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The object of this research is the preparation of a catalogue listing the technical and tactical characteristics of all U. S. Army non-radar type electronic equipments, in particular, the radio interference producing and susceptibility characteristics.

1 SEPTEMBER 1961 to 15 FEBRUARY 1963

PLACED BY THE U. S. ARMY
ELECTRONICS RESEARCH AND DEVELOPMENT LABORATORY
FORT MONMOUTH, NEW JERSEY
# TABLE OF CONTENTS

<table>
<thead>
<tr>
<th>Section</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>I. PURPOSE</td>
<td>1</td>
</tr>
<tr>
<td>II. ABSTRACT</td>
<td>3</td>
</tr>
<tr>
<td>III. PUBLICATIONS, LECTURES, REPORTS AND CONFERENCES</td>
<td>5</td>
</tr>
<tr>
<td>IV. FACTUAL DATA</td>
<td>7</td>
</tr>
<tr>
<td>A. Measurement Techniques for Microwave Communications Equipment</td>
<td></td>
</tr>
<tr>
<td>1. Introduction</td>
<td>7</td>
</tr>
<tr>
<td>2. Transmitter Tests and Test Procedures</td>
<td>9</td>
</tr>
<tr>
<td>2.1. Power Output</td>
<td>9</td>
</tr>
<tr>
<td>2.2. Frequency Stability</td>
<td>12</td>
</tr>
<tr>
<td>2.3. Spurious and Harmonic Emissions</td>
<td>15</td>
</tr>
<tr>
<td>2.4. Intermodulation</td>
<td>19</td>
</tr>
<tr>
<td>2.5. Modulator Bandwidth</td>
<td>21</td>
</tr>
<tr>
<td>2.6. Modulation Characteristics</td>
<td>24</td>
</tr>
<tr>
<td>2.7. Sideband Splatter</td>
<td>25</td>
</tr>
<tr>
<td>3. Receiver Tests and Test Procedures</td>
<td>26</td>
</tr>
<tr>
<td>3.1. Noise Figure</td>
<td>26</td>
</tr>
<tr>
<td>3.2. Sensitivity</td>
<td>30</td>
</tr>
<tr>
<td>3.3. Selectivity</td>
<td>32</td>
</tr>
<tr>
<td>3.4. Spurious Responses</td>
<td>34</td>
</tr>
<tr>
<td>3.5. Intermodulation</td>
<td>37</td>
</tr>
<tr>
<td>4. Pulse Position Multiplexer Terminal Equipment Tests and Test Procedures</td>
<td>40</td>
</tr>
<tr>
<td>4.1. Bandwidth</td>
<td>40</td>
</tr>
<tr>
<td>4.2. Crosstalk</td>
<td>43</td>
</tr>
</tbody>
</table>
# TABLE OF CONTENTS (Continued)

<table>
<thead>
<tr>
<th>Section</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>5. System Tests</td>
<td>44</td>
</tr>
<tr>
<td>5.1. Error Rate</td>
<td>44</td>
</tr>
<tr>
<td>5.2. Proposed System Tests</td>
<td>57</td>
</tr>
<tr>
<td>B. Mixer Testing and Evaluation.</td>
<td>58</td>
</tr>
<tr>
<td>1. Introduction</td>
<td>58</td>
</tr>
<tr>
<td>2. Experimental Results</td>
<td>59</td>
</tr>
<tr>
<td>3. Mixer Improvement</td>
<td>80</td>
</tr>
<tr>
<td>4. Response Prediction</td>
<td>80</td>
</tr>
<tr>
<td>5. Conclusions</td>
<td>89</td>
</tr>
<tr>
<td>C. Spurious Response Bandwidth and Rejection</td>
<td>89</td>
</tr>
<tr>
<td>1. Introduction</td>
<td>89</td>
</tr>
<tr>
<td>2. Spurious Response Bandwidth</td>
<td>90</td>
</tr>
<tr>
<td>3. Spurious Response Rejection</td>
<td>94</td>
</tr>
<tr>
<td>3.1. Sensitivity</td>
<td>94</td>
</tr>
<tr>
<td>3.2. Selectivity</td>
<td>94</td>
</tr>
<tr>
<td>3.3. Mixer Parameters</td>
<td>95</td>
</tr>
<tr>
<td>4. Conclusions</td>
<td>96</td>
</tr>
<tr>
<td>D. Data Prepared or Published During the Contract Period</td>
<td>96</td>
</tr>
<tr>
<td>E. Automatic Data Reduction</td>
<td>98</td>
</tr>
<tr>
<td>V. CONCLUSIONS</td>
<td>101</td>
</tr>
<tr>
<td>VI. OVERALL CONCLUSIONS</td>
<td>103</td>
</tr>
<tr>
<td>VII. RECOMMENDATIONS</td>
<td>107</td>
</tr>
<tr>
<td>VIII. IDENTIFICATION OF KEY TECHNICAL PERSONNEL</td>
<td>109</td>
</tr>
<tr>
<td>IX. REFERENCES</td>
<td>115</td>
</tr>
</tbody>
</table>
APPENDIX - Determination of Coefficients for Frequencies Generated by a Mixer with Two Input Signals

1. Definitions and Assumptions ............................... 117
2. Procedures .................................................. 118
3. Formation of Frequency Term Coefficients ............... 126
# LIST OF FIGURES

<table>
<thead>
<tr>
<th>Figure No.</th>
<th>Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.</td>
<td>Block Diagram of Power Output Test Setup for Transmitters Rated Up to 10 Watts</td>
<td>11</td>
</tr>
<tr>
<td>2.</td>
<td>Block Diagram of Power Output Test Setup for Transmitters Rated from 10 Watts to 1 Kilowatt</td>
<td>11</td>
</tr>
<tr>
<td>3.</td>
<td>Block Diagram of Frequency Stability Test Setup for Transmitters</td>
<td>14</td>
</tr>
<tr>
<td>4.</td>
<td>Block Diagram of Spurious and Harmonic Emissions Test Setup for Transmitters Rated up to 10 Watts</td>
<td>18</td>
</tr>
<tr>
<td>5.</td>
<td>Block Diagram of Spurious and Harmonic Emissions Test Setup for Transmitters Rated from 10 Watts to 1000 Watts</td>
<td>18</td>
</tr>
<tr>
<td>6.</td>
<td>Block Diagram of Intermodulation Test Setup for Transmitters</td>
<td>23</td>
</tr>
<tr>
<td>7.</td>
<td>Block Diagram of Modulator Bandwidth for Transmitters</td>
<td>23</td>
</tr>
<tr>
<td>8.</td>
<td>Block Diagrams of Noise Figure Test Setups for Receivers</td>
<td>29</td>
</tr>
<tr>
<td>9.</td>
<td>Block Diagram of Sensitivity Test Setup for Receivers</td>
<td>31</td>
</tr>
<tr>
<td>10.</td>
<td>Block Diagram of Intermodulation Test Setup for Receivers</td>
<td>41</td>
</tr>
<tr>
<td>11.</td>
<td>Block Diagram of Test Setup for Bandwidth Test of Terminal Equipments</td>
<td>41</td>
</tr>
<tr>
<td>12.</td>
<td>Block Diagram of Error Rate Test Setup</td>
<td>48</td>
</tr>
<tr>
<td>13.</td>
<td>Block Diagram of Sampling Device</td>
<td>50</td>
</tr>
<tr>
<td>14A</td>
<td>Square Wave Generator Input Circuit to TH-5/TG Telegraph Terminal Equipment</td>
<td>52</td>
</tr>
<tr>
<td>14B</td>
<td>Simplified Schematic of Resistive Adder</td>
<td>52</td>
</tr>
<tr>
<td>15.</td>
<td>Schematic Diagram of Sampling Device</td>
<td>53</td>
</tr>
<tr>
<td>16.</td>
<td>Power Supply for Error Detector</td>
<td>56</td>
</tr>
<tr>
<td>17.</td>
<td>Schematic Diagram of 1N82A Test Mixer</td>
<td>60</td>
</tr>
<tr>
<td>18.</td>
<td>Harmonic and Response Levels Versus Forward Bias for Single 1N82A Diode Mixer With No Frequency Discrimination</td>
<td>64</td>
</tr>
<tr>
<td>Figure No.</td>
<td>Description</td>
<td>Page</td>
</tr>
<tr>
<td>-----------</td>
<td>------------------------------------------------------------------------------</td>
<td>------</td>
</tr>
<tr>
<td>19</td>
<td>Harmonic Level Versus Reverse Bias For Single 1N82A Diode Mixer With No Frequency Discrimination</td>
<td>68</td>
</tr>
<tr>
<td>20</td>
<td>Harmonic Level Versus Forward Bias For Single 1N82A Diode Mixer With No Output Frequency Discrimination</td>
<td>69</td>
</tr>
<tr>
<td>21</td>
<td>Characteristic of Anode Current in Diodes for Small Conduction Angles</td>
<td>71</td>
</tr>
<tr>
<td>22</td>
<td>Graphical Solution of the Expression for $C_3$ or Third Harmonic Component of Diode Current</td>
<td>74</td>
</tr>
<tr>
<td>23</td>
<td>Plot of Absolute Value of the Third Harmonic Component of Diode Current</td>
<td>75</td>
</tr>
<tr>
<td>24</td>
<td>Logarithmic Plot of Absolute Value of Third Harmonic Component of Diode Current</td>
<td>76</td>
</tr>
<tr>
<td>25</td>
<td>Response Level Versus Reverse Bias For Single 1N82A Diode Mixer With No Frequency Discrimination</td>
<td>77</td>
</tr>
<tr>
<td>26</td>
<td>Response Level Versus Forward Bias For Single 1N82A Diode Mixer With No Frequency Discrimination</td>
<td>78</td>
</tr>
<tr>
<td>27</td>
<td>Harmonic and Response Levels Versus Forward Bias For Single 1N82A Diode Mixer With Capacitor Across Output</td>
<td>79</td>
</tr>
<tr>
<td>28</td>
<td>Comparison of Harmonic and Response Levels Versus Cathode Voltage For Balanced Infinite Input Impedance Diode Mixer</td>
<td>81</td>
</tr>
<tr>
<td>29</td>
<td>Comparison of Harmonic and Response Levels Versus Plate Supply Voltage For Balanced Infinite Input Impedance Diode Mixer</td>
<td>82</td>
</tr>
<tr>
<td>30</td>
<td>Measurable Response Level Versus Cathode Voltage for Balanced Infinite Input Impedance Diode Mixer</td>
<td>83</td>
</tr>
<tr>
<td>31</td>
<td>Pascal's Triangle of Binomial Coefficients Divided by $2^{(n-1)}$ where $n$ is the Power to Which the Binomial is to be Raised</td>
<td>85</td>
</tr>
<tr>
<td>32</td>
<td>Spurious Response Triangle For a Mixer Having a Polynomial Representation of the Tenth Degree</td>
<td>87</td>
</tr>
</tbody>
</table>
## LIST OF TABLES

<table>
<thead>
<tr>
<th>Table</th>
<th>Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>I.</td>
<td>RECORD OF HARMONIC LEVEL FOR STANDARD IF RESPONSE</td>
<td>62</td>
</tr>
<tr>
<td>II.</td>
<td>BANDWIDTHS FOR DIFFERENT P AND Q AM RECEIVER</td>
<td>93</td>
</tr>
<tr>
<td>III.</td>
<td>RECEIVER TYPES TESTED AND MANUSCRIPT VOLUME NUMBERS</td>
<td>97</td>
</tr>
<tr>
<td>IV.</td>
<td>LIST OF TRANSMITTERS EVALUATED</td>
<td>97</td>
</tr>
<tr>
<td>V.</td>
<td>GENERAL MANUSCRIPTS</td>
<td>97</td>
</tr>
</tbody>
</table>

### APPENDIX TABLES

<table>
<thead>
<tr>
<th>Appendix</th>
<th>Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>IA.</td>
<td>COSINE $\theta$ RAISED TO THE TENTH POWER</td>
<td>118</td>
</tr>
<tr>
<td>IIA.</td>
<td>VARIOUS COSINE PRODUCT COMBINATIONS</td>
<td>120</td>
</tr>
<tr>
<td>IIIA.</td>
<td>FREQUENCIES GENERATED BY TERMS UP TO $n = 10$ IN MIXER POLYNOMIAL</td>
<td>127</td>
</tr>
<tr>
<td>IVA.</td>
<td>HARMONICS TO $K = 10$ WITH ONE INPUT ONLY</td>
<td>132</td>
</tr>
<tr>
<td>VA.</td>
<td>ADDITIONAL TERMS FROM COMBINATIONS CONTRIBUTING TO HARMONIC GENERATION</td>
<td>133</td>
</tr>
</tbody>
</table>
I. PURPOSE

The purpose of this project is to conduct a comprehensive investigation to determine the characteristics of U. S. Army communications equipments deemed necessary for the prediction and minimizing of radio interference. Emphasis shall be placed on pulse communications type equipments. Measurement techniques for obtaining the required data and a format for a directory of these data shall be developed. Computer methods shall be developed for processing these data to obtain outputs useful in determining optimum characteristics for communications equipments operating in prescribed interference environments.

The areas of investigation on this project are divided into two tasks as follows:

I. The development of tests and procedures for the evaluation of the interference susceptibility and emanation of pulse communications equipments.

II. The development of computer methods for constructing and processing mutual interference matrices and the application of computer methods to the processing of laboratory data.
II. ABSTRACT

Tests and test procedures for obtaining spectrum signature data for U. S. Army microwave communications equipment were developed. System and terminal equipment tests and test procedures which are not included in existing spectrum signature collection plans are described. Bandwidth and crosstalk tests and test procedures developed for the pulse position modulated-time division multiplexer AN/TCC-13 are discussed. The system error rate test is presented in detail, and additional proposed system tests are briefly outlined.

The results of mixer studies are presented. Data taken on vacuum tube and diode mixers verify existing mixer theories and indicate a theory involving the effect of conduction angle. A prediction method for estimating the levels of mixer responses is outlined, but additional work is needed to establish its validity.

The results of receiver spurious response studies are presented. The spurious response bandwidth is found to be inversely proportional to the integer 'q' in the spurious response equation. The difficulties in obtaining useful spurious response data are discussed.

Three computer programs for solving certain types of interference problems are reviewed. The volumes of the Manuscript of Catalogue containing these computer programs were published during the contract period.
III. PUBLICATIONS, LECTURES, REPORTS & CONFERENCES

On September 14-15, 1961, Dr. I. E. Perlin and Mr. R. N. Bailey visited Fort Monmouth for a technical discussion.

Mr. E. W. Wood and Mr. R. N. Bailey visited Empire Devices, Inc., to attend a seminar on October 16-17, 1961.

Mr. R. N. Bailey, Dr. I. E. Perlin, Mr. J. G. Holey, Mr. P. Johnson, Jr., Mr. W. C. Knapp, and Mr. E. W. Wood attended the Seventh Tri-Service Conference on Electromagnetic Compatibility on November 7-9, 1961, in Chicago, Illinois. The papers presented were: "A Comparison of the Spectrum Signatures of AM, FM, and SSB Communications Equipments," by Mr. E. W. Wood; and "A Discussion of Test Procedures For Obtaining Spectrum Signatures of Pulse Communications Equipment," by Mr. R. N. Bailey.

Mr. R. N. Bailey, Mr. E. W. Wood, Mr. P. Johnson, Jr., Mr. J. G. Holey, Dr. I. E. Perlin, Mr. J. L. Cundiff, Mr. T. T. Spengler, and Mr. R. Techo visited Fort Monmouth on January 8-10, 1962, to present a seminar on computer programs.

Mr. R. N. Bailey and Dr. I. E. Perlin visited Fort Huachuca, Arizona, on February 12-16, 1962, to attend the Department of Defense Symposium on Technical Problems of the Electromagnetic Compatibility Program.

Mr. E. W. Wood and Mr. T. T. Spengler visited USASRDIL, Fort Monmouth, on March 21-23, 1962, for a discussion on computer programs.

On March 23, 1962, Mr. E. W. Wood attended the meeting at Fort Monmouth of the DOD committee on Interference Instrumentation and Measurements.

Mr. R. N. Bailey attended the IRE International Convention, New York City, March 26, 1962.
Mr. Sidney Weitz, Mr. Warren Kesselman, and Mr. Dick Markham visited Georgia Tech on May 4-5, 1962, for a technical discussion.

Mr. E. W. Wood attended the Instrument Society of American Show held in Huntsville, Alabama, on May 9-10, 1962.

Mr. R. N. Bailey attended the PGRFI Symposium held at San Francisco, California, on June 28-29, 1962.

Mr. R. N. Bailey and Mr. E. W. Wood visited Fort Monmouth on July 11, 1962, to discuss project progress and regions of emphasis for the ensuing contract period.

Mr. R. N. Bailey attended the DOD Symposium on "Measurement Techniques and Test Instrumentation" held at ECAC, Annapolis, Maryland, on August 29-30, 1962.

Mr. R. N. Bailey, Mr. D. W. Robertson, Mr. J. L. Cundiff, and Mr. C. W. Stuckey visited Fort Monmouth on October 16-17, 1962, for a presentation and demonstration of computer techniques.

Mr. R. N. Bailey, Mr. J. L. Cundiff, Mr. R. D. Trammell, Jr., and Mr. J. R. Walsh, Jr., attended the Eighth Tri-Service Conference on Electromagnetic Compatibility held in Chicago, Illinois, on October 30-31, and November 1, 1962. The paper, "Computer Reduction of Laboratory Data for Construction of MIC's," was presented.

Mr. J. L. Cundiff also presented this paper to the Mid Southeastern Chapter of the Association of Computing Machines in Nashville, Tennessee, on December 14, 1962.

A Seminar on "Measurement Techniques and Pitfalls Associated with Spectrum Signature Data Collection," was held at Georgia Tech on November 28-29, 1962.
IV. FACTUAL DATA

A. Measurement Techniques for Microwave Communications Equipment

1. Introduction

The development and verification of tests and test procedures for the evaluation of the interference susceptibility and emanation characteristics of U. S. Army communications equipments operating in the frequency range of 1 to 10 kmc are divided into four major categories as follows:

- (1) Transmitters
- (2) Receivers
- (3) Terminal Equipments
- (4) System Tests

Since most interference characteristics are traceable to transmitters and receivers, the extension, modification, and verification of the tests and test procedures developed earlier for transmitters and receivers form the basis for the microwave tests. However, because of the complexity of microwave systems, other tests are essential.

Many microwave links are necessary for reliable long range communications. Multiplex and demultiplex terminal equipments provide methods whereby a large number of time-shared or frequency-shared information channels may be sent over a single communication link. The terminal equipments may contribute to the interference problem by creating additional interference or by adding distortion to the desired signals. Therefore, it is required that the terminal equipments be investigated for their interference producing or susceptibility characteristics.
Normally microwave transmitters and receivers are operated with auxiliary terminal equipments, and system tests are desirable to determine the overall system performance. It is necessary that closed system tests be conducted in the laboratory.

An investigation of characteristics of the U. S. Army communications equipments which operate at frequencies above 1 kmc has shown that at the present time the Signal Corps uses primarily the PPM-FM, PCM-FM, or other composite "double-modulation" type systems. The only microwave equipments available for interference studies were the AN/TRC-29 and the AN/GRC-50(V) sets. The AN/TRC-24 set was tested, but no multiplier-amplifier or converter heads for the range above 1 kmc were available.

