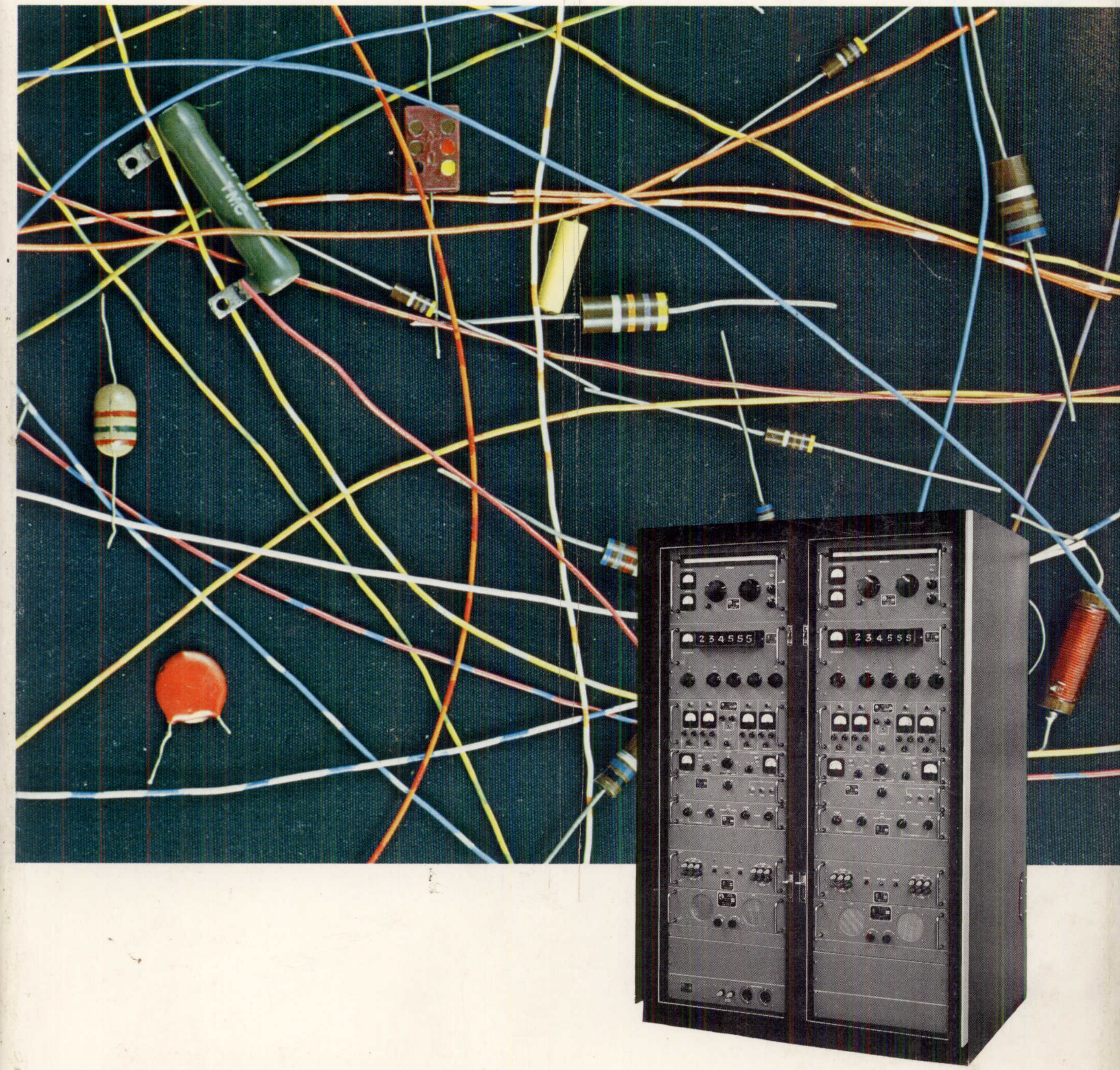
SINGLE SIDEBAND HANDBOOK



THE TECHNICAL MATERIEL CORPORATION

In 1962 the Technical Materiel Corporation instituted a factory training program for Field Engineers and customer personnel. This program has been enthusiastically received.

Each trainee attending a course at the school receives a copy of "A FIRST PRIMER DESCRIBING SSB", published by TMC (Canada). The "First Primer" is extracted from a paper delivered by Mr. D. V. Carroll, President and Managing Director of TMC (Canada), to the Quebec section of the Institute of Electronic and Electrical Engineers. It describes the advantages of Single Sideband Communications systems over conventional AM systems, and briefly discusses the techniques involved.

Persons attending the TMC factory courses have widely divergent backgrounds; these range from the novice to the highly trained and experienced. While the "First Primer" contains the basic essentials necessary to a cursory understanding of SSB techniques, it was felt that trainees and other interested personnel should be furnished with a more comprehensive SSB reference, oriented toward TMC equipment and techniques, and written from a compltely practical standpoint. This work is the result. It has been designed to serve not only as a "home study" text for the beginner, but as a ready reference for the experienced SSB Field Engineer or Technician, who must be ready at all times to conduct "on the job" training.

It is anticipated that, as new equipments are produced, and as new techniques are developed, this book will be updated. Comments, constructive criticisms, suggestions, and articles of interest are encouraged. Communications should be addressed to:

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Director of Engineering Services
700 Fenimore Road
Mamaroneck, New York

William P. Henneberry
September, 1963

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THE TECHNICAL MATERIEL CORPORATION

MAMARONECK, NEW YORK

O C T O B E R 1 9 6 5

ERRATA: THE SINGLE SIDEBAND HANDBOOK.

Dust Jacket: flyleaf; 3rd paragraph; 10th line; correct spelling of word: completely

Page 6 paragraph 2; change: F_s to f_s

Page 8 after: Alternate Formulas For Simple Symmetrical Sine Wave Modulation are; change: M to m

Page 11 beneath Alternately: change M^2 to m^2

Page 17 Captions, Figure 2-10. Change:
AUDIO FREQUENCIES: 2 KC, 4 KC, 6 KC
to: AUDIO FREQUENCY: 6 KC

Change: SINGLE TONES to SINGLE TONE

After the word Determine: change first two lines to read:

the upper side frequency
the lower side frequency

In the paragraph which begins: If the power efficiency is 60%, add the word: intelligence after the word "total" in the first line.

Delete the following two lines:

fc + 2 kc = 3 mc + 2 kc = 3.002 mc
fc + 4 kc = 3 mc + 4 kc = 3.004 mc

change the last line to read:

fc + 6 kc = 3 mc + 6 kc = 3.006 mc

Page 18 Change: a) The Lower Side Frequencies to read:
a) The Lower Side Frequency

Delete the following two lines:

fc - 2 kc = 3 mc - 2 kc = 2.998 mc
fc - 4 kc = 3 mc - 4 kc = 2.996 mc

Page 51	paragraph 5-3c: first word: change: The to the
Page 54	first formula after last paragraph; add equal sign after PAVG
Page 55	change $\frac{.5}{8}$ =.625 to $\frac{.5}{8}$ =.0625
Page 65	4th paragraph; 3rd line from bottom; change: "follower" to followed.
Page 66	paragraph beginning: A preemphasis network; correct spelling of network
Page 83	second paragraph; 2nd line; correct spelling of the word: conservatively
Page 87	change .003% to .0003%
Page 89	Figure 6-8; caption on REGENERATIVE DIVIDER block; change 250 mc to 250 KC
Page 105	under Figure 6-18E; change wording to read: SUPPRESSION DB = 20 LOG $\frac{160}{40}$ = 20 LOG 4 = 20 x .6 = 12DB
Page 112	Figure 7-3; Caption C-216 refers to 100-550 uuf balance capacitor rather than diode, as shown.
Page 113	subparagraph (4); second line; correct spelling of the word intermodulation.
Page 114	Figure 7-5; primary of T-2702 should be shown center tapped; center tap going to B Plus.
Page 132	fold out schematic following; label as Fig. 8-2
Page 143	formulas above Fig. 8-13; change: 10.35 + j(149.9 - 24) to read: 10.35 + j(149.9 - 24.9)
	formula Q = $\frac{125}{10.35}$ = .12 remove decimal point from .12
Pages 146, 147	in Figures, whenever Vp is shown, change to r_p
Page 152	last paragraph; 1st line; correct spelling of the word: bandpass
Page 155	line above: Determination of Reactances at 38 KC: change spelling of word: frequency

Page 156 Figure, top of page; E ASSUMED 21 KV, not 21 KΩ

Page 160 reverse captions, figures 9-6, 9-7

Page 162 second paragraph after: THE QUARTZ CRYSTAL AS A RESONANT ELEMENT change: followed to follow

in formula above Figure 9-11, change 159 X 10^{3-} to 159 X 10^{-3}

Page 163 second line after formulas at top of page; insert the word: in after connected

Page 164 Figure 9-13; change value of L-2 to 11.5 mhy.

below Figure 9-13 caption, change remainder of page to read:

At a frequency below the series resonant point, branches 2 and 3 will be capacitive, and branch 1 will be inductive. A new resonant frequency, f_{a-1}, will occur when:

$$X_{L2} = \frac{X_{C2} \times (X_{C1} - X_{L1})}{X_{C2} + (X_{C1} - X_{L1})}$$

Solving for L-2, assuming f_{a-1} to be 459 KC,

$$L_2 = \frac{1-\omega^2 L_1 C_1}{\omega^2 C_1 + C_2 - \omega^2 L_1 C_1 C_2}$$

In this case, L-2 is 11.5 mhy.

Page 165 Delete the first two lines on this page.

Page 166 second paragraph after Fig. 9-16; third line; correct spelling of word: circuit

Page 168 4th line from top; correct spelling of word: additional

Page 171 move up last line

Page 172 Figure 10-2; under 100 KC SELECTOR; opposite 7: change: 3.0 to 3.1

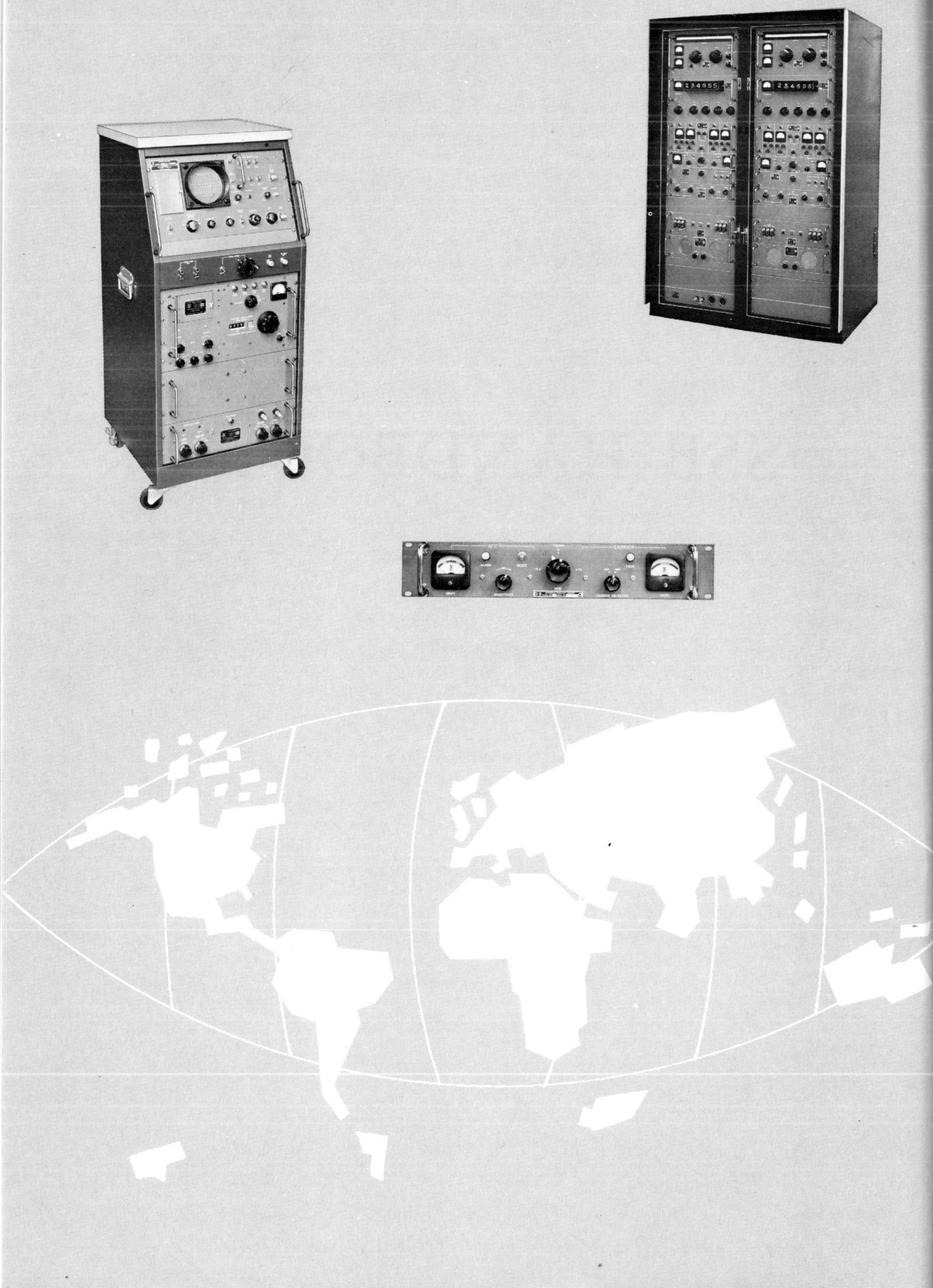
Page 177 paragraph (h): second line; change 32.5 to 3.25

Page 179 equation immediately above Figure 10-5 caption; second term should be

$$-\ (f_2 - f_1)$$

Page 184 Figure 11-1; caption on control linking RF AMPS and HFO: correct spelling of word: TUNING

Page 188	after: (2) Automatic Frequency Control; 2nd paragraph, 4th line; change: As this point to: At this point
Page 190	Figure 11-4; to right of plate, pin 9, first capacitor should be 110 uuf, rather than 110 uf
Page 194	subparagraph (j) change 100,000 uv to 10,000 uv
Page 196	subparagraph (1) line 2; change 12.7 to 12.17 mcs.
Page 205	subparagraph (g) line 3; remove parenthesis from before: See point A
Page 207	Figure 11-18; left hand column of calculations; change: 246,785 to 246,735
Page 210	Figure 11-15, fold out after page 210; change AFC-3 output from 2 mc $\pm$KC to 2 mc $\pm$3 KC change B_2 product detector injection from 254.710 to 243.710 captions, product detector injections; change caption reading: A1, B2 to: A1, B1.



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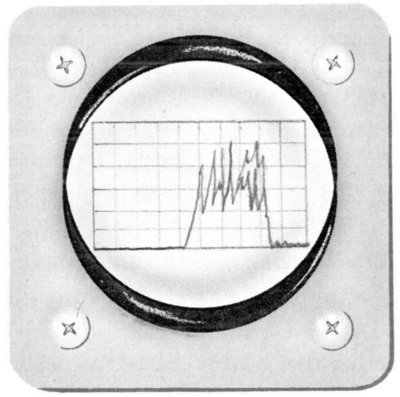

BY

WILLIAM P. HENNEBERRY

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WORLD WIDE SUPPLIERS OF ELECTRONIC COMMUNICATIONS EQUIPMENT

THE TECHNICAL MATERIEL CORPORATION

700 FENIMORE ROAD MAMARONECK, NEW YORK

PREFACE

The author has had a wide variety of experience in the application of theory, in teaching and in maintaining electronic equipment during his tenure of duty in the U. S. Navy.

Mr. Henneberry was instrumental in formulating lesson plans, study material, laboratory experiments, etc. for synthesized single sideband transmitters and receivers. He has been instrumental in imparting this information to hundreds of students, military and commercial, who have attended the single sideband training program at The Technical Materiel Corporation. His knowledge of complex circuitry coupled with his capability to put this information into understandable and simple language has been a major contributing factor to the success of our training program.

Although much of the material contained in this book centers on some of the products of the Technical Materiel Corporation, the application of these ideas, as well as the presentation of the subject material makes this book a desirable training aid for both classroom and extension course applications.

D. W. (Nick) Carter
Director of Engineering Services
THE TECHNICAL MATERIEL CORPORATION

December, 1964

DOMINIC "DOM" COSTANTINO
SENIOR TMC FIELD ENGINEER

DEDICATION

To those men who provide assistance in the installation, operation and maintenance of electronic equipment all over the world, to the traveling field engineers, this book is dedicated.

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ACKNOWLEDGEMENTS

The author expresses his thanks to Messrs. Jon Gilbertson, Richard Panasuk and Paul Grove who undertook the burden of proofreading and checking of the final manuscript; to Messrs. D. W. Carter and B. D. Pritchard, for their encouragement and technical suggestions. To Mr. Harold Johnson, who has convinced the author that it is easier to produce a finished manuscript than it is to publish a finished work.

The author wishes, also, to convey his appreciation to Mr. Ray H. dePasquale, President, Technical Materiel Corporation and to Mr. William Galione, Executive Vice President, whose votes of confidence have made this work possible.

CHAPTER 1

AN INTRODUCTION TO SSB COMMUNICATIONS

1-1 The Need for SSB Techniques

The science of communications has made rapid strides in the last half century. The invention of the amplifying vacuum tube in 1906 opened fantastic new vistas for exploration in this field; these vistas have since been widened by the increased applications found for solid state devices.

Most advances in the field of communications can be traced to the *need* for more reliable, secure and swift methods of communication from one place to another. This "advance due to need" is applicable to all branches of electronics. The well known multivibrator circuit was invented by two Frenchmen in the 1900's. The circuit remained a laboratory secret until the advent of Radar; then, the need for this device made its application commonplace.

The first known single sideband experiments were conducted in 1915; the results of these experiments were ignored by all but the telephone engineers who needed this new technique because of incredible telephone system expansion. In the last ten years, the need for the advantages of single sideband communications systems has resulted in an improvement in techniques and in component hardware necessary for this type of communication.

Most long distance radio communication is carried on in the high frequency spectrum, covering the range from approximately 2 to 30 megacycles. Some long distance communication is carried out at frequencies as low as 15 KCS; this requires transmitters of extremely high power and large physical size, with attendant costly antenna systems and real estate.

Since the high frequency spectrum is limited, it is mandatory that the most efficient use be made of the available space. Services that may be accommodated by VHF, UHF, microwave, landlines and other means, have been removed and still the HF spectrum remains crowded.

To utilize the available space to the maximum extent:

a) guard channels between adjacent bands must be made as narrow as possible. The guard channels are required to allow for frequency drift of transmitters.

b) transmitter stability must be improved. This follows, naturally, because of the reduction in width of the guard channels. In addition, multiplexing, becoming more and more common, requires almost absolute transmitter stability.

c) transmitter bandwidths must be kept as narrow as possible yet they must carry a maximum amount of intelligence.

d) spurious radiations outside the transmitter bandwidth must be eliminated to prevent spillover into adjacent channels.

The present state of the art of single sideband communications fulfills these needs. In addition, rapid advances are being made which will make even better use of the available high frequency spectrum space. These advances include radio controlled transmitter stabilization, improved frequency selective filters and sophisticated multiplexing arrangements.

1-2 Brief History of Single Sideband Communications

In 1914 it was mathematically deduced that an amplitude modulated wave consisted of a carrier frequency and a pair of sideband frequencies for each audio modulating frequency. This information has been accepted for years; but in 1914 violent argument raged. The "anti" sideband groups contended that the sidebands were mathematical fiction and did not exist in fact.

In 1915 an important experiment was conducted by Mr. H. P. Arnold at the U. S. Naval Radio Station at Arlington, Virginia. Mr. Arnold tuned an antenna to pass one sideband, while attenuating the other. Thus it was established that the carrier and sidebands were separate entities.

Subsequently, Mr. John R. Carson of the American Telephone and Telegraph Company conducted further tests and reached the conclusions that:

a) both sidebands contain identical intelligence, and are, in fact, mirror images of each other.

b) of the intelligence radiated by the antenna, the greatest amount of power was radiated by the carrier.

c) the carrier contained no useful intelligence, and served only as a reference for the sidebands.

The inference here is clear. If the carrier consumes most of the power, yet serves no useful purpose, and if both sidebands are identical, why not eliminate the carrier and one sideband, and transmit a single sideband?

In the receiver, the carrier could be recombined with the received single sideband signal, and the original audio intelligence recovered.

That is exactly what Mr. Carson did, except that he used both sidebands, because, for his applications, the additional spectrum was required.

He developed an electronic circuit called a "Balanced Modulator", which mixes a carrier with audio intelligence frequencies, then balances out the carrier, leaving the sidebands intact. This device was patented in 1923. Using the Carson techniques, the American Telephone and Telegraph Company established a submarine cable in 1927. With multiplexing arrangements, the sidebands carried many channels of information. No power was wasted in the carrier, and the available spectrum was used to the best possible advantage.

Engineers continued to improve on sideband techniques. Most developments were oriented toward overseas cable and radiotelephone service. The development of SSB systems as we know them was hampered because of poor frequency stability, poor filter selectivity, and because low distortion linear power amplifiers were unknown.

Radio Amateurs entered the SSB field in the early 1930's. By the early 1950's the art had progressed sufficiently for the military services to "go sideband".

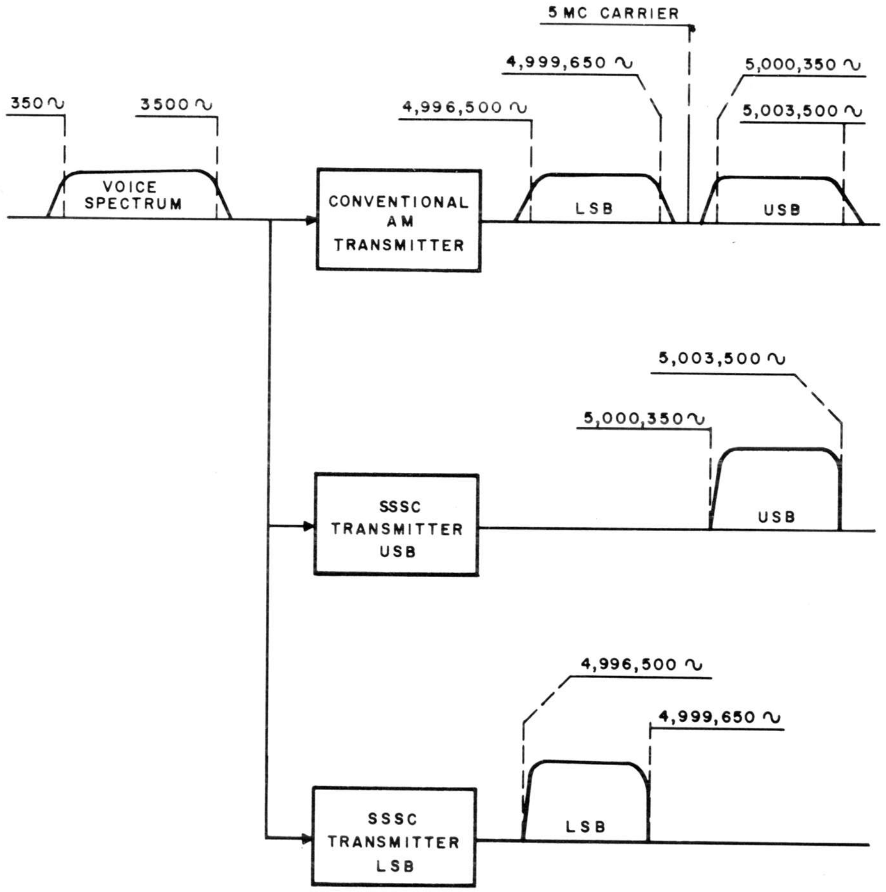
1-3 Tentative Definition of a Single Sideband Signal

The definition to be presented here is tentative, and will not conflict with later definitions. A tentative definition is necessary because the term "Single Sideband", (SSB), has become generalized and refers also to sideband signals which are not truly "Single Sideband, Suppressed Carrier", (SSSC), signals.

A SINGLE SIDEBAND SUPPRESSED CARRIER (SSSC) SIGNAL IS A BAND OF AUDIO INTELLIGENCE FREQUENCIES WHICH HAS BEEN TRANSLATED TO A BAND OF RADIO FREQUENCIES WITHOUT DISTORTION OF THE INTELLIGENCE.

Note that no mention is made of a carrier frequency. Note also that no mention is made of the terms: "Upper Sideband", (USB), or "Lower Sideband", (LSB), since either one of these may be transmitted or received.

1-4 Pictorial Representation of a Typical Voice Spectrum, with Resultant Conventional AM Spectrum and SSSC Spectra. The assigned carrier frequency is 5.0 mcs.



CHAPTER 2

A REVIEW OF AMPLITUDE MODULATION

2-1 Definition of Amplitude Modulation

Amplitude Modulation is defined as the process by which the amplitude of the radiated wave is varied in accordance with the intelligence to be transmitted.

2-2 A Discussion of the Equation of an Amplitude Modulated Wave Employing Single Sine Wave Modulation.

The equation of an amplitude modulated wave employing single sine wave modulation is presented below:

$$e = Eo \sin 2\pi ft + \frac{m\,Eo}{2} \cos 2\pi(f - fs)t - \frac{m\,Eo}{2} \cos 2\pi(f + fs)\,t$$

where: e is the instantaneous amplitude of the radiated wave.

Eo is the maximum value of the carrier amplitude.

f is the carrier frequency.

m is the degree of modulation, ordinarily expressed as a percentage.

fs is the single sine wave modulating frequency.

The first term: Eo sin $2\pi ft$ tells us that the carrier frequency's instantaneous amplitude varies sinusoidally with time, and that the carrier amplitude is independent of the degree of modulation. The carrier is present, then, with or without modulation, and maintains a constant average amplitude at a frequency, f.

The second term: $\frac{m\,Eo}{2}$ cos $2\pi(f-fs)t$ tells us that the radiated wave contains a new frequency, (f—fs). This frequency is lower than the carrier frequency by an amount, Fs. The amplitude of this component depends on the degree of modulation, and, when m is 100%, it is half the carrier amplitude. This is called the lower side frequency. The third term: $\frac{m\,Eo}{2}$ cos 2π (f+fs)t tells us that the radiated wave contains a new frequency, (f+fs). This frequency is higher than the carrier frequency by an amount, fs. The amplitude of this component depends on the degree of modulation, and, when m is 100%, it is half the carrier amplitude.

Note that the original modulating frequency is not contained in the radiated wave.

The original modulating frequency, fs, is present at the plate of the final RF amplifier of the transmitter, but the response of the tuned circuits following is such that this component is lost.

The equation becomes complex when more than one sine wave is used to modulate the carrier, or when non sinusoidal modulating frequencies are employed.

When the carrier is modulated with two sinusoidal frequencies, two upper and two lower side frequencies are produced. When the carrier is modulated with voice signals, two sidebands are produced; one above the carrier and one below it. For each individual voice tone, there are two sidetones, each displaced from the carrier by an amount equal to the tone frequency. Since the amplitude of each tone of a voice pattern is different, the degree of modulation of each pair of resultant sidetones is different.

A pictorial representation illustrating three frequency spectrums is shown in figure 2-1.

From the pictorial representation of figure 2-1 an important additional fact about amplitude modulation may be noted:

THE BANDWIDTH REQUIRED FOR AN AMPLITUDE MODULATED SIGNAL IS TWO TIMES THE HIGHEST MODULATING FREQUENCY EMPLOYED

At "A", the modulating frequency is 1 KC; the total bandwidth required is 2 KCS.

At "B" and "C", the highest modulating frequency is 10 KCS; the total bandwidth required is 20 KCS.

2-3 The Degree of Modulation (See Figure 2-2)

Assume that a transmitter is operating at a carrier frequency of 4.0 mcs, and that the output is being monitored on a conventional oscilloscope.

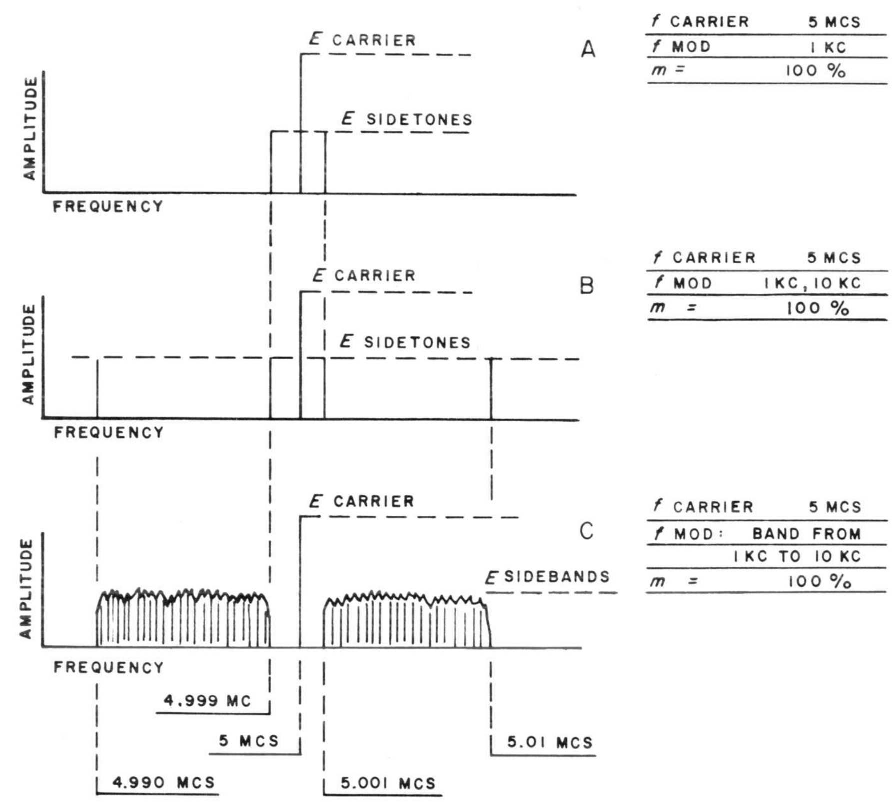


Figure 2-1.

The oscilloscope presentation is shown at "B". Note that before time t-1, the pattern is a simple sine wave with a maximum amplitude, Eo, of 500 volts. This represents the carrier without modulation. At "A", t-0, a carrier vector is shown of 500 volts amplitude with no side frequencies.

At time t-1, the modulation frequency of 2 KCS is applied; two side frequencies are created; the side frequency vectors "ride" the carrier vector. The USB vector travels in a positive direction counterclockwise, and the LSB vector travels clockwise. At time t-1, the side frequency voltages cancel, leaving the resultant wave at 500 volts.

At time t-2, the side frequency vectors and the carrier vector add to 1000 volts. This corresponds to E max at "B".

At time t-3, the side frequency vectors again cancel, and the voltage at "B" is Eo.

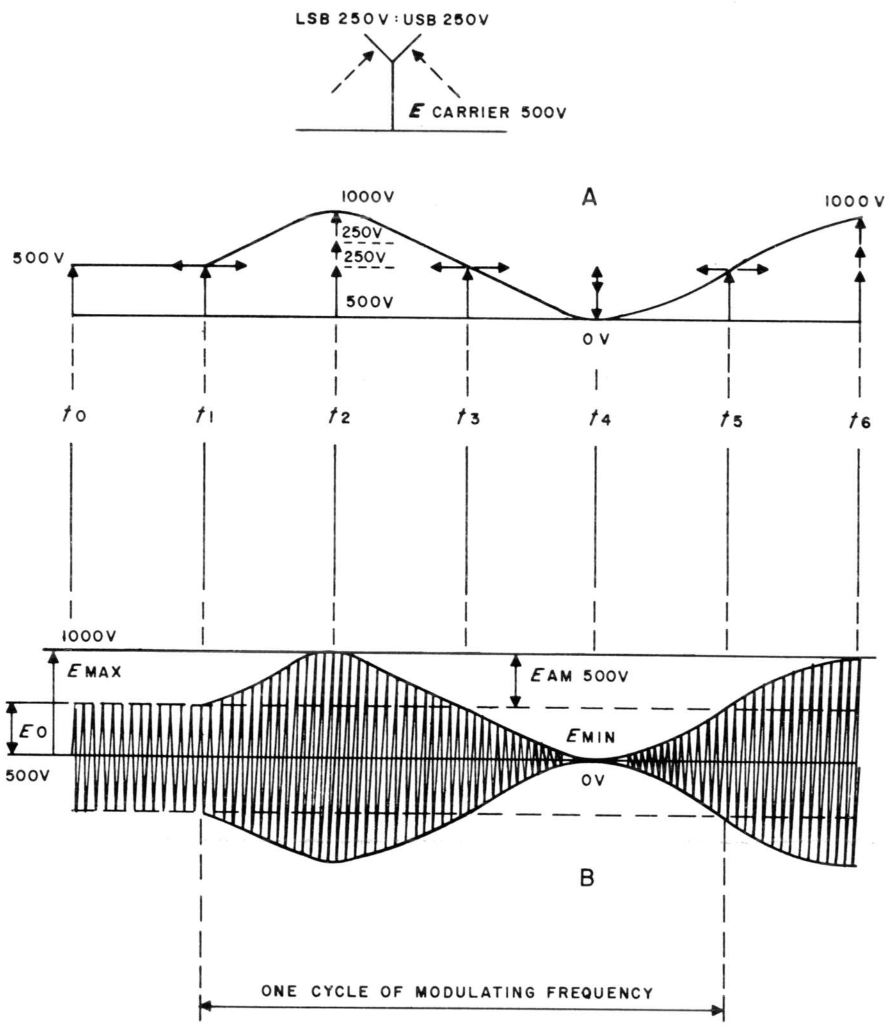
At time t-4, the side frequency vectors add in such a manner as to cancel out the carrier vector. This corresponds to E min at "B".

The composite waveform at "B" is the resultant of the in and out of phase addition of the three discrete frequencies making up the radiated wave.

The degree of modulation, m, expressed as a percentage, for simple sinusoidal modulation, is given by:

$$m = \frac{Eo - E\ min}{Eo} \times 100$$

Figure 2-2. Plot of a Radiated Wave Modulated by a Single Symmetrical Sine Wave Frequency.



Alternate formulas for simple symmetrical sine wave modulation are:

$$M = \frac{\text{E max} - \text{E min}}{2\ \text{Eo}} \times 100 \quad \text{and} \quad m = \frac{\text{E max} - \text{Eo}}{\text{Eo}} \times 100$$

For any kind of modulating signal:

$$m = \frac{\text{E max} - \text{E min}}{\text{E max} + \text{E min}} \times 100$$

For the particular values given in Figure 2-2, each formula yields: m = 100%.

$$m = \frac{\text{Eo} - \text{E min}}{\text{Eo}} \times 100 = \frac{500 - 0}{500} \times 100 = 100\%$$

$$m = \frac{\text{E max} - \text{E min}}{2\ \text{Eo}} \times 100 = \frac{1000 - 0}{1000} \times 100 = 100\%$$

$$m = \frac{\text{E max} - \text{Eo}}{\text{Eo}} \times 100 = \frac{1000 - 500}{500} \times 100 = 100\%$$

$$m = \frac{\text{E max} - \text{E min}}{\text{E max} + \text{E min}} \times 100 = \frac{1000 - 0}{1000 + 0} \times 100 = 100\%$$

The maximum amplitude of the original modulating voltage is 500 volts (E am). This is evenly distributed: 250 volts per sideband. This would indicate that the total modulating power is distributed evenly in each sideband.

Note that each sideband voltage is: $\frac{\text{m Eo}}{2}$ as indicated in the equation presented in Paragraph 2-2.

Notice also that E am equals Eo for 100% modulation. When E am is less than Eo, m is less than 100%. This is illustrated in Figure 2-3.

Since all of the intelligence is contained in the sidebands, the higher the degree of modulation, the higher the "intelligence power" transmitted.

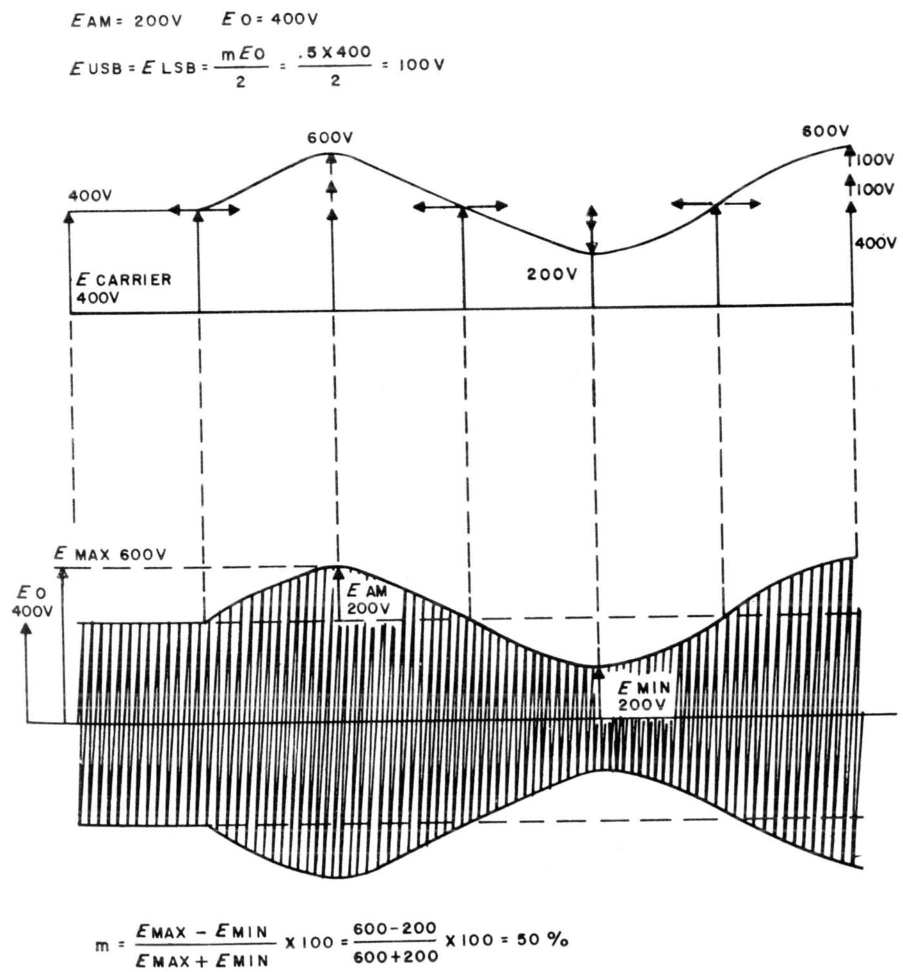


Figure 2-3. Plot of a Radiated Wave Modulated by a Single Symmetrical Sine Wave Frequency. The degree of Modulation, m, is 50%.

2-4 Overmodulation

With no modulation voltage applied, m is 0%. When E am equals Eo, the envelope crest is 2 Eo and the envelope trough is 0. This represents 100% modulation. When E am is less than Eo, but greater than 0, the modulation percentage is less than 100%.

Should E am exceed Eo, the trough amplitude (which cannot be negative) remains at O for a longer period than it should; distortion results. This is illustrated below, in Figure 2-4.

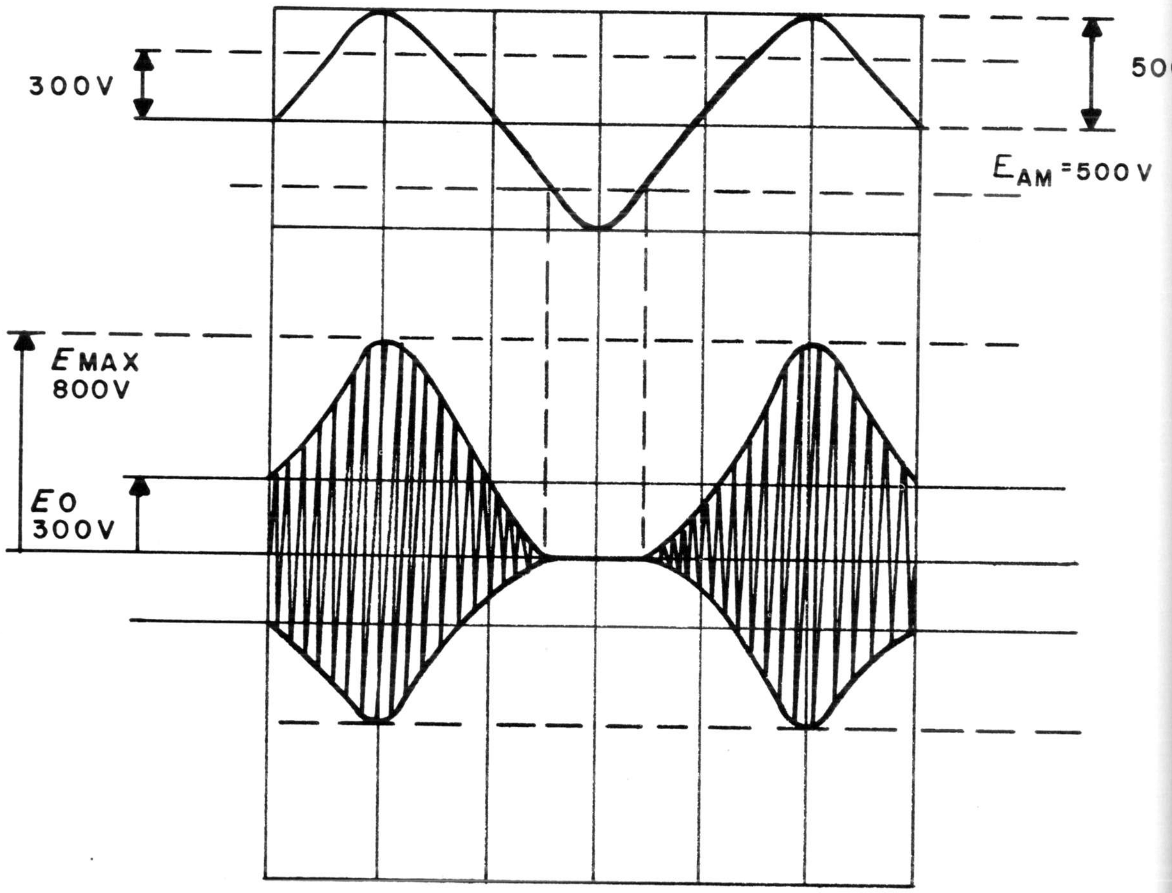


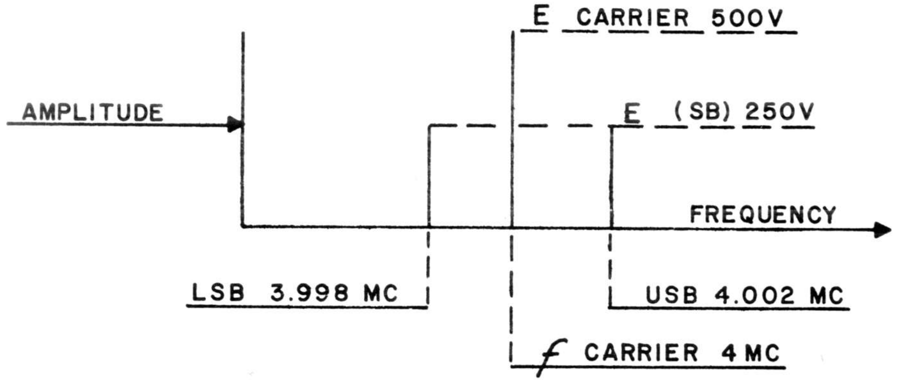
Figure 2-4. Overmodulation.

