160°F 0 to 160°F	-40 to 120°F -20 to 130°F	-40 to 160°F 0 to 130°F	-40 to 140°F 32 to 130°F	-40 to 140°F 32 to 130°F	-40 to 140°F 32 to 130°F	-65 to 160°F -40 to 130°F
6. Temperature vs capacity	Poor at low temperature	Good at low temperature compared to Leclanche	Fair to good at low temperature	Fair at low temperature	Good at high temperature Poor at low temperature	Poor at low temperature	Poor at low temperature	Excellent
7. Shelf life at 63°F to 80% initial capacity (in years)	2 to 3	2 to 3	3 to 5	2 to 3	2 to 3	2 to 3	2 to 3	3 to 5 (estimated)
8. Shape of discharge curve	Sloping	Sloping	Sloping	Fairly flat	Flat	Flat	Flat	Flat

TABLE 2.5.3 SECONDARY ELECTROCHEMICAL CELLS

	Gelled-electrolyte lead-acid	Nickel-Cadmium	Silver-Cadmium	Silver-Zinc
1. Energy output Watt-hours per lb Watt-hours per $in.^3$	 9 1.1	 12 to 16 1.2 to 1.5	 22 to 34 1.5 to 2.7	 40 to 50 2.5 to 3.2
2. Nominal cell voltage	2.12	1.2	1.1	1.5
3. Cycle life	200 to 500	500 to 2,000	150 to 300	80 to 100
4. Temperature range Store Operate	 -76 to 140°F -76 to 140°F	 -40 to 110°F -20 to 140°F	 -85 to 165°F -10 to 165°F	 -85 to 165°F -10 to 165°F
5. Shelf life at 68°F to 80% of capacity	8 months (with lead-calcium grids)	2 weeks to 1 month	3 months	3 months
6. Internal resistance	low	low	very low	very low
7. Discharge curve	sloping	flat	flat	flat
8. Relative cost, rated on a scale of 4 for maximum	1	2	4	3

2.6 Packaging and Interconnects

Much of the overall size and weight of the existing packet radio design can be attributed to its intended mission and experimental nature. Designing packet radio for an experimental system required building in flexibility and versatility. It is the ability to change parameters, modes of operation, thresholds, etc., and to observe the system's response that was paramount in the experimental system design.

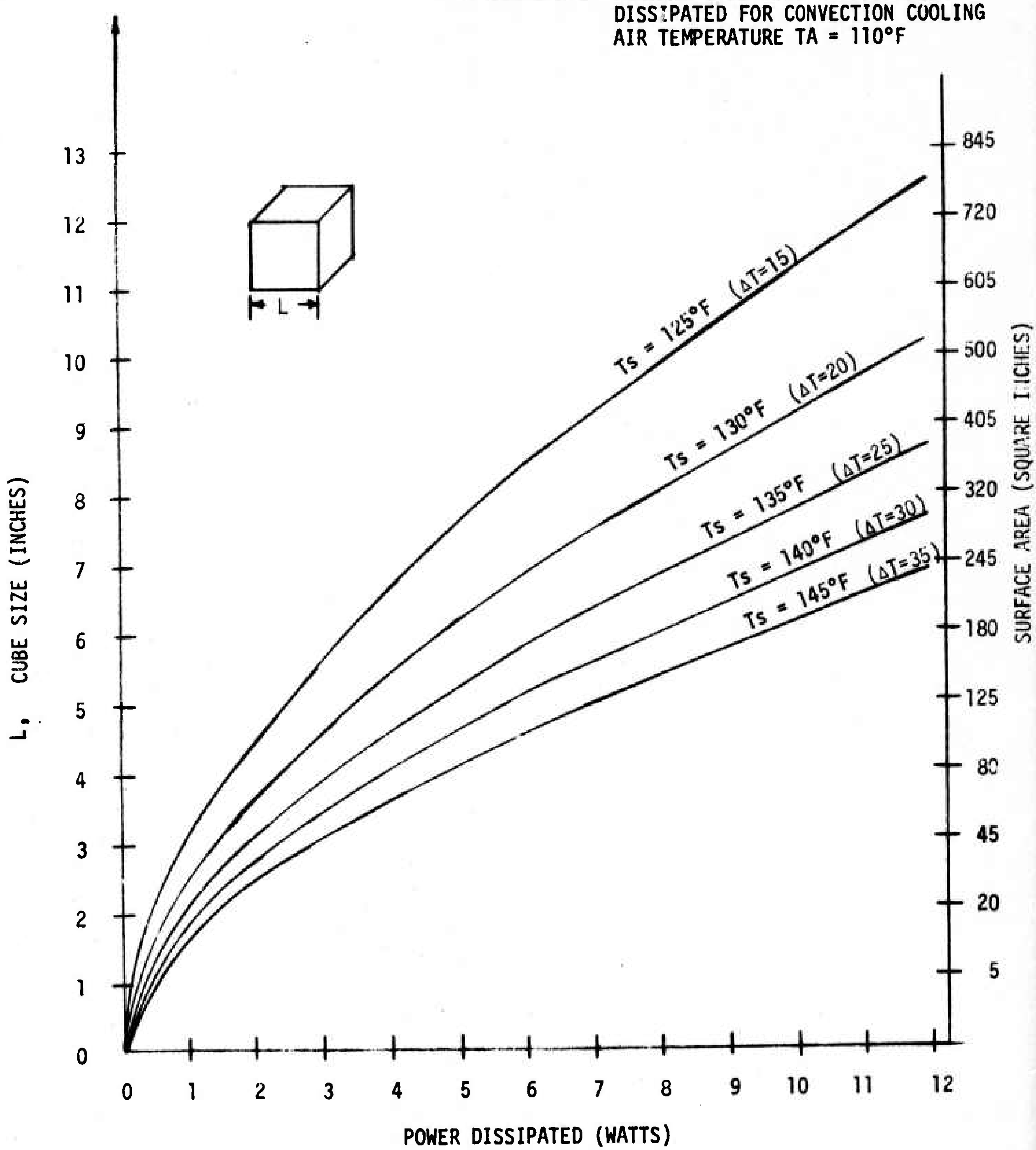
One design approach to assuring a high degree of flexibility is modular subassemblies. Each major function (frequency synthesis, bit synchronization, data detection, etc.) becomes a separate plug-in module. Desired changes in major function areas may be incorporated by simply replacing the redesigned module, resulting in minimal impact on the rest of packet radio. In a system using modular subassemblies, the module size is fixed or nearly fixed with one dimension allowed to change, usually in discrete steps. Since all circuit functions do not require the same amount of space, an optimum module size is chosen where some circuit functions are hard pressed to be contained while other functions may consume little more than half the available space. Therefore, modularity restricts the optimum allocation of volume.

The path and physical supporting structure by which a signal moves from one processing point to the next is an interconnect. Interconnects occupy considerable space in the packet radio. Moreover, the problem of space consuming interconnects is increased by the modular approach since a whole system of interconnects must be provided to allow the plug-in modules to talk to each other.

The thermal design of the packet radio package has considerable impact on the size of the system. A sufficient heatsink must be provided to ensure that component life or performance is not adversely affected. For a given set of conditions, the power dissipated and the allowable temperature rise, the size of the heatsink depends on the thermal impedance between source and sink. Therefore, minimizing thermal impedance for minimum heatsink size is a basic concern of the package design.

For a small packet radio, especially a hand-held unit, the surface temperature of the heatsink may allow acceptable performance for the radio unit but maybe unacceptable from a human factors standpoint. Figure 2.6.1 illustrates the cube size and surface area required to dissipate a given amount of power and keep the surface temperature within a given range of ambient (only convection cooling from five surfaces of the cube is considered in the calculations). For four watts of power dissipated in the cube, the surface area required, to maintain the surface temperature within 30 degrees of ambient, is about 85 square inches. Reducing the allowed differential by one-half increases the surface area by more than 2-1/2 times. This demonstrates one of the most fundamental constraints to realization of a reduced size packet radio. Transferring the heat out of a small volume and maintaining a ΔT that allows a man to handle the device is central to the discussion of constraints.

FIGURE 2.6.1 MINIMUM CUBE SIZE VERSUS POWER DISSIPATED FOR CONVECTION COOLING AIR TEMPERATURE TA = 110°F



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3.0 ALTERNATIVE SOLUTIONS TO SIZE REDUCTION IN A PACKET RADIO

Limitations and constraints to size reduction in the packet radio for several functional areas were identified in Section 2.0. The functions considered the most difficult to reduce in size are RF power amplification, frequency generation, filtering, energy storage, and packaging. The purpose here is to discuss solutions and alternatives in several of these functional areas. In general, these solutions and alternatives entail relaxing or eliminating certain capabilities of the packet radio, employing new technologies for more efficient circuit realizations, and exploiting the potentials of developing technologies.

When the capabilities of the existing packet radio unit are considered, that is, its ability to function as a station front-end, repeater, or terminal in an experimental packet radio network, clearly the available volume has been well utilized. Therefore, since it is not possible to maintain all the capabilities of the existing packet radio and because it is very desirable to have a small packet radio terminal, one avenue to size reduction is to define a set of desirable features for terminal operation only. While this allows the selection of a subset of the capabilities of the existing packet radio for the terminal unit, other features such as an on-board power source and the input-output capability become important. Although many of the means to size reduction are general and would be applicable to any packet radio element, regardless of its network function, the underlying assumption in the discussions that follow is a packet radio intended for terminal operation.

The solutions and alternatives are presented in the following six subsections. Subsection 3.1 describes the effect of reduced RF power out and mobile operation on repeater-to-terminal range. For three watts output power, the worst case range limit for 99 percentile detection probability is 0.4 miles. Subsection 3.2 proposes using lumped element components and chip-carrier circuitry to achieve reduced size in the power amplifier chain. Using these techniques a 35 dB gain 6 watt power out amplifier chain operating at 2.2 GHz occupying less than one square inch of circuit space has been reported in the literature.

Hybrid large scale integration appears to be the answer to smaller temperature compensated crystal oscillators described in Subsection 3.3. A developing technology, the surface acoustic wave oscillator, is also described and is seen to possess many attractive features for a small packet radio. How size reduction can be achieved by multiplexing certain functions between the transmit and receive modes is discussed in Subsection 3.4. Subsection 3.5 describes the implementation of relatively small filters by minimizing the number of resonant sections and unloaded Q required by proper circuit utilization. Both surface acoustic wave and charge-coupled device matched filters are discussed in Subsection 3.5. Charge-coupled device matched filters are shown to have considerable potential for the matched filtering application in packet radio.

3.1 Relaxation of System Requirements

The relaxation of some system requirements does not necessarily imply an inferior performance for the reduced-size packet radio. A fundamental requirement of the reduced-size packet radio is RF compatibility with existing packet radios. Although the concept of size reduction in a packet radio is generally the primary objective of relaxing or eliminating certain capabilities is a packet radio intended for terminal operations. For example, terminal operation of a packet radio would require the radio to be mobile for maximum utilization. This means the radio must perform in multipath environments which restrict data rates in urban areas to 200 Kbs or less. Therefore, a single data rate of 100 Kbs for terminal operation is not an elimination of a system requirement in the pursuit of size reduction, but a necessary constraint on terminal operation.

Other relaxations are intended primarily for size reduction such as reduced transmit power level, elimination of multichannel capability and power control. These suggested changes do not degrade the basic performance of the packet radio. Obviously, system operational flexibility is reduced--the less transmit power, the range is decreased and the more repeaters needed for a given area coverage. A reduced power output level is necessary since maintaining a ten watt power output capability in a small hand-held unit is impossible due to heat dissipation and power source limitations.

Reduced transmit power reduces the terminal-repeater link distance. In addition, the terminal-repeater link suffers because the terminal is assumed to be mobile and the received signal is fading at a rate proportional to the speed of the terminal. For equivalent bit error rate performance, the signal-to-noise ratio must be greater in the fading environment [1]. The terminal-repeater link has been estimated for reduced power out and mobile operation. The maximum allowable path loss has been calculated using the following relation [2]:

$$C/N(\mathrm{dB}) = \mathrm{ERP} - \text{path loss} + G_R - 10 \log KTB - NF$$

where ERP is the effective radiated power, G_R is the receiver antenna gain, B is the noise bandwidth, and NF is receiver noise figure. Using the following assumptions, a maximum allowable path loss of 138 dB was calculated:

1. Rayleigh fading environment (E_b/N_o = 20.4 dB)
2. 3 watt transmit power
3. 0 dBi transmit antenna gain
4. 9 dBi receive antenna gain
5. 20 MHz noise bandwidth
6. 8 dB noise figure

Figure 3.1.1 presents the expected path loss as a function of distance or the terminal-to-repeater link. The slope of the attenuation curves go at 40 dB/dec which is double the slope of the free space attenuation curve. The 99 percentile detection range was determined using Ott's statistics [3]. Ott's measurements indicate a standard deviation of about 8 dB for distances on the order of a few miles. From Figure 3.1.1, the worst case range limit for the 99 percentile detection probability is 0.4 mile. This compares to 1.4 mile for repeater-to-repeater links with non-fading signals. In order to double the worst case range limit, the maximum allowable path loss must increase 12 dB. It is possible that the range limit of 0.4 mile could be increased through a combination of directive antennas, and lower receiver noise figures.

To summarize, considerable size reduction can be obtained by eliminating some system requirements such as channel changeability, dual data rates, power control, etc., and the reduction of others such as transmitter power. Of the system requirements considered for changing for a terminal packet radio, the reduced transmitter power impacts the packet radio system the most. Clearly, the range reduction is the biggest price paid from a packet radio system standpoint for a small packet radio.

3.2 RF Power Amplification

Two basic problems existing in reducing the size of the transmit chain amplifiers. The first involves reduction of the size of the electronic circuitry and the second involves optimization of the thermal design to dissipate heat in as small an area as possible. The overall design must incorporate techniques to minimize size both electrically and thermally. Several different circuit techniques can be considered for size reduction of the electronic circuitry in the transmit amplifier chain. Each approach has advantages as well as limitations:

1) Microwave Integrated Circuits

Microwave integrated circuits (MIC) require elements that are fractions of a wavelength due to the distributed nature of such circuits. Conventional dielectrics such as alumina cause wavelengths of inches at 2 GHz and thus there is a basic lower bound to the size of MIC circuit elements. The size reduction obtainable by using higher dielectric materials varies only as $1/\sqrt{\varepsilon}$ where ε is the dielectric constant. GaAs yields an improvement of about 13% while ferrite yields 21% reduction in line lengths. In addition, the loss tangent for alumina is about 1×10^{-4} whereas the loss tangent for GaAs is 16×10^{-4}, thus GaAs is more lossy than alumina and this is undesirable for power amplifier circuits at microwave frequencies.

2) Thin Film Amplifiers and Filtering

Thin film amplifiers show considerable potential for size reduction since gains of 15 dB and octave bandwidths can be obtained on substrates measuring .25 x .25 inches. This small size is in part achieved by utilizing lumped elements, particularly inductors, and MOS capacitors.

