INTERFERENCE REDUCTION TECHNIQUES EMPLOYING ACTIVE DEVICES

Hugh W. Denny
Robert A. Byers
Carl R. Driskell
Charles S. Wilson
Georgia Institute of Technology

TECHNICAL REPORT NO. RADC—TR—68—8
February 1968

This document has been approved for public release and sale; its distribution is unlimited.

NOTICE
This document is not to be used by anyone.

Prior to 4/18 1970 without permission of the Research Sponsor and the Experiment Station Security Office.

Rome Air Development Center
Air Force Systems Command
Griffiss Air Force Base, New York
When US Government drawings, specifications, or other data are used for any purpose other than a definitely related government procurement operation, the government thereby incurs no responsibility nor any obligation whatsoever; and the fact that the government may have formulated, furnished, or in any way supplied the said drawings, specifications, or other data is not to be regarded, by implication or otherwise, as in any manner licensing the holder or any other person or corporation, or conveying any rights or permission to manufacturer, use, or sell any patented invention that may in any way be related thereto.

Do not return this copy. Retain or destroy.
INTERFERENCE REDUCTION TECHNIQUES EMPLOYING ACTIVE DEVICES

Hugh W. Denny
Robert A. Byers
Carl R. Driskell
Charles S. Wilson

Georgia Institute of Technology

This document has been approved for public release and sale; its distribution is unlimited.
FOREWORD

This final report was prepared by Hugh W. Denny, Robert A. Byers, Carl R. Driskell and Charles S. Wilson of Georgia Institute of Technology, Atlanta, Georgia, under Contract F30602-67-C-0066, project number 4540, task number 454003. Reporting period covered 1 December 1966 to 30 November 1967. RADC project engineer is George A. Long (EMCVI-2).

This report has been reviewed and is approved.

Approved:

GEORGE A. LONG
Interf Analysis & Control Sec
Vulnerability Reduction Br

Approved:

RICHARD M. COSEL
Colonel, USAF
Chief, Communications Division

FOR THE COMMANDER:

IRVING J. GABELMAN
Chief, Advanced Studies Group
ABSTRACT

This report discusses the development and performance of several techniques employing active devices for the reduction of adjacent and co-channel interference in receivers. For operational situations where the interfering source is co-sited with the receiver, an automatic phase control system is described which detects the differential phase shift between the interference path and the direct path from the source and supplies a phase corrected signal for cancellation of the interfering signal. For the more general situation where the interference is not a cosite source, a feed forward system and a dual loop, phase locked system were developed. For 60 kHz spaced signals at 300 MHz, the feed forward system typically provides 55 dB suppression to CW interference and 40 dB suppression to AM interference. The dual loop cancellation system, which generates an auxiliary signal to cancel the interfering signal, is shown to be capable of suppressing an interfering signal by 50 dB when the interference signal is greater than an audio bandwidth from the desired signal. The improved performance of passive preselectors that can be obtained through Q multiplication is demonstrated with a Q multiplier device for the UHF region. Effective Q's of approximately 7,000 with relatively small coaxial cavities are realized with this device. Active and passive circuit configurations which incorporate the extremely narrow passbands of quartz crystals are discussed. When used as band stop filters, interference rejection levels greater than 50 dB are possible with these filters.
# TABLE OF CONTENTS

<table>
<thead>
<tr>
<th>SECTION</th>
<th>PAGE</th>
</tr>
</thead>
<tbody>
<tr>
<td>I. INTRODUCTION</td>
<td>1</td>
</tr>
<tr>
<td>II. INTERFERENCE CANCELLATION</td>
<td></td>
</tr>
<tr>
<td>A. Automatic Phase Control System</td>
<td>4</td>
</tr>
<tr>
<td>B. Feed Forward System</td>
<td>10</td>
</tr>
<tr>
<td>C. Dual Loop AM System</td>
<td>17</td>
</tr>
<tr>
<td>III. UHF Q MULTIPLICATION</td>
<td>20</td>
</tr>
<tr>
<td>IV. CRYSTAL INTERFERENCE FILTERS</td>
<td>44</td>
</tr>
<tr>
<td>V. CONCLUSIONS</td>
<td>63</td>
</tr>
<tr>
<td>VI. REFERENCES</td>
<td>65</td>
</tr>
<tr>
<td>VII. SCHEMATIC DIAGRAMS</td>
<td>66</td>
</tr>
</tbody>
</table>
**LIST OF FIGURES**

<table>
<thead>
<tr>
<th>FIGURE</th>
<th>PAGE</th>
</tr>
</thead>
<tbody>
<tr>
<td>1. Illustration of Typical Adjacent Channel Interference Situation</td>
<td>2</td>
</tr>
<tr>
<td>2. Vector Diagram Illustrating the Principles of the RF Cancellation Technique</td>
<td>5</td>
</tr>
<tr>
<td>3. Block Diagram of RF Cancellation Technique for Co-sited Transmitter-Receiver Pairs</td>
<td>6</td>
</tr>
<tr>
<td>4. Block Diagram of Automatic Phase Control System</td>
<td>7</td>
</tr>
<tr>
<td>5. Photograph of Automatic Phase Control Device</td>
<td>9</td>
</tr>
<tr>
<td>6. Simplified Block Diagram of the Feed Forward Cancellation Technique</td>
<td>11</td>
</tr>
<tr>
<td>7. Cancellation of a CW Signal</td>
<td>13</td>
</tr>
<tr>
<td>8. Cancellation of an AM Signal</td>
<td>14</td>
</tr>
<tr>
<td>9. Cancellation of a Closely Spaced CW Interfering Signal</td>
<td>15</td>
</tr>
<tr>
<td>10. Cancellation of a Closely Spaced AM Interfering Signal</td>
<td>16</td>
</tr>
<tr>
<td>11. Block Diagram of the Dual Loop AM Cancellation System</td>
<td>18</td>
</tr>
<tr>
<td>12. Block Diagram of UHF Q Multiplier</td>
<td>21</td>
</tr>
<tr>
<td>13. Multiplication Factor as a Function of the Midband Loop Gain</td>
<td>25</td>
</tr>
<tr>
<td>14. The Normalized Bandwidth of the Q Multiplier as a Function of the Midband Loop Gain</td>
<td>27</td>
</tr>
<tr>
<td>15. Pole-Zero Plot of a Simple All-Pass Network Function</td>
<td>30</td>
</tr>
<tr>
<td>16. Pole-Zero Plot of An Ideal Phase Shifter</td>
<td>30</td>
</tr>
<tr>
<td>17. Block Diagram of An Ideal Phase Shifter</td>
<td>33</td>
</tr>
<tr>
<td>18. Block Diagram of Voltage Controlled Phase Shifter</td>
<td>33</td>
</tr>
<tr>
<td>19. Phase Shift Versus Control Voltage at Selected Frequencies Between 275 and 350 MHz</td>
<td>34</td>
</tr>
<tr>
<td>FIGURE</td>
<td>PAGE</td>
</tr>
<tr>
<td>--------</td>
<td>------</td>
</tr>
<tr>
<td>20. Expanded Block Diagram of the UHF Q Multiplier</td>
<td>35</td>
</tr>
<tr>
<td>21. Photograph of the UHF Q Multiplier</td>
<td>36</td>
</tr>
<tr>
<td>22. Response Characteristics of a Coaxial Cavity With and Without Q Multiplication</td>
<td>37</td>
</tr>
<tr>
<td>23. Transmission Characteristics of the Combination of a High Q UHF Filter and a Small Cavity of Multiplied Q</td>
<td>38</td>
</tr>
<tr>
<td>24. Comparative Levels of Two Closely Spaced Signals at 325 MHz Before and After Filtering With the Combined UHF Filter and Q Multiplier</td>
<td>39</td>
</tr>
<tr>
<td>25. Relative Suppression of An Undesired Signal Spaced 1.5 MHz from the Desired Signal at 325 MHz</td>
<td>40</td>
</tr>
<tr>
<td>26. Typical Intermodulation Generation in the Q Multiplier</td>
<td>42</td>
</tr>
<tr>
<td>27. Audio Output Spectral Displays which Show the Effectiveness of the Q Multiplier in Reducing Cross Modulation</td>
<td>43</td>
</tr>
<tr>
<td>28. Rejection Characteristics of Simple Notch Filter</td>
<td>45</td>
</tr>
<tr>
<td>29. Rejection Characteristics of a 100 MHz Bridge Notch Filter</td>
<td>47</td>
</tr>
<tr>
<td>30. Rejection Characteristics of a 220 MHz Bridge Notch Filter</td>
<td>48</td>
</tr>
<tr>
<td>31. Rejection Characteristics of Hybrid Notch Filter</td>
<td>50</td>
</tr>
<tr>
<td>32. Rejection Characteristics of Low Pass Filter</td>
<td>52</td>
</tr>
<tr>
<td>33. Rejection Characteristics High Pass Filter</td>
<td>53</td>
</tr>
<tr>
<td>34. Rejection Characteristics of Two Cyrstals in a Simple Notch Filter</td>
<td>55</td>
</tr>
<tr>
<td>35. Block Diagram of Negative Resistance Circuit</td>
<td>56</td>
</tr>
<tr>
<td>36. Effects of Negative Resistance on a Crystal Notch Filter Response</td>
<td>58</td>
</tr>
<tr>
<td>37. Q Multiplication of a Crystal Response</td>
<td>59</td>
</tr>
<tr>
<td>FIGURE</td>
<td>PAGE</td>
</tr>
<tr>
<td>--------</td>
<td>------</td>
</tr>
<tr>
<td>38.</td>
<td>Block Diagram of a Crystal Controlled Impedance</td>
</tr>
<tr>
<td>39.</td>
<td>Response of a Hybrid Filter With a Crystal Controlled Impedance</td>
</tr>
</tbody>
</table>
EVALUATION

The objective of this effort was to develop specialized amplifiers using such techniques as Q multiplication and negative feedback which produce extremely narrow bandpass and band reject filters. These filters are to be used in solving specific interference problems especially those in the UHF region from 225 MHz to 400 MHz and more specifically adjacent channel interference due to too close spacing of transceiver frequencies.