Thus, the maximum percentage of modulation allowed for an AM transmitter is 100. The percentage or degree of modulation is limited by the peak amplitude of the strongest modulating signal, with respect to the carrier amplitude.

The MODULATION CAPABILITY of an AM transmitter is the maximum percentage to which the transmitter may be modulated before spurious sidebands are generated in the output, or before distortion becomes objectionable.

2-5 Power Distribution in an Amplitude Modulated Wave

Consider an amplitude modulated wave applied to a flat 50 ohm transmission line in an antenna feed system. For simplicity, the conditions of Figure 2-2 will be used. The carrier amplitude, Eo, is 500 volts. The modulation voltage, E am is 500 volts. The percent of modulation is 100. Each sideband voltage is 250 volts. The frequency spectrum is shown below.



$$P\ \text{carrier} = \frac{E\ \text{carr}^2}{R} = \frac{(5 \times 10^2)^2}{5 \times 10^1} = \frac{25 \times 10^4}{5 \times 10^1} = 5\ \text{KW}$$

$$P\ \text{usb} = \frac{E\ \text{usb}^2}{R} = \frac{(2.5 \times 10^2)^2}{5 \times 10^1} = \frac{6.25 \times 10^4}{5 \times 10^1} = 1.25\ \text{KW}$$

$$P\ \text{lsb} = \frac{E\ \text{lsb}^2}{R} = \frac{(2.5 \times 10^2)^2}{5 \times 10^1} = \frac{6.25 \times 10^4}{5 \times 10^1} = 1.25\ \text{KW}$$

Alternately:

$$P\ \text{each sb} = \frac{M^2\ P\ \text{carr}}{4} = \frac{1 \times 5 \times 10^3}{4} = 1.25\ \text{KW}$$

The radiated wave contains a total power of:

$$P\ \text{total} = P\ \text{carrier} + P\ \text{usb} + P\ \text{lsb} = 7.5\ \text{KW}$$

The power in *each* sideband is $\frac{1.25\ \text{KW}}{7.5\ \text{KW}}$ or 16⅔% of the *total radiated power*.

The *total sideband power is* $\frac{2.50 \text{ KW}}{7.5 \text{ KW}}$ or 33⅓% of the *total radiated power.*

The ratio of Power Output Modulated to Power Output Unmodulated is $\frac{7.5 \text{ KW}}{5.0 \text{ KW}}$ or 1.5 (See Figure 2-5)

Return for a moment to Figure 2-2. With no modulation voltage, only the carrier is present. This consumes 5 KW of power *whether we are transmitting intelligence or not.*

At the crest of the modulation cycle, the envelope voltage reaches 2 Eo or 1000 volts. This represents a peak power of:

$$P \text{ peak} = \frac{E \max^2}{R} = \frac{(1 \times 10^3)^2}{5 \times 10^1} = \frac{1 \times 10^6}{5 \times 10^1} = .2 \times 10^5 = 20 \text{ KW}$$

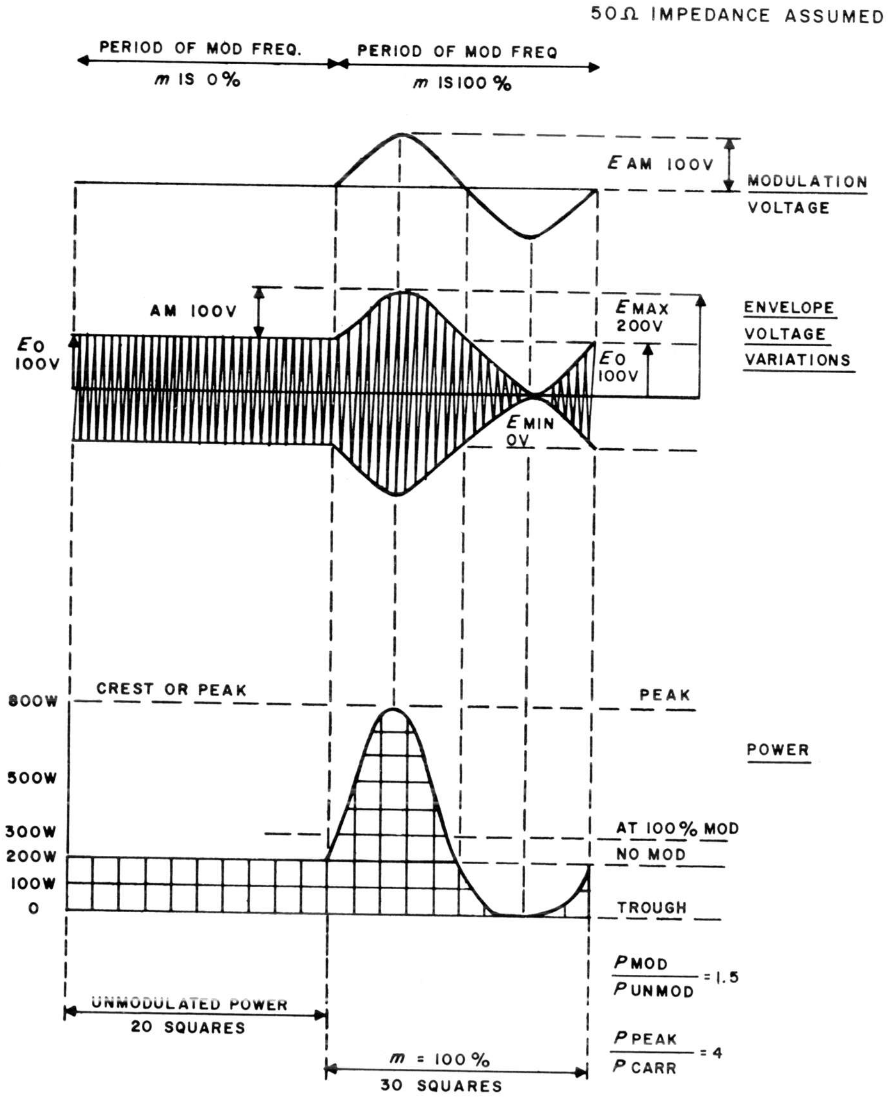


Figure 2-5. Power Relations in an Amplitude Modulated Wave.

Thus, for a total output of 7.5 KW, of which 2.5 KW contains useful intelligence, the transmitter must be designed to handle power peaks of 20 KW, assuming 100% efficiency of the system. Actually, the transmitter must be capable of handling more.

Let us be generous and assume that the power efficiency of the system is 70%. Then, for 20 KW out, the system must supply:

$$\frac{20 \times 10^3}{7 \times 10^{-1}} = 28.6 \text{ KW (on peaks) or } \frac{7.5\text{KW}}{.7} = 10.7 \text{ KW (ave.)}$$

Let us now express as a percentage, the ratio of useful power to total power input:

$$\frac{\text{Useful Power Out}}{\text{Total Power In}} = \frac{2.5 \text{ KW}}{\frac{7.5 \text{ KW}}{.7}} = 23.4\%$$

2-6 Increase of Antenna Current and Power Output for Various Degrees of Amplitude Modulation

The sketch below shows a Class B Audio Power Amplifier being used to modulate a Class C RF Amplifier. An ammeter is inserted in the antenna circuit to indicate Io, the antenna current without modulation, and I am, the antenna current with amplitude modulation.

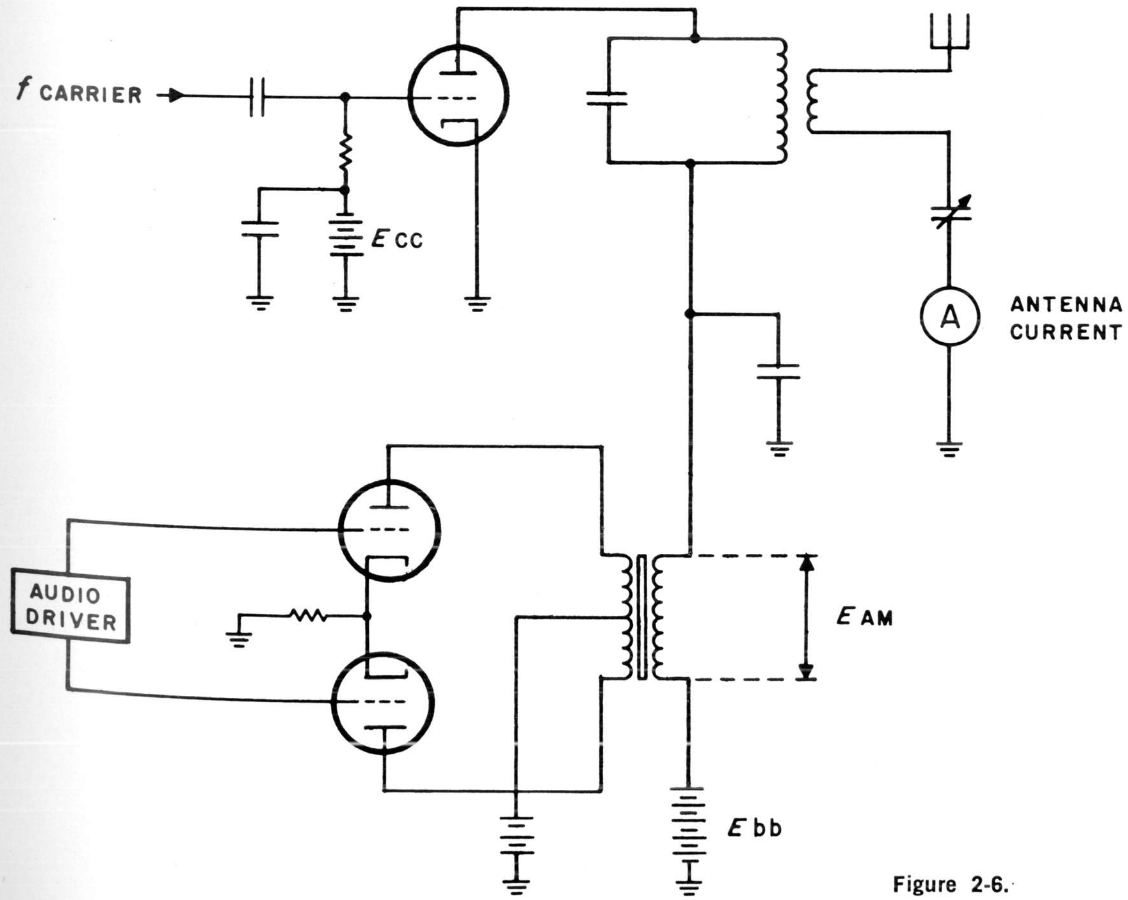


Figure 2-6.

With no modulation, m is 0%. The output power is the carrier power. Generally:

$$P = I^2R \quad \text{and} \quad I = \sqrt{\frac{P}{R}}$$

Io, the antenna current with no modulation is given by:

$$Io = \sqrt{\frac{Pc}{Ra}}$$

where: Pc is the carrier power (without modulation)
Ra is the antenna resistance.

When modulation is applied, the total sideband power is:

$$\frac{m^2\ Pc}{2}$$

Then, I am, the antenna current with modulation, is:

$$I\ am = \sqrt{\frac{Pc + \frac{m^2Pc}{2}}{Ra}} \quad \text{Factoring, } I\ am = \sqrt{\frac{Pc\left(\frac{m^2}{2} + 1\right)}{Ra}}$$

Then:

$$I\ am = Io\sqrt{\frac{m^2}{2} + 1}$$

For $m = 0$ $\quad I\ am = Io$

For $m = 50\%$ $\quad I\ am = Io\sqrt{1.125} = 1.061\ Io$

For $m = 100\%$ $\quad I\ am = Io\sqrt{1.5} = 1.225\ Io$

Thus: as m is increased from 0 to 50%, antenna current rises from Io to 1.061 Io. As m is increased to 100%, antenna current rises to 1.225 Io.

Since $P = I^2R$ and, since Ra remains constant, output power is:

1.125 Pc at 50% amplitude modulation.

1.5 Pc at 100% amplitude modulation.

An examination of Figure 2-5 will verify this.

2-7 Use of the Conventional Oscilloscope to Determine the Percent of Amplitude Modulation, and to Examine the Output for Overmodulation

A conventional oscilloscope may be used to determine the percent of amplitude modulation, and to examine the output envelope for evidence of distortion. There are two general methods:

a) method A provides the conventional elliptical pattern.
b) method B provides a trapezoidal pattern.

Method A is illustrated in figure 2-7.

A loose coupling from the antenna circuit is connected to the vertical deflection plates of the oscilloscope direct. The lower plate should be

grounded and the hot lead connected to the upper plate. If the coupling is loose enough, the connection may be made to the vertical input jacks on the front panel of the scope. The connection will depend on the power of the transmitter, the coupling, and the type of input attenuators used in the scope.

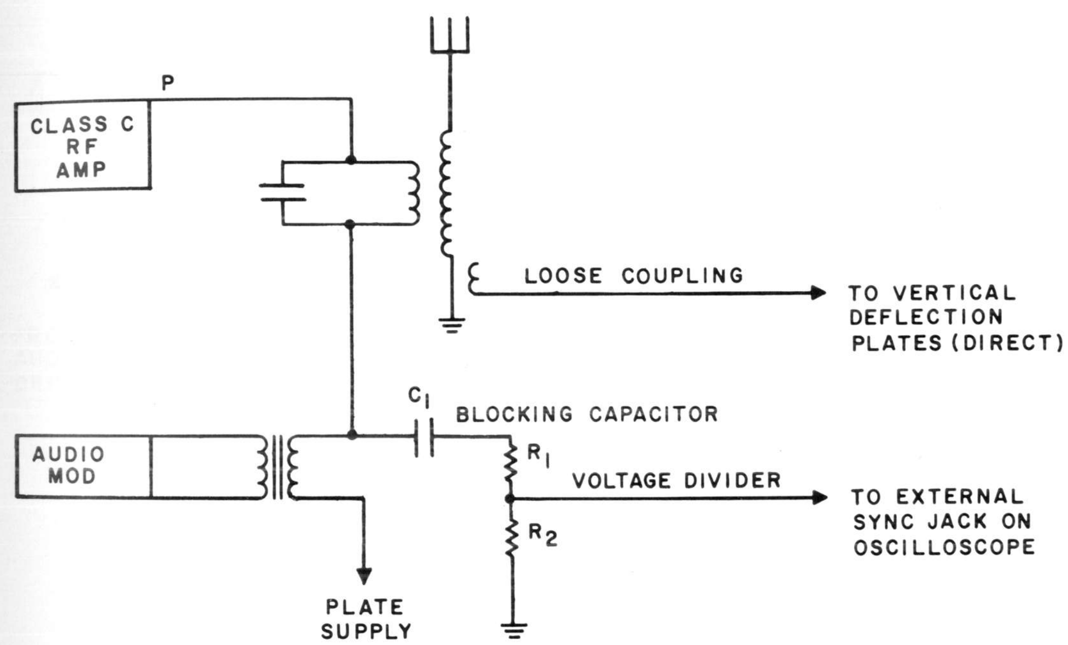


Figure 2-7.

A blocking capacitor and voltage divider circuit is connected to the high side of the modulation transformer. The blocking capacitor must be of such value as to withstand two times the voltage of the plate supply. The values of R-1 and R-2 should be such as to furnish just sufficient modulator voltage to "sync in" the waveform.

The internal sweep generator should be used. The sweep frequency should be about half the modulation frequency, so that two cycles will appear on the scope.

When controls have been adjusted properly, a waveform similar to that shown in figure 2-8 will be observed.

To determine the percent of modulation, m:

a) count the greatest number of squares covered by the modulation envelope, and let them equal E max.
b) count the least number of squares covered by the modulation envelope, and let them equal E min.
c) determine the percent of modulation from:

$$m = \frac{E\ max - E\ min}{E\ max + E\ min} \times 100$$

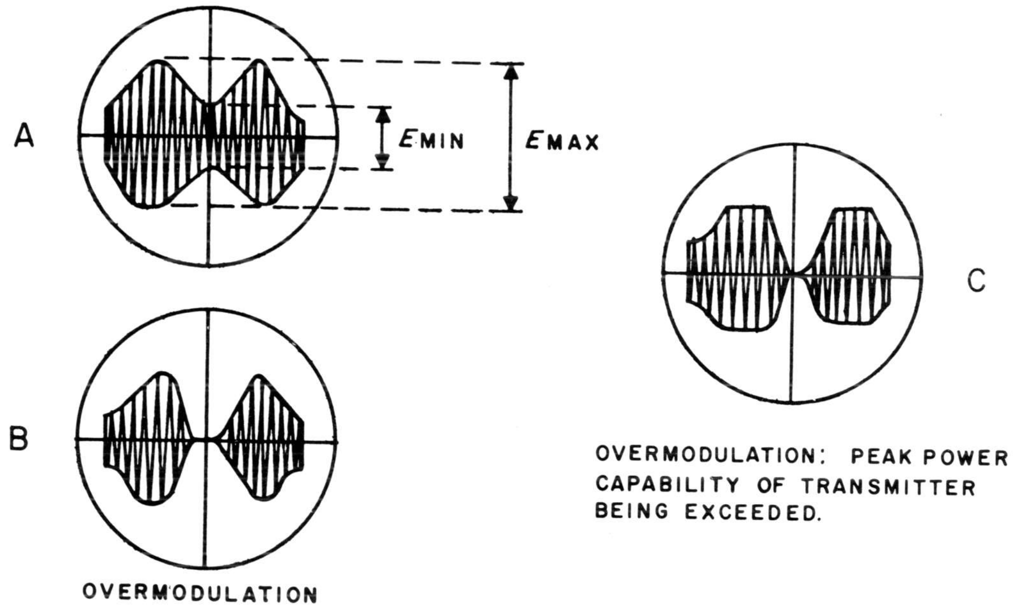


Figure 2-8.

METHOD B:

In method "B", the loose coupling from the antenna circuit is connected as before. Modulation voltage is taken from the voltage divider and fed to the horizontal amplifiers of the oscilloscope.

Thus, the output modulation envelope drives the vertical plates of the oscilloscope and the modulating signal drives the horizontal plates. The internal sawtooth sweep circuits are disabled. Vertical and horizontal gains are adjusted to provide a pattern similar to the one shown in figure 2-9.

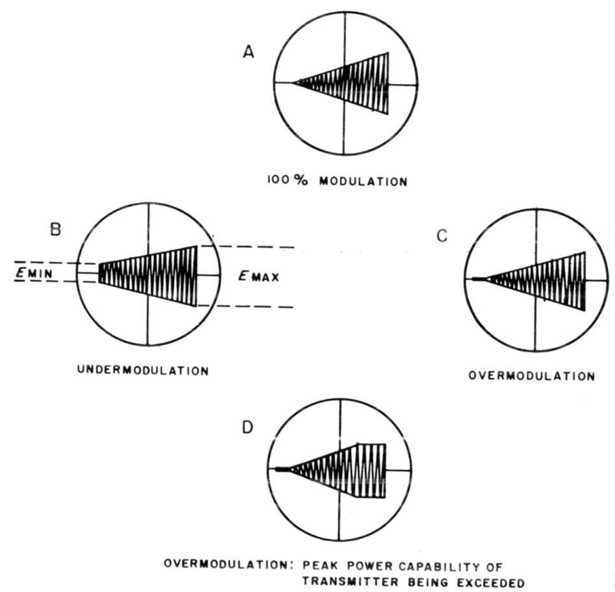


Figure 2-9.

2-8 Review Problems on Amplitude Modulation, with Answers

1. A Class B Modulator and a Class C RF Amplifier are connected as shown in figure 2-10.

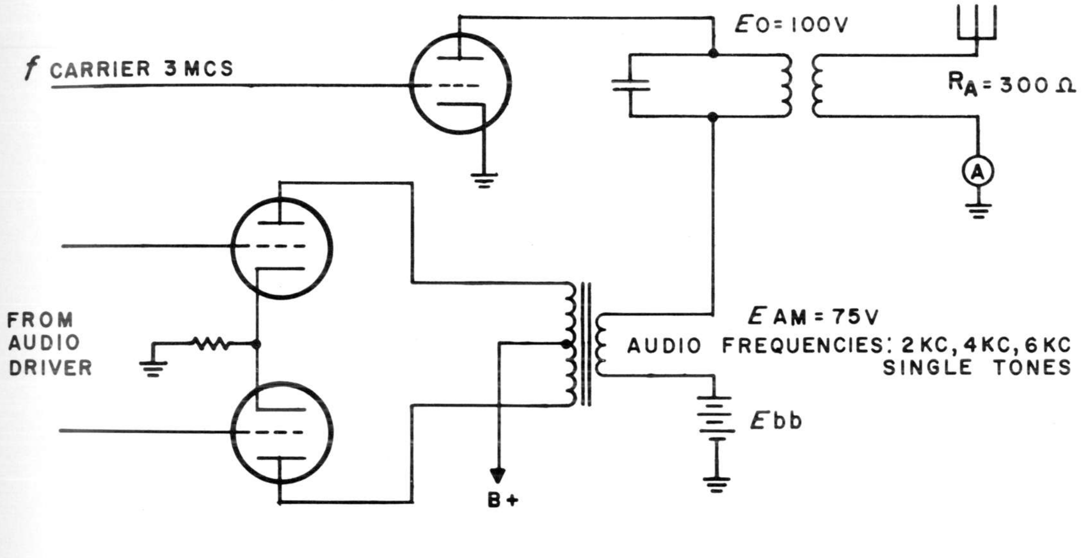


Figure 2-10.

Determine: the upper side frequencies
the lower side frequencies.
the total bandwidth required.
E max.
E min.
m, the % of modulation.
P carrier.
P each sideband.
P total output under modulation.
Io, antenna current without modulation.
I am, antenna current with modulation.
If the power efficiency is 60%, find the ratio of total power output to total power input, and express it as a percentage.

a) *The Upper Side Frequencies:*

fc + 2 kc = 3 mc + 2 kc = 3.002 mc

fc + 4 kc = 3 mc + 4 kc = 3.004 mc

fc + 6 kc = 3 mc + 6 kc = 3.006 kc

b) *The Lower Side Frequencies:*

fc — 2 kc = 3 mc — 2 kc = 2.998 mc

fc — 4 kc = 3 mc — 4 kc = 2.996 mc

fc — 6 kc = 3 mc — 6 kc = 2.994 mc

c) Total Bandwidth Required:

b w = 2x the highest mod. frequency = 2 x 6 kc = 12 kc

d) E max

E max = Eo + E am = 100v + 75v = 175v

e) E min:

E min = Eo — E am = 100v —75v = 25v

f) % of Modulation m

$$m = \frac{\text{E max} - \text{E min}}{\text{E max} + \text{E min}} \times 100 = \frac{175-25}{175+25} \times 100 = \frac{150}{200}$$

$$= \frac{3}{4} \times 100 = 75\%$$

g) P carrier:

$$\text{P carr.} = \frac{\text{Eo}^2}{\text{Ra}} = \frac{(1\times10^2)^2}{3\times10^2} = \frac{1\times10^4}{3\times10^2}$$

$$= .333 \times 10^2 = 33.3 \text{ watts}$$

h) P each Sideband:

$$\text{P each sideband} = \frac{\text{m}^2\ \text{Pc}}{4} = \frac{(.75)^2 \times 33.3}{4} = 4.68 \text{ watts}$$

i) P Total Output under Modulation:

Pt = Pc + P usb + P lsb = 33.3 + 4.68 + 4.68 = 42.66 watts

j) Io, Antenna Current without Modulation:

$$\text{Io} = \frac{\text{Eo}}{\text{Ra}} = \frac{100}{300} = .333 \text{ amps}$$

Alternately:

$$\text{Io} = \sqrt{\frac{\text{Pc}}{\text{Ra}}} = \sqrt{\frac{33.3}{300}} = \sqrt{.111} = .333 \text{ amps}$$

k) I am, Antenna Current with Modulation:

$$\text{I am} = \text{Io}\sqrt{\frac{\text{m}^2}{2}+1} = .333\sqrt{\frac{(.75)^2}{2}+1} =$$

$$.333 \times 1.131 = .377 \text{ amps.}$$

Alternately:

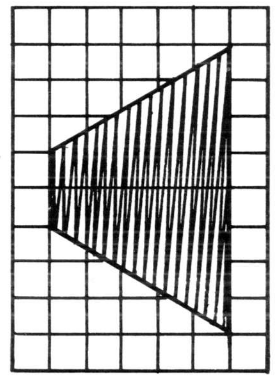
$$\text{I am} = \sqrt{\frac{\text{P total}}{\text{Ra}}} = \sqrt{\frac{42.66}{300}} = \sqrt{.1422} = .377 \text{ amps}$$

l) Percent Useful Output over Total Input:

$$\text{P input} = \frac{\text{P total out}}{60\%} = \frac{42.66}{.6} = 71.1 \text{ watts}$$

$$\frac{\text{P useful out}}{\text{P total in}} \times 100 = \frac{9.36 \text{ watts}}{71.1 \text{ watts}} \times 100 \cong 13\%$$

2. Determine the percent of modulation from the oscilloscope pattern given below.



$$m = \frac{8-2}{8+2} \times 100 =$$

$$\frac{6}{10} \times 100 = 60\ \%$$

3. A 100% modulated AM transmitter has the following characteristics: Eo, 500 volts. E am, 500 volts. Ra, 600 ohms. Assuming a 100% power efficiency, what peak power must the transmitter be prepared to handle?

E max = Eo + E am = 500v + 500v = 1000v

$$\text{P peak} = \frac{\text{E max}^2}{\text{Ra}} = \frac{(1\times10^3)^2}{6\times10^2} = \frac{1\times10^6}{6\times10^2}$$

$$= .166 \times 10^4 = 1.66 \text{ kw}$$

4. If the transmitter of Problem #3 has a power efficiency of 80%, what peak power must the transmitter be prepared to handle?

1.66 kw = 80% peak power actually supplied

$$\text{actual peak power supplied } \frac{1.66 \text{ kw}}{.8} = 2.08 \text{ kw}$$

5. A transmitter has an unmodulated antenna current, Io, of 4 amperes. The antenna current with modulation, I am, is 4.5 amperes. Find the percent of modulation.

$$\text{I am} = \text{Io}\sqrt{\frac{\text{m}^2}{2}+1} \qquad \text{I am}^2 = \text{Io}^2\left(\frac{\text{m}^2}{2}+1\right)$$

$$\text{I am}^2 = \frac{\text{Io}^2\ \text{m}^2 + 2\ \text{Io}^2}{2} \qquad 2\ \text{I am}^2 = \text{Io}^2\ \text{m}^2 + 2\ \text{Io}^2$$

$$\frac{2\ \mathrm{I\,am}^2 - 2\ \mathrm{Io}^2}{\mathrm{Io}^2} = m^2 \qquad m = \sqrt{\frac{2\ \mathrm{I\,am}^2 - 2\ \mathrm{Io}^2}{\mathrm{Io}^2}}$$

$$m = \sqrt{\frac{40.5 - 32}{16}} = \sqrt{2.531 - 2} = \sqrt{.531} = 72.8\%$$

6. Assuming that a system with an efficiency of 100% could be built, and amplitude modulated 100%. What is the maximum theoretical percentage of:

$$\frac{\text{Power Out (useful)}}{\text{Power Input (total)}} \ ?$$

$$\frac{\text{P useful in}}{\text{P input}} = \frac{\text{P total sideband}}{\text{P carrier + sidebands}}$$

P total sidebands $= .5$ P carrier

P carrier $= 1$

P input $= 1.5$

$$\% \ \frac{\text{P useful out}}{\text{P input}} = \frac{.5}{1.5} \times 100 = 33.3\%$$

CHAPTER 3

ELEMENTS OF SINGLE SIDEBAND

3-1 A Note Concerning Definitions and Terms

"Single Sideband" (SSB), is a general descriptive term encompassing a wide variety of sideband concepts. A true single sideband signal is, actually, a "Single Sideband Suppressed Carrier" (SSSC) signal. Such a signal contains one sideband, with the carrier suppressed to the point of non existence.

Let us consider a service operated with two sidebands, each sideband containing separate intelligence. For example, the lower sideband might contain 16 FSK teletype tone channels, and the upper sideband might contain a full voice spectrum. The carrier might be suppressed 20 DB below the peak sideband power, to operate an automatic frequency control unit at the receiving point. This type of operation is not SSSC, neither is it AM. It is given the name: "Independent Sideband, Reduced Carrier," (ISB, Reduced Carrier). This system falls within the general heading of SSB operation, but, since it is not truly SSSC operation, it must take another title.

It is possible to obtain other variations of SSB operation. These variations will be discussed in detail in subsequent chapters. This chapter deals with essentials, that is, with true SSSC signals.

Subsequently, the term SSB will be used to indicate, in general, any type of "sideband" operation.

3-2 Essential Elements of a SSSC System

A single sideband suppressed carrier (SSSC) transmitter translates intelligence at audio frequencies to desired radio frequencies. The upper or lower sideband may be transmitted; the carrier is suppressed to the point of non existence. See Figure 3-1, below:

Figure 3-1 shows an audio spectrum, from 350 to 3500 cycles, being applied to two separate SSSC transmitters. The translation to the assigned

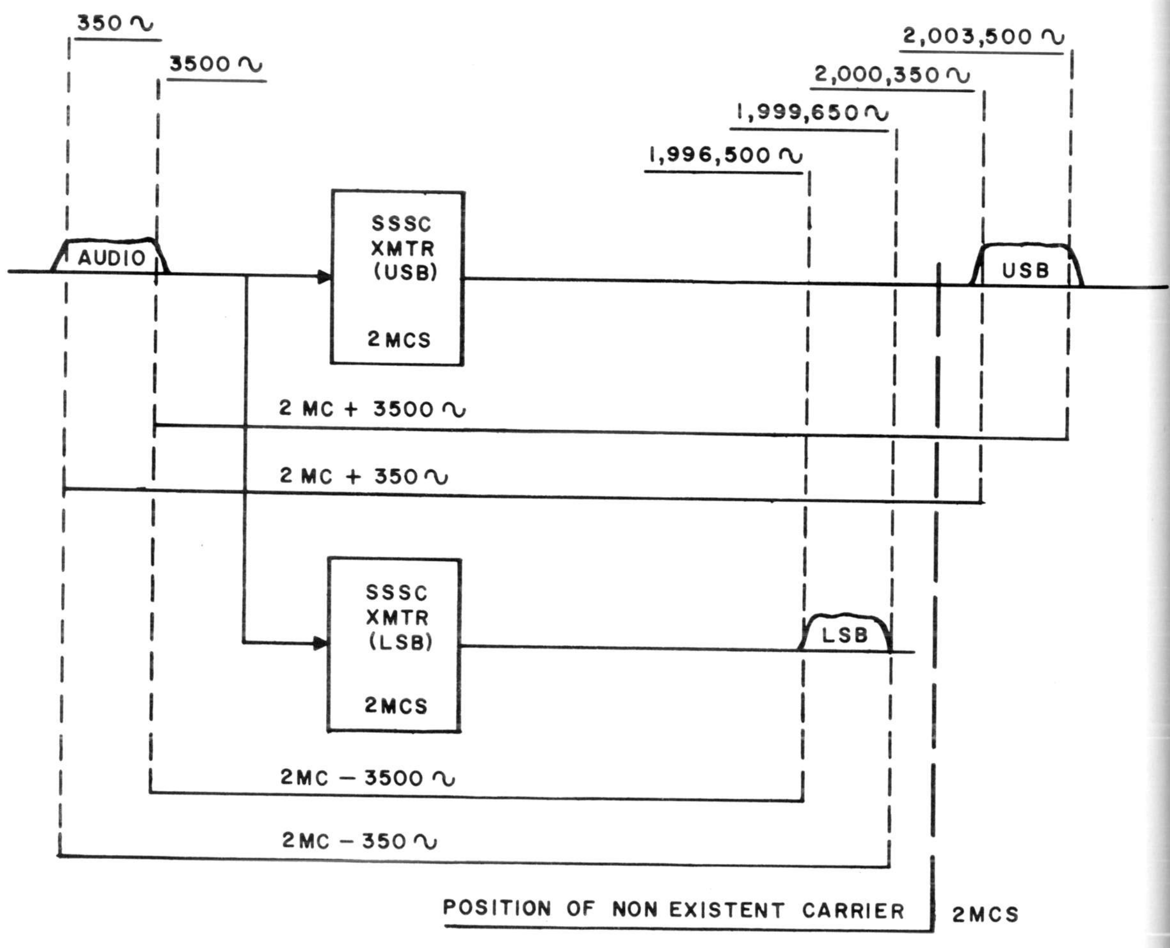


Figure 3-1.

RF frequency of 2 mc is shown. The heavy dotted line between the sidebands is the non existent carrier.

A single sideband receiver accepts the transmitted upper or lower sideband signal. A local "carrier" simulating the original carrier is re-inserted and the original audio spectrum is recovered. Figure 3-2, below, shows an elementary single sideband receiver.

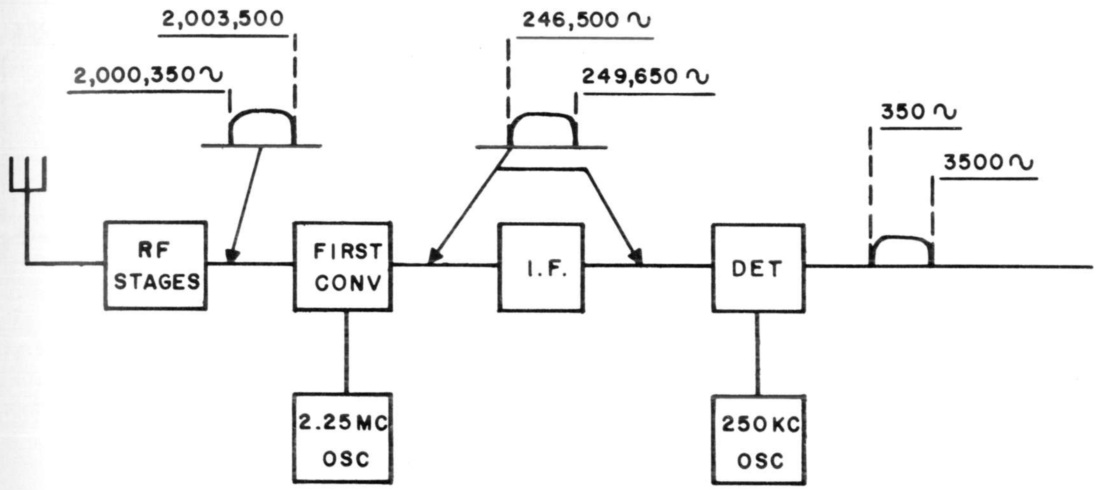


Figure 3-2.

The simple receiver shown in Figure 3-2 is configured to receive the upper sideband signal transmitted in Figure 3-1. The receiver is a superheterodyne, with an intermediate frequency of 250 KCS. The local oscillator operates 250 KC above the incoming signal frequency. The output of the first converter contains, as usual, the two original input frequencies, their sum and their difference.

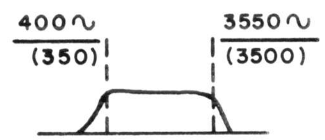
The output circuit of the converter will be tuned to the region of 250 KCS. This means that only the difference output will be passed. Note that the difference output is now a lower sideband; this is of no significance in this particular receiver, because the detector will not know the difference. It is a point to remember, however, for it will assume great importance in the discussions of practical receivers which will follow.

The I.F. signal is passed to the detector, which also receives an injection frequency of 250 KCS. The difference output of the detector circuit is the original audio spectrum.

3-3 Stability Requirements of a SSB System

Suppose that the 250 KC oscillator in our elementary receiver of Figure 3-2 had an error of plus 50 cycles. Its frequency, then, would be 250.050 KCS. The resultant output of the detector, then, would be:

250,050	250,050
− 249,650	246,500
400 ∿	3 550 ∿



The original audio spectrum has been severely distorted. Such distortion with multiplexed narrow band teletype signals would be disastrous. Suppose that the transmitter oscillator and the receiver local oscillator also had appreciable frequency error; the resultant audio spectrum might well be useless.

It has been found that, as far as voice signals are concerned, "naturalness" disappears with an error of about 50 cycles, even though the message may be understandable. The signal becomes unintelligible at an error of about 100 cycles. Certain types of intelligence require almost absolute stability.

IT IS IMPERATIVE, THEN, THAT ALL FREQUENCY GENERATING CIRCUITS OF A SSB SYSTEM BE DESIGNED FOR EXTREME STABILITY.

So important is the attribute of stability that frequency generating circuits are often controlled by central frequency standards. TMC synthesized systems currently employ a 1 mc standard with an accuracy of 1 part in 100,000,000 per day.

3-4 Generation of SSB Signals

As stated previously, a SSB transmitter translates an audio intelligence spectrum to a desired radio frequency. The initial generation of SSB signals in commercial and military equipments is usually carried out at a relatively low frequency. In current TMC equipment, the initial generation is performed:

a) at 17 KC in one Model of the SBE

b) at 250 KC in another model of the Model SBE and

c) at 250 KC in the SBG exciters

Frequencies of 100 KC and 455 KC are common in equipments produced by other commercial manufacturers. Such frequencies are often referred to as "sub carrier" frequencies. At these frequencies it is less difficult to construct the critical filters required than at the higher frequencies.

There are two general methods of generating SSB signals:

a) the Filter method.

b) the Phase Shift method. (in this method, the initial generation of SSB signals may be accomplished at higher frequencies than in the filter method).

The Technical Materiel Corporation uses the Filter method exclusively. Each method has its advantages and disadvantages; each will be discussed in detail in subsequent paragraphs.

3-5 The Balanced Modulator

Both the Filter and the Phase Shift methods of generating single sideband signals employ an electronic circuit called a "Balanced Modulator". There are many variations of this circuit but all perform essentially the

same functions. The balanced modulator mixes two signals, passes the sum and difference frequencies, and attenuates the two original frequencies.

The two original frequencies are the "carrier" and the "modulation" frequencies. The modulation frequencies are lost because of the bandpass characteristics of the circuits; the carrier is balanced out electronically. Figure 3-3, below, shows a typical balanced modulator circuit.

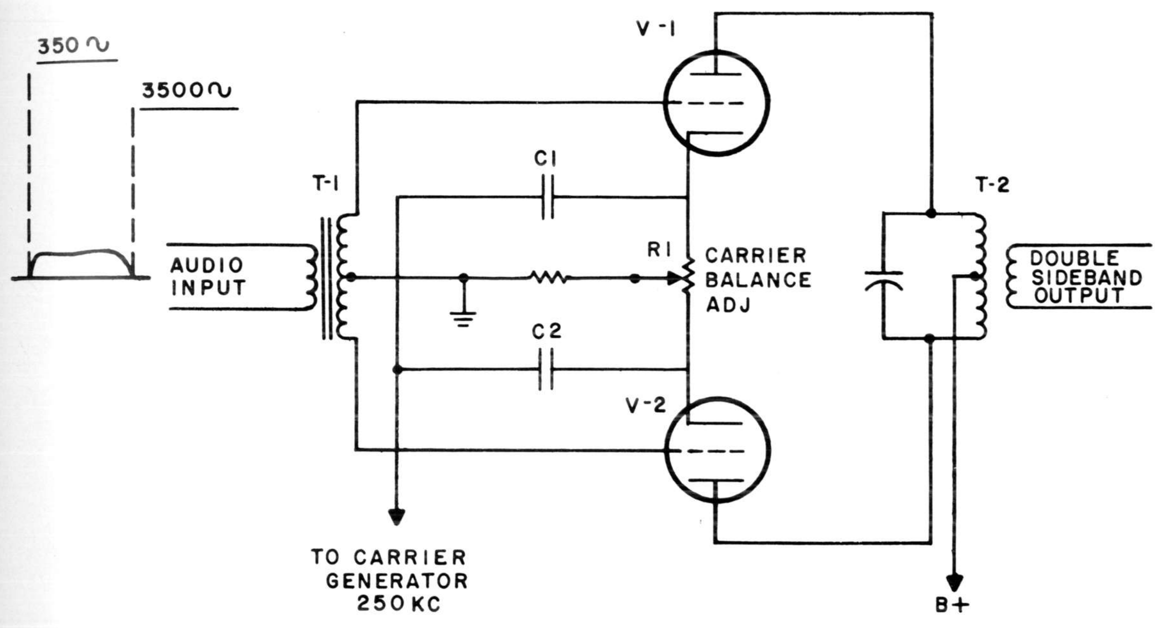


Figure 3-3.

The carrier frequency, 250 KC, is applied to the cathodes of V-1 and V-2 in parallel. Thus, as the carrier voltage swings alternately positive and negative, plate current in both tubes changes by the same amount, and in the same direction. The plate currents at the carrier frequency, then, are 180 degrees out of phase in the primary of tuned transformer T-2; no carrier appears in the secondary.