3) Lumped Element Circuits with Chip Components and Chip Carriers

Lumped element circuit techniques utilizing chip transistors and designs tailored to the gain, power and bandwidth requirements

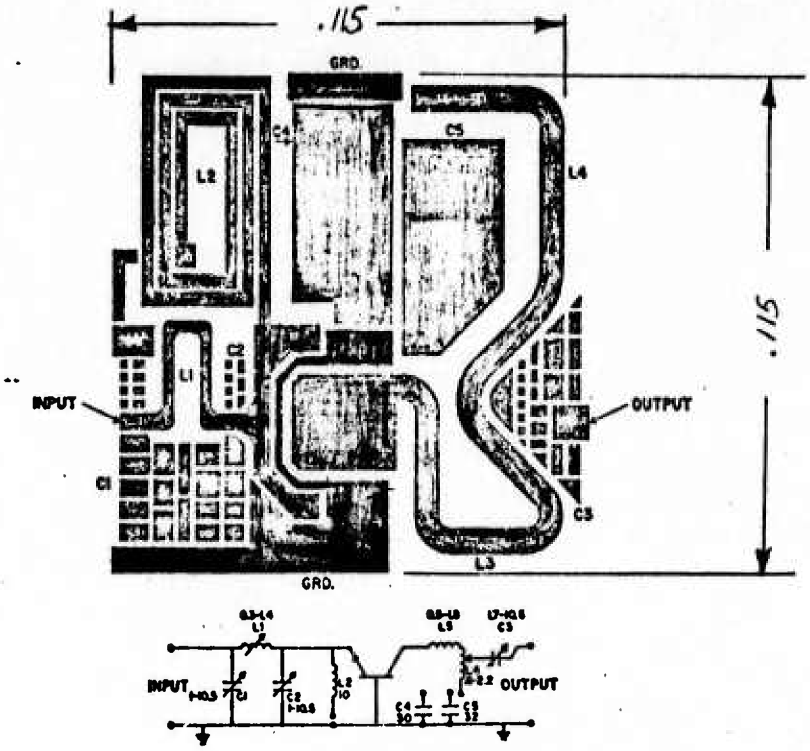


FIGURE 3.2.1 CIRCUIT DIAGRAM AND METALLIZATION PATTERN FOR SINGLE-STAGE LUMPED ELEMENT 2.25 GHZ AMPLIFIER

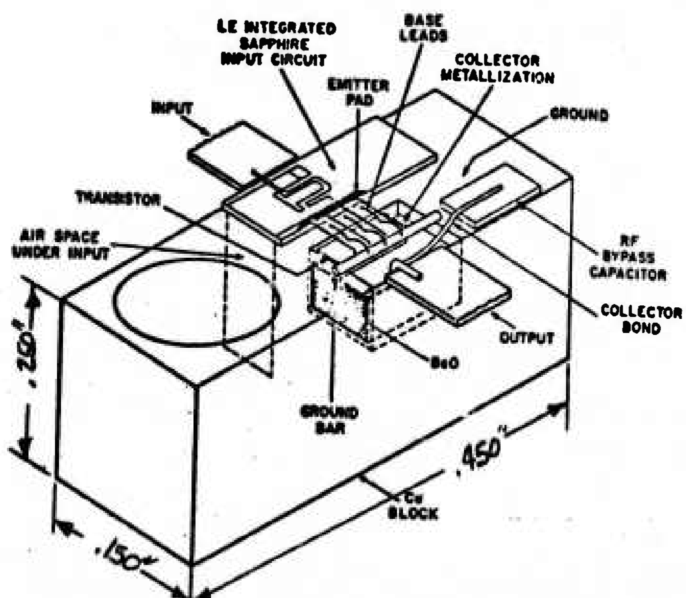


FIGURE 3.2.2 CONSTRUCTION DIAGRAM OF 7W 2.25 GHZ TRANSISTOR AMPLIFIER

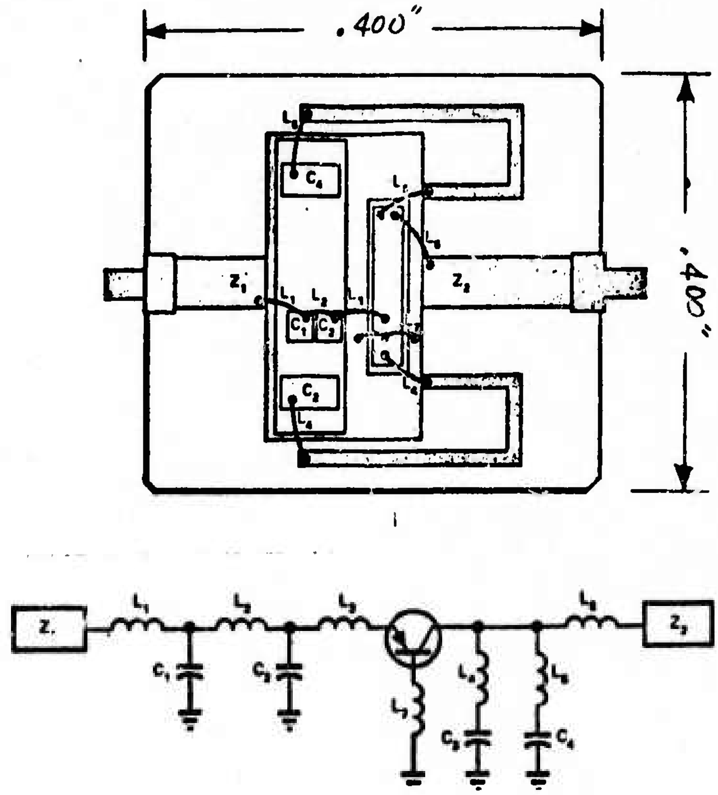


FIGURE 3.2.3 SCHEMATIC RELATIONSHIP OF THE AMPAC 1114-15 SHOWING INTERNAL MATCHING

appear to offer the greatest promise for reduced size. This is because lumped elements are much smaller than MIC elements at L and S bands since they are much smaller than a wavelength on the dielectric. Lumped elements are formed by a three-layer sandwich of metal-dielectric-metal on a substrate such as sapphire. Capacitors can be formed either by using the center dielectric or by interdigital techniques; inductors can be formed by deposition of the metal directly on the substrate. Inductors of several tens of nanohenries and capacitors of several tens of picofarads are obtainable up to several gigahertz with Q's of 100. Examples of such elements are shown in the amplifier of Figure 3.2.1. Typical values are 5nH coils with a diameter of 32 mils and 50pF capacitors 45 mil square.

Lumped element circuit techniques can also be used in high power circuits as shown in Figure 3.2.2. Here the chip is mounted on a BeO block which is in turn imbedded into a copper block; this provides excellent thermal conductivity. The input circuit is formed by a lumped element π section while 5 mil bondwires shunt the collector to ground through an SiO_2 capacitor. This technique provides for minimization of parasitics and since the base is attached directly to the copper block by bondwires, an excellent ground is achieved.

4) Monolithic Integration

In the monolithic circuit, active and passive devices performing the amplifier gain and matching functions are grown in a semiconducting substrate. Candidate substrate materials at microwave frequencies are silicon or gallium arsenide. Since tuned lines for impedance matching on either silicon or GaAs at 1800 MHz are substantially long, a lumped element approach is preferred. However, monolithic integration, in addition to utilization of lumped elements, does not appear to offer any additional reduction in size since the lumped elements themselves have the most significant impact.

Power Amplifier

Assuming operation at a reduced transmit power level (3 watts to the antenna), the output power of the power amplifier must be 4.5 to 5 watts to overcome the anticipated losses between it and the antenna. At an operating frequency of approximately 1700 MHz with a bandwidth of 30 to 40 MHz, an efficiency of 50% or greater and a gain of 8 dB or better is desirable. Also, the device (circuit) should be able to withstand a very high VSWR (greater than 10:1) since loss or damage to the antenna could be a higher probability event in a hand-held packet radio. Microwave Semiconductor Corporation (MSC) has a 6 watt, internally matched device that covers 1.7 to 2.0 GHz with a 9 dB gain and efficiency of 50% (see Figure 3.2.3). This device has been tested at 15 to 18 volts and will meet the above specifications at 18V with a gain degradation to about 8 dB. These devices are optimized for 24 volt operation, and for 24 volts devices the saturated output power point is 7 watts; therefore, a reduced supply voltage (15 to 18 volts) should still permit 5 watts of output power.

Since a reasonable gain for the output power stage is 8 dB, approximately 37 dB of gain will be required preceding this stage. Caulton [4, 5] and others have reported multistage lumped element amplifiers with approximately such gains at S-band. Caulton developed an amplifier to accept a 1 watt 15 mil x 15 mil transistor chip. Tuning for different impedance levels was accomplished by paralleling capacitors and paralleling or tapping strip inductors with bondwire as shown in Figure 3.2.2. When operated as a Class-C amplifier the circuit provided 10 dB gain with greater than 1-W output and greater than 30 percent efficiency at 2.25 GHz. This circuit measured 115 mils x 115 mils. Three of these stages were operated in cascade with the first two stages biased for Class-A operation and the third for Class-C operation. These stages were adjusted to give 1W of power with 30 dB gain and 20 percent efficiency with a 5 percent bandwidth. Caulton then used this three stage amplifier with a power stage built on a special holder (see Figure 3.2.2) to achieve 35 dB gain, 6 watts of power out, a 26% efficiency over a 6% bandwidth at 2.2 GHz. The overall amplifier package measured 7/8" x 1".

Heatsink

As described above the circuitry for the transmit amplifiers can be reduced substantially. Based on the results described, RF power amplifier chain could be built in an area approximately 1 inch square with the exact layout depending on module constraints. However, the overall size of the amplifier is determined by the heatsinking required for the transistors since the heatsink must extend below the plane of the substrates.

An estimate of the thermal rise ΔT_J of each stage can be made by using representative values of the thermal resistance, θ_{JC}, for each device and the average power dissipated.

	θ_{JC}	P_D	ΔT_J (°C)
Stage 1	200°C/W	4.5mw	0.9
Stage 2	200°C/W	45mw	9.0
Stage 3	100°C/W	0.450w	45
Driver 4	50°C/W	0.395w	19.8
Power Amp	11°C/W	2.5w	28

Note that these are ΔT_J values based on average power dissipated at 50% duty cycle. The peak ΔT_J values will be twice this.

The objective in heatsinking the devices is to maintain the junction below some temperature value which will allow reasonable operating life; this value is usually chosen to be about 150°C. Calculations have shown that 14 square inches of surface area will provide adequate cooling by convection to maintain all junctions below 150°C based on an operating range of 0 to 70°C. Therefore, the heatsink and not the actual transmitter circuitry is the constraining influence to size reduction of the transmitter chain.

3.3 Frequency Generation

As pointed out in Subsection 2.3, the direct (multiplication) or indirect (phase lock loop) method may be used to implement the local oscillator system in the packet radio. No matter which method is used a temperature-compensated crystal oscillator is required to achieve a ± 1 ppm stability of the LO system. The direct method which multiplies the crystal oscillator frequency to arrive at the LO frequency consumes less power, but has the problem of eliminating harmonics of the multiplier stages. The indirect method which uses a phase-locked loop easily lends itself to programmability but suffers from large size and high power consumption. A third approach to the LO system is the surface acoustic wave oscillator which has several attractive features including small size and low power consumption but currently exhibits poor long term stability.

Presently, there does not appear to be a straightforward solution to implementing the LO system in small size with low power consumption. As indicated, either approach has its advantages and disadvantages. The following discussion describes the advantages and disadvantages of the several approaches possible for the implementation of an LO system for a small packet radio.

Crystal Oscillator

A temperature-compensated crystal oscillator (TCXO) consists of a crystal oscillator circuit, and a temperature compensation network. Conceptually, temperature compensation is achieved by pulling the oscillator frequency to compensate for the nonlinear frequency-temperature changes of the crystal and the oscillator elements over temperature. In general, the optimum choice for the crystal resonator is the AT cut of quartz whose temperature characteristics are shown in Figures 2.3.2 and 2.3.3.

One candidate approach [18] to temperature compensation of the crystal oscillator uses a dual varactor scheme and one or two "3 point" network segments employing a thermistor in each leg as shown in Figures 3.3.2 and 3.3.2. Figure 3.3.2 presents the complete circuit schematic of the TCXO.

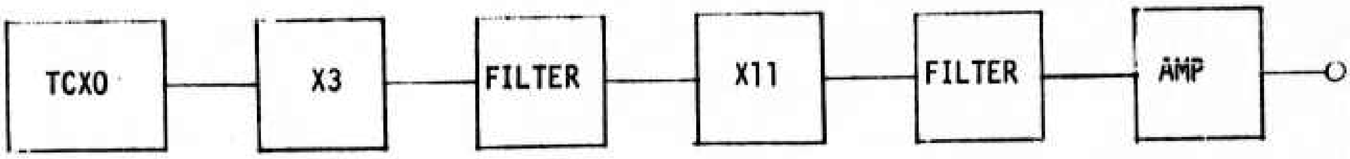


FIGURE 3.3.4 DIRECT MULTIPLICATION APPROACH TO FREQUENCY SYNTHESIS

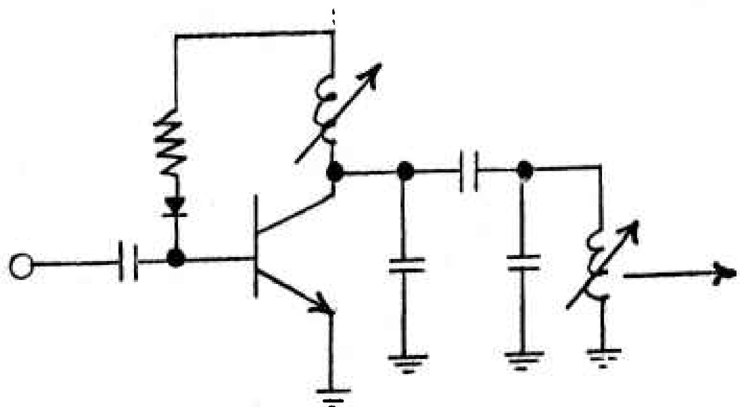


FIGURE 3.3.5 LOW ORDER ACTIVE MULTIPLIER STAGE

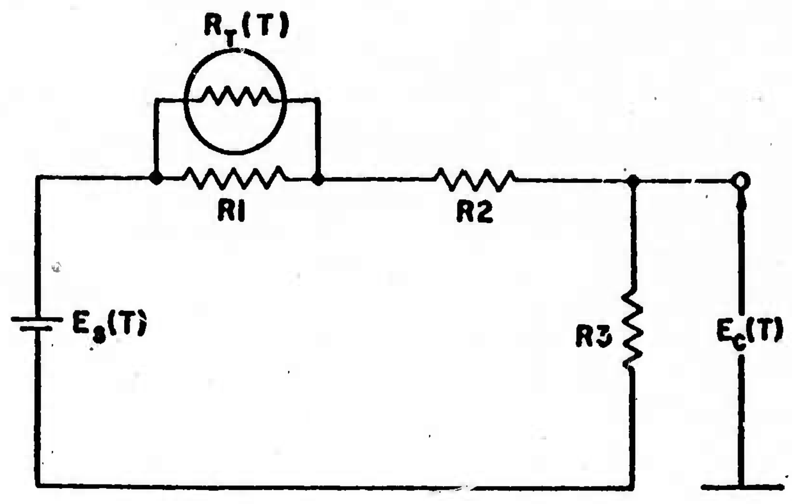


FIGURE 3.3.1 THREE POINT THERMISTOR - RESISTOR NETWORK

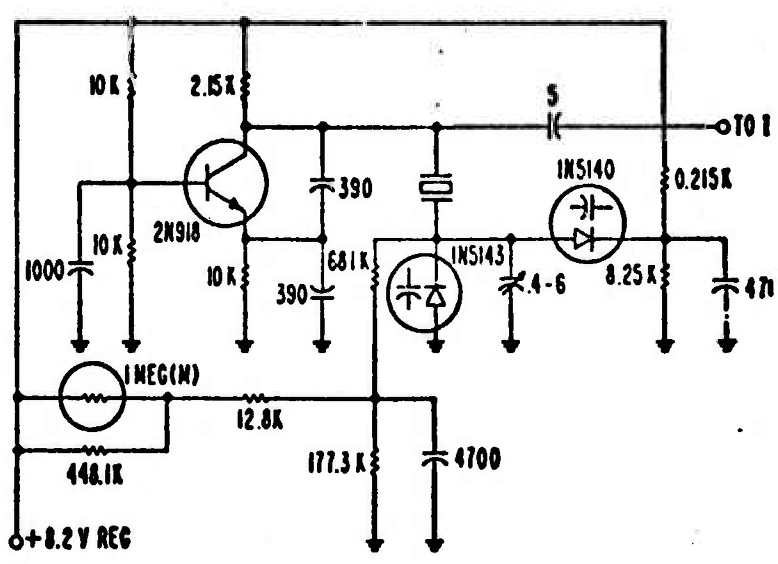


FIGURE 3.3.2 CIRCUIT SCHEMATIC OF A HIGH STABILITY TCXO

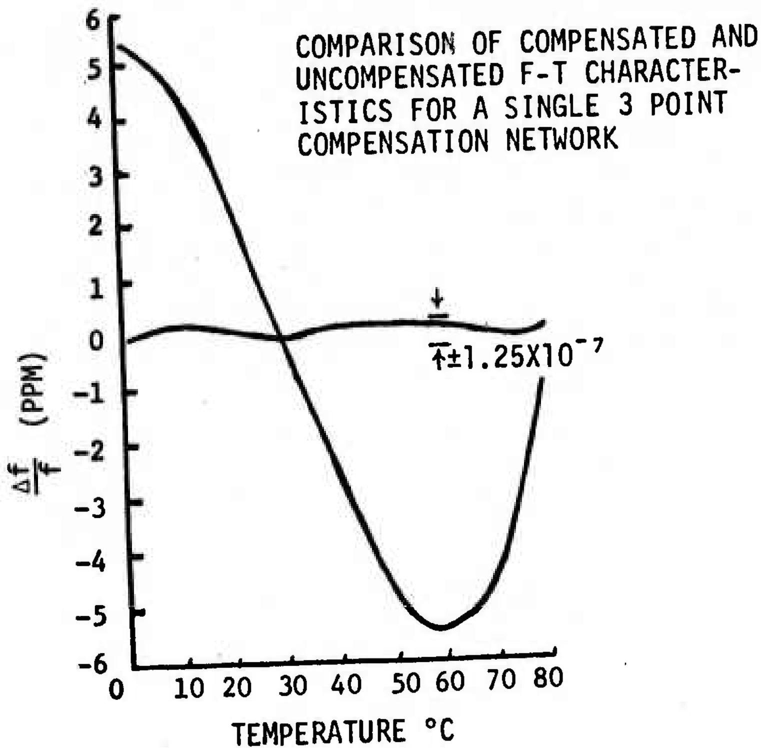


FIGURE 3.3.3 TEMPERATURE - COMPENSATED CRYSTAL OSCILLATOR STABILITY

The circuit, excluding crystal, is envisioned on a .5 x .5 inch thick film ceramic substrate. All components would be of chip form, excluding the resistors and thermistor. Also included in the package will be a monolithic chip voltage regulator (such as the Motorola MC 7812) as well as an additional chip amplifier stage for buffering.