The feasibility of Q multiplication at UHF has been demonstrated in this contract. However the model developed requires considerable improvement in power handling capability and noise figure before becoming a practical device.

The rejection capabilities of crystal notch filters have been thoroughly explored and, unless better crystals are produced, further reduction of their insertion loss is highly improbable.

To summarize, some very useful techniques in Q multiplication, phase shifting, and interference cancellation have resulted which can be utilized in specific interference situations.

GEORGE Ai LONG
Interf Anal & Control Sec
Vulnerability Reduction Branch
SECTION I
INTRODUCTION

The rapidly expanding demand for more communication channels in a fixed amount of available spectrum space has resulted in the assignment of communications channels with very little frequency separation. At the same time, the advancing state-of-the-art in receiver design has continued to lower the minimum signal level that can be detected. This combination of close channel spacing and increased sensitivity has created serious adjacent channel interference problems.

An insight into the nature of the adjacent channel interference problem can be gained by the consideration of a typical operational situation. In the UHF region of 200 to 400 MHz, transmitters capable of producing output powers of 200 watts (+53 dBm) or more are quite common and typical receivers in this frequency range have sensitivities of -100 dBm. An operational situation might be that as illustrated in figure 1, where a receiver is tuned to a relatively weak (-90 dBm) signal at 201 MHz while the strong undesired signal is at 200 MHz. Operational conditions often require that transmitters and receivers share the same antenna. In that case, if a 10 dB signal-to-noise ratio is required in the receiver's rf stage to prevent excessive information loss, then the interfering signal must be attenuated 153 dB to -100 dBm. The typical receiver preselector can be expected to provide approximately 50 dB of attenuation to the undesired signal and an additional 20 dB of attenuation is typically supplied by antenna multicouplers. The remaining 83 dB of attenuation necessary to reduce the +33 dBm signal out of the multcoupler to the -50 dBm permitted at the receiver's antenna terminals requires a filter technique exhibiting an effective attenuation slope of 16,600 dB per octave for an equivalent low-pass or high-pass configuration.

In the past, selective filters, employing only passive elements, have been effective in reducing or eliminating the undesirable effects of large adjacent channel signals. However, in some situations, such as that discussed above, passive selective devices are often unable to provide the required rejection to the interfering signal. In these situations, other techniques must be sought.

This report discusses the findings of a study program which emphasized the application of active devices and techniques to the reduction of adjacent channel interference. The general areas of investigation were:

1) cancellation of the undesired signal;
2) multiplication of the Q of passive devices; and
3) the application of quartz resonators as interference suppression filters.

The suppression of a strong CW signal by subtracting a phase locked auxiliary signal was demonstrated under a previous contract.\(^1\) Efforts to
Figure 1. Illustration of Typical Adjacent Channel Interference Situation.
adapt this device to the suppression of AM signals were successful. The performances of an automatic phase control system and a feed forward system add considerable impetus to the active cancellation approach for the suppression of AM signals in the 225 to 400 MHz region.

At low frequencies, particularly in the audio range, the application of active devices to counter the inherent circuit losses associated with passive devices has permitted a high degree of selectivity to be obtained in resonant structures. This principle of Q multiplication was successfully extended to the UHF region where extremely high Q responses with moderately sized passive devices were obtained.

Because of their very narrow passbands, quartz crystal resonators present an attractive approach to the problem of differentiating between two very closely spaced signals. Several active and passive circuit configurations incorporating crystal resonators were evaluated for interference suppression capabilities.

The description and evaluation of these techniques for the suppression of adjacent channel interference is contained in this report.
SECTION II
INTERFERENCE CANCELLATION

The adjacent channel interference problem may be conveniently represented by the vector diagram of figure 2. A low level desired signal, represented by the short vector, \( S_1 \) is effectively masked by the high level undesired signal which is represented by the longer vector, \( S_2 \). Although the two signals are close in frequency, they are not at the same frequency. Consequently, the angular velocities of rotation of the two vectors are not equal. Cancellation of the undesired signal may be obtained by combining an auxiliary signal, represented by the dotted vector, \( S_3 \), with the original two signals.

The auxiliary signal must identically match the undesired signal in amplitude and frequency and the phase angle between the two must be maintained at \( \pi \) radians.

Three possible cancellation techniques were investigated. The characteristics of the operational situation will determine which of these approaches is most appropriate for a specific interference problem.

A. Automatic Phase Control System

If the interfering transmitter or antenna is physically located nearby, a sample of the interference signal may be obtained and directed to the receiver in the manner shown in figure 3. Referring to the vector diagram of figure 2, the solid vector, \( S_2 \), represents the interference signal that reaches the receiver through the antenna lead, and the dotted vector, \( S_3 \), represents the signal on the auxiliary path.

Appropriate phase and amplitude corrections must be performed on the signal sample for cancellation of the undesired signal before entering the receiver front end. The angular frequencies, \( \omega_2 \) and \( \omega_3 \), of \( S_2 \) and \( S_3 \) are identical but the phase shift through the two paths will not necessarily be the same. For cancellation, a net difference of 180 degrees must exist in the phase shift through the two paths. A simple type of phase shifter such as an appropriate length of transmission line may be used to adjust the path length so that \( \theta = \pi \) radians. However, drift in the transmitter frequency or environmental effects such as temperature and humidity which detune the multicouplers slightly can result in a significant change in the phase of this signal at the receiver terminals. To maintain the proper phase relationship, a voltage controlled phase shifter with automatic tracking offers an attractive solution. The block diagram of a system to perform this function is shown in figure 4. A portion of the interfering transmitter signal sample is fed to a mixer and heterodyned with a local oscillator to produce a 30 MHz IF signal. A portion of the signal-plus-interference is coupled to a second mixer where it is heterodyned with the local oscillator to produce another
Figure 2. Vector Diagram Illustrating the Principles of the RF Cancellation Technique.
Figure 3. Block Diagram of RF Cancellation Technique for Co-sited Transmitter-Receiver Pairs.
Figure 4. Block Diagram of Automatic Phase Control System.
The 30 MHz IF signal. The two 30 MHz signals are amplified, compared in a phase detector and the resulting phase error voltage is used to adjust the phase shifter. A manual shifter acts as a coarse phase control and is adjusted to obtain the phase required for initial cancellation of the interfering signal. The voltage controlled phase shifter corrects for incremental changes to maintain the required phase shift.

Initially, a 200 to 400 MHz transistorized local oscillator was employed in a configuration which directed portions of the local oscillator signal through a broad band hybrid to the two 30 MHz IF subchassis for mixing. Unfortunately, this system exhibited poor local oscillator frequency stability. Poor stability of the local oscillator can cause an erroneous phase error voltage to be generated since any drift in the frequency of the local oscillator with respect to the RF input signal can produce a differential phase shift through the two IF amplifiers. The low output impedance of the local oscillator transistors excessively loaded the tuned circuit. The low impedance could not be properly matched over the octave bandwidth because of the wide variation in transistor parameters. Consequently, a Nuvistor vacuum tube version of the local oscillator was incorporated into the system and the free running frequency stability of the output signal was improved to approximately 5 parts in $10^6$ when averaged over a one minute time interval. The instabilities were further reduced by constructing the two balanced mixers on the local oscillator subchassis. By separating the mixers from the IF amplifiers, total 30 MHz gain is divided between two subchassis. An automatic frequency control (AFC) loop added further improvements to the local oscillator stability. A portion of the 30 MHz IF signal is heterodyned with a crystal controlled 24 MHz signal in the manner shown in figure 4. The resulting 6 MHz IF signal is fed to a limiter and a discriminator which provide an error signal to control the frequency of the local oscillator. With AFC applied, the AFC loop gain was sufficient to increase the stability to 5 parts in $10^8$ when locked to an input signal derived from a frequency standard. AFC between the local oscillator and the reference input signal was obtained at signal levels as low as 5μV.