In the simplified sketch of figure 3-4, the carrier voltage is shown instantaneously negative. Carrier plate current in each tube is increasing by the same amount. The resultant carrier CEMF's in the primary of T-2 are shown in opposition.

For optimum balancing out of the carrier, tubes and components must be matched. Final balance is achieved by a careful adjustment of R-1, the carrier balance adjust potentiometer. Circuits of this type can be expected to suppress the carrier by 20 to 30 db. Since a total suppression in a transmitter system of about 50 db is desired, further suppression must be accomplished in other circuits.

The audio input is applied to the control grids of V-1 and V-2 in a push pull arrangement. Thus, as the audio plate current of V-1 increases, the audio plate current of V-2 decreases; the CEMF's of these components add in the primary of T-2. The audio is lost, however, because T-2 is tuned to the region of 250 KC to accept the sidebands.

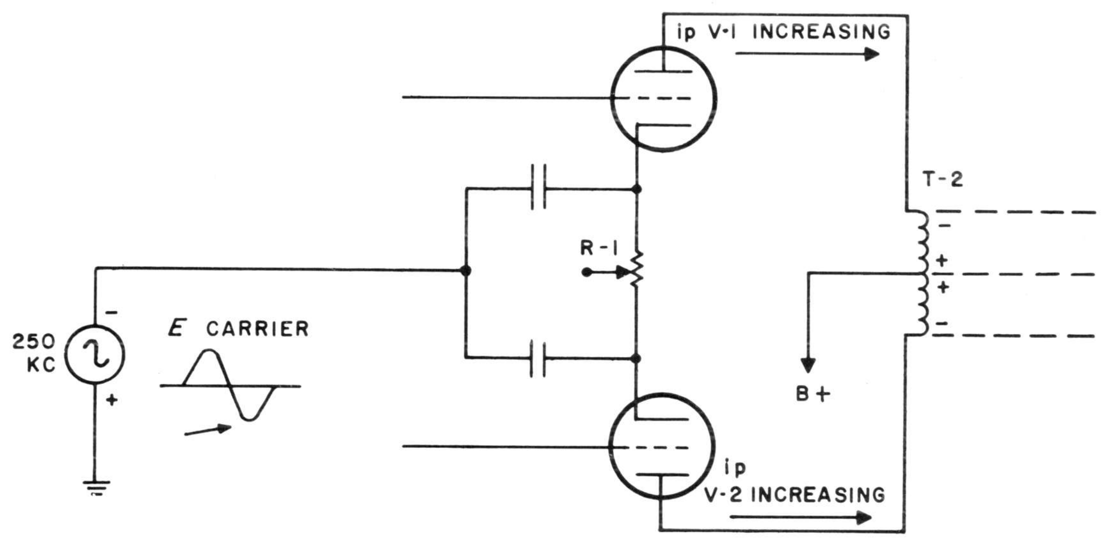


Figure 3-4.

The usual mixing action of the audio and carrier frequencies produces sidebands, that is, sum and difference frequencies in the region of 250 KCS. The width of the output spectrum will depend on the input audio spectrum and the bandpass characteristics of T-2.

Thus, if an audio spectrum from 350 to 3500 cycles is applied to the balanced modulator designed for a "carrier" injection of 250 KCS, the output will be as shown below:

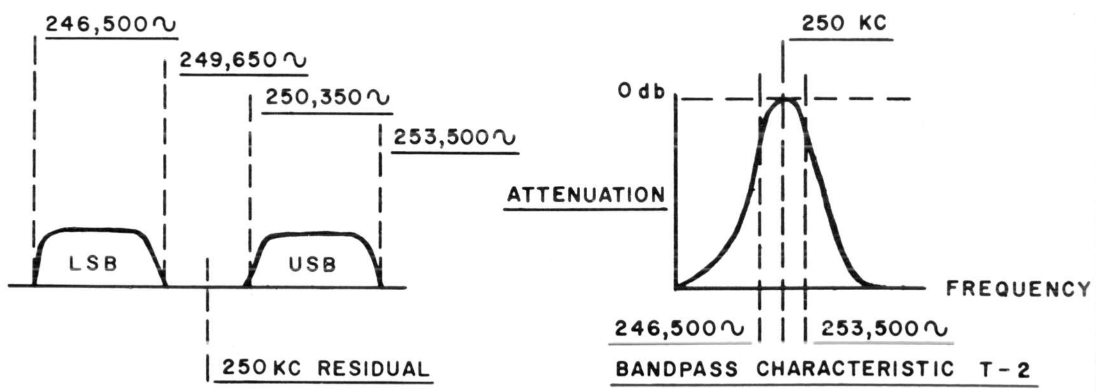


Figure 3-5.

3-6 The Filter Method of Sideband Generation

The Filter method of SSB generation employs, essentially, a balanced modulator and a selective filter.

Consider the output of the balanced modulator of Figure 3-5; this is a double sideband output. Each sideband contains the same intelligence. Since it is our intention to transmit a true SSSC signal, the unwanted sideband must be eliminated. This is accomplished, in the low power level stages of the transmitter, for reason of convenience and economy. The vehicle of elimination will be a selective filter.

The selective sideband filter will be configured to pass only the desired sideband. This filter will also attenuate the residual "carrier" output of the balanced modulator. A perfect filter would have zero insertion loss, straight vertical skirts, and a flat top, that is, equal response for all frequencies in the passband. The sketch below shows the perfect, though unattainable, characteristic of a filter designed to pass the USB in our system.

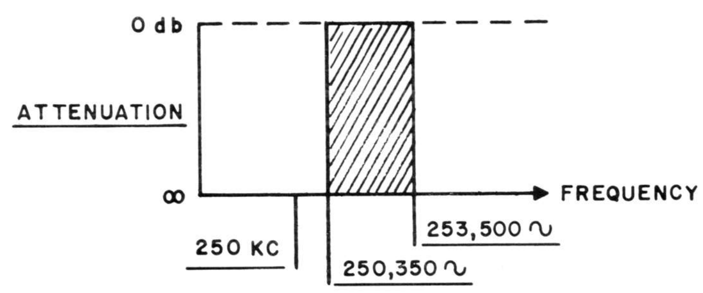


Figure 3-6

3-7 The Phase Shift Method of Sideband Generation

The phase shift method does not require expensive critical selective sideband filters; therefore, the initial sideband generation may take place at a higher frequency than with the Filter method.

Figure 3-7 shows the block diagram of a phase shift system designed to pass the lower sideband (LSB).

The explanation of the system will be carried out by means of quasi-vector algebra. Notice that the system is composed of four phase shift networks, two balanced modulators, and a resistive combining network. This is only one of many possible phase shift configurations.

An audio signal of 1 KC at a phase angle of 0 degrees is applied to two phase shift networks. One network shifts the phase of the audio signal by plus 45 degrees, the other shifts it by minus 45 degrees.

The sub carrier signal of 250 KC, at angle 0 degrees, is applied to two phase shift networks. One network shifts the phase of the 250 KC signal by plus 45 degrees; the other shifts it by minus 45 degrees.

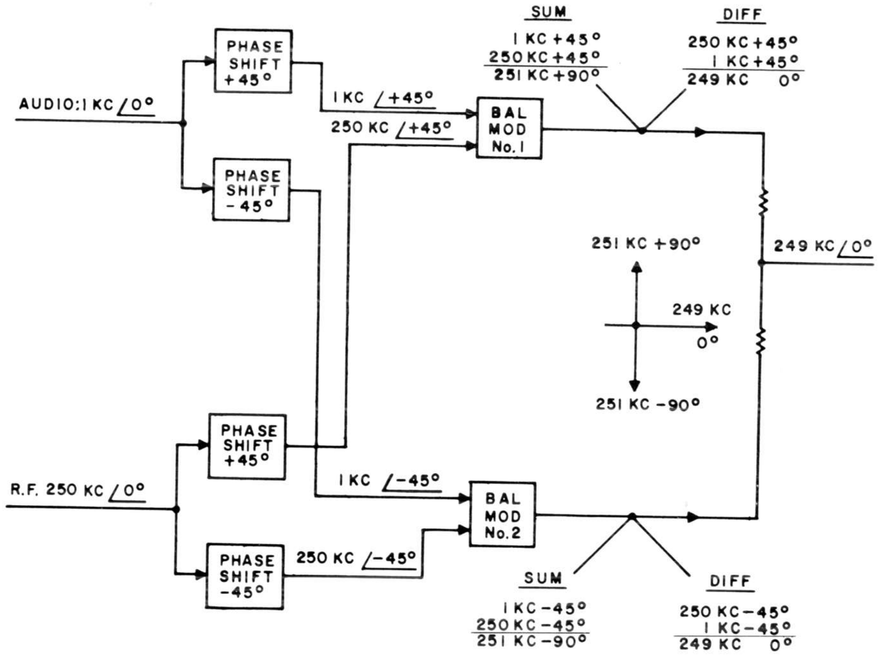


Figure 3-7.

Balanced modulator #1 receives:

a) a 1 KC signal at plus 45 degrees.
b) a 250 KC signal at plus 45 degrees.

The balanced modulator passes the sum and difference frequencies:

a) 251 KC at plus 90 degrees.
b) 249 KC at 0 degrees.

These sum and difference frequencies are fed to the resistive combining network.

Balanced modulator #2 receives:

a) a 1 KC signal at minus 45 degrees.
b) a 250 KC signal at minus 45 degrees.

The balanced modulator passes the sum and difference frequencies:

a) 251 KC at minus 90 degrees.
b) 249 KC at 0 degrees.

These sum and difference frequencies are fed to the resistive combining network.

The signals at 251 KC are 180 degrees out of phase, and add vectorally to zero.

The signals at 249 KC are in phase, and re-inforce each other.

3-8 Translation to the Operating Frequency

We have thus far succeeded in generating the single sideband suppressed carrier at a relatively low frequency. It now remains to translate the SSSC signal to the high frequency desired for transmission. There are many systems and methods for accomplishing this. The system used will depend on the type of service for which the transmitter has been designed. For example:

a) radio amateur might operate on one or more "ham bands".
b) a radiotelephone transmitter might operate on a single frequency.
c) a military system might require continuous coverage over the range 2 mc to 30 mcs.

Obviously, the greater the coverage, the more complex the arrangements for translation of the sideband. Since TMC transmitters are generally designed for continuous coverage from 2 to 28 mcs, we will base our translations on this requirement.

Figure 3-8 illustrates one scheme to accomplish the translation.

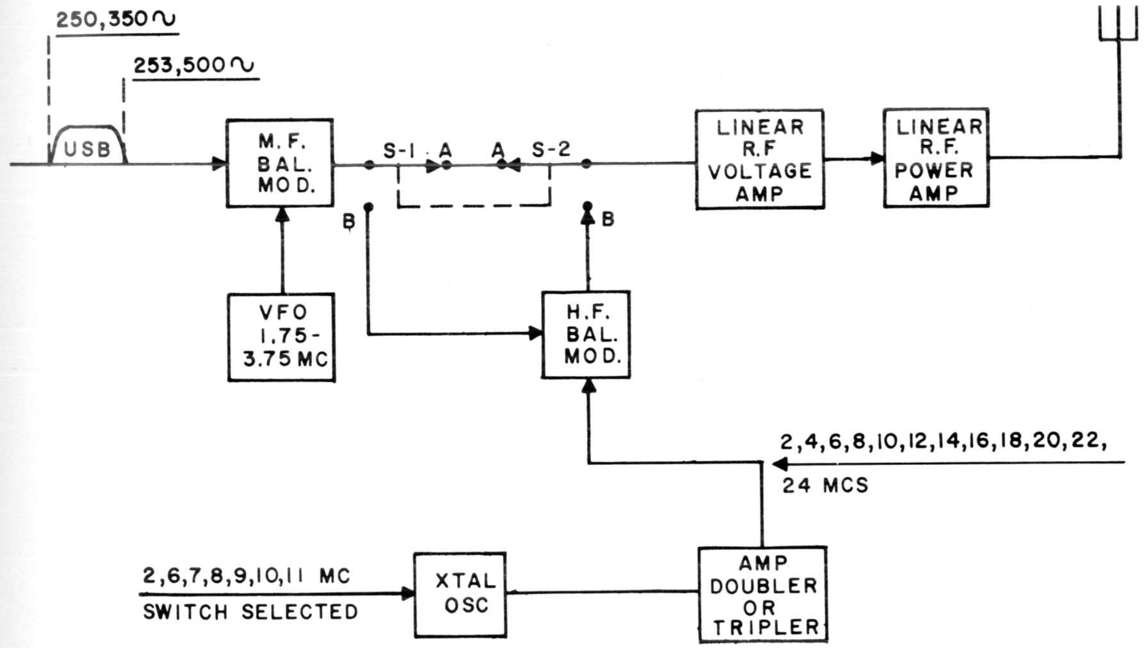


Figure 3-8.

Figure 3-8 is a simplified block diagram of a system designed to receive a SSSC signal, USB, at 250 KC, and to translate the signal to any RF frequency in the range 2 to 28 mcs. No attempt has been made to show the switching, tuning, frequency control, or tuned circuit arrangements. We will concern ourselves now only with the translation itself.

The SSSC signal, an upper sideband, is presented to a "medium Frequency" balanced modulator. The second input to this balanced modulator is a "sub carrier" frequency, in the range 1.75 mcs to 3.75 mcs. This is provided by a stable, tunable VFO.

The output circuits of the medium frequency balanced modulator are designed to pass only the SUM frequencies from the balanced modulator. Thus, no sideband inversion takes place in this circuit.

When the VFO feeding the MF balanced modulator is at 1.75 mc, the sideband will be translated to 2.0 mcs. When the VFO is at 3.75 mcs, the sideband will be translated to 4.0 mcs. This is shown in the simplified sketches below:

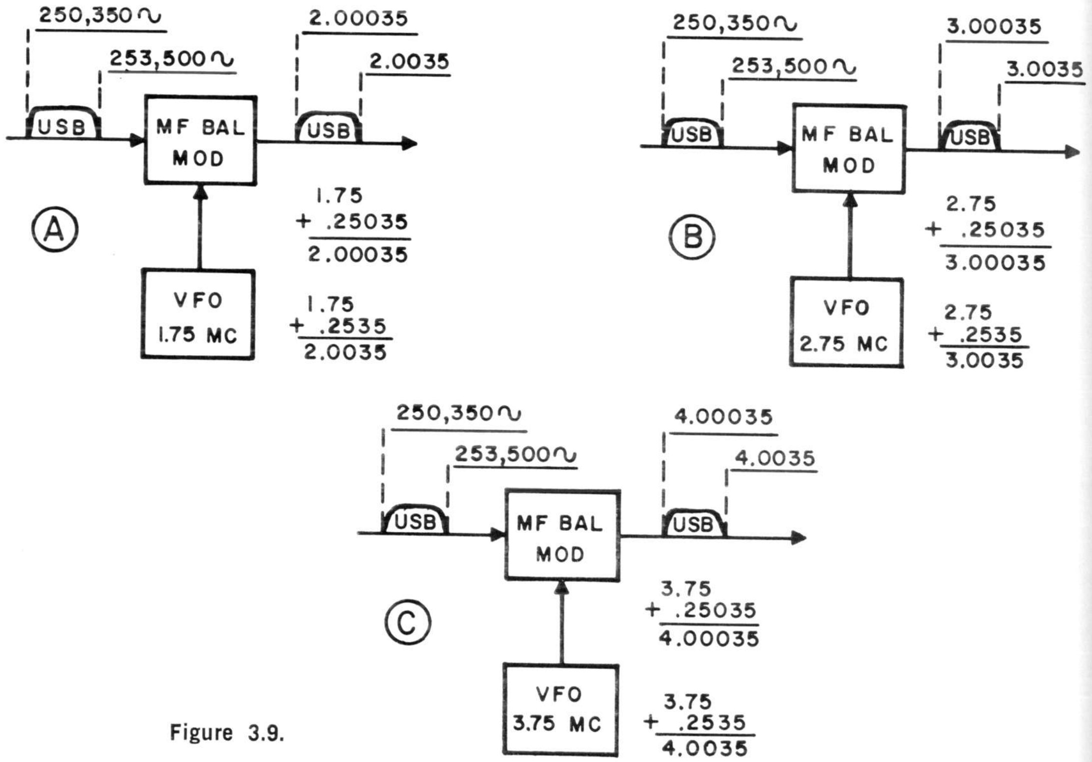


Figure 3.9.

The output of the medium frequency circuits, then, is an USB, always in the range from 2 to 4 mcs.

The output frequency range of the transmitter is 2 to 28 mcs. This range will be divided into 13 bands, each band covering a 2 mc range, as follows:

BAND	FREQUENCY RANGE
1	2 — 4 mcs
2	4 — 6 mcs
3	6 — 8 mcs
4	8 — 10 mcs
5	10 — 12 mcs
6	12 — 14 mcs
7	14 — 16 mcs
8	16 — 18 mcs
9	18 — 20 mcs
10	20 — 22 mcs
11	22 — 24 mcs
12	24 — 26 mcs
13	26 — 28 mcs

On Band #1 range, switches S-1 and S-2 are shown at "A" in Figure 3-8. The upper sideband, in the range 2 — 4 mcs, from the MF circuits, is applied to a chain of linear RF voltage amplifiers and a final power amplifier. From this point the signal is applied to the antenna.

On all other bands, #2 through #13, switches S-1 and S-2 are thrown to position "B". (Figure 3-8)

Under these conditions, the output of the MF circuits is applied to a high frequency (HF) balanced modulator. The second input to the HF balanced modulator is an injection frequency, derived from a crystal oscillator-amplifier-doubler-tripler arrangement.

Selector switches will cause the crystal oscillator circuit to operate at one of 7 possible frequencies. The same selector switches will cause the amplifier-doubler-tripler circuit to feed the desired frequency to the HF balanced modulator.

See Table 3-1 below.

TABLE 3.1

Band	*Frequency Range*	*Crystal Osc.*	*Amp - Doubler - Tripler*
1	2 — 4 mcs	Not Applicable	Not Applicable
2	4 — 6 mcs	2 mcs	2 mcs
3	6 — 8 mcs	2 mcs	4 mcs
4	8 — 10 mcs	6 mcs	6 mcs
5	10 — 12 mcs	8 mcs	8 mcs
6	12 — 14 mcs	10 mcs	10 mcs
7	14 — 16 mcs	6 mcs	12 mcs
8	16 — 18 mcs	7 mcs	14 mcs
9	18 — 20 mcs	8 mcs	16 mcs
10	20 — 22 mcs	9 mcs	18 mcs
11	22 — 24 mcs	10 mcs	20 mcs
12	24 — 26 mcs	11 mcs	22 mcs
13	26 — 28 mcs	8 mcs	24 mcs

The crystal oscillator can thus be operated at 7 reasonably low frequencies, to produce 12 required injection frequencies.

The HF balanced modulator output circuits are tuned to the SUM frequencies; the upper sideband is again passed without inversion.

The output of the HF balanced modulator, an upper sideband at the assigned RF frequency, is passed to the chain of linear voltage amplifiers and the final power amplifier, for eventual transfer to the antenna.

Examples of Operation:

Note: In SSB operation the term "carrier frequency" refers to that position in the spectrum reserved for the carrier. Later a definite distinction will be made between the "carrier" frequency and the "assigned" frequency.

Example #1:

Given: assigned frequency is 5.25 mcs, USB.

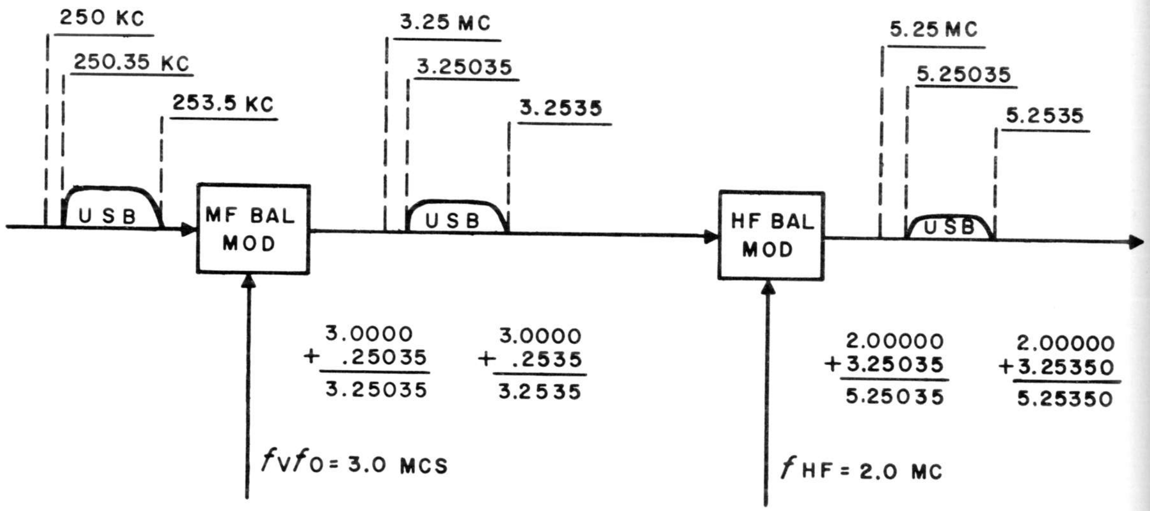
From Table 3-1:

f hf injection = 2 mcs

f mf = f assigned − f hf injection

$$f \text{ mf} = \begin{array}{r} 5.25 \\ -\ 2.00 \\ \hline 3.25 \text{ mcs} \end{array}$$

$$f \text{ vfo} = f \text{ mf} - 250 \text{ kc} = \begin{array}{r} 3.25 \\ -\ .25 \\ \hline 3.00 \text{ mcs} \end{array}$$



Example #2:

Given: assigned frequency is 17 mcs, USB.

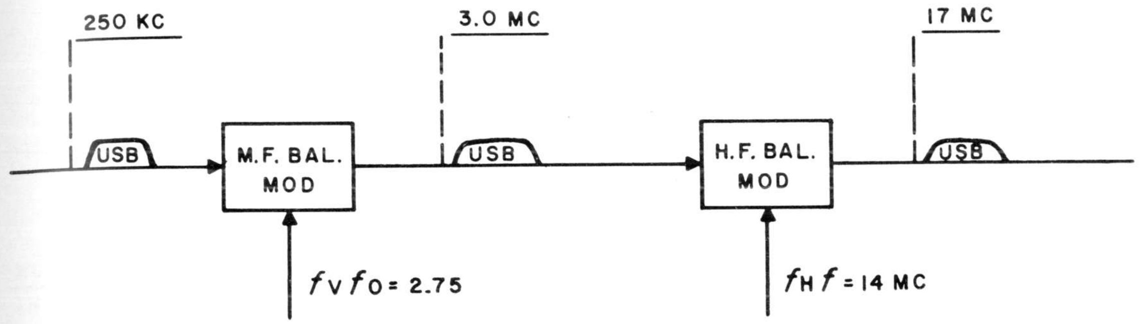
From Table 3-1:

f hf injection = 14 mcs

f mf = f assigned − f hf injection

$$f \text{ mf} = \begin{array}{r} 17.0 \\ -\ 14.0 \\ \hline 3.0 \text{ mc} \end{array}$$

$$f \text{ vfo} = f \text{ mf} - 250 \text{ kc} = \begin{array}{r} 3.0 \\ -\ .25 \\ \hline 2.75 \text{ mcs} \end{array}$$



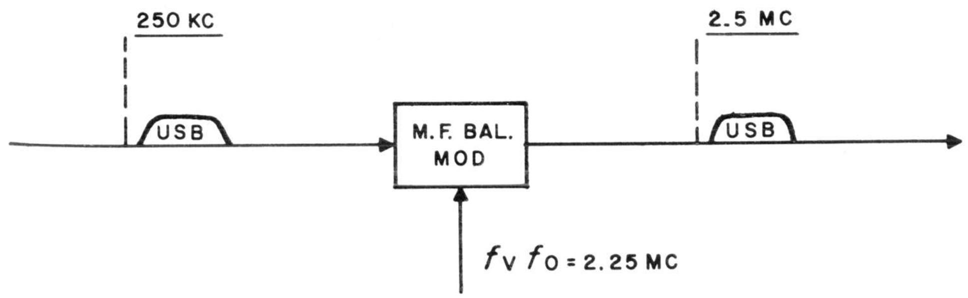
Example #3:

Given: assigned frequency is 2.5 mcs

From Table 3-1: the HF balanced modulator is bypassed.

f assigned = f mf

$$f \text{ vfo} = f \text{ mf} - 250 \text{ kc} = \begin{array}{r} 2.500 \\ -\ .250 \\ \hline 2.250 \text{ mcs} \end{array}$$



CHAPTER 4

TYPES OF SINGLE SIDEBAND OPERATION

4-1 Introductory Note

Thus far, we have considered only Single Sideband Suppressed Carrier, (SSSC) operation. Chapter Four will introduce modes of operation, which, while not true SSSC methods, must be classified under the heading of Single Sideband, (SSB) operation. In addition, new terms will be introduced and defined.

4-2 Single Sideband Reduced Carrier Operation

It was established in Chapter Three that SSSC systems require stabilities of the highest order. Unless the injection frequency in the detector of the SSSC receiver bears a precise relationship to the received SSSC signals, the audio output spectrum is distorted. Thus, at any instant, the relationship of the re-inserted carrier to the received sideband frequencies must be the same as the relationship of the transmitter carrier to the transmitted sideband frequencies. This imposes the most stringent frequency stability requirements.

In the SSB Reduced Carrier mode of operation, one sideband is transmitted as before; in addition a low level carrier, suppressed approximately

10 to 20 db below the peak sideband power, is transmitted. Such a carrier is termed a "PILOT CARRIER".

In the receiver, the carrier is separated from the sideband, amplified, and re-inserted in the detector. Thus, the re-inserted carrier is identical, for all practical purposes, to the transmitted carrier. A simplified illustration of such a scheme is presented in Figure 4-1 and 4-2.

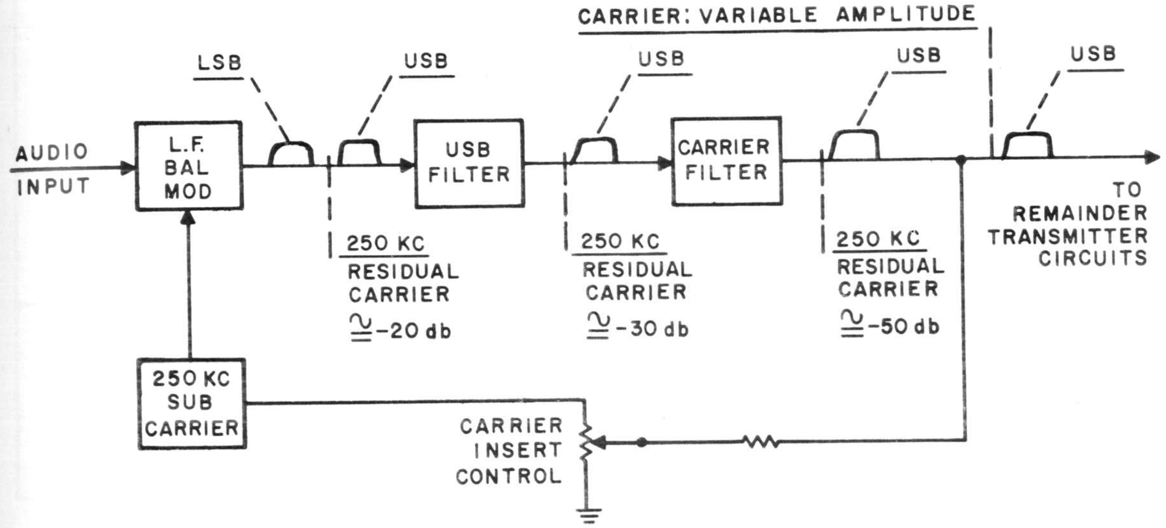


Figure 4-1.

Figure 4-1 shows a method of re-inserting the carrier in the transmitted signal if desired. A low frequency balanced modulator receives an audio input and a 250 KC injection frequency. The output of the balanced modulator is a double sideband signal, with the carrier suppressed approximately 20 db. The USB filter attenuates the LSB and passes the USB; the carrier is further attenuated here. The USB filter is followed by a carrier filter. This filter has a decided notch at 250 KC, to attenuate the 250 KC carrier to the maximum possible degree, (approximately 50 db).

At this point, the output of a carrier re-insert circuit is connected. The control potentiometer can be adjusted for any degree of carrier suppression, from minimum to maximum. The carrier control is usually a front panel control, calibrated in db or percent of carrier suppression. We now have a transmitter capable of SSSC operation, or SSB, reduced carrier operation.

Figure 4-2 shows an USB signal, with pilot carrier, at a frequency of 5.0 mcs, emanating from an SSB transmitter. The signal is picked up by the receiving antenna, and is converted to an I.F. frequency of 250 KCS.

The I.F. contains both the pilot carrier and the sideband frequency. The sideband has been inverted, but, in this particular receiver, this is of no significance.

Filter Z-2 is a bandpass filter, which will pass only the sideband frequencies. These are amplified in the I.F. amplifier, the output of which is applied to the detector.

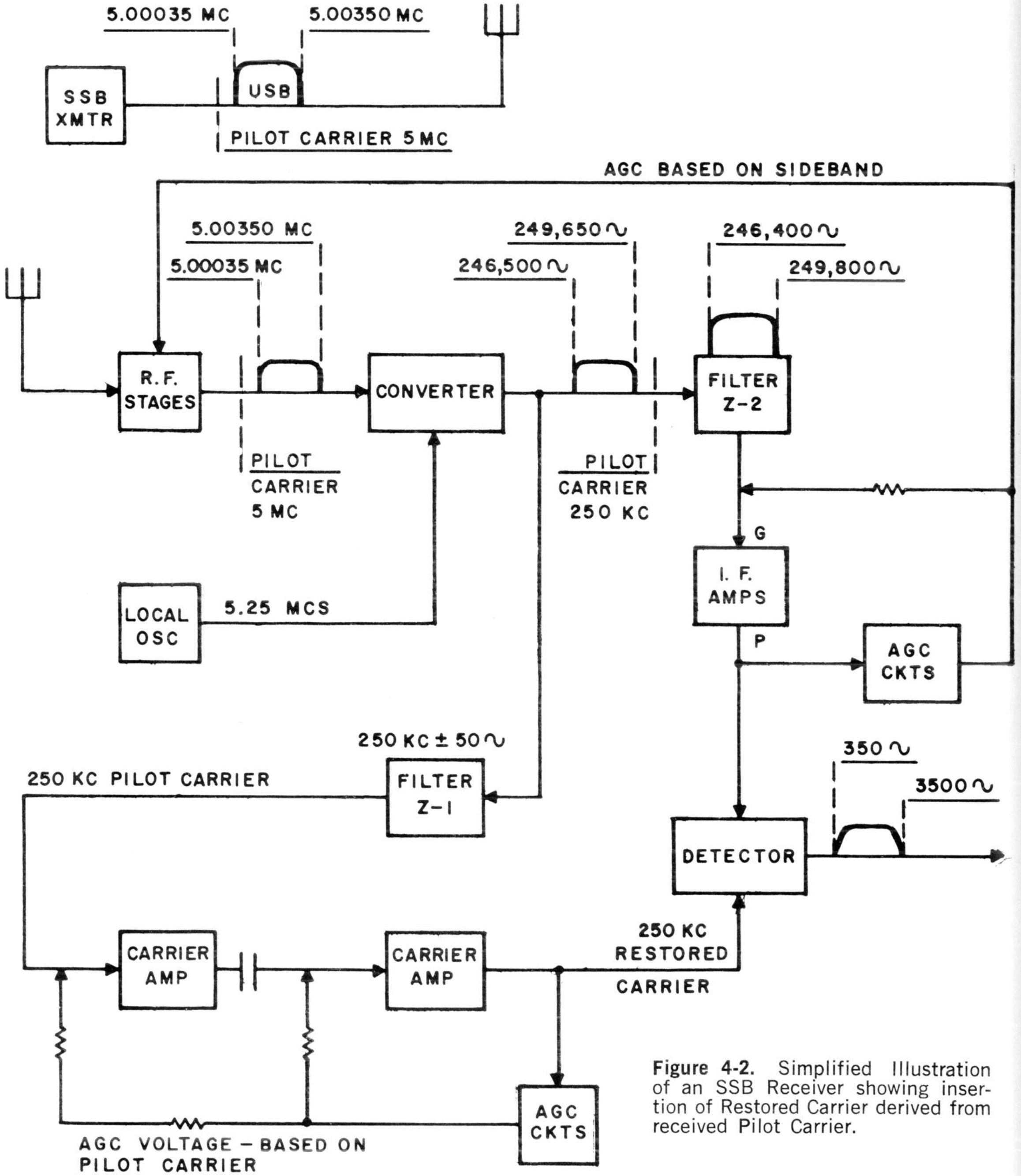


Figure 4-2. Simplified Illustration of an SSB Receiver showing insertion of Restored Carrier derived from received Pilot Carrier.

An AGC voltage, based on the sidebands, controls the I.F. amplifier and the R.F. stages.

Z-1 is a narrow band filter, which will pass only those frequencies in the immediate vicinity of 250 KC. Filter Z-1, then, passes the pilot carrier stripped of the sidebands.

The pilot carrier is amplified by two high gain cascade stages of amplification. The output of these stages, called the RESTORED CARRIER, is also fed to the detector. An AGC voltage, based on the pilot carrier, controls the carrier voltage amplifiers.

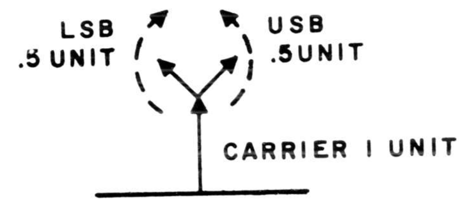
The detector output is the difference between the sideband input and the restored carrier input. Thus, the original audio spectrum is recovered.

The two AGC systems allow for independent fading of the carrier or sidebands.

The RESTORED CARRIER may also be called: EXALTED CARRIER, ENHANCED CARRIER, or RECONDITIONED CARRIER.

4-3 Single Sideband, Full Carrier Operation

A true SSSC transmitter is not compatible with a conventional AM receiver. A conventional AM detector requires that the amplitude and phase of the carrier bear a definite relationship to the sidebands. In a 100% modulated AM signal the sideband vectors "ride" the carrier vector, rotating in opposite directions. The carrier amplitude, Eo, is twice the amplitude of each sideband.



If one sideband is lost, the AM detector will reproduce the original audio without appreciable distortion. A SSB reduced carrier signal is not compatible with an AM detector unless the principle of EXALTED CARRIER is used.

The simplest compatible system employs SSB full carrier. Only one sideband is transmitted, together with full carrier. The carrier is transmitted at from 4 to 6 db below the peak output power of the transmitter. This type of operation is known as COMPATIBLE SSB (CSSB), or COMPATIBLE AM.

The transmitter of Figure 4-1 can be used to transmit SSB full carrier. The signal may be received on a conventional AM set.

4-4 Independent Sideband Operation (ISB)

Amplitude modulated waves have distinct characteristics; two of these are:

a) the total bandwidth required equals twice the highest modulating frequency.

b) the sidebands are mirror images; each contains identical information.

If a voice spectrum in the range 300 - 3000 cycles is transmitted in the AM mode, the total bandwidth required is approximately 6 KC.

A SSB transmitter will transmit the same audio spectrum and require half the AM bandwidth, or approximately 3 KC.

The tendency in military communication systems is to transmit as much intelligence simultaneously as possible. By means of multiplexing a single sideband can be made, for example, to carry 16 different channels of teletype information. New techniques will increase this capability. If one sideband carries 16 channels, then two sidebands will carry 32 channels.

Independent Sideband (ISB) operation makes use of both sidebands, with different information on each sideband. This may be accomplished with complete carrier suppression, or with any degree of carrier insertion desired. This type of operation makes maximum use of the available spectrum space, but less power is available per channel than with true SSB operation.

Figure 4-3 illustrates a method of generating ISB with variable carrier re-insertion.

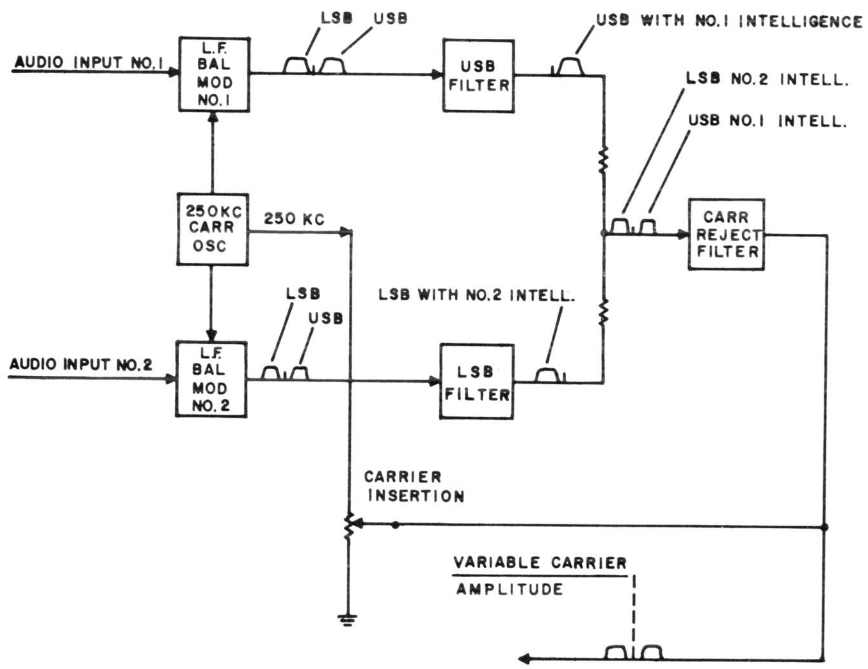


Figure 4-3.

Two balanced modulators are employed. Each receives:

a) audio signals from different sources.

b) a sub carrier (250 KC in this case) from a local, stabilized oscillator.

The output of both balanced modulators is a double sideband signal, with carrier suppressed.

The output of balanced modulator #1 is applied to an USB filter, which passes the USB containing intelligence from audio source #1.

The output of balanced modulator #2 is applied to a LSB filter, which passes the LSB containing intelligence from audio source #2.

The outputs are combined; the signal is now an ISB signal; that is, a double sideband signal with independent information on each sideband.

A carrier filter is incorporated at this point to suppress the residual carrier to the maximum extent possible.

The 250 KC output of the local stabilized oscillator is applied to a carrier re-insert potentiometer. The wiper of this control determines the degree of carrier insertion in the output Independent Sideband (ISB) signal.

The output signal is applied to frequency translation circuits and linear amplifiers, as with any SSB transmitter.

The arrangement shown provides a certain amount of versatility, in that:

a) with no #1 audio input, a LSB may be transmitted with any degree of carrier.

b) with no #2 audio input, an USB may be transmitted with any degree of carrier.

c) with different audio signals at each input, an ISB signal may be transmitted with any degree of carrier.

Thus, the system is capable of:

a) SSSC, USB operation.

b) SSSC, LSB operation.

c) SSB, reduced carrier, either sideband operation.

d) SSB, full carrier (Compatible AM) operation.

e) ISB, with variable carrier suppression.

The receiver is a double conversion superheterodyne, with IF frequencies of 1.75 mcs and 250 KCS. Assuming that the receiver operates in the range 2 to 32 mcs, the first local oscillator will operate in the range 3.75-33.73 mcs; that is, 1.75 mcs above the incoming signal. See Figure 4-4.

Assume that an USB signal, at a carrier frequency of 5 mcs, containing an audio spectrum, is being received. The output of the first converter is an inverted sideband signal, as shown in figure 4-4A.

This signal is applied to an IF amplifier, and then to the second converter. The second converter injection frequency is 2 mcs, derived from a stabilized oscillator. The sideband is again inverted, and is now displaced in its normal place in the spectrum, as shown in figure 4-4B.

This signal, with or without carrier, is applied:

a) to a carrier filter, which passes a very narrow band of frequencies in the immediate vicinity of 250 KC.

b) to two filters: one, a LSB filter and the other an USB filter.

The Sideband Selector switch is placed in position #2, to pass the received USB signal. With the switch in this position, the "B" section is "blank".

The USB is processed in IF amplifier "A", the output of which is applied to Detector "A". The second input to detector "A" is a 250 KC injection frequency, derived either from an internal stabilized oscillator, or from the reconstructed pilot carrier, if one is being transmitted. The selection is made by the "RCC — INT OSC" switch.