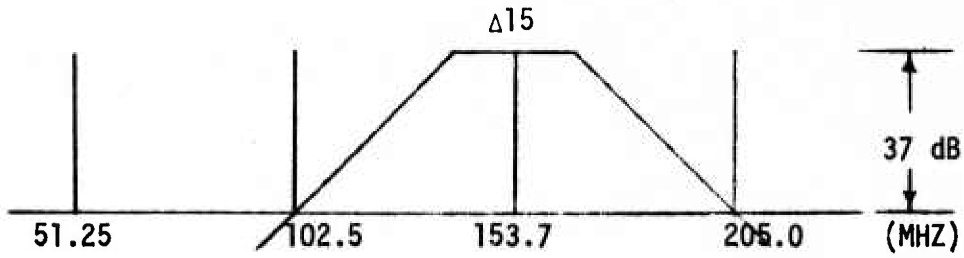
Figure 3.3.3 illustrates the performance of a TCXO using this approach with conventional components. A stability of ± .13 ppm can be achieved over a temperature range of 0 to 80°C.

Direct Multiplication Frequency Synthesis

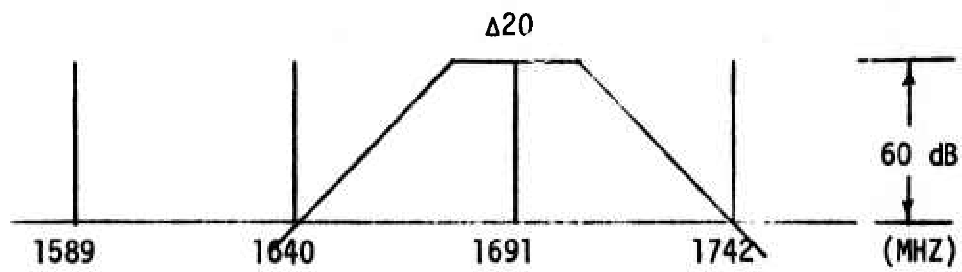
Direct multiplication may be a viable approach to a single frequency local oscillator system. This approach is shown in Figure 3.3.4. In this configuration a TCXO having an output frequency in the 51-52 MHz range, establishes the frequency standard for the system. The TCXO output is multiplied sequentially in two multiplier stages for a total multiplication factor of 33. Spurious frequencies are rejected via interstage filter networks. Lower order frequency multiplication in the first stage of the chain may be accomplished with an active multiplier to provide conversion gain. A typical X3 stage is illustrated in Figure 3.3.5.

The bandpass filter used for spurious rejection in the first multiplier is realized with lumped element chips. Coil manufacturers, such as the Cambion 556-1000 Series have microminiature inductors available as fixed and variable monolithic chips. The trade-off associated with inductor shrinkage is loss of unloaded Q. Manufacturer's data indicate that a Q minimum of 30 is achievable for a .100 x .160 variable monolithic chip inductor. Assuming a loaded Q of 10, the insertion loss is calculated to be about 4.5 dB. This power loss must be balanced by the conversion gain of the active stage.

The skirt selectivities of the interstage filters are shown geometrically in Figure 3.3.6. The first stage filter is a 2 pole, 10% bandwidth chebyshev filter followed by a 4 pole YIG filter.



FIRST STAGE FILTER



YIG FILTER

FIGURE 3.3.6 FILTER CHARACTERISTICS FOR DIRECT MULTIPLICATION FREQUENCY SYNTHESIS

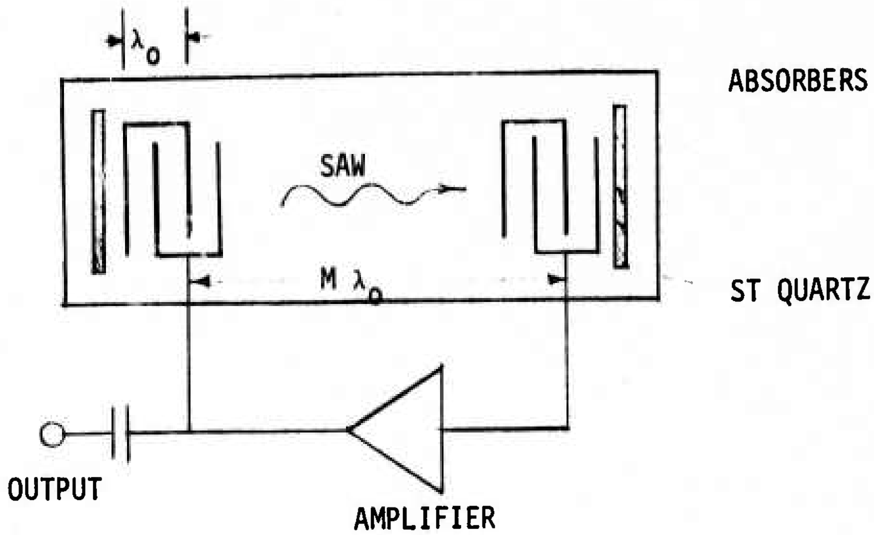


FIGURE 3.3.7 BASIC SURFACE ACOUSTIC WAVE OSCILLATOR STRUCTURE

The X11 stage may use a step recovery diode for best efficiency. Using this approach, the incoming signals from the X3 multiplier are converted to impulses via the snap action of the diode. A tank circuit resonant at the eleventh harmonic loads the diode to achieve the desired response. Due to the power losses incurred in the X11 multiplier (about 7 dB) and the insertion loss of the YIG filter (3 dB), a lumped element tuned amplifier stage is used to provide 10 dB gain at the LO frequency.

In summary, the major drawback to the direct multiplication approach for single channel frequency synthesis is the extra filters and amplifier necessary to achieve the desired spurious rejection levels. However it is believed that proper filtering, bypassing and shielding techniques may be implemented effectively in small package dimensions using miniature chip components. The major advantage is the low power consumption level. The local oscillator system outlined in Figure 3.3.4 would be expected to consume less than 0.5 watts.

Direct Frequency Synthesis Using Surface Acoustic Wave Device

The surface acoustic wave (SAW) oscillator is a device of simple structure that may prove to have application in packet radio. The basic oscillator structure is shown in Figure 3.3.7. The oscillator consists of a SAW delay line in the feedback loop of an amplifier. If the gain of the amplifier exceeds the delay line insertion loss, oscillation will be sustained according to [1]

$$\frac{wL}{v} + \phi_E = 2N\pi$$

where $L = M\lambda_0$ is the SAW path length, v is the acoustic wave velocity, N is an integer and ϕ_E is the electrical phase shift associated with the amplifiers and transducers, and w is the frequency of oscillation. From the set of frequencies satisfying the above relation, the transducer finger pitch, λ_0 determines the frequency of operation f_0.

It is here that the value of the SAW oscillator lies; the frequency of oscillation is not related to the basic crystal dimensions. Fundamental frequency of operation over the range 20 MHz to 1 GHz is readily accomplished using conventional photolithography, while an extension to 2 or 3 GHz can be made with electron beam techniques. In addition, SAW oscillators can be made small and rugged, such that at frequencies above approximately 300 MHz the entire oscillator can be housed in a single T08 can.

The short term stability of present-day SAW oscillators approaches that of bulk crystal oscillators. The effective Q of the SAW oscillator is determined by the delay line length. The values achievable lie in the range of 100 to 10,000. Although these values are lower than those obtainable with bulk crystal oscillators, the SAW oscillator has a much higher power handling capability thus giving comparable short term stability and fm noise [13]. The effective Q of the surface acoustic wave oscillator may be improved by replacing the delay line with a SAW resonator. The SAW resonator consists of one or more transducers located between two surface acoustic wave reflectors. The reflectors are simply metal strips etched on the surface of the quartz. Q's of 10,000 are reported for the SAW resonator with Q's of 100,000 believed to be within the capability of the technology. [14]

The medium term stability is determined by the temperature characteristics of ST quartz. From Figure 2.3.2, the SAW oscillator would be expected to exhibit a stability of ± 25 ppm across 0 to 70°C temperature range. However, due to the smallness of the SAW delay line and the good thermal contact possible, very simple temperature control can be used to improve stability. If, for example, the temperature of the SAW oscillator is maintained within ± 4°C of the temperature inversion point ($\sim$20°C for ST quartz but can be made any suitable value), the frequency variation is less than 1 ppm.

It is the long term stability of SAW oscillators which limits the application of these devices. Aging rates of about 3 parts per million per month can be expected using normal fabrication procedures. This compares with a few tenths of a part per million per month for bulk crystals.

Indirect Frequency Synthesis

The indirect approach to frequency synthesis employs an oscillator that is phase locked to a frequency standare, thereby assuming the stability of the frequency standard. The block diagram of Figure 2.3.4 presents the basic components of the indirect frequency synthesizer. Programmability is achieved by the ability to change the value of N, the divide ratio, to choose a given frequency within the range of the voltage controlled oscillator. For non-programming applications, N is fixed.

The means to size reduction and lower power consumption in the phase-locked loop appear to be few. One technique is to use a single loop filter/amplifier in which size reduction can be achieved using conventional chip carrier bonding techniques with deposited resisters and chip capacitors. A second technique that may reduce the power of the countdown circuitry for a non-programmable LO system is the use of binary (powers of 2) counters and a combination of fast and slow logic. That is, the front-end divider stage would be very fast logic, but would deliver an output that could be handled by slower and less power consuming dividers. A third consideration proposes using a passive X2 multiplier at the phase-locked loop output to obtain the 1600 MHz LO frequency. The passive X2 multiplier, while incurring some conversion loss, does not consume any power.

Summary

For the near term, the direct method to frequency synthesis appears the best route to implementation of an LO system for a small packet radio. Although difficult, spurious responses produced during the multiplication process can be adequately attenuated, and the low power consumption makes the direct approach the optimum choice.

Future developments relating to the long term stability of the SAW oscillator will be of interest. When the long term stability of the SAW oscillator is improved to a few tenths of a port per millim per month, the SAW oscillator will be the answer to the frequency generation problem in a small packet radio.

3.4 Multiplexed Functions

Since the packet radio operates half-duplex (does not simultaneously transmit and receive), functions common to both the transmitter and receiver were reviewed to determine whether time sharing was possible. A scheme which multiplexes the mixer and also allows the filtering requirements to be met is shown in Figure 3.4.1. Controlled switches are used to switch the signal paths through the mixer. The bandpass filter could be a narrowband four-pole filter to achieve the desired image rejection on receive. When transmitting the bandpass filter's higher insertion loss can be tolerated since the power levels at the mixer output are relatively low. The bandpass filter may not provide enough attenuation to the LO; therefore, LO-RF isolation in the mixer would be needed. The double balanced mixer can be designed to have good LO-RF isolation (25 dB minimum at packet radio frequencies) and can be implemented in small size. Construction of a double balanced mixer can be accomplished using two torid transformers and a diode quad. Diode quads of 50 mil square are available and high frequency torid transformers operating up to 3 GHz can be used to build a mixer in a space of 0.25" X 0.5". The necessary LO power can be reduced by employing low potential barrier mixer diodes such as the recently developed zero-bias Schottky barrier diode.

Using a single mixer for both the down conversion and up conversion, the mixer noise figure is likely to be worse since optimizing the up conversion loss would be desirable. The degraded mixer noise figure would not necessarily result in a degraded receiver noise figure since the RF amplifier gain may be increased because of the narrow operating bandwidth.

Estimates of the percentage savings in volume and power were derived for the scheme of Figure 3.4.1 by assuming that the techniques to size reduction described in this document were applied to a dual conversion approach without circuit multiplexing. The volume and power requirements for the dual conversion circuitry were compared to the volume and power requirements for the above described approach. Volume savings were estimated at 60% and power savings at 50%.

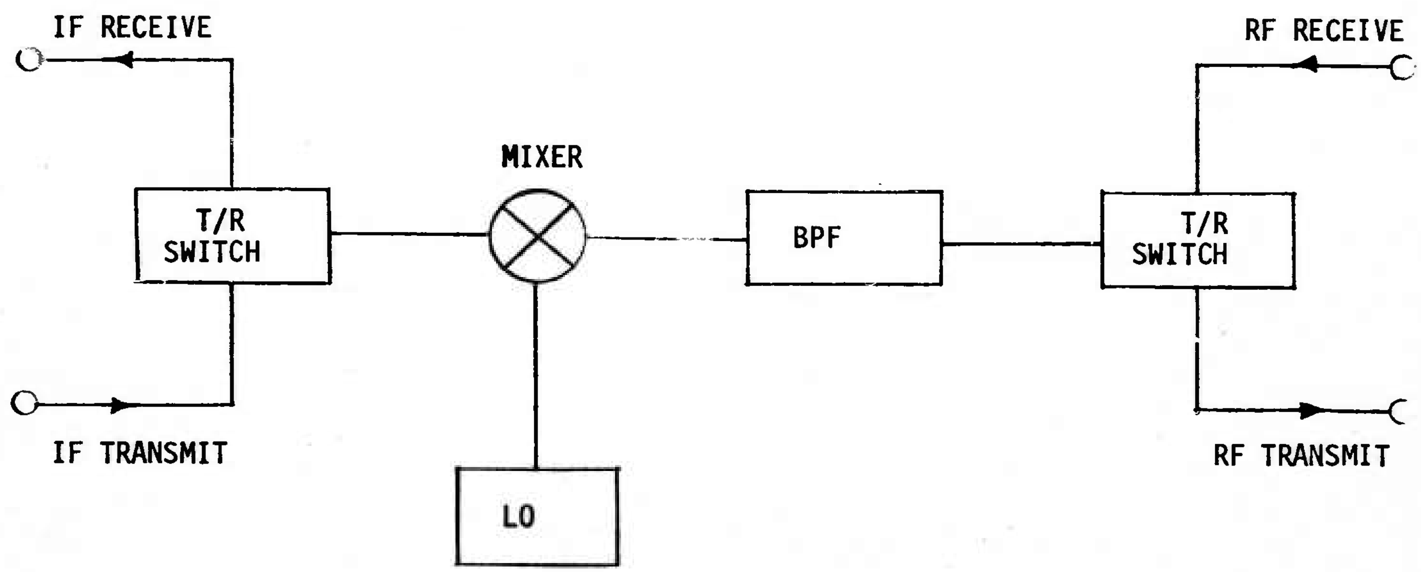


FIGURE 3.4.1 SINGLE MIXER SCHEME TO TRANSCEIVER REALIZATION

3.5 Filter Functions

Frequency Filters

As discussed in Section 2.5, the key to realizing small low loss filters is obtaining resonant elements in small size with high Q. This becomes increasingly difficult as the frequency increases. Therefore, alternatives are sought such as described in Subsection 3.4. Through a combination of efficient circuit design and using the best approach to small high Q resonant elements. The filtering requirements of a small packet radio can be met. Low loss filters that do not provide the required out of band filtering can be used as the front-end in combination with higher loss filters preceding the RF power amplifier chain.