The AN/TRC-29 is basically a microwave FM radio relay set which can be used with the PPM Multiplex Set AN/TCC-13, the PCM Multiplex Set AN/TCC-15, a frequency-division multiplex equipment, or a commercial television terminal equipment for transmission of television signals. The PPM multiplex set AN/TCC-13 was provided for testing and procedure evaluation with the AN/TRC-29 group.

The AN/GRC-50 is also basically an FM radio relay set. It is designed specifically for system use and generally is not used without supporting PCM or FDM multiplex equipment; however, no terminal equipment was available. The AN/GRC-50 sets were not available until November 1, 1962; therefore, only a few transmitter and receiver interference studies have been made on these units.

Since no pulsed-carrier type equipments were available for use in the development and verification of the microwave tests and test procedures, the
directory is slanted toward FM type equipment. The use of waveguides for communications equipments is limited, and the procedures do not include waveguide techniques.

2. **Transmitter Tests and Test Procedures**

The following tests have been developed for the evaluation of microwave transmitter spectrum signature characteristics:

1. Power Output
2. Frequency stability
3. Spurious and Harmonic Emissions
4. Intermodulation
5. Modulator Bandwidth
6. Modulation Characteristics
7. Sideband Splatter

The sequence of these tests has been selected for maximum testing efficiency, but is not mandatory. Other tests, including case radiation, have not been verified.

2.1. **Power Output**

2.1.1. **General**

Since the power output is a good indication of the transmitter's overall condition, it should be determined with a high degree of accuracy. In general, the power output of a microwave transmitter varies a few decibels within a band and from klystron to klystron or PA tube to PA tube, etc. The transmitter should be tuned according to the technical manual for the set.

If a bolometer-type power meter is used to measure the power output, extreme or varying ambient temperatures should be avoided. The power meter should be shielded from transmitter radiations. Power correction factors must be added when pulsed outputs are measured with average-power indicating devices. A low-pass filter may be needed if the transmitter harmonics are
found to be strong; however, the insertion loss of the filter at the tuned frequency must be known as well as the overall attenuation of the path from the transmitter to the power measuring device. The bolometer element should not be overloaded or improperly biased and should provide the proper termination impedance to match the line. Impedance matching devices may have to be utilized. If an NFIM or frequency selective voltmeter is used, the bandwidth of the equipment must be greater than the transmitter output spectrum occupancy.

AFC circuits should be operating properly when this test is performed. Since klystrons can operate in many different modes, the correct mode must be selected. This usually results in the largest power output.

The standard test frequencies are the 5 percent, 50 percent, and 95 percent frequencies within each band.

Several methods for measuring power output are shown in Figures 1 and 2.

2.1.2. Required Equipment

The required test equipments for the different methods are listed below.

1. Calibrated attenuators similar to Microlab AD-10N and AB-05N.
2. Bolometer mount with suitable frequency range.
3. Power meter similar to HP-430C.
4. Directional Coupler similar to Narda 3022.
5. Terminating load similar to Microlab TD-5FN.
6. NFIM and substitution signal generator if necessary.
7. Connecting cables.
8. Screen Room.

2.1.3. Test Setup See block diagrams, Figures 1 and 2.
Figure 1. Block Diagram of Power Output Test Setup for Transmitters Rated Up to 10 Watts.

Figure 2. Block Diagram of Power Output Test Setup for Transmitters Rated from 10 Watts to 1 Kilowatt.
2.1.4. Test Procedures

(1) Arrange and connect the equipment as shown in Figure 1 or 2 depending on the power output. For initial tuning the transmitter should be connected to the dummy load only.

(2) Select a calibrated attenuator(s) which limits the input power to the bolometer to a safe value (see power meter instruction manual), but which permits a mid or high scale reading on the power meter.

(3) Adjust the pulsed output for nominal pulse width and repetition rate. An unmodulated output is used for an FM transmitter.

(4) Tune the transmitter to a standard test frequency, \( f_0 \), at rated output in accordance with the tuning procedure outlined in the technical manual.

(5) Record the power indicated by the power meter. Apply correction factors to obtain power output for the pulsed-type output. There are no correction factors for FM measurements. Convert the power to dbm.

(6) Add the attenuation, in db, corrected for frequency, of the attenuators and coupler to the value obtained in (5). The sum is the power output of the transmitter at \( f_0 \).

(7) Repeat (4) through (6) at other standard test frequencies.

2.1.5. Presentation of Test Data These data are presented in the following tabular form. The heading of the column should include the serial number of the transmitter.

<table>
<thead>
<tr>
<th>Frequency (Mc)</th>
<th>Power Output (dbm)</th>
</tr>
</thead>
</table>

2.2. Frequency Stability

2.2.1. General A serious problem common to all microwave
users is that of operating frequency stability. Since most microwave relay sets are the continuous duty type, this test determines the behavior of the AFC circuits over a long period of time. The transmitter is tuned to the mid-frequency in each band. Pulsed-type transmitters should be operated at the nominal repetition rate and pulse width.

The frequency measuring device should not be overloaded or have insufficient levels for proper measurement. Variations of power line voltages on the transmitter and test equipment should be at a minimum. A defective klystron or oscillator tube may cause frequency pulling or drift. The klystron should be tuned to the proper mode.

2.2.2. Required Equipment

(1) Same as Power Output Test.

(2) Frequency Counter and/or Converter similar to Beckman/Berkeley models 7370 and 7580.

2.2.3. Test Setup See Figure 3.

2.2.4. Test Procedures

(1) Tune the transmitter to rated power output at the mean test frequency.

(2) Ascertain that all AFC circuits are operating and that all unstable conditions which may affect the operating frequency are minimized.

(3) Record the frequency as soon as the transmitter becomes operational.

(4) Record the frequency at intervals of not more than 15 minutes for about 4 hours.

2.2.5. Presentation of Test Data The test data are presented in tabular form as shown below. The heading of each table
Figure 3. Block Diagram of Frequency Stability Test Setup for Transmitters.
should include the serial number of the transmitter, tuned frequency, and power. A graph of frequency versus time may be beneficial.

<table>
<thead>
<tr>
<th>Operating Time (Minutes)</th>
<th>Measured Frequency (Mc)</th>
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</thead>
</table>

2.3. **Spurious and Harmonic Emissions**

2.3.1. **General** The purpose of this test is to scan the spectrum from 0.15 Mc to 44 kmc so as to evaluate any spurious and harmonic outputs of the transmitter under test. The majority of these outputs are related to the fundamental and/or harmonics of the master oscillator, multipliers, driver, and power amplifier frequencies. Other frequencies which are not harmonically related to the master oscillator may be present due to mixing of the various frequencies present in a transmitter. These intermodulation products are usually harmonically related to oscillators used in control systems and similar auxiliary circuits. Also, in some transmitters containing both of the above groups, sums and differences of some of these frequencies are present. This test is performed with the transmitter properly terminated. Although communications type transmitters are not used with the same antenna at all times, the test provides a basis for the comparison of different transmitter types.

When performing this test, precautions must be taken to avoid recording spurious responses of the NFIM rather than the transmitter outputs. Spurious outputs of frequency counters and other test equipment may appear as transmitter outputs. The measurements should be accurate to 1 part in
The power meter and other nonlinear devices in the line should be disconnected since these may cause spurious outputs for a large input signal. The length of lines at microwave frequencies is important, and all lines should be properly terminated. A stub tuner may be necessary to properly match the transmitter to the dummy load.

Several methods for proper identification of spurious emissions are as follows:

1. Low-pass filters are used when measuring frequencies below the tuned frequency.

2. High-pass filters are used when measuring frequencies above the tuned frequency. However, many microwave high-pass filters have increased attenuation at 2.5 times cutoff frequency.

3. Band-reject or bridged-T filters are used to attenuate the fundamental level.

4. Band-pass filters may be utilized at the spurious frequency in question.

5. Absorption type wavemeters may produce an indication at the spurious frequency.

6. Another NFIM with a different conversion process may be helpful if a receiver spurious response is suspected.

7. A 3 db pad inserted at the receiver input will usually cause a 3 db reduction in the spurious output level. A decrease greater than 3 db indicates a NFIM spurious response.

The attenuation of the path from the transmitter to the measuring device should be measured at each spurious frequency. Most attenuators do not have a constant attenuation over a wide frequency range.

2.3.2. Required Equipment

1. Same as in Power Output Test.
(2) Spectrum analyzer similar to Panoramic SPA-4a or an NFIM similar to the Empire Devices NF-112.

(3) Substitution signal generators (0.15 Mc to 44 kmc).

(4) Signal sampler similar to Microlab HY-10C or directional coupler similar to Narda 3022.

2.3.3. Test Setup See Figures 4 and 5.

2.3.4. Test Procedures

(1) Tune the transmitter to rated power output at a standard test frequency, \( f_o \). See Power Output Test before proceeding.

(2) Select a directional coupler or signal sampler and an attenuator or attenuators which will limit the power input to the measuring device to a safe value.

(3) Set the spectrum analyzer or NFIM input attenuator to minimum and tune to 0.150 Mc.

(4) Slowly tune the spectrum analyzer or NFIM from 0.150 Mc to 44 kmc and each time a response is noted, adjust the input attenuator to obtain an on scale reading; substitute the signal generator output at the same frequency. Adjust the level appearing on the analyzer screen or NFIM indicator to the same level found using the transmitter output. Spurious responses of the analyzer or NFIM must be recognized and avoided and the spurious frequency unquestionably verified.

(5) The power output of the transmitter harmonic or spurious frequency is the reading determined in (4) plus the attenuation, in \( \text{db} \), of the path from the transmitter to the receiver at the harmonic or spurious frequency.

(6) Repeat (1) through (5) at other standard frequencies.
Figure 4. Block Diagram of Spurious and Harmonic Emissions Test Setup for Transmitters Rated Up to 10 Watts.

Figure 5. Block Diagram of Spurious and Harmonic Emissions Test Setup for Transmitters Rated from 10 Watts to 1000 Watts.
2.3.5. **Presentation of Test Data** These data are presented in tabular form. The heading of each table should include the tuned frequency and the serial number of the transmitter.

<table>
<thead>
<tr>
<th>Measured Frequency (Mc)</th>
<th>Calculated Frequency (Mc)</th>
<th>Power Output (dbm)</th>
<th>Identification</th>
</tr>
</thead>
</table>

It is desirable to plot the spurious output data on semi-log graph paper with dbm versus frequency.

2.4. **Intermodulation**

2.4.1. **General** This test evaluates the intermodulation-generating properties of the output stage of a transmitter. The level of intermodulation products obtained when an external signal is coupled into a transmitter output circuit depends on the selectivity of the coupling circuit, the level of the interfering signal and the nonlinearity of the output stage. Spurious frequencies will then be generated in accordance with the equation

\[ f_s = nf_o + mf_i \]

where \( f_s \) = generated spurious frequency,

\( f_o \) = frequency of transmitter under test,

\( f_i \) = interfering signal frequency, and

\( m,n = 0,1,2,3, \ldots \) .

The spurious frequencies generated, which are relatively near the fundamental frequency of the transmitter, will be radiated at much higher levels due to the selectivity of the output circuit. For example, if a signal from an interfering transmitter is coupled into the desired trans-
mitter and \( f_1 = f_0 + \Delta f \), the spectrum of frequencies generated will include
\( f_0 + \Delta f, f_0 + 2\Delta f, f_0 + 3\Delta f, \ldots, f_0 + n\Delta f \). In order to be radiated, these
frequencies must be passed by the output circuit of the transmitter.

Therefore, the measurable output spectrum will be terminated at \( f_0 + n\Delta f \)
where \( n \) may be large for low Q output circuits, high amplitude interfering
signals and small \( \Delta f \)'s. The sum and difference of \( f_i \) and \( f_0 \) are generally
strong.

One possible source of error in these measurements is that due to
intermodulation in the NFIM. The identification techniques for transmitter
intermodulation frequencies are similar to the techniques described in
Section 2.3.1.

2.4.2. Required Equipment

(1) Same as Power Output Test.

(2) Additional transmitter with termination for
interfering source.

(3) Signal sampler or directional coupler.

2.4.3. Test Setup  See Figure 6.

2.4.4. Test Procedures

(1) Tune the transmitter under test to rated output
at a standard test frequency, \( f_0 \). See Power
Output Test Section.

(2) Tune the interfering transmitter to rated output
at a frequency which is 1 percent higher than \( f_0 \).
This corresponds to a spacing of 1 percent.

(3) Connect the equipment as shown in Figure 6. The
value of the coupling attenuator is such that the
ratio of the power coupled from the output at \( f_1 \)
to the output at \( f_0 \) is approximately 20 db. This
corresponds to a coupling of 20 db.
(4) Measure and record the outputs of \( f_0 \) and \( f_i \) using signal generator substitution at the NFIM.

(5) Tune the NFIM in steps of \( \Delta f \) above \( f_i \) and in steps of \( \Delta f \) below \( f_0 \) and measure the intermodulation products. Caution should be taken to insure that no receiver intermodulation occurs.

(6) Correct the measurements in (4) and (5) for attenuation at each measured frequency and record the output, in dbm, of \( f_i \), \( f_i \), and of the intermodulation products which appear at the output of the transmitter under test.

(7) Repeat (3) through (6) for couplings of 40 and 60 db.

(8) Repeat (2) through (7) for spacings of 5 percent and 10 percent.

(9) Repeat the entire test at other desired standard test frequencies.

(10) If insufficient data are obtained with the spacings and coupling values outlined above, other values should be tried.

2.4.5. Presentation of Test Data  The data are to be presented in the following tabular form. The heading of each table should include the serial number of the transmitter under test.

<table>
<thead>
<tr>
<th>Measured Frequency (Mc)</th>
<th>Calculated Frequency (Mc)</th>
<th>Output (dbm)</th>
<th>Intermodulation (Order)</th>
<th>Coupling (db)</th>
<th>Spacing (Mc)</th>
</tr>
</thead>
</table>

2.5. Modulator Bandwidth

2.5.1. General  This test determines the modulator bandwidth for the transmitter. FM carrier transmitters are adjusted for 30 percent of rated deviation. If a deviation meter is not available, the zero count method \(^4\) may be useful in determining the required deviation.
2.5.2. Required Equipment

(1) Same as Section 2.3.2.

(2) Deviation Detector or Spectrum Analyzer

(3) Video Generator

2.5.3. Test Setup  See Figure 7.

2.5.4. Test Procedures

(1) Arrange the equipment as shown in Figure 7.

(2) Adjust the video signal generator output to obtain approximately 30 percent modulation.

(3) Vary the generator frequency over the range from 30 cps to 15 Mc while observing the modulation monitor. When the maximum percentage modulation point has been established, leave the generator tuned to that frequency and adjust the generator output control to obtain 30 percent modulation. Record this frequency and the signal generator output level.

(4) Increase the generator output by one decibel.

(5) Increase the generator frequency until the percentage modulation decreases to 30 percent. Record this frequency.

(6) Lower the generator frequency until the percentage modulation decreases to 30 percent. Record this frequency.

(7) Repeat (4) through (6) for level increases of 3, 6, 10, 20, 30 ... decibels until the generator output limitations are reached.

2.5.5. Presentation of Test Data  These data are presented in tabular and graphical form.

<table>
<thead>
<tr>
<th>Modulating Frequency (kc)</th>
<th>Modulator Response (db relative to maximum)</th>
</tr>
</thead>
</table>
Figure 6. Block Diagram of Intermodulation Test Setup for Transmitters.

Figure 7. Block Diagram of Modulator Bandwidth for Transmitters.
2.6. Modulation Characteristics

2.6.1. General This test is a check of the modulation characteristics for FM carrier transmitters. A curve of percent rated deviation is plotted versus video input level. A similar test for pulsed-carrier transmitters may not be meaningful since the modulator gates the magnetron on and off at all video levels above a certain threshold level.

A test for modulation characteristics for pulsed-type transmitters is possible by varying the video input frequency while maintaining constant amplitude; but since a multiplexer with fixed video frequency may be utilized, such a test does not necessarily contribute information for interference studies.

2.6.2. Required Equipment

Same as Section 2.5.2.

2.6.3. Test Setup Same as shown in Figure 7.

2.6.4. Test Procedures

(1) Tune the transmitter to rated power output at a standard test frequency.

(2) Connect the equipment as shown in Figure 7 and apply a mean test signal.

(3) Vary the video input level to the modulator in convenient steps and record the percent of rated deviation at the transmitter output for each step.

(4) The video input level is in dbm and the deviation is in percent of rated deviation.

2.6.5. Presentation of Test Data The data are presented in the following tabular form and as a graph of percent of rated deviation versus video input. The heading of each table should include the trans-
mitter serial number, power output, tuned frequency, modulating frequency, rated deviation, etc.

<table>
<thead>
<tr>
<th>Video Input (dBm)</th>
<th>Deviation (percent of rated)</th>
</tr>
</thead>
</table>

2.7. **Sideband Splatter**

2.7.1. **General** Sideband splatter is defined as those portions of the sidebands which fall outside the assigned channel and is caused by nonlinearities in the modulator and power amplifier stages.

FM carrier transmitters are adjusted for 90 percent of rated deviation by varying the video input amplitude to the modulator. Pulsed output transmitters are adjusted for 100 percent modulation at the mean pulse rate.

2.7.2. **Required Equipment**

(1) Same as Section 2.3.2.

(2) Camera

2.7.3. **Test Setup** See Figure 7

2.7.4. **Test Procedures**

(1) Tune the transmitter to rated power output at a standard test frequency, \( f_0 \). See Power Output Test.

(2) Connect the equipment as shown in Figure 7.

(3) The FM transmitter is adjusted for 90 percent of rated deviation.

(4) Tune the spectrum analyzer to \( f_0 \) and adjust the sweep width so that all sideband components appear on the screen.

(5) Adjust the analyzer for a 40 dB dynamic range. Photograph the presentation.

(6) Repeat (5) for a 60 dB dynamic range if available.

(7) Repeat the test at other standard test frequencies.
2.7.5. Presentation of Test Data The test data are presented as photographs. Each photograph should be identified with power output, sweep width, dynamic range, sweep rate, etc.

3. Receiver Tests and Test Procedures

The tests which have been developed for microwave receivers are as follows:

(1) Noise Figure
(2) Sensitivity
(3) Selectivity
(4) Spurious Responses
(5) Intermodulation

Other tests including cochannel and close-channel tests have been written, but not verified.

3.1. Noise Figure

3.1.1. General The fundamental limitation of receiver sensitivity is the noise generated within the receiver. Receiver noise occurs in all stages, but since noise generated in the "front end" is amplified by all succeeding amplifier stages it is the principal noise source.

The receiver noise figure as measured by either of the two methods described here is essentially an evaluation of the noise generated by the tubes and circuitry in the input stages. A low noise figure is indicative of low noise generation and therefore good receiver design, whereas a high noise figure indicates either poor design or misalignment. A receiver with a low noise figure generally has good sensitivity.

Two methods for measuring noise figure are outlined. The second is designed specifically for use with the Airborne Instruments Laboratory
Automatic Noise-Figure Indicator, AIL Type 74. The connection from the IF output to the noise figure instrument must not disturb the normal operation of the test receiver. When such an instrument is not available, the noise figure may be measured by the first method, which is similar to the procedures described in IRE Standard 59 IRE 20, S1.