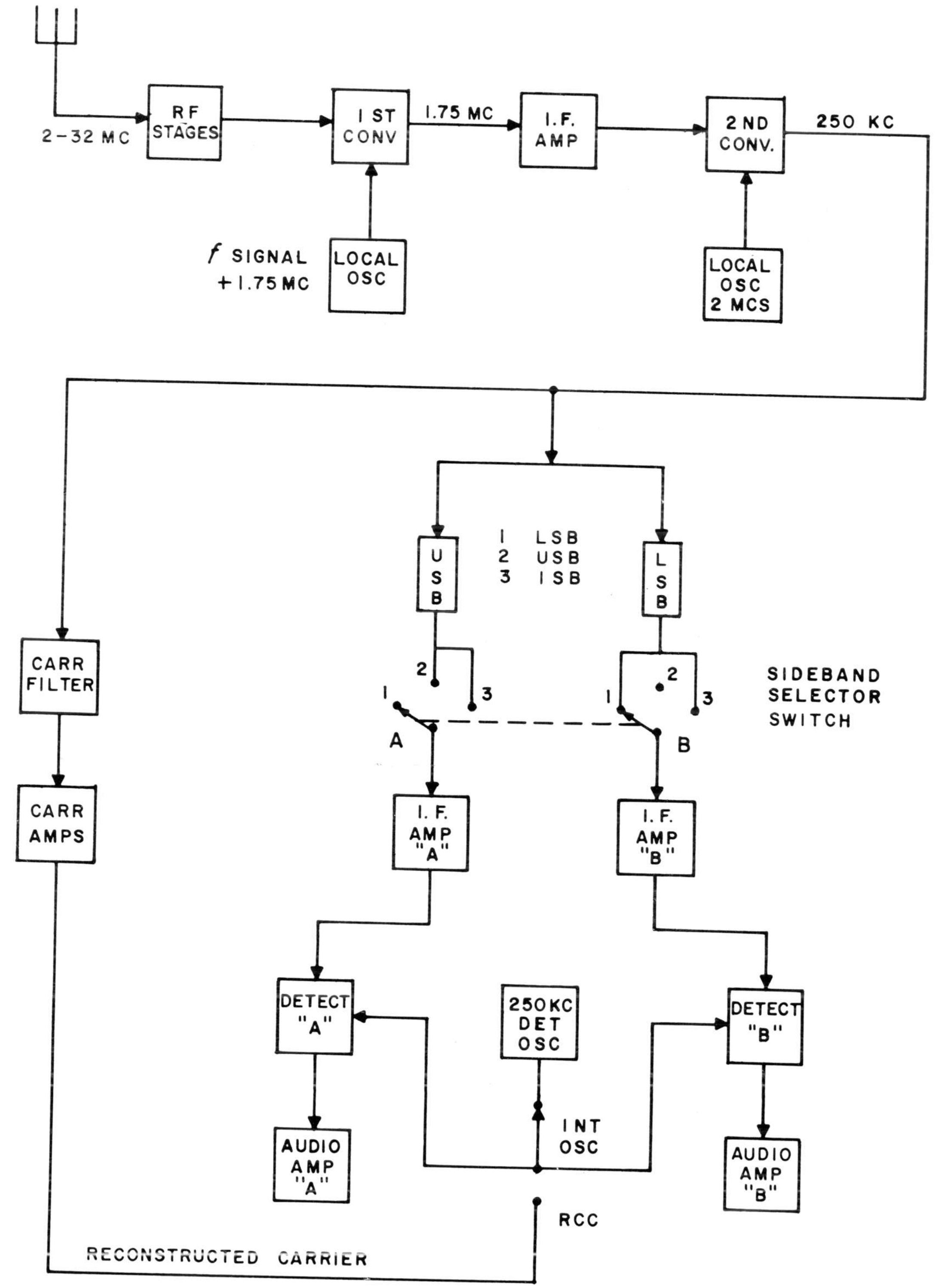


Figure 4-4. Simplified Block Diagram of a Receiver Compatible with the Transmitter of Figure 4-3.

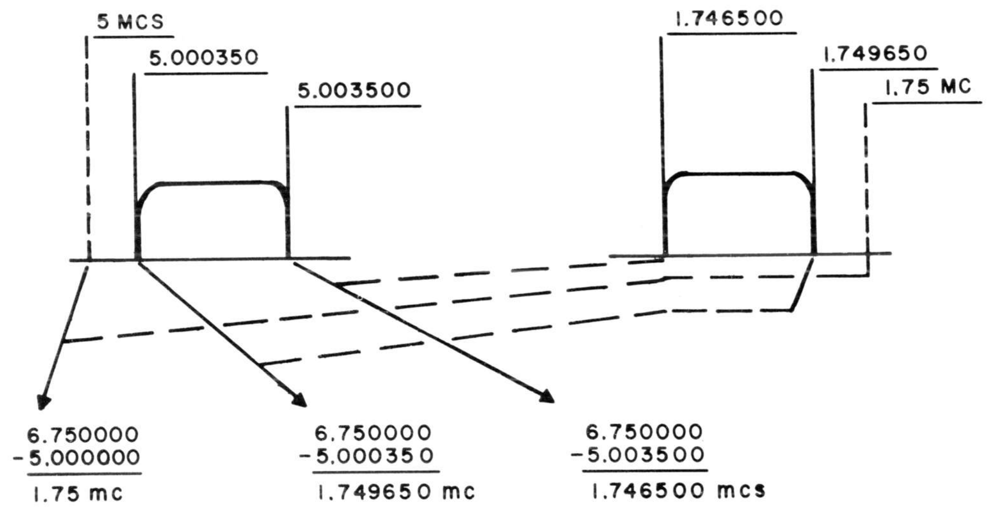


Figure 4-4A.

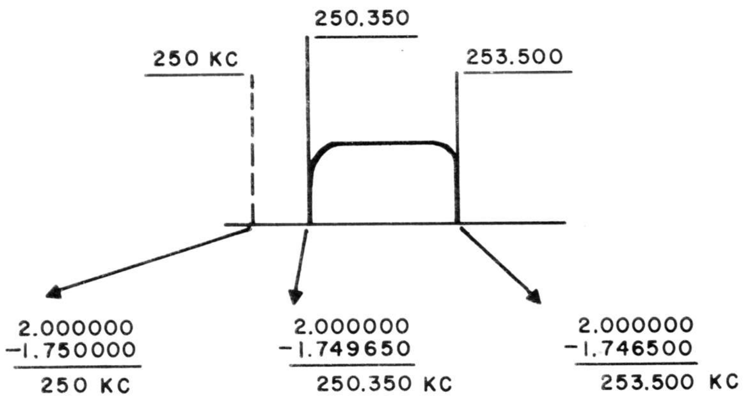


Figure 4-4B.

The output of detector "A" is applied to audio amplifier "A".

With the sideband selector switch in position #1, as shown, the LSB is being selected. In this case, IF amplifier "B", detector "B" and audio amplifier "B" would be processing the signal.

With the sideband selector switch in position #3, both sideband filters, both IF amplifiers, detectors and audio amplifiers process the separate intelligence. This corresponds to ISB operation.

4-5 ISB Operation Utilizing Multiplexers

Figure 4-5 shows a simplified scheme for imposing four separate intelligence channels on two sidebands in ISB operation. The frequencies selected are not necessarily representative, and have been selected for illustration only.

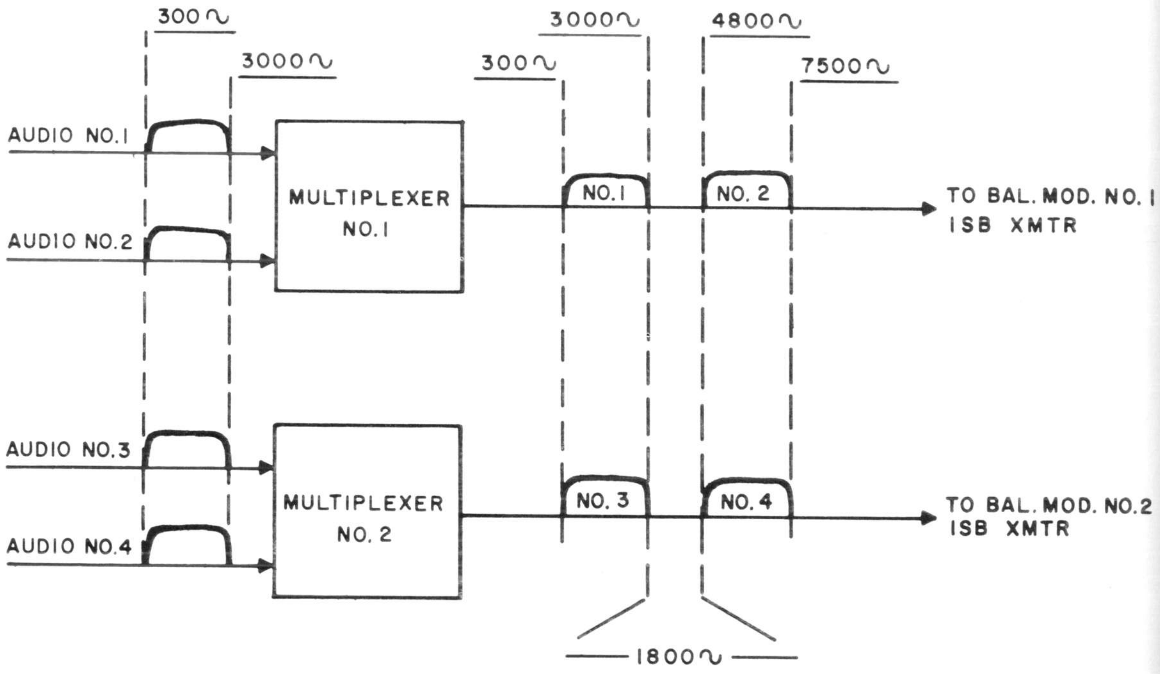
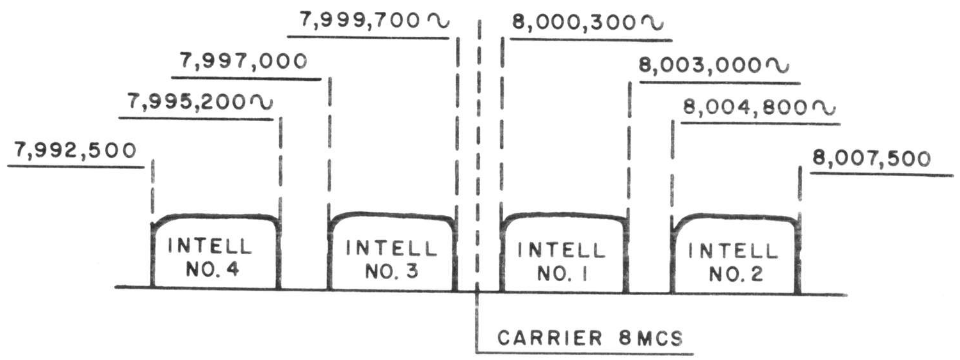


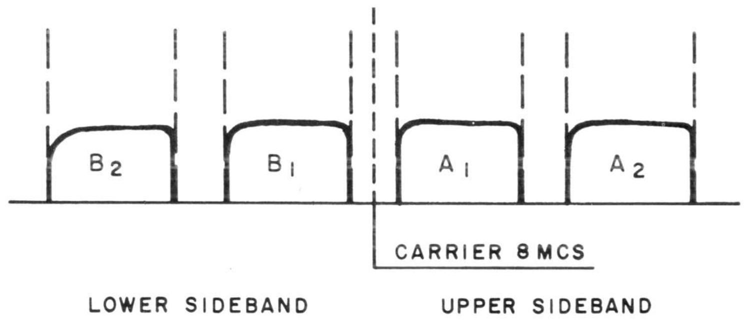
Figure 4-5.

Multiplexer #1 receives two independent audio spectrums from 300-3000 cycles. Spectrum #1 is processed "straight through"; spectrum #2 is translated upward to occupy a higher position in the audio spectrum. Multiplexer #2 performs the same functions with two additional independent audio spectrums.

The output of each multiplexer is applied to a low frequency balanced modulator in an ISB transmitter of the type shown in Figure 4-3. It follows that, in such a system, the sideband filters must have a bandpass of at least 7200 cycles. The output of such a transmitter, at an assigned frequency of 8 mcs, is shown below.



The intelligence "slots", as they are colloquially called, are usually identified as shown in the sketch on the following page.



4-6 Miscellaneous Systems Not Classified as True SSB Systems

a) Double Sideband (DSB) Operation:

In a Double Sideband System, two sidebands containing identical information, are transmitted without carrier. These sidebands are mirror images of each other. A true DSB signal, then, is similar to an AM signal except for the absence of carrier. A DSB transmitter usually employs a high level modulator which balances out the carrier.

In a DSB receiver, a special phase locked oscillator is employed to re-insert the carrier in the same phase relationship to the sidebands as in the original modulation at the transmitter.

Note: Certain TMC equipment is described as having "DSB" capability. This is not the same type of DSB operation referred to above.

In the DSB mode of TMC transmitting equipment, identical audio is applied to both low frequency balanced modulators of an ISB configuration. Thus, two identical sidebands are transmitted without carrier. These sidebands, however, have been processed through sideband filters which change the phase relationships between the sidebands.

b) Controlled Carrier Operation:

In controlled carrier operation, a low level carrier is transmitted with one (or two) sidebands during periods of voice modulation. During pauses in speech, nearly full carrier is transmitted.

In the receiving system, automatic frequency control systems with long time constants hold the exalted carrier on frequency.

Note: This type of "controlled carrier" operation should not be confused with conventional AM controlled carrier.

CHAPTER 5

THE NATURE OF SINGLE SIDEBAND SIGNALS

5-1 Introductory Note

In this chapter, elementary SSB waveforms will be exhibited, relationships between peak and average powers will be examined; basic diode balanced modulator action will be discussed. New terms will be introduced and defined.

5-2 Generating an Elementary DSB Waveform

Figure 5-1 shows a basic balanced modulator circuit using diodes. Because of the diode configuration, the circuit is called a "balanced bridge" modulator.

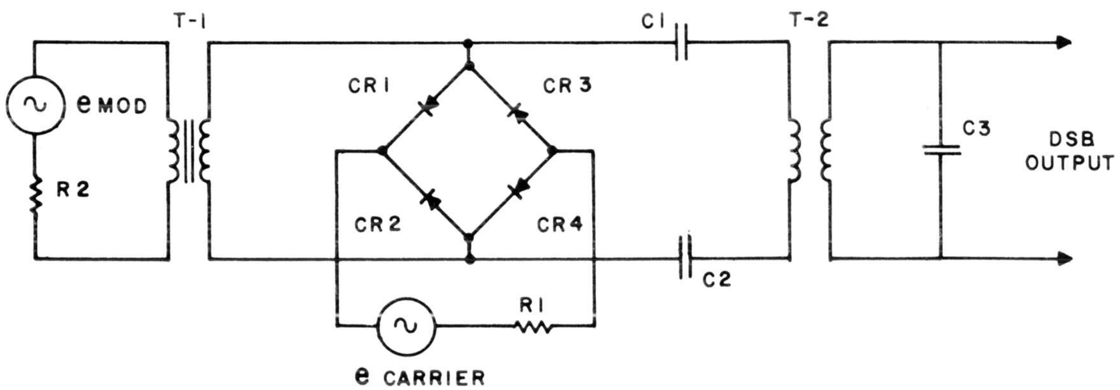


Figure 5-1.

Transformer T-1 couples the audio input signal to the modulator circuit. The secondary winding of T-2 resonates with C-3 to form a tuned circuit, the center frequency of which is the "carrier" frequency. The bandwidth of the entire system must be such as to pass twice the highest audio modulating frequency.

Capacitors C-1 and C-2 have appreciable reactance at audio frequencies and negligible reactance at the output sideband frequencies. The carrier voltage amplitude is 8 to 10 times the amplitude of the audio input voltage.

Diodes CR-1, CR-2, CR-3 and CR-4 may be of the selenium or copper oxide rectifier type, or they may be vacuum tubes or crystals. These diodes must be very carefully matched if proper carrier suppression is to be achieved. R-1 and R-2 are the internal resistances of the carrier and audio sources, respectively.

Assume that the carrier is a 100 KC sine wave at 10 volts, and that the modulating signal is a 1 KC sine wave at 1 volt. Figure 5-2 shows the frequency, time and amplitude relationships between these two signals for one cycle of the modulating voltage.

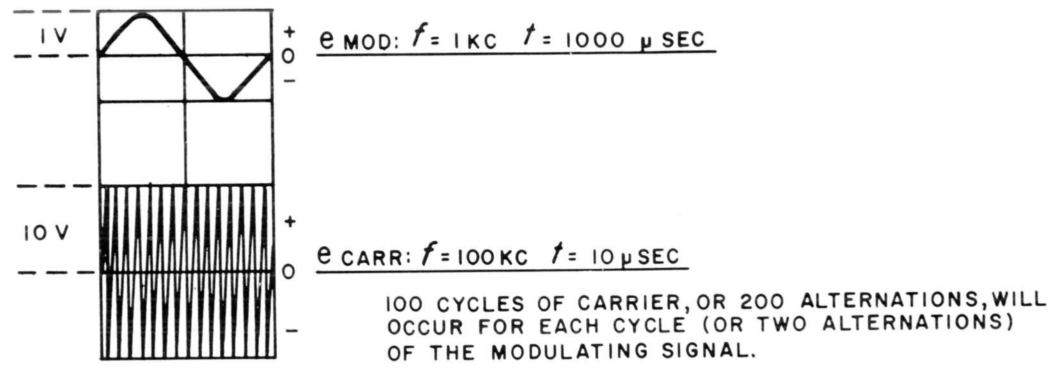


Figure 5-2.

Consider the action of the balanced bridge modulator circuit with carrier signal only. See Figures 5-3 and 5-4. For this explanation, the diodes will be considered to have zero forward resistance and infinite back resistance.

With the carrier generator polarity as shown in Figure 5-3, all diodes are forward biased and conduct equally. No potential difference exists between points A and B, since the bridge is balanced. No output appears at T-2.

With the carrier generator polarity as shown in Figure 5-4, all diodes are back biased; no current flows. At the crest of the carrier cycle 10 volts is impressed. The inverse voltage distribution is shown at this instant. Again, there is no potential difference between points A and B, and no output appears at T-2.

When modulation voltage is applied, and the carrier polarity causes the rectifiers to conduct, the bridge will be unbalanced and two side frequencies will appear in the output. When modulation is applied and the carrier polarity back biases the rectifiers, there will be no output because the back bias greatly exceeds the modulation voltage.

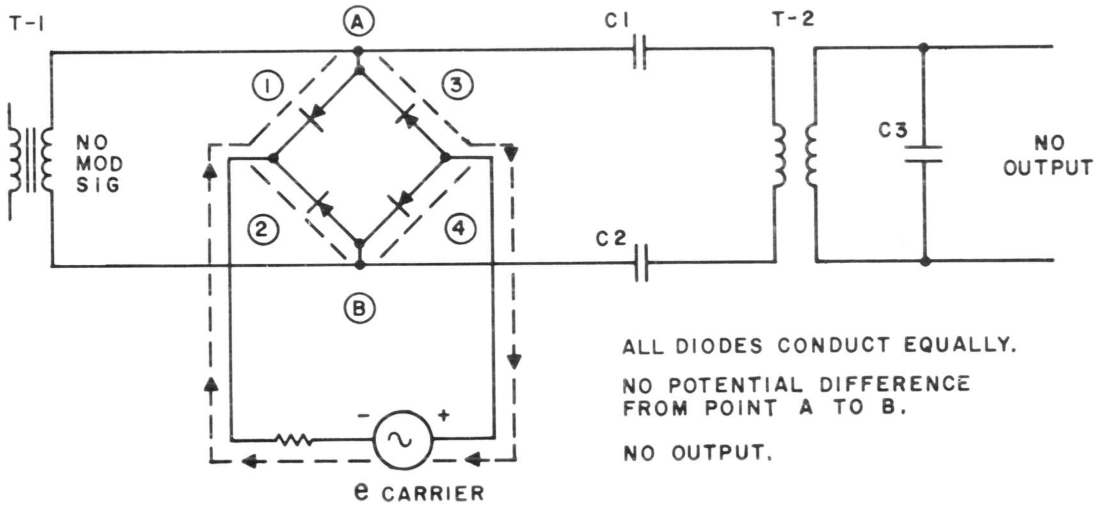


Figure 5-3.

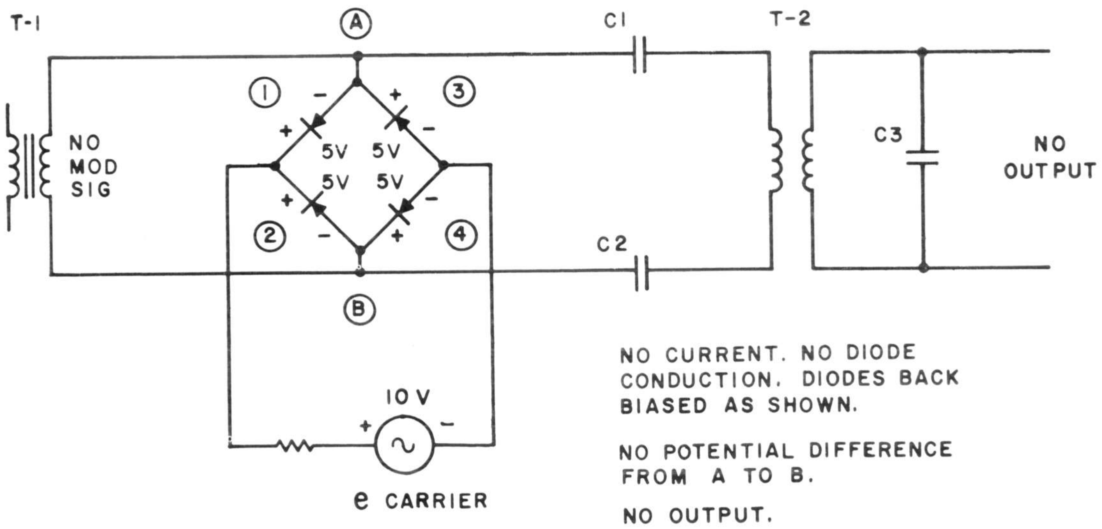


Figure 5-4.

Figure 5-5 shows the action when both modulation voltage and carrier voltage is applied to the modulator circuit. The carrier voltage alternations are shown greatly out of scale, in order that individual alternations may be seen. With the left side of the carrier generator negative, and the positive alternation of the modulating voltage being applied, diodes #1 and #4 increase conduction, because of the polarity of the modulation voltage. This unbalances the bridge; current flows as shown by the heavy dotted line. Current pulse amplitude of the current pulses follows the envelope of the modulation voltage. When the carrier generator polarity is positive at the left side, the diodes are back biased and no current flows.

It should be remembered that the carrier actually undergoes 50 negative alternations during the positive alternation of the modulating voltage, and that Figure 5-5 has been purposely distorted to show the action of the waveforms.

It should be apparent from an examination of Figure 5-5 that the current pulses occur at the repetition rate of the carrier, and at an amplitude corresponding to the instantaneous amplitude of the modulating voltage.

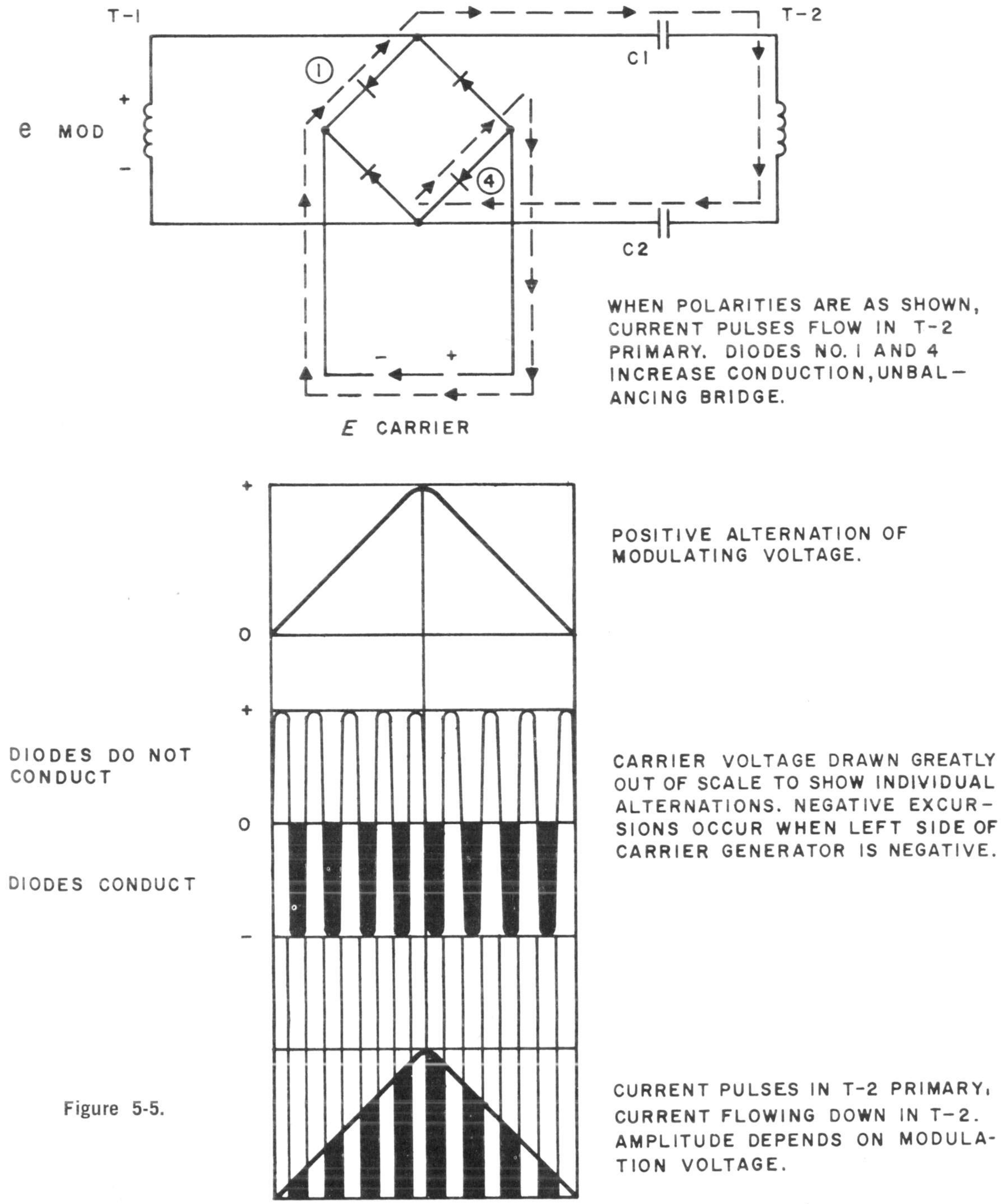


Figure 5-5.

The *current* pulses in the primary of T-2 cause pulses in the secondary of T-2; the secondary of T-2 is a high Q resonant circuit, tuned to the carrier frequency. The current pulses may be compared to the pulses of plate current in a Class C RF amplifier; these pulses "kick" the resonant tank, causing an alternating RF voltage to appear across the tank. The flywheel effect supplies the missing voltage alternations. The voltage output of T-2, then, resembles the illustration below:

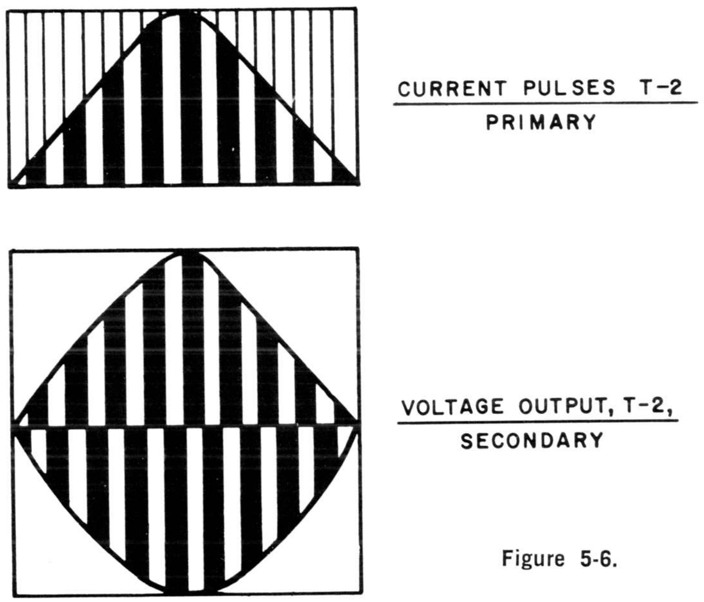


Figure 5-6.

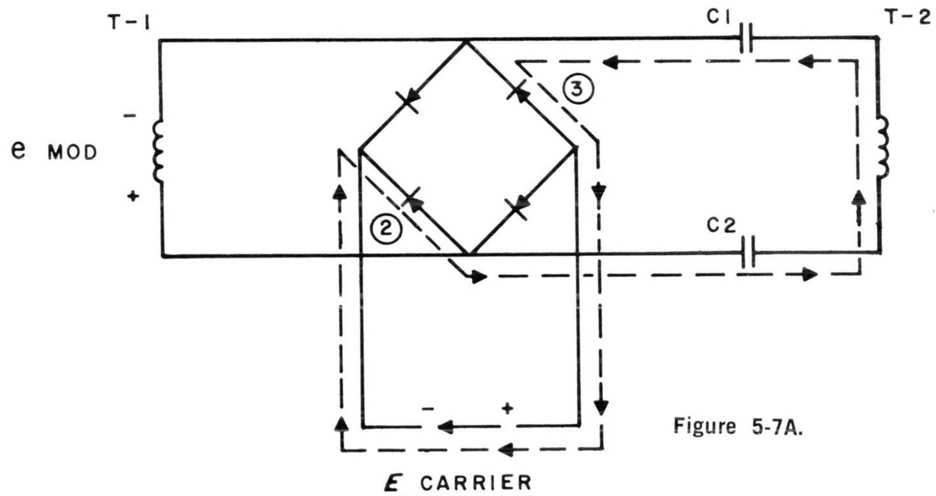


Figure 5-7A.

Refer to Figure 5-7A. When the polarity of the audio modulating signal reverses, and when the polarity of the carrier generator is such that the diodes conduct, diodes #2 and #3 increase their conduction due to the polarity of the modulation voltage; the bridge is again unbalanced; pulses of current flow in the primary of T-2, in a direction opposite to that shown

in Figure 5-5. An envelope, similar to that shown in Figure 5-6, appears at the secondary of T-2.

Figure 5-7B, below, shows the relationship of the the complete RF output envelope from T-2 as a result of one complete cycle of the modulation voltage.

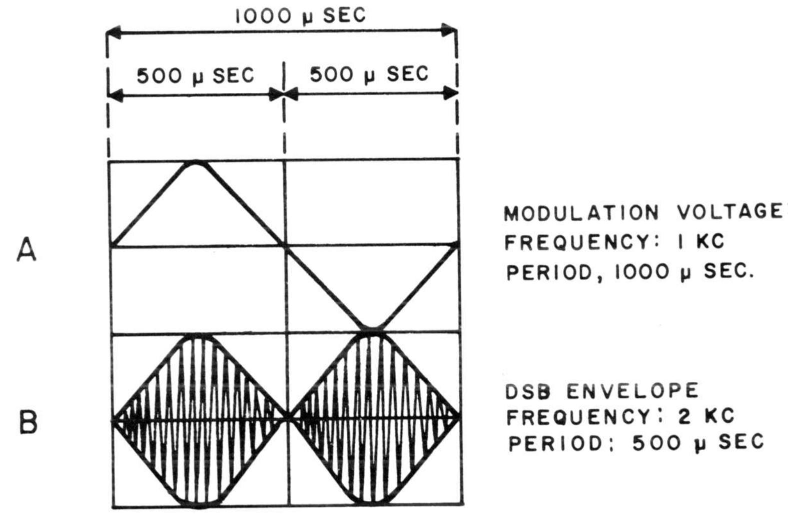


Figure 5-7B.

The waveform of Figure 5-7B is a double sideband suppressed carrier, (DSSC) signal. The waveform is composite, representing two frequencies:

f carrier plus f modulation: 101 KC
f carrier minus f modulation: 99 KC

The repetition rate of the waveform, 2 KC, is the difference in frequency between the two sideband components.

If this waveform is applied to two highly selective filters, as shown below, the two sideband components, sinusoidal in shape, will be recovered.

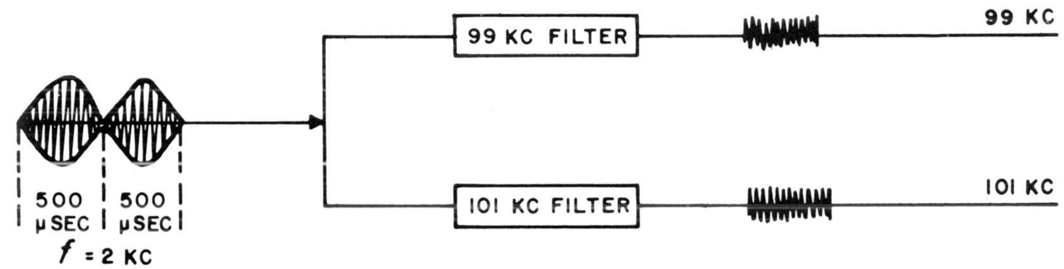


Figure 5-8.

If either the 99 KC output or the 101 KC output is supplied with a re-inserted carrier equal in amplitude to the RF tone, a DSB envelope similar in shape to that of Figure 5-8 will be observed. However, this new envelope will have a repetition frequency of 1 KC, the difference in frequency between the tone and the re-inserted carrier. This is shown in Figure 5-9, below.

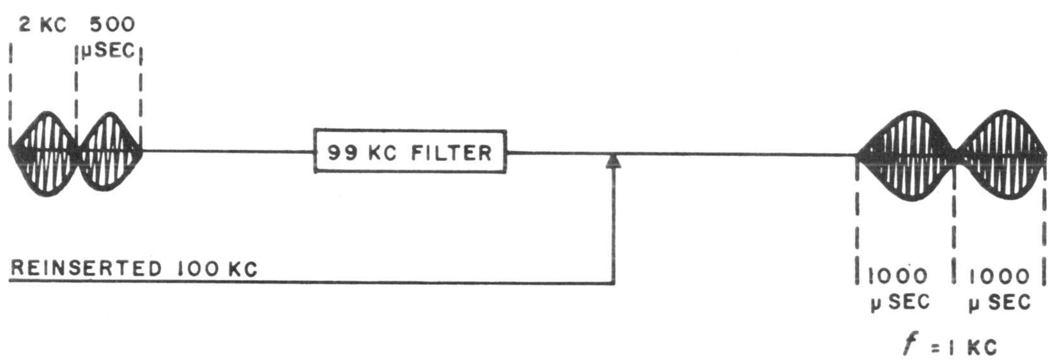


Figure 5-9.

In Figure 5-9, if the re-inserted carrier is not equal in amplitude to the single RF tone, the characteristic two tone waveform will not be observed. The sketch below shows the resultant waveforms for various degrees of carrier re-insertion.

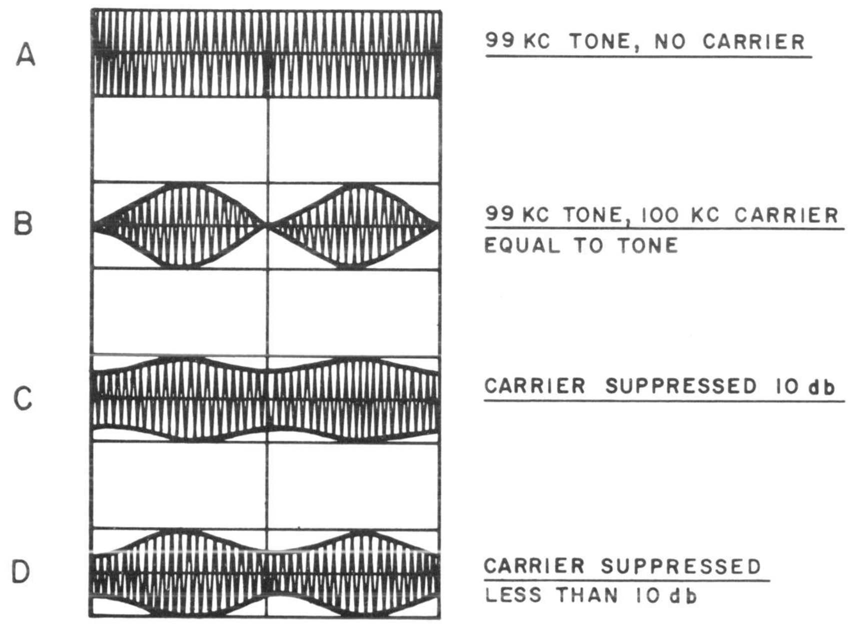


Figure 5-10.

Figure 5-11 below shows a two tone DSB envelope developed from a single tone and carrier in a balanced modulator. The figure also shows the same envelope with carrier "leakage", or with a small amount of purposely re-inserted pilot carrier.

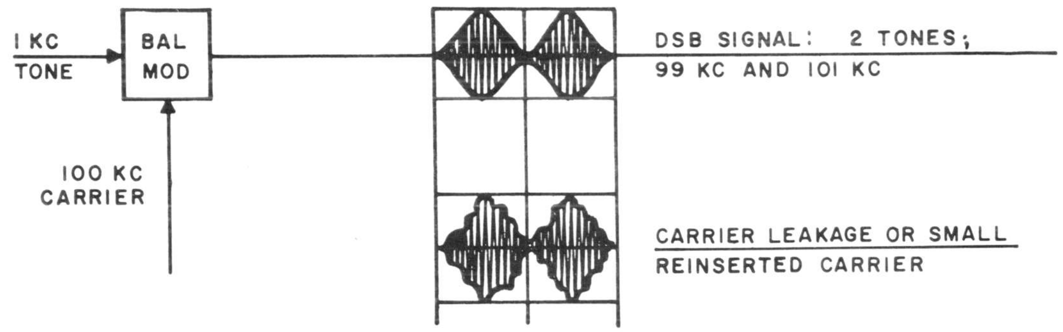


Figure 5-11.

5-3 Multitone Operation of SSSC Transmitters

The discussion which follows has both practical and impractical aspects. The discussion is impractical in that it involves a theoretical treatment evolving from a laboratory experiment which cannot be duplicated under field conditions. The discussion is extremely practical in that it will provide a thorough understanding of:

a) the term: PEAK ENVELOPE POWER (PEP), as applied to SSB transmitters.

b) the relationships between PEAK ENVELOPE POWER and AVERAGE POWER in SSB transmitters.

c) The insurmountable difficulties encountered when attempts are made to transmit square waves with SSB transmitters.

Examine the experimental test set up presented in the sketch of Figure 5-12.

The tone generator is capable of generating 16 discrete sinusoidal tones. Each tone unit may be turned on and off independently. There is a fixed frequency difference between consecutive tones; therefore, all tones may be started "in phase". The tone frequencies are of equal amplitude.

The output of the tone generator is applied to a SSSC exciter, which passes only the upper sideband, without carrier or any unwanted radiations. The SSSC exciter is terminated in a perfect resistive dummy load, of the proper value. The output is monitored by:

a) a high quality oscilloscope.

b) a high quality spectrum analyzer.

c) a peak reading RF VTVM, calibrated to read RMS.

Note: The experiment being discussed was actually performed in the TMC Engineering laboratory; however it was impossible to achieve all of the conditions called for. The 16 tone generator was a Telesig Corp. 16 channel TTY generator, the frequencies of which differ from those specified in Figure 5-12. The tones did not have the frequency stability required, nor could they be phase locked. Sufficient information was obtained, however, to develop this theoretical treatment.

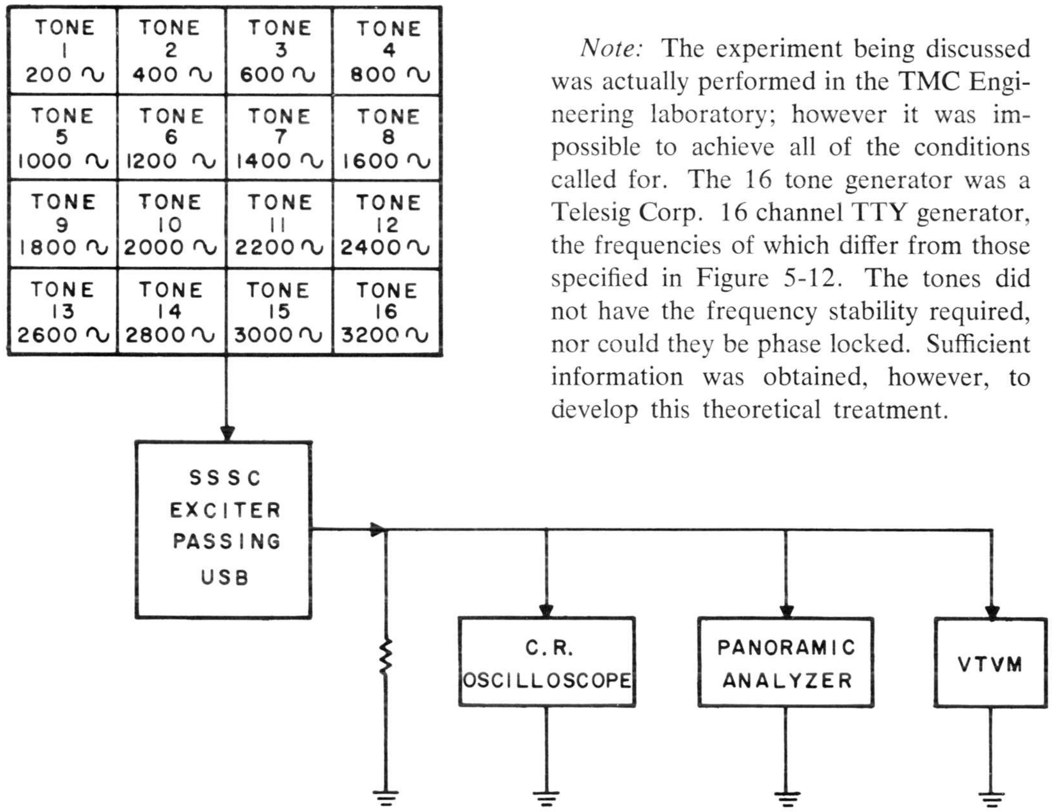


Figure 5-12.

Results with Tones #1 and #2 applied:

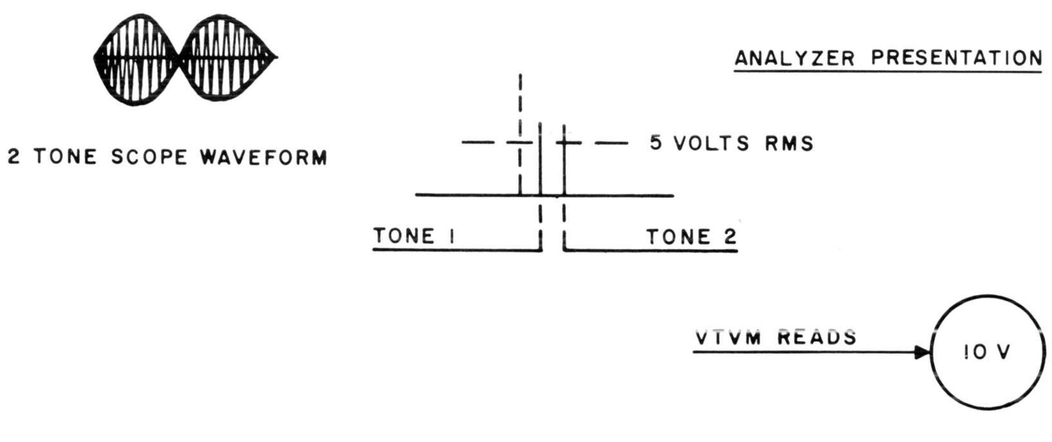


Figure 5-13.