Several filter types have been investigated and appear attractive as candidates for the front-end filter application. These include the end-coupled resonator, the parallel coupled resonator, the hairpin filter and the comb line filter. The end coupled resonator would require a lumped element low pass filter to attenuate the spurious response near 2fo, yet the total volume, assuming an ε of 7, is estimated to be 0.224 cable inches. The realization of the other filters mentioned also can be done in a similar size or less.

Since the minimum unloaded Q requirement for the RF filter is 500, conventional stripline and microstrip techniques are ruled out (unless air dielectric is used which results in relatively large size). Table 3.5.1 compares the characteristics of three filter types that could be used for the RF filter.

Interdigital line and combline filters with air dielectric have been designed and used for more than a decade. The unloaded Q's have been measured and calculated [26] at approximately 1200 and 600 respectively. The estimated volumes, when designed [26-28] with ground plane spacing of 0.2" and with 0.060" thick resonant bar for 4 pole, 2% bandwidth, are 1.3 cubic inches and 0.65 cubic inches respectively.

TABLE 3.5.1 COMPARISON OF CANDIDATE FILTER TYPES

	Interdigital Line	Comb Line	YIG
Attenuation characteristics 0 - 2fo	4-Pole Chebyshev		
Spurious ≥ 2 fo	3 fo, 5 fo ...	≥ 5 fo	65 dB below
Qu	≥ 1000	500-650	500-700
Insertion Loss	3 dB maximum	6 dB maximum	6 dB maximum
Temp. Sensitivity	≤1MHz for -20+71°C	≤5MHz for -20+71°C	≤5MHz for 0-60°C
Tunability	Mechanical tuning convenient	Mechanical tuning convenient	Magnetic biasing Inconvenient
Cost	Inexpensive	Inexpensive	Expensive
Fabrication	Easy	Easy	Complicated
Weight	≤ 3 Oz		
Size (estimated)	1.8"X2."X0.36"	0.9"X2."X0.36"	0.75" cube
Shield	Not required		Magnetic shield
Ability to take environmental punishment	More rugged		Less rugged

The YIG filter has been used for applications where electrical tunability is desirable. The tunability is achieved through using DC bias current to change the biasing magnetic filed which in turn changes the resonant frequency of the YIG spheres. For fixed frequency operation a permanent magnet can be used. A YIG sphere is very small, about 20-40 mil in diameter, and has very high unloaded O. Another important feature of the YIG filter is the limiting characteristics of the YIG crystal. Pure YIG (no gallium) has a limiting threshold of -23 dBm for frequencies below 3.6 GHz. Since the limiting characteristics are dependent on the amount of gallium doping of the material, it is possible to lower the usable frequency range by increasing the amount of gallium doping, so that the limiting threshold of +10 dBm can be obtained at 1800 MHz.

As may be concluded from Table 3.5.1 the combline filter appears to be the best compromise among the features compared. The YIG filter has the smallest size, but is more difficult to build and is less rugged. The interdigital line filter, while having the lowest loss, has the undesirable third harmonic response and is comparatively large.

Matched Filters

In the course of the previous discussion on matched filters, it was pointed out that the major cause of performance degradation in surface acoustic devices (SAWD) used as matched filters in packet radio is phase error that may exist across the device resulting from the center frequency of the incoming signal differing from the effective center frequency of the device. As the center frequency increases, the offset frequencies resulting from temperature variation and crystal misorientation increase, thus increasing the phase error and degrading performance. If some degree of flexibility in the SAWD center frequency is to be achieved to allow system optimization based on this parameter, the center frequency variation due to temperature and crystal misorientation must be reduced or compensated for. Temperature control techniques are beginning to be developed which may permit control of the substrate temperature with small power consumption and little additional increase in size. Greenhouse and McGill [29] have reported building a temperature controlled crystal oscillator by utilizing hybrid microcircuit substrate heater circuits. With approximately .2 watts steady state applied power they controlled the oscillator to +93°C $\pm$ .1° over -40 to +75°C. These results would indicate that the SAWD could be controlled to within $\pm$6° with perhaps one-tenth watt. Such control would probably require constant power drain since rapid turn-offs and turn-ons would demand substantially more power to establish temperature equilibrium. Other techniques for temperature control for SAW devices have been described in the 1975 IEEE Ultrasonics Symposium proceedings.

Center frequency error due to crystal misorientation can be reduced. The manufacturing process can produce crystals with an accuracy of +15 minutes from the true ST cut; fortunately, crystals can be measured with an accuracy of $\pm$ 3 minutes. Therefore, selectively grouping crystals based on the measured angle and using a different photomask for each group, misorientation errors can be partially compensated for. Another means of compensating for a center frequency error due to crystal misorientation is through adjustment of the receiver local oscillator. This method could, in effect, reduce errors resulting from misorientation to zero. If a single LO system is used for both transmit and receive, a reasonable approach would seem to be separate tuning elements (perhaps varactors) in the oscillator so that in transmit all terminals transmit as near as possible to the ideal center frequency but in receive the oscillator is switched to a slightly different frequency. Such a technique could also help minimize the effects of SAWD aging (20 to 30 ppm per year) since the receiver could be "retuned" rather than replacing the SAW device. Even with some rudimentary temperature control and selective crystal grouping, the performance of the SAWD remains questionable as the center frequency approaches 300 MHz. Still, the employment of these techniques would allow more flexibility in the center frequency than is currently allowed.

The actual crystal dimensions are approximately .3"W x 3."L x .025"H, therefore considerable size reduction can be made if the package design is optimized. A key to reduction of package size is to incorporate integrated matching networks adjacent to the SAW device terminals. In particular, lumped element spiral inductors can provide inductance and Q's approaching values obtainable with discrete coils. Caulton, [30], and other authors have reported lumped element techniques for both inductors and capacitors. In addition, these techniques have been applied to match SAW devices. These inductors were fabricated with copper on teflon glass boards, with overall dimensions of about .1" x .16". Therefore the networks could be placed on the SAWD substrate adjacent to the SAWD transducers. By employing matching techniques and designing an optimized package, the overall package size could be reduced to about .5" x 3.5" x .2".

Charge-Coupled Device Matched Filters

An attractive alternate approach to matched filtering in packet radio which avoids many of the trade-offs associated with the detection of the MSK waveforms using SAWD's makes use of charge-coupled devices (CCD) for differential coherent detection of MSK at baseband. Recent developments in the technology of charge-coupled devices indicate that baseband detection of MSK signals can be done using CCD matched filters [32].

Charge-coupled devices operate by transferring packets of minority charge, which represent analog signals, from one potential well to another [3]. These potential wells are formed by a linear array of deep depleted MOS capacitors, either on a uniformly doped substrate (surface channel) or a substrate with a thin, depleted layer of opposite conductivity at the surface (buried channel). Figure 3.5.1 presents the structure of a surface channel device. In operation, minority charge (electrons) exist in a thin inversion layer at the oxide-semiconductor interface. Since charges always move to the local potential minimum, charge is transferred from one potential well to the other by application of appropriate voltages to the electrodes. There are two measures that primarily determine device performance: 1) the time necessary to transfer charge from one storage site to the next, and 2) the fraction of the original charge that gets transferred. The two mechanisms responsible for charge transfer and thermal diffusion and fringing field drift. Thermal diffusion alone limits the clocking speed to about 1 MHz. Charge transfer speed is enhanced due to a small component of the electric field in the direction of charge transfer resulting from the externally applied gate voltages. The fringing field is responsible for clock rates to about 10-15 MHz in surface channel devices.

Charge transfer efficiency is determined by 1) the extent to which potential barriers exist between two potential wells and 2) the loss of carriers due to surface trapping effects. Overlapping silicon gates have helped to minimize charge transfer inefficiency due to potentital barriers [33]. Charge lost to fast interface surface states is minimized in the buried channel structure by ensuring that the depletion region wells start to form below the surface within the bulk of the semiconductor. Figure 3.5.2 shows the structure of the buried channel CCD. Improvement in the charge transfer efficiency of the buried channel CCD is accompanied by an improvement in clocking speed. The buried channel structure places the channel a greater distance from the gates thus increasing the relative importance of the fringing field drift. Clocking speeds above 200 MHz have been reported [32] and charge transfer efficiencies of 99.999% have been measured in buried channel devices [31].

The unique properties of charge-coupled devices (CCD) make them attractive in a wide variety of applications: Their light sensitivity makes them useful for self-scanning solid state imagers; their small size and high packing density are good attributes for high volume digital memory; and their analog nature makes them useful for many signal processing applications. As near ideal analog shift registers, the CCD permits the fabrication of highly complex functional devices very simply. One very attractive application of charge-coupled devices to packet radio is matched filtering. Packet radio employs a differentially coherent minimum-shift-keyed (MSK) modulation technique in a spread spectrum system. Charge-coupled device matched filters can be used in the MSK demodulation process.

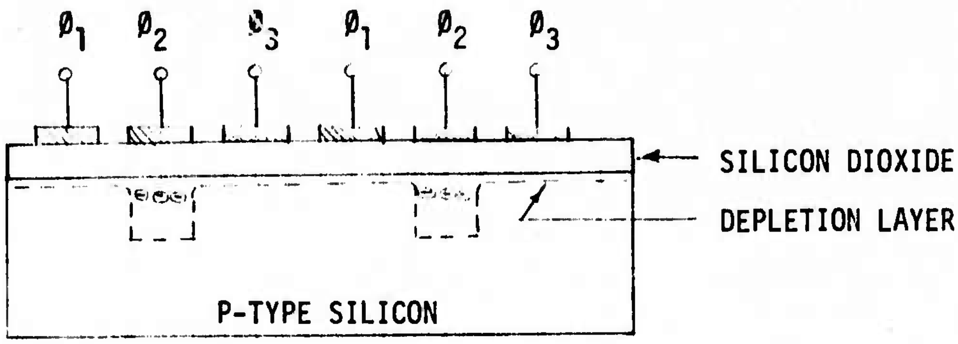


FIGURE 3.5.1 THREE-PHASE SURFACE CHANNEL CHARGE-COUPLED DEVICE

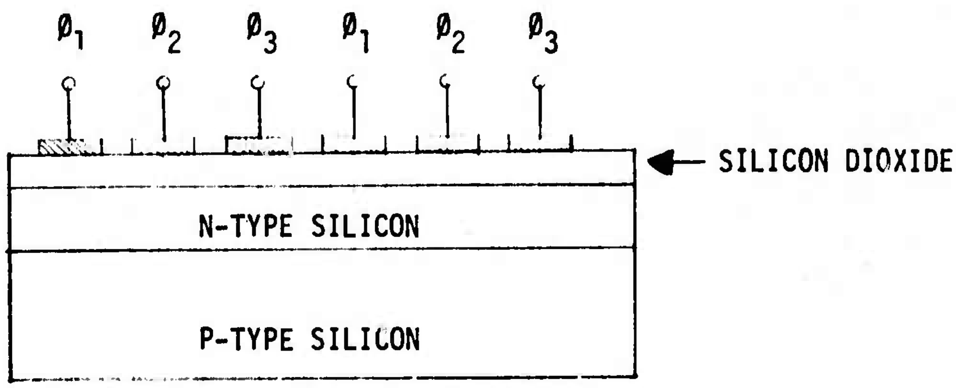


FIGURE 3.5.2 THREE-PHASE BURIED CHANNEL CCD

MSK is a digital modulation technique that generates a constant amplitude, phase continuos signal. MSK can be viewed as a special case of offset quadrature phase-shift-keying with sinusoidal symbol weighting and can be mathematically expressed as

$$y(t) = C_I \mathrm{Cos}(\frac{\pi t}{2T_c}) \mathrm{Cos} w_o t + C_Q \mathrm{Sin}(\frac{\pi t}{2T_c}) \mathrm{Sin} w_o t \qquad (1)$$

where C_I and C_Q represent the inphase and quadrature subchannel sequences and w_o the carrier frequency. Figure 3.5.3 provides a general pictorial description of MSK waveforms. Data is transmitted in a 128-chips-per-bit differential format at 100 KBS. This implies that the quadrature subchannels are carrying independent chip sequences of 64 chips per bit each.

A matched filter is one whose impulse response is the time inverse of the signal waveform to which it is matched. If $V_i(t)$, a voltage waveform that exists over the interval [0, T] is the input to a linear time-invariant filter with impulse response h(t), then the output is given by

$$V_o(\tau) = \int_0^T V_i(t)\, h(\tau - t)dt \quad 0 \leq \tau \leq T \qquad (2)$$

Now, if the filter impulse response is the time inverse of $V_i(t)$, the output voltage is a maximum at $\tau = T$.

$$V_{max} = V_o(T) = \int_0^T V_i(t)V_i(t - T)dt \qquad (3)$$

Figure 3.5.4 provides a conceptual diagram of the implementation of a matched filter using a charge-coupled device. The matched filter consists of a sampling circuit followed by N delay stages. The input voltage $V_i(t)$ is sampled and a charge proportional to the sampled voltage is shifted to the right at the occurance of each clock pulse. Each delay stage is tapped, multiplied by the weighting coefficients, w_k, and summed. When the coefficients, w_k are selected to achieve an impulse response which is the time inverse of $V_i(t)$, then equation (3) is approximated; the actual output is given by