In the first method a diode noise generator is used for the receiver input. By determining the diode current $I_d$ and the noise power outputs of the receiver under test, the noise figure is computed from the formula

$$F = 20 \frac{I_d}{R_s} \left( \frac{1}{P_2/P_1 - 1} \right),$$

where $R_s$ is the frequency independent load resistance of the noise generator, $P_1$ is the receiver noise power output with the noise generator output at zero (except for its thermal noise at standard temperature) and $P_2$ is the receiver noise power output with the generator noise applied.

If $R_s$ is equal to 50 ohms and $P_2$ is equal to twice the value of $P_1$, the above equation reduces to

$$F = 1000 I_d.$$

If $I_d$ is measured in milliamperes, then

$$F = 10 \log_{10} I_d \text{ in decibels.}$$

This test should be performed at the 5 percent, 50 percent, and 95 percent frequencies in each band.

3.1.2. **Required Equipment**

1) Shielded Test Location
(2) Short connecting cables and adapters

(3) Noise generator or automatic noise figure instrument similar to AIL Type 74.

(4) True RMS voltmeter

3.1.3. Test Setup  See Figure 8.

3.1.4. Test Procedures  Method I

(1) Tune the receiver to the lowest standard test frequency. Control positions are standard, except AVC and AFC are off.

(2) With the noise generator output reduced to zero, measure the noise voltage at the output of the receiver.

(3) Increase the noise generator output until the noise voltage at the receiver output is increased 3 decibels.

(4) Measure the diode current in the noise generator, in milliamperes, and record as $I_d$.

(5) Calculate the average noise figure

$$F = 10 \log_{10} I_d.$$

(6) Repeat (1) through (5) at other standard test frequencies.

Method II

(7) Turn the automatic noise figure indicator (ANFI) and receiver on and allow at least 15 minutes of warm-up time.

(8) Tune the receiver to a standard test frequency.

(9) Connect the high and low voltage cables to the argon discharge noise generator, or plug in the noise diode head, whichever is provided with the instrument. Switch to the appropriate head and noise generator.
Figure 8. Block Diagrams of Noise Figure Test Setups for Receivers.
(10) Connect the output of the noise generator to the input of the receiver to be tested. The noise generator must be properly terminated.

(11) Connect the IF output of the receiver to the IF input of the ANFI.

(12) Press the noise generator AUTO NF ON and the meter indication 2nd DET CURRENT pushbuttons. Select either the gas or diode noise generator. The meter then reads the rectified second-detector current at the output of the IF amplifier. The meter should read 0.5 ma, indicating proper AGC action.

(13) Press the noise generator AUTO NF ON and meter indication NF RANGE HIGH or LOW pushbuttons, and read the noise figure on the meter.

(14) Repeat the procedures at other test frequencies.

3.1.5. Presentation of Test Data The noise figure test data are presented in tabular form as shown below.

<table>
<thead>
<tr>
<th>Frequency (Mc)</th>
<th>Noise Figure (db)</th>
</tr>
</thead>
</table>

3.2. Sensitivity

3.2.1. General Good sensitivity indicates that all amplifiers are performing normally, and that the alignment and tracking are satisfactory. Poor sensitivity on all bands indicates loss of amplification due to tube failure or loss of operating voltage at some point, providing that the alignment is satisfactory.

This test provides a reference point for level comparisons in the spurious response and intermodulation tests in addition to producing the receiver sensitivity characteristics. The standard response for FM receivers is a 20 db quieting output ratio. However, a 6 db quieting ratio response
is also desirable. Since most signal generators have power outputs less than 0 dbm, insufficient selectivity and spurious response data may be obtained using the 20 db quieting ratio.

This measurement is taken at all standard test frequencies. The measuring point is the center of the passband.

3.2.2. Required Equipment

(1) Signal Generator
(2) Frequency Meter
(3) Voltmeter or Output Indicator
(4) Screen Room
(5) Connecting Cables

3.2.3. Test Setup See Figure 9.

Figure 9. Block Diagram of Sensitivity Test Setup for Receivers.
3.2.4. Test Procedures

(1) Tune the receiver to a standard test frequency in accordance with the receiver technical manual.

(2) Tune a CW signal to the center of the receiver tuned frequency passband.

(3) Adjust the signal generator level to obtain 6 db of quieting at the receiver output. Record the generator level.

(4) Repeat (3) for 20 db of quieting.

(5) Repeat (1) through (4) at other standard test frequencies.

3.2.5. Presentation of Test Data The data are present in the following tabular form.

<table>
<thead>
<tr>
<th>Band</th>
<th>Frequency (Mc)</th>
<th>Power Input (6 db quieting) (dbm)</th>
<th>Power Input (20 db quieting) (dbm)</th>
</tr>
</thead>
</table>

3.3. Selectivity

3.3.1. General The selectivity as measured by this test shows the response at frequencies slightly removed from the tuned frequency. In general the selectivity characteristics are determined by the IF amplifier tuned circuits. The selectivity is also a measure of the receiver's ability to discriminate against off-channel radiations, and in reality is a measure of the receiver's passband characteristics. An entire passband curve is usually needed to determine the center tuned frequency since double or triple peaked IF curves are common for wideband receivers. This information is helpful in the spurious response test to determine the exact identification of a spurious response.
The most important part of this test is that of frequency measurement. For this reason the frequencies should be measured with an accuracy of at least 1 part in $10^6$.

This measurement may be taken at the IF, and the signal may be coupled in after the mixer stage. Special adapter circuits are usually necessary for this arrangement, and care must be taken to ensure that auxiliary circuits do not disturb the normal IF characteristics.

3.3.2. Required Equipment

Same as Section 3.2.2.

3.3.3. Test Setup Same as Figure 9.

3.3.4. Test Procedures

(1) Tune the receiver to the mid test frequency in a band.

(2) Apply a CW signal at the antenna or IF input terminals. Tune the signal generator frequency to approximately the mid-frequency of the passband.

(3) Adjust the signal generator output level to obtain 6 db of quieting at the receiver output. Record the generator output level and frequency.

(4) Move the generator frequency in small increments, and at each increment record the generator output level necessary to obtain 6 db quieting until the center passband characteristics are determined.

(5) Obtain the outside 3, 6, 10, 20, 30... db response points until the signal generator output limits are reached.

(6) Plot a smooth curve of response in db versus frequency to check for gross errors in measurements.

3.3.5. Presentation of Test Data The data are presented in tabular and graphical form.
3.4. Spurious Responses

3.4.1. General The receiver spurious responses, a major source of interference, are involved, directly or indirectly, in nearly every type of interference that has been observed in the laboratory.

In general, the majority of spurious responses are due to the first mixer and may be calculated by

\[ F_{SR} = \left| \frac{p f_{LO_1} \pm f_{IF_1}}{q} \right| \]

where \( f_{SR} \) is the frequency of the spurious response, \( f_{LO_1} \) is the first local oscillator fundamental frequency, \( f_{IF_1} \) is the first intermediate frequency, and \( p \) and \( q \) are positive integers (zero excluded, in the case of \( q \)).

The magnitude of the spurious response is in part a function of the RF amplifier filter circuits and in part due to the characteristics of the mixer. Unusually large responses may be caused by poor tracking in the RF amplifier tuned circuitry or by improper operation of the mixer. Other unusually large responses, especially at frequencies much higher than the receiver tuned frequency, are due to leakage around the tuned circuits and to spurious resonances.

Multiple-conversion receivers may have spurious responses due to the second and third conversion stages, but they are not usually as large as
those due to the first conversion stage and may often be neglected in inter-
ference studies. They are calculated, by the above equation, with the
proper local oscillator and IF frequencies substituted.

Spurious responses are ordinarily evaluated by a single-signal
generator test in the following manner. With the receiver tuned to one
of the standard test frequencies, the signal generator should be tuned
over a wide range of frequencies to discover receiver output responses
at frequencies other than the one to which it is tuned. These other fre-
quencies are called spurious responses. Each spurious response frequency
is measured and the signal level is adjusted to give some prescribed
output indication. The level of this signal may be compared to that level
of desired signal at the tuned frequency which produces the same output
indication. The ratio of these two levels, usually expressed in decibels,
is called the spurious response rejection ratio.

If calibrated attenuators are used in the transmission line from
the signal generator to the receiver, the attenuator frequency range
must be considered since it is easy to leave one attenuator in the line
during the entire test.

Wideband receivers usually have multi-peaked spurious responses
resembling the IF characteristics. If the spurious response frequencies
are measured accurately, an exact identification of each may be obtained.
The frequency measurement accuracy should be at least 1 part in $10^6$. The
tuned frequency should be measured occasionally for the possibility of
drift. The tuned frequency, local oscillator frequency, and IF should be
measured and recorded.
Low-pass filters which have at least 50 db of attenuation at the signal generator second harmonic should be utilized. Signal generator harmonics may cause erroneous identification of spurious response frequencies.

3.4.2. Required Equipment

(1) Same as Section 3.3.2.

(2) Signal generators covering the frequency range of 0.150 Mc to 44 kmc.

(3) NFIM

3.4.3. Test Setup Same as Figure 9.

3.4.4. Test Procedures

(1) Tune the receiver to a standard test frequency in a band.

(2) Measure and record the local oscillator and tuned frequencies using an NFIM, signal generator, and frequency meter.

(3) Tune the signal generator to the tuned frequency and adjust the generator output level to obtain 6 db of quieting. Record the level.

(4) Tune the generator away from the tuned frequency and increase the level to full output. Insert a low-pass filter with a cut-off frequency between the generator frequency and the second harmonic.

(5) Slowly tune the signal generator, beginning at 0.150 Mc until a frequency is found at which the receiver responds. Adjust the generator output to produce 6 db quieting. Record the frequency and level.

(6) Measure and record all responses in the range from 0.150 Mc to 44 kmc. A large number of responses are generally present near the tuned frequency, and care must be taken when measuring these responses.
(7) Identify each response. Responses which are not readily identified should be measured again.

(8) Repeat the procedures at other standard test frequencies.

(9) Correct the power levels for path attenuation to obtain the power level at the receiver antenna terminals.

3.4.5. Presentation of Test Data The data are presented in tabular form.

<table>
<thead>
<tr>
<th>Measured Frequency (Mc)</th>
<th>Calculated Frequency (Mc)</th>
<th>Identification (p, q and sign)</th>
<th>Power Input (dbm)</th>
</tr>
</thead>
</table>

3.5. Intermodulation

3.5.1. General The intermodulation characteristics of a receiver are of primary importance because they give an indication of the interference possibilities when the receiver is used in the presence of two off-channel signals. Assuming that these signals have not been mixed before arriving at the receiver, some mixing may be expected in the RF amplifier tubes and/or the first mixer. If one of the extraneous signals generated in this manner happens to fall at the tuned frequency with sufficient amplitude, interference of a cochannel nature is the result.

Usually, it is assumed that the third order mix is potentially the most serious type of intermodulation because both signals may be within the passband of the input circuits. The frequency relationships for this type of mix are given by

\[ f_o = 2f_a - f_b \]
where \( f_o \) is the receiver tuned frequency and \( f_a \) and \( f_b \) are the interfering signal frequencies.

It is also possible that higher order intermodulation products caused by two interfering signals near the tuned frequency could produce interference.

The second order, or primary, mix is defined by

\[
 f_o = f_b - f_a.
\]

Although one or both interfering signals must be far removed from the pass-band, they may still develop enough voltage at the grid of the first tube to produce strong interference. Therefore, all frequency ranges should be investigated for low RF attenuation points.

Care must be taken in conducting this test to insure that intermodulation does not occur within the signal generators themselves. If the signal generators are of a type such that generator intermodulation is a possibility, it is essential that a device be used to couple the generator outputs to the receiver which will provide considerable isolation between the generators. Filter techniques outlined in section 2.3.1 are extremely useful.

3.5.2. **Required Equipment**

(1) Frequency Meter.

(2) RF signal generators.

(3) Filters, isolation networks, etc.

(4) Voltmeter or output indicator.

3.5.3. **Test Setup** See Figure 10.
3.5.4. Test Procedures

(1) Tune the receiver to a standard test frequency.

(2) Tune \( f_a \) to the receiver tuned frequency. Record the level necessary to obtain a 6 db quieting ratio. The level should be corrected by subtracting the path attenuation to the receiver antenna terminal.

(3) If third order mixing is to be observed, tune slightly off the tuned frequency and increase the generator output to near maximum. Move \( f_a \) until receiver desensitization is not apparent. Record \( \Delta f = f_a - f_o \).

(4) Tune \( f_b \) to a frequency to cause receiver intermodulation and adjust the power outputs of \( f_a \) and \( f_b \) generators to equal levels to obtain a 6 db quieting ratio.

(5) Test for the possibility of signal generator intermodulation.

(6) Correct the power levels for path attenuation to the receiver antenna terminals. Record the corrected power levels and \( f_a \).

(7) Test for all third order mixing possibilities on both sides of the tuned frequency recording \( \Delta f \), power input, etc.

(8) If primary mixing is to be observed, set \( f_a = f_b - f_o \) and test for all possibilities on both sides of \( f_o \). Again \( \Delta f = f_a - f_o \).

(9) Test all other mixing possibilities, fifth order, etc., on both sides of \( f_o \).

(10) Repeat the procedures at other standard test frequencies.

3.5.5. Presentation of Test Data

The data are presented in tabular form. A graph of Power Input versus \( \Delta f \) for each intermodulation order is desirable.

<table>
<thead>
<tr>
<th>Intermodulation Order</th>
<th>( \Delta f ) (Mc)</th>
<th>Power Input (dbm)</th>
</tr>
</thead>
</table>

-39-
4. Pulse Position Multiplexed Terminal Equipment Tests and Test Procedures

The two tests and test procedures developed for pulse position terminal equipments determine the bandwidth and crosstalk information.

The following definitions and standards have been adopted specifically for the tests described in this section. The noise and signal sources used in these tests should have an output impedance of 600 ohms and be capable of producing an output level of zero dbm. The indicator for these tests is the multiplexed test set oscilloscope. The standard pulse deviation is that deviation in microseconds produced as the result of a 1000 cps modulating signal at rated input level for the equipment.

4.1. Bandwidth

4.1.1. General The bandwidth of the multiplexer as measured by this test indicates the level of any given audio frequency required to produce the standard pulse deviation in microseconds. This bandwidth is determined by the passband characteristics of the pulse position modulator of the channel under test. The upper audio frequency limit will ultimately be determined by the multiplexer pulse repetition rate regardless of the modulator passband characteristics.

4.1.2. Required Equipment

(1) Audio Signal Generator.

(2) Connecting Cables.

(3) Multiplexer Test Set or Equivalent.

(4) Audio frequency voltmeter.

4.1.3. Test Setup The setup should be as indicated in Figure 11.
Figure 10. Block Diagram of Intermodulation Test Setup for Receivers.

Figure 11. Block Diagram of Test Setup for Bandwidth Test of Terminal Equipments.
4.1.4. Test Procedures

(1) Place the multiplexer in operation with master output to RF connected to master input from RF with designated length of cable.

(2) Connect the test set probe to the multiplexer output pulse terminal and adjust the test set for observance of the pulse belonging to the particular channel being tested. Adjust the time base to a convenient but known number of microseconds per division. The pulse should lie with leading edge on center vertical scribe.

(3) Set the audio generator to 1000 cps at rated input level to the multiplexer audio input terminals for the channel under test. Read and record the pulse deviation in microseconds from the original unmodulated position.

(4) Increase the audio generator level by 3 db and tune toward a lower frequency until the pulse deviation is again standard. Record the frequency, level, and pulse deviation.

(5) With this same level, tune toward a frequency higher than 1000 cps until the pulse deviation is again standard. Record frequency, level, and pulse deviation.

(6) Repeat (4) and (5) for a 6 db increase in generator output.

(7) With rated generator output, sweep the established bandwidth, stopping at the frequency of greatest pulse deviation. If this frequency differs from 1000 cps, set the audio generator level for a standard pulse deviation and record frequency, level, and deviation.

4.1.5. Presentation of Data The data are presented in tabular form as shown below:

<table>
<thead>
<tr>
<th>Standard Deviation (μ-sec)</th>
<th>Frequency (cps)</th>
<th>Level (dbm)</th>
</tr>
</thead>
</table>

-42-
4.2. Crosstalk

4.2.1. General  Crosstalk as measured by this test indicates whether or not signals modulating a given multiplex channel spill over into adjacent or alternate channels. The test is qualitative in nature and only indicates whether or not crosstalk exists in the multiplex equipment. The test is frequency independent since the modulated channel is fed by a white noise or band-limited white noise signal in the audio range of the multiplexer.

4.2.2. Required Equipment

(1) Noise Generator.

(2) Connecting Cables.

(3) Multiplexer test set or equivalent.

(4) Audio frequency voltmeter.

4.2.3. Test Setup  The test setup is the same as for the bandwidth test with the noise generator substituted for the signal generator.

4.2.4. Test Procedures

(1) Arrange the equipment as in (1) and (2) in Section 4.1.4 except that the pulse display should include the adjacent and first alternate channels on each side with the modulated pulse centered.

(2) Set the noise generator to the rated input level and observe that the modulated channel pulse is being deviated to about the standard deviation.

(3) Examine the pulses of the adjacent and alternate channels. If noise or jitter is seen to occur in the position of these pulses when the noise generator is on, there is some crosstalk present. Record deviation in $\mu$-sec if readable or photograph pulse display.
4.2.5. Presentation of Test Data The data are presented in tabular form as shown below or with photographs showing the desired information.

| Standard Deviation (μ-sec) | Adjacent Channel Deviation Right (μ-sec) | Left (μ-sec) | 1st Alternate Channel Deviation Right (μ-sec) | Left (μ-sec) |

5. System Tests

The only system test which has been completely developed and checked is the error rate test. Since the overall philosophy and circuitry involved in this test have not been reported earlier, this section differs from the format and content of the preceding sections, which describe tests and test procedures for transmitters, receivers, and terminal equipments.

Section 5.2 contains a discussion of additional proposed system tests utilizing a speech system test set and various types of interference.

5.1. Error Rate

5.1.1. General It is desirable from the standpoint of system vulnerability to be able to measure the effects of various types of interfering signals on a particular system. It may be found that some systems are more vulnerable to a given type interference than others are. It would be desirable to know how the various types of interference degrade a given system's performance.

In the ultimate situation, a system would be checked in an actual operating setup using the antennas and other associated equipment necessary
for complete system operation. To develop a system test, it is desirable to conduct the experiment in the laboratory. It is more convenient to eliminate the antenna systems and use a coaxial connection between the transmitter and receiver and include the attenuation necessary to provide the proper signal level at the receiver. Interfering signals can be introduced at the receiver terminals at reasonable signal levels for the tests. Although there are system characteristics which these procedures eliminate such as antenna pattern effects and propagation, etc., this is a good first approximation to an operational system.