The two 30 MHz IF amplifiers were constructed with transistors having forward AGC characteristics which permitted IF gain control through the use of a remote potentiometer. The net gain of the interference channel amplifier was variable from 15 dB to 38 dB, and the adjustable range of the signal-plus-interference channel amplifier gain was from 33 dB to 57 dB. Both channels exhibited similar bandwidths of 1.5 MHz.

Filtering of the phase detector output voltage was accomplished with a low-pass filter with a variable cutoff frequency. Selection of cutoff frequencies of 2.5, 10, 100 or 1000 Hz was provided through a front panel switch. The phase detector was followed by a operational amplifier with a maximum gain of 60 dB which allowed an increase of the maximum phase detector error voltage to ±10 Volts for an RF input signal of -85 dBm at the signal-plus-interference port.

The completed automatic phase control device for the RF cancellation system is shown in the photograph of figure 5.
Figure 5. Photograph of Automatic Phase Control Device.
B. Feed Forward System

In many operational situations, the interfering transmitter and susceptible receiver are not located at the same site. In these situations, the acquisition of a sample of the interfering signal at its source is either impossible or impractical. For such cases, a sample of the interfering signal must be obtained at the antenna, appropriately modified in an auxiliary feed forward path, and subsequently recombined with the original signal for cancellation prior to the receiver. A simplified block diagram of this type of system is shown in figure 6.

In the feed forward system, the difference in the phase shift around the sampling loop and the phase shift in the straight through path must be 180 degrees at the frequency of the interference signal but should be as small as possible (0 degrees, ideally) at the tuned frequency of the receiver. Since the two frequencies are close together, an abrupt phase change in the auxiliary path must occur over a narrow frequency range. The rate of phase shift directly determines the minimum frequency separation between the desired and interfering signals at which cancellation of the interference is possible. A cavity resonator is appropriate for approximating the desired phase response because it exhibits a rate of change of phase with frequency that is directly proportional to Q. Since high Q cavity resonators are readily available, they represent the most practical means for obtaining the required rapid phase shift.

The sampling loop cannot be tightly coupled into the main line without attenuating the desired signal unnecessarily. Consequently, amplification is necessary within the sampling loop to permit the adjustment of the amplitude of the cancellation signal to match the amplitude of the interference signal at the point of cancellation. The gain of the amplifier in the system is sufficient to overcome the attenuation through the two couplers plus any losses exhibited by the phase shifter.

With a broadband transistor amplifier in the system, greater than 60 dB suppression of a CW signal was obtained by cancellation. However, only about 40 dB of suppression of an AM signal was possible. The capability of the feed forward system to suppress an AM signal is directly related to the linearity of the transfer characteristics of the amplifier. The amplifier nonlinearities prevent the components of the amplified interference sample from exactly matching the components of the original signal and thus complete cancellation can not be obtained in practice.

The output level at which saturation of the amplifier occurred was approximately 0 dBm. With a 10 dB coupler as the summation device, the system is restricted to the cancellation of interfering signals at levels of less than -10 dBm. In an attempt to provide a higher power handling capability, another broadband amplifier was constructed with a transistor of higher collector dissipation in the output stage. In spite of the increased output capacity of the amplifier, the undistorted output power was not significantly improved over the initial unit.
Figure 6. Simplified Block Diagram of the Feed Forward Cancellation Technique.
An increased output capability with improved linearity was obtained with a Nuvistor vacuum tube amplifier. This amplifier was constructed as a relatively narrow bandwidth tuned amplifier. A tuned amplifier is not a significant limitation to the use of the system since there are other elements in the system which are tunable.

Gain adjustments were performed with a current controlled attenuator. This type of attenuator is attractive in that it can be constructed in a small physical configuration and can be continuously varied over a wide range. A network configuration of PIN diodes produced the desired attenuation characteristic.

Prototype models of 10 dB directional couplers were constructed from published design data. Directional couplers are used as signal samplers in preference to other techniques, such as resistive coupling, because the directivity of the couplers attenuates feedback components which tend to cause oscillations around the loop.

The suppression of a CW signal with the feed forward system is demonstrated in figure 7. Figure 7A shows the spectrum analyzer display of a 290 MHz signal through the system with the auxiliary path disconnected. Figure 7B shows that this signal can be suppressed greater than 60 dB with the system. The cancellation of AM signals is illustrated in figure 8. Figure 8B illustrates 55 dB of suppression of the AM signal shown in 8A.

Since this cancellation technique is primarily of interest as a possible solution to an adjacent channel interference problem, its ability to improve the reception of a weak signal in the presence of a strong undesired signal is important. Figure 9A depicts a situation where an undesired CW signal only 60 KHz away from a desired signal is 40 dB higher than the level of the desired signal. In figure 9B, the undesired signal is suppressed greater than 60 dB so that its level is approximately 15 dB less than the level of the desired signal. Although the desired signal is also attenuated by about 10 dB, relatively speaking, there is a 55 dB improvement in the ratio of the desired to undesired signal levels.

Figure 10 shows the relative suppression of an AM signal located only 60 KHz away from the desired signal. For these two signal tests, a cavity with a Q of 800 was used to permit the close spacing of 60 KHz without undue attenuation of the desired signal. The insertion loss of this cavity is about 6 dB. The cavity resonator used in the single signal tests had a lower Q of about 150 but its insertion loss was less than 1 dB. The greater insertion loss of the higher Q cavity created a higher gain requirement in the auxiliary loop than the Nuvistor amplifier was able to supply. A commercial power amplifier of higher gain was necessary for the two signal tests. The greater nonlinearities of this amplifier limited the degree of cancellation obtainable for an AM signal.
A. BEFORE CANCELLATION

B. AFTER CANCELLATION

CENTER FREQUENCY: 290 MHz
SPECTRUM WIDTH: 30 KHz/cm

Figure 7. Cancellation of a CW Signal.
A. BEFORE CANCELLATION

B. AFTER CANCELLATION

CENTER FREQUENCY: 290 MHz
SPECTRUM WIDTH: 30 KHz/cm

Figure 8. Cancellation of an AM Signal.
Figure 9. Cancellation of a Closely Spaced CW Interfering Signal.

A. BEFORE CANCELLATION

B. AFTER CANCELLATION

CENTER FREQUENCY: 290 MHz
Figure 10. Cancellation of a Closely Spaced AM Interfering Signal.
C. Dual Loop AM System

In both of the previous systems, no closed feedback loops for the adjustment of the amplitude of the cancellation signal are necessary because once initial adjustments are made and cancellation obtained any subsequent amplitude variations in the interfering signal source occur in the cancellation signal also. These systems are ultimately limited, however, either to some minimum frequency separation between the desired and undesired signals because of the finite phase slope obtainable with practical resonators or because an auxiliary signal cannot be obtained from the source.

In the absence of a signal sample from the source, the cancellation technique shown in the block diagram of figure 11 permits closer spacing of the desired and undesired signals than is permitted by the feed forward system. The allowed frequency spacing between the desired and undesired signals is determined in this dual closed loop system only by the characteristics of the phase lock loop and the relative amplitudes of the two signals.

In the system of figure 11, a cancellation signal is generated by controlling the frequency, phase and amplitude characteristics of an auxiliary oscillator so that they match those of the interference signal. A sample of the interference signal is obtained from the output side of the summation junction to provide the reference for the dual closed loop system. The phase of the interference sample is compared in a phase lock loop with the phase of the cancellation signal. The error voltage which is proportional to the phase difference between the two signals is used to correct the phase of the cancellation oscillator. In addition to the phase lock loop, a closed amplitude correction loop detects the amplitude of the interference signal. The level of the detected audio signal controls a high level amplitude modulator which adjusts the output of the cancellation oscillator to match the level of the interference signal at the summation junction.

The AM cancellation filter consists of two 200-400 MHz double conversion receivers sharing a common local oscillator. From the directional coupler, the RF input signal is fed to a cascode Nuvistor preselector which supplies approximately 15 dB or RF gain and establishes the system noise figure. The amplified RF signal and the cancellation signal are heterodyned in balanced mixers with the first local oscillator to produce 31 MHz first IF signals. Two 31 MHz amplifiers provide a gain of 58 dB with a nominal bandwidth of 1 MHz.