$$V_o(t) = \sum_k w_k V_i(t - (k - 1)\, T)$$

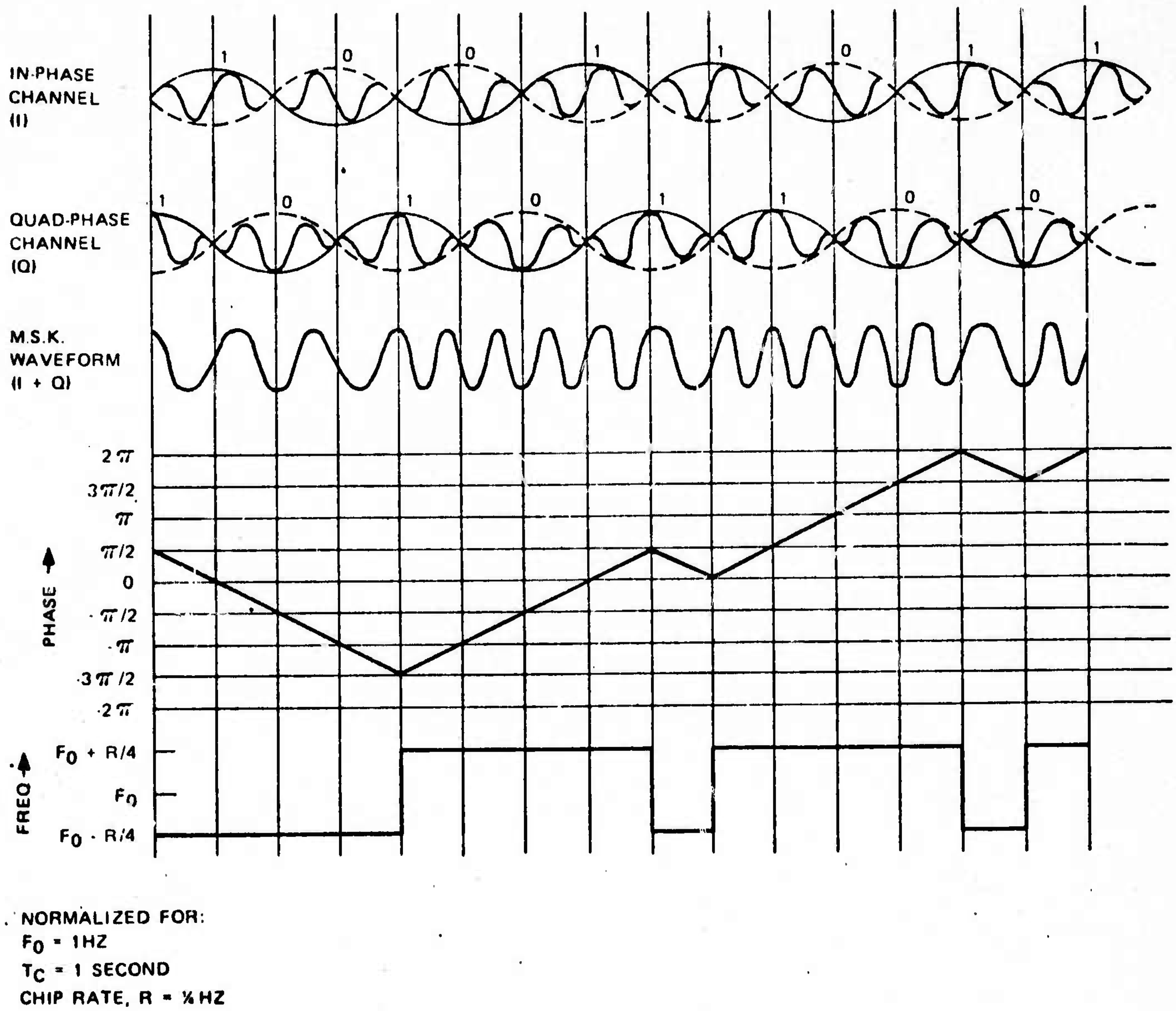


FIGURE 3.5.3 EXAMPLE OF MINIMUM-SHIFT-KEYED (MSK) WAVEFORMS

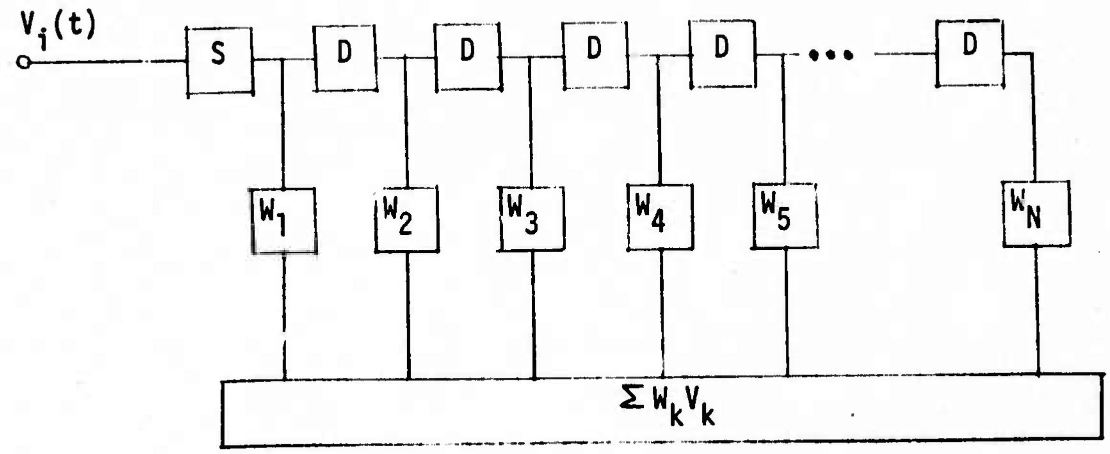


FIGURE 3.5.4 IMPLEMENTATION OF THE MATCHED FILTER WITH CHARGED-COUPLED DEVICE

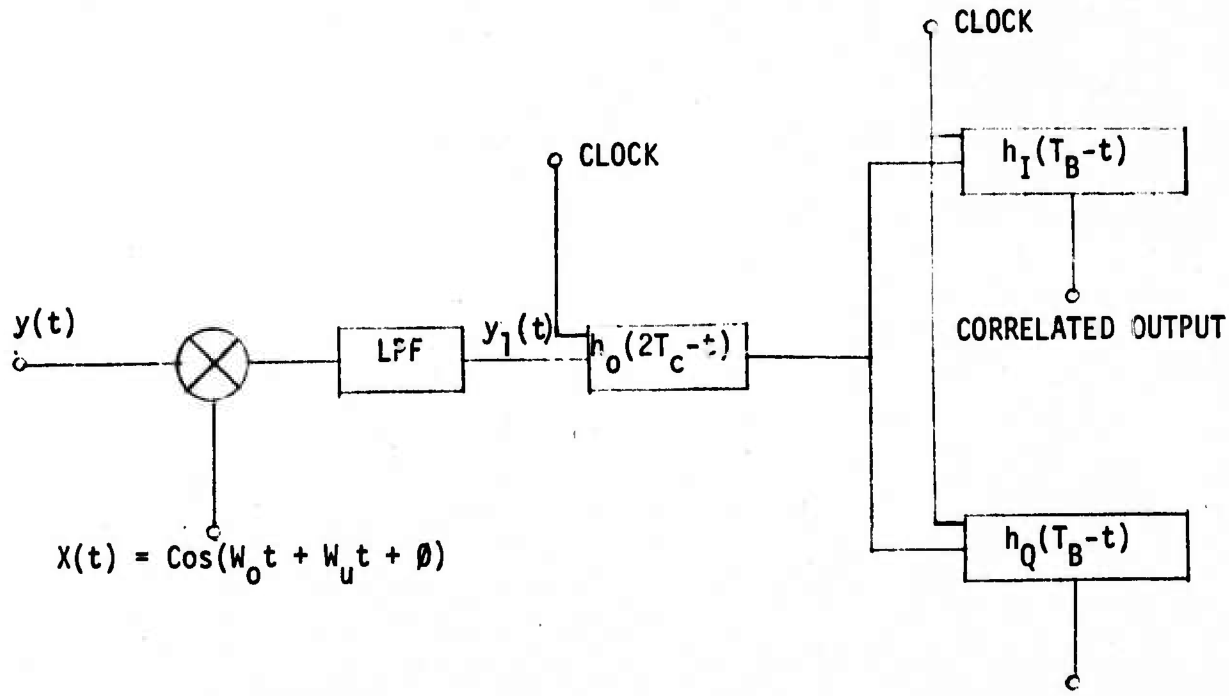


FIGURE 3.5.5 MSK DETECTION AT BASEBAND USING CCDs

Because of the frequency limitations of CCD's, they are inherently baseband devices. This means translating the MSK modulated carrier from 1800 MHz to baseband or equivalently, reducing the number of RF cycles from 140 3/4 cycles per chip to 1/4 cycle per chip. Since MSK modulation consists of quadrature subchannels carrying independent chip sequences at one-half the total chip rate, separation of the two channels is necessary. Complete separation, however, of the subchannels can be done only in a phase coherent system. Since packet radio is non-coherent at the chip level and differentially coherent at the bit level, the implementation of a CCD matched filter detection must provide this capability. The implications of non-coherent detection at baseband of MSK waveforms can be explained with the help of block diagrams of Figure 3.5.5. The received signal at RF, y(t), is mixed with the local oscillator, X(t).

$$X(t) = \cos(w_o t + w_u + \emptyset)$$

where w_o is the RF carrier frequency, w_u represents the uncertainty in the received signal frequency and $\emptyset$ is its unknown phase. w_u results from both transmit and receive local oscillator instability and doppler shifts in a mobile environment. The phase of the received signal, in general, is a random variable uniformly distributed over [0, 2π]. The unknown phase function

$$\theta(t) = w_u t + \emptyset$$

is assumed to be a slowly varying function of time such that the phase uncertainty is essentially a constant from bit to bit. From the low pass filter, the received signal has the form

$$y_1(t) = C_I \mathrm{Cos}\left(\frac{\pi t}{2T_c}\right) \cos\theta + C_Q \mathrm{Sin}\left(\frac{\pi t}{2T_c}\right) \mathrm{Sin}\theta \qquad (5)$$

Generally, $y_1(t)$ contains both inphase and quadrature signal components and depending on the value of the phase constant θ, one subchannel may be present without the other. Therefore, a phase invariant system requires using two matched filters, one for each subchannel. It also becomes evident from Equation (5) that quadrature mixing of the incoming signal is required to ensure that one of the subchannels is never completely lost. This requires parallel processing of $y_1(t)$.

For convenience in the description of the demodulation process, Figure 3.5.5 shows a matched filter $h_0(2T_c - t)$ matched to the cosine weighting of the MSK waveforms followed by filters matched to the inphase and quadrature chip sequences. When matched filters are actually implemented for MSK detection h_0 is likely to be incorporated into h_I and h_Q. The output of h_0 is the correlation of h_0 and $y_1(t)$ and is expected to be a series of correlation pulses defined over $[-2T_c, 2T_c]$. These correlation pulses which represent the chip sequences are clocked into h_I and h_Q. Maximum output is observed from h_I and h_Q when the input chip sequence matches the "built in" chip sequence of the matched filter.

In order to characterize the performance of charge-coupled devices for MSK signal demodulation, the effects of the frequency uncertainty w_u and sampling rate have been investigated. Figure 3.5.6 shows the degradation in the magnitude of the correlation pulse from h_0 for various values of w_u. Curve 2 (k = 1) represents an w_u of 1.8 MHz which results in about 0.5 dB degradation. Clearly, w_u would have to be very large to severely degrade the correlation magnitude. The limitation on w_u is determined by the reasonable phase error that can be sustained across h_I and h_Q. Allowing 5 dB degradation in signal-to-noise ratio establishes this limit at 0.4 radians which corresponds to maximum frequency error of 6.4 KHz.

Insight to the effect of the sampling rate on the signal-to-noise ratio of the CCD detection process can be gained by assuming the CCD to be an ideal sampler (Figure 3.5.7) and the noise to be zero mean white additive noise. Let

$$S(t) = \sqrt{\frac{E}{T_c}} \sin\left(\frac{\pi t}{2T_c}\right) \quad 0 \leq t \leq 2T_c$$

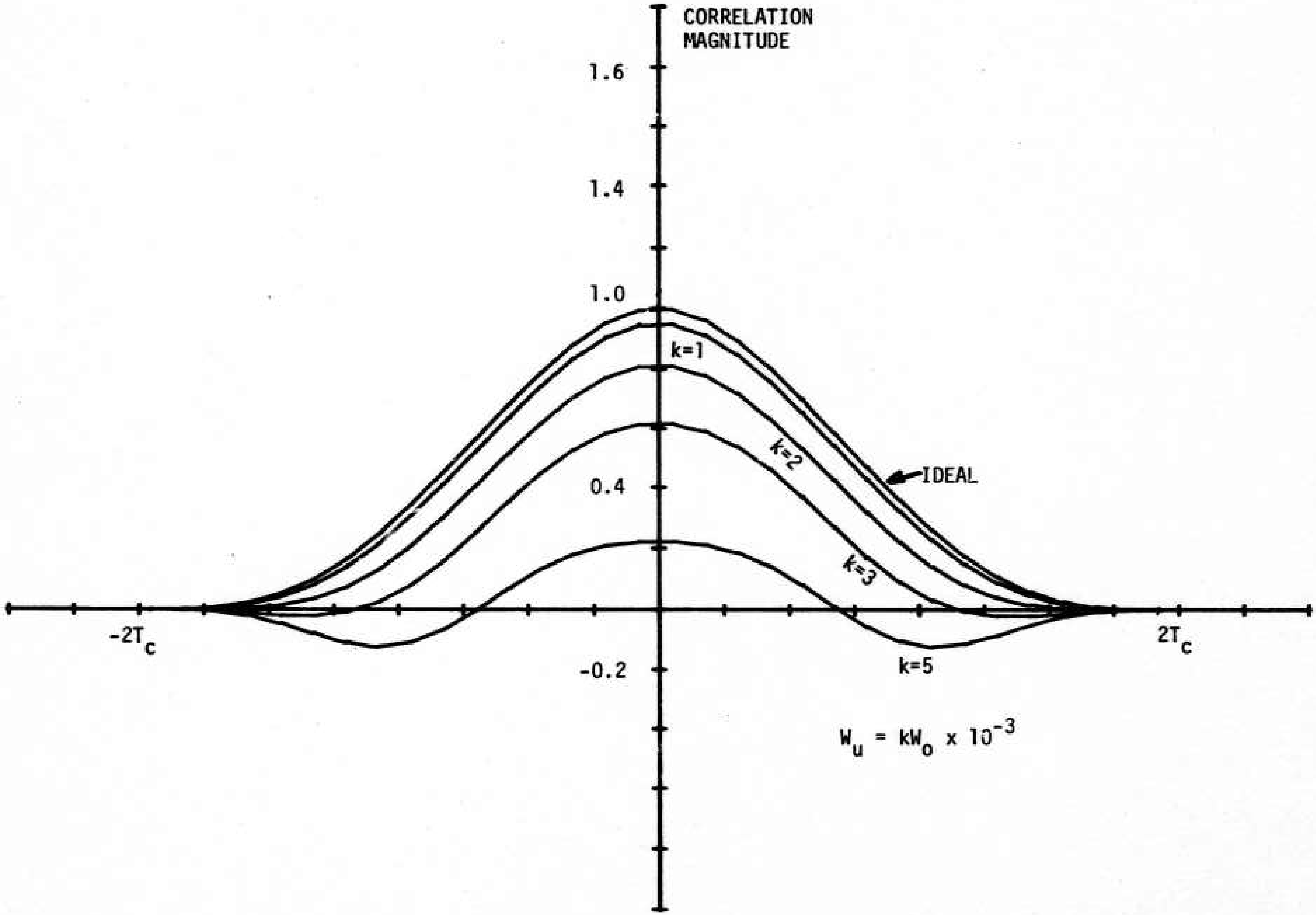


FIGURE 3.5.6 CORRELATION MAGNITUDE VERSUS FREQUENCY ERROR (Ø = 0)

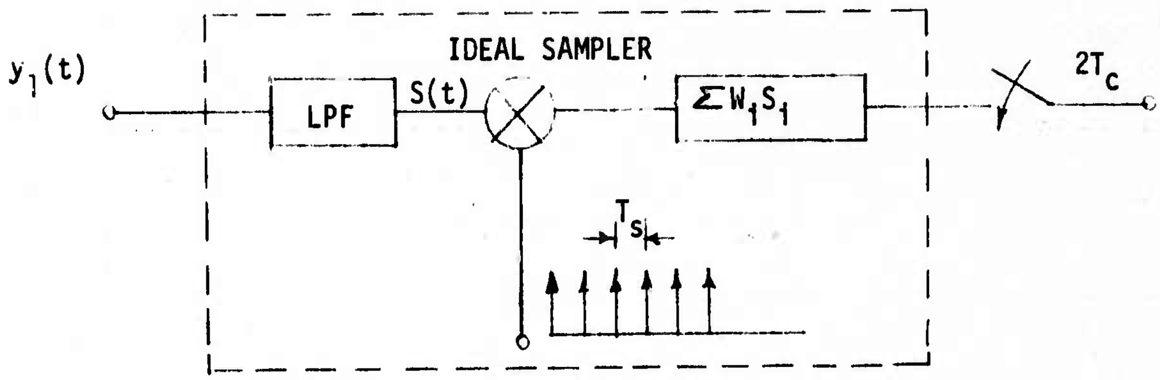


FIGURE 3.5.7 CCD MODEL FOR SNR DERIVATION

Then the signal energy is

$$E = \int_0^{2T_c} S^2(t)dt$$

The matched filter h_o has weighting coefficients $w_i = S[(i - 1/2)T_s]T_s$ $(i = 1, 2, . . . N)$ where T_s, the sampling frequency, is $\frac{2T_c}{N}$. Since the signal samples can be sampled with some error Δt with respect to the matched filter samples, the signal samples can be expressed as

$$S_i = S[\Delta t + (i - 1/2)T_s] \quad -\frac{1}{2}T_s \leq \Delta T \leq \frac{1}{2}T_s$$

The signal output for N samples is

$$E_s = \sum_{i=1}^{N} w_i S_i$$

The low pass filter establishes the noise bandwidth at B, with a variance of $N_o B$. Then the output noise is

$$n = \sum_{i=1}^{N} U_i \; S[(i - 1/2)T_s]T_s \quad i = 1, 2 . . . N$$

where U_i is the ith noise sample.