There are basically two types of multiplex systems. One is a frequency division system wherein the various channels are stacked in the base-band in frequency. That is, channel one would occupy a given frequency band and channel two would occupy the next adjacent frequency band and so forth. The other multiplex system makes use of time sharing and is known as time division multiplexing. In a time division system the input channels to be transmitted are sampled, usually at regular intervals, and information from each of the input channels is transmitted only a fraction of the total time. There are sampling requirements that must be met regarding the maximum channel frequency that can be transmitted with a given sampling rate and the video bandwidth requirements for the system.

There has been a great deal of work performed on interference associated with the frequency division type systems. In particular, frequency division multiplex-frequency modulation systems have been studied. This is the system used primarily by the telephone company in its microwave relay systems and hence a large amount of effort has been expended on it.
One of the principal sources of interchannel interference in a frequency division multiplex-frequency modulation or phase modulation system is the variation of attenuation and phase shift of the transmission path with frequency. An echo produces a simple form of such a variation. Echoes can arise from reflections in the equipment or multipath transmission. Tests performed on this type system make use of a random noise modulating signal. Random noise of appropriate bandwidth and power adequately simulates a composite speech signal composed of many telephone channels. For studies involving interchannel interference, the noise energy corresponding to some particular telephone channel is removed from the broadband noise modulating voltage. When such a wave is impressed on the frequency modulator and the resulting FM wave is transmitted, detected and finally demodulated, the received output in the originally clear channel represents interchannel interference. By comparing the reception level with and without the band-elimination filter, the distorting signal-to-desired signal ratio is established.

The system that has been available for evaluation is a time division multiplex-frequency modulation system. The time division multiplexing is further described as pulse position modulation, wherein the position in time of a particular channel pulse is made to vary in accordance with a modulating voltage. The time of occurrence of the channel pulse is advanced or retarded relative to its no modulation position as a function of the amplitude and polarity of the input channel signal at the instant that this signal is sampled.
In order to perform a test on such a system, there must first be some scheme of detecting any degradation to the transmitted information. An approach that has been tried is to count the number of errors made in a telegraph type signal in a given time with a known RF signal level and a given RF interfering level applied to the receiver antenna terminals. To inject a telegraph type signal into the multiplexing equipment described above which is used to multiplex voice frequency channels (that is a channel with a bandwidth of 300 to 3000 cps) a TH-5/TG telegraph terminal is used. This terminal takes a telegraph type signal such as a teletype signal and converts the mark, or circuit closed condition, to a frequency of 1325 cps and the space, or circuit open condition, to a frequency of 1225 cps. The TH-5/TG also performs the inverse operation of reconversion of the voice-frequency signals to a dc signal capable of driving a teletype system.

5.1.2. Test Setup and Discussion

Figure 12 is a block diagram of the complete system for the error rate test setup. The signal source used for the test was a square-wave generator which provided a simple on-off binary test signal. The period of this signal was comparable with the mark space period of a 60 wpm teletype signal.

The comparison of the output pulses from such a system to those at the input would be very simple if there were no delay introduced by the system itself. The fact is that there is delay in any system. In particular in this system, there is delay intentionally introduced in the telegraph terminal equipment because of carrier suppression and auxiliary
Figure 12. Block Diagram of Error Rate Test Setup.
circuits. The use of a simple "exclusive or" circuit which produces no output if both of its inputs are in the same state and produces an output when its inputs are in different states would suffice in the case where there is no system delay. In the practical case that is encountered, the "exclusive or" circuit cannot be used without additional refinements because the system delay introduces an error indication at the beginning and end of any pulse passed through the system, providing that the time delay is less than the pulse width which is the situation in this case. Additional auxiliary circuits, such as trigger shaping circuits, delay circuits, and gating circuits, are resorted to in order that both channels can be sampled simultaneously at a time when the states of the signal in each channel would be the same if no system errors are produced.

The sampling gate is generated by deriving a set of triggers for a monostable delay circuit from the input pulse signal to the telegraph terminal equipment. Figure 13 is a block diagram of the circuitry involved. A low-pass filter is used in the signal channel to remove high frequency transients from the circuit and preclude the possibility of false gates. The Schmitt circuit is used to generate a waveform with sharp edges so that the process of differentiation produces usable triggers. The polarity of the triggers produced is made the same in the phase inverter and rectifier. Delay is provided by the monostable multivibrator which is triggered each time the input signal changes states. A second monostable multivibrator is used as a gate generator. Delay through the circuit and the gate width can be controlled by controlling the period of
Figure 13. Block Diagram of Sampling Device.
the two monostable multivibrators. Since the two output gates are driven
from the same gate generator their outputs must occur simultaneously. The
outputs from these gates are supplied to the "exclusive or" circuit in
which the detection of a missing pulse is made. The number of errors
produced is counted with a digital counter.

The schematic diagram in Figure 14 shows the method used to key the
TH-5/TG telegraph terminal equipment which is designed to work with a set
of teletype contacts plugged into the send jack. In order to key this
circuit electronically a resistive adder is used. The positive voltage
from the supply in the TH-5/TG keeps the input tube conducting and results
in the output of the TH-5/TG being in the mark condition. When the teletype
contact opens, or in this case when the square wave generator output goes
negative, the input tube V12 is cut off and the output of the TH-5/TG is
in the space condition. The square wave generator used has an output
which goes from zero volts or ground to a negative voltage of approxi-
mately 60 volts.

Figure 15 is a schematic diagram of the delay and sampling circuitry
used to eliminate the delay problems introduced by the system under test.
Input 1 is connected to the output terminals of the square wave generator
used to key the TH-5/TG telegraph terminal. It is this signal that is
used to trigger the delay and gating circuits. From input 1, the signal
is passed through a low-pass filter composed of a series 10K ohm resistor
and shunt 0.05 µf capacitor. This filter is provided to remove any transients
which may be present at the input and eliminate the possibility of these
transients producing a false output. A Schmitt trigger circuit, composed
Figure 14A. Square Wave Generator Input Circuit to TH-5/TG Telegraph Terminal Equipment.

Figure 14B. Simplified Schematic of Resistive Adder.
Figure 15. Schematic Diagram of Sampling Device.
of transistors Q1 and Q2, is then used to produce a waveform with sharp edges. The level at which this circuit triggers is set by the variable potentiometer in the base circuit of the first transistor. Differentiation of the output waveform of the Schmitt circuit is accomplished by the RC coupling combination into transistor Q3. A phase inverter, Q3, is used to drive the rectifiers D2 and D3 and produce a unidirectional pulse for each transition of the input waveform. These pulses are then used to trigger a monostable delay multivibrator composed of Q4 and Q5. The length of the fixed delay produced by this circuit is adjusted to provide sampling when both input 1 and input 2 are in the same state. Triggers are derived from the output of the delay multivibrator to trigger the gate multivibrator. The gate length is fixed and is adjusted to provide a sampling period comparable with system delay considerations. The gate multivibrator is again a monostable multivibrator composed of transistors Q6 and Q7. Gating pulses from this circuit are coupled to the two gates Q8 and Q9. The signal from input 1 is applied to the gate composed of Q8. Signals from input 2, after passing through a low-pass filter, are applied to the gate composed of Q9. Inputs 1 and 2 are therefore sampled simultaneously. Gated signals from the gates are passed through emitter followers Q10 and Q11 to drive an "exclusive or" circuit composed of transistors Q12 through Q15. The "exclusive or" circuit is described in detail in the literature. Outputs produced by the "exclusive or" circuit represent errors and are applied to a digital counter. A low-pass filter is used at the input to the counter to eliminate the possibility of any transients on the "exclusive or" circuit output waveform from triggering the counter.
falsely. The power supply for the electronic circuitry described is shown in Figure 16.

The missing pulse detector was originally intended to work with a teletype signal. However, it was found that, because of the five digit code plus a start pulse and a stop pulse, this original circuit did not yield an error conditions everytime the output printer produced errors. This was due to the sampling pulse being produced after a transition of state in the input signal. A teletype signal can remain in one state for several code intervals without a transition in between intervals. Because of this, the simple gate generator circuit as it stands is not suitable for counting all the errors in a teletype signal. It could be made to do so by deriving synchronizing pulses from the teletype equipment at the expense of additional complexity.

5.1.3. Conclusions

Some flexibility is possible in the error detection system with the binary type input signal. The length of the pulse can be lengthened to allow for longer delays. The gate length of the pulse can be increased to provide a better gate length to pulse length ratio and therefore a relay or similar device could be operated by the pulse from the demultiplexing equipment. Lengthening the pulse is limited by the time available for observation if a certain probability and confidence limit is placed on the results. Any requirements that necessitates more observation time must be weighed against the ability of the test equipment to maintain the desired output conditions over the period of time required.
Figure 16. Power Supply for Error Detector.
To date, the system has been used with cochannel interference in a TDM-FM system with telegraph type signals. As expected with an FM system, the output error rate is zero until the interfering RF cochannel signal amplitude approaches that of the system signal. The error rate then rapidly changes from zero to 100 percent errors as the signal level is increased several decibels. It appeared that as the frequency of the interfering signal was changed in the passband of the receiver that the amount of interfering signal level required to produce a given error rate changed. Whether this is due to passband characteristics of the receiver or some aspect of the multiplexing system bears further investigation.

It might be assumed at first that the number of errors produced by such a system is the total number of binary digits, on or off operations, in the time interval measured. Care must be exercised, however, because if the multiplex system fails to produce an output signal, the telegraph terminal equipment is designed to force its output conditions to be a mark. Thus under such circumstances, the maximum measurable error rate would be one-half the maximum possible error rate.

It is recommended that a noise loaded test as used with frequency division multiplexing be used. The test would be on a telephone channel basis rather than on a telegraph channel basis.

5.2. Proposed System Tests

The General Electronics Laboratories Speech Systems Test Set can be used in further evaluation of the AN/TRC-29 and its associated terminal equipments. The General Electronics Laboratories, through a Signal Corps contract, developed the GEL Speech Systems Test Set to replace, in special
cases, a group of trained human listeners. This set measures the degradation in understandability of a speech sample that passes through a noisy intelligence channel by computing a "pattern correspondence index" (PCI). The instrument is designed such that the PCI is monotonically related to an articulation score obtained under identical conditions for a wide variety of interference types. The test set is intended to serve as a standard listener capable of providing synthesized listener scores for the types of interference with which microwave systems can be tested. The types of interference which can be used include cochannel CW, pulsed, and FM signals.

B. Mixer Testing and Evaluation

1. Introduction

The purpose of this study relates to finding means of reducing or eliminating spurious responses due to unwanted mixer actions or product formation within the mixer stages of receiving equipments. In order to eliminate these unwanted products, the nature of their occurrence must be thoroughly investigated through the testing and analysis of presently available mixer configurations. After a complete understanding of the formation of spurious products is established, mixer design can be altered to provide better rejection to unwanted products.

Considerable work toward this goal had been performed by other organizations prior to the studies described here. Previous studies, however, have not encompassed a large collection of data nor have they experimentally demonstrated a basic nature of spurious product occurrence for all the known means of product generation.
The studies thus far have included the taking of data for a grounded grid mixer of the type found in the 225 to 400 Mc tuning head of the AN/TRC-24, a single 1N82A diode mixer described in Figure 17, and an experimental mixer configuration using a pair of triodes as infinite input impedance diode rectifiers. The data thus far assembled have demonstrated the need for more data on the earlier types tested and for confirming data from other mixer types yet to be built and tested. The latter is chiefly to ascertain, if possible, that all mixer types form spurious products via the same means, thus establishing a basic theory common to all mixer types.

The study to date indicates that most of the significant spurious responses are generated exclusively within the mixer active element and are not produced by harmonics of either input signal as injected by the signal source. This fact had been established through previous research and was known prior to this study. Harmonics present in the input signal cause troublesome spurious responses only when the mixer generated responses are extremely low or when the level of the injected harmonics approaches that of the signal fundamental. This study has led to a theoretical method of response prediction which may prove to be a practical tool.

2. Experimental Results

In addition to the verification of previous experimentation, the study indicates that there is little if any correlation between the level of a given spurious response and the levels of the harmonics which could combine to form the spurious response even as measured directly within the circuit loop containing the active element. Verification of this phenomenon has not yet been thoroughly established through experimentation but the
Figure 17. Schematic Diagram of 1N82A Test Mixer.
theory that such harmonics are not the significant source of spurious responses is supported by mathematical analysis.* Substantiating data, taken as a sample, may be seen in Table I. Harmonic substitution in the first mixer types tested also indicates that there is no correlation in the above mentioned levels.

The data of Table I were taken on the single 1N82A diode mixer having no tuned circuits. The test was made with the local oscillator fixed at 10 Mc and +13 dbm and with the difference frequency of 1 Mc held to a standard NFIM response. Another NFIM was used to measure the levels of the desired harmonics. The diode was held at zero bias. The local oscillator harmonic level was not recorded in this sample test.

The test setup used to obtain the data of Table I requires two signal generators and two NFIM receivers. The first signal generator is tuned to the selected local oscillator frequency and the level is set at +13 dbm and held constant. The test is conducted by choosing a given local oscillator harmonic, tuning the second signal generator to this harmonic frequency plus or minus the desired difference frequency, and adjusting the level of the second generator to obtain a standard response at the difference frequency. This is the signal input level recorded in the first column of data. The NFIM is then used to measure the level of the signal from the second generator at the input and output sides of the 1N82A and these data appear in the second data column of Table I under $q = 1$.

*See Appendix, page 117.
### TABLE I

**Record of Harmonic Level for Standard IF Response**

<table>
<thead>
<tr>
<th>LOCAL OSCILLATOR AT 10 mc, +13 dBm</th>
<th>BNC NO. 2</th>
<th>BNC NO. 4</th>
<th>BNC NO. 2</th>
<th>BNC NO. 4</th>
<th>BNC NO. 2</th>
<th>BNC NO. 4</th>
<th>BNC NO. 2</th>
<th>BNC NO. 4</th>
<th>BNC NO. 2</th>
<th>BNC NO. 4</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>P=2</strong></td>
<td></td>
<td></td>
<td></td>
<td></td>
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<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>BNC NO. 2</td>
<td>75</td>
<td>+30</td>
<td>88.5</td>
<td>-12</td>
<td>NSR</td>
<td>NSR</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>BNC NO. 4</td>
<td></td>
<td>+27</td>
<td></td>
<td>-6</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td><strong>P=3</strong></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>BNC NO. 2</td>
<td>54</td>
<td>+20</td>
<td>96</td>
<td>+21.5</td>
<td>&gt;101</td>
<td>NSR</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>BNC NO. 4</td>
<td></td>
<td>+10</td>
<td></td>
<td>+7</td>
<td></td>
<td></td>
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<td></td>
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<tr>
<td><strong>P=4</strong></td>
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<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>BNC NO. 2</td>
<td>77</td>
<td>+38</td>
<td>90</td>
<td>+13.5</td>
<td>&gt;101</td>
<td>NSR</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>BNC NO. 4</td>
<td></td>
<td>+30</td>
<td></td>
<td>-1</td>
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<td><strong>P=5</strong></td>
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<td>59</td>
<td>+20</td>
<td>96</td>
<td>+13.5</td>
<td>100</td>
<td>NSR</td>
<td></td>
<td></td>
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<td></td>
</tr>
<tr>
<td>BCN NO. 4</td>
<td></td>
<td>+11.5</td>
<td></td>
<td>-2</td>
<td></td>
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<td></td>
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</tr>
<tr>
<td><strong>P=6</strong></td>
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<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>BNC NO. 2</td>
<td>76</td>
<td>+40</td>
<td>89</td>
<td>+6.5</td>
<td>&gt;101</td>
<td>&gt;101</td>
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<td>BNC NO. 4</td>
<td></td>
<td>+31</td>
<td></td>
<td>-10</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

NSR = No Standard Response Was Obtained
The second signal generator is then tuned to a subharmonic of the frequency just previously selected, thus making \( q = 2, 3, \ldots, n \), and the level adjusted for the same standard response. This is the signal input level recorded in the 3rd, 5th, etc. columns of data. Again the NFIM is used to measure the same frequency as selected previously, i.e. that frequency corresponding to \( q = 1 \), or that frequency which is different from the chosen local oscillator harmonic by the desired difference frequency.

If the responses were due to mixing the frequency corresponding to \( q = 1 \) and the associated local oscillator harmonic, the level of the former, for a constant \( p \), should be constant for any value of \( q \) that is chosen. The data sample of Table I does not exhibit a constant level at the frequency corresponding to \( q = 1 \) for any local oscillator harmonic tested. The theory that spurious product generation is the direct result of the two fundamental frequencies caused by the nonlinear element is therefore verified. The harmonics generated in the process are simply by-products of the same nonlinearity and may or may not form products. Such products, if produced, would be so overshadowed by those formed in the direct process that they would be insignificant.

Another factor favoring this theory is seen in Figure 18. A high correlation exists between the relative level of a given spurious response and that of the local oscillator harmonic generated by the same terms in the series expansion for the nonlinear element of the mixer. It is noted, for example, that as the diode bias is varied, the changes in the relative levels of both the desired response and the second harmonic are nearly identical. This is true for the response \( (p = m, q = n) \) and
Figure 18. Harmonic and Response Levels Versus Forward Bias for Single 1N82A Diode Mixer with No Frequency Discrimination.
for the \((m + n)\)th harmonic in general. This phenomenon has been observed in prior experimentation with vacuum tube mixers and the same theoretical conclusions were deduced. Differences in inflection points of the correlated curves are probably due to several factors, one being that of equipment limitations.

The method followed in obtaining the data may cause variations in the position of such inflection points since the harmonic data were taken using only one input signal, that of the local oscillator. A close look at the series representation discloses that the coefficients of each of the higher ordered terms are altered by the injection of the second signal. In many cases, the value of the higher ordered terms is not negligible and the variation caused by the presence of the second signal may bring about some of the inflection point differences noted.

For example, with one input, the expression for the second harmonic is:

\[
E_1 \cos 2\omega t = \left[ \frac{1}{2} e_a^2 K_2 + \frac{1}{2} e_a^4 K_4 + \frac{15}{32} e_a^6 K_6 + \frac{7}{16} e_a^8 K_8 + \ldots \right] \cos 2\omega t
\]

and with two inputs, the expression is:

\[
E_2 \cos 2\omega t = \left[ \frac{1}{2} e_a^2 K_2 + \left( \frac{1}{2} e_a^4 + \frac{3}{2} e_a^2 e_b^2 \right) K_4 + \left( \frac{15}{32} e_a^6 + \frac{7}{16} e_a^8 + \frac{105}{8} e_a^4 e_b^4 \right) K_6 + \left( \frac{105}{8} e_a^2 e_b^2 + \frac{35}{3} e_a^2 e_b^6 \right) K_8 + \ldots \right] \cos 2\omega t.
\]
The difficulty in obtaining good correlation may be resolved by establishing the two input frequencies, holding their levels constant, and measuring both the harmonic content and the level of the signal produced by the difference in the two frequencies for each change in the variable parameter of the test. All tests to date have been performed by establishing a signal input level sufficient for a standard response and then plotting the data for presentation as if the input had been held constant and the IF response measured. In this process, the signal input level is not a constant. If the harmonics were measured with the signal required for the standard response present, the result should be the same as for the first proposed solution.

In regard to the errors created by measuring harmonics without the second signal present, it is noted that with the standard response type test, the second signal is at a maximum in the nulls of the responses and thus in the nulls of the harmonics. It is at these points that the correlation is the poorest.