With two tones applied, the characteristic two tone oscilloscope waveform is obtained. The spectrum analyzer shows the two tones of equal amplitude at 5 volts RMS, and the VTVM shows the RMS value of the peak voltage, 10 volts.

Average Power for a number of tones is the sum of the powers of each tone. Then, for two tones,

$$P_{AVG} = \frac{E_1^2}{R} + \frac{E_2^2}{R} = \frac{5^2}{50} + \frac{5^2}{50} = .5 + .5 = 1 \text{ watt}$$

PEAK ENVELOPE POWER is defined as the RMS power developed at the peak of the modulation envelope.

The peak is developed when the two tones are in phase; therefore, the peak envelope voltage, PEV, is 10.

↑5V
↑5V PEV = 10V

Peak envelope power, then, is:

$$PEP = \frac{(E_1 + E_2)^2}{R} = \frac{10^2}{50} = \frac{100}{50} = 2 \text{ watts}$$

We now state, without proof, the following theoretical relationships:

$$\frac{P_{AVG}}{PEP} = \frac{1}{2} = .5 = \frac{1}{N}$$

where N is the number of tones

$$\frac{P_{EACH\ TONE}}{PEP} = \frac{.5}{2} = .25 = \frac{1}{N^2}$$

PEP is 2 $\times$ P_{AVG} in a two tone condition.

The case of the two tone envelope is extremely important; in this case, THEORETICAL MEASUREMENTS AND PRACTICAL MEASUREMENTS WILL BE IDENTICAL. This is the basis for the widely used TWO TONE TEST in SSB transmitting equipment.

Results With Tones #1, #2 and #3 applied:

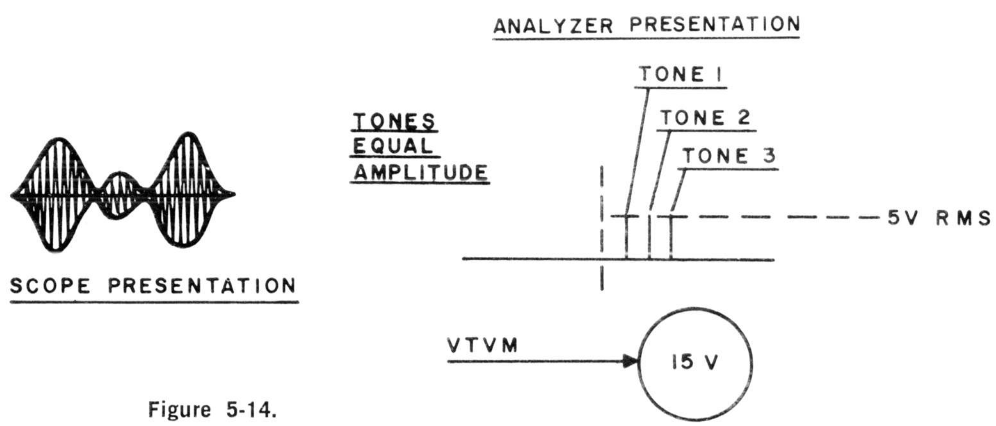


Figure 5-14.

Notice that, on the oscilloscope, the peak amplitude has increased, and that a small "bubble" appears in the presentation. The two original peaks have been "squeezed" together slightly, to make room for the "bubble". The spectrum analyzer shows three tones of equal amplitude, at 5 volts RMS. The VTVM now reads 15 volts.

$$P_{AVG} = \frac{E_1^2}{R} + \frac{E_2^2}{R} + \frac{E_3^2}{R} = .5 + .5 + .5 = 1.5 \text{ watts}$$

$$P_{EV} \begin{matrix} 5v \\ 5v \\ 5v \end{matrix} = 15v$$

$$PEP = \frac{(E_1 + E_2 + E_3)^2}{R} = \frac{15^2}{50} = \frac{225}{50} = 4.5 \text{ watts}$$

$$\frac{P_{AVG}}{PEP} = \frac{1}{N} = \frac{1.5}{4.5} = \frac{1}{3}$$

$$\frac{P_{EACH\ TONE}}{PEP} = \frac{1}{N^2} = \frac{.5}{4.5} = \frac{1}{9}$$

Results With Tones #1, #2, 3# and #4 applied:

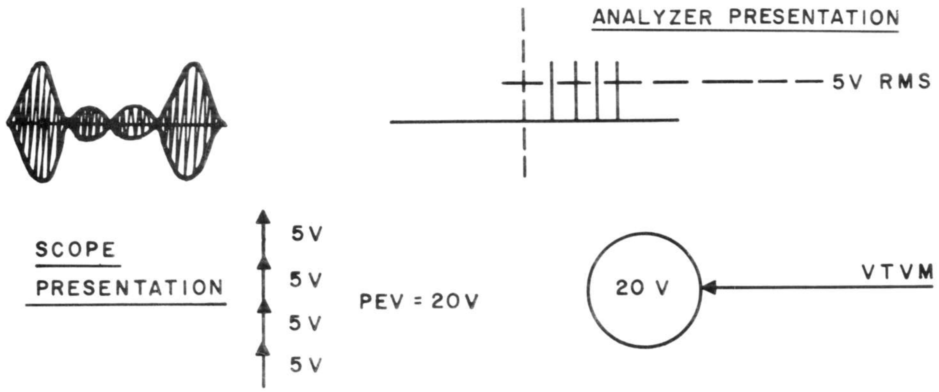


Figure 5-15.

The peak amplitude on the oscilloscope presentation has again increased. A second "bubble" appears in the waveform. The two original peaks have been shortened further, and their sides are now steeper. The spectrum analyzer shows four tones of equal amplitude at 5 volts RMS. The VTVM now reads 20 volts.

$$P_{AVG} \frac{E_1^2}{R} + \frac{E_2^2}{R} + \frac{E_3^2}{R} + \frac{E_4^2}{R} = .5 + .5 + .5 + .5 = 2 \text{ watts}$$

$$PEP = \frac{(E_1 + E_2 + E_3 + E_4)^2}{R} = \frac{(2 \times 10^1)^2}{5 \times 10^1}$$

$$= \frac{4 \times 10^2}{5 \times 10^1} = 8 \text{ watts}$$

$$\frac{P_{AVG}}{PEP} = \frac{1}{N} = \frac{1}{4} = .25 = \frac{2}{8} = \frac{1}{4} = .25$$

$$\frac{P_{EACH\ TONE}}{PEP\cdot} = \frac{1}{N^2} = \frac{1}{16} = .0625 = \frac{.5}{8} = 0.625$$

Results With Tones #1, #2, #3, #4 and #5 applied:

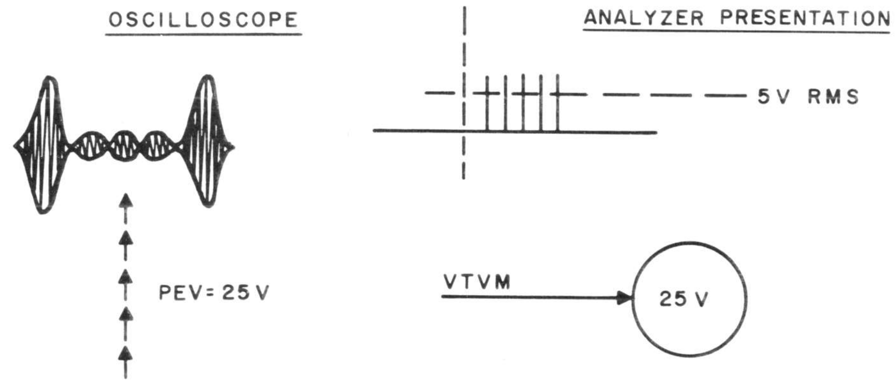


Figure 5-16.

The peak amplitude of the oscilloscope presentation has again increased; this is compensated for by adjustment of the vertical VOLTS/CM control. A third "bubble" appears, and the main peaks have been squeezed together, with attendant steepening of their sides. The spectrum analyzer shows five tones of equal amplitude at 5 volts RMS, and the VTVM reads 25 volts.

$$P_{AVG} = \frac{E_1^2}{R} + \frac{E_2^2}{R} + \frac{E_3^2}{R} + \frac{E_4^2}{R} + \frac{E_5^2}{R} = 2.5 \text{ watts}$$

$$PEP = \frac{(E_1 + E_2 + E_3 + E_4 + E_5)^2}{R} = \frac{25^2}{50} = 12.5 \text{ watts}$$

$$\frac{P_{AVG}}{PEP} = \frac{1}{N} = \frac{1}{5} = .2 = \frac{2.5}{12.5} = .2$$

$$\frac{P_{TONE}}{PEP} = \frac{1}{N^2} = \frac{1}{25} = .04 = \frac{5 \times 10^{-1}}{12.5 \times 10^0} = .04$$

An Examination of the Patterns Thus Far Obtained:

The sketches of the oscilloscope waveforms thus far presented were actually obtained in the TMC experiment; difficulty was experienced in synchronizing the oscilloscope presentation after five tones because, as previously stated, the test conditions did not meet exactly the conditions specified.

It is apparent that, as the number of tones is increased, the peak envelope power increases, and the main peaks, (at a repetition rate determined by the tone spacing), became narrower, with steeper sides. It may be expected that, with the 16 tones applied, the following theoretical conditions would be observed.

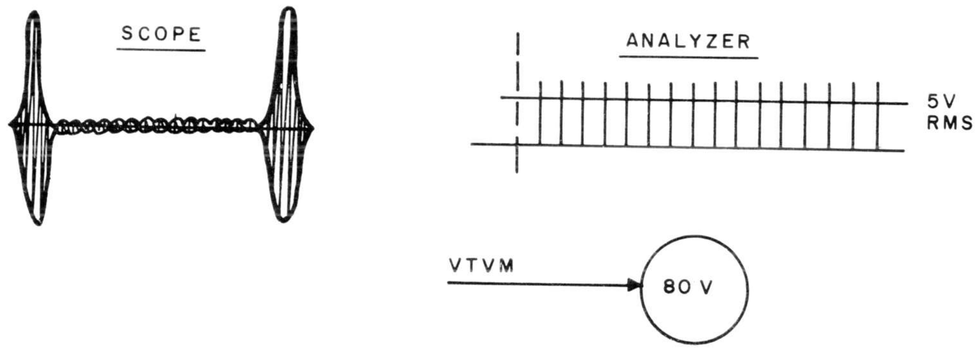


Figure 5-17.

$$P_{AVG} = \frac{E_1^2}{R} + \frac{E_2^2}{R} + \ldots \frac{E_{16}^2}{R} = 8 \text{ watts}$$

$$PEP = \frac{(E_1 + E_2 + \ldots E_{16})^2}{R} = 128 \text{ watts}$$

$$\frac{P_{AVG}}{PEP} = \frac{1}{N} = \frac{1}{16} = .0625 = \frac{8}{128} = .0625$$

$$\frac{P_{TONE}}{PEP} = \frac{1}{N^2} = \frac{1}{256} = .00391 = \frac{.5}{128} = .00391$$

In this particular system, then, each tone consumes an AVERAGE power of .5 watts. The total AVERAGE power is 16 x .5 or 8 watts.

However, the theoretical voltage peak at the crest of the modulation cycle is 80 volts. The PEAK ENVELOPE POWER corresponding to this voltage peak is 128 watts.

Thus, the system must be capable of furnishing 128 watts on peaks if distortion is to be avoided.

Consider, for a moment, a transmitter rated at 10 KW PEP, 5 KW Average with two tones applied.

With two tones applied, PEP is 2 $\times$ P average. IF, WITH 16 TONES APPLIED, THE SYSTEM IS OPERATED AT 5 KW AVERAGE, THE SYSTEM MUST BE CAPABLE OF HANDLING A PEAK ENVELOPE POWER OF 80 KILOWATTS !!

$$\frac{P_{AVG}}{PEP} = \frac{1}{N} = \frac{1}{16} \quad PEP = P_{AVG} \cdot N = 80 \text{ kw.}$$

Consider, now, the waveform change from 2 to 16 tones. The tones are harmonically related: tone #1, at 200 cycles, is followed by the second, third, fourth, etc. to the 16th harmonic. It is reasonable to assume that, if more harmonics were added, the main peaks would become steeper. If an infinite number of harmonics were added, the waveform might look like the sketch of figure 5-18. This waveform would require AN INFINITE BANDWIDTH and INFINITE POWER.

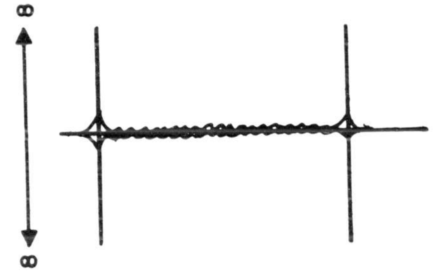


Figure 5-18.

When a fundamental and an infinite number of even and odd harmonics are combined, in the proper phase relationship and amplitude, the result is a sawtooth wave. When a fundamental and an infinite number of odd harmonics are combined in the proper phase relationship, the result is a square wave.

It should be apparent, then, that perfect waves of this type cannot be transmitted by a SSB system without distortion. This fact will take on more significance when the processing of voice waveforms, or any waveforms with a high peak to average ratio, is discussed.

PRACTICAL ASPECTS OF THE TONE EXPERIMENT

Table 5-1 shows the relationship of Average Power to Peak Envelope Power as a function of the number of tones. The theoretical values are obtained if all tones are as specified in the opening paragraph. The actual values obtained in the experiment are shown also. Figure 5-19 is a graph of Table 5-1.

Note that at two tones the theoretical and practical values are identical, and that such is the case until the fifth tone is added. This was the point in the experiment at which it became extremely difficult to obtain the desired pattern on the oscilloscope.

It must be stressed again that the possibilities are extremely remote that a transmitter will be expected to handle 16 absolutely stable, phase locked tones with equal frequency spacing. For example:

Consider a sideband signal carrying 16 FSK teletype channels. At any instant, each channel may be in a MARK, SPACE or OFF condition. These tones are not phase locked, do not have equal frequency spacing, and may be subject to slight drift. This presents a combination of events impossible to predict; the action is, at best, random.

Now consider a SSB signal carrying audio voice intelligence from 300 to 3000 cycles. This is a conglomeration of fundamentals, overtones and combinations at widely different amplitudes. Consider also that no two individuals will generate identical voice spectrums. The voice spectrum has a high peak to average ratio, and the highest peaks occur at a random rate.

Thus, the tone experiment serves to illustrate the relationships among Peak Envelope Power, Average Power and a number of tones in an ideal situation, but it does not produce a clear cut formula for setting up a transmitter to insure optimum operation without distortion.

TABLE 5-1: RATIO OF AVERAGE POWER TO PEAK ENVELOPE POWER AS A FUNCTION OF SINGLE TONES

Number of Tones	*Theoretical*	*Actual*
1	1.00	1.00
2	.50	.50
3	.333	.333
4	.250	.250
5	.200	.210
6	.166	.175
7	.143	.155
8	.125	.140
9	.111	.120
10	.100	.110
11	.090	.105
12	.083	.100
13	.077	.100
14	.0715	.100
15	.0666	.100
16	.0625	.100

To Summarize:

a) overall transmission efficiency depends on the amount of AVERAGE POWER radiated.

b) maximum power is limited to the PEAK POWER capability of the transmitter.

c) PEAK ENVELOPE POWER is the RMS value of power at the crest of the modulation envelope. With certain intelligence, the repetition rate of the peak envelope can be determined, as can the ratio of AVERAGE POWER to PEAK ENVELOPE POWER. With most intelligence, the peaks occur at a random rate, and at random amplitude.

d) excessively high peaks may occur occasionally, yet not often enough to warrant reducing the average transmitter power.

e) a high ratio of AVERAGE POWER TO PEAK ENVELOPE POWER is desirable.

f) a method must be incorporated to transmit as much average power as possible, while limiting the occasional high peaks of the envelope to a point within the capability of the transmitter at which a minimum of distortion is generated.

5-4 Automatic Load and Drive Control (ALDC)

Automatic Load and Drive Control, ALDC, is one method used in TMC transmitters to limit high peaks of the modulation envelope in the final

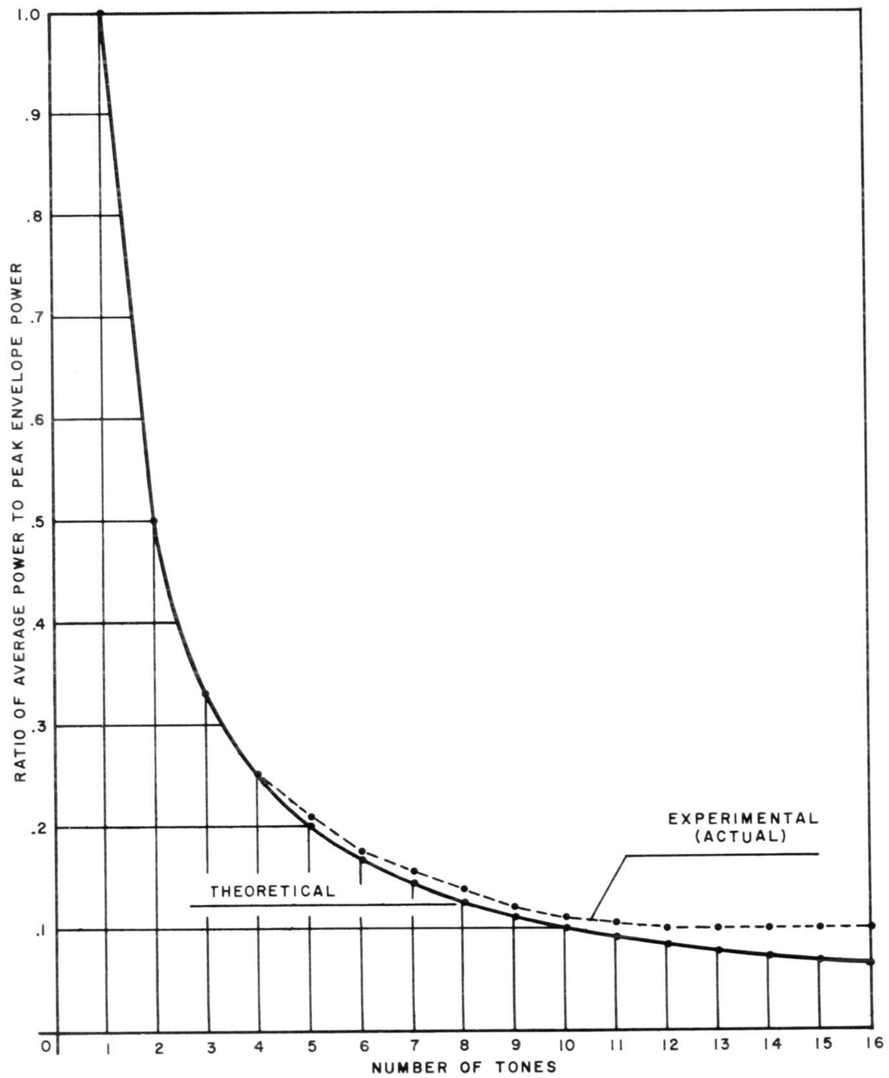


Figure 5-19. Ratio of P average to PEP as a Function of Tones.

stages of the transmitter. Essentially, the ALDC system samples the output envelope, and develops a bias when the envelope crest exceeds a certain value. This bias is returned to a previous stage to reduce the drive. A simplified sketch of an ALDC arrangement is presented in figure 5-20.

A small sample of the output SSB envelope is coupled to the ALDC circuits by C-1. CR-1 is the ALDC rectifier, the cathode of which is back biased by a variable DC voltage. R-1, the ALDC control, determines the back bias on CR-1 and, therefore, the point at which the diode will conduct. When the peak value of the SSB envelope exceeds the bias, CR-1 conducts, developing a negative voltage which is stored in C-2. The negative voltage is filtered and applied to the control grid circuit of a Class A amplifier, which feeds the driver stage. With such an arrangement, the transmitter may be operated near its maximum power capability without being overdriven on high signal peaks. The circuit should have a fast attack time, so that the overload is relieved at once. A typical TMC circuit of this kind will be discussed in a subsequent chapter.

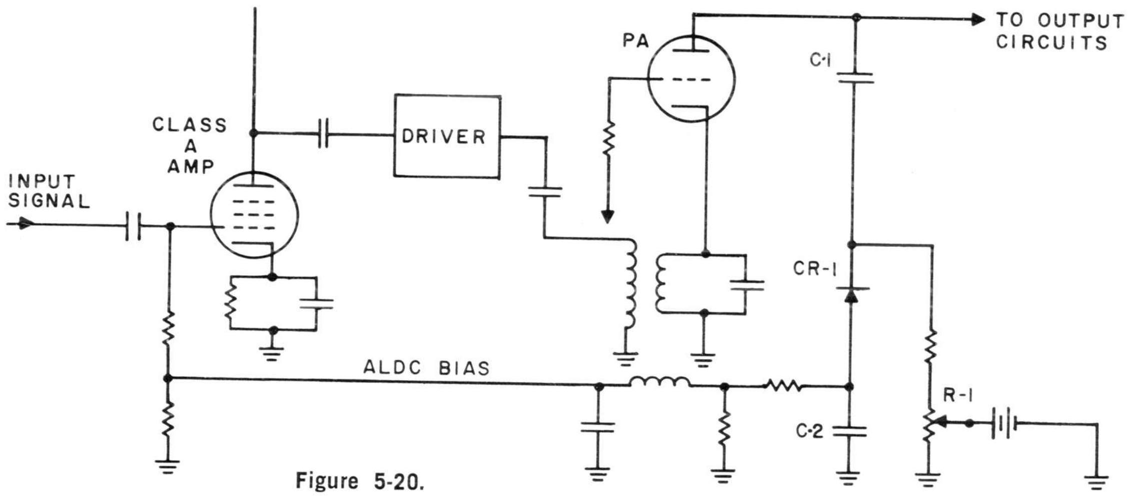


Figure 5-20.

5-5 Speech Processing for Single Sideband Operation

INTRODUCTORY NOTE

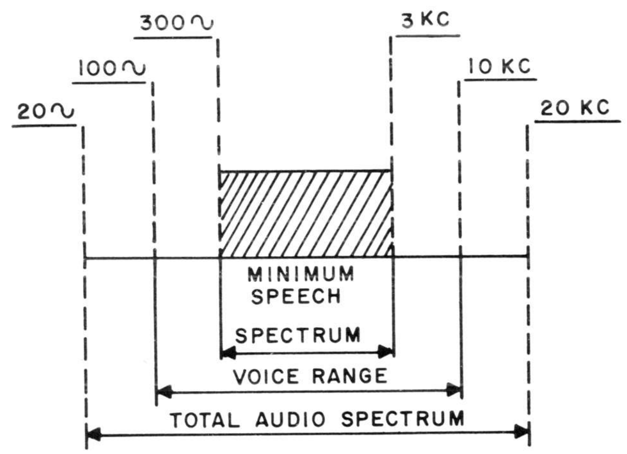
Thus far the discussions concerning Peak Envelope Power and Average Power have centered around the transmission of single tones or groups of tones having equal amplitude; no mention has been made of special requirements for the transmission of speech. When speech is to be transmitted, optimum results will be obtained only if the initial speech intelligence has been properly processed. This processing may be accomplished by one or more techniques, each of which will subsequently be described. Before discussing these techniques, an elementary examination of certain characteristics of speech and speech waveforms will be undertaken.

THE FREQUENCY BANDWIDTH REQUIRED FOR ARTICULATE SPEECH

The complete audio spectrum covers the range from about 20 cps to 20,000 cps. For "high fidelity" applications, encompassing the full range of voice and orchestra, this complete spectrum is involved. When speech intelligence is to be transmitted by commercial or military means, the only criteria is "readibility", that is, how well the information contained in the speech intelligence is understood at the point of reception. The extent to which a transmission system is capable of reproducing the *original speech meaning* is measured in terms of *ARTICULATION*. Articulation should not be confused with naturalness.

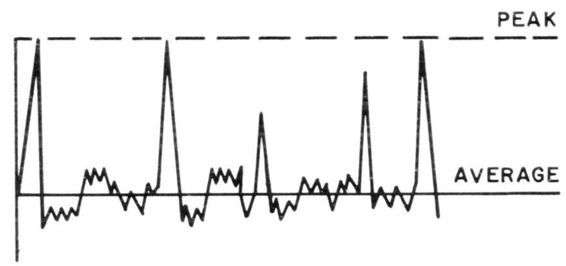
The sounds of speech lie in the range 100 cps- 10KCS, but articulate speech may be conveyed in a spectrum from about 300 cps - 3 KCS. Since the average power must be distributed throughout the entire spectrum transmitted, it is most economical to employ the narrowest feasible bandwidth. Thus, we will be concerned with the speech spectrum from about 300 cps to 3000 cps.

Figure 5-21A is a plot of the total audio spectrum, the total speech spectrum, and the optimum audio spectrum for articulate speech.



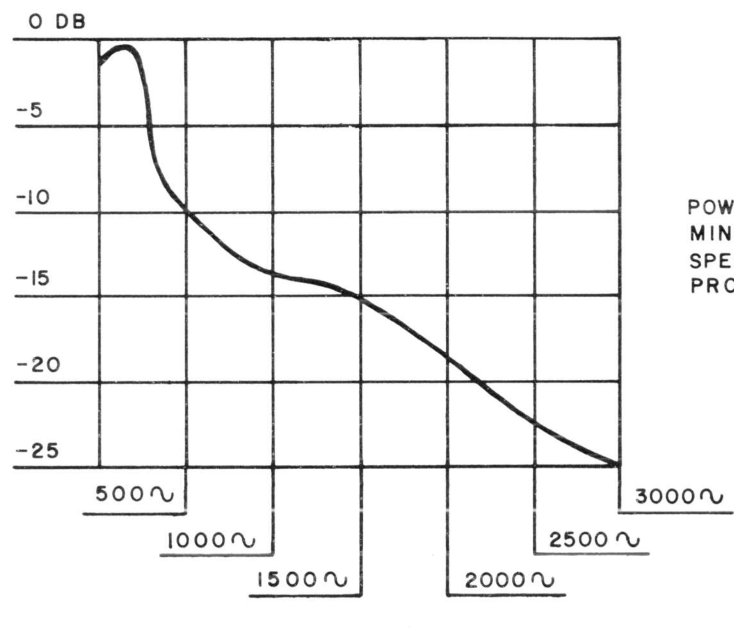
THE COMPLETE AUDIO SPECTRUM, THE COMPLETE SPEECH SPECTRUM, AND THE MINIMUM SPEECH SPECTRUM FOR GOOD ARTICULATION

Figure 5-21A.



A TYPICAL SPEECH PATTERN, SHOWING THE RELATIVELY HIGH PEAK TO AVERAGE RATIO.

Figure 5-21B.



POWER DISTRIBUTION IN THE MINIMUM FEASIBLE SPEECH SPECTRUM PRIOR TO PROCESSING.

Figure 5-21C.

THE PEAK TO AVERAGE RATIO OF A SPEECH WAVEFORM

The peak to average ratio of a pure sine wave is 2:1; that is, the average power in a pure sine wave is one half the peak power.

The average speech waveform is characterized by intense peaks of short duration occurring at a high repetition rate. The peaks are due to the high harmonic content of the speech waveform. Thus, the peak to average ratio of a speech waveform is much greater than that of a sine wave with the same peak power. Some means must be devised, then, to "compress" the dynamic range of the human voice to make it more compatible with the requirements of an electronic communication system.

Figure 5-21B shows a typical speech pattern. The peak power is much greater than the average power because of the recurring peaks of high intensity. There is little "talk power" in this waveform, because, if the transmitter is operated at its rated average power, the peak envelope power will exceed the power handling capability of the transmitter, with resultant "splatter" and distortion.

POWER DISTRIBUTION IN THE VOICE SPECTRUM

Figure 5-21C shows the distribution of power in the minimum feasible voice spectrum. Note that a peak occurs at about 200 cycles, and that power diminishes rapidly until, at 3000 cycles, the power is down 25 DB from the power at the 200 cycle peak.

It has been found that the vowel sounds, a, e, i, o, u, add little to articulation, and that the consonant sounds are extremely important to articulation. The vowels are associated with the low frequencies, and the consonants are associated with the high frequencies. If the high power, low frequency components are attenuated, and the low power high frequency components are emphasized, articulation is not adversely affected, and the peak to average ratio of the speech waveform is considerably reduced.

FREQUENCY BAND COMPRESSION

Frequency Band Compression, applied to speech, is a technique whereby electronic circuits are designed to pass the minimum feasible speech spectrum, and to attenutate all other frequencies.

SPEECH COMPRESSION

Speech Compression is a technique which improves the peak to average ratio of a speech waveform by means of AGC action. Speech compression also compensates for different speech intensities produced at the microphone by individuals with widely different voice characteristics.

The compressor circuit is usually an automatic variable gain amplifier employing a threshold control which determines the input level at which compression takes place. A portion of the output signal is usually rectified, and applied, as bias, to a previous stage. The audio output is thus kept at

a reasonably constant level, and the high peaks characteristic of the speech waveform are reduced in amplitude. An elementary amplifier and compressor arrangement is shown in the sketch below:

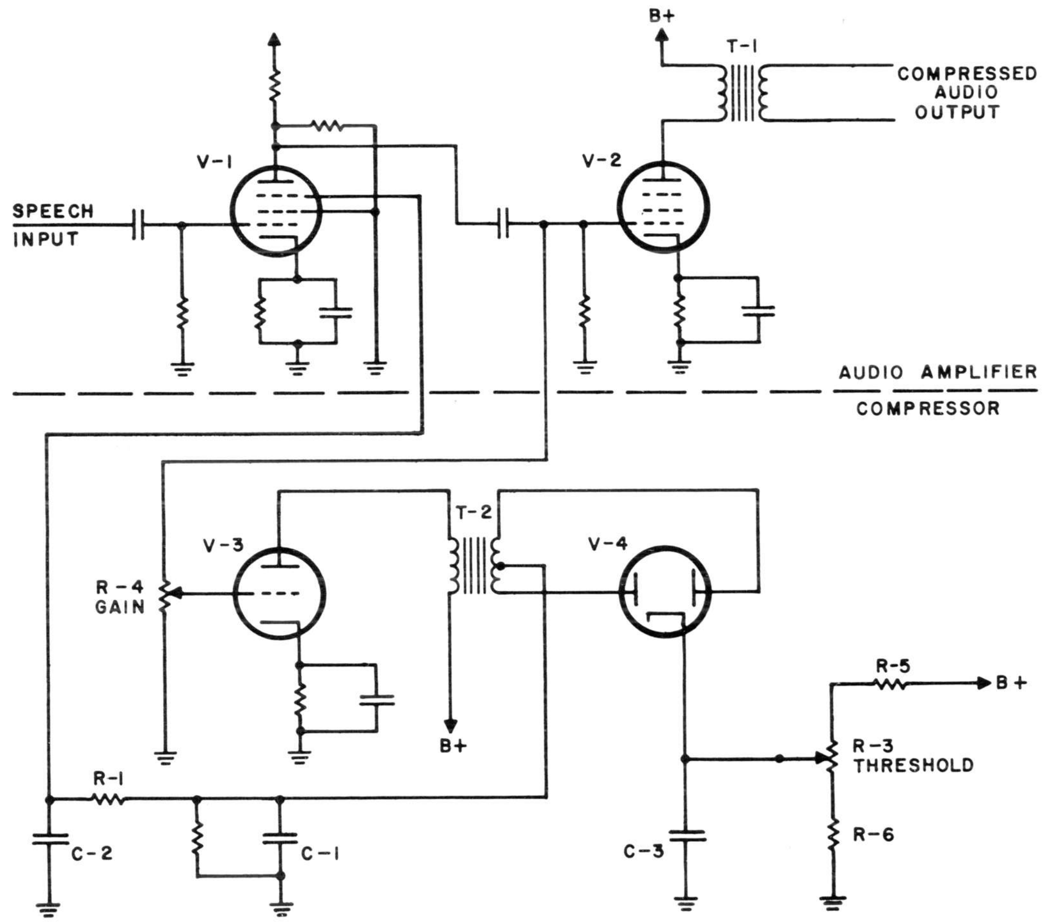


Figure 5-22.

In the circuit of Figure 5-22, V-1 is a voltage amplifier, connected in cascade with power amplifier stage V-2. The suppressor grid of V-1 is supplied with a negative voltage from the compressor circuit to regulate the overall gain.

The audio signal applied to the control grid of V-2 from the first stage is also applied to the control grid of compressor amplifier V-3, via a GAIN control, R-4.

The output of compressor amplifier V-3 is applied, via T-2, to full wave rectifier V-4. On each alternation of polarity at the secondary of T-2, one half of V-4 conducts, developing a negative voltage at C-1. This voltage is filtered by R-1, C-2, and is applied to the supressor grid of V-1.

The cathode of rectifier V-4 is held at some positive potential by the voltage divider network R-5, R-3, R-6. Capacitor C-3 holds this voltage constant with changes in the signal. R-3 sets the "threshold" of compression; that is, it determines the amplitude of signal from V-3 which will cause the rectifier to conduct and compression to take place.

R-4, the gain control in the grid circuit of V-3, determines the rate at which the gain is reduced with increasing signal level, for any given setting of R-4.

It should be emphasized that the compressor circuit does not *clip* signal peaks; it merely compresses them. Figure 5-23, below, shows a typical speech pattern before and after compression.

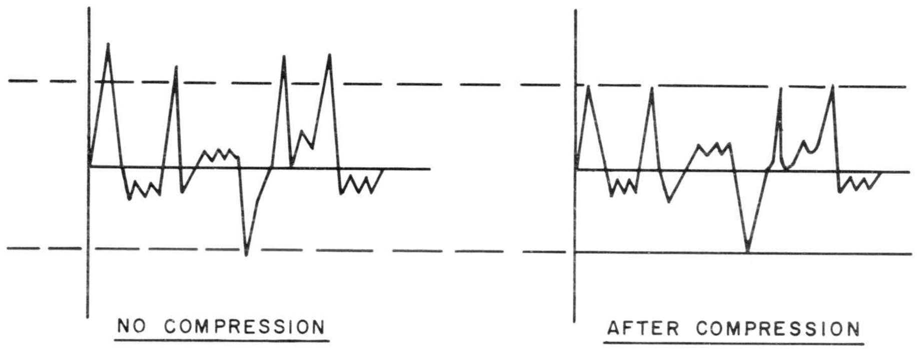


Figure 5-23.

PEAK CLIPPING

Peak Clipping is a technique which reduces the peak to average ratio of a speech waveform by clipping signal peaks which exceed a given reference. There are many types of speech clipping circuits; for this explanation, the most elementary circuit has been chosen for simplicity. Consider Figurès 5-24A and 5-24B.

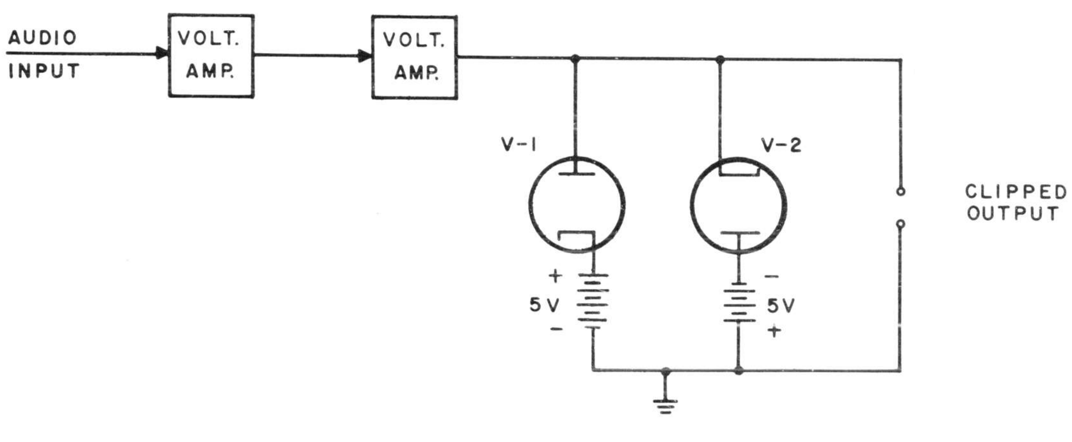


Figure 5-24A.

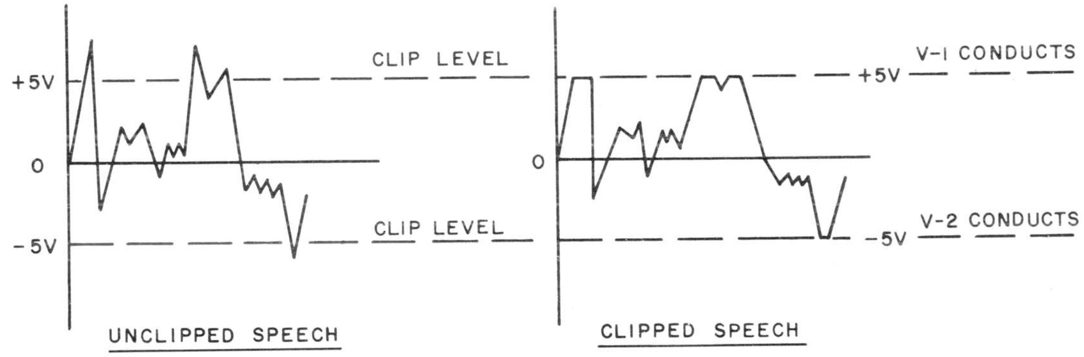


Figure 5-24B.

Figures 5-24A and B show an elementary clipper circuit and a speech waveform before and after clipping.

As long as the signal level remains below the clipping level, (5 volts in this case), neither diode will conduct, since each is back biased by a 5 volt supply. The diodes, then, are open circuits, and the output waveform is a replica of the input waveform.

Now consider the first positive excursion of the unclipped input waveform; when the signal level reaches 5 volts diode V-1 conducts; since the diode can be considered as a short circuit, the positive terminal of the 5 volts is effectively connected to the output and the level is maintained at this value until the signal input level falls below 5 volts. V-2 cannot conduct during this period because the increased positive voltage on its cathode is driving it further into cutoff. V-2 operates in the same manner as V-1 except on input negative peaks.

Clipping has one distinct disadvantage; it generates harmonics. Consider the speech waveshape after clipping: the clipped waveform is approaching a rectangular shape, with flat top and steep sides; this is indicative of a fundamental frequency and a large number of harmonics. Return for a moment to Figure 5-18. It was previously established that a square wave applied to a SSB transmitter would require infinite power and infinite bandwidth; it is apparent that clipping aggravates the very condition it is designed to correct. The answer lies in Frequency Band Compression, or filtering. The clipping circuit must be follower by a filter circuit which passes only the minimum feasible audio frequencies required for speech, and attenuates all other frequencies.

CLIPPING LEVEL

Clipping Level is generally given in decibels. For example, a system may employ "nominal 10 db clipping". This means that, if the original peak amplitude is 10 volts, the resulting maximum amplitude will be 3.16 volts, which is 10 DB down from 10 volts.

Tests have shown that 6 DB of clipping is just noticeable; 12 DB of clipping is acceptable, even though the "naturalness" of the voice suffers, and that 20 to 25 DB of clipping is tolerable in the presence of interference.

PREEMPHASIS

Preemphasis is a technique which compensates for the uneven distribution of power in the speech spectrum. Return to Figure 5-21-C for a moment. Note that the highest power is concentrated in the lower frequencies and that little power is contained in the higher frequencies. The vowel sounds, associated with the lower frequencies, are not essential to good speech articulation but the reverse is true of the high frequencies, which are associated with the consonant sounds. In addition, the higher the frequency the lower the signal to noise ratio, because of the increased power attenuation of the speech at high frequencies.

A preemphasis nework or circuit emphasizes the high frequencies and attenuates the lows; this has little effect on speech articulation, even though the "natural" quality of the voice may suffer. Preemphasis, then:

a) improves the peak to average ratio of the speech waveform by reducing it.

b) improves the signal to noise ratio by "lifting" the high frequencies above the noise level.

c) provides a more reasonable power density over the speech spectrum.

Note: Preemphasis is widely used in FM and TV broadcasting; but, since complete fidelity rather than speech articulation is the criterion with such systems, a de-emphasis network must be incorporated in the receiver to return the emphasized portions of the transmitted spectrum to their proper levels. SSB receivers do not require de-emphasis because speech articulation, rather than complete fidelity, is the criterion here.

The amount of preemphasis is usually stated in terms of "DB per OCTAVE".

The term: "OCTAVE" is usually associated with speech or music. An Octave contains twelve notes, each successive note being 1.06 times the frequency of the preceding. An Octave represents a frequency ratio of 1:2; thus, 1 Octave separates 500 cycles and 1 KC.

If 500 cycles is multiplied by 1.06, the result is 530 cycles. If 530 cycles is multiplied by 1.06, the result is 561.8 cycles. If this operation is carried out a total of 12 times, the resultant is 1 KC, one octave removed from 500 cycles. Similarly, 1 octave separates 40 and 80 cycles, 1200 and 2400 cycles, and so forth.

To say that a circuit provides a preemphasis of 6 DB per Octave means that a frequency f-2, one octave higher than another frequency f-1, will be either accentuated or amplified 6 DB above f-1. Another frequency, f-3, 1 Octave higher than f-2, will be either accentuated or amplified 6 DB above f-2 and 12 DB above f-1. This is shown in Figure 5-25.

Preemphasis circuits may either amplify or accentuate. For example, consider Figures 5-26A and B.

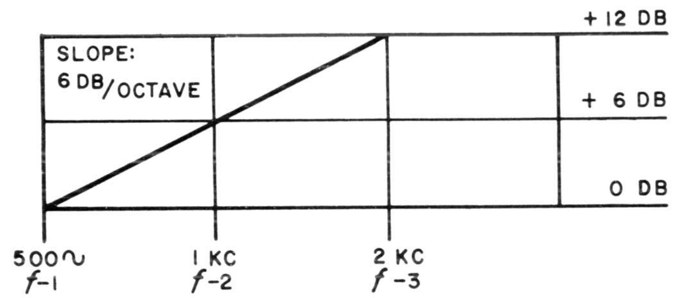


Figure 5-25.