It follows that

$$E\{n^2\} = \sum_{i=1}^{N} N_o B T_s^2 S^2[(i - 1/2)T_s]$$

is the output noise power for n samples. The output signal noise ratio becomes

$$SNR = \frac{E_s^2}{E[n^2]} = \frac{\left\{\sum_{i=1}^{N} S[(i - 1/2)T_s^2 \; S[\Delta t + (i - 1/2)T_s)T_s\right\}^2}{N_o \; BT_s^2 \sum_{i=1}^{N} S^2[(i - 1/2)T_s]} \qquad (6)$$

SAMPLING RATE $R = \frac{1}{T_s}$	# OF SAMPLES N	$\Delta T = T_s/4$		$\Delta T = T_s/2$	
		SNR	DEGRADATION dB	SNR	DEGRADATION dB
$1.5\ R_c$	3	$1.866\ \frac{E}{N_o}$	-0.3	$1.500\ \frac{E}{N_o}$	-1.25
$2\ R_c$	4	$1.924\ \frac{E}{N_o}$	-0.16	$1.707\ \frac{E}{N_o}$	-0.688
$2.5\ R_c$	5	$1.950\ \frac{E}{N_o}$	-0.11	$1.809\ \frac{E}{N_o}$	-0.400
$3.0\ R_c$	6	$1.965\ \frac{E}{N_o}$	-0.077	$1.860\ \frac{E}{N_o}$	-0.315
$4.0\ R_c$	8	$1.981\ \frac{E}{N_o}$	-0.041	$1.922\ \frac{E}{N_o}$	-0.170
$R \to \infty$	$N \to \infty$	$2.000\ \frac{E}{N_o}$	0	$2.000\ \frac{E}{N_o}$	0

Table 3.5.2 OUTPUT SIGNAL-TO-NOISE RATIO OF CCD MATCHED FILTER

As the number of samples N goes to infinity and Δt goes to zero, the SNR approaches $\frac{2E}{N_o}$. Table 3.5.1 contains the calculated SNR's using Equation (6) for several values of sampling rates and sample timing errors. Worst case degradation at a sampling rate that is twice the chip rate or 4 samples per $2T_c$ period is 0.688 dB. Twice the chip rate corresponds to a 25.6 MHz sampling rate and requires a CCD matched filter containing 256 cells to detect a 64 chip sequence.

A Multifunctional Charge-Coupled Device

Since the inphase and quadrature subchannels carry independent chip sequences, the implementation of a packet radio receiver using charge-coupled devices would require designing and fabricating two separate devices. A more attractive approach is to build a single programmable device which could perform a multitude of functions that would be useful to packet radio. Figure 3.5.8 presents a block diagram of a multi-functional programmable device. Appropriately programmed, this device will 1) perform match filtering for each inphase and quadrature subchannels, 2) provide a delayed version of the input signal, and 3) deliver the integrated output of the signal existing in both halves of the device.

In order to perform the match filtering function, the appropriate chip sequence is reversed and clocked into the shift register at one-fourth the basic clock rate or 6.4 MHz. Using a clock rate of 25.6 MHz to the CCD, four cells are required to provide adequate matching to the cosine weighting of the MSK waveform. In other words, taps from the CCD cells are cosine weighted and summed. This correlated output represents an MSK chip which is multiplied by the appropriate bit in the shift register. Summing over all 64 stages of the shift register implements the matched filtering function.

The delay function is implemented simply by providing an output network for the charge-coupled array.

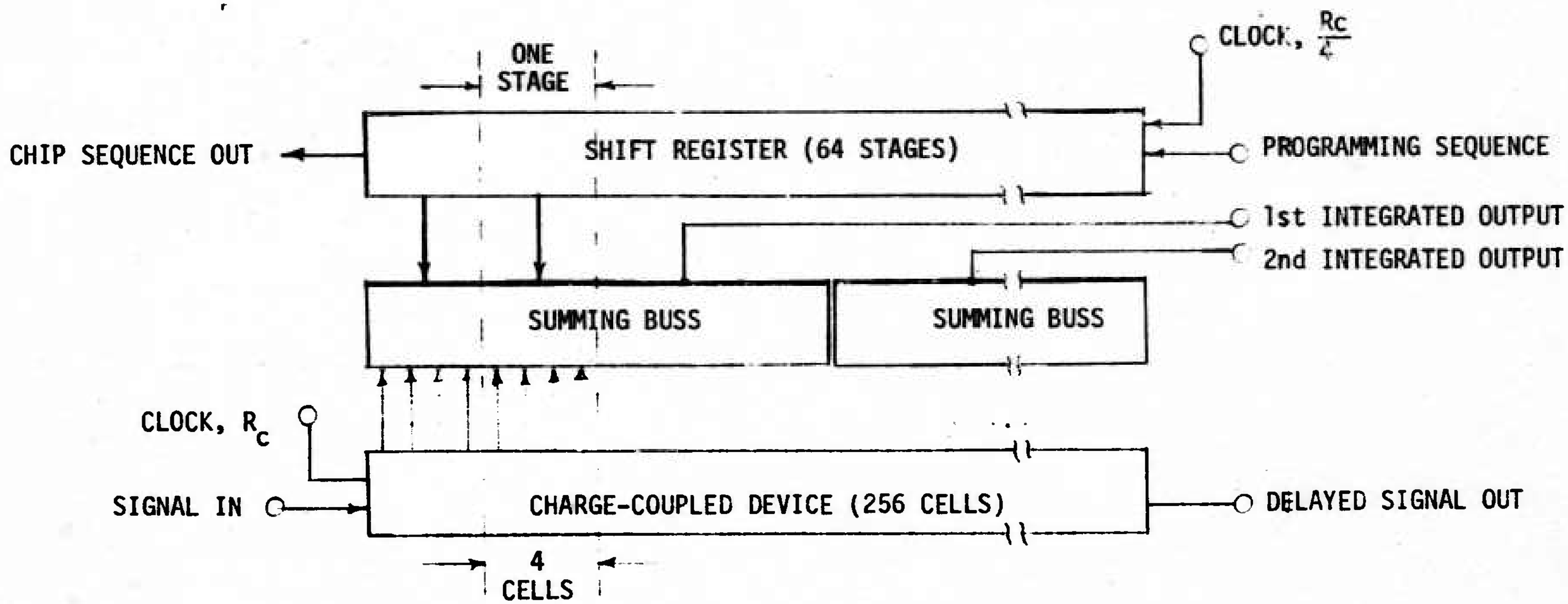


FIGURE 3.5.8 MULTIFUNCTIONAL CHARGE-COUPLED DEVICE FOR SIGNAL PROCESSING

Figure 3.5.8 as shown illustrates "cosine-weighted" integration with an integrated output for the first 128 cells of the CCD and an integrated output for the second 128 cells. The integrated output is accomplished by setting the shift register to all ones. In order for the input signal to be truly integrated, the weights on all cells would have to be unity. In packet radio, where the integration function could used for signal processing purposes, it is believed that a cosine weighted integrated output would suffice for the intended application.

The following are desirable features for the device described in Figure 3.5.8.

1) A clocking frequency of 25.6 MHz;

2) Total power dissipation including dissipation in the clock driver circuitry on the order of 100 MW;

3) The clocking voltages be logic compatible; that is, TTL, ECL, etc;

4) A dynamic range of at least 40 dB;

5) Output circuitry such that the delayed signal out has zero dB loss with respect to the input signal.

6) Excluding effects of the codes used, the ratio of correlation peak to the highest side lobe should be about 40 dB.

Receiver Implementation Using CCD's

Charge-coupled devices as intrinsic low noise analog signal processing devices based on the well-developed silicon integrated circuit technology should find a wide range of applications in packet radio. Charge-coupled device applications to packet radio can offer reduced size, reliability, flexibility and improved performance.

Figure 3.5.9 presents a block diagram of one approach to realizing a packet radio receiver that employes charge-coupled devices in a variety of functions which will result in reduced-size packet radio. In Figure 3.5.9 the received signal is translated to baseband by quadrature mixing, followed by low pass filtering before being detected by charge-coupled device matched filters. The multipath integration function following the detection process can be implemented using a charge-coupled device. Also, shown in Figure 3.5.9 is a recirculating, coherent filter designed to improve the signal-to-noise ratio at the data decision point. The coherent filter consists of the positive feedback loop around a one-bit delay line with a loop gain $\alpha<1$. This recirculating loop adds the correlation pulses from the matched filters coherently while the noise is added incoherently, thereby improving the signal-to-noise ratio by $\frac{1+\alpha}{1-\alpha}$.[36] The delay line is a charge-coupled device as well as the initial delay necessary to allow the data decision to "inverse modulate" the received signal so that the coherent filter receives pulses all of the same polarity. The CCD delay line in the coherent filter is tapped to provide input to the bit sync timing circuitry and the automatic gain control circuitry.

A charge-coupled device approach to receiver implementation for packet radio could offer many attractive features including the following:

1) For baseband detection, the image rejection problem is non-existant; therefore, front-end filter requirements are relaxed.

2) Charge-coupled devices operate over a wide temperature range with very little performance degradation.

3) Increased processing gain can be realized very easily. A 3 dB increase at 100 KBS or 6 dB or more at slower bit speeds can be accomplished.

4) The CCD is a device that exploits the technology of large-scale integration. It is quite possible that the entire matched filter section including the multiplication and summing functions can be realized on a single chip. Thus, the matched filter section has the potential of being very much smaller with higher performance than an equivalent surface acoustic wave device. Large-scale integration of the post detection signal processing functions is more readily permitted by using CCD performing a variety of functions.

5) The local oscillator stability can be relaxed, but not enough it appears, to avoid temperature compensation of the crystal.

6) The insertion loss of a CCD matched filter is determined primarily by the input and output circuits. In a good CCD signal attenuation due to charge-transfer in efficiency can be small for matched filter applications. The overall insertion loss is likely to be much less than equivalent SAW devices.

7) The tap weights of the CCD matched filter can be programmed in a very straightforward manner to allow dynamic code changing, providing the capability for increased spoof protection.

Although CCDs have many attractive features for signal processing applications, they have some drawbacks. The single most undesirable characteristic of CCDs is the large power requirements of the clock drivers. CCDs themselves consume very little power; however, their clocking requirements demand 10 to 15 volt swings. At high clocking frequencies, the power requirements of the clocking circuits could make CCDs unacceptable for small packet radio applications.

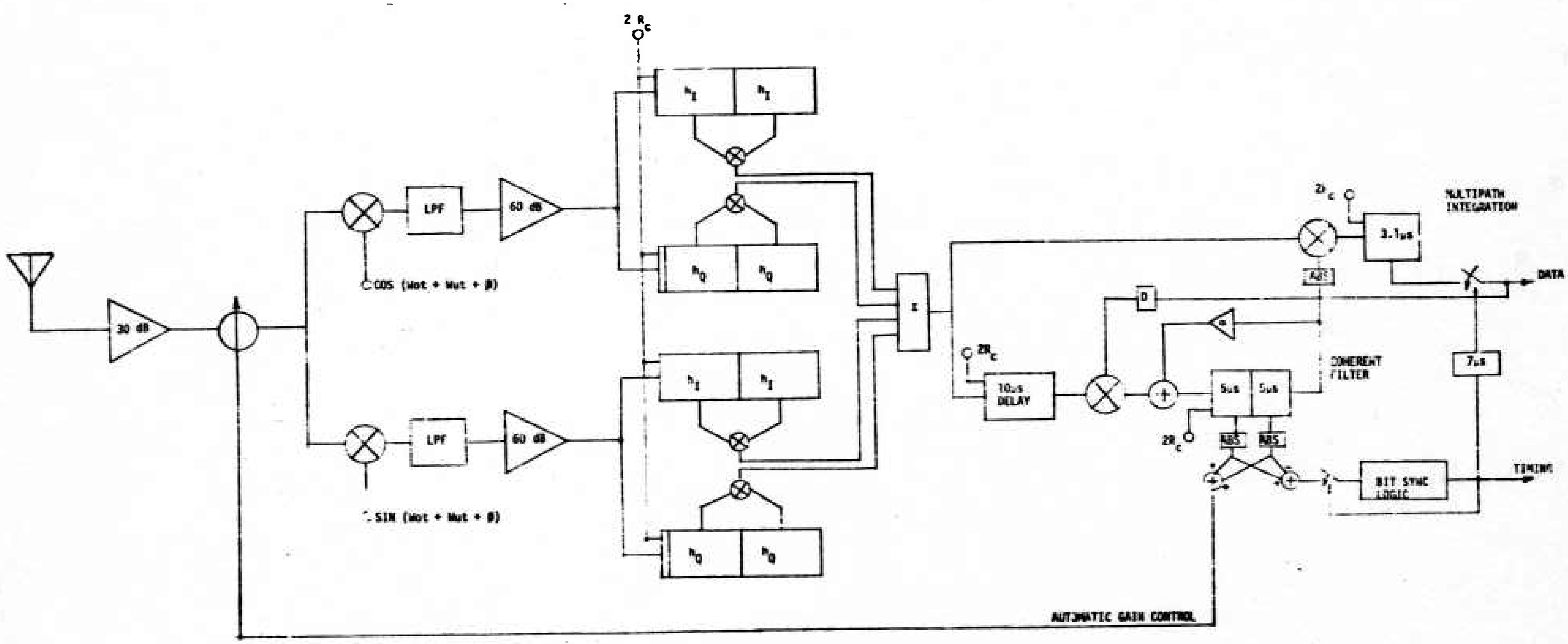


FIGURE 3.5.9 DIFFERENTIALLY COHERENT MSK RECEIVER USING CHARGE-COUPLED DEVICES

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4.0 HAND HELD PACKET RADIO: DESIGN EXERCISE

The fundamental constraints to realization of a reduced size packet radio and potential solutions to relief from these constraints has been explored. In an effort to focus the problem on today's technology and give a state-of-the-art data point, a design exercise for a hand held packet radio was conducted. This section contains a design plan description of a hand held device and it includes general system description, block diagrams, specification summary, and an artist's conception of the unit. This is organized as follows: Transceiver description, digital section description, power source, antenna description, I/O device description, and summary. It should be noted that this design exercise does not incorporate charge coupled device matched filters or the SAWD oscillator as described in Section 3.0. The incorporation of these two techniques would further enhance the hand held design.

4.1 Transceiver Description

The salient operational and mechanical features pursuant to size reduction in the RF section are summarized below:

1) A single passive mixer scheme to be used for transmit and receive signal conversion;

2) Circuit multiplexing by incorporation of PIN diode switching networks to permit time sharing of common system functions;

3) A direct synthesis approach to the LO system for low power consumption;

4) Chip-carrier circuitry to be used in the transmit, receive, and LO sections for high package density;

5) Lumped element components to be used at microwave frequencies for impedance matching and filtering as opposed to distributed techniques;

√

6) A rugged package design to permit ease of handling. The case would be hermetically sealed and impervious to damage from the environment.

A candidate approach to the RF section is shown in the functional block diagram of Figure 4.1.1 and Figure 4.1.2 illustrates how the RF section may be physically organized.

During reception, RF signals from the antenna are injected to a 2-pole band-pass filter section primarily designed to suppress intermodulation responses and harmonic outputs from the transmitter. The filtered signal is switched via SPDT diode network to a narrow-band chip-carrier low noise amplifier. The amplifier stage provides approximately 23 dB of RF gain, sufficient enough to overcome the losses of the second switch, YIG filter, and mixer to establish the receiver noise figure at less than 8 dB. From the low noise amplifier the RF signal is switched again and filtered in a fixed tuned 4-pole YIG filter. This stage provides image rejection of at least 60 dB as well as selective limiting of high level in band signals. Once filtered the received signal is high side injected to a passive mixer stage and down converted to 67 MHz. At the IF, approximately 60 dB of gain is added in the noncoherent AGC stage prior to injection into the matched filter.