In the past, it has been assumed that the coefficients of the terms for higher orders of curvature are negligible when demonstrating that the harmonic curves and response curves should bear the above relationships to one another. Experimental data demonstrate that such is not always the case because of the effects due to the angle of conduction which cause both harmonics and responses to null at certain values of conduction angle. A null of this type results in all higher ordered harmonics becoming greater than the one which vanished and under former assumptions would indicate that the primary generating term of the nulled harmonic had vanished.
An inspection of the terms generating any given harmonic will show that if the primary generating term did indeed vanish, the higher orders would still produce the harmonic. This is true because the higher ordered harmonics are abundantly present and the associated constant coefficients are larger for the lower ordered harmonic in most cases. The inconsistency in analysis can only be overcome by assuming that the harmonic vanishes because of out-of-phase addition of the various terms contributing to the vanishing harmonic. This fact, however, does not invalidate the theory since it is noted that the series for any given response and that for the corresponding harmonic contain terms of like order having a like number of similar coefficients relating to the input levels.* The slight differences in the coefficients may also be a factor contributing to differences noted in the correlated curves.

The variations in level of the harmonics of the local oscillator for changes in the bias on the mixing diode are plotted in Figures 19 and 20. The significant range of positive and negative bias is covered. Note that any given harmonic falls into sharp nulls across the full range of bias values and that the number of these nulls is two less than the order of the harmonic. An exception is noted in the 8th through 10th harmonics where it is presumed that the sensitivity was not sufficient to determine the last nulls of each, and in the sixth where an additional null was discovered at high forward bias, making the number of nulls only one less than the harmonic order.

* See Appendix, page 117.
Figure 19. Harmonic Level Versus Reverse Bias for Single 1N32A Diode Mixer with No Frequency Discrimination.
Figure 20. Harmonic Level Versus Forward Bias for Single 1N82A Diode Mixer with No Output Frequency Discrimination.
This phenomenon is characteristic of the variations in harmonic generation due to changing the conduction angle of a sinusoidal input, a condition which exists within a diode mixer when the diode bias is altered. A mathematical analysis of the effect on harmonic content due to conduction angle demonstrates this phenomenon. As an example, suppose that current flows in the diode for an angle \(2\phi\) as shown in Figure 21, then the amplitude of the \(p\)th harmonic component, \(C_p\), can be obtained from:

\[
C_p = \frac{1}{\pi} \int_{-\phi}^{\phi} i_p(\theta) \cos p\theta \, d\theta = \frac{2}{\pi} \int_{0}^{\phi} i_p(\theta) \cos p\theta \, d\theta.
\]

If the diode characteristic in the conducting region has a polynomial representation of degree \(K\), then for a pure cosine driving signal the plate current:

\[
i_p(\theta) = \sum_{n=0}^{K} b_n \cos^n(\theta) = \sum_{n=0}^{K} a_n \cos(n\theta).
\]

Substituting this expression for \(i_p\) in the first expression:

\[
C_p = \frac{2}{\pi} \int_{0}^{\phi} \sum_{n=0}^{K} \left[ a_n \cos(n\theta) \right] \cos p\theta \, d\theta.
\]

Expanding the above expression inside the integral:

\[
C_p = \frac{2}{\pi} \int_{0}^{\phi} a_p \cos^2 p\theta \, d\theta + \frac{2}{\pi} \sum_{n=0}^{K} \frac{1}{2} a_n \left[ \cos(n-p)\theta + \cos(n+p)\theta \right] d\theta.
\]
Figure 21. Characteristic of Anode Current in Diodes for Small Conduction Angles.
Separating the expression into two terms permits evaluation of the case where \( n = p \). Carrying out the indicated integration:

\[
C_p = \frac{2}{\pi} \left\{ \frac{a_p}{2} \sin 2p \phi + \frac{a_p}{4p} \sin 2p \phi \right\} + \frac{2}{\pi} \sum_{n=0}^{K} a_n \left[ \frac{\sin(n + p) \phi}{2(n + p)} + \frac{\sin(n - p) \phi}{2(n - p)} \right].
\]

By choosing the harmonic and the degree of the polynomial representing the current pulse, it is possible to determine the amplitude of this harmonic for any value of the conduction angle. Because of the number of variables involved, a graphical solution is a more practical approach and provides a curve which can be related to the experimental data.

An example for the case of the third harmonic with a third-degree polynomial representation of the diode forward characteristic:

\[
i_p = 10^{-1} + 10^{-1} \cos \phi + 10^{-2} \cos 2\phi + 10^{-3} \cos 3\phi
\]

will yield the following expression for the third harmonic:

\[
C_3 \approx \frac{10^{-1}}{12 \pi} \left[ 3 \sin 4\phi + 6 \sin 2\phi + 8 \sin 3\phi \right]
\]

where \( p = 3 \), \( K = 3 \), \( a_0 = 10^{-1} \), \( a_1 = 10^{-1} \), \( a_2 = 10^{-2} \), and \( a_3 = 10^{-3} \).

Figure 22 is a graphical representation of the expression for \( C_3 \). \( C_3 \) is represented by the solid curve and is the summation of the three dashed curves. Since the data obtained on the mixer shown in Figures 19 and 20 are for the absolute value of the harmonic in decibels, a better relation can be observed if the curves for \( C_3 \) are drawn without regard for sign.
This curve is displayed linearly in Figure 23 and logarithmically in Figure 24. The curve in the latter display shows great similarity to the 3rd harmonic curve of Figures 19 and 20.

Zero conduction angle corresponds to large reverse diode bias where both the mixer third harmonic and $C_3$ are seen to be at a null. As the conduction angle is increased, both the harmonic and $C_3$ go through a maximum and into a second null. Further increase in conduction angle carries the diode into forward bias operation where the two curves of interest pass through another maximum and then another null. Notice that the larger values of conduction angle (forward bias) produce very little 3rd harmonic and that the small nulls at the high values of conduction angle are lost. The diode at high forward bias acts like a closed switch and the third harmonic content of the input signal feeds through and obscures the small nulls.

Figures 25 and 26 show the responses which correspond to the harmonics of Figures 19 and 20. It is obvious that the responses follow the same null pattern as the harmonics for like orders of generation. Although no analysis has yet been accomplished showing that conduction angle affects response products in the same manner as the harmonics are affected, it is expected that such analysis will yield similar results simply on the basis of the experimental correlation thus far obtained.

Figure 27 shows curves for response and harmonic data taken under the same conditions as those of Figure 18 except that the diode output was resonated at the intermediate frequency. The effect is a spreading and partial cancellation of the nulls obtained without the tuned circuit.
Figure 22. Graphical Solution of the Expression for $C_3$ or Third Harmonic Component of Diode Current.
Figure 23. Plot of Absolute Value of the Third Harmonic Component of Diode Current.

\[ |C_3| = \frac{10^{-1}}{12\pi} \left( 3\sin 4\phi + 6\sin 2\phi + 8\sin 3\phi \right) \]
Figure 24. Logarithmic Plot of Absolute Value of Third Harmonic Component of Diode Current.
Figure 25. Response Level Versus Reverse Bias for Single 1N82A Diode Mixer with No Frequency Discrimination.

L.O.: 10 Mc, +13 dbm.
RESPONSES ARE NUMBERED AT INFLATION POINTS WITH THE VALUE OF P.
Q = 1 FOR ALL RESPONSES SHOWN.

MEASUREMENT LIMIT: 12 db
Figure 26. Response Level Versus Forward Bias for Single LN82A Diode Mixer with No Frequency Discrimination.
Figure 27. Harmonic and Response Levels Versus Forward Bias for Single 1N32A Diode Mixer with Capacitor Across Output.
This is due to the change in effective diode bias at the IF rate since with the tuned circuit present, the IF voltage is developed in series with the diode bias supply. In the light of this fact, the vacuum tube mixers tested so far might yield better correlated data if the data were taken with the tuned circuits removed.

3. Mixer Improvement

An experimental mixer employing two triodes connected as high input impedance diodes was tested to determine its characteristics. This diode mixer was particularly poor in the response versus harmonic correlation since there was little correlation between any set except the desired response and the second harmonic. This is demonstrated with the curves in Figures 28 and 29. It should be further pointed out that this experimental diode mixer exhibits excellent rejection to spurious product formations as can be seen from the curves in Figure 30. With 40 volts on the cathode, the spurious responses are some 80 db down. A further breakdown and study of the twin high input impedance mixer in the light of the knowledge gained from the 1N82A mixer experiments still remain to be done. Such tests, however, should yield considerable insight into the design of spurious-response-free mixers.

4. Response Prediction

A prediction method has evolved from the mixer study in an effort to find a basis for the dependency of response levels on the value of both \((p + q)\) and \((q - p)\). The mixer study indicates that all responses are formed due to nonlinearities (except IF responses) which can be grouped into one polynomial expression of degree \(n\). The expansion of such a polynomial when the input variable is the sum of two sinusoidal signals yields a multiplicity
Figure 28. Comparison of Harmonic and Response Levels Versus Cathode Voltage for Balanced Infinite Input Impedance Diode Mixer.
NOTE: Curves plotted for best comparison between given harmonic and response. Harmonics nor responses are in proper relation to one another.

Figure 29. Comparison of Harmonic and Response Levels Versus Plate Supply Voltage for Balanced Infinite Input Impedance Diode Mixer.
Figure 30. Measurable Response Level Versus Cathode Voltage for Balanced Infinite Input Impedance Diode Mixer.
of terms derived through the binomial expansion of the input signals. If the binomial expansion coefficients are divided by \(2^{(n-1)}\), where \(n\) is the polynomial term order or the value of \((p + q)\), and are then arranged into Pascal's Triangle, each position in the triangle will represent the constant coefficient of one of the possible mixer responses. Such a triangle is shown in Figure 31. It is noted that each coefficient is precisely the one associated with the lowest order term generating a given response frequency.* Any horizontal line contains coefficients associated with one particular term of the mixer polynomial or value of \((p + q)\) and the order increases from top to bottom. About the vertical axis, the triangle is even-valued such that each column of coefficients is associated with a given value of \((p - q)\) or \((q - p)\) depending upon the direction from center. The 45-degree lines represent lines of constant \(p\) or \(q\) such that at any intersection is found the coefficient associated with the response \((p = x, q = y, \dagger)\), where \(x\) and \(y\) are the intersecting 45-degree line values. The outside 45-degree values are the harmonic coefficients and the coefficient at the apex is a dc term having no significance in this discussion.

Before discussing the use of this triangle, some facts about the triangle should be mentioned. The triangle demonstrates that the value of a response will depend upon both \((p + q)\) and \((p - q)\) since \((p + q)\) is the order of the primary generating term and \((p - q)\) indicates a variation of the constant coefficients associated with the generating term. Such a relationship has been previously suspected from experimental data.12

Secondly, the triangle is related only to the mixer activity and is

* See Appendix, page 117.
Figure 31. Pascal's Triangle of Binomial Coefficients Divided by $2^{(n-1)}$ where $n$ is the Power to which the Binomial is to be Raised.
independent of any preselection and filtering prior to the mixer. Furthermore, the triangle does not account for the mixer polynomial coefficient values nor for the relative levels of the local oscillator and the input signal.

These drawbacks, however, should be surmountable with a knowledge of the selectivity of the preselector and filters and perhaps some knowledge of the mixer characteristics. The method of obtaining the unknown parameters is covered in the following discussion.

If the values of the triangle are converted to decibels, causing the apex term to become infinitely attenuated since its frequency is zero, the chart of Figure 32 is generated. This chart or spurious response triangle may be used in conjunction with a few simple response measurements to determine the receiver mixer characteristics. Prior to this, however, a response or selectivity curve of the preselector must be obtained so that the input signal may be corrected for losses due to the pre-mixer filtering.

The constants, or variables as the case may be, which relate to the level of the local oscillator and the level of the response, hereafter designated as $e_p$ and $e_q$ respectively, are raised to the power of the respective $p$ and $q$ values for any given product as designated by the triangle. For example, the response $2, 2, \uparrow$ has the factors $\frac{3}{4} e_p^2 e_q^2 K_4$ associated with it.

For this reason, the ratio of any two adjacent products in any row of the chart will yield a known constant times the ratio of $e_p/e_q$ or $e_q/e_p$ and will thus be independent of the associated polynomial coefficient. In fact, the ratio of any two products in the same row is independent of the polynomial, but the ratio of $e_p/e_q$ will be raised to a power which equals the number of
Figure 32. Spurious Response Triangle for a Mixer Having a Polynomial Representation of the Tenth Degree.
spaces separating the two products forming the ratio. If the two products have the same spacing from the center of the chart, the constant factor will be unity.

If the signal input level to the mixer is held constant, any two responses obtainable in a given row will yield data reducible to the ratio of the effective $e_p$ and $e_q$ as presented to the mixer nonlinearity. Obviously the preselector characteristics must be known in order to correct the input levels to provide equal signals to the mixer at both responses of a given row. It is not necessary to maintain the fixed input level between rows since this factor may be corrected by forcing the data to yield an equal $e_p/e_q$ ratio in all rows.

After the response values have all been normalized to obtain identical ratios in all rows, the response values may be compared to their respective chart values and the values of $LK_n$ may be determined. In the case of the desired response, normalization is accomplished by setting its corrected input level to the same value as that used in the $LK_3$ row and then normalizing each value to the ratio established in the $LK_3$ row.

Once the values of $LK_n$ are established, the mixer attenuation to any response may be calculated. Adding the preselector attenuation at the frequency of the response produces the overall receiver attenuation to the response. When this value is normalized to the desired response, the result is overall receiver rejection to the spurious response in question.

If a table of the mixer polynomial coefficients for the receiver were available, the measurements could be reduced to taking only two spurious
response and one desired response measurements. With these measurements, the ratio can be established and the resulting calculated responses can be normalized with respect to the desired response, thus yielding the overall receiver rejection to any possible response.

The above method is, however, only theoretical and has not yet been verified through experiment. The approach is promising but further effort is needed to prove whether or not there are practical limitations to performing the indicated measurements and calculations.

5. Conclusions

The mixer studies have yielded valuable data regarding the primary causes of spurious responses. However, there is still no established theory relating the generation of spurious responses to the conduction angle. The experimental mixer, which exhibits low spurious response susceptibility, has not yet been fully investigated. Further study is required if the effort expended thus far is to culminate in some usable form for the reduction or elimination of spurious responses.

The prediction method resulting from this study has not been tested in a practical situation. Such testing is essential to both the method of prediction and to the verification of the theories evolved. Further study is required to establish the method through its utilization with receiving equipments already available.

C. Spurious Response Bandwidth and Rejection

1. Introduction

One of the most serious forms of interference associated with heterodyne receivers is that resulting from spurious responses. These
responses are caused primarily by the direct interaction of the two mixer input fundamental frequencies due to the nonlinear characteristics of mixers. It has been determined that harmonics of the local oscillator and of the signal which are generated in the mixer as the mixing takes place have little significance in the generation of spurious responses. Balanced mixers and others which utilize feedback may not exhibit these characteristics. The spurious frequencies may be readily identified in terms of the local oscillator, intermediate frequency and integer multipliers. The additional information required for a complete analysis of spurious response behavior is the spurious signal level, the limits of the integer multipliers $p$ and $q$ and the bandwidth of the responses. When this information is known, it is possible to optimize the intermediate frequency of a receiver for the best spurious response rejection and to determine which responses are significant. If these factors can be approximated from analyses of prior practical and theoretical considerations, it becomes possible to design RF and IF stages for improved spurious response rejection.

The problem in accomplishing this goal is discussed in the following sections.

2. Spurious Response Bandwidth

The bandwidth of an individual response is influenced primarily by the IF passband characteristics; but response symmetry is not expected because of other factors. Abrupt variations in RF selectivity, slight misalignment, and front-end overload may affect the bandwidth as well as the amplitude of the responses.
If an IF tuned circuit is considered which has a passband with a center frequency and incremental frequency, ΔIF, on each side, and the general spurious response equation \( F_{SR} = \frac{p\ell_{LO} \pm IF}{q} \) for a single conversion receiver is considered; it is possible to illustrate the manner in which the spurious response bandwidth (SRB) should vary.

\[
F_{SR} = \left| \frac{p\ell_{LO} \pm IF}{q} \right|
\]

\[
SRB \propto \left| \frac{p\ell_{LO} \pm (IF + \Delta IF)}{q} - \frac{p\ell_{LO} \pm (IF - \Delta IF)}{q} \right|
\]

\[
SRB \propto \left| \frac{\pm IF \pm \Delta IF \mp IF \pm \Delta IF}{q} \right|
\]

\[
SRB \propto \left| \frac{\pm 2\Delta IF}{q} \right|
\]

Therefore, the spurious response bandwidth is:

\[
SRB \propto \frac{BM(IF)}{q}
\]

This result illustrates that the spurious response bandwidth is inversely proportional to the integer multiplier q. Summarizing, an input frequency change will be multiplied by the integer q which causes the spurious response IF bandwidth to be effectively reduced by an amount that is dependent upon the integer q.

Experimental data reported earlier\(^{13}\) show that the bandwidth of a spurious response definitely does decrease as q values increase even for different response levels. Little difference in the characteristics is
noted for (+) and (-) mixes. Bandwidth change as a function of q is almost identical for different response levels and different mixes. These data substantiate the theoretical considerations.

Since it has been experimentally verified that the SRB is inversely proportional to q, it only remains to determine the constant of proportionality. The process by which this may be done is relatively simple. Assume that the bandwidth is:

$$\text{SRB} = \frac{K}{q}.$$ 

By fitting a smooth curve to experimental data points, the constant K can be determined. By conventional procedures,

$$K = \frac{\sum_{i=1}^{n} q_i B_{q_i}}{n}$$

where $q_i = q$ at which bandwidth data are available, $B_{q_i} = \text{bandwidth}$ associated with each $q_i$, and $n$ is the number of data points. To determine K consider the 3 db response bandwidths listed in Table II;

$$K = \frac{1(108) + 4(23) + 4(26) + 5(18) + - - -}{14}$$

$$K = \frac{1423}{14} \approx 102 \text{ kc}$$

and $\text{SRB}_{3} = \frac{102}{q} \text{ kc}.$

Similarly for the 10 and 20 db bandwidths

$$\text{SRB}_{10} = \frac{199}{q} \text{ kc}$$

and $\text{SRB}_{20} = \frac{325}{q} \text{ kc}.$
<table>
<thead>
<tr>
<th>P</th>
<th>Q</th>
<th>Sign</th>
<th>3 db Bandwidth (kc)</th>
<th>10 db Bandwidth (kc)</th>
<th>20 db Bandwidth (kc)</th>
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</thead>
<tbody>
<tr>
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<td>108</td>
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<td>284</td>
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<tr>
<td>1</td>
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<td>+</td>
<td>x</td>
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<td>26</td>
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<td>x</td>
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<td>-</td>
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<td>x</td>
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<td>-</td>
<td>x</td>
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<td>x</td>
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<td>18</td>
<td>-</td>
<td>2</td>
<td>x</td>
<td>x</td>
</tr>
</tbody>
</table>
A comparison of the calculated values of K with the measured spurious response bandwidths shows little difference. Similar results were obtained for another receiver. It can therefore be concluded that the spurious response bandwidth behaves in the manner predicted by the initial derived equation, \( SRB = \frac{BW(IF)}{q} \).