In Figure 5-26A, the amplification increases with frequency. Since inductive reactance increases with frequency, and amplification depends on total plate impedance, the higher frequencies will offer increased amplification.

In Figure 5-26B, no amplification takes place; the preemphasis circuit is a simple coupling circuit; the values of C and R are chosen so that the coupling capacitor offers high reactance at low frequencies and decreasing reactance at higher frequencies. Thus, the high frequencies are emphasized by attenuating the low frequencies.

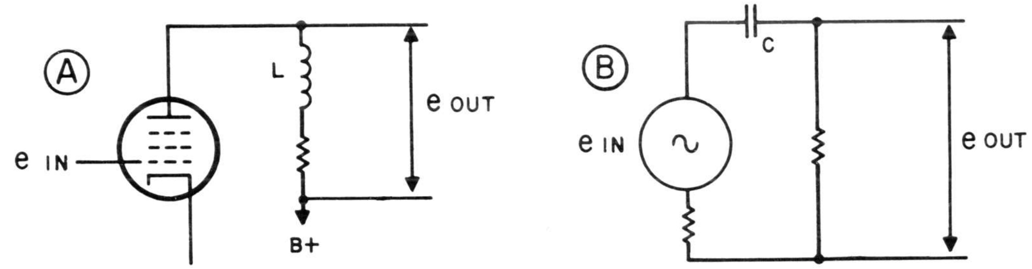
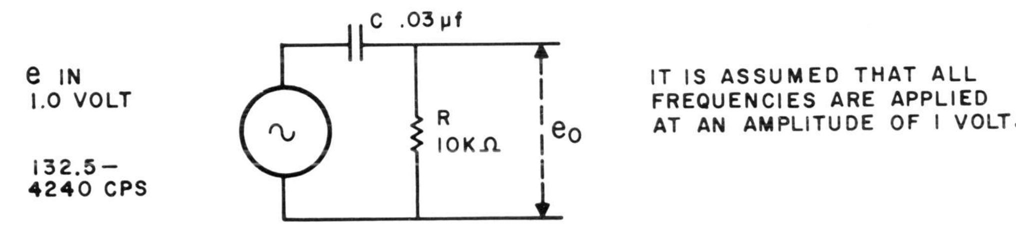


Figure 5-26.

Consider the circuit shown below:



The low frequency "roll off" occurs when the reactance of the capacitor equals the resistance. Then:

$$f = \frac{159\times10^{-3}}{CR} = \frac{159\times10^{-3}}{3\times10^{-8}\times1\times10^{4}} = \frac{159\times10^{-3}}{3\times10^{-4}} = 530 \text{ cycles}$$

$$Zt = R - j\,Xc = 10k\Omega - j\,10k\Omega = 14.14\ k\Omega\ \angle -45^\circ$$

$$Eo = \frac{E \text{ in } R}{Zt} = \frac{10k\Omega}{14.14k\Omega} = .707v$$

$$db = 20 \log \frac{e}{e} = 20 \log \frac{1.00}{.707} = -3db$$

This is the —3 db point on the graph presented in Figure 5-27. Frequencies above this frequency will be accentuated; frequencies below this frequency will be attenuated.

At one Octave below 530 cycles, (265 cycles):

$$Xc = 20k\Omega \quad Zt = 10k\Omega - j20k\Omega = 22.4k\Omega \angle -63°4$$

$$Eo = \frac{R}{Zt} = \frac{10k\Omega}{22.4k\Omega} = .447v$$

$$db = 20 \log \frac{e}{e} = 20 \log \frac{1.00}{.447} = 20 \log 2.24 = -7db$$

At one Octave above 530 cycles, (1060 cycles):

$$Xc = 5k\Omega \quad Zt = 10k\Omega - j5k\Omega = 11.2k\Omega \angle -26.°6$$

$$Eo = \frac{R}{Zt} = \frac{10k\Omega}{11.2k\Omega} = .893 \text{ volts}$$

$$db = 20 \log \frac{e}{e} = 20 \log \frac{1.00}{.893} = 20 \log 1.12 = -1db$$

At one Octave below 265 cycles (132.5 cycles):

$$Xc = 40K\Omega \quad Zt = 10K\Omega - j40K\Omega = 41K\Omega \angle -75.°9$$

$$Eo = \frac{R}{Zt} = \frac{10K\Omega}{41K\Omega} = .244 \text{ volt}$$

$$db = 20 \log \frac{1.00}{.244} = 20 \log 4.1 = -12.26 \text{ db}$$

At one Octave above 1060 cycles (2120 cycles):

$$Xc = 2.5K\Omega \quad Zt = 10K\Omega - j\ 2.5K\Omega = 10.28K\Omega \angle -14.°1$$

$$Eo = \frac{R}{Zt} = \frac{10K\Omega}{10.28K\Omega} = .974 \text{ volts}$$

$$db = 20 \log \frac{1.00}{.974} = 20 \log 1.028 = -.24db$$

At one Octave above 2120 cycles (4240 cycles):

$$Xc = 1.25K\Omega \quad Zt = 10K\Omega - j\ 1.25K\Omega = 10.05K\Omega \angle -7.°14$$

$$Eo = \frac{R}{Zt} = \frac{10K\Omega}{10.05K\Omega} = .995 \text{ volts}$$

$$db = 20 \log \frac{1.00}{.995} = 20 \log 1.005 = -.06 \text{ db}$$

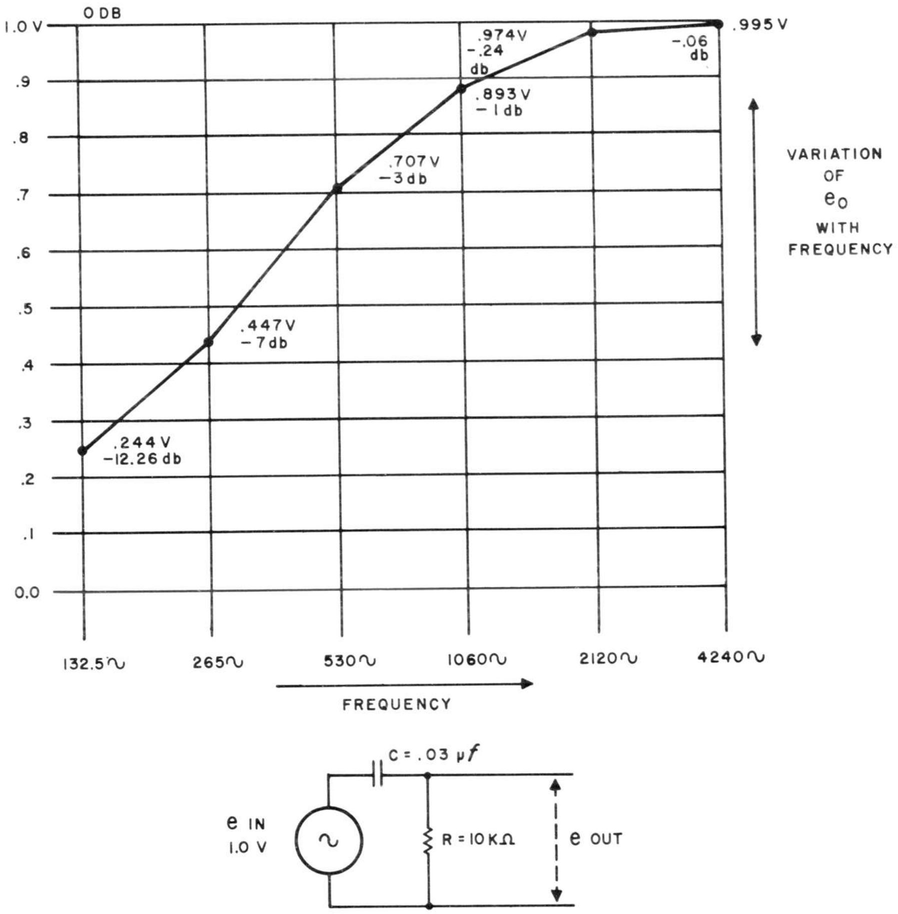


Figure 5-27.

Figure 5-27 illustrates the action of a simple RC Preemphasis circuit. It is assumed that the input voltage is a constant 1.0 volt irrespective of the frequency.

5-6 TMC Model SPU Speech Processing Unit

The TMC Speech Processing Unit Model SPU is a compact, fully transistorized device which may be used with voice modulated AM or SSB transmitters, to provide increased overall efficiency. Articulation tests have shown that this unit will provide 50 to 65% improvement of intelligibility under adverse signal to noise conditions. The unit may also be used with a receiver; in this case it operates as a constant level audio amplifier. Figure 5-28, below, shows the front panel of the Model SPU.

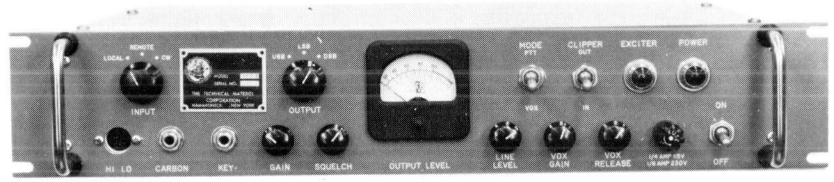


Figure 5-28.

TECHNICAL SPECIFICATIONS

Refer to Figures 5-29, 5-30 and 5-31. Figure 5-29 shows the front panel of the unit, with all controls, indicators and input jacks numbered.

Figure 5-30 shows the rear connections. Figure 5-31 is a simplified block diagram of the unit.

INPUTS:

Audio line at 0 DBM, plus or minus 20 DBM, balanced or unbalanced.

Carbon microphone at —25DB. Connections on the front panel to Western Electric type 309 ring tip sleeve plug or 6 connector microphone plug. Terminal board connections at rear. The required DC voltage is supplied internally.

High and Low impedance microphones at —55 DB. Front panel 6 connector jack, or rear terminal board connections.

CW Key input. Front panel Western Electric type 309 jack.

Squelch (anti VOX) input at rear panel.

Push to Talk input at rear panel. The PTT feature is also incorporated with the six connector microphone jack.

PREEMPHASIS:

Provides a 6 DB per Octave slope peaked at 2500 cycles.

CLIPPING:

A nominal 12 DB of clipping is provided when the CLIPPER switch is placed in the IN position.

AUDIO OUTPUT:

Two 0 DBM, 600 ohm sets of contacts at rear panel. A front panel switch allows selection of USB, LSB or DSB.

DISTORTION:

Nominal 5%.

KEYING CONTROL: (Dry Contact)

A keying relay, incorporating two sets of operating contacts, may be actuated by the Push to Talk feature, by voice operated (VOX) control, or by the INPUT switch when placed in the CW position. This feature permits an associated exciter to be enabled, or an associated receiver to be muted.

DYNAMIC RANGE:

The unit incorporates automatic level control circuitry to maintain the output within plus or minus 2 DB for input variations of 40 DB. The speed of response for increasing audio levels is at a syllabic rate. This is a most effective form of volume compression.

FREQUENCY RESPONSE:

200-3000 cycles plus or minus 1.5 DB. Rolloff, 30 DB per Octave above 3000 cycles.

METERING:

A front panel meter monitors the output levels at the 600 ohm output lines.

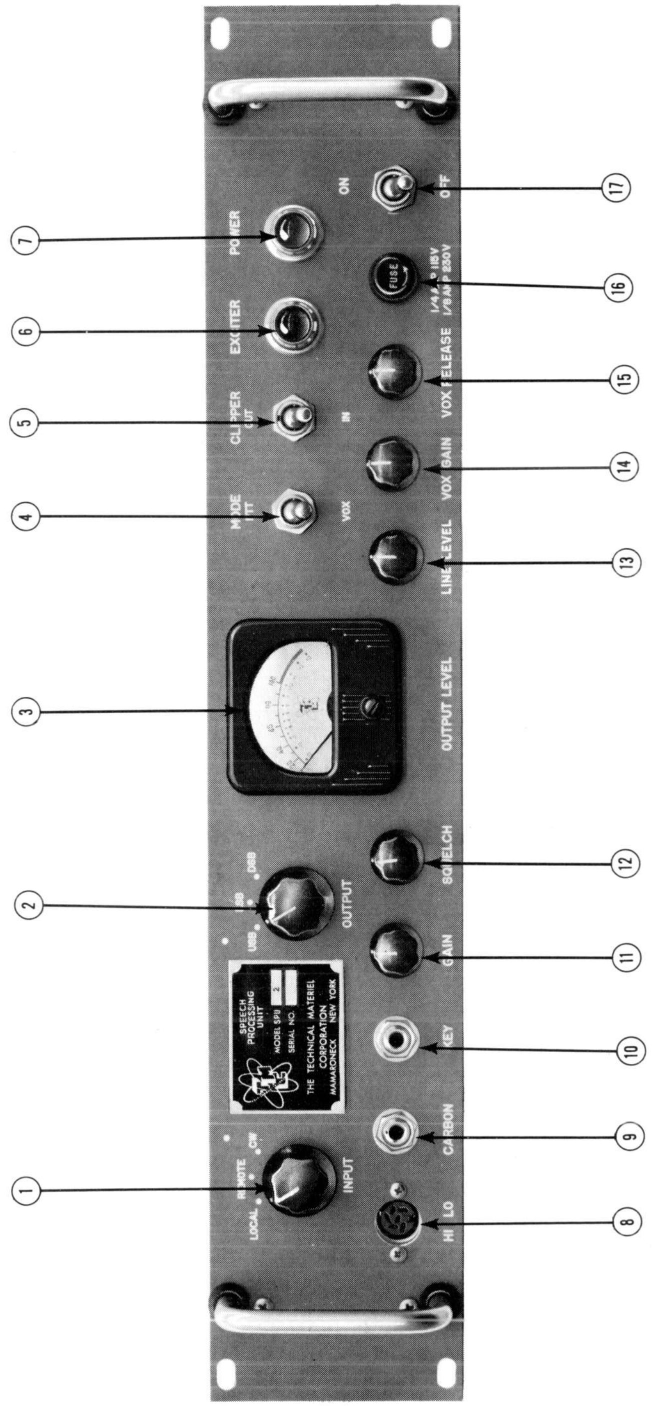


Figure 5-29. Panel View of Model SPU, showing Controls and Indicators.

DISCUSSION OF FRONT PANEL OPERATING CONTROLS, INPUTS AND INDICATORS. (See Figure 5-29)

1. INPUT switch:

In the LOCAL position, intelligence from a Carbon microphone, or from a low or high impedance microphone is selected. These sources may be connected at the front panel at jacks No. 8, No. 9, or they may be connected from external sources at the rear panel. If the source plugged into the 6 connector microphone jack has an associated ear piece, this also will be activated.

In the REMOTE position, intelligence from an external 600 ohm audio line is selected. This intelligence may be for transmission or reception.

In the CW position, the keying relay is activated. A CW key, plugged into jack No. 10, is connected to the associated transmitter via the rear panel.

2. OUTPUT switch:

In USB position, the output of the Model SPU is routed to the output terminals designated USB.

In LSB position, the output of the Model SPU is routed to the output terminals designated LSB.

In DSB position, the output of the Model SPU is routed to both USB and LSB output terminals.

3. OUTPUT LEVEL meter:

This meter is connected to the output 600 ohm line. It is calibrated in two scales: an arbitrary scale, indicating 0 to 100%, and a VU scale, calibrated from minus 20 DB to plus 3 DB.

4. MODE switch:

In the PTT position, the push to talk button of an associated microphone activates the keying relay.

In the VOX position, the keying relay is activated only in the presence of intelligence.

5. CLIPPER switch:

With this switch in the IN position, the clipper and preemphasis circuits are activated. The switch would normally be left in this position for voice transmission.

With this switch in the OUT position, the clipper and preemphasis circuits are bypassed. The switch would normally be left in this position when the SPU is used for reception.

6. EXCITER indicator:

This indicator is illuminated when the keying relay is activated by either the PTT feature, the VOX feature, or when the INPUT switch is in the CW position.

7. POWER indicator:

This indicator is illuminated when the POWER ON OFF switch, No. 17, is placed in the ON position.

8. HI-LO jack:

This is the six connector jack for use with a high or low impedance microphone and associated earpiece.

9. CARBON microphone jack:

This jack is designed to receive a carbon microphone.

10. KEY jack:

This jack is used for insertion of a telegraph key, when it is desired to employ the CW mode with an associated transmitter.

11. GAIN control:

This is a potentiometer control inserted in the amplifier chain to control the gain during transmission or reception.

12. SQUELCH control:

This is a potentiometer control, inserted in the squelch (anti VOX) input circuit; the input voltage arrives from the audio section of an associated receiver. The squelch circuit prevents this associated receiver from triggering the VOX control circuit. The SQUELCH control adjusts the magnitude of the audio signal from that receiver.

13. LINE LEVEL control:

This is a potentiometer volume control inserted in the final audio output stage; it regulates the output level as indicated on the OUTPUT LEVEL meter, No. 3.

14. VOX GAIN control:

This control regulates the sensitivity of the voice control circuit.

15. VOX RELEASE control:

This control regulates the rapidity with which the keying relay de-activates after a cessation of voice intelligence.

16. PRIMARY FUSE HOLDER:

This contains a 115 volt, ¼ ampere fuse, or a 230 volt, ⅛ ampere fuse, depending on the primary power connection. This fuse is connected in the primary circuit of the power supply.

17. POWER ON OFF switch:

This is a double pole, single throw toggle switch, which connects or removes primary power.

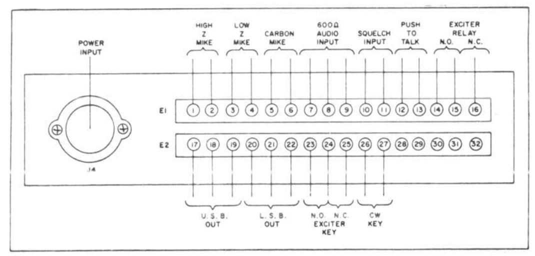


Figure 5-30. Diagram Showing Rear Panel Connections, SPU.

DISCUSSION OF THE SIMPLIFIED BLOCK DIAGRAM, Model SPU

a) *Input Circuits:*

Connections for a low impedance microphone, a high impedance microphone, a carbon microphone, a 600 ohm audio line and a push to talk feature may be brought in at the rear panel. The 600 ohm line may be connected to a source for transmission, or it may be connected to the output of a receiving system for reception.

Connection may be made at the front panel at J-1 for the high or low impedance microphone and at J-2 for a carbon microphone. The six connector jack, J-1, has provisions for an earpiece.

The high impedance microphone input is applied to preamplifier stage Q-2 via T-1, an impedance matching transformer. The low impedance microphone input connects directly to the base of Q-2.

The carbon microphone connects to the base of Q-2 via C-1, R-2. This connection is necessary because of the polarizing DC voltage used with the carbon microphone and because less amplification is required.

b) *Input Switch S-1:*

With the INPUT switch, S-1, in the LOCAL position, as shown, the signal impressed on the base of Q-2 is also applied to the earpiece, for local monitoring. The output of preamplifier Q-1 is applied to Q-2, an emitter follower stage, via GAIN control R-8.

With the INPUT switch in position No. 2, REMOTE, the intelligence on the 600 ohm audio line is coupled via T-5 and S-1 to the earpiece terminal of J-1 and to the base of emitter follower Q-2.

With the INPUT switch in position No. 3, CW:

(1) the earpiece connection at J-1 is grounded.

(2) the audio amplifiers receive no input.

(3) a ground is applied to keying relay K-1 to energize it.

c) *The VOX SQUELCH circuits:*

The output of emitter follower stage Q-2 is applied to the VOX SQUELCH circuits. Q-3 and Q-4 are VOX amplifiers; their output is controlled by the strength of the incoming signal and by the setting of the VOX GAIN control. This output is rectified; the rectifier produces a negative voltage, which is applied to the base of Q-5.

A SQUELCH input from an associated receiver circuit is applied, via the SQUELCH control, to a rectifier circuit which produces a positive voltage; this positive voltage is applied to the base of Q-5.

Thus, the VOX circuit attempts to actuate relay K-1 via the action of Q-6, and the SQUELCH circuit attempts to de-activate it.

Q-5 contains the VOX DECAY control; this changes an RC time constant, which determines the time required for K-1 to de-energize after removal of audio intelligence.

Relay K-1 may also be energized in two other ways:

(1) with the MODE switch, S-4, in the PTT position, a ground is applied to the base of Q-6, nullifying the action of the VOX circuits, and the push to talk button activates K-1.

(2) with the INPUT switch, S-1, in the CW position, a ground is applied to the low side of K-1, causing it to energize.

d) *The Audio Amplifier Chain:*

The output of preamplifier Q-2 is also applied to a push pull amplifier stage, Q-7, Q-8. The push pull stage incorporates a clipper circuit, CR-5, CR-6. The signal is further amplified in Q-9; emitter follower stage Q-10 sends the signal to the CLIPPER switch, S-2.

With the CLIPPER switch in the OUT position as shown, the signal bypasses the Preemphasis and clipping networks, and is applied to emitter follower stage Q-15. This stage is followed by Z-1, an audio low pass filter which severely attenuates frequencies above 3000 cycles.

With the CLIPPER switch in the IN position, the audio from emitter follower Q-10 is processed in a preemphasis network and a clipping circuit. The CLIPPER switch would normally be left in the IN position when the Model SPU-2 is being used to process a voice signal for transmission.

Q-16 is the power amplifier, the output of which is controlled by the LINE LEVEL control. The audio output amplitude may be monitored on the front panel LINE LEVEL meter. The OUTPUT switch selects USB, LSB, or DSB operation.

e) *The AGC Arrangement:*

The output of emitter follower Q-10 is also applied to the AGC circuits, which determine the dynamic range of the unit. These circuits might also be termed: "Volume Compression" circuits.

Q-11 and Q-12 are AGC amplifiers; Q-13 and Q14 are temperature compensators. The AGC output of Q-14 is applied to the push pull amplifier circuit, Q-7 and Q-8, to keep the gain of the circuits reasonably constant over a wide range of input amplitudes.

CHAPTER 6

DISTORTION IN SSB TRANSMITTERS; SIGNIFICANT TESTS OF TRANSMITTERS

6-1 Introductory Note

We have thus far assumed that balanced modulator, mixer, linear voltage amplifier and linear power amplifier circuits have performed their functions ideally, that is, without distortion.

It is true that distortion in modern SSB transmitters can be held to a low value, but this is the result of careful design, alignment and operation.

In this chapter distortion in SSB transmitters will be discussed. New terms will be introduced and a method described for determining the magnitude of the most objectionable form of distortion in the output. Transmitter frequency stability and tests concerning the measurement of peak envelope output power will be described.

6-2 Spurious Mixer Products

Consider the simple mixer circuit shown in Figure 6-1.

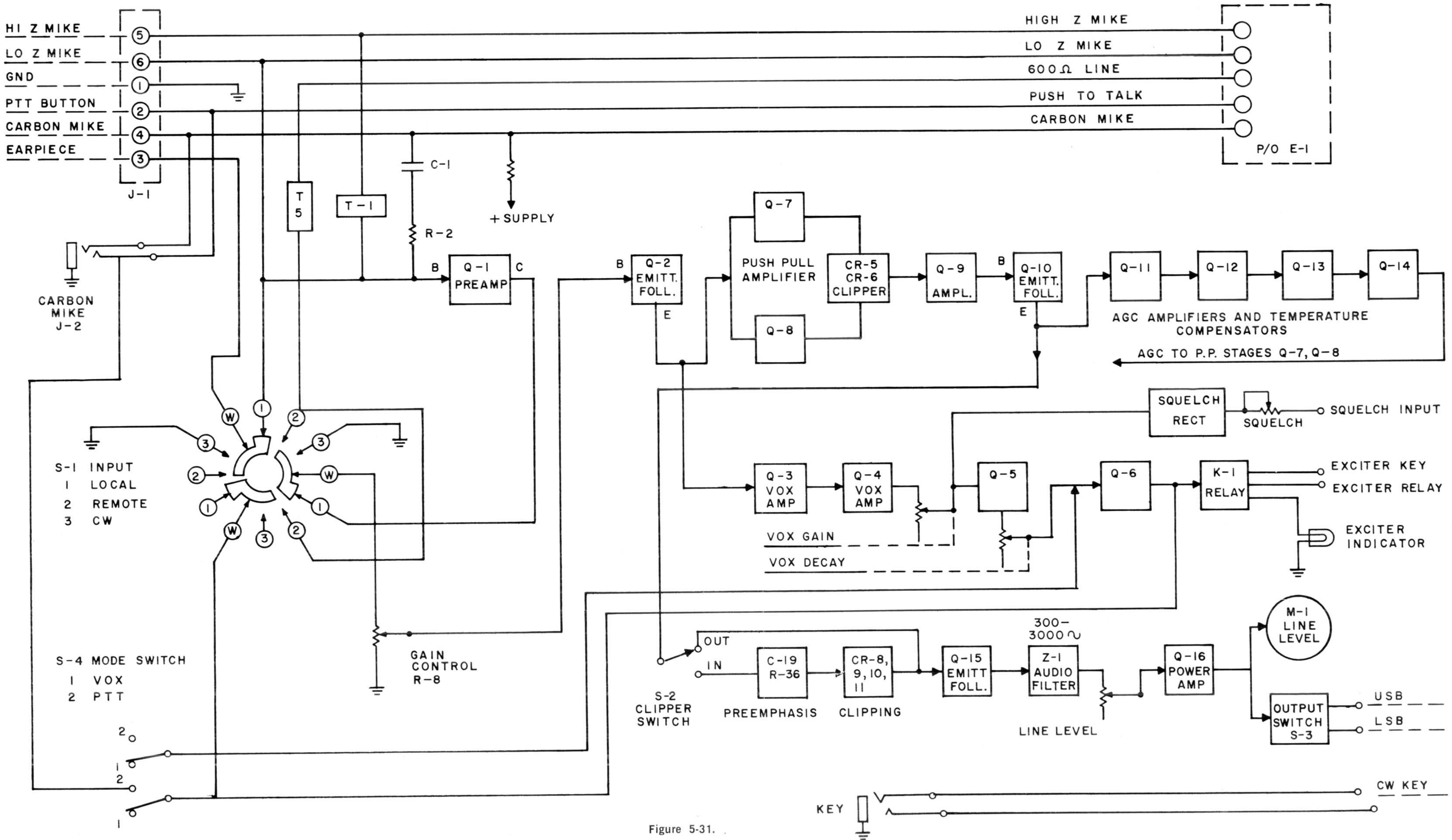

HI Z MIKE
LO Z MIKE
GND
PTT BUTTON
CARBON MIKE
EARPIECE
J-1
HIGH Z MIKE
LO Z MIKE
600Ω LINE
PUSH TO TALK
CARBON MIKE
P/O E-1
C-1
R-2
T 5
T-1
+SUPPLY
CARBON MIKE J-2
Q-1 PREAMP
Q-2 EMITT. FOLL.
Q-7
PUSH PULL AMPLIFIER
Q-8
CR-5 CR-6 CLIPPER
Q-9 AMPL.
Q-10 EMITT. FOLL.
Q-11
Q-12
Q-13
Q-14
AGC AMPLIFIERS AND TEMPERATURE COMPENSATORS
AGC TO P.P. STAGES Q-7, Q-8
SQUELCH RECT
SQUELCH
SQUELCH INPUT
S-1 INPUT
1 LOCAL
2 REMOTE
3 CW
Q-3 VOX AMP
Q-4 VOX AMP
Q-5
Q-6
K-1 RELAY
EXCITER KEY
EXCITER RELAY
EXCITER INDICATOR
VOX GAIN
VOX DECAY
GAIN CONTROL R-8
S-4 MODE SWITCH
1 VOX
2 PTT
OUT
IN
S-2 CLIPPER SWITCH
C-19 R-36
PREEMPHASIS
CR-8, 9, 10, 11
CLIPPING
Q-15 EMITT FOLL.
300–3000∿
Z-1 AUDIO FILTER
LINE LEVEL
Q-16 POWER AMP
M-1 LINE LEVEL
OUTPUT SWITCH S-3
USB
LSB
KEY
CW KEY

Figure 5-31.

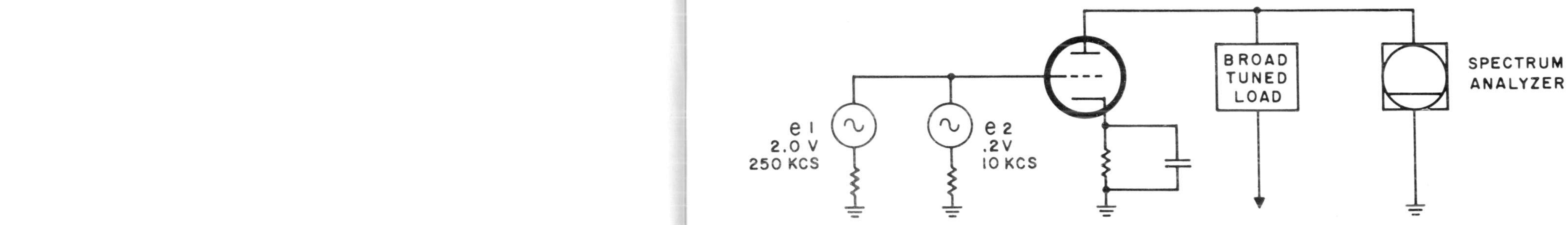


Figure 6-1.

Both signals are applied to the control grid. Normally, the output of a mixer or balanced modulator circuit is sharply tuned to either the sum or difference frequency spectrum, but, in this case, the plate impedance is a broadly tuned circuit, in order to examine the full output spectrum on the attached Panoramic Analyzer.

Mixer circuits must be operated with a certain amount of nonlinearity, since the second order sum and/or difference products are desired in the output. It is virtually impossible to design such a circuit without introducing factors which will produce higher order products. Even a Class A voltage amplifier, operated carefully, will produce distortion because the plate current characteristic is not absolutely linear.

The Spectrum Analyzer connected as shown in Figure 6-1 might show outputs at:

a) 250 KC, 500 KC, 750 KC and 1 MC. These frequencies represent the original 250 KC injection frequency and the 2nd, 3rd and 4th harmonics.

b) 10 KC, 20 KC, 30 KC and 40 KC. These frequencies represent the original 10 KC signal frequency and the 2nd, 3rd and 4th harmonics.

c) 240 KC and 260 KC, the sum and difference frequencies.

d) harmonics of the 240 KC difference frequency, at 480 KC, 720 KC, etc.

e) harmonics of the 260 KC sum frequency, at 520 KC, 780 KC, etc.

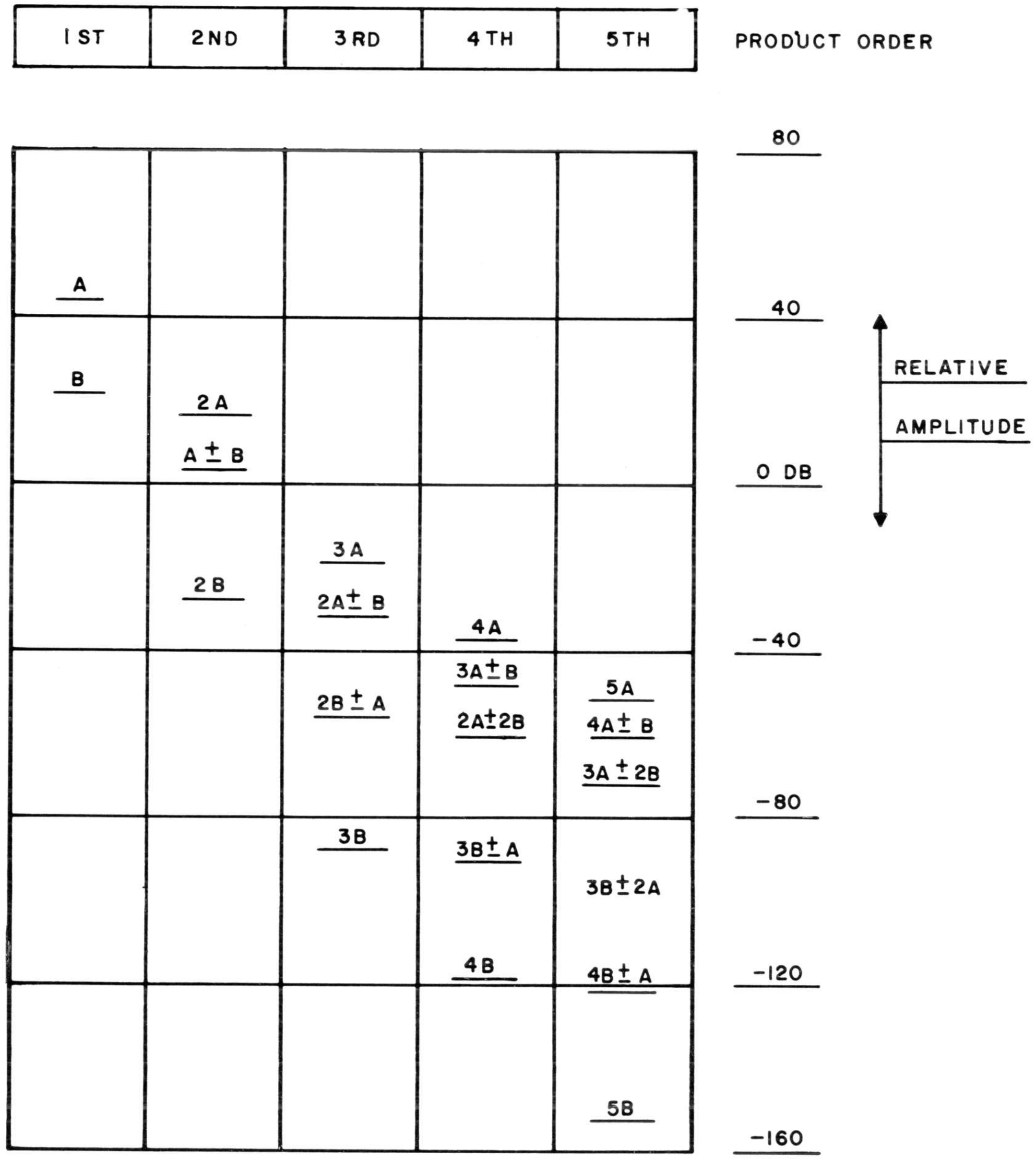
f) many additional frequencies resulting from beats between pairs of frequencies listed in (a), (b), (c), (d) and (e) above.

Thus, with only two discrete frequency inputs, a great number of new frequencies are created; these new frequencies appear with wide amplitude variations. For example, the original 250 KC has the largest output; following is the original 10 KC input. Even the second harmonic of 250 KC may appear at a greater amplitude than the desired sum and difference frequencies. Generally, amplitude decreases as the product order increases.

Table 6-1, on the following page, shows a plot of the significant output frequencies and their relative amplitudes from the simple mixer circuit just described.

The 250 KC injection frequency will be identified as "A".

The 10 KC signal frequency will be identified as "B".



A = 250 KC INJECTION

B = 10 KC INJECTION

TABLE 6-1. A PLOT OF THE RELATIVE AMPLITUDES OF THE SIGNIFICANT OUTPUT FREQUENCIES FROM THE MIXER CIRCUIT SHOWN IN FIGURE 6-1.

Zero DB reference is taken as the amplitude of the normally desired sum and difference frequencies, A ± B.

6-3 Examination of the Spectrum of a Practical Balanced Modulator Circuit

Consider the circuit shown in Figure 6-2.

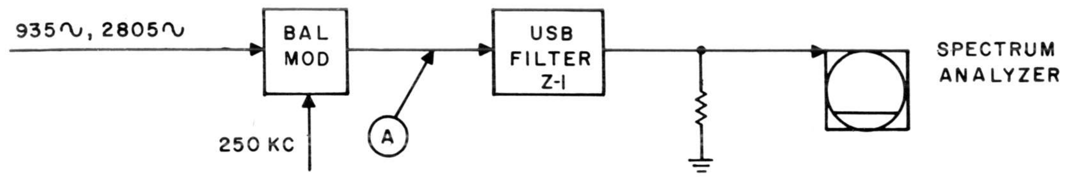


Figure 6-2.

Two discrete frequencies, at 935 cycles and 2805 cycles, are applied to a well designed balanced modulator with a carrier injection frequency of 250 KC. The balanced modulator is followed by a highly selective USB filter, Z-1, which will pass only frequencies in the range 250,300 cycles to 257,500 cycles. The output of this arrangement is monitored on a Spectrum Analyzer. The balanced modulator will greatly attenuate the 250 KC carrier frequency, and, consequently, its harmonics. The filter, Z-1, will further attenuate the carrier. The harmonics of the audio input frequencies are far removed from the filter bandpass characteristics, hence, these may be discounted.

For all practical purposes, only the sum frequencies, 250,935 cycles, and 252,805 cycles, with a very small residual carrier, should appear at the output. This *ideal* condition is shown in Figure 6-3.

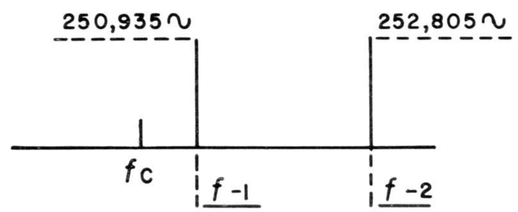


Figure 6-3.

Note that the sum frequency 250,935 cycles, has been designated as f-1 and that 252,805 cycles has been designated as f-2.

The Spectrum Analyzer, however, shows a pattern differing from the ideal. The pattern actually observed is shown in Figure 6-4, below:

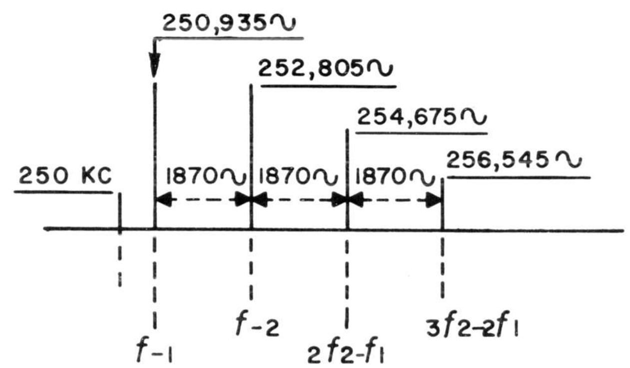


Figure 6-4.

The additional outputs are the *odd order products,* which fall within the passband. These odd order products result from beats as follows:

a) the second harmonic of f2 beats with f1, to produce (2f2-f1).

	2f2	505,610	
	f1	— 250,935	
2f2 — f1		254,675	This is the 3rd order product.

b) the third harmonic of f2 beats with the second harmonic of f1, to produce (3f2 — 2f1).

	3f2	758,415	
	2f1	— 501,870	
3f1 — 2f2		256,545	This is the 5th order product.

c) the fourth harmonic of f2 beats with the 3rd harmonic of f1, to produce (4f2 — 3f1)

	4f2	1,011,220	
	3f1	— 752,805	
4f2 — 3 f1		258,415	This is the 7th order product;

in this particular case, it is outside the bandpass.

d) this procedure could be carried on to show additional odd order products. For example, the 9th order product would be (5f2 — 4f1). Note, on Figure 6-4, that the frequency separation is always the difference between f1 and f2.

e) as a corollary, odd order products also exist on the other side of the carrier for the harmonics of f1 beating with f2 and its harmonics.

For example:

2f1 — f2	2f1	501,870	
	f2	— 252,805	
		249,065	This is the other 3rd order product.

similarly, the 5th order product is 3f1 — 2f2
the 7th order product is 4f1 — 3f2
the 9th order product is 5f1 — 4f2

In this particular case, because of the selective filter Z-1, these frequencies fall outside the bandpass, and will not be observed on the Spectrum Analyzer. They appear at point "A", on Figure 6-2, however, because a double sideband signal exists here.

When the output of the circuit of Figure 6-2 is translated to a higher frequency in another mixer or balanced modulator, the filtering will most probably be accomplished with tuned circuits. Depending on the selectivity, it is likely that odd order products will be observed on both sides of the two tones out to the 5th or 7th order.

SUMMARY

Many new frequencies are created in a balanced modulator. Most of these are easily eliminated by tuned circuits and filters. The most objectionable form of distortion occurring in a SSB transmitting system is the distortion due to odd order products. This distortion falls within the passband and adjacent spectrum. Once created, it cannot be eliminated.

Fortunately, in the exciter section of a well designed system, the odd order products can be held down to a very low value. For example, in a rectifier balanced modulator the carrier can be held down to at least —40 DB and the 3rd order products to about —50 DB, with corresponding higher attenuation of the other odd order products. By using linear voltage amplifiers and carefully operated linear power amplifiers, the amplitude of third order products in the transmitter final output can be held to about —40 DB at full power, with resulting higher attenuation of the other odd order products. These odd order products are often termed: INTERMODULATION PRODUCTS.

6-4 Undesired Transmitter Output

Generally speaking, there are two kinds of undesired power responses in the output of a transmitter:

a) spurious responses outside the passband, to which the transmitter is ostensibly tuned. These are mainly harmonics.

b) intermodulation, incidental amplitude and phase modulation products in or near the passband.

Most of this distortion can be minimized or eliminated except for the intermodulation distortion, once it is generated. The principal objection to this odd order product distortion is not that it appreciably degrades the signal in the passband, but that it splatters on both sides of the passband to interfere with adjacent channels.

The final tuned tank of a transmitter usually has a high unloaded "Q" and a relatively low loaded "Q", usually about 12. This increases the effective bandwidth. Even though the actual intelligence spectrum may occupy only 3 KC or 7 KC, there may be appreciable odd order product splatter outside the intelligence passband (and inside it), to interfere seriously with communications on adjacent channels. Recall, for a moment, one of the principal advantages of SSB: "AN SSB SIGNAL REQUIRES ONE HALF THE BANDWIDTH OF A COMPARATIVE AM SIGNAL". Appreciable splatter can nullify this advantage.

Return, for a moment, to Figure 6-4. An appreciable number of odd order products can be generated with only *two tones* applied. Consider, for a moment, the number of odd order products that may be generated with a *full speech spectrum* is fed into the system.