The two 31 MHz signals are heterodyned with a 24 MHz, crystal controlled, second local oscillator to produce 7 MHz second IF signals. An additional gain of 50 dB is supplied by second IF amplifiers over a bandwidth of 100 KHz. A portion of one of the 7 MHz output signals is used to develop an AFC voltage for stabilizing the frequency of the first local oscillator. An AFC voltage for the cancellation oscillator is obtained from the other 7 MHz signal. This signal is applied to a discriminator to develop an error signal which aids in the acquisition of phase lock of the cancellation signal to the interfering signal. The outputs from the two 7 MHz IF amplifiers are phase compared and the resultant phase error voltage is added to the AFC voltage and both are applied to the cancellation oscillator.
Figure 11. Block Diagram of the Dual Loop AM Cancellation System.
The adjustable phase shifter at the output of the 7 MHz IF amplifier in the cancellation oscillator loop has a range in excess of 360 degrees. Adjustment of the phase of the 7 MHz IF signal results in the phase of the cancellation oscillator being shifted by the same amount since the phase lock loop maintains a quadrature relationship between the two 7 MHz IF signals. The fixed 90 degree phase shift shown on the block diagram of figure 11 results in the 7 MHz inputs to the amplitude detector being in phase since the two phase detector signals are in quadrature.

The output of the amplitude detector is coupled to a DC amplifier which supplies the drive signal to the high level modulator in the output of the cancellation oscillator. The basic modulator consists of a balanced bridge configuration of HPA 3001 PIN diodes. The high on-off impedance ratio of these diodes permits greater than 98 per cent modulation of the cancellation signal. Since nonlinearities in the modulator generate unwanted components in the cancellation signal, a closed loop modulation system samples the output of the modulator and compares it with the characteristics of the audio signal which is derived from the interference sample. As a result, the level of the cancellation signal is made to follow the amplitude variations of the interfering signal.

To aid in the initial adjustment of the system, a detector and audio amplifier with speaker is used to monitor the level of the interference signal. A null in the audio output signal indicates when cancellation is achieved.

An attenuator is provided between the directional coupler and the pre-selector to prevent a very high level interfering signal from saturating the preselector. If the preselector saturates, the correct amplitude characteristics of the interfering signal can not be discerned. As suppression of the interfering signal is approached, the attenuation is decreased which permits a greater degree of cancellation to be obtained.

The operation of this cancellation filter in a breadboard configuration achieved approximately 50 dB suppression of an AM signal. During the performance of this test, a high degree of RF leakage from the cancellation oscillator into other parts of the system, particularly into the interference signal channel, was observed. A much higher degree of cancellation should be possible with a more adequately shielded configuration.
Although signal rejection techniques such as active cancellation have proven effective in reducing adjacent channel interference, these techniques are most appropriate for situations involving a small number of interfering signals. As the number of interference sources increase, the complexity of the cancellation system must grow accordingly. For those situations which involve several interfering signals, a simpler approach is to provide a sufficiently narrow band pass filter at the front end of the receiver. In this way, the desired signal is passed and all other undesired signals are attenuated.

One major difficulty encountered in the construction of passive element filters having the sufficiently small percentage bandwidths necessary for the rejection of undesired adjacent channel signals is the limited Q factor available in passive devices. This Q limitation has been overcome, in some instances, by the application of an active device connected in such a manner as to supply the inherent circuit losses associated with the passive circuit elements. This technique in effect multiplies the Q of the resonant circuit and produces a high degree of selectivity in the filter.

Basically, the technique of Q multiplication involves the controlled application of positive feedback around a resonant circuit. The block diagram of figure 12 illustrates the operational principles of a method which proved feasible at UHF.

In the course of the developmental program, certain features were found to be desirable in the various components of the system. For example, to avoid unpredictable loading effects on the antenna which would cause the input voltage, $e_i$, to fluctuate with adjustments in the system, a high degree of isolation was required between $e_i$ and the output voltage, $e_1$, of the summing junction and between $e_i$ and the feedback voltage, $e_4$. An active hybrid as the summing junction supplied the required isolation and in addition, provided some gain to both $e_i$ and $e_4$.

The basic resonant structure of figure 12 whose Q is to be multiplied is designated as $A(s)$, a second order response function. To minimize the order of multiplication required to achieve a specific Q with a small sized structure and thus avoid the inherent instabilities associated with high degrees of multiplication, a coaxial cavity having a loaded Q of approximately 800 was constructed. This high Q was obtained at the expense of a 6 dB insertion loss through the cavity. Sufficient gain was provided in the remainder of the system to overcome this loss.

The sampling ratio, $k$, was selected as approximately 0.1 which represents a compromise between the need to avoid excessive attenuation of $e_2$, the voltage
Figure 12. Block Diagram of UHF Q Multiplier.
output of $A(s)$, and the need to provide an adequate voltage sample, $e_3$.

$B(s)$ represents the variable gain and phase in the feedback path. The division of $B(s)$ into essentially independent phase and gain functions resulted in greater ease in tuning. A voltage controlled phase shifter that exhibited very little gain variation with phase adjustments supplied the required phase characteristic. The independent gain adjustment was obtained with the use of a current controlled attenuator that exhibited a low phase shift versus current characteristic. Because of its independent gain and phase characteristics, $B(s)$ can be considered as a gain function, $G(s)$, and an independent phase function, $\phi(s)$.

The behavior of the UHF Q multiplier is best explained in terms of its voltage transfer characteristic. To derive the voltage transfer characteristic of the system, first observe in figure 12 that

$$e_1 = a(e_i + e_4) \tag{1}$$

where $a$ is the voltage gain between each junction input and its output; and further, that

$$e_2 = A(s)e_1 \tag{2}$$

$$e_3 = k e_2 \tag{3}$$

$$e_4 = B(s) e_3 \tag{4}$$

and, finally, that the output voltage

$$e_0 = (1 - k)e_2 \tag{5}$$

Substituting the expression for $e_2$ from equation (2) into equation (5) gives

$$e_0 = (1 - k)A(s)e_1 \tag{6}$$

From equation (1), obtain the expression for $e_1$ and substitute in (6):

$$e_0 = (1 - k)A(s) \left[ a \ (e_i + e_4) \right] \tag{7}$$
Expansion of equation (7) and subsequent substitutions from equations (4) and (3), respectively, result in

\[ e_o = a(1 - k) A(s) e_1 + ak(1 - k) A(s) B(s) e_2 \]  \hspace{1cm} (8)

However, since

\[ e_o = \frac{1}{1 - k} e_o \]  \hspace{1cm} (9)

from equation (5), equation (8) may be written as

\[ e_o = a(1 - k) A(s) e_1 + ak A(s) B(s) e_o \]  \hspace{1cm} (10)

Rearrangement of the terms of (10) reveals that the voltage transfer, \( T(s) \), characteristic of the system is

\[ T(s) = \frac{e_o}{e_i} = \frac{a(1 - k) A(s)}{1 - ak A(s) B(s)} \]  \hspace{1cm} (11)

The transfer characteristic of a single cavity, coaxial resonator, tuned to \( f_0 = \frac{\omega_0}{2\pi} \) and characterized by \( Q_u \gg Q_L \), where \( Q_u \) is the unloaded \( Q \) and \( Q_L \) is the loaded \( Q \), is

\[ A(s) = \frac{\frac{\omega_0}{Q_u}}{s^2 + \frac{\omega_0}{Q_u} s + \omega_o^2} \]  \hspace{1cm} (12)

Substituting this expression for \( A(s) \) into equation (11) gives

\[ T(s) = a(1 - k) \frac{\frac{\omega_0}{Q_u} s}{s^2 + \left[1 - ak B(s)\right] \frac{\omega_0}{Q_u} s + \omega_o^2} \]  \hspace{1cm} (13)
which relates the overall transfer characteristic of the system to the basic characteristics of the resonator. For convenience, the loaded Q of the resonator is designated as simply "Q" in equation (13).

The multiplied Q of the response function of equation (13) can be defined as

$$ Q' = \frac{Q}{1 - ak B(s)} $$

(14)

Defining the ratio of the multiplied Q to the natural Q as the multiplication factor, $M_f$, gives

$$ M_f = \frac{Q'}{Q} = \frac{1}{1 - ak B(s)} $$

(15)

Since the behavior of the system is primarily of interest near the tuned frequency, $jw_o$ may be substituted for $s$ in equation (15). Equation (15) then becomes

$$ M_f = \frac{1}{1 - ak B(jw_o)} $$

(16)

For proper multiplication, the phase of $B(jw_o)$ is adjusted so that $B(jw_o) = + G(jw_o)$, where $G(jw_o)$ is the midband gain in the feedback loop. Figure 13 shows how the multiplication factor behaves as a function of the gain for selected values of the ak product. As would be expected, when the product $ak G(jw_o)$ approaches unity, the multiplication factor increases without bound or, in other words, the system approaches self sustained oscillations. Note that for a specific multiplication factor the slope of the curve is less at smaller ak values. Although more gain is required to produce a specific multiplication factor with smaller ak values, the system will be less sensitive to incremental gain variations in the less sloping regions of the curves.