During the transmit mode the transmitter accepts a data packet in serial bit form from the microprocessor, encodes the bits with the spread spectrum codes, converts this digital chip signal to constant amplitude, phase continuous MSK, and upconverts to the RF band via high side injection to the single passive mixer stage. The YIG filter provides at least 45 dB attenuation to the LO and spurious mixer products. Once up-converted and filtered the transmit signal is amplified by a factor of 45 dB in the pre-amp, driver and power amplifier stages for a minimum transmit power of 3 watts.

The local oscillator system utilizes an oscillator-multiplier configuration and achieves the necessary stability. A crystal oscillator with an output frequency of 51.25 MHz serves as the frequency standard. This frequency is sequential multiplied with active and passive chip multiplier stages to produce an LO of 1691 MHz. Harmonic filtering is achieved initially via a miniature filter

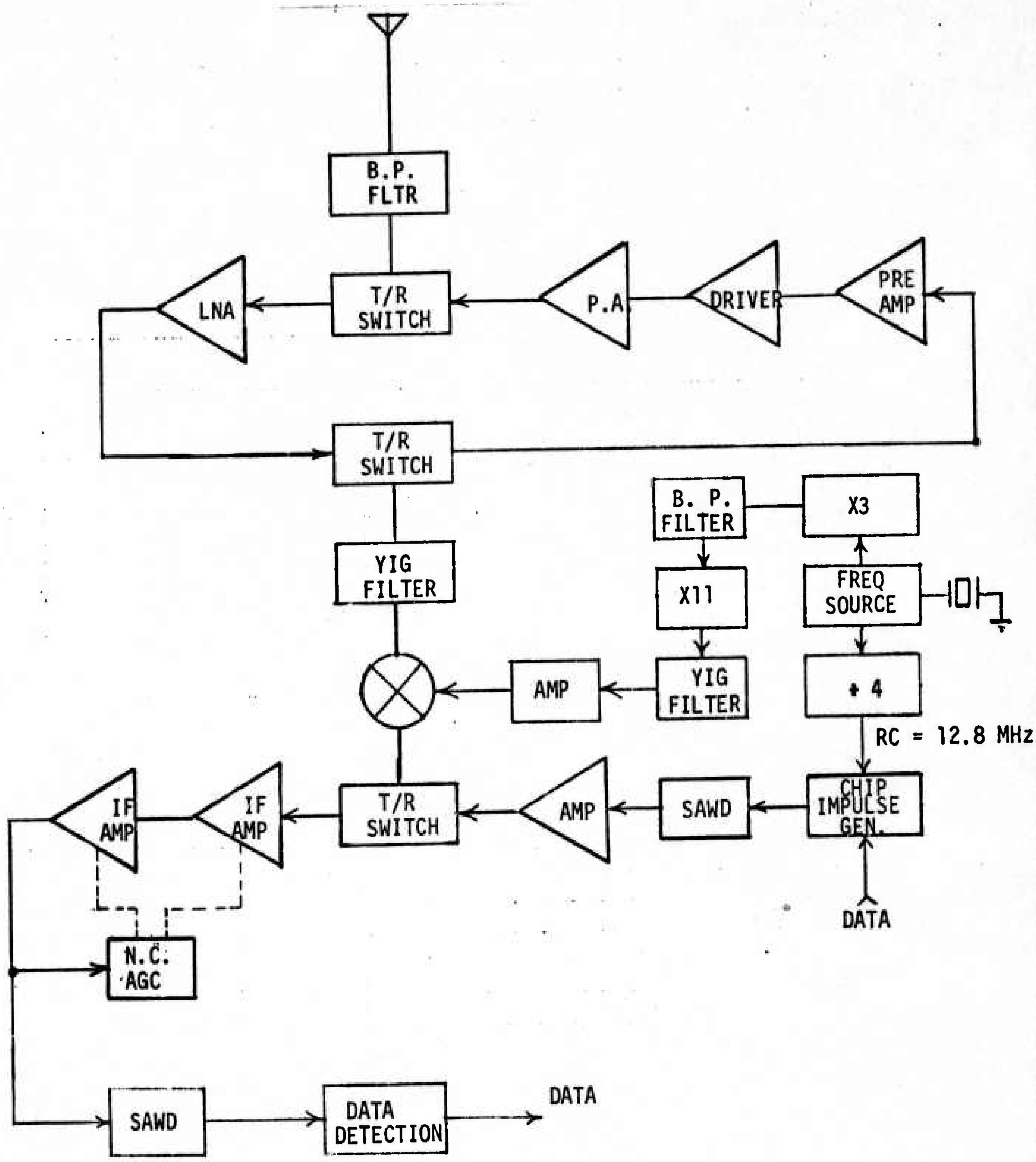


FIGURE 4.1.1 FUNCTIONAL BLOCK DIAGRAM OF RF SECTION

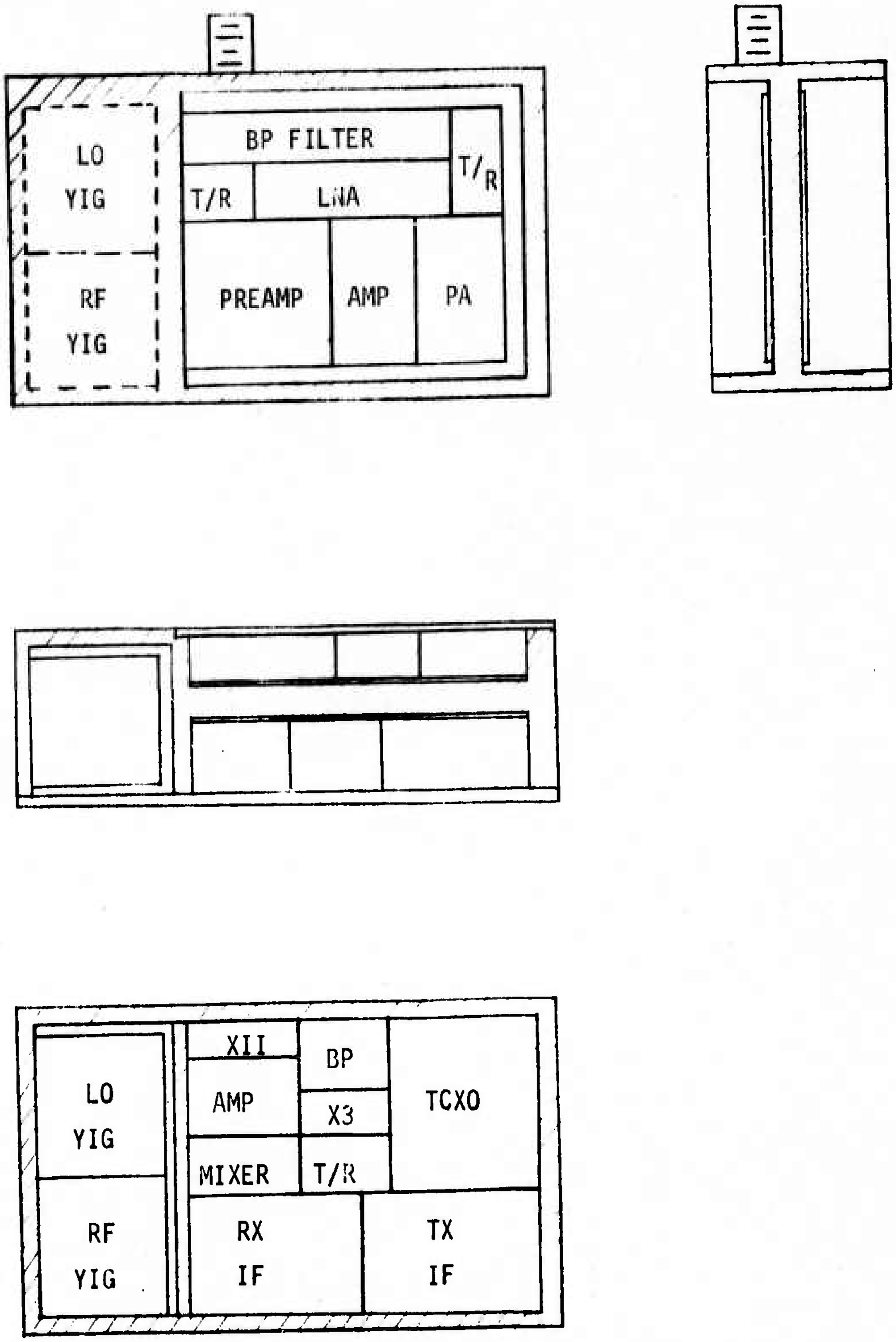


FIGURE 4.1.2 POSSIBLE ORGANIZATION OF RF SECTION

TABLE 4.1.1 REDUCED-SIZE PACKET RADIO RF SECTION PERFORMANCE

TRANSMITTER	
Frequency (20 MHz bandwidth)	1758 MHz
Power Output	3 watts (minimum)
Spurious Outputs	≥ 50 dB below carrier
RECEIVER	
Frequency (20 MHz bandwidth)	1758 MHz
Noise Figure	8 dB
Image Rejection	60 dB (minimum)
RF/IF Gain	15 dB (minimum)
IF Frequency	67 MHz
LOCAL OSCILLATOR SYSTEM	
Frequency	1691 MHz
Power Output	0 dBm (minimum)
Spurious Outputs	≥ 50 dB below

structure utilizing monolithic chip inductors and MOS capacitors. A 4-pole YIG filter provides final filtering prior to injection to the mixer. The 12.8 MHz chip clock is derived from the crystal by dividing down.

Table 4.1.1 summarizes the performance measures of the RF section of a reduced power, single channel, 20 MHz bandwidth, single data rate packet radio.

4.2 Digital Section Description

Progress toward miniaturization in digital electronics has been significant. The number of circuits per unit area almost doubles each year. Transistor gates can now be realized in less than 5 square mils, providing thousands of devices per chip. Chip densities are expected to continue to increase [1].

Large-scale integrated circuit chip sizes; i.e., most single chip microprocessors, are on the order of 200 X 200 mils. This is considered a large chip in today's integrated circuit technology. It has been observed that chip yield falls off at a less than exponential rate as active area increases [2]. There is no fundamental reason why yield should not be 100 percent; for every rejected circuit represents a flaw that theoretically could have been avoided [3]. Improvements in fabrication and processing procedures are expected to decrease defect densities and thereby improve yield and allow larger and larger integrated circuit chips. As a result, the implementation of digital functions in packet radio is not viewed as an impediment to size reduction. It is not at all unreasonable to assume that at some point in time all the digital functions in packet radio can be realized on a single monolithic integrated circuit chip.

Current technology in digital electronics will allow all the digital functions necessary for control and processing and in a reduced-size packet radio terminal to be implemented in volume of less than 2 cubic inches and consume less than one watt of power. Although not able to boast of the density or speed of the other technologies, complimentary-metal-oxide semiconductor (CMOS) technology has the lowest power consumption. As the importance of power consumption in a reduced-size packet radio is paramount, the entire digital section in packet radio is envisioned to be implemented in CMOS. Where the speed of CMOS will not suffice, CMOS silicon-on-sapphire would be employed, since operation above 40 MHz has been reported for this technology [4].

A single chip CMOS microprocessor from RCA has the following characteristics:

1) word length - 8 bits
2) machine cycle time - less than 4 μs
3) average power consumption - 60 mw
4) supply voltage - 5 volts
5) on chip DMA channel

CMOS random access memory (RAM) and CMOS read only memory (ROM) have power consumptions on the order of 50 nanowatts per bit in the standby mode and 15 microwatts per bit during operation. Access times are less than 800 ns. Current integration level for CMOS RAM is one kilobit per chip. However, 4 kilobit static CMOS memory is expected to appear soon [4]. Memory consisting of 4K 16-bit words of CMOS RAM and 1K 16-bit words of CMOS ROM is expected to consume a maximum of 200 mw of power.

Other necessary digital functions such as address decode, timing and control, input-output channel, and radio-digital interfaces can be implemented in CMOS with an estimated power consumption less than 500 mw. The resulting 20-22 chips could be integrated using hybrid thick or thin film technology to arrive at a volume of less than 2 cubuc inches.

4.3 Power Source

The estimated requirements on the power source is five watts of continuous power plus 15 watts of peak power at a 15% duty cycle for ten hours. A 65 watt-hour source will meet this requirement. Supply voltages of +15 and +5 are needed.

There are two approaches to provide the necessary voltages. As higher capacity cells tend to have greater energy densities, DC-to-DC conversion may

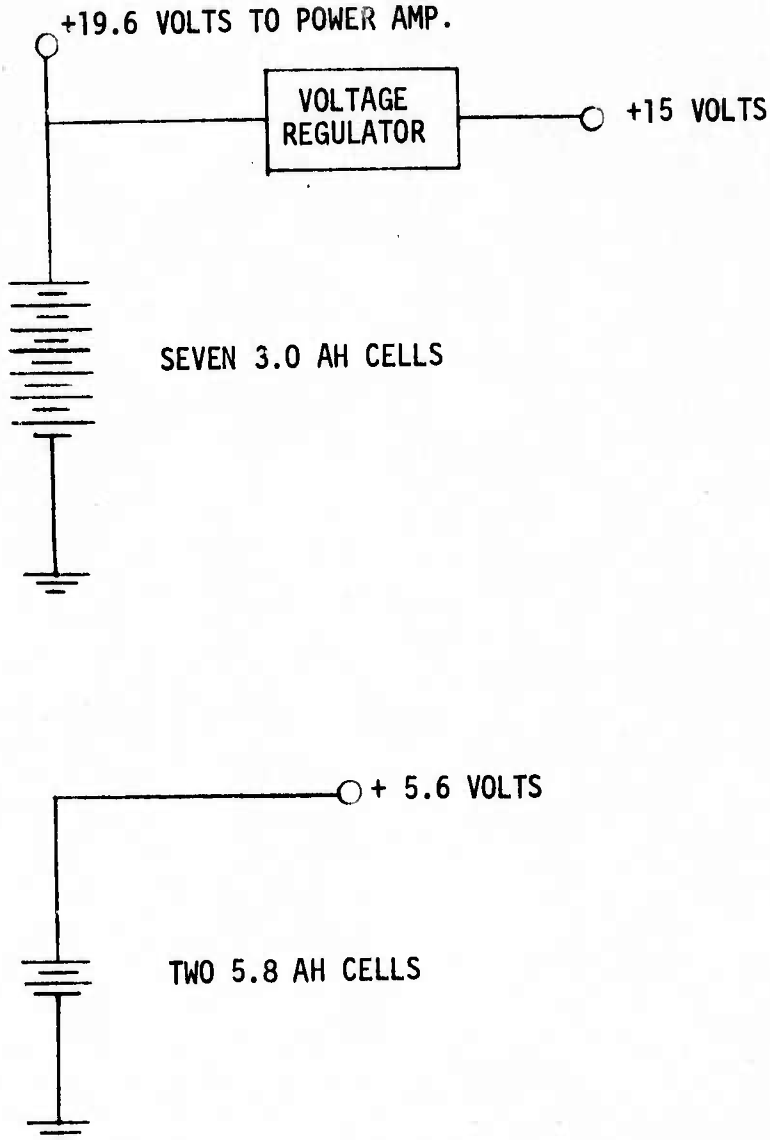


FIGURE 4.3.1 POWER SUPPLY SYSTEM USING LITHIUM CELLS

appear to take advantage of the greater energy densities of higher capacity cells as opposed to stacking smaller cells to achieve the supply voltages. However, the poor efficiency of DC-to-DC converters more than cancels the increased energy density high voltage cells when low supply voltages are also required.