The incidental amplitude and phase modulation results from imperfect stabilization of the master oscillator or synthesizer system from which translation injection frequencies are generated. This type of distortion can be minimized to the point of non existence with modern frequency and control synthesizers.

REJECTION OF THE UNWANTED SIDEBAND

It has been assumed thus far that the undesired sideband has been completely attenuated. Such attenuation will be a function, principally, of the quality of the sideband filters employed and the selectivity of the system. Unwanted sideband suppression in a well designed system can be considered complete, for all practical purposes. For example. a transmitter might have the following specification:

"UNDESIRED SIDEBAND REJECTION: A 500 CYCLE TONE IN THE DESIRED SIDEBAND IS DOWN, IN THE UNDESIRED SIDEBAND, 60 DB FROM FULL PEP OUTPUT".

If: $\quad DB = 10 \log P/P$

Then: $\quad \frac{60}{10} = \log P/P$

$$6 = \log P/P$$

$$P/P = 1 \times 10^6$$

Assume that full PEP output of the transmitter is 200 KW; then the tone in the unwanted sideband is:

$$\frac{200{,}000}{1{,}000{,}000} = 200 \text{ milliwatts.}$$

This is shown in Figure 6-5.

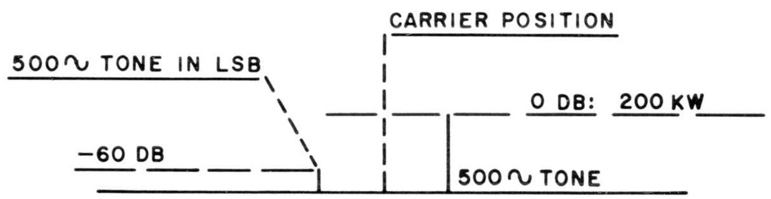


Figure 6-5.

DISTORTION AS RELATED TO THE CAPABILITY OF THE HIGH VOLTAGE POWER SUPPLY

Usually, the high voltage power supply for the plate of the final linear power amplifier is also used, with regulator circuits, for the final screen voltage. Tetrode linear power amplifiers require "stiff" screen regulation; in addition, the plate supply must be able to furnish wide variations of power, from "no signal" to "full PEP" condition.

Suppose that a single tone is applied through the entire system, and that levels are adjusted so that this tone is at, say 10 DB below the rated PEP output of the transmitter.

Suppose, now, that a second tone is introduced, at a level corresponding to rated PEP. If the power supply is extremely "healthy", that is, if it has been designed for such a contingency, and if the power amplifier tube is conservatively rated, there will be no noticeable "compression" of the first tone. If, however, the power supply cannot deliver the required power, the first tone will be reduced in amplitude or "compressed', and the second tone will probably be "clipped".

Compression is used in this sense to describe a characteristic of a particular power amplifier and its associated high voltage power supply. It is actually a measure of the variation in power supply output voltage for conditions of greater than normal loading.

Consider now a linear power amplifier and power supply which will sustain a rated 10 KW PEP, but which, being conservatively rated, will actually handle a maximum of 15 KW PEP without serious degradation.

Suppose that two equal amplitude tones are applied through the system.

Before proceeding, let us reestablish an important fact:

PEAK ENVELOPE POWER IS TWO TIMES AVERAGE POWER WHEN TWO TONES ARE APPLIED. THIS RELATIONSHIP WILL NOT HOLD TRUE UNDER OTHER CONDITIONS. REFER TO TABLE 5-1 or FIGURE 5-19. THE TWO TONE TEST IS USED EXTENSIVELY IN SSB SYSTEMS BECAUSE:

a) THE RELATIONSHIP BETWEEN PEP AND AVERAGE POWER IS DEFINITE FOR TWO TONES.
b) IF THE TWO TONES HAVE A 3:1 to 5:1 RATIO, THEY CAN BE USED TO DETERMINE THE EXTENT OF INTERMODULATION DISTORTION.

Let us increase the transmitter drive until the 15 KW PEP point is exceeded. This will correspond to 7.5 KW average power. Beyond this point, the tones will be clipped; that is, the extremities of the waveform will be flattened. A great many new frequencies will be generated. The final tuned circuits will eliminate all but the intermodulation products. In band distortion will increase, and splatter on either side of the passband will become objectionable.

DISTORTION CAUSED BY IMPROPER LINEAR AMPLIFIER OPERATION

Linear power amplifiers will be discussed in another chapter; however, an important point regarding the operation of these devices will be taken up here while the subject of distortion is being examined.

Linear power amplifiers are usually of the triode or tetrode types, operated Class AB-1 or AB-2. They may be grid driven or cathode driven. In any event the operating point on the eg ip transfer characteristic must be selected with great care, often for a *particular* tube. In addition, other operating conditions must be carefully controlled. These include:

a) neutralization circuits.
b) degeneration circuits.
c) regulated screen voltage circuits.
d) input and output impedance conditions. These should be designed for optimum constancy.

Obviously, if the circuits concerned with the operation of the final linear amplifier are not rigidly designed and controlled, and, in addition, *carefully monitored,* excessive distortion may be introduced.

6-5 SSB Transmitter Monitoring and Testing

It has already been determined that:

a) there is no convenient formula to convert average power to peak envelope power for random intelligence.

b) a transmitter operator has no *positive* assurance that he is transmitting the maximum possible power with the least amount of distortion using the panel meter indicators only.

c) an operator cannot "see" splatter on the panel meters.

The number and scope of the panel indicating devices incorporated in a transmitter will usually depend on the power capability of the unit and the service for which it is designed.

A radio amateur with more ingenuity than funds operating SSB voice on one or more bands will "make do" with the minimum of test and monitoring facilities. A military service station with requirements for full HF coverage, high power, stability and versatility of transmitted intelligence will employ every possible means to keep the output "clean". This is especially true when Independent Sideband transmission is utilized, with many independent multiplexed channels. The point to be made is this: a facility or agency which spends many thousands of dollars for a modern SSB transmitter should expend an additional few thousands by procuring at least a minimum of adequate test and monitoring equipment.

SIGNIFICANT TRANSMITTER TESTS

The complete test and alignment procedure for a SSB transmitting system encompasses much more detail than will be discussed here. The following paragraphs refer to significant tests frequently made by operator or maintenance personnel.

When a transmitter is tuned and put "on the air", the radiated signal has three characteristics of the utmost importance. They are:

a) the carrier or assigned frequency; its accuracy and long term stability.

b) the output power; that is, the peak envelope power for a SSB system.

c) the amplitude and frequencies of unwanted radiations; that is, the amount and spectrum placement of any radiation not concerned with the carrier or the original intelligence. This includes harmonics, odd order products in the passband, parasitics, modulation products outside the passband, etc.

THE CARRIER FREQUENCY

As previously stated, the carrier frequency is an assigned frequency which acts as a reference for an intelligence signal in the frequency spectrum. It is an infinitely thin position in the spectrum corresponding to the absolute position of the carrier frequency whether the carrier is present or not. Certain authorities define "assigned frequency" as the frequency corresponding to the center frequency of the sideband being utilized. For

example, a 3 KC USB spectrum might contain an "assigned frequency" 1.5 KC above the nominal carrier frequency. This is shown in Figure 6-6.

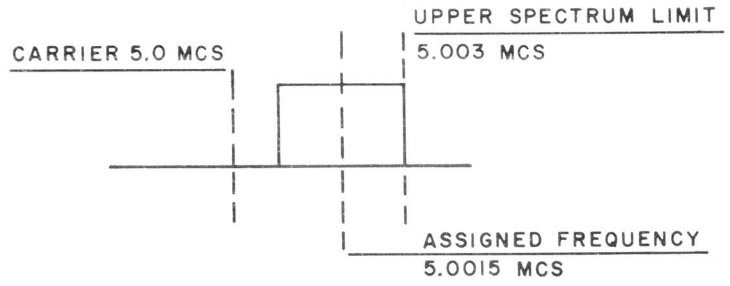


Figure 6-6.

For Independent sideband operation, the carrier frequency and the assigned frequency would be the same, since both are at the center of the total bandpass. The idea of the "assigned frequency" will require further clarification, since different manufacturers employ different bandpasses. TMC currently supplies exciters and transmitters which provide a response of either 3 KC or 7.5 KC per sideband, for a total bandpass of either 6 KC or 15 KC for ISB or AM operation.

The author suggests that readers carefully consult all pertinent directives relating to frequency assignment to determine, for their particular organization, exactly what is meant by "assigned frequency". In the absence of any information to the contrary, the carrier frequency should be taken as the assigned frequency. In this publication, the carrier frequency and the assigned frequency are the same.

STABILITY OF THE TRANSMITTED FREQUENCY SPECTRUM

There are two general types of frequency control. For want of better names, these will be termed:

a) Precision Frequency Control.

b) Non Precision Frequency Control.

These systems are sometimes referred to as synthesized or non-synthesized systems, respectively. This terminology is not always accurate. Precision frequency control is often, but not always, accomplished by means of frequency synthesizers. In addition, some systems employing frequency synthesis cannot be classified in the modern sense, as precision controlled systems. The terms: Precision and Non Precision are themselves relative. For example, in current practice, if a 20 mc carrier drifts plus and minus 20 cycles, the frequency control would be considered "non precision". However, this is "precision" control when compared with systems of ten or fifteen years ago.

To illustrate the difference between the two systems, two TMC equipments will be examined. One, the Model SBE-3 Sideband Exciter, is an equipment considered to have non precision control. The other, the

Model SBG, Sideband Generator, is a good example of a precision controlled system.

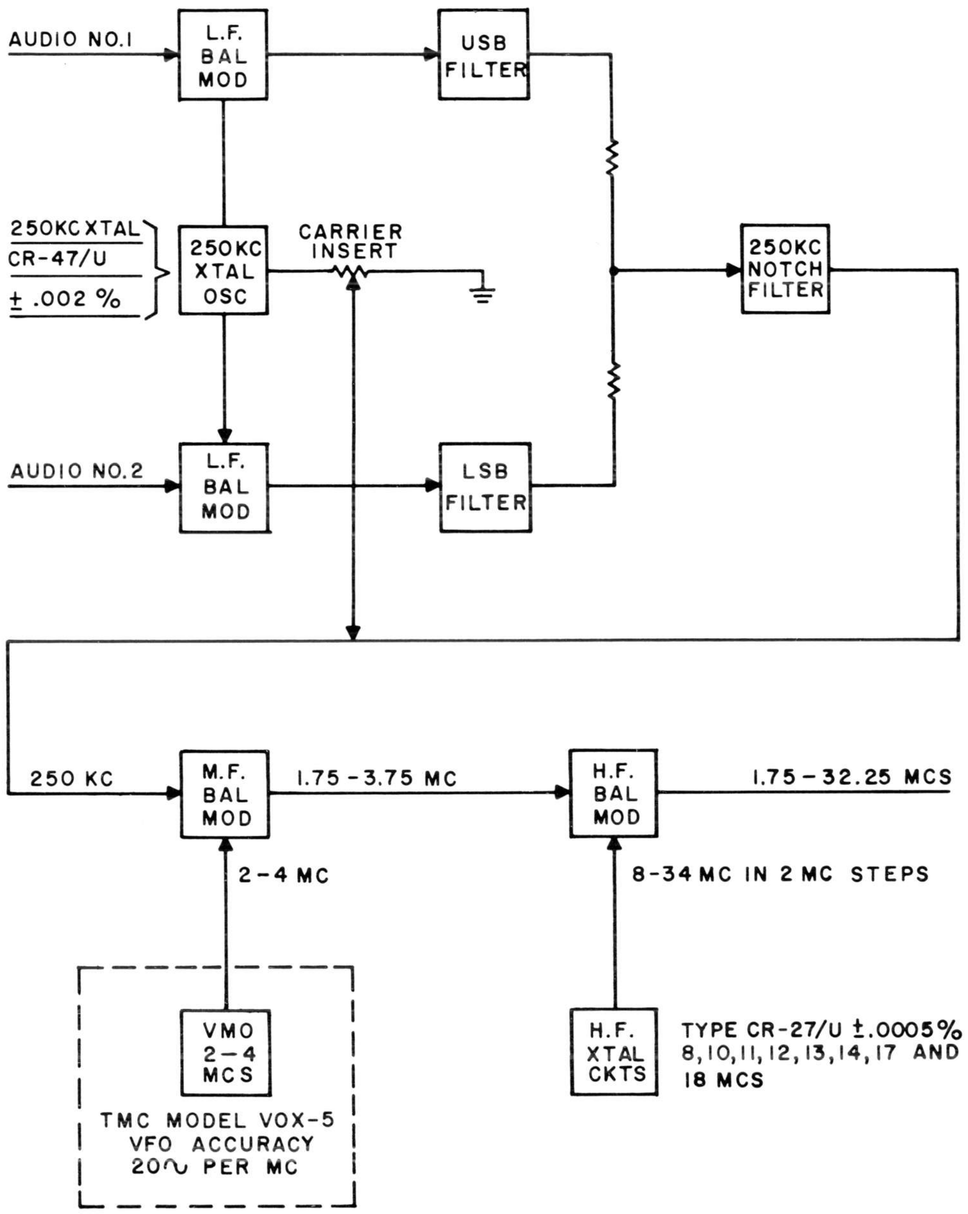


Figure 6-7. Simplified Block Diagram of the TMC Model SBE Sideband Exciter, showing only Essential Elements.

Refer to Figure 6-7. Assume that audio signals are applied to each low frequency balanced modulator. Both low frequency balanced modulators are fed with a 250 KC injection frequency supplied by an independent 250 KC crystal oscillator. The accuracy of this crystal is plus or minus .002%. This represents a maximum error of plus or minus 5 cycles.

Assuming a maximum error in the positive direction, the actual injection is 250,005 cycles.

The carrier frequency presented to the medium frequency balanced modulator, then, is 250,005 cycles.

The medium frequency balanced modulator is fed with an injection frequency in the range 2 to 4 mcs. This is usually furnished by a TMC Model VOX, Variable Frequency Oscillator. The accuracy of this device is plus or minus 20 cycles per megacycle in the basic range of 2 to 4 mcs.

Assume that a medium frequency of 3.0 mcs is desired. Then the VMO will be adjusted for a frequency of 3.25 mcs, that is, 250 KC above the medium frequency. The maximum error here is 3.25×20, or 65 cycles. Assume that this error is in the maximum negative direction. The VMO injection frequency, then, is:

$$\begin{array}{rr} & 3{,}250{,}000 \\ - & 65 \\ \hline & 3{,}249{,}935 \end{array} \text{ cycles.}$$

The high frequency balanced modulator is supplied with an injection frequency in the range 8 - 34 mcs, in two mc steps. The basic frequencies are shown; these are fed straight through, or doubled or tripled, as required. The high frequency injection is above the desired output frequency, because the output of the high frequency balanced modulator is a difference frequency.

Thus, if an output frequency of 27 mcs is desired, the high frequency injection is 30 mcs. The basic crystal oscillator frequency is 10 mcs, and this is tripled to provide the 30 mc frequency. The accuracy of the high frequency crystal is .0005%, or 5×10^{-6}.

At 10 mcs, the maximum error is 50 cycles. Due to tripling, this error is increased to 150 cycles. Assume that in this case the error is subtractive. The actual high frequency injection, then, is:

$$\begin{array}{rr} & 30{,}000{,}000 \\ - & 150 \\ \hline & 29{,}999{,}850 \end{array} \text{ cycles.}$$

The output "carrier", then, is:

$$\begin{array}{rr} & 29{,}999{,}850 \\ - & 2{,}999{,}930 \\ \hline & 26{,}999{,}920 \end{array} \text{ cycles.}$$

This is 80 cycles removed from 27 mcs.

This represents an error of $\frac{80}{27{,}000{,}000}$ or .003%.

This is within the allowed tolerance for the high frequency spectrum.

The above example has been presented to show a typical case of non precision frequency control. It is actually an extreme case; the frequency control of the Model SBE-3 is actually better. The 250 KC, 2-4 mc, and 8-34 mcs circuits are controlled by thermostatic ovens, and every precaution is taken to keep the frequency constant. Even so, the facts remain that:

a) the three injection frequencies are independent of each other, and the frequency drift will be random.
b) the amount and direction of frequency drift of each oscillator is independent.

Now refer to Figure 6-8, which shows only the essential elements of the TMC Model SBG Sideband Generator, which may be considered a good example of precision frequency control.

a) the heart of the system is a 1 megacycle standard, which supplies a frequency of 1 mc, with an accuracy of 1 part per 100,000,000 per day, to:
   (1) the VMO control circuits.
   (2) the HF control circuits.
   (3) a regenerative divider circuit.
b) The VMO Synthesizer control circuits are variable in steps of 100 cycles; these are applied to the variable master oscillator, with a range of 2-4 mcs. The output of the VMO, is a frequency in the range 2 - 4 mcs, in 100 cycle steps, *locked to the 1 mc standard.*
c) The regenerative divider is a precision countdown circuit, which produces a 250 KC frequency, locked to the 1 mc standard. This is applied to a sideband exciter where it is used as a subcarrier. The output of the sideband exciter is a nominal 250 KC frequency, which may be in the ISB, SSB, or AM mode. This output is applied to the medium frequency balanced modulator.
d) The second input to the medium frequency balanced modulator is the VMO frequency in the range 2-4 mcs. The medium frequency is in the range 1.75 - 3.75 mcs, the difference frequency of the two inputs. Since both inputs are based on the 1 mc standard, the medium frequency may be considered locked to the 1 mc standard.
e) the high frequency control circuits receive the signal from the 1 mc standard, and a sample of the output of the high frequency injection oscillator circuits. These circuits send a correction voltage to the high frequency oscillator circuits, to maintain the oscillator at the proper frequency.
f) the high frequency balanced modulator receives the medium frequency and the stable high frequency injection, in the range 10 mcs - 30 mcs, locked to the 1 mc standard.
g) the output of the high frequency circuits is a frequency in the range 1.75 - 33.75 mcs, in 100 cycle steps, locked to the 1 mc standard.
h) three synchronization indicators are included in the system:
   (1) SYNC indicator #1 is a small monitor oscilloscope, which indicates proper operation of the VMO control circuits.
   (2) SYNC indicator #2 is a neon indicator in conjunction with a front panel meter on the variable master oscillator, which indicates the proper operation of the VMO injection circuits.

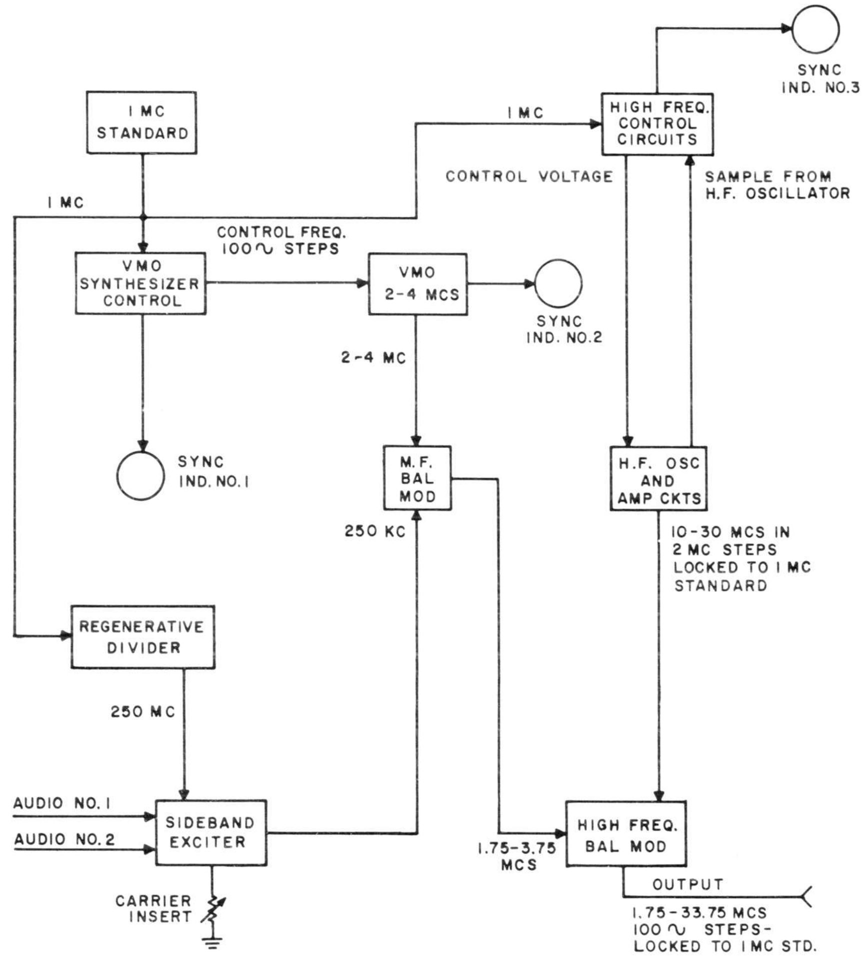


Figure 6-8. Essential Elements of the TMC Model SBG Precision Frequency Controlled Sideband Generator.

   (3) SYN indicator #3 is a neon indicator, which shows whether the the high frequency synchronizing circuits are providing the proper high frequencies to the HF balanced modulator.
i) the system furnishes 320,000 discrete output frequencies, in 100 cycle increments, based on the accuracy of the 1 mc standard, which is 1 part in 100,000,000 per day. This represents an accuracy of .01 cycle.

A high quality frequency counter, with a resolution of 1 cycle, connected to the output of the system, will read the desired output frequency exactly.

At the present time, this is considered to be precision frequency control of the highest order. However, the demands for even greater stability are being felt. It is possible that in the near future stabilities on the order of 1 part per 10^9 or 10^{10} may be required and obtained.

FREQUENCY MEASUREMENT

Because of the extreme stability requirements of a SSB system, the measurement of frequency is critical. In many cases, the frequency stability of the SSB system is better than that of the frequency meter being employed. This is particularly true of precision frequency controlled systems.

For example: frequencies transmitted from WWV and WWVH are accurate to within 1 part in 10^8. Frequencies received from these stations may be as accurate as those transmitted for several hours per day during total light or total darkness over the transmission path. Errors in the received frequencies may vary approximately between plus and minus 3 parts in 10^7. During ionospheric storms, conditions may cause momentary changes as large as 1 part in 10^6.

In general, the following methods of frequency measurement may be used to good effect:

a) a high quality frequency counter with a resolution of one cycle or better, capable of operation at all frequencies in the range to be measured.

There are many fine frequency counters available; the most notable are those manufactured by Hewlett-Packard, Beckman-Berkeley, Lavoie, and others. Frequency is read directly on a read-out indicator. The counter must be capable, by itself or with suitable converter attachments, to indicate any frequency in the range under consideration. The accuracy of the reading depends primarily on the accuracy of the internal standard in the counter, and on the resolution of the readout. The resolution refers to the last significant figure that can be read accurately.

The minimum signal amplitude required for an accurate reading will depend on the type of counter employed. Generally, the newer counters require less than 1 volt RMS. If the frequency is to be sampled from the transmitter output, sensitivity poses no problem, since the sampling will be done with an attenuator or voltage divider arrangement. In any event, only the carrier frequency should be present when the reading is taken. Until recently, frequency counters were considered to be laboratory instruments. They are becoming increasingly numerous in the field, due to developments in miniaturization. In certain cases, certain alignments demand that a counter be used. It is probable that, in the near future, a miniature frequency counter will be considered as a necessary item of "field" test equipment.

b) a phase comparator circuit. These circuits find particular use in precision controlled frequency systems. The 1 mc frequency standards employed in the TMC Models SBG-1, SBG-2, and DDR-5 precision controlled systems are accurate to 1 part in 10^8 per day. When the system is aligned and operating properly, frequency control will be as accurate as the frequency standard employed. Figures 6-9 and 6-10 show the phase comparator circuits used in the Models DDR-5 and SBG-1, respectively.

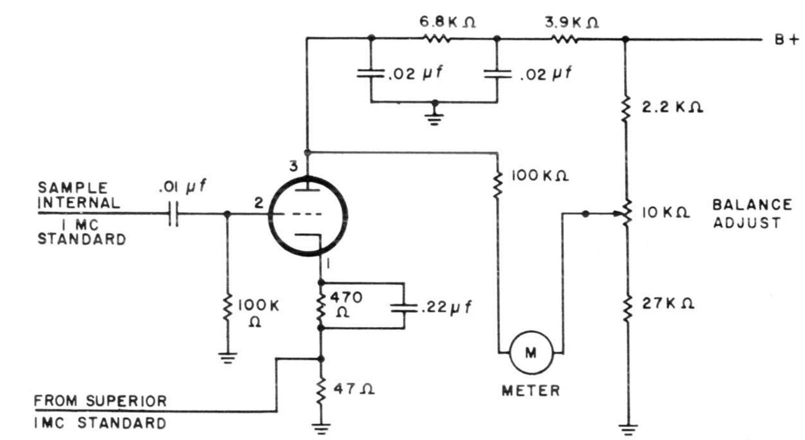


Figure 6-9. One Megacycle Comparator Circuit Employed in the Model DDR-5 Receivers (AN/FRR-60(v)).

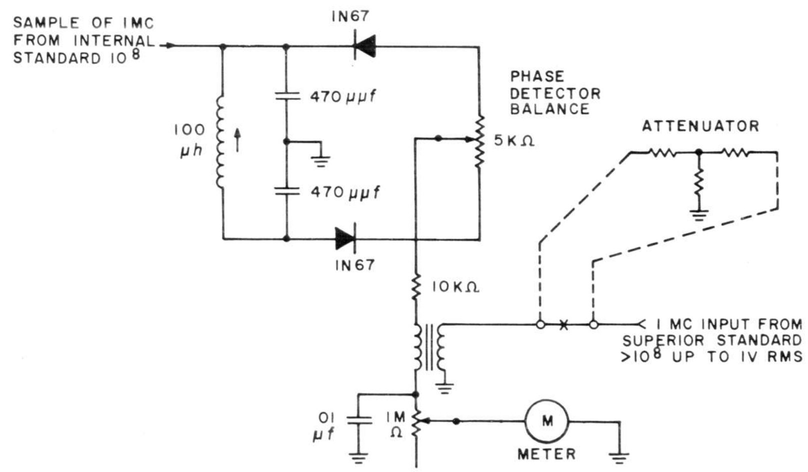


Figure 6-10. One Megacycle Comparator Circuit Employed in the Model SBG-1, SBG-2 and CPO-1 Units.

In Figure 6-9, a sample of the 1 mc output of the internal 1 mc standard is applied to the control grid of the phase comparator tube. The 1 mc output of a superior standard is connected in the cathode circuit. The plate load contains a low pass filter circuit which will present maximum impedance at the lowest audio frequencies only. A zero center meter is connected from the plate to B Plus via a voltage divider circuit. The audio beat between the two ostensibly identical 1 mc signals will be observed as a meter swing around zero as the two signals go in and out of phase. The number of complete swings per second is the error in CPS. No swing will be observed:

(1) if the two frequencies differ appreciably from 1 mc.

(2) if the two frequencies are identical.

The 10 K ohm balance potentiometer adjusts the meter to zero center with the input from the external standard removed. The internal standard is adjusted to bring the two frequencies into exact coincidence.

In Figure 6-10, the sample from the internal 1 mc standard is applied to a simple phase comparator circuit. The input from the superior external 1 mc standard is applied to the wiper of the balance pot via a transformer. An attenuator must be used if the input from the external standard exceeds 1 volt RMS. When the two signals are identical in frequency and 90 degrees out of phase, the DC voltage at the wiper of the balance pot is zero; if the signals swing in and out of phase a positive or negative voltage is developed at the wiper. The DC voltage is applied, via a sensitivity control, to the zero center meter. The number of sweeps of the meter per second indicates the error in CPS. The phase detector balance pot is adjusted for zero on the meter with the external standard removed.

In both the systems just described: if the internal standards are properly adjusted and the system properly tuned, the entire system should be phase locked to the standard. All frequencies generated may be assumed as correct.

c) a good quality heterodyne frequency meter. This usually takes the form of a stable variable frequency oscillator acting in conjunction with a secondary frequency standard, calibration and indicator circuits. The secondary standard is usually a very stable single frequency crystal oscillator housed in a carefully controlled oven; this crystal oscillator circuit is used to calibrate the variable frequency oscillator at a "check point" near the frequency to be measured. The frequency to be measured and the output of the variable frequency oscillator are then mixed in a non linear impedance and the difference frequency, usually in the audio range, is recorded on some type of indicator. The variable frequency oscillator is then adjusted for "zero beat"; its carefully calibrated dial then reads the frequency being measured. Some heterodyne frequency meters have sufficient output to allow them to be used as signal generators. The accuracy of the HFM depends, to a great degree, on:

(1) the accuracy of the secondary standard crystal oscillator circuit, and the amount of time elapsed since its last calibration with a superior standard.

(2) the skill of the operator in reading the indicators to determine exact "zero beat".

(3) the drift of the variable frequency oscillator during the measuring process.

(4) the extent to which the variable frequency oscillator dial is calibrated and the degree of "vernier" tuning.

Generally, heterodyne frequency meters do not have the accuracy to be expected from frequency counters or phase comparator circuits. It is difficult to obtain perfect tracking of a VFO over a wide range. The accuracy of two older U.S. Navy HFM's, the Models LR and LM, is about 1 part in 10,000. Newer equipments have decidedly better accuracy. Heterodyne frequency meters are manufactured to cover all frequencies in the communications spectrum.

d) a crystal calibrator circuit. The operation of this circuit is similar to the HFM just described, but it is generally used to calibrate a variable master oscillator in a relatively narrow frequency range. Figure 6-11 shows the block diagram of the crystal calibrator scheme used in the TMC:

(1) Model VOX Variable Frequency Oscillators.

(2) Model GPT-750 Transmitters.

(3) Model PMO Test Equipments.

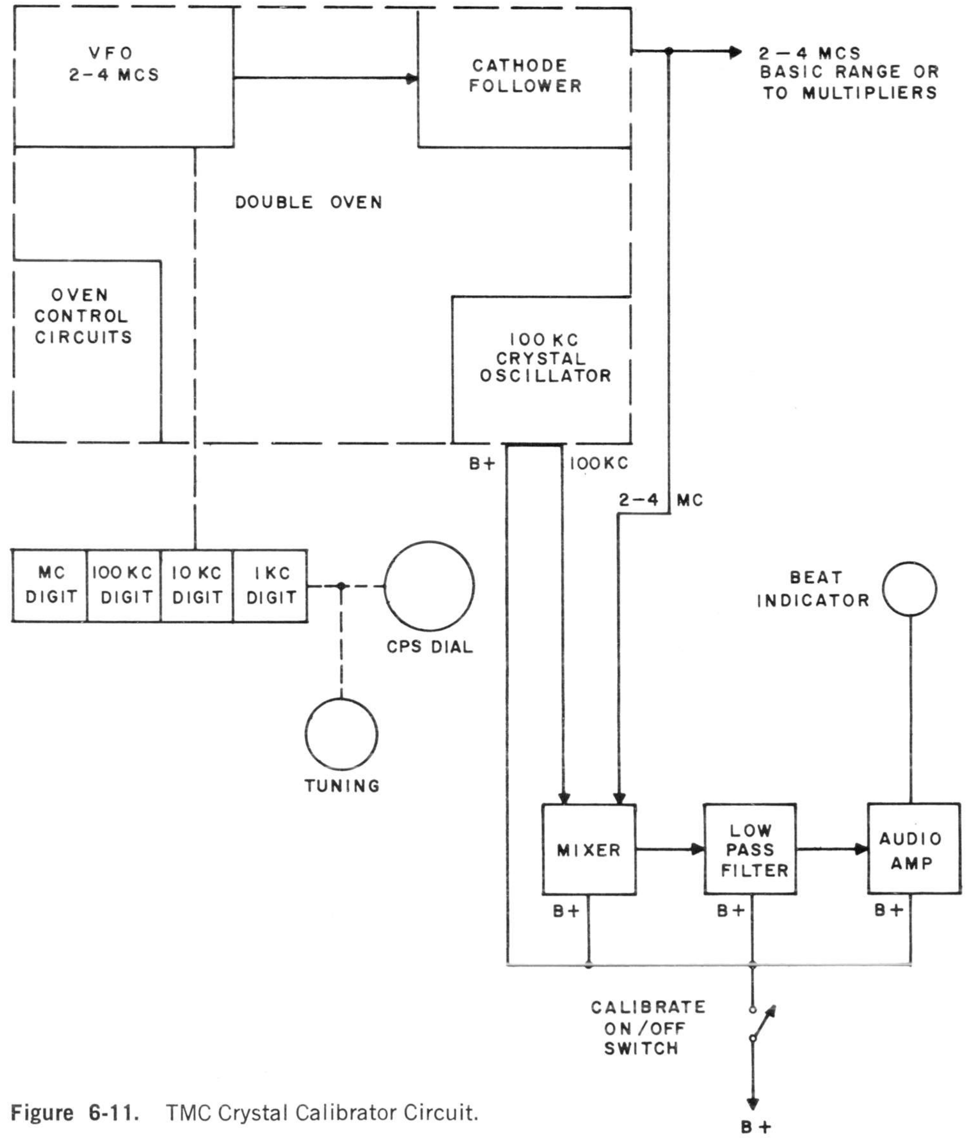


Figure 6-11. TMC Crystal Calibrator Circuit.

The Crystal Calibrator circuit operates as follows:

(1) a double oven contains:
  (a) the VFO, operating in the range 2-4 mcs.
  (b) a cathode follower, for isolation of the VFO from external circuitry.
  (c) a 100 KC crystal oscillator circuit.
  (d) oven control circuitry to maintain constant temperature and to shut down primary oven power in case of malfunction of the inner oven thermostat.

(2) the VFO is tuned by a vernier dial ganged to a counter and auxiliary dial. The MC, 100 KC, 10 KC and 1 KC digits of the VFO frequency are displayed directly. The auxiliary dial displays the remainder of the frequency. The auxiliary dial is marked in 100 cycle steps, but more precise frequencies may be obtained through interpolation. The cathode follower output is delivered to a mixer tube and to external circuitry for straight through operation or multiplication, depending on the particular equipment.

(3) the mixer also receives the output of the 100 KC crystal oscillator. This is applied to the control grid. Since the cathode of the mixer is grounded, grid clamping takes place. This is shown in Figure 6-12.

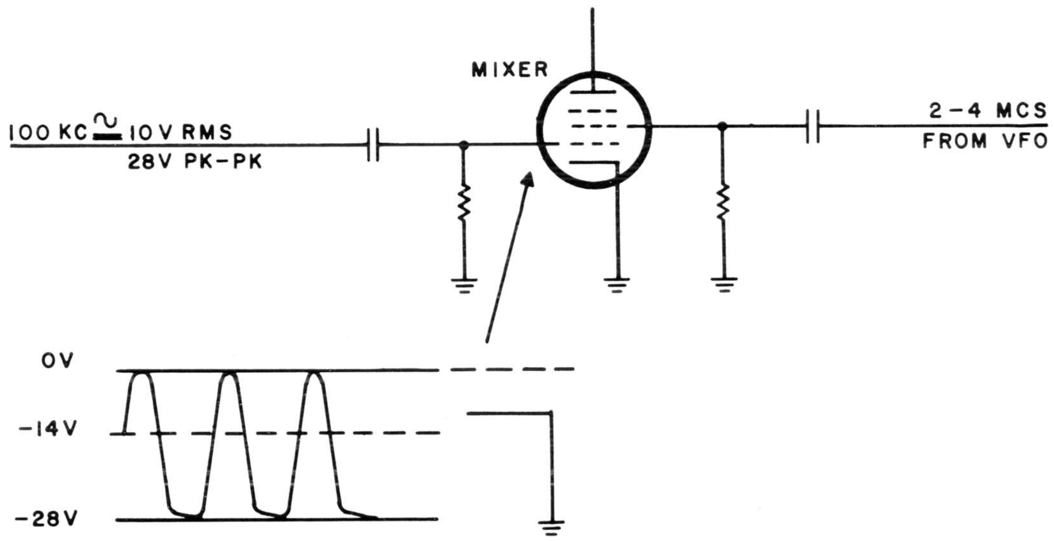


Figure 6-12.

(4) the mixer plate contains a low pass filter designed to pass only the lowest audio frequencies. Audio beats will be obtained at 100 KC points from 2 to 4 mcs. The output of the low pass filter is applied to an audio amplifier designed only for optimum gain and low frequency response. The output of the audio amplifier connects to a neon beat indicator, and, in some cases, to a telephone jack for aural beat indication.

(5) a CALIBRATE ON OFF switch removes plate supply from the crystal oscillator, mixer and audio amplifier in the OFF position.

(6) trimmer adjustments are included to reset the VFO at the high and low ends, when the "resettability" of the VFO counter requires re-alignment. Another trimmer allows adjustment of the 100 KC crystal oscillator frequency to exactly 100 KC.

(7) the accuracy of the system depends on:
  (a) VFO stability. This is 20 cycles per megacycle.
  (b) the accuracy of the 100 KC crystal frequency.
  (c) the skill of the operator in determining exact zero beat, and in manipulating the dial and counter.
  (d) the mechanical-electrical design and adjustment of the dial, counter, and associated tuning capacitors. Every effort has been made to have these components "track" between 100 KC check points.

e) an oscilloscope, preferably one with extended frequency response, high gain, calibrated time base and calibrated vertical amplifiers. This instrument can be used to good effect in many frequency measuring applications, not only by means of Lissajous patterns, but by means of direct measurement. This is especially true of non sinusoidal waveforms at submultiples of the internal standard frequency in precision controlled systems. Such frequencies can be accurately measured with ease.

For example, in the Model DDR-5 Receiver, a precision frequency control system is used. The internal 1 mc standard, accurate to 1 part in 100,000,000, is counted down to 500 KC, 100 KC, 10 KC, 1 KC and 100 CPS. The resultant pulses at 100 KC, 10 KC, 1 KC and 100 CPS are generated by Phantastron circuits. These circuits can be accurately set with the type of oscilloscope described. In addition, these pulses can be used as precise references on the horizontal plates of the oscilloscope while unknown frequencies are applied to the vertical plates, and the Lissajous patterns used to adjust the unknown frequencies for precise measurement or adjustment.

THE MEASUREMENT OF PEAK ENVELOPE POWER

When two tones of equal amplitude are applied to a SSB system, the ratio of PEP to Average Power is 2:1. The average power can be determined from the panel meters; thus, if a 70 ohm resistive dummy load is connected to the output of the transmitter, and the antenna current meter reads approximately 8.4 amperes, the average power being dissipated by the load is 5000 watts. At this time, the PEP is 10,000 watts. This relationship is valid for two tones only, however, and some means must be found to determine PEP regardless of the type of intelligence carried by the system. At the same time, some convenient means must be found to determine when the peak envelope power capability of the transmitter is being exceeded. Two instruments will be found invaluable in this regard:

a) a Modern, high gain, oscilloscope with calibrated time base and calibrated vertical amplifiers, such as the TEKTRONIX series.
b) a VTVM such as the Hewlett-Packard Model 410B, which is a peak reading voltmeter with its dial calibrated in terms of RMS value.

Assume that two equal amplitude tones are applied to a SSB transmitter and that an oscilloscope is connected to a voltage divider or monitor jack at the transmitter output. The voltage at the monitor jack will be greatly attenuated, but this is of no significance. A peak reading, RMS calibrated VTVM such as the Hewlett Packard Model 410B may be used in lieu of or to augment the oscilloscope. Figure 6-13 shows the arrangement.

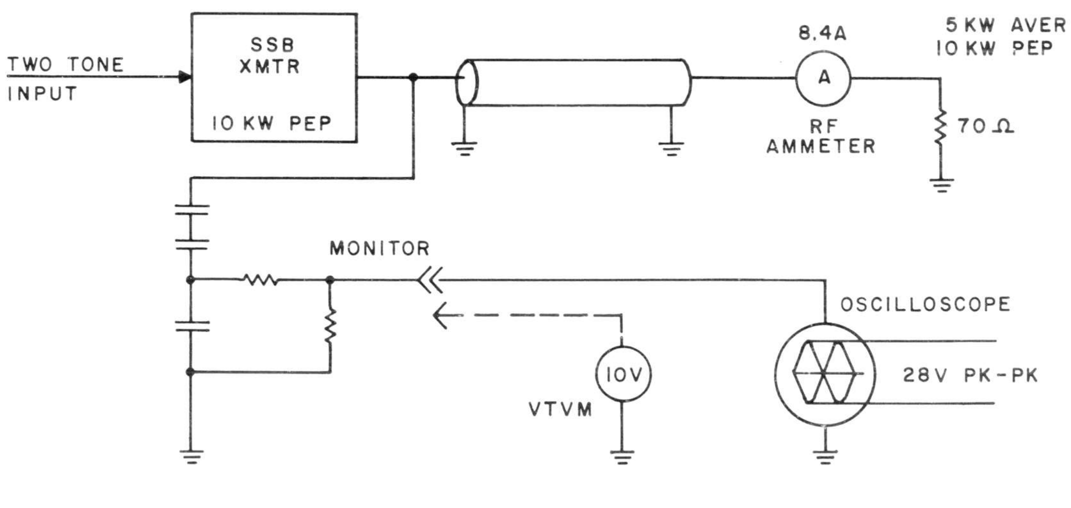


Figure 6-13.

With 8.4 amperes into 70 ohms, the average RF power is 5 KW, and, since two tones are applied, the PEP is 10 KW. We will assume that the oscilloscope shows a two tone waveform with a peak to peak reading of 28 volts. The VTVM would read the RMS value of the envelope peaks, or 10 volts. Remember that peak envelope power is defined as the RMS value of power developed at the crest of the modulation cycle.

We have established two important points:

a) a visual reference, on the oscilloscope, representing PEP of 10 KW. We do not wish to exceed this reference (in the case of a 10 KW PEP transmitter).
b) a numerical reference (28 volts Pk-Pk or 10 volts RMS), representing the absolute value of PEP. Remember that this reference is only an example, used for illustration.

Since, generally, P equals E^2/R: if R is held constant, P will vary as the square of the voltage.