Perhaps a more useful form for the multiplication factor is in terms of the change in bandwidth rather than in terms of Q. For example, suppose a cavity resonator with some natural Q is available. Then, the 3 dB bandwidth, $\Delta f$, at a particular frequency, $f$, is

$$ \Delta f = \frac{f}{Q} $$

(17)

If the Q of the resonator is multiplied, the modified bandwidth is
Figure 13. Multiplication Factor as a Function of the Midband Loop Gain.
If a normalized bandwidth, $\bar{B}$, is defined such that

$$\Delta f' = \frac{\Delta f}{M_f Q}$$

(18)

then it follows from (17) and (18) that

$$\bar{B} = \frac{\Delta f'}{\Delta f}$$

(19)

The normalized bandwidth as a function of the gain, $G(j\omega_0)$, is shown in figure 14 for selected $a_k$ values. These curves again emphasize the greater system stability which results from lower $a_k$ values because bandwidth variations are less for specific gain changes at the lower values of the $a_k$ product.

In actual practice, incremental changes in system gain and phase can be expected which will cause bandwidth variations. In order to separate the effects of changes in the magnitude and phase of the loop gain on the normalized bandwidth, $\bar{B}$, equation (20) can be written in the form

$$\bar{B} = 1 - ak B(j\omega)$$

(20)

This division of the feedback characteristic into separate gain and phase functions is realistic since they are essentially independent in the actual system. The variation of the normalized bandwidth due to changes in gain or phase that may be caused by environmental factors such as temperature, humidity, and others can now be described by differentiating (21) to obtain an expression for the total differential, $d\bar{B}$. Since bandwidth fluctuations as a function of environmental factors are of greatest concern at a fixed frequency of operation, the frequency dependence of the gain and phase may be ignored. Then $G(s)$ and $\phi(s)$ may be simply written as $G$ and $\phi$, respectively. Under these circumstances, performing the necessary differentiation gives

$$d\bar{B} = -ak \left[ Gd\phi + \phi dG \right]$$

(22)

To illustrate the significance of (22), suppose the phase function is adjusted so that $\phi = 1.0$ and suppose that $d\phi = 0$. Changes in $\bar{B}$ because of changes in
Figure 14. The Normalized Bandwidth of the Q Multiplier as a Function of the Midband Loop Gain.
The characteristics desired in the device which generates the $\phi(s)$ function are those possessed by an ideal phase shifter. An ideal phase shift network permits the phase of the signal through the network to vary without affecting the amplitude of the signal. An example of an ideal phase shifter is the variable length of transmission line which is commonly used in the VHF and UHF regions. A principle limitation to the use of this type of shifter in the Q multiplication device is the long lengths of line required in the 200 to 400 MHz range. For example, at 300 MHz, a length variation of 50 centimeters is necessary to obtain at least 180 degrees of phase shift.

Since one of the primary objectives in developing the Q multiplication technique is to realize large effective Q's in smaller physical systems, such long lengths of transmission lines are not desirable. Consequently, the development of a phase shifter more compatible with the overall Q multiplication concept was necessary.

A length of transmission line is an example of an all-pass network, that is, the amplitude of the transfer function is a non-varying function of frequency while the phase does vary as a function of frequency. The pole-zero (p-z) plot of an elementary all pass function is shown in figure 15. The transfer function represented by this p-z plot is

$$T(s) = \frac{s - z_1}{s - p_1} = \frac{s - \sigma_1}{s + \sigma_1} . \quad (24)$$

Evaluating (24) at $s = j\omega_1$, point A, yields

$$T(j\omega_1) = \frac{j\omega_1 - \sigma_1}{j\omega_1 + \sigma_1} . \quad (25)$$

The magnitude of the transfer function is
\[ |T(j\omega_1)| = \left| \frac{j\omega_1 - \sigma_1}{j\omega_1 + \sigma_1} \right| = \sqrt{\frac{\omega_1^2 + \sigma_1^2}{\omega_1^2 + \sigma_1^2}} = 1 \quad . \tag{26} \]

Note that as point A moves to a new frequency, \( s = j\omega_2 \), the value of \( |T(j\omega)| \) remains unity.

The net phase shift, \( \theta_T \), through the network of the signal at frequency \( \omega_1 \) is given by

\[ \theta_T = \theta(z_1) - \theta(p_1) = \pi - \alpha - \alpha = \pi - 2\alpha = \pi - 2 \tan^{-1} \frac{\omega_1}{\sigma_1} \quad . \tag{27} \]

It can be seen from equation (27) that as the frequency of point A changes from \( \omega_1 \) to another frequency \( \omega_2 \) the phase shift through the network changes.

The objective of a phase shifter, however, is to vary the net phase between two points in a transmission path at a fixed frequency (point A not moving as above) with minimum perturbation to the amplitude. If instead of the fixed p-z pair of figure 15, suppose a pair of complex poles and zeros move on a line parallel to the \( j\omega \) axis as shown in figure 16. The transfer function of this p-z function is

\[ T(s) = \frac{(s - z_1)(s - z_1^*)}{(s - p_1)(s - p_1^*)} = \frac{(s - \sigma_1 - j\omega_2)(s - \sigma_1 + j\omega_2)}{(s + \sigma_1 - j\omega_2)(s + \sigma_1 + j\omega_2)} \quad . \tag{28} \]

The amplitude of \( T(s) \) evaluated at \( s = j\omega_1 \) can be shown to be

\[ |T(j\omega)| = \frac{\sqrt{\sigma_1^2 + (\omega_1 + \omega_2)^2}}{\sqrt{\sigma_1^2 + (\omega_2 - \omega_1)^2}} \times \frac{\sqrt{\sigma_1^2 + (\omega_1 + \omega_2)^2}}{\sqrt{\sigma_1^2 + (\omega_2 - \omega_1)^2}} = 1 \quad . \tag{29} \]

Equation (29) shows that as the complex poles and zeros move up and down parallel to the \( j\omega \) axis the amplitude is invariant.

The net phase shift at point A is

\[ \theta_T = \theta(z_1) + \theta(z_1^*) - \theta(p_1) - \theta(p_1^*) \quad . \tag{30} \]
Figure 15. Pole-Zero Plot of a Simple All-Pass Network Function

Figure 16. Pole-Zero Plot of An Ideal Phase Shifter.
From equation (28) the expression for the total phase shift can be written as

$$\theta_T = \tan^{-1}\left(\frac{w_1 - w_2}{\sigma_1}\right) + \tan^{-1}\left(\frac{w_1 + w_2}{\sigma_1}\right) - \tan^{-1}\left(\frac{w_1 - w_2}{\sigma_1}\right) - \tan^{-1}\left(\frac{w_1 + w_2}{\sigma_1}\right).$$  \hspace{1cm} (31)

Equations (29) and (31) show that the p-z configuration of figure 16 represents the network function of an ideal phase shifter since the phase angle but not the amplitude of the transfer function is a function of frequency.

Another way of expressing equation (28) is

$$T(s) = \frac{s^2 - \frac{w_o}{Q} s + w_o^2}{s^2 + \frac{w_o}{Q} s + w_o^2}. \hspace{1cm} (32)$$

Being an improper fraction, $T(s)$ may be written as one plus a proper fraction:

$$T(s) = 1 + \frac{P(s)}{N(s)}, \hspace{1cm} (33)$$

or

$$T(s) - 1 = \frac{P(s)}{N(s)}. \hspace{1cm} (34)$$

Performing operation (34):

$$\frac{P(s)}{N(s)} = \frac{s^2 - \frac{w_o}{Q} s + w_o^2}{s^2 + \frac{w_o}{Q} s + w_o^2} - 1 = -\frac{2}{Q} \frac{w_o}{s}. \hspace{1cm} (35)$$

The form of the expression (35) for $P(s)/N(s)$ is immediately recognized as the second order response such as that obtained with a conventional LC circuit. Consequently, this analysis indicates that a practical phase shifter should
be obtainable with a proper combination of relatively simple networks such as that shown in figure 17.

Figure 18 is a detailed block diagram of a voltage controlled phase shifter which was constructed for the frequency range of 300 to 350 MHz. The input signal is divided into two paths of equal amplitude with the power divider. Channel one consists of a wideband amplifier with appropriate resistive padding to provide constant gain over the desired frequency range. Channel two contains a voltage controlled, single tuned amplifier with appropriate buffers to stabilize the input and output impedance of the channel as the tuning is varied. From equation (35) the net gain at resonance through the tuned amplifier channel must be twice that of the other channel and the net phase difference between the two must be 180 degrees. Consequently, the net gain of channel two is adjusted to be twice the gain of channel one. To insure that 180 degrees of phase difference is maintained at the summing junction over the desired frequency range, a short piece of transmission line is incorporated in channel one to equalize the time delay through both channels. Figure 19 shows that this system permits greater than 180 degrees of phase adjustment over the frequency range from 275 to 350 MHz.