3. Spurious Response Rejection

The level of a spurious response is influenced by receiver characteristics such as sensitivity, selectivity, overload characteristics, and mixer parameters. To illustrate the difficulty in estimating spurious response rejection, each of these characteristics will be discussed in the following paragraphs.

3.1. Sensitivity

The sensitivity of a receiver while not necessarily affecting the spurious response rejection ratios does affect the absolute level of signals required to create a given response. Other factors being equal, a more sensitive receiver will generally be more susceptible to spurious responses.

3.2. Selectivity

The selectivity of a receiver can be closely approximated from theoretical considerations for those frequencies close to the tuned frequency. The selectivity at frequencies appreciably removed from the tuned frequency is influenced by spurious resonances, the type and number of tuned circuits, and leakage. Consideration of these factors in the prediction of the selectivity is difficult to accomplish. Stray capacitance,
inductive effects and shielding all affect the selectivity. Saturation effects can further cause appreciable variation from the theoretical.

Estimation of selectivity characteristics can result in errors in attenuation of several orders of magnitude at frequencies far removed from the tuned frequency. Measured selectivity curves for a receiver over a wide frequency range, while tedious to obtain, give a good valid knowledge suitable for use in spurious response analysis; however, the curves will be valid only for that particular receiver. Other receivers of the same type may exhibit different selectivity characteristics.

The overload characteristics of receivers are essentially the same as the single-signal characteristics but are different when two or more signals are present. With two signals present at the input, the effective overload characteristics are a function of both the level of the desired signal and the interfering signal. These factors make it difficult, if not impossible, to accurately estimate receiver selectivity from theoretical considerations alone.

3.3. Mixer Parameters

Mixer parameters have been discussed and analyzed by many investigators. Unfortunately, most of the analysis has been based on simple diode detectors and on fixed-bias fixed signal operation. As shown in Figure 18 of this report, the harmonic levels and spurious product levels are drastically influenced by bias and signal level. This effect has also been noted by Steiner\(^{14}\), who states, "The calculated values are accurate only if the total input power to the mixer is approximately 0 dBm. For greater values the mixer tends to saturate and for lesser values it apparently does not
behave in accordance with the mathematical power series model". For reasons yet to be determined, low input levels result in intermodulation product amplitudes which may exceed the calculated level by as much as 20 db. This effect can definitely result if the coefficients are assumed to remain constant for different signal levels.

It is also usually assumed that spurious mixes are directly related to the lowest generating order. The discussion and data in Section B of this report illustrate that this assumption may not be valid. To substantiate this assumption, it is usually argued that the higher order coefficients are negligible. There is evidence in the data of Section B that such is not always the case.

4. Conclusions

The foregoing discussion illustrates the difficulties encountered in the estimation of spurious response rejection. Determination of significant responses and p and q limits are equally difficult. However, for many purposes the order of magnitude estimates may be needed; therefore, analyses are being made to determine a method of estimating spurious response rejection and limits on the multiplier integers p and q.

D. Data Prepared or Published During the Contract Period

Tables III through V list the volumes of the Manuscript of Catalogue which were prepared and published during the contract period.
### TABLE III
RECEIVER TYPES TESTED AND MANUSCRIPT VOLUME NUMBERS

<table>
<thead>
<tr>
<th>Receiver Type</th>
<th>Type of Reception</th>
<th>Operating Frequencies</th>
<th>Manuscript Volume Number</th>
</tr>
</thead>
<tbody>
<tr>
<td>R-417/TRC-24 (A and D Bands)</td>
<td>FM</td>
<td>50 to 600 Mc</td>
<td>211</td>
</tr>
<tr>
<td>R-417/TRC-24 (A Band with filters)</td>
<td>FM</td>
<td>50 to 100 Mc</td>
<td>211 Supplement</td>
</tr>
<tr>
<td>RT-209/PRC-21</td>
<td>FM</td>
<td>152 to 174 Mc</td>
<td>212</td>
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### TABLE IV
LIST OF TRANSMITTERS EVALUATED

<table>
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<tr>
<th>Transmitter Type</th>
<th>Power Output</th>
<th>Frequency Range</th>
<th>Type of Emission</th>
<th>Manuscript Volume Number</th>
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<td>152 to 174 Mc</td>
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<td>T-302/TRC-24 (A Band with filters)</td>
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<td>326 Supplement</td>
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### TABLE V
GENERAL MANUSCRIPTS

<table>
<thead>
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<th>Title</th>
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<tbody>
<tr>
<td>Mutual Interference Chart Constructor Program</td>
<td>6</td>
</tr>
<tr>
<td>Computer Reduction of Laboratory Data for Input to MIC Construction and Frequency Selection Programs</td>
<td>7</td>
</tr>
<tr>
<td>A Computer Technique for the Selection of Compatible Frequencies for Communications Equipment Lash-Ups</td>
<td>8</td>
</tr>
</tbody>
</table>
E. Automatic Data Reduction

Two data reduction techniques were developed and checked during this project interim: "A Computer Reduction of Laboratory Data for Input to MIC Construction and Frequency Selection Programs," described in Volume 7 of the Manuscript of Catalogue, and "A Computer Technique for the Selection of Compatible Frequencies for Communications Equipments Lash-Ups," described in Volume 8 of the Manuscript of Catalogue.

The former program was developed to analyze data as it comes from the laboratory and to arrange the data in a proper format for use in MIC Construction and Frequency Selection Programs. Operating requirements were simplified to permit large usage and the program was written so that the analyses would be consistent. The output may be obtained in the form of punched cards, printout, or magnetic tape.

The technique for selection of frequencies for lash-ups was based on an analysis of the lash-up problem as it is currently defined. It is assumed that a list of frequencies which are available for use will be known. The resulting program uses these known frequencies and the interference characteristics of the equipment to be used in the specific lash-up to assign compatible frequencies. The technique does not require the construction of MIC's but uses the actual interference lines. The program also specifies the antenna coupling required for specific frequencies to be usable. The output from this program may be in the form of punched cards or printout.

Volume 6 of the Manuscript of Catalogue was also completed and published during the contract period. The volume contains a computer program which provides a flexible method of constructing and printing a Mutual Interference
Chart of any size from a list of interference lines belonging to any transmitter-receiver pair.
V. CONCLUSIONS

Tests and test procedures were developed and verified using FM microwave transmitters and receivers and the multiplex set AN/TCC-13 which employs pulse position modulated-time division multiplexing. No pulsed output transmitters were available for evaluation purposes, but the tests and test procedures are adaptable for use in evaluation of pulsed-type equipments. Most of the tests and test procedures for transmitters and receivers are similar to the ones described in existing spectrum signature collection plans. Additional techniques were developed where needed for testing in the microwave region.

The system error rate test provides a relative measure of the effects of various types of interfering signals on a particular system. As expected with an FM system the output error rate is zero until the interfering RF cochannel signal amplitude approaches that of the system signal. The error rate then rapidly changes from zero to 100 percent as the signal level is increased a few decibels. As the frequency of the interfering signal is changed in the receiver passband, the amount of interfering signal level required to produce a given error rate is changed.

The terminal equipment crosstalk test provides a method for determining crosstalk information in the terminal set. No crosstalk interference was measurable in the multiplex set AN/TCC-13.

The sensitivity of NFIM's which operate at frequencies above 1 kmc is approximately -80 dbm while the sensitivity of NFIM's below 1 kmc is usually better than -100 dbm. Therefore, fewer transmitter spurious emission and intermodulation data points are measurable at frequencies
above 1 kmc. The limited dynamic range of the microwave spectrum analyzers prevents the observance of transmitter output spectrum components more than 40 db below the carrier level. The development of some of the tests and test procedures was impaired by the unavailability of test equipments such as wideband FM signal generators and FM deviation detectors which operate at microwave frequencies.

The mixer studies have involved the taking of data for a limited number of mixers. The effort expended so far has resulted in the development of a mixer which has excellent spurious response rejection possibilities. The gathered data have led to the verification of existing mixer theories and have indicated a theory involving the effect of conduction angle. An attempt to establish relationships between the various p and q combinations has resulted in a possible prediction method for certain type mixers.

The bandwidth of an individual spurious response is determined primarily by the IF passband characteristics. The level of a spurious response is influenced by receiver characteristics such as RF selectivity and mixer parameters.
VI. OVERALL CONCLUSIONS

Two computer programs were written in order to simplify data processing and manuscript preparation for the receiver spurious response and transmitter spurious emission data. However, accurately measured spectrum signature data are essential. Previous methods of identifying the frequencies according to the applicable mathematical equation and listing them involve excessive amounts of engineering and data recording time. A typical instance required more than 8 hours of an engineer's time and several days of typist's time. Through the use of a Burroughs 220 computer, this time was reduced to approximately one engineering man-hour and one typist man-hour.

The cumulated transmitter and receiver spectrum signature data were analyzed to determine trends in the data. It was found that many transmitter and receiver characteristics did not change significantly with tuned frequency over any single band. In fact, the change in most cases was considerably less than the set-to-set variations.

A manual method for analyzing intermodulation interference in a lash-up was developed. The method is intended for use in those cases where the number of intermodulation products is so large that an insufficient number of intermodulation interference-free channels is obtainable, thus requiring a certain separation of two cosite frequencies to eliminate a product frequency. The method can also be used to point out specific areas where equipment improvement is desirable.

Because of the importance of transmitter emission bandwidth and side-band splatter, the harmonic spectrums of FM transmitters were investigated.
The spectrum symmetry and levels of each modulated harmonic signal reflect the manner in which the transmitter is tuned. It was found that the spectrum occupancy increases as the harmonic number of the FM signal increases, but not as an integral multiple of the fundamental spectrum width.

The investigation of carrier cancellation for extending the dynamic range of spectrum analyzers resulted in the development of a device which successfully reduces the carrier amplitude 40 decibels. The device provides sufficient selectivity to allow measurement of transmitter radiation occurring 20 kc from the carrier with very little error. Several fields of study exist in which the cancellation device should prove useful. An important and relatively untouched field is that of studying the frequency characteristics of carrier noise.

The spectral behavior of communications receiver spurious responses is found to be highly predictable. The bandwidth of each spurious response can be calculated by dividing the IF bandwidth by the 'q' used in the response equation to identify the response.

The experimental work on the spurious response bandwidth and rejection ratios has been beneficial to the automatic data reduction programs by providing refinements in regard to the expected guardband widths and level requirements of interference lines. This study, combined with the mixer studies, has resulted in a better understanding of the methods of spurious response generation, prediction, and suppression.

The computer programs for automatic data reduction show the feasibility of solving many co-sited interference problems by utilizing transmitter and receiver spectrum signature data. Methods are presented
for deriving interference lines used in lash-up frequency selection schemes. The procedures developed in the lash-up program contain the computing techniques for solving the majority of lash-up interference cases. Two practical examples were also presented. The computer presently used has insufficient storage capacity to handle large scale lash-up problems using the present "universal" library procedures. Larger lash-up problems may be written for the available computer by departing from the universal procedures and writing each program individually. Large scale lash-up programs may be written using the library procedures outlined in Volume 8, Manuscript of Catalogue, for computers with greater core-storage capacity (for example, Burroughs 220 with 10,000 word core-storage capacity).
VII. RECOMMENDATIONS

It is recommended that additional microwave equipment be tested for further verification of the tests and test procedures outlined in this report. Terminal equipments employing various multiplexing schemes should be investigated since only the multiplex set AN/TCC-13 was available for testing.

System tests involving various types of cochannel and close-channel interference need to be developed. Preliminary tests which were made on the AN/TRC-29 and AN/TCC-13 sets indicate that the General Electronics Laboratories Speech Systems Test Set may be of value in further evaluation of system cochannel interference.

It is recommended that a study be made for the development and improvement of microwave test equipment including wideband FM signal generators and FM deviation detectors.

It is further recommended that the mixer studies be continued. The investigations have yielded valuable data regarding the primary causes of spurious responses. However, there is still no established theory relating the generation of spurious responses to the conduction angle. The spurious response prediction method resulting from these studies has not been tested in a practical situation. Further study is required if the effort expended thus far is to culminate in some usable form for the reduction or elimination of spurious responses.

The approach described in Volume 8 of the Manuscript of Catalogue toward the solution of cosite interference problems for lash-ups is both
practical and feasible. However, the computer program for a specific lash-up will differ from the program for any other specific lash-up primarily because of the difference in the number of transmitter and/or receivers involved, and the difference in the number of assignments of simplex and duplex pairs of equipments. Since only two examples were constructed and presented, it is recommended that additional work be done in solving more complex lash-up interference problems.
VIII. IDENTIFICATION OF KEY TECHNICAL PERSONNEL

<table>
<thead>
<tr>
<th>Name</th>
<th>Title</th>
<th>Approximate Hours</th>
</tr>
</thead>
<tbody>
<tr>
<td>Robert N. Bailey</td>
<td>Project Director (Through December 1962)</td>
<td>1482</td>
</tr>
<tr>
<td>John L. Cundiff</td>
<td>Research Assistant</td>
<td>680</td>
</tr>
<tr>
<td>Hugh W. Denny</td>
<td>Research Assistant</td>
<td>528</td>
</tr>
<tr>
<td>John G. Holey</td>
<td>Research Assistant</td>
<td>681</td>
</tr>
<tr>
<td>Alton P. Jensen</td>
<td>Asst. Research Engineer</td>
<td>138</td>
</tr>
<tr>
<td>Walter C. Knapp</td>
<td>Research Assistant</td>
<td>336</td>
</tr>
<tr>
<td>Sterlin P. Lenoir, Jr.</td>
<td>Special Research Engineer</td>
<td>92</td>
</tr>
<tr>
<td>I. E. Perlin</td>
<td>Research Professor</td>
<td>673</td>
</tr>
<tr>
<td>D. W. Robertson</td>
<td>Head, Communications Branch (1 Oct. 1962 to Present)</td>
<td>185</td>
</tr>
<tr>
<td>Charles W. Stuckey</td>
<td>Asst. Research Engineer</td>
<td>643</td>
</tr>
<tr>
<td>Robert Techo</td>
<td>Research Engineer</td>
<td>289</td>
</tr>
<tr>
<td>Robert D. Trammell, Jr.</td>
<td>Asst. Research Engineer</td>
<td>1351</td>
</tr>
<tr>
<td>Joseph R. Walsh, Jr.</td>
<td>Research Engineer</td>
<td>1296</td>
</tr>
<tr>
<td>E. Wendell Wood</td>
<td>Project Director (1 Jan. 1963 to Present)</td>
<td>1656</td>
</tr>
<tr>
<td>William B. Wrigley</td>
<td>Head, Communications Branch (November 1959 through September 1962)</td>
<td>179</td>
</tr>
</tbody>
</table>

The background and qualifications of these men are presented in the following paragraphs.

Mr. Bailey served as Project Director through December 1962. He has been associated with the project since 5 January 1959. He holds the degree
of M.S. in E.E. from the Georgia Institute of Technology. His previous experience includes four years as radio technician in the USAF; two years as Instrument Engineer for the E. I. duPont de Nemours & Co., Savannah River Plant. He is a registered Professional Engineer in the State of Georgia.

Mr. Cundiff received a B.S. of E.P. degree from Auburn University in 1959. His previous experience includes RCA programmed missile data reduction programs for the FLAC and IBM 704, 709, 7090, 1401 computers. These programs reduced data from missile tracking systems such as PPS-16 radars, MARK 2 azusa, GE Guidance system, Ballistic Cameras, fixed cameras, and tracking cameras. Other duties at RCA included the setting up and programming of the Automatic Operating System for the 7090 presently used at the Atlantic Missile Range. The work done on the 1401 consisted of implementing the 1401 as the peripheral equipment to support the 7090 data reduction system. He has also taught courses as Brevard Engineering College.

Mr. Denny joined the project in October 1962. He received a B.E.E. degree from Tennessee Polytechnic Institute in 1960 and is presently working toward an M.S. degree in the same field from the Georgia Institute of Technology. His previous experience includes work in design, construction, and testing of transistorized, crystal-controlled, VHF oscillators. As an Army Officer, he served with a combat-area Signal Batallion in Texas and Germany as a Wire Communications Officer. While at Florida Power and Light Company, he worked in the Systems Control Division in the planning of load control systems and in the Relay Division.
Mr. Holey received the B.E.E. degree in June 1956, and a M.S. degree in the same field from the Georgia Institute of Technology. His previous experience includes three years in the USAF as a Communications Officer, with approximately one year in single-sideband receiver and transmitter testing and evaluation. Prior to serving in the Air Force, he was employed by the Chrysler Corp. Missile Operation as an electrical engineer. Mr. Holey was a member of the project from August 1960 until June 1962.

Mr. Jensen received an A.S. degree in Industrial Technology and also an A.S. degree in Mechanical Technology from Southern Technical Institute in 1952. He received his B.S. degree in Mechanical Engineering from the Georgia Institute of Technology in 1956 and is presently working toward a M.S. degree in the same field. His previous experience includes two years as a Junior Engineer with the Rohm and Haas Research Division in Huntsville, Alabama, where his primary responsibility was in field testing of small free-flight, rocket-type, weapon systems. He is currently director of a project to determine the component information requirements of design engineers and to establish a system for classifying, storing, and retrieving engineering data using existing equipments.

Mr. Knapp joined the project on July 1, 1961, and was employed on a full time basis until April of 1962. He received a B.S. degree in Physics from Georgia State College in June 1961. Mr. Knapp has worked on various research projects as a Research Technician since January 1955. He has also been engaged in construction and operation of various radar systems and communication equipments.
Mr. Lenoir received an M.S. in E.E. degree from the Georgia Institute of Technology in 1949. His previous experience includes four years with the U.S. Navy Reserve; one year as Instructor in Physics, Mississippi State College; one year as Graduate Assistant in E.E., Georgia Institute of Technology; and two years as electronic scientist, Armed Forces Security Agency. Mr. Lenoir has been associated with various projects at Georgia Tech since 1951.

Dr. Perlin received his Ph.D. in Mathematics from the University of Chicago in 1935. His prior experience includes five years as Professor of Mathematics at the Illinois Institute of Technology; 17 years in the U. S. Navy Reserve, Gunnery. He has been working on a full-time basis with Georgia Tech since 1946. He is presently a Professor of Mathematics and also a Research Professor at the Engineering Experiment Station.

Mr. Robertson assumed duties as Head of the Communications Branch on October 1, 1962. He has been associated with the Georgia Institute of Technology since 1947. He received a B.S. degree in E.E. in 1950 and a M.S. degree in the same field in 1957. From 1941 to 1947, he was employed by Civil Service where his work included maintenance and installation of airborne radio and radar equipments. At Georgia Tech, he has been a staff member and director of various projects.

Mr. Stuckey received a B.S. in I.E. in 1957 and a M.S. in I.E. in 1961 at Georgia Institute of Technology. Since 1957 he has been working on various projects at the Engineering Experiment Station and teaching in the Mathematics Department. He has been associated with a project with Rome Air Development Center in the field of articulation and intelligibility measurement. In particular, he has developed a statistical method for the precise comparison and evaluation of articulation team performance.
Dr. Techo received a M.S. in Ch.E. in 1958 and a Ph.D. degree in the same field from the Georgia Institute of Technology in 1961. His previous experience includes three years as Engineering Officer in the U. S. Navy; and he has been associated with Georgia Tech since 1956.