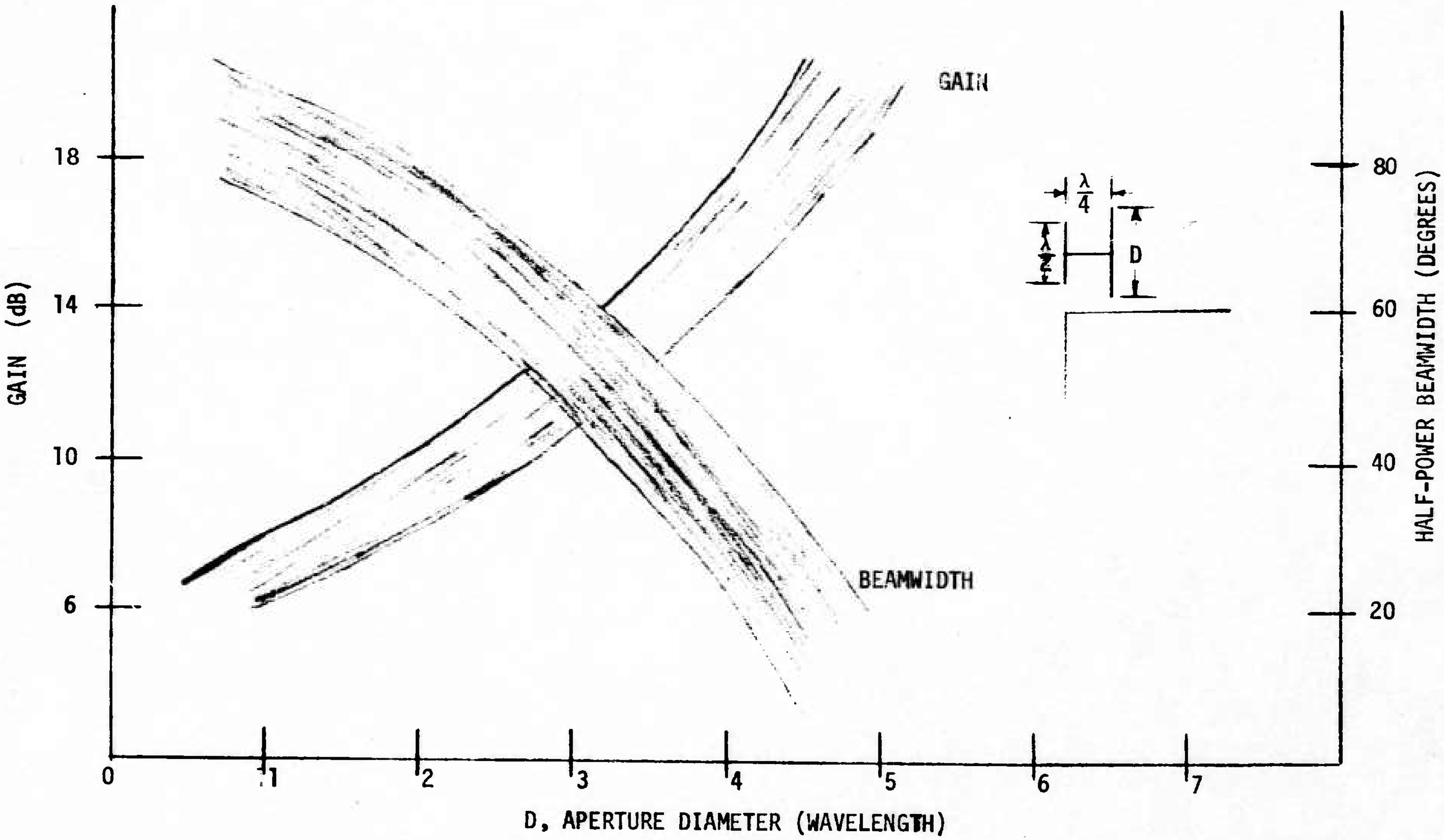
The disadvantage of voltage regulation may be avoided for the +5 volt supply providing power to the digital circuits and +15 volt supply providing power to the power amplifier, and since the supply voltages are relatively low, the most efficient source is one that uses the stacking arrangement. Using Lithium cells (highest energy density of any available electrochemical system) the power system could appear as shown in Figure 3.3.1. Two separate supplies are shown. One power supply provides +19.6 volts for the power amplifier and +15 volts regulated for analog circuitry. As the nominal voltage of Lithium cells is 2.8 volts, 7 cells of 3.0 AH capacity would be required. The second power supply would provide +5.6 volts for the digital circuitry and probably not need regulation. Two 5.8 AH cells would suffice. The estimated size of this approach to a power system using Lithium cells is 13 cubic inches and weighing about 13 ounces. This supply system has an effective energy density of 112 watt-hours/lb or 7 watt-hours/in^3.

4.4 Antenna Description

The antenna for the reduced-size packet radio must be compatible with package size of the unit. It is desirable for the antenna system to provide as much gain as possible, consistent with the constraint that size of the antenna system more than doubles for every 3 dB increase in antenna gain. Omnidirectional and directional antennas may be considered for a small packet radio. Directional antennas could provide increased gain in a given direction to offset the reduced range of a small packet radio.

For a vertically polarized omnidirectional response, a quarter-wave monopole (about 2" at 1800 MHz) would provide about 1.5 dB gain above an isotropic source. A single half-wave dipole would allow a little better than 2 dBi power gain but would stand 9 inches (including RF choke). The omnidirectional gain

FIGURE 4.4.1 ELECTRICAL CHARACTERISTICS OF HALF-WAVE DIPOLE IN FRONT OF GROUND PLANE



could be increased by stacking dipoles but the size of the antenna begins to grow very fast. For example, two colinear dipoles would give 5 dBi gain while standing 15 inches or more.

The most practical approach to a directional antenna for a small packet radio appears to be a half-wave dipole in front of a ground plane. Figure 4.4.1 illustrates the gain and half-power beamwidth characteristics as a function of aperture diameter (effectively the ground plane size) of such an antenna. Better than 6 dBi gain can be realized for an aperture diameter of one wavelength. An acceptable beamwidth of better than 70 degrees can also be obtained. If the ground plane diameter is reduced to the order of a half-wave length, the gain is not seriously degraded; however, the front-to-back lobe ratio deteriorates.

For general use in a mobile environment, an omnidirectional antenna is recommended due to the random nature of the arrival angle of the received signal; however, for a hand held application not requiring operation on a moving platform, a directional antenna compatible with small size can be used to extend the operational range of the unit.

4.5 Input/Output Device Description

The demands on the input/output capability of a small packet radio intended for terminal operation could encompass a broad range of desirable features. That is, the terminal device provides the interface between the user and the packet radio network. A minimum requirement on the input/output capability of a very small packet radio would be a hand calculator-type keyboard input and a display with reasonable character capacity. Clearly, the keyboard should allow input of all alphanumeric characters in addition to specific operation and instructions to the packet radio.

The display for packet radio should, as a minimum requirement, have the capability to provide a visual representation of all alphanumeric characters. The display must necessarily be small, lightweight, and consume small amounts of power. Although considerable development is underway in various display technologies, only light emitting diodes (LED's) and liquid crystal displays (LCD's) appear to have application to packet radio in the near term.

LED's are available in several different colors (red, orange, yellow, and green); doubly doped junctions can emit a variable spectrum depending on the level of the drive current. In addition, LED's have microsecond response times, long life, and operate over wide temperature ranges. The one major disadvantage of LED displays is power consumption requirements for portable applications. Although improvements in efficiency and lower drive currents (from a new milliamperes down to 0.1 to 0.5 ma per segment) are making LED's more attractive to low power applications [1], their power requirements may still exclude them from use in very small packet radios.

The recent announcement by Hewlett Packard of success in fabricating half inch square monolithic arrays of junction-isolated Gallium-Arsenide Phosphide diodes is a development of interest [2]. Each chip contains a 30 X 36 array of LED's. Assembling four chips edge-to-edge forms a 1/2" X 2" display. The complete panel can continuously display 36 alphanumeric characters. Consisting of 4328 diodes, the display strobed a 60 hertz rate, a column at a time has a maximum dissipation of 2 watts. A display appropriate for packet radio 1/2" X 4", would, therefore, consume 4 watts.

Liquid crystal displays are passive; that is, they are light-modulating devices and must depend on ambient light or an external light source for their operation. Liquid crystals are organic compounds (usually a mixture of materials called Butylonilines) whose molecules are usually long and rod-like in shape and possess, in addition to the solid and isotropic liquid phases of normal liquids, a third phase which occurs at temperatures between those associated with the solid and isotropic liquid phases. In this intermediate phase, the liquid crystal flows like a liquid, but has the long-range properties of a solid. The liquid crystal display is formed by confining the liquid crystal between two glass plates. The normal liquid crystal thickness is about 10 to 20 μm.

The single most attractive feature of liquid crystal displays is their low power consumption. Typically, the dynamic scattering LCD dissipates 0.1 to 1 mw/cm^2 at drive voltage of 15 volts at 30 Hz [3]. The twisted-neumatic LCD consumes even less, about 1 μw/cm^2. An eighty-character display consisting of 2 lines of 40 characters would consume less than 100 mw of power [4].

Limitations of LCD's include relatively short life, slow response times, narrow temperature range and the need for a pure AC drive signal. LCD lifetime has shown recent improvements, from about 15,000 hours up to about 50,000 hours [1]. An average DC component shortens the life of an LCD. LCD's exhibit response times from 100 to 200 ms. Since the response time is related to the material viscosity, response time is a strong function of temperature. The slow response times make multiplexing LCD's very difficult. The operating temperature range is the range of temperatures over which the liquid crystal material remains an anisotropic liquid. This range is normally 0-70°C.

The 80-character display referred to above would be expected to have dimensions 1" X 4" and weigh less than 6 ounces.

4.6 Summary

To summarize the description of the hand-held packet radio device, Table 4.6.1 shows the cubic inch volume and average power requirements by function in the packet radio.

Figure 4.6.1 illustrates an artist's conception of the device. The projected weight would be between three and four pounds with the weight concentrated in the power source attached to the end of the unit. This exercise demonstrates that a reasonable solution to the realization of a hand-held packet radio of the current design is feasible and attainable with today's technology, with promise for further improvement tomorrow.

TABLE 4.6 1 ESTIMATED VOLUME AND POWER REQUIREMENTS

Component	Size (Cubic Inches)	Power (Watts)
Front end filter	0.16	-
T/R Switches (3)	0.18	0.23
Power chain	0.50	12.70*
Rx RF amp	0.10	0.15
Mixer	0.65	-
RF YIG	0.45	-
LO YIG	0.45	-
LO system	0.45	0.42
TCXO (LO)	0.50	.05
Rx Matched filter	0.50	-
Non-Coherent AGC	0.75	0.50
Tx IF amp	0.05	0.07*
Encoder/Modulator	0.50	0.50*
Data detector	0.10	0.10
Bit Sync	0.10	0.50
Digital section	2.0	1.0
a) Microprocessor		
b) 4k RAM plus 1k ROM		
c) Address decode		
d) Timing control		
e) I/O channel		
Keyboard	4.0	-
Display	2.0	0.10
Power Supply	13.0	-
Total:	25.84	3.05 16.22*

*Peak power

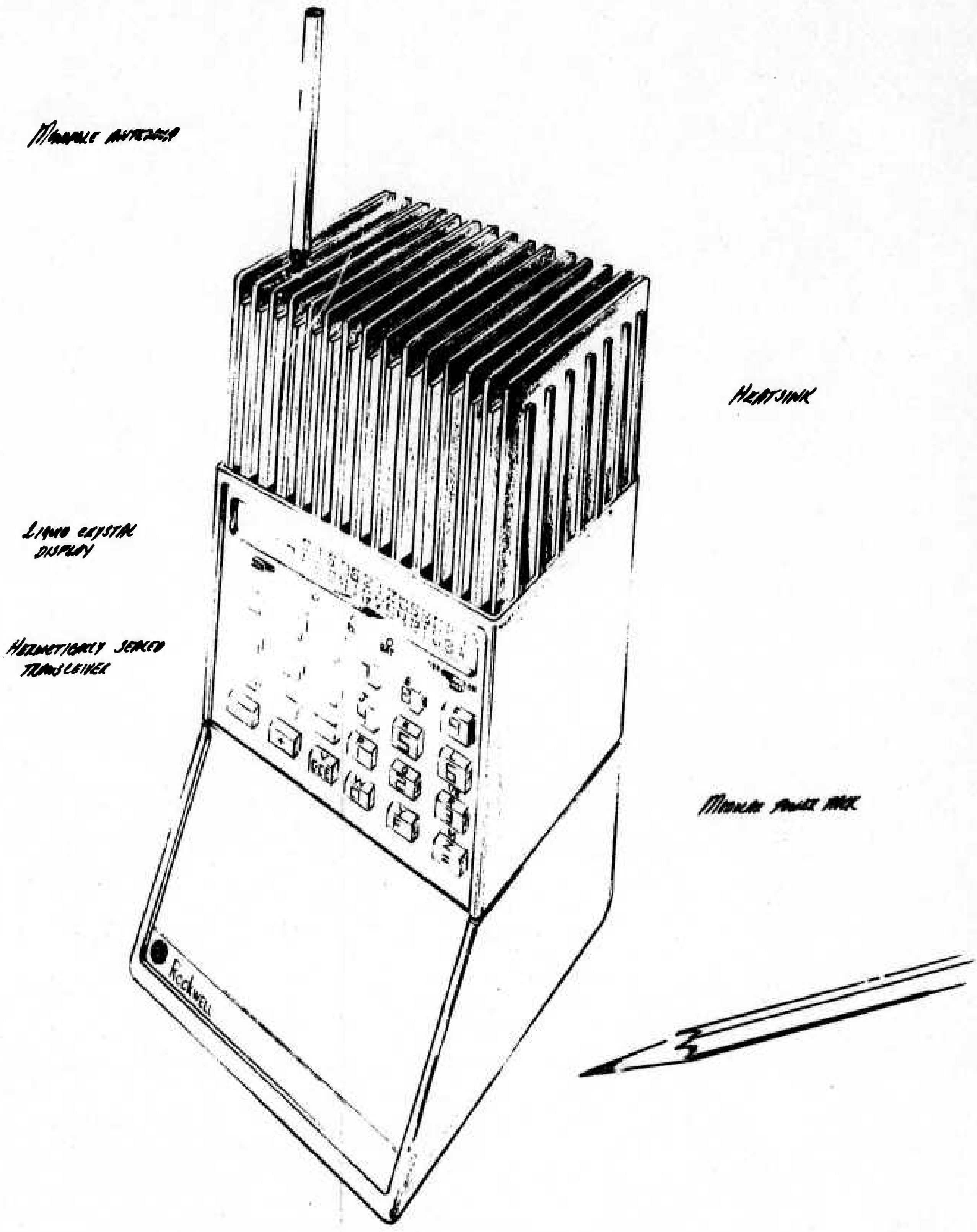


FIGURE 4.6.1 ARTIST'S CONCEPTION OF HAND-HELD PACKET RADIO

REFERENCES SECTION 4.4

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Unclassified
Security Classification

DOCUMENT CONTROL DATA - R & D

(Security classification of title, body of abstract and indexing annotation must be entered when the overall report is classified)

1. ORIGINATING ACTIVITY (Corporate author) Collins Radio Group Rockwell International	2a. REPORT SECURITY CLASSIFICATION Unclassified 2b. GROUP None
3. REPORT TITLE Quarterly Technical Report for the Project "Packet Radio Communications"	
4. DESCRIPTIVE NOTES (Type of report and inclusive dates) Quarterly Technical Report	
5. AUTHOR(S) (First name, middle initial, last name) Collins Radio Group Rockwell International	

6. REPORT DATE 1 May 1976	7a. TOTAL NO. OF PAGES 108	7b. NO. OF REFS 59
8a. CONTRACT OR GRANT NO. DAHC15-73-C-0192 b. PROJECT NO. c. d.	9a. ORIGINATOR'S REPORT NUMBER(S) 523-0602169-001C3L 9b. OTHER REPORT NO(S) (Any other numbers that may be assigned this report)	
10. DISTRIBUTION STATEMENT Unlimited		
11. SUPPLEMENTARY NOTES	12. SPONSORING MILITARY ACTIVITY Advanced Research Projects Agency Department of Defense	

13. ABSTRACT

The results of an investigation of size reduction of the packet radio are present in this document. This results include:

1) The identification of the limitations and constraints to size reduction of the packet radio,

2) Solutions and alternatives to these constraints, and

3) Design exercise for a small packet radio.

DD FORM 1 NOV 65 1473

Unclassified
Security Classification

Unclassified

Security Classification

14 KEY WORDS	LINK A		LINK B		LINK C	
	ROLE	WT	ROLE	WT	ROLE	WT
1. Packet radio communications						
2. Surface wave devices						
3. Hand-held terminal						
4. Charge-coupled devices						
5. Matched filters						
6. Frequency Synthesis						
7. RF power amplification						
8. Small energy sources						

Security Classification