An expanded block diagram of the UHF Q multiplier is shown in figure 20.

The completed device is shown in the photograph of figure 21.

Figure 22 shows the response curve of an unmodified resonator with a natural Q of approximately 700 at 325 MHz. The response curve of the cavity with Q multiplication is also shown in figure 22. The 3 dB bandwidth of approximately 50 KHz represents an effective Q of 6500 or a multiplication of approximately 9.3. These curves show that the system not only narrows the bandwidth but also relatively increases the skirt attenuation to signals which are essentially out of band.

The attenuation as a function of bandwidth is demonstrated in figure 23 for four resonator combinations. The transmission characteristics of the cascaded combination of a Collins UHF Filter, Model 56C-2, and the Q multiplied cavity shows the narrower passband with improved skirt rejection that is obtained over just the Collins Filter alone.

The ability of the Q multiplier to effectively discriminate against one signal while allowing another closely spaced signal to pass is illustrated in figure 24. When the two equal amplitude signals shown in figure 24A are passed through the Q multiplier and the UHF filter, their relative amplitudes are as shown in figure 24B. One of the signals is attenuated approximately 40 dB while the other is not affected.

Figure 25A depicts a situation where an undesired signal, spaced 1.5 MHz away, is more than 40 dB stronger than the desired signal. Note that the desired signal can not be discerned from the spectrum analyzer noise. When the signals are passed through the UHF filter and the Q multiplier, the desired signal is amplified and the undesired signal is attenuated. As evident in figure 25B, the resulting desired signal level is 20 dB greater than the
Figure 17. Block Diagram of An Ideal Phase Shifter.

Figure 18. Block Diagram of Voltage Controlled Phase Shifter.
Figure 19. Phase Shift Versus Control Voltage at Selected Frequencies Between 275 and 350 MHz.
Figure 20. Expanded Block Diagram of the UHF Q Multiplier.
Figure 21. Photograph of the UHF Q Multiplier.
Figure 22. Response Characteristics of a Coaxial Cavity With and Without Q Multiplication.
Figure 23. Transmission Characteristics of the Combination of a High Q UHF Filter and a Small Cavity of Multiplied Q.
A. BEFORE FILTERING

B. AFTER FILTERING

Figure 24. Comparative Levels of Two Closely Spaced Signals at 325 MHz Before and After Filtering With the Combined UHF Filter and Q Multiplier.
Figure 25. Relative Suppression of An Undesired Signal Spaced 1.5 MHz from the Desired Signal at 325 MHz.
undesired signal level.

Being an active device, the Q multiplier might be expected to generate unwanted responses which could become a source of interference in themselves. Of particular concern are the third order intermodulation products between the desired and undesired signals which might be generated in the active hybrid summing junction. Intuitively, any third order intermodulation products should, at most, be no higher than the level of the interfering signal itself out of the multiplier because the multiplier's response characteristics should discriminate against the unwanted products as effectively as against the undesired signal. In fact, the typical level of intermodulation products generated in the system are as shown in figure 26 which shows that, for 160 KHz spacing between the two input signals, the third order product is better than 50 dB below the desired signal level.

Figure 27 permits an evaluation of the relative cross modulation properties of the system to be made. Figure 27A is a spectral display of the audio components in the output of a highly selective receiver (Singer Metrics, NF 105) when two signals with 250 KHz spacing at 325 MHz are applied to the receiver's input. The desired signal is amplitude modulated at 400 Hz and the undesired is modulated at 1000 Hz. The other audio components are generated by cross modulation within the receiver. With the UHF Filter tuned to the desired signal (400 Hz modulation) some attenuation of the 1000 Hz component is evident as shown in figure 27B. Some reduction in the level of the harmonics of 400 Hz also occur along with a decrease in the level of the cross modulation components. Figure 27C shows that filtering the desired signal through the Q multiplier effectively removes the undesired signal. Equally important, this display shows no evidence of cross modulation in the Q multiplication device.

Although no particular effort was applied to developing a low noise device, an input signal level of -85 dBm is sufficient to develop a 10 dB signal-to-noise ratio at the output of the Q multiplier.

These performance results verify that the technique of Q multiplication for the realization of narrow passband filters is feasible at UHF. The resulting device represents a practical approach to the reduction of adjacent channel interference in the 200 to 400 MHz range.
Figure 26. Typical Intermodulation Generation in the Q Multiplier.
RECEIVER: NF105, SINGER METRICS.
DESIRABLE SIGNAL:
$ f_o = 325$ MHz
400 Hz MODULATION
500 $\mu$V AT RECEIVER
UNDESIRABLE SIGNAL:
250 KHz FROM DESIRED
1000Hz MODULATION
500 $\mu$V AT RECEIVER

Figure 27. Audio Output Spectral Displays which Show the Effectiveness of the Q Multiplier in Reducing Cross Modulation.
SECTION IV
CRYSTAL INTERFERENCE FILTERS

Quartz crystal resonators are a class of passive devices having extremely narrow transmission bandwidths. The exceptionally high Q values exhibited by quartz crystals suggest the application of these devices to the solution of adjacent channel interference problems. However, crystal resonators suffer certain limitations which restrict their direct application to interference situations. For example, a single resonator cannot be used to cover a wide frequency range of situations because of the fixed frequency nature of the overtone responses. Furthermore, only limited use can be made of a single quartz resonator in a bandpass configuration because the extremely narrow passband may cause appreciable distortion when the desired signal drifts slightly in frequency. Normally, several resonators must be used to produce a filter with sufficient bandwidth to prevent distortion. In this regard, considerable success has been achieved in constructing multi-element bandpass filters in crystal lattice configurations below about 120 MHz. Unfortunately, these devices are not currently available in the 225 to 400 MHz frequency range.

While quartz resonators have only limited applications as bandpass filters at VHF, they may be used with fewer limitations in a band stop configuration. When used as a band reject filter to attenuate an interfering signal, small variations in the attenuation over the band occupied by the interfering signal or a slight drift in the interfering signal frequency are unimportant so long as the attenuation always exceeds the amount required to produce interference-free reception of the desired signal. In many instances, the required rejection can be supplied with a single resonator. With this advantage and the fact that a single resonator in a band reject configuration may be effectively operated at higher overtone frequencies (approaching the UHF region), the band reject configuration is more desirable than the bandpass configuration.

The schematic diagram of a simple bandpass filter which has a sharp rejection notch at the peak of the bandpass response is shown in figure 28. The rejection characteristics of this crystal notch filter at 100 MHz are also shown. While the peak of the response occurs at the parallel resonant frequency of L and \((C + C_o)\), where \(C_o\) is the crystal parasitic capacitance, the sharp notch occurs at the series resonant frequency of the crystal. The depth of the notch depends directly on the ratio of the series resonant resistance of the crystal to the parallel resonant impedance of L and \((C + C_o)\).

A primary limitation of this filter configuration is the large number of undesirable spurious responses. If the desired signal falls within the region of these spurious responses, it may be subject to severe distortion and attenuation. Therefore, a crystal free of spurious responses or a circuit configuration capable of suppressing the spurious responses is needed. Since the development of crystals free of spurious responses was outside the realm of this study, efforts were directed toward the development of circuit
Figure 28. Rejection Characteristics of Simple Notch Filter.
configurations which suppress spurious responses.

One configuration found to be quite effective in suppressing the undesired spurious responses is the balanced bridge. This configuration takes advantage of the fact that normally the series resistance at a spurious response is higher than the resistance at the main response. By balancing the bridge at the resistance of the main response, the effectiveness of the main response is enhanced and the effects of the spurious responses are degraded. The response characteristics of the bridge type notch filter which are shown in figure 29 reveal a significant reduction in the number and magnitude of the spurious responses as compared to those of the simple notch filter. An additional advantage is gained by the use of the bridge configuration in that the depth of the notch depends directly on the precision of the bridge balance rather than on the ratio of the crystal resistance to the parallel resonant impedance of the tuned circuit.

Figure 30 illustrates the performance of a balanced bridge filter near 220 MHz. The maximum response ripple in the passband, 3 kHz above and below the notch, is less than 2 dB. In this region, a desired signal would be subject to little attenuation from the notch filter. The passband insertion loss with this configuration at 220 MHz is on the order of 10 dB compared to 2 dB at 100 MHz. Unfortunately, the insertion loss of this filter further increased at the higher overtone frequencies as shown in Table I.