Mr. Trammell joined the project staff in May 1962. He received a B.S. in E.E. from the Georgia Institute of Technology in 1956, and an M.S. in the same field in 1959. His previous experience includes work in Analog Computation Programming, and design and construction of special purpose medical data reduction computers for Emory University. As an Army Officer, he served at Fort Knox, Kentucky, as Company Executive Officer and as Regimental Motor Officer in a basic training regiment in USATC, Armor.

Mr. Walsh also joined the project staff in May 1962. He received a B.S. in E.E., from Georgia Institute of Technology in 1949, and an M.S. in the same field in 1961. His previous experience includes work with the Civil Aeronautics Administration where he installed and tested VOR and ILS navigation systems. At Georgia Tech, he was an assistant director of a project to design and construct a radar system for studying characteristics of ground clutter and target return; and was concerned with applications of these results in an experimental radar system. He has been associated with the design and development of electronic circuitry of a radar system for study of the polarization and statistical properties of sea return, and with field operation of this equipment.

Mr. Wood assumed duties as Project Director on 1 January 1963. He holds a B.E.E. degree from the Georgia Institute of Technology and has completed requirements for an M.S. degree in the same field. His previous
experience includes three years with the Federal Communications Commission at the Atlanta Field Office where he worked with radio interference cases; fifteen months as a Test Assistant for the Georgia Power Company; two years as a Radio Engineer with an Atlanta broadcasting company; and two years as a radio and TV technician in Thomas County, Georgia.

Mr. Wrigley was Head of the Communications Branch from November 1959 until September 1962. He has been associated with the U.S. Army Reserve and Georgia Air National Guard for 21 years and presently holds the rank of Colonel. His previous experience includes two years as Development Engineer with RCA; four years as Ordnance and Anti-Aircraft Officer, U. S. Army; one year as Plant Superintendent and Chief Engineer, U. S. Homes, Inc.; one year as Development Engineer, E. I. Sarbacher & Associates; one year as Head of Electromagnetic Propagation Group, A. F. Cambridge Research Center; and has been associated with the Engineering Experiment Station, Georgia Institute of Technology since 1948. Mr. Wrigley is presently associated with a commercial firm in Atlanta.

Respectfully submitted:

[Signature]
E. W. Wood
Project Director

Approved:

[Signature]
D. W. Robertson, Head
Communications Branch

[Signature]
M. W. Long, Chief
Electronics Division
IX. REFERENCES


3. PRD Reports, Volume 5, No. 3., July 1957.


13. Ibid.

APPENDIX

Determination of Coefficients for Frequencies Generated by a Mixer with Two Input Signals

1. Definitions and Assumptions

Assume that the mixer characteristic may be represented by a polynomial of degree n:

\[ i_{mo} = K_0 e_{i1} \cos \theta + K_1 e_{i1} \cos \phi + K_2 e_{i1}^2 + K_3 e_{i1}^3 + \cdots + K_n e_{i1}^n \]

where:

- \( i_{mo} \) = mixer output current,
- \( e_{i1} \) = mixer input voltage, and
- \( K_n \) = constant coefficients for the given mixer parameters.

The input signal to a mixer may be assumed to be the sum of two cosine waves (that of the local oscillator and that of the carrier signal) represented by:

\[ e_i = e_a \cos a + e_b \cos b, \]

where:

- \( a = 2\pi f_{LO} t \),
- \( b = 2\pi f_{CS} t \),
- \( f_{LO} \) = Local oscillator frequency,
- \( f_{CS} \) = Carrier signal frequency,
- \( e_a \) = Local oscillator signal voltage,
- \( e_b \) = Carrier signal voltage, and
- \( t = \) Time in seconds.
2. Procedures

Substituting the quantity $e_a \cos a + e_b \cos b$ for $e_i$ in the polynomial will yield a mixer output current consisting of a sum of various frequency terms in accordance with the binomial expansion of $\left( e_a \cos a + e_b \cos b \right)^n$ for each term in the polynomial.

In order to facilitate the expansion, one may first raise $\cos \theta$ to the highest power needed to reach the degree of the polynomial desired. The terms $a$ or $b$ may then be substituted for $\theta$ in forming the necessary products. In this example, the degree is carried to ten; thus $\cos \theta$ is raised to the tenth power and all intermediate steps are recorded and shown in Table I-A.

---

**TABLE I-A**

**COSINE $\theta$ RAISED TO THE TENTH POWER**

$\cos \theta = 1 \cos \theta$

$\cos^2 \theta = \frac{1}{2} + \frac{1}{2} \cos 2\theta$

$\cos^3 \theta = \frac{3}{4} \cos \theta + \frac{1}{4} \cos 3\theta$

$\cos^4 \theta = \frac{3}{8} + \frac{1}{2} \cos 2\theta + \frac{1}{8} \cos 4\theta$

$\cos^5 \theta = \frac{5}{8} \cos \theta + \frac{5}{16} \cos 3\theta + \frac{1}{16} \cos 5\theta$

$\cos^6 \theta = \frac{5}{16} + \frac{15}{32} \cos 2\theta + \frac{3}{16} \cos 4\theta + \frac{1}{32} \cos 6\theta$

(Continued)
TABLE I-A (Continued)

COSINE θ RAISED TO THE TENTH POWER

\[
\begin{align*}
\cos^7 \theta &= \frac{35}{64} \cos \theta + \frac{21}{64} \cos 3\theta + \frac{7}{64} \cos 5\theta + \frac{1}{64} \cos 7\theta \\
\cos^8 \theta &= \frac{35}{128} + \frac{7}{16} \cos 2\theta + \frac{7}{32} \cos 4\theta + \frac{1}{16} \cos 6\theta + \frac{1}{128} \cos 8\theta \\
\cos^9 \theta &= \frac{63}{128} \cos \theta + \frac{21}{64} \cos 3\theta + \frac{9}{64} \cos 5\theta + \frac{9}{256} \cos 7\theta + \frac{1}{256} \cos 9\theta \\
\cos^{10} \theta &= \frac{63}{256} + \frac{105}{256} \cos 2\theta + \frac{15}{64} \cos 4\theta + \frac{45}{512} \cos 6\theta + \frac{5}{256} \cos 8\theta + \frac{1}{512} \cos 10\theta
\end{align*}
\]

When the various product combinations of the above are formed, the terms belonging to the binomial expansion of \((\cos a + \cos b)^n\) for all of the terms of the polynomial are obtained as shown in Table II-A. It is not necessary to perform both sides of any binomial expansion since the coefficients for \(\cos a \cos^2 b\) may be obtained from the expansion of \(\cos^2 a \cos b\), etc. In Table II-A, the binomial coefficient, polynomial coefficient, and the product of the input amplitudes raised to the appropriate powers are parenthesized and shown below the term with which they are associated in the binomial expansion.
### TABLE II-A

**VARIOUS COSINE PRODUCT COMBINATIONS**

<table>
<thead>
<tr>
<th>Expression</th>
<th>Expanded Form</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\cos a \cos b = \frac{1}{2} \cos(a + b) + \frac{1}{2} \cos(a - b)$</td>
<td>(2e $a \cdot b$)</td>
</tr>
<tr>
<td>$\cos^2 a \cos b = \frac{1}{2} \cos b + \frac{1}{4} \cos(2a + b) + \frac{1}{4} \cos(2a - b)$</td>
<td>(3e $a \cdot b$)</td>
</tr>
<tr>
<td>$\cos^2 a \cos^2 b = \frac{1}{4} \cos(2a + 2b) + \frac{1}{4} \cos(2a - 2b)$</td>
<td>(6e $a \cdot b$)</td>
</tr>
<tr>
<td>$\cos^3 a \cos b = \frac{3}{8} \cos(a + b) + \frac{3}{8} \cos(a - b) + \frac{1}{8} \cos(3a + b) + \frac{1}{8} \cos(3a - b)$</td>
<td>(4e $a \cdot b$)</td>
</tr>
<tr>
<td>$\cos^3 a \cos^2 b = \frac{3}{16} \cos a + \frac{1}{8} \cos 3a + \frac{3}{16} \cos(a + 2b) + \frac{3}{16} \cos(a - 2b)$</td>
<td>(10e $a \cdot b$)</td>
</tr>
<tr>
<td>$\cos^3 a \cos^3 b = \frac{9}{32} \cos(a + b) + \frac{9}{32} \cos(a - b) + \frac{3}{32} \cos(a + 3b) + \frac{3}{32} \cos(a - 3b)$</td>
<td>(20e $a \cdot b$)</td>
</tr>
</tbody>
</table>

-120-
TABLE II-A (Continued)

\[
\cos^4 a \cos b = \frac{3}{8} \cos b + \frac{1}{4} \cos(2a + b) + \frac{1}{4} \cos(2a - b) + \frac{1}{16} \cos(4a + b)
\]

\[
(5e^{4a e b}_{K_5}) + \frac{1}{16} \cos(4a - b)
\]

\[
\cos^4 a \cos^2 b = \frac{3}{16} + \frac{1}{4} \cos 2a + \frac{1}{8} \cos 4a + \frac{1}{8} \cos(2a + 2b) + \frac{1}{8} \cos(2a - 2b)
\]

\[
(15e^{4a e b}_{K_6}) + \frac{1}{32} \cos(4a + 2b) + \frac{1}{32} \cos(4a - 2b) + \frac{3}{16} \cos 2b
\]

\[
\cos^4 a \cos^3 b = \frac{9}{32} \cos b + \frac{3}{32} \cos 3b + \frac{3}{16} \cos(2a + b) + \frac{3}{16} \cos(2a - b)
\]

\[
(35e^{4a e b}_{K_7}) + \frac{1}{16} \cos(2a + 3b) + \frac{1}{16} \cos(2a - 3b) + \frac{3}{64} \cos(4a + b)
\]

\[
+ \frac{3}{64} \cos(4a - b) + \frac{1}{64} \cos(4a + 3b) + \frac{1}{64} \cos(4a - 3b)
\]

\[
\cos^4 a \cos^4 b = \frac{9}{64} + \frac{3}{16} \cos 2a + \frac{3}{64} \cos 4a + \frac{3}{16} \cos 2b + \frac{3}{64} \cos 4b
\]

\[
(70e^{4a e b}_{K_8}) + \frac{1}{8} \cos(2a + 2b) + \frac{1}{8} \cos(2a - 2b) + \frac{1}{32} \cos(4a + 2b)
\]

\[
+ \frac{1}{32} \cos(4a - 2b) + \frac{1}{32} \cos(2a + 4b) + \frac{1}{32} \cos(2a - 4b)
\]

\[
+ \frac{1}{128} \cos(4a + 4b) + \frac{1}{128} \cos(4a - 4b)
\]

\[
\cos^5 a \cos b = \frac{5}{16} \cos(a + b) + \frac{5}{16} \cos(a - b) + \frac{5}{32} \cos(3a + b)
\]

\[
(6e^{5a e b}_{K_6}) + \frac{5}{32} \cos(3a - b) + \frac{1}{32} \cos(5a + b) + \frac{1}{32} \cos(5a - b)
\]
\[
\cos^5 a \cos^2 b = \frac{5}{16} \cos a + \frac{5}{32} \cos 3a + \frac{1}{32} \cos 5a + \frac{5}{32} \cos(a + 2b)
\]
\[
\cos^5 a \cos 3b = \frac{15}{64} \cos(a + b) + \frac{15}{64} \cos(a - b) + \frac{15}{128} \cos(3a + b)
\]
\[
\cos^5 a \cos 4b = \frac{15}{64} \cos a + \frac{15}{128} \cos 3a + \frac{3}{128} \cos 5a + \frac{5}{32} \cos(a + 2b)
\]

TABLE II-A (Continued)

\[
\cos^5 a \cos 2b = \cos^5 a \cos 2b + \cos^5 a \cos 2b + \cos^5 a \cos 2b
\]
\[
\cos^5 a \cos 3b = \cos^5 a \cos 3b + \cos^5 a \cos 3b + \cos^5 a \cos 3b
\]
\[
\cos^5 a \cos 4b = \cos^5 a \cos 4b + \cos^5 a \cos 4b + \cos^5 a \cos 4b
\]
TABLE II-A (Continued)

\[
\cos^5 a \cos^5 b = \frac{25}{128} \cos(a + b) + \frac{25}{128} \cos(a - b) + \frac{25}{256} \cos(a + 3b) \\
(252a e \_b 5K_{10}) \\
+ \frac{25}{256} \cos(a - 3b) + \frac{5}{256} \cos(a + 5b) + \frac{5}{256} \cos(a - 5b) \\
+ \frac{25}{256} \cos(3a + b) + \frac{25}{256} \cos(3a - b) + \frac{25}{512} \cos(3a + 3b) \\
+ \frac{25}{512} \cos(3a - 3b) + \frac{5}{512} \cos(3a + 5b) + \frac{5}{512} \cos(3a - 5b) \\
+ \frac{5}{256} \cos(5a + b) + \frac{5}{256} \cos(5a - b) + \frac{5}{512} \cos(5a + 3b) \\
+ \frac{5}{512} \cos(5a - 3b) + \frac{1}{512} \cos(5a + 5b) + \frac{1}{512} \cos(5a - 5b) \\
\]

\[
\cos^6 a \cos b = \frac{5}{16} \cos b + \frac{15}{64} \cos(2a + b) + \frac{15}{64} \cos(2a - b) + \frac{3}{32} \cos(4a + b) \\
(76a e \_b K_{11}) \\
+ \frac{3}{32} \cos(4a - b) + \frac{1}{64} \cos(6a + b) + \frac{1}{64} \cos(6a - b) \\
\]

\[
\cos^6 a \cos^2 b = \frac{5}{32} + \frac{15}{64} \cos 2a + \frac{3}{32} \cos 4a + \frac{1}{64} \cos 6a + \frac{5}{32} \cos 2b \\
(286a e \_b 2K_{8}) \\
+ \frac{15}{128} \cos(2a + 2b) + \frac{15}{128} \cos(2a - 2b) + \frac{3}{64} \cos(4a + 2b) \\
+ \frac{3}{64} \cos(4a - 2b) + \frac{1}{128} \cos(6a + 2b) + \frac{1}{128} \cos(6a - 2b) \\
\]

\[
\cos^6 a \cos^3 b = \frac{15}{64} \cos b + \frac{45}{256} \cos(2a + b) + \frac{45}{256} \cos(2a - b) + \frac{9}{128} \cos(4a + b) \\
(846a e \_b 3K_{9}) \\
+ \frac{9}{128} \cos(4a - b) + \frac{3}{256} \cos(6a + b) + \frac{3}{256} \cos(6a - b) \\
+ \frac{5}{64} \cos 3b + \frac{15}{256} \cos(2a + 3b) + \frac{15}{256} \cos(2a - 3b) \\
\]
TABLE II-A (Continued)

\[
\begin{align*}
+ \frac{3}{128} \cos(4a + 3b) &+ \frac{3}{128} \cos(4a - 3b) + \frac{1}{256} \cos(6a + 3b) \\
+ \frac{1}{256} \cos(6a - 3b) &
\end{align*}
\]

\[
\begin{align*}
\cos^6 a \cos^4 b &= \frac{15}{128} + \frac{45}{256} \cos 2a + \frac{9}{128} \cos 4a + \frac{3}{256} \cos 6a + \frac{5}{32} \cos 2b \\
(210)\quad e^6 \cdot e^4 K^{10} &+ \frac{15}{128} \cos(2a + 2b) + \frac{15}{128} \cos(2a - 2b) + \frac{3}{64} \cos(4a + 2b) \\
&+ \frac{3}{64} \cos(4a - 2b) + \frac{1}{128} \cos(6a + 2b) + \frac{1}{128} \cos(6a - 2b) \\
&+ \frac{5}{128} \cos 4b + \frac{15}{512} \cos(2a + 4b) + \frac{15}{512} \cos(2a - 4b) \\
&+ \frac{3}{256} \cos(4a + 4b) + \frac{3}{256} \cos(4a - 4b) + \frac{1}{512} \cos(6a + 4b) \\
&+ \frac{1}{512} \cos(6a - 4b)
\end{align*}
\]

\[
\begin{align*}
\cos^7 a \cos b &= \frac{35}{128} \cos(a + b) + \frac{35}{128} \cos(a - b) + \frac{21}{128} \cos(3a + b) \\
(8)\quad e^7 \cdot e K^8 &+ \frac{21}{128} \cos(3a - b) + \frac{7}{128} \cos(5a + b) + \frac{7}{128} \cos(5a - b) \\
&+ \frac{1}{128} \cos(7a + b) + \frac{1}{128} \cos(7a - b)
\end{align*}
\]

\[
\begin{align*}
\cos^7 a \cos^2 b &= \frac{35}{128} \cos a + \frac{21}{128} \cos 3a + \frac{7}{128} \cos 5a + \frac{1}{128} \cos 7a \\
(36)\quad e^7 \cdot e^2 K^9 &+ \frac{35}{256} \cos(a + 2b) + \frac{35}{256} \cos(a - 2b) + \frac{21}{256} \cos(3a + 2b) \\
&+ \frac{21}{256} \cos(3a - 2b) + \frac{7}{256} \cos(5a + 2b) + \frac{7}{256} \cos(5a - 2b) \\
&+ \frac{1}{256} \cos(7a + 2b) + \frac{1}{256} \cos(7a - 2b)
\end{align*}
\]

-124-
\begin{align*}
\cos^7 a \cos^3 b &= \frac{105}{512} \cos(a + b) + \frac{105}{512} \cos(a - b) + \frac{63}{512} \cos(3a + b) \\
&\quad + \frac{63}{512} \cos(3a - b) + \frac{21}{512} \cos(5a + b) + \frac{21}{512} \cos(5a - b) \\
&\quad + \frac{3}{512} \cos(7a + b) + \frac{3}{512} \cos(7a - b) + \frac{35}{512} \cos(a + 3b) \\
&\quad + \frac{35}{512} \cos(a - 3b) + \frac{21}{512} \cos(3a + 3b) + \frac{21}{512} \cos(3a - 3b) \\
&\quad + \frac{7}{512} \cos(5a + 3b) + \frac{7}{512} \cos(5a - 3b) + \frac{1}{512} \cos(7a + 3b) \\
&\quad + \frac{1}{512} \cos(7a - 3b) \\
\cos^8 a \cos b &= \frac{35}{128} \cos b + \frac{7}{32} \cos(2a + b) + \frac{7}{32} \cos(2a - b) + \frac{7}{64} \cos(4a + b) \\
&\quad + \frac{7}{64} \cos(4a - b) + \frac{1}{32} \cos(6a + b) + \frac{1}{32} \cos(6a - b) \\
&\quad + \frac{1}{256} \cos(8a + b) + \frac{1}{256} \cos(8a - b) \\
\cos^8 a \cos^2 b &= \frac{35}{256} + \frac{7}{32} \cos 2a + \frac{7}{64} \cos 4a + \frac{1}{32} \cos 6a + \frac{1}{256} \cos 8a \\
&\quad + \frac{35}{256} \cos 2b + \frac{7}{64} \cos(2a + 2b) + \frac{7}{64} \cos(2a - 2b) \\
&\quad + \frac{7}{128} \cos(4a + 2b) + \frac{7}{128} \cos(4a - 2b) + \frac{1}{64} \cos(6a + 2b) \\
&\quad + \frac{1}{64} \cos(6a - 2b) + \frac{1}{512} \cos(8a + 2b) + \frac{1}{512} \cos(8a - 2b)
\end{align*}
TABLE II-A (Continued)

\[
\cos^9 a \cos b = \frac{63}{256} \cos(a + b) + \frac{63}{256} \cos(a - b) + \frac{21}{128} \cos(3a + b) \\
(10e_a^9 e_b^{K_{10}}) + \frac{21}{128} \cos(3a - b) + \frac{9}{128} \cos(5a + b) + \frac{9}{128} \cos(5a - b) \\
+ \frac{9}{512} \cos(7a + b) + \frac{9}{512} \cos(7a - b) + \frac{1}{512} \cos(9a + b) \\
+ \frac{1}{512} \cos(9a - b)
\]