Let PEP(max) = 10 KW and let PEP(meas) = PEP at any other time.
Let E(max) = the voltage corresponding to PEP(max) and let E (meas) = the voltage at any other time.
Then: at any time:

$$\text{PEP(meas)} = \frac{\text{PEP(max)}}{\left(\frac{\text{E max}}{\text{E meas}}\right)^2}$$

Examples:

1. Oscilloscope reads 14V Pk-Pk

$$\text{PEP(meas)} = \frac{10\ \text{KW}}{\left(\frac{28}{14}\right)^2} = \frac{10\ \text{KW}}{4} = 2.5\ \text{KW}$$

2. VTVM reads 12 V:

$$\text{PEP(meas)} = \frac{10\ \text{KW}}{\left(\frac{10}{12}\right)^2} = \frac{10\ \text{KW}}{.694} = 14.4\ \text{KW}$$

3. Oscilloscope reads 25 V Pk-Pk

$$\text{PEP(meas)} = \frac{10\ \text{KW}}{\left(\frac{28}{25}\right)^2} = \frac{10\ \text{KW}}{1.25} = 8\ \text{KW}$$

AN ALTERNATE METHOD FOR INSURING THAT PEP CAPABILITY IS NOT EXCEEDED

a) this method utilizes the ALDC circuits, which are turned off initially.
b) two audio tones of equal amplitude are applied to the system; the transmitter is tuned, loaded and adjusted for rated average power output as indicated by the panel meters. With two tones applied, PEP is two times average power.
c) The ALDC circuits are now energized. The ALDC control on the power amplifier is adjusted so that the output power just commences to decrease. This indicates that the RF applied to the ALDC circuits just exceeds the bias on the ALDC rectifier. The ALDC control is "backed off" this point.
d) the signal drive from the exciter stages is reduced to zero. The audio and exciter stages are adjusted to receive the intelligence to be transmitted, and the carrier is re-inserted as required.
e) the exciter drive to the transmitter is gradually increased. The meters indicating RF output power are carefully watched.
f) as the exciter drive is slowly increased, a point will be reached when the output power just starts to decrease. This represents the point of peak envelope power capability located in step (c).

g) the exciter output is reduced just below the point located in step (f).

h) the ALDC circuits should now insure that the peak envelope power capability of the transmitter will not be exceeded.

i) should the input intelligence be changed, or the amount of carrier insertion be altered, the exciter output may be adjusted to meet the new conditions.

MEASUREMENT OF ODD ORDER (INTERMODULATION) DISTORTION PRODUCTS

a) the generation and reduction of odd order products has been discussed in considerable detail elsewhere in this chapter. To recapitulate:

(1) odd order products are most objectionable because they fall within the transmitted passband.

(2) if two discrete audio frequencies are injected into a SSB system, a set of products will be generated above and below the audio frequencies. These products will be spaced at intervals corresponding to the difference in frequency between the two original audio tones.

(3) the amplitude and frequency of these odd order products may be observed on a spectrum analyzer connected to the transmitter output.

b) two audio tones are used to measure this distortion because:

(1) with two equal amplitude tones applied and the transmitter adjusted for rated average power output, the transmitter is also operating at rated PEP.

(2) with more than two tones injected, the spectrum analyzer presentation becomes confused.

c) the following equipment is required for the odd order product distortion measurements:

(1) a spectrum analyzer capable of covering the frequency range of the transmitter. The sweep width should be sufficient to scan the entire transmitted spectrum.

(2) a stable RF injection oscillator for the spectrum analyzer. The spectrum analyzers usually are operated at a fixed IF frequency, necessitating a VFO, which may or may not be an integral part of the analyzer.

(3) a two tone audio generator. This instrument must generate two audio tones with a frequency ratio of from 3:1 to 5:1, with an absolute minimum of distortion.

d) the measurement is carried out in the following manner:

(1) the transmitter is tuned, loaded, and adjusted for rated average power output with the two equal amplitude tones applied to the system. The transmitter, then, is operating at rated PEP.

(2) a sample of the RF output is taken from the final power amplifier monitor jack; this is connected to the signal input of the spectrum analyzer.

(3) the VFO associated with the spectrum analyzer is manipulated to provide a presentation on the analyzer screen.

(4) the gain and attenuator controls on the analyzer are adjusted to show the two original tones at O DB reference level. (The vertical axis of the analyzer is usually calibrated in DB).

(5) the sweep width controls on the analyzer are adjusted to provide sufficient resolution to observe the third, fifth, and perhaps the seventh order products. These will be found on both sides of the two original tones, at intervals corresponding to the frequency spacing of the two original tones.

(6) the level of the odd order products should be noted. The third order products are usually the most objectionable.

(7) at this point, the transmitter may be "fine tuned" to reduce the spurious responses to a minimum.

(8) should it be impossible to reduce the distortion to the limits required by the transmitter specifications, the input to the spectrum analyzer may be shifted to the monitor output jack of the preceding stages. This technique should "pinpoint' 'the unit responsible for the excessive distortion.

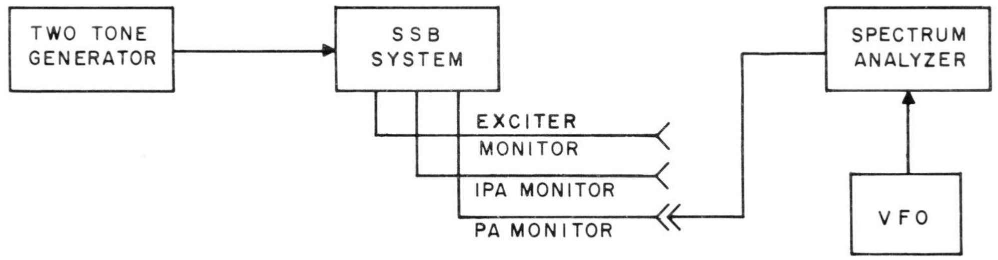


Figure 6-14.

Figure 6-15 shows an idealized plot on an analyzer screen when two audio tones at 935 and 2805 cycles per second are applied to an SSB system. The completely suppressed carrier is at 4 mcs, and the two tones are in the lower sideband. The signal attenuation has been decreased; this places the maximum amplitude of the desired tones off the screen, and accentuates the odd products. In a well adjusted system, the third order products will be down about 40 db from the desired tones.

A METHOD FOR MEASURING ODD ORDER DISTORTION PRODUCTS IN TRANSMITTERS EMPLOYING MULTI-CHANNEL OPERATION

a) the "two tone" test method for measuring intermodulation distortion does not provide adequate information for transmitters operated under multi-channel loading conditions.

b) intermodulation distortion levels of the order of 15 db below signal

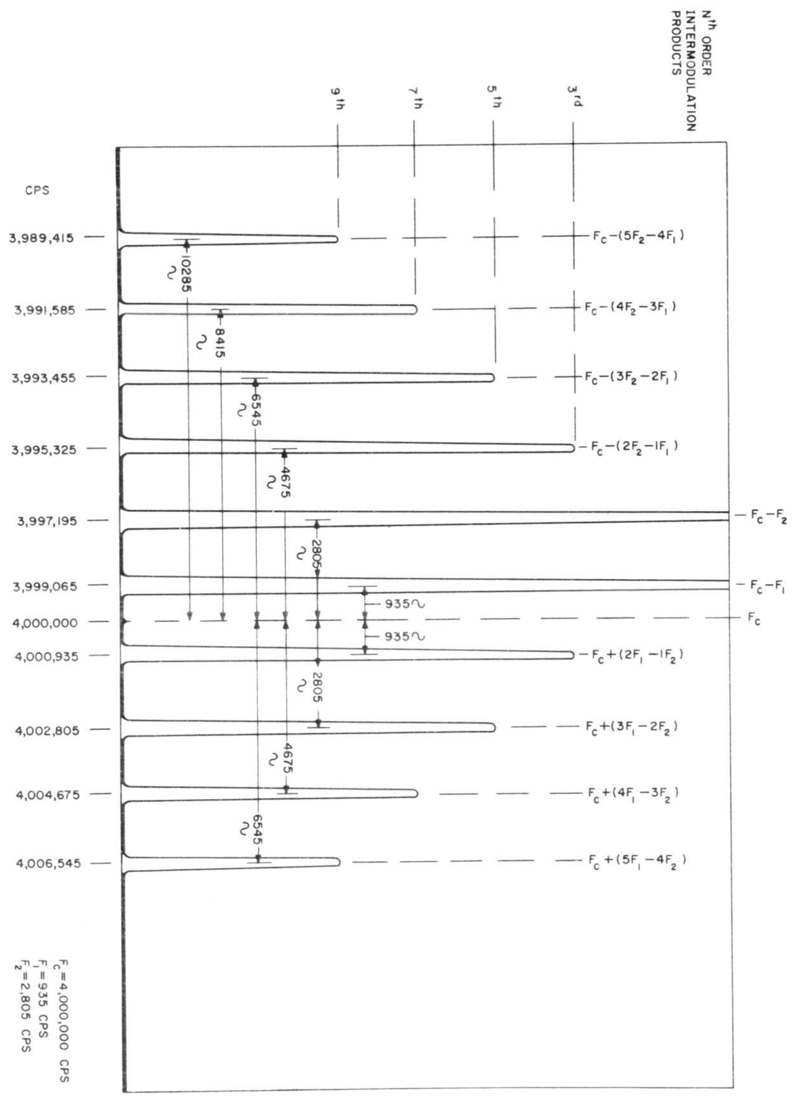


Figure 6-15.

level may be tolerated in multi-channel systems employing FSK without seriously degrading the performance of the system; however, voice channels, often placed adjacent to TTY tone channels, require a much higher signal to distortion ratio.

c) the following basic procedure outlines a method for dynamic measurement of intermodulation distortion in multi-channel systems:

(1) a noise generator is used in conjunction with two 2 channel multiplexers to feed an SSB transmitter system.

(2) each channel occupies a bandwith of about 2700 cycles. Two channels are dispersed on either side of the assigned frequency. This represents a total bandwith of about 12 KC.

(3) the output of the noise generator is fed into *three* channels.

(4) the transmitter output is monitored on a spectrum analyzer. The sweep width, attenuation and resolution controls are adjusted to present a convenient presentation of the total 12 KC transmitted spectrum.

(5) the transmitter is tuned to the assigned frequency and the output is adjusted for rated average power output.

(6) the intermodulation distortion in the *unused* channel is measured with respect to the signal level in the occupied channels.

THE TMC MODEL PTE SPECTRUM ANALYZER

a) the TMC Model PTE Spectrum Analyzer is a versatile item of test equipment used primarily to observe and measure intermodulation distortion products at the output of exciters and transmitters in the frequency range 2.5-30 mcs. It is capable, without additional equipment, of spectrum analysis in the range up to 63.5 mcs, and, with additional frequency generating equipment, up to about 1000 mcs. The Model PTE may be used for:

(1) single sideband studies.
(2) hum level analysis.
(3) adjacent channel interference investigations.
(4) band occupancy studies.
(5) residual carrier and sideband level measurements.
(6) spurious oscillation and modulation measurements.
(7) FM deviation measurements.

b) the device is an automatic scanning superheterodyne receiver with associated cathode ray indicator, which permits analysis and identification of one or more radio frequencies at one time. Each signal in the frequency range being scanned is displayed on a cathode ray indicator in a manner which indicates its amplitude and frequency. A single frequency produces a single "pip". A modulated signal shows frequency distribution and level.

c) the vertical (Y) axis of the indicator is calibrated in terms of amplitude. The horizontal (X) axis is calibrated in terms of frequency.

d) the indicator provides visual means of examining such items as the effects of power supply variations, thermal changes, humidity, component variations, shock, vibrations and load changes upon frequency. Both magnitude and direction of the frequency drift are indicated. Parasitic oscillations which normally pass unnoticed can quickly be detected and identified. The system permits intermodulation distortion readings in the passband to —60 DB.

e) figure 6-16 shows a front view of the unit. Figures 6-17A and B show the operating controls. A complete discussion of this equipment is beyond the scope of this book; from these illustrations, and from the brief discussion which follows, the reader may gain some knowledge of the versatility of the instrument.

f) the hinged lid may be used for a writing surface; this opens on a storage compartment, which may be used to house accessories.

Figure 6-16. Front view of the Model PTE Spectrum Analyzer.

g) the spectrum analyzer proper is mounted at an angle for convenient operator viewing. Figure 6-17A shows the detail of this unit. In addition to the usual oscilloscope controls, provision is made for linear or logarithmic amplitude measurements, variable sweep width, automatic frequency control on certain sweep widths, a wide range of attenuation, 5 KC markers, and many other features. Signal input and VFO input connections are also available at the rear of the analyzer.

h) the control panel contains jacks for sampling the audio tones, VFO, and RF tone outputs.
The AUTO - MANUAL sweep switch is normally kept in the AUTO position; however, when switched to the manual position, the sweep may be moved very slowly by means of the hand crank.

i) the Variable Frequency Oscillator Model VOX provides an injection frequency for the mixer in the spectrum analyzer. The analyzer operates at a fixed IF frequency of 500 KC, and the VFO is set up at a frequency 500 KC higher than the center frequency being observed. The range of the Model VOX is 2 - 64 mcs. A calibration circuit is included to calibrate the oscillator at any 100 KC check point in the basic range of 2 - 4 mcs.

j) the Two Tone Generator Model TTG provides two audio tones, at

935 and 2805 cycles. It also provides two RF tones at 1.999 and 2.001 mcs for calibration of the spectrum analyzer. These tones are furnished at variable levels, and may be turned on and off at will. The output of the TTG is exceptionally clean and free of noise.

MEASUREMENT OF CARRIER SUPPRESSION WITH AN OSCILLOSCOPE

If a Spectrum Analyzer is not available, carrier suppression may be estimated quite closely with an oscilloscope connected to the output of an exciter or power amplifier. This method is excellent for transmitters which do not require complete carrier suppression; it is less effective when the carrier suppression is almost complete, since it requires the reading of peak to peak amplitudes.

a) with a single tone applied to one sideband, a single sine wave output should be observed on the oscilloscope. The waveform will show no modulation with zero carrier leakage. See Figure 6-18A.

b) carrier leakage will show as modulation, as shown in Figure 6-18B and C.

c) if the peaks and valleys of the waveform are read, the carrier suppression can be estimated quite closely with the following formula:

$$\text{Suppression in DB} = 20 \log \frac{A + B}{A - B}$$

where A and B are the oscilloscope readings of the peaks and valleys, respectively. When the carrier is inserted equal in amplitude to the tone, the classic two tone envelope is observed, and the carrier suppression is zero. See Figure 6-18D.

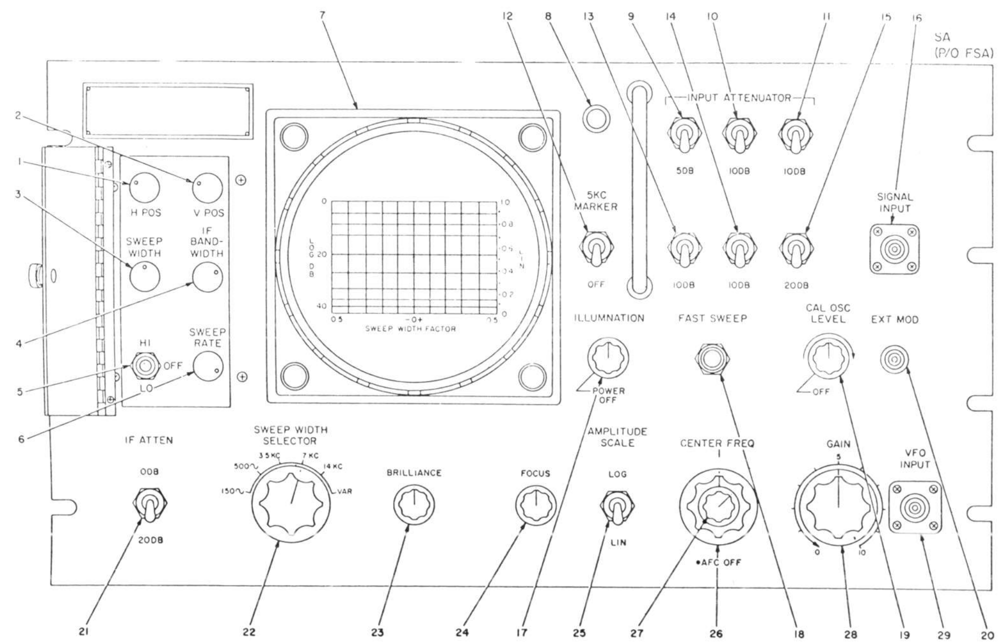
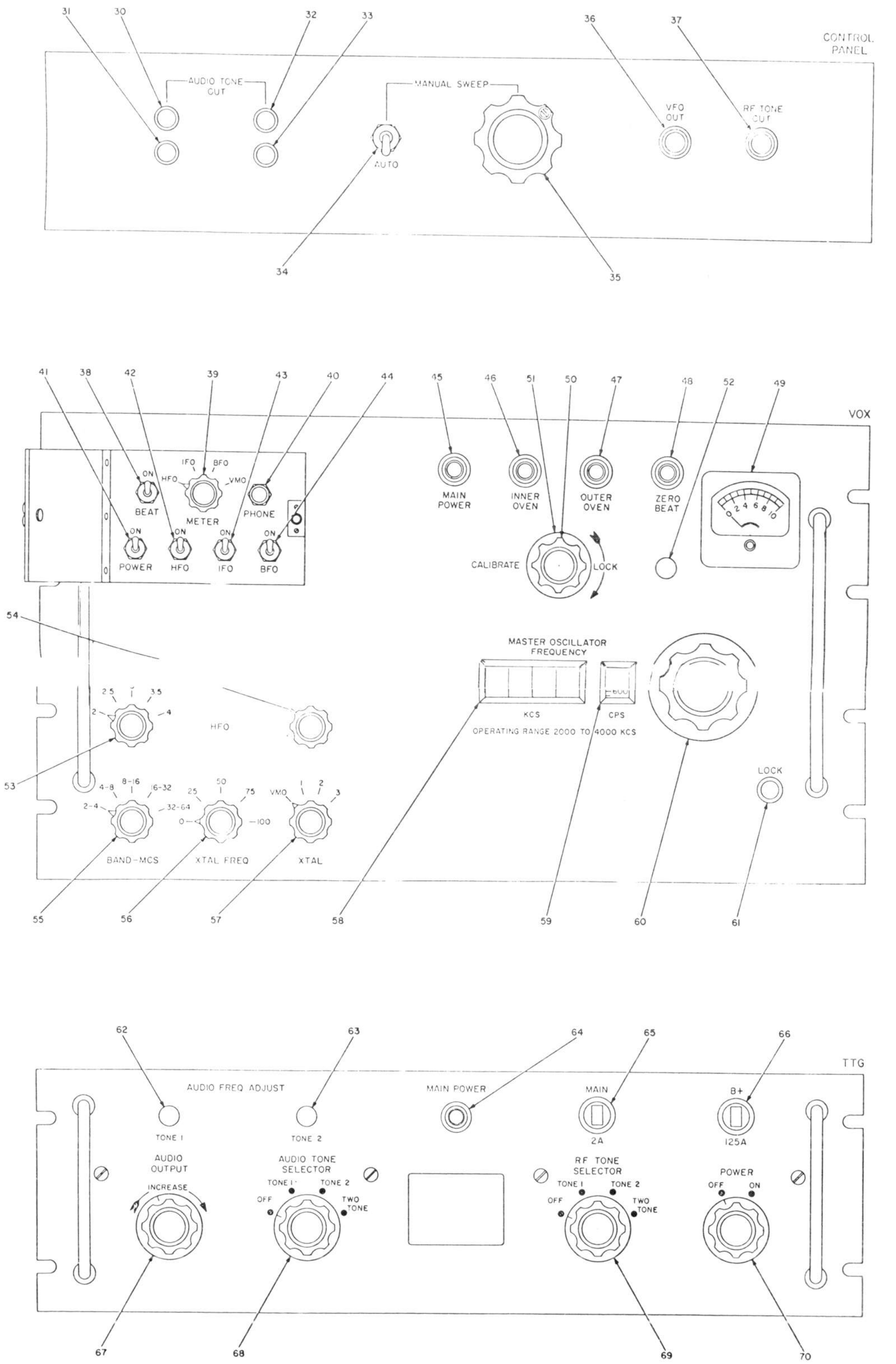
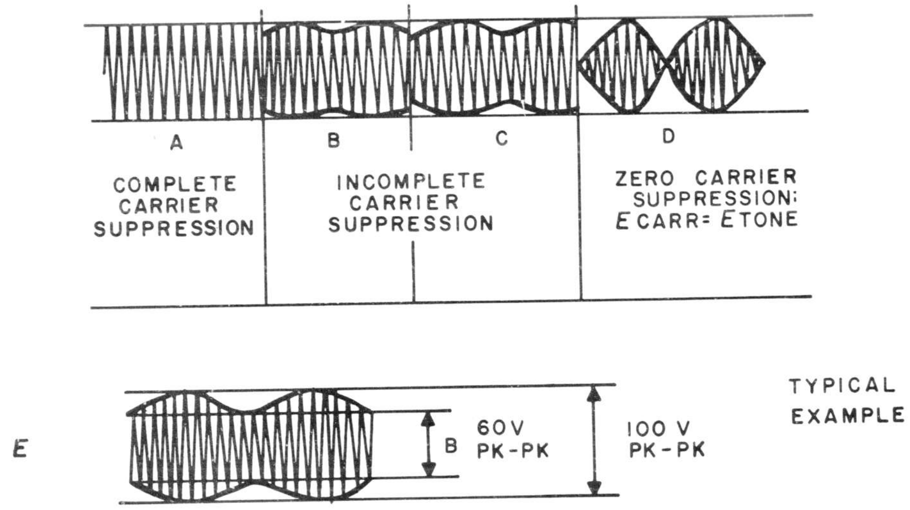


Figure 6-17A. Front view of the Model PTE Spectrum Analyzer.



Panel View of PTE, Showing Operating Controls

Figure 6-17B.



$$\text{SUPPRESSION DB} = 20 \text{ LOG } \frac{160}{40} = 20 \text{ LOG } 4 - 20 \times .3 = 6\text{DB}$$

Figure 6-18.

Caution: Spurious Outputs other than carrier leakage will also cause modulation.

LINEARITY MEASUREMENTS OF SSB POWER AMPLIFIERS, UTILIZING ENVELOPE DETECTORS AND AN OSCILLOSCOPE

a) in a linear power amplifier, with a single RF tone applied, the plate signal, ep, is 180 degrees out of phase with the grid signal, eg.

b) figure 6-19 shows the resulting oscilloscope pattern when the grid and plate signals (single tone) are fed to the horizontal and vertical

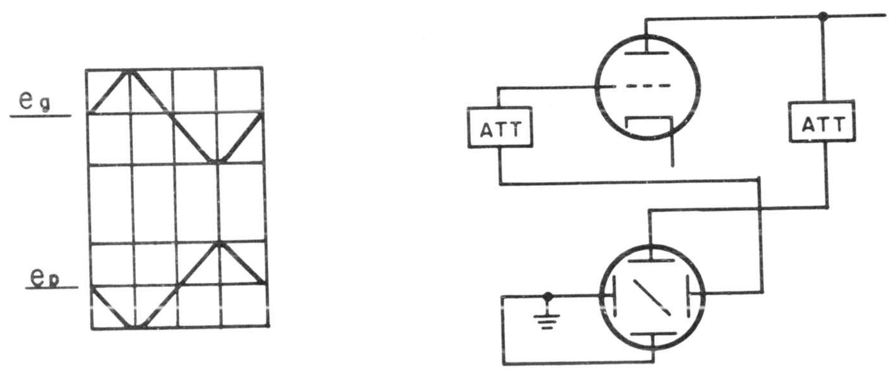


Figure 6-19.

amplifiers of a wide band oscilloscope. This is the classic Lissajous pattern for two sine waves 180 degrees out of phase.

c) it has been pointed out that, for distortion checks, two equal amplitude tones should be applied to a system. This results in the classic two tone SSB envelope. Even though the instantaneous values of the waves making up the composite envelope are 180 degrees out of phase at the grid and plate of the final amplifier, the *envelopes* at these points are *in* phase. This is shown in Figure 6-20.

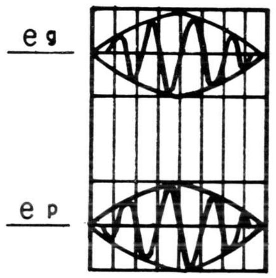


Figure 6-20.

d) if, with two equal amplitude RF tones applied, the horizontal and vertical plates of the wide band oscilloscope are connected to the grid and plate circuits of a linear power amplifier via suitable attenuators and envelope detectors, the oscilloscope will show the classic Lissajous pattern for two signals in phase. This is shown in Figure 6-21.

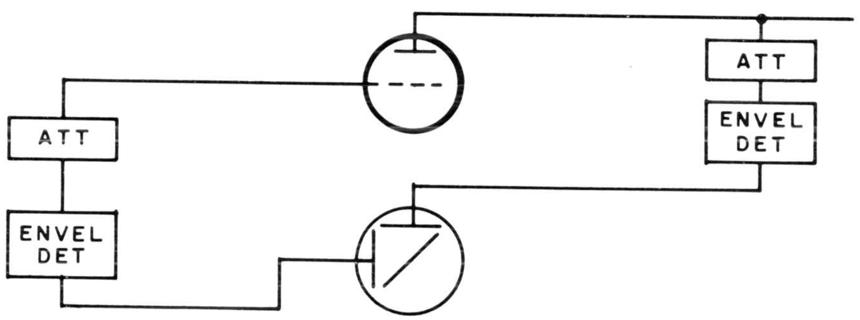


Figure 6-21. Pattern Obtained on Wide Band Oscilloscope with two Equal Amplitude RF Tones Applied to Linear Power Amplifier: No Distortion.

e) the remaining figures show other patterns, and the cause.
   (1) effect of inadequate vertical response of oscilloscope vertical amplifier.
   (2) effect of phase shift. Can be corrected with a simple phase shift network in series with one of the envelope detectors.
   (3) effect of amplifier overloading. The grid drive signal should be reduced.
   (4) effect of inadequate static plate current. Bias should be reduced.
   (5) effect of poor grid regulation when grid current is drawn. The driving signal should be reduced.

f) Figure 6-22 below shows the schematic of an envelope detector. The applied signals should be on the order of 1 volt or more.

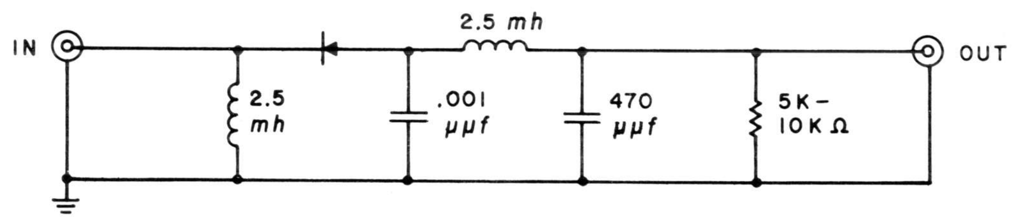


Figure 6-22.

CHAPTER 7

BALANCED MODULATORS

7-1 Introductory Note

Balanced modulator action has already been discussed in considerable detail in sections 3-5 and 5-2; hence, the purpose and general theory of the balanced modulator should, at this time, be well understood. An attempt will be made in this chapter to classify these circuits and to describe TMC configurations for each commonly used classification. Information on the adjustment of these circuits will also be included.

7-2 Classifications of Balanced Modulators

a) balanced modulators may be classified in two general divisions as far as the modulating signal is concerned:
   (1) circuits whose operation depends on the *polarity* of the modulating signal.
   (2) circuits whose operation depends on the *amplitude* of the modulating signal.
b) balanced modulators may be classified in three general divisions as far as physical construction is concerned:
   (1) rectifier balanced modulators.

   (2) multi-electrode vacuum tube balanced modulators.
   (3) nonlinear reactance modulators.
c) rectifier balanced modulators may be classified in three general divisions, in accordance with the way in which the rectifiers are connected:
   (1) Ring or Lattice type.
   (2) Series type.
   (3) Shunt type.
d) multielectrode balanced modulators may be classified in two general divisions, in accordance with the manner in which they are operated:
   (1) square law type, which depends on the nonlinearity of the plate current characteristic.
   (2) product modulator type, which depends on the action of the resultant plate current, due to the use of different control grids for signal injection.

7-3 Rectifier Balanced Modulators

a) rectifier balanced modulators have decidedly advantageous characteristics if they are operated continously under the environmental conditions for which they were designed.
   (1) they are stable.
   (2) they have no power requirements and generate no heat (except for those circuits employing vacuum tube rectifiers).
   (3) they can be made extremely compact.
   (4) they require little maintenance after initial adjustment.
b) any type of rectifier may be employed, including diode vacuum tubes, selenium, copper oxide, germanium crystal, etc.; however, the choice will depend mainly on the power level, the range of frequencies to be translated, and the space available. Modern technique employs carefully matched diode pairs or groups. In addition, a balance potentiometer and sometimes a balance capacitor is included for optimum matching and to compensate for circuit "aging".
c) generally, a carrier suppression of more than 40 db may be obtained. Third order products may be down 50 db, with attendant greater suppression of the higher order products.
d) rectifier balanced modulators depend principally on the amplitude of the modulating signal; the "carrier" signal is used for switching. A carrier to signal amplitude ratio of about 10:1 is common.
e) the action of the "Balanced Bridge" modulator was discussed in detail in Chapter 5. The action of the other types is similar. The Ring or Lattice type has the highest efficiency, because it is a full wave circuit, that is, it produces twice the output voltage as the shunt or series type. In any event, it should be noted that rectifier balanced modulator circuits cannot amplify.
f) Figure 7-1, on the following page, illustrates the basic rectifier balanced modulator configurations.

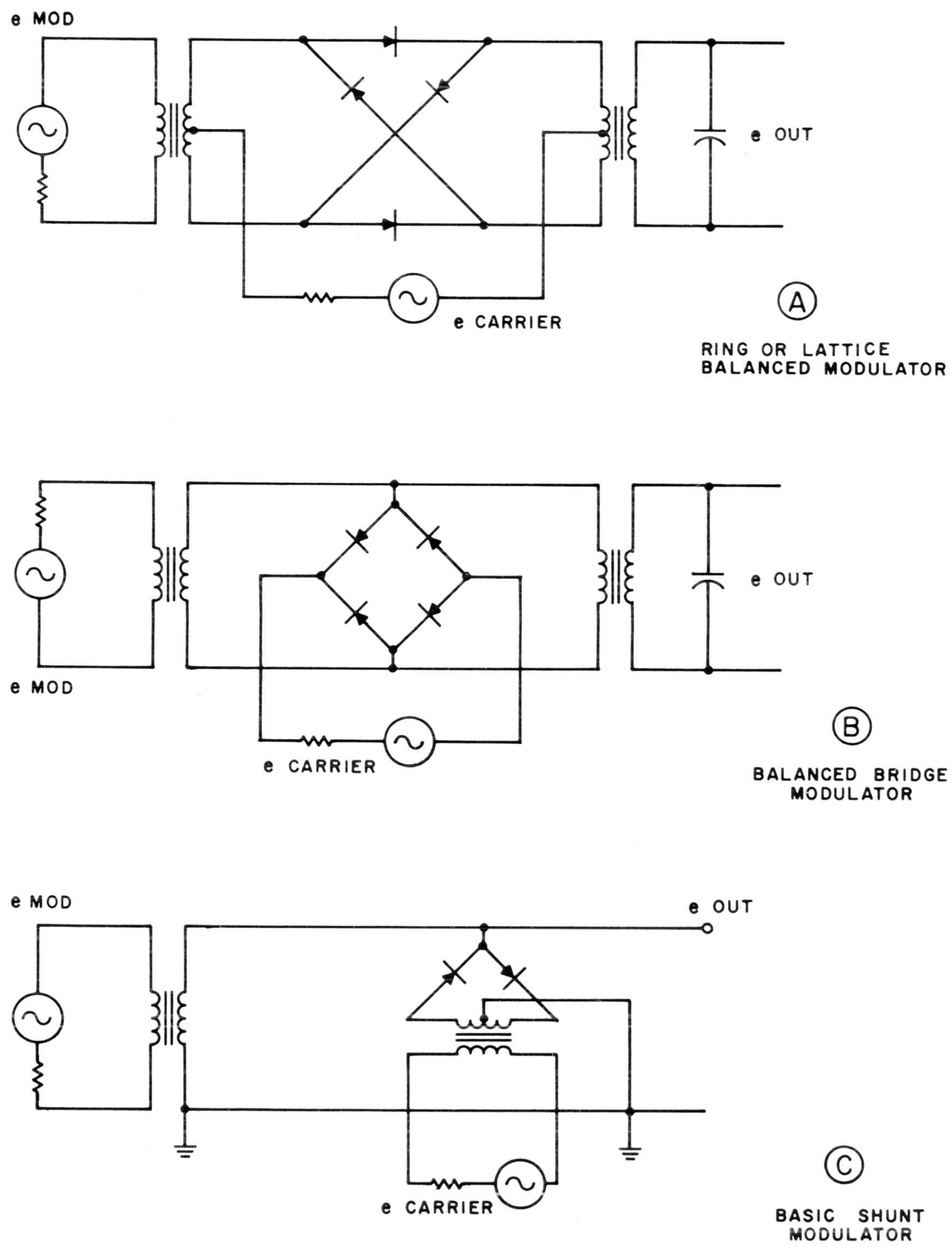


Figure 7-1. Types of Rectifier Balanced Modulators.

7-4 TMC Rectifier Balanced Modulator Configurations

a) High Frequency Balanced Modulator employed in the Model SBE-3 Exciter (AN/URA-28). See Figure 7-2, below:

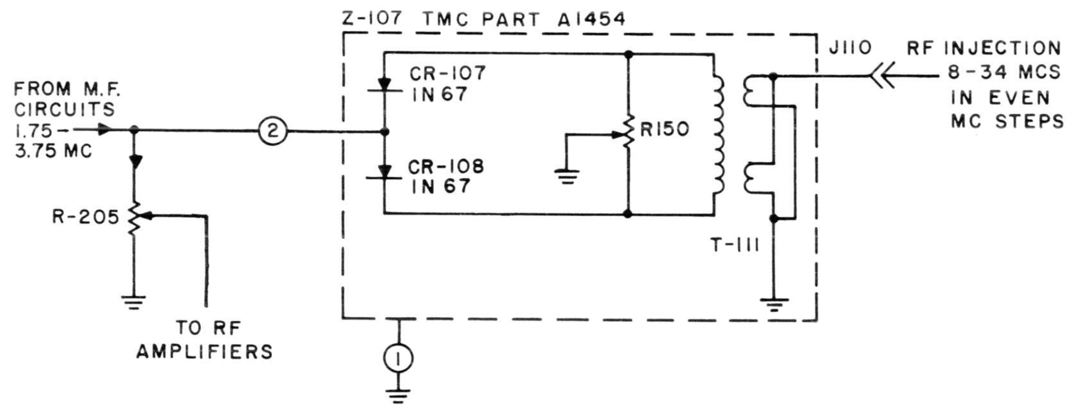


Figure 7-2.

(1) the sketch of Figure 7-2, above, shows the complete circuit, which is enclosed in a metal shield can of approximately 3 cubic inches total volume. Connections are brought out at terminals #1 and #2, and at miniature coaxial jack J-110. The wiper screw of the balance adjust pot is accessible from the top of the can, and is fitted with a locking nut.

(2) terminal #2 serves as both an input and an output connection. The input here is a medium frequency in the range 1.75 - 3.75 mcs. The output is a high frequency, in the range 4.25 - 32.25 mcs.

(3) the input at J-110 is an RF injection frequency, switch selected, in the range 8 - 34 mcs, in even megacycle steps.

(4) the output from terminal #2 is applied to an RF linear amplifier chain via the RF OUTPUT control, R-205.

(5) the circuit is adjusted as follows: the medium frequency input is disabled; a sensitive RF VTVM or oscilloscope is connected at the "high" side of R-205. The BAND MCS switch on the equipment is set for an RF injection frequency of 18 mcs at J-110. The balanced condition will be reached when R-150 is adjusted for minimum indication on the RF indicator connected at R-205. After balancing, the adjust pot is locked.

b) Low Frequency Balanced Modulator employed in the Model CBE-1, 2; (0-714/UR) Sideband Exciter. See Figure 7-3.

(1) the sketch of Figure 7-3 shows the complete circuit; actually, the Sideband Exciter contains two such units; this is the USB balanced modulator. The LSB balanced modulator is identical except for symbol numbers.

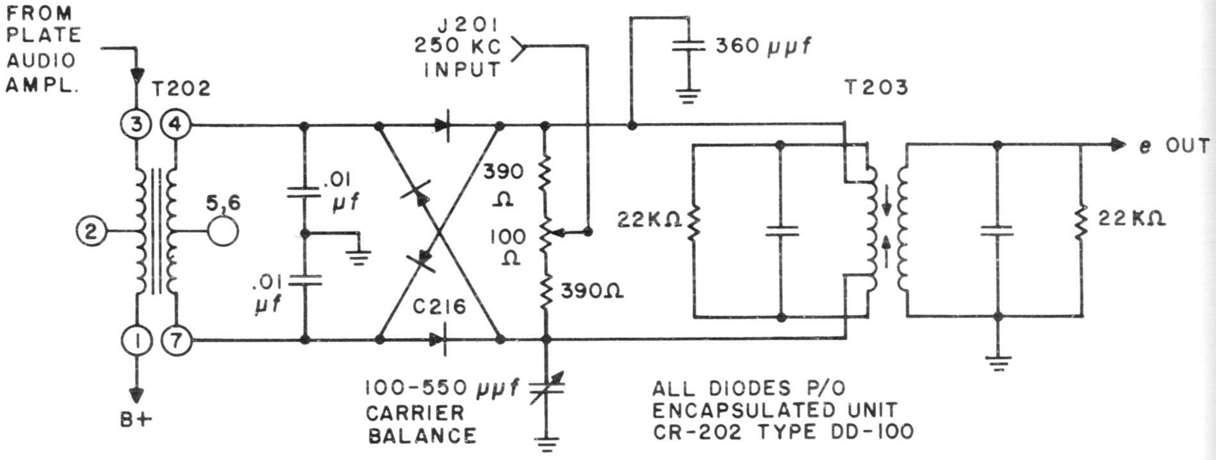


Figure 7-3.

(2) it may be determined at a glance that this balanced modulator is of the Ring or Lattice type. The diodes are contained in a single encapsulated unit, which plugs into a miniature 7 pin tube socket.

(3) the audio input arrives from the plate of the audio amplifier at terminal 3 of T-202. The 250 KC carrier is applied to the wiper of the 100 ohm carrier balance potentiometer.

(4) note that this circuit incorporates two adjustments for carrier balance: the 100 ohm potentiometer, and a trimmer capacitor, C-216.

(5) the sideband output is applied to a double tuned transformer, which uses 22 K ohm swamping resistors to assure the proper bandpass.

(6) the circuit is adjusted as follows:

  (1) a sensitive RF VTVM or oscilloscope is connected to the plate of the RF amplifier following the balanced modulator circuit.

  (2) the carrier balance potentiometer is turned to either extreme to purposely inject 250 KC to the double tuned transformer, T-203.

  (3) the top and bottom slugs of T-203 are adjusted for maximum indication on the RF indicator.

  (4) a check is made to insure that no signal is arriving at T-202 from the audio circuits.

  (5) the carrier balance pot and the balance capacitor, C-216, are adjusted alternately in small increments for minimum indication on the RF indicator connected to the plate of the RF amplifier stage following the balanced modulator. Both adjustments are then locked.

c) Intermediate Frequency Balanced Modulator employed in the Model CHG-2 Frequency Amplifier (AM-2505A/URA-31). See Figure 7-4.

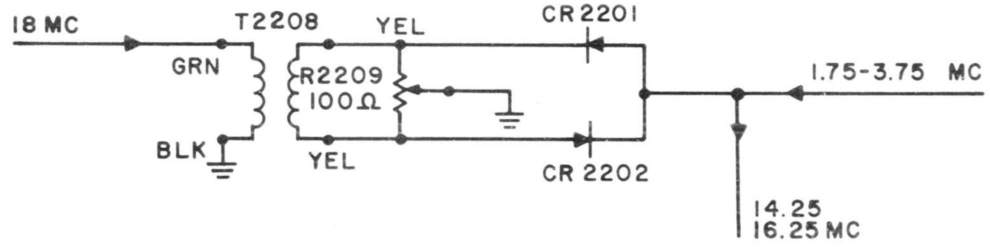


Figure 7-4.

(1) this is a simple circuit, designed to translate a frequency in the range 1.75 - 3.75 mcs, to an intermediate frequency in the range 14.25 - 16.25 mcs. Note its similarity to the circuit of Figure 7-2.

(2) an 18 mc injection frequency is applied at the GREEN dot of T-2208. The frequency to be translated, in the range 1.75 - 3.75 mcs, is applied at the junction of the diodes. From this point the new frequency, in the range 14.25 - 16.25 mcs, is taken off.

(3) the carrier balance adjust potentiometer is adjusted for minimum 18 mc output signal.

7-5 Multielectrode Vacuum Tube Balanced Modulators

a) multielectrode vacuum tube balanced modulators have one distinct advantage over the rectifier types: they provide conversion gain. Because these devices are capable of amplification, care must be taken in their design to prevent oscillation.

b) as stated previously, these circuits are generally one of two different types:

  (1) those which depend on square law operation, or the nonlinear characteristic of the plate current curve.

  (2) those which depend on the resultant plate current when the two frequencies to be mixed are applied to different injection grids. In such circuits, nonlinearity is not necessary. Such circuits are often called: "PRODUCT MODULATORS".

c) multielectrode vacuum tube circuits of this kind pose special problems:

  (1) balance may change with signal level.

  (2) the carrier level for minimum intermodulation distortion is critical.

  (3) due to the aging of vacuum tubes, and because of heating effects, these circuits generally will not remain in balance as long as comparable diode rectifier circuits.

  (4) because balance may change with signal level, and because the carrier level is critical for a minimum of intermoduation distortion, the level of the signal and carrier must be carefully controlled.

d) in the paragraphs which follow, several TMC circuits employing multielectrode vacuum tubes as balanced modulators will be presented.