Table I

<table>
<thead>
<tr>
<th>Crystal Overtone</th>
<th>Frequency (MHz)</th>
<th>Insertion Loss (dB)</th>
<th>Rejection Level (dB)</th>
<th>Approximate 3 dB Bandwidth (kHz)</th>
<th>Approximate Q</th>
</tr>
</thead>
<tbody>
<tr>
<td>3rd</td>
<td>59.8583</td>
<td>2.8</td>
<td>48</td>
<td>8.8</td>
<td>6,800</td>
</tr>
<tr>
<td>5th</td>
<td>99.7747</td>
<td>2.0</td>
<td>46</td>
<td>9.6</td>
<td>10,400</td>
</tr>
<tr>
<td>7th</td>
<td>139.6858</td>
<td>3.0</td>
<td>44</td>
<td>8.8</td>
<td>15,900</td>
</tr>
<tr>
<td>9th</td>
<td>179.5951</td>
<td>5.2</td>
<td>42</td>
<td>11.0</td>
<td>16,300</td>
</tr>
<tr>
<td>11th</td>
<td>219.5019</td>
<td>9.0</td>
<td>41</td>
<td>14.9</td>
<td>14,700</td>
</tr>
<tr>
<td>13th</td>
<td>259.4062</td>
<td>14.0</td>
<td>37</td>
<td>16.3</td>
<td>15,900</td>
</tr>
<tr>
<td>15th</td>
<td>299.3070</td>
<td>20.0</td>
<td>30</td>
<td>26.3</td>
<td>11,400</td>
</tr>
<tr>
<td>17th</td>
<td>339.2261</td>
<td>26.0</td>
<td>&gt; 25</td>
<td>44.4</td>
<td>7,700</td>
</tr>
<tr>
<td>19th</td>
<td>379.1300</td>
<td>28.0</td>
<td>&gt; 23</td>
<td>57.6</td>
<td>6,600</td>
</tr>
</tbody>
</table>

46
Figure 29. Rejection Characteristics of a 100 MHz Bridge Notch Filter.
Figure 30. Rejection Characteristics of a 220 MHz Bridge Notch Filter.
While part of the increase in insertion loss may be attributed to the larger resonant resistance of the crystal at the overtone frequencies, a large part of the insertion loss was believed to be due to the increase in core loss and decrease in transformer coupling which occurs at the higher frequencies. A filter insertion loss of 18.5 dB at 380 MHz with the bridge unbalanced indicated that the transformers were contributing a large part of the loss. However, an insertion loss of 19.5 dB at 260 MHz with the original transformers replaced with low loss UHF transformers indicated that additional factors were responsible for the high value of loss.

Further study of the bridge configuration revealed that the insertion loss was related to the series resonant impedance of the crystal at the overtone response frequency. With a crystal in one arm of the bridge and an impedance in the other arm matched to the series resonant impedance of the crystal and the shunt capacity of the holder, cancellation of the signal in the frequency range of crystal resonance produces an effective notch filter. Off resonance, the crystal becomes a high impedance which unbalances the bridge. However, the resistance and capacitance necessary for balancing the crystal resonant impedance remain in series with the signal path and determine the passband insertion loss of the filter. In general, at the higher overtone responses, the series resonant impedance of the crystal increases and the result is that the overall insertion loss of the filter rises. Consequently, low insertion loss with the balanced bridge configuration depends upon low series impedance at the overtone responses. Crystal resonators with a sufficiently low series impedance at overtone responses up to 400 MHz were not available to permit any significant reduction in the passband insertion loss of this filter configuration.

Since broadband hybrid junctions were available that had lower loss characteristics in the VHF region than conventional transformers, a crystal notch filter was developed which employed a hybrid junction in the manner shown in figure 31. In the hybrid junction, a signal applied to the summing port is split between the two side arm ports. Any impedance mismatch at the side arms reflect the signals to the output difference port. With one side arm containing the crystal and the other side arm containing an impedance matched to the series resonant impedance of the crystal and the shunt capacity of the holder, cancellation of the signal at the crystal series resonant frequency produces a notch characteristic as shown by the response curve of figure 31. Off resonance, the high impedance of the crystal unbalances the hybrid and the amount of unbalance determines the insertion loss of the filter. Typically, this filter configuration exhibited a 3 dB lower insertion loss than the simple balanced bridge.

The depth of the notch obtainable with the balanced hybrid filter depends upon the precision of match between the side arm impedances. The passband insertion loss, however, depends upon the degree of mismatch between these impedances. Minimum insertion loss with maximum rejection at the notch frequency will, consequently, be obtained with a crystal unit of lowest series resonant impedance. The low balancing impedance required to produce the notch characteristic at crystal resonance then exhibits a large mismatch to the crystal off-resonance impedance which results in a lower insertion loss.
Figure 31. Rejection Characteristics of Hybrid Notch Filter.
By decreasing the potentiometer setting to minimum resistance, the passband insertion loss was reduced to 3.5 dB at the 220 MHz measurement frequency. Naturally, this procedure unbalanced the bridge and degraded the notch rejection value but it verified that a lower value of crystal resonant impedance is necessary to reduce the passband insertion loss of the notch filter configuration.

One limitation of crystal interference filters is the lack of flexibility because of their fixed tuned characteristics. A way of circumventing this limitation is to have available a large number of crystals with response frequencies appropriate to the interference environment. For maximum flexibility in meeting changes in the environment, the crystal elements should be interchangeable with minimum perturbation to the overall performance of the filter. The bridge configuration is limited in this regard as it must be balanced at each frequency of operation. Increased flexibility can be obtained by incorporating some of the crystal parameters into the integral design of a broadband filter configuration such as a high-pass or low-pass. For example, the shunt capacity, $C_o$, is primarily a function of the holder and can be integrated into the filter design. Then, as crystals of this holder group are interchanged to obtain responses at different frequencies, the overall bandpass response will not be materially changed.

Figure 32 illustrates the behavior of a low-pass filter which was designed to incorporate the crystal holder capacity. The crystal resonant impedance provides a shunt path across the transmission line at the series resonant frequency. The amount of attenuation provided by this shunt path depends upon the ratio of the characteristic impedance of the filter to the series resonant impedance of the crystal. Consequently, the impedance level of the filter should be at the maximum practical value and the series impedance of the crystals should be as low as possible. To provide maximum attenuation at crystal resonance, the low-pass filter was designed for a 200 ohm impedance level. The decrease in the attenuation, which is evident at the higher overtone response frequencies, is caused by the higher shunting impedance at the crystal overtone responses.

A crystal resonator is characterized by a number of parallel resonant frequencies which alternate between the conventional overtone and spurious response frequencies. The impedance level at parallel resonance is high compared to the series impedance. This high impedance can be used to obtain narrow bands of high insertion loss in the passband of a high-pass filter such as the one shown in figure 33. Although rejection ratios of greater than 20 dB are evident at 100 MHz, the rejection ratio rapidly deteriorates to a negligible value at 200 MHz indicating a limited applicability of this technique at UHF.

Two quartz crystals having the same fundamental response will not, in general, exhibit the same pattern of spurious responses. If the spurious response frequencies are not the same, at the particular frequency where one crystal has a low impedance path, the other crystal should exhibit a high impedance path. It might be expected, then, that a filter containing two
Figure 32. Rejection Characteristics of Low Pass Filter.
Figure 33. Rejection Characteristics High Pass Filter.
series crystal elements having the same fundamental response frequency would have fewer spurious responses than a single crystal filter. The transmission characteristics of a bandpass filter designed to include two crystal elements in series are shown in figure 31. Also shown for comparison purposes are the transmission characteristics of the filter with each crystal individually in the filter. Rather than having fewer spurious responses, this filter with two crystals in series displayed a number of spurious responses approximately equal to the sum of the responses with each crystal separately in the filter. It is possible that these two crystals were cut from the same blank and were plated by identical processes. If this is true, the spurious response of the two units would tend to coincide and both units would exhibit low impedance paths at the same frequencies. However, this test would indicate that the cascading of two crystal elements of the same fundamental frequency is not a very promising technique for reducing the number of spurious responses.

The achievement of crystal interference filters having high attenuation in the stop band and low attenuation in the passband is limited by the high series impedance of typical quartz resonators in the 200 to 400 MHz region. A reduction of this impedance is necessary for a low insertion loss in the bridge type filters and a high rejection ratio in such filters as the integrated low-pass. One way to reduce the effective series impedance is to supply some of the energy loss with an active device. By connecting the active device into the circuit in such a manner as to generate a negative resistance, the net impedance at crystal resonance can be reduced.