3. Formation of Frequency Term Coefficients

When the terms of like frequency are combined into a sum of terms containing coefficients of like ordered K's, the frequencies and associated coefficients shown in Table III-A are obtained. These frequencies are those generated by the action of a nonlinearity, described by a polynomial of the tenth degree, on two sinusoidal signals added immediately prior to the nonlinear device. For ease of comparison, the harmonic terms shown in Table IV-A are those for the harmonics generated with only one input signal. Those shown in Table V-A are additional terms that are formed when the second input signal is introduced. The frequencies shown in Table III-A are roughly half of the frequencies formed by the mixer with both input signals present. The selection shown is sufficient to determine all of the frequencies by substituting a for b, etc.
TABLE III-A

FREQUENCIES GENERATED BY TERMS UP TO n = 10 IN MIXER POLYNOMIAL

| cos a | \[ e_{a_{1}} K_{1} + K_{3} \left( \frac{3}{4} e_{a}^{3} + \frac{3}{2} e_{a} e_{b}^{2} \right) + K_{5} \left( \frac{5}{8} e_{a}^{5} + \frac{15}{4} e_{a}^{3} e_{b}^{2} + \frac{15}{8} e_{a} e_{b}^{4} \right) + K_{7} \left( \frac{35}{64} e_{a}^{7} + \frac{315}{32} e_{a}^{4} e_{b}^{3} + \frac{105}{16} e_{a}^{2} e_{b}^{5} \right) + K_{9} \left( \frac{315}{128} e_{a}^{6} + \frac{945}{32} e_{a}^{4} e_{b}^{2} + \frac{315}{16} e_{a}^{2} e_{b}^{4} \right) + \frac{315}{32} e_{a}^{2} e_{b} + \frac{315}{256} e_{a} e_{b}^{8} + \frac{7}{8} e_{a} e_{b}^{6} + \frac{315}{64} e_{a}^{2} e_{b} \right) + \ldots + \] |
| cos 2a | \[ \frac{1}{2} e_{a}^{2} K_{2} + K_{4} \left( \frac{1}{2} e_{a}^{4} + \frac{3}{2} e_{a} e_{b}^{2} \right) + K_{6} \left( \frac{15}{32} e_{a}^{6} + \frac{15}{4} e_{a}^{4} e_{b} + \frac{105}{16} e_{a}^{2} e_{b}^{4} + \frac{105}{8} e_{a} e_{b}^{4} \right) + K_{8} \left( \frac{7}{16} e_{a}^{8} + \frac{105}{32} e_{a}^{6} e_{b} + \frac{4725}{128} e_{a}^{2} e_{b}^{6} \right) + K_{10} \left( \frac{105}{128} e_{a}^{10} + \frac{525}{16} e_{a}^{4} e_{b} + \frac{315}{32} e_{a} e_{b}^{8} + \frac{1575}{256} e_{a}^{2} e_{b}^{8} \right) + \ldots + \] |
| cos 3a | \[ \frac{1}{4} e_{a}^{3} K_{3} + K_{5} \left( \frac{5}{16} e_{a}^{5} + \frac{5}{4} e_{a}^{3} e_{b}^{2} + \frac{21}{64} e_{a}^{7} + \frac{105}{32} e_{a}^{3} e_{b}^{4} \right) + K_{7} \left( \frac{21}{64} e_{a}^{7} + \frac{315}{32} e_{a}^{4} e_{b}^{3} + \frac{105}{16} e_{a}^{2} e_{b}^{5} \right) + K_{9} \left( \frac{21}{64} e_{a}^{9} + \frac{945}{64} e_{a}^{5} e_{b}^{4} + \frac{105}{16} e_{a}^{3} e_{b}^{6} \right) + \frac{189}{32} e_{a}^{7} e_{b}^{2} + \ldots + \] |
TABLE III-A (Continued)

\[
\begin{align*}
\cos 4a & \left[ \frac{1}{8} e_a K_4 + K_6 \left( \frac{3}{16} e_a + \frac{15}{16} e_a e_b \right) + K_8 \left( \frac{7}{32} e_a + \frac{105}{8} e_a e_b \right. \\
& \left. \quad + \frac{21}{8} e_a e_b \right) + K_{10} \left( \frac{15}{64} e_a + \frac{245}{64} e_a e_b \right) + \frac{525}{64} e_a e_b \\
& \left. \quad + \frac{315}{64} e_a e_b \right) + - - - + \\
\cos 5a & \left[ \frac{1}{16} e_a K_5 + K_7 \left( \frac{7}{64} e_a + \frac{21}{32} e_a e_b \right) + K_9 \left( \frac{9}{64} e_a + \frac{189}{64} e_a e_b \right. \\
& \left. \quad + \frac{63}{32} e_a e_b \right) + - - - + \\
\cos 6a & \left[ \frac{1}{32} e_a K_6 + K_8 \left( \frac{1}{16} e_a + \frac{7}{16} e_a e_b \right) + K_{10} \left( \frac{45}{512} e_a + \frac{315}{128} e_a e_b \right. \\
& \left. \quad + \frac{45}{32} e_a e_b \right) + - - - + \\
\cos 7a & \left[ \frac{1}{64} e_a K_7 + K_9 \left( \frac{9}{256} e_a + \frac{9}{32} e_a e_b \right) + - - - + \\
\cos 8a & \left[ \frac{1}{128} e_a K_8 + K_{10} \left( \frac{5}{256} e_a + \frac{45}{256} e_a e_b \right) + - - - + \\
\cos 9a & \left[ \frac{1}{256} e_a K_9 + - - - + \\
\cos 10a & \left[ \frac{1}{512} e_a K_{10} + - - - +
\end{align*}
\]
### TABLE III-A (Continued)

\[
\begin{align*}
\cos(a \pm b) & \left[ e a e b K_2 + \left( \frac{3}{2} e a e b + \frac{3}{2} e a e b \right) K_4 + \left( \frac{45}{8} e a e b + \frac{15}{8} e a e b \right) K_6 + \left( \frac{105}{8} e a e b + \frac{105}{8} e a e b \right) K_8 + \left( \frac{1575}{32} e a e b \right) K_{10} + \ldots \right] \\
\cos(2a \pm b) & \left[ \frac{3}{4} e a e b K_3 + \left( \frac{15}{8} e a e b + \frac{15}{8} e a e b \right) K_5 + \left( \frac{105}{16} e a e b \right) K_7 + \left( \frac{315}{16} e a e b \right) K_9 + \ldots \right] \\
\cos(2a \pm 2b) & \left[ \frac{3}{4} e a e b e a e b K_4 + \left( \frac{15}{8} e a e b + \frac{15}{8} e a e b \right) K_6 + \left( \frac{35}{4} e a e b \right) K_8 + \left( \frac{1575}{64} e a e b \right) K_{10} + \ldots \right] \\
\cos(3a \pm b) & \left[ \frac{1}{2} e a e b K_4 + \left( \frac{15}{8} e a e b + \frac{15}{16} e a e b \right) K_6 + \left( \frac{105}{16} e a e b \right) K_8 + \left( \frac{1575}{64} e a e b \right) K_{10} + \ldots \right]
\end{align*}
\]
### TABLE III-A (Continued)

<table>
<thead>
<tr>
<th>[ \cos(3a \pm 2b) ]</th>
<th>[ \left[ \frac{5}{8} e_a^3 e_b^2 K_6 + \left( \frac{35}{16} e_a^3 e_b^4 + \frac{105}{64} e_a^5 e_b^2 \right) K_7 + \left( \frac{315}{32} e_a^5 e_b^4 + \frac{315}{64} e_a^3 e_b^6 + \frac{189}{64} e_a^7 e_b^2 \right) K_9 \right] ]</th>
</tr>
</thead>
<tbody>
<tr>
<td>[ \cos(3a \pm 3b) ]</td>
<td>[ \left[ \frac{5}{8} e_a^3 e_b^3 K_6 + \left( \frac{35}{16} e_a^3 e_b^5 + \frac{35}{16} e_a^5 e_b^3 \right) K_8 + \left( \frac{1575}{256} e_a^5 e_b^5 + \frac{315}{64} e_a^3 e_b^7 + \frac{315}{64} e_a^7 e_b^3 \right) K_9 \right] ]</td>
</tr>
<tr>
<td>[ \cos(4a \pm b) ]</td>
<td>[ \left[ \frac{5}{16} e_a^4 e_b^2 K_5 + \left( \frac{105}{64} e_a^4 e_b^4 + \frac{21}{32} e_a^6 e_b^2 \right) K_7 + \left( \frac{315}{64} e_a^4 e_b^5 + \frac{189}{32} e_a^6 e_b^3 + \frac{63}{64} e_a^8 e_b \right) K_9 \right] ]</td>
</tr>
<tr>
<td>[ \cos(4a \pm 2b) ]</td>
<td>[ \left[ \frac{15}{32} e_a^4 e_b^2 K_6 + \left( \frac{35}{16} e_a^4 e_b^4 + \frac{21}{16} e_a^6 e_b^2 \right) K_8 + \left( \frac{1575}{256} e_a^4 e_b^5 + \frac{315}{32} e_a^6 e_b^4 + \frac{315}{128} e_a^8 e_b^2 \right) K_9 \right] ]</td>
</tr>
<tr>
<td>[ \cos(4a \pm 3b) ]</td>
<td>[ \left[ \frac{35}{64} e_a^4 e_b^3 K_7 + \left( \frac{315}{128} e_a^4 e_b^5 + \frac{63}{32} e_a^6 e_b^3 \right) K_8 \right] ]</td>
</tr>
<tr>
<td>[ \cos(4a \pm 4b) ]</td>
<td>[ \left[ \frac{35}{64} e_a^4 e_b^4 K_8 + \left( \frac{315}{128} e_a^4 e_b^6 + \frac{315}{128} e_a^6 e_b^4 \right) K_9 \right] ]</td>
</tr>
<tr>
<td>[ \cos(5a \pm b) ]</td>
<td>[ \left[ \frac{3}{16} e_a^5 e_b^2 K_6 + \left( \frac{21}{16} e_a^5 e_b^3 + \frac{7}{16} e_a^7 e_b \right) K_8 + \left( \frac{315}{64} e_a^5 e_b^5 + \frac{315}{64} e_a^7 e_b^3 + \frac{45}{64} e_a^9 e_b \right) K_9 \right] ]</td>
</tr>
</tbody>
</table>

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-130-
| \( \cos(5a \pm 2b) \) | \( \frac{21}{64} e_a e_b^2 K_7 + \left( \frac{63}{32} e_a^2 e_b + \frac{63}{64} e_a e_b^2 \right) K_9 + \quad + \) |
| \( \cos(5a \pm 3b) \) | \( \frac{7}{16} e_a e_b^3 K_8 + \left( \frac{315}{128} e_a^2 e_b + \frac{105}{64} e_a e_b^2 \right) K_{10} + \quad + \) |
| \( \cos(5a \pm 4b) \) | \( \frac{1}{2} e_a e_b^4 K_9 + \quad + \) |
| \( \cos(5a \pm 5b) \) | \( \frac{1}{2} e_a e_b^5 K_{10} + \quad + \) |
| \( \cos(6a \pm b) \) | \( \frac{7}{64} e_a e_b K_7 + \left( \frac{63}{64} e_a e_b^3 + \frac{105}{64} e_a e_b^2 \right) K_9 + \quad + \) |
| \( \cos(6a \pm 2b) \) | \( \frac{7}{32} e_a e_b^2 K_8 + \left( \frac{105}{64} e_a e_b + \frac{45}{256} e_a e_b^2 \right) K_{10} + \quad + \) |
| \( \cos(6a \pm 3b) \) | \( \frac{21}{64} e_a e_b^3 K_9 + \quad + \) |
| \( \cos(6a \pm 4b) \) | \( \frac{105}{256} e_a e_b^4 K_{10} + \quad + \) |
| \( \cos(7a \pm b) \) | \( \frac{1}{4} e_a^2 e_b K_8 + \left( \frac{45}{64} e_a e_b^3 + \frac{45}{256} e_a e_b^2 \right) K_{10} + \quad + \) |
| \( \cos(7a \pm 2b) \) | \( \frac{5}{64} e_a^2 e_b^2 K_9 + \quad + \) |
| \( \cos(7a \pm 3b) \) | \( \frac{15}{64} e_a^2 e_b^3 K_{10} + \quad + \) |
| \( \cos(8a \pm b) \) | \( \frac{5}{256} e_a^2 e_b K_9 + \quad + \) |
| \( \cos(8a \pm 2b) \) | \( \frac{45}{512} e_a e_b^2 K_{10} + \quad + \) |
TABLE III-A (Continued)

\[
\cos(9a \pm b) \left[ \frac{5}{256} e_a^9 e_b^9 K_{10} - - - + \right]
\]

TABLE IV-A

HARMONICS TO K = 10 WITH ONE INPUT ONLY

\[
\begin{align*}
\cos a & \left[ e_{a1} + \frac{3}{4} e_a^3 K_3 + \frac{5}{8} e_a^5 K_5 + \frac{35}{64} e_a^7 K_7 + \frac{63}{128} e_a^9 K_9 + - - - + \right] \\
\cos 2a & \left[ \frac{1}{2} e_a^2 K_2 + \frac{1}{2} e_a^4 K_4 + \frac{15}{32} e_a^6 K_6 + \frac{7}{16} e_a^8 K_8 + \frac{105}{256} e_a^{10} K_{10} + - - - + \right] \\
\cos 3a & \left[ \frac{1}{4} e_a^3 K_3 + \frac{5}{16} e_a^5 K_5 + \frac{21}{64} e_a^7 K_7 + \frac{21}{64} e_a^9 K_9 + - - - + \right] \\
\cos 4a & \left[ \frac{1}{8} e_a^4 K_4 + \frac{3}{16} e_a^6 K_6 + \frac{7}{32} e_a^8 K_8 + \frac{15}{64} e_a^{10} K_{10} + - - - + \right] \\
\cos 5a & \left[ \frac{1}{16} e_a^5 K_5 + \frac{7}{64} e_a^7 K_7 + \frac{9}{64} e_a^9 K_9 + - - - + \right] \\
\cos 6a & \left[ \frac{1}{32} e_a^6 K_6 + \frac{1}{16} e_a^8 K_8 + \frac{45}{512} e_a^{10} K_{10} + - - - + \right] \\
\cos 7a & \left[ \frac{1}{64} e_a^7 K_7 + \frac{9}{256} e_a^9 K_9 + - - - + \right] \\
\cos 8a & \left[ \frac{1}{128} e_a^8 K_8 + \frac{5}{256} e_a^{10} K_{10} + - - - + \right] \\
\cos 9a & \left[ \frac{1}{256} e_a^9 K_9 + - - - + \right] \\
\cos 10a & \left[ \frac{1}{512} e_a^{10} K_{10} + - - - + \right]
\end{align*}
\]
TABLE V-A
ADDITIONAL TERMS FROM COMBINATIONS CONTRIBUTING TO HARMONIC GENERATION

\[
\begin{align*}
\cos a & \left[ \frac{3}{2} e_a e_b K_3 + \frac{15}{4} e_a e_b K_4 + \frac{15}{8} e_a e_b K_5 + \frac{315}{32} e_a e_b K_7 + \frac{105}{16} e_a e_b K_9 + \frac{945}{32} e_a e_b K_8 + \frac{35}{16} e_a e_b K_9 + \frac{315}{16} e_a e_b K_9 \\
& + \frac{315}{32} e_a e_b K_9 + \frac{315}{128} e_a e_b K_9 + \cdots \right] \\
\cos 2a & \left[ \frac{3}{2} e_a e_b K_4 + \frac{15}{4} e_a e_b K_4 + \frac{105}{8} e_a e_b K_6 + \frac{45}{16} e_a e_b K_6 + \frac{525}{16} e_a e_b K_10 + \frac{315}{32} e_a e_b K_10 + \cdots \right] \\
\cos 3a & \left[ \frac{5}{4} e_a e_b K_5 + \frac{105}{32} e_a e_b K_7 + \frac{105}{32} e_a e_b K_7 + \frac{945}{64} e_a e_b K_9 + \frac{189}{32} e_a e_b K_9 + \cdots \right] \\
\cos 4a & \left[ \frac{15}{16} e_a e_b K_6 + \frac{105}{32} e_a e_b K_6 + \frac{21}{8} e_a e_b K_8 + \frac{945}{64} e_a e_b K_10 + \frac{315}{64} e_a e_b K_10 + \cdots \right] \\
\cos 5a & \left[ \frac{21}{32} e_a e_b K_7 + \frac{189}{64} e_a e_b K_9 + \frac{63}{32} e_a e_b K_9 + \frac{525}{64} e_a e_b K_9 + \cdots \right] \\
\cos 6a & \left[ \frac{7}{16} e_a e_b K_8 + \frac{315}{128} e_a e_b K_8 + \frac{45}{32} e_a e_b K_8 + \cdots \right]
\end{align*}
\]
<p>| | | | |</p>
<table>
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<tr>
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</tr>
</thead>
<tbody>
<tr>
<td>cos $7a$</td>
<td>$\frac{9}{32} e_a^7 e_b^2 K_9 + - - +$</td>
<td></td>
<td></td>
</tr>
<tr>
<td>cos $8a$</td>
<td>$\frac{45}{256} e_a^8 e_b^2 K_{10} + - - +$</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

$\cos 9a$: None to degree $n = 10$

$\cos 10a$: None to degree $n = 10$
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The results of receiver spurious response studies are presented. The spurious response bandwidth is found to be inversely proportional to the integer 'q' in the spurious response equation. The difficulties in obtaining useful spurious response data are discussed.

Three computer programs for solving certain types of interference problems are reviewed. The volumes of the Manuscript of Catalogue containing these computer programs were published during the contract period.

Tests and test procedures for obtaining spectrum signature data for U.S. Army microwave communications equipment were developed. System and terminal equipment test lists and test procedures which are not included in existing spectrum signature collection plans are described. Test sets and procedures developed for the pulse position modulated-time division multiplexer (PPMTD) are discussed. The system error rate test is presented in detail, and additional proposed system tests are briefly outlined.

The results of receiver spurious response studies are presented. Data taken on vacuum tube and diode mixers verify existing error theories and indicate a theory involving the effect of conduction angle. A prediction method for estimating the levels of mixer responses is outlined, and additional work is needed to establish its validity.

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