The diagram of figure 35 illustrates a network arrangement which can exhibit negative resistance properties. The net impedance, \( R \), at the input terminals of the circuit can be determined by first assuming an input voltage, \( V \), then calculating the input current, \( I_1 \), and finally obtaining the ratio,

\[
R = \frac{V}{I_1} \quad .
\]  

(36)

Suppose \( R_1 \) defines the amplifier input impedance as reflected through the 1:1 transformer, and \( R_2 \) designates the terminating impedance of the phase shifter. Further assume that the output impedance of the phase shifter is 50 ohms, and that \( G \) is equal to the net closed loop gain. The circuit loop equations can now be written as:

\[
(R_1 + R_2) I_1 - R_2 I_2 = V \quad \text{(37)}
\]

\[
- R_2 I_1 + (R_2 + 50) I_2 = G(I_2 R_1) \quad .
\]  

(38)

The current \( I_2 \) is the second loop current as illustrated in figure 35. After solving equations (37) and (38) for \( I_1 \) and substituting into equation (36),
Figure 34. Rejection Characteristics of Two Crystals in a Simple Notch Filter.
Figure 35. Block Diagram of Negative Resistance Circuit.
the input impedance is found to be

$$R = \frac{50 (R_1 + R_2) + (1 - G) R_1 R_2}{R_2 + 50}.$$  \hfill (39)

If $R_1 = R_2 = 50$ ohms, equation (39) becomes

$$R = 25 (3 - G).$$  \hfill (40)

Equation (40) indicates that for $G > 3$, the circuit input impedance is negative which represents an unstable condition. However, when this negative resistance is connected in series with a crystal element, stability is assured so long as the magnitude of $R$ remains less than the resonant impedance of the crystal.

The effectiveness of the negative resistance circuit in the improvement of the rejection ratio of a crystal band reject filter is illustrated in figure 36. An increase of more than 35 dB in the rejection of this simple filter configuration is evident in this figure. Although not apparent in figure 36, a second response was observed at 221 MHz. This response probably resulted from the series resonance of the crystal holder capacitance with an inductive component of the negative resistance circuit. Two attractive features of the series combination of the crystal and negative resistance are its inherent suppression of spurious responses and its relative freedom from intermodulation products. Because the series impedances at the spurious responses are much higher than at the main response, the negative impedance will effectively cancel only the main response impedance. The effect is similar to that described earlier in connection with the balanced bridge configuration in that the filter's response is enhanced at the main overtone response, but the response at the spurious resonances is not affected. The crystal element itself helps to prevent the generation of undesirable intermodulation products in the active portion of the circuit. Since the crystal is in series with the active device, two signals with a frequency separation greater than the crystal's bandwidth are greatly attenuated prior to the active device and thus do not produce intermodulation products.

Another way to overcome the losses caused by the crystal series impedance and, at the same time, reduce the effect of the crystal's spurious responses is to incorporate the crystal in a Q multiplier configuration. Figure 37 illustrates the effect of Q multiplication on a crystal bandpass filter. The curves represent a Q multiplication of approximately 6.5 and show a suppression of spurious response levels by 12 to 16 dB.

The diagram of figure 38 illustrates a technique for generating a frequency sensitive impedance. The input impedance, $R(s)$, of the coupler is 50 ohms at all frequencies except those within the crystal bandwidth. When the
Figure 36. Effects of Negative Resistance on a Crystal Notch Filter Response.
Figure 37. Q Multiplication of a Crystal Response.
Figure 38. Block Diagram of a Crystal Controlled Impedance.
loop is adjusted such that the loop gain is real with a magnitude of unity at
the crystal resonant frequency, the input impedance becomes very high. Off
crystal resonance, the loop gain is much less than unity and $R(s)$ then be-
comes approximately equal to 50 ohms.

This frequency sensitive impedance can be employed in conjunction with
a hybrid junction to develop an improved, narrow bandpass filter. The basic
filter configuration is shown in figure 39 along with its transmission char-
acteristics. The off resonance attenuation of the filter results from the
hybrid junction being balanced at all ports because $R(s)$ is approximately
equal to 50 ohms. At crystal resonance, however, the high value of $R(s)$
unbalances the hybrid and maximum transmission through the filter occurs.
This filter provides greater off resonance attenuation than does simply a
crystal element in series with the transmission line. In addition, the
effects of the crystal spurious responses are suppressed.
Figure 39. Response of a Hybrid Filter With a Crystal Controlled Impedance.
Active cancellation techniques may be effectively employed to reduce adjacent channel and co-channel interference. The three techniques developed during this study are applicable to a wide variety of operational situations. When the interfering source is sufficiently close by to obtain a sample of the interfering signal, the automatic phase lock of the first system can effectively maintain the degree of suppression of the undesired signal. If a sample cannot be obtained from the source, as is generally the case, the second or feed forward system may be used to suppress the interference. A suppression capability of greater than 60 dB for CW signals and greater than 50 dB for AM signals was demonstrated with this system. However, when the frequency spacing between the interfering signal and the desired signal is less than 50 kHz, the feed forward system attenuates the desired signal excessively. For those situations involving very closely spaced signals, the third or dual loop cancellation system is suggested. As long as the phase and amplitude characteristics of the interfering signal can be discerned from those of the desired signal, this system typically provides 50 dB of suppression to the interference.

In their present state of development, the active cancellation systems can effectively handle interfering signal levels up to 100 milliwatts. In view of the high power level of some interfering signals, the operating power level of these systems should be increased through such techniques as power division and distributed amplification.

The degree of suppression of the interference is limited to approximately 60 dB by the nonlinearities of the active components in the systems. Further work should be directed to increasing this degree of suppression by the application of techniques such as negative feedback to the development of more linear active components.

The technique of Q multiplication was successfully applied in the UHF region. With this technique, filter characteristics exhibiting very narrow bandwidths were realized using relatively small passive resonators. Q values approaching 7,000 were obtained in a relatively compact device. In the present state of development, this device can only handle power levels on the order of -25 dBm. In view of the promising results obtained with the device, further efforts should be directed to improving its power handling capabilities.

The overtone responses of quartz crystals may be used to produce very narrow bandpass and band reject responses in the VHF region. Balanced bridge and hybrid configurations are capable of rejection levels as great as 50 dB. Both configurations are effective in reducing the number and level of the crystal spurious responses. However, these filters exhibit insertion losses of 7 to 10 dB at 200 MHz and exhibit even greater insertion losses at higher
frequencies. The major portion of this insertion loss is due to the high resonant impedance of the crystal. Active techniques can reduce the effect of this high impedance. A negative resistance configuration achieved a band reject response with only 3 dB insertion loss in the passband. The technique of Q multiplication when applied to a bandpass configuration of a crystal filter increased the skirt rejection by approximately 20 dB. Both active techniques enhance the suppression of spurious responses. Unfortunately, they both produce an undesirable narrowing of the already very narrow crystal response. Further study should be made of methods for increasing the crystal response width while maintaining a low filter insertion loss and a high rejection ratio.
SECTION VI

REFERENCES


SECTION VII

SCHEMATIC DIAGRAMS
Diagram 1. UHF Local Oscillator, Mixers and Power Supply For Automatic Phase Control System.
Diagram 2. IF Amplifiers For Automatic Phase Control System.
Diagram 3. Second Local Oscillator, Mixer, Limiter, Discriminator and DC Amplifiers For Automatic Phase Control System.
Diagram 4. First Local Oscillator, Cancellation Oscillator, Buffer Amplifier and AM Modulator For AM Cancellation Filter
Diagram 5. First and Second IF Amplifiers For AM Cancellation Filter.
Diagram 6. Variable Phase Shifter, Phase and Amplitude Detectors and Signal Channel Limiter For AM Cancellation Filters.
Diagram 7. Second Local Oscillator and APC Loop Amplifiers For AM Cancellation Filters
Diagram 9. System Control Functions of Q Multiplier.

*These pieces of coax must be exactly the same length.
Diagram 10. Voltage Controlled Phase Shifter.
Diagram 11. FET Attenuator and Active Hybrid Combiner.

ALL RESISTANCES IN OHMS
### Interference Reduction Techniques Employing Active Devices

**Abstract**

This report discusses the development and performance of several techniques employing active devices for the reduction of adjacent and co-channel interference in receivers. For operational situations where the interfering source is co-sited with the receiver, an automatic phase control system is described which detects the differential phase shift between the interference path and the direct path from the source and supplies a phase corrected signal for cancellation of the interfering signal. For the more general situation where the interference is not a co-sited source, a feed forward system and a dual loop, phase locked system were developed. For 60 kHz spaced signals at 300 MHz, the feed forward system typically provides 55 dB suppression to CW interference and 40 dB suppression to AM interference. The dual loop cancellation system, which generates an auxiliary signal to cancel the interfering signal, is shown to be capable of suppressing an interfering signal by 50 dB when the interference signal is greater than an audio bandwidth from the desired signal. The improved performance of passive preselectors that can be obtained through Q multiplication is demonstrated with a Q multiplier device for the UHF region. Effective Q’s of approximately 7,000 with relatively small coaxial cavities are realized with this device. Active and passive circuit configurations which incorporate the extremely narrow passbands of quartz crystals are discussed. When used as band stop filters, interference rejection levels greater than 50 dB are possible with these filters.
<table>
<thead>
<tr>
<th>KEY WORDS</th>
<th>LINK A</th>
<th></th>
<th>LINK B</th>
<th></th>
<th>LINK C</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Q multiplier</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Crystal Filters</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Interference rejection</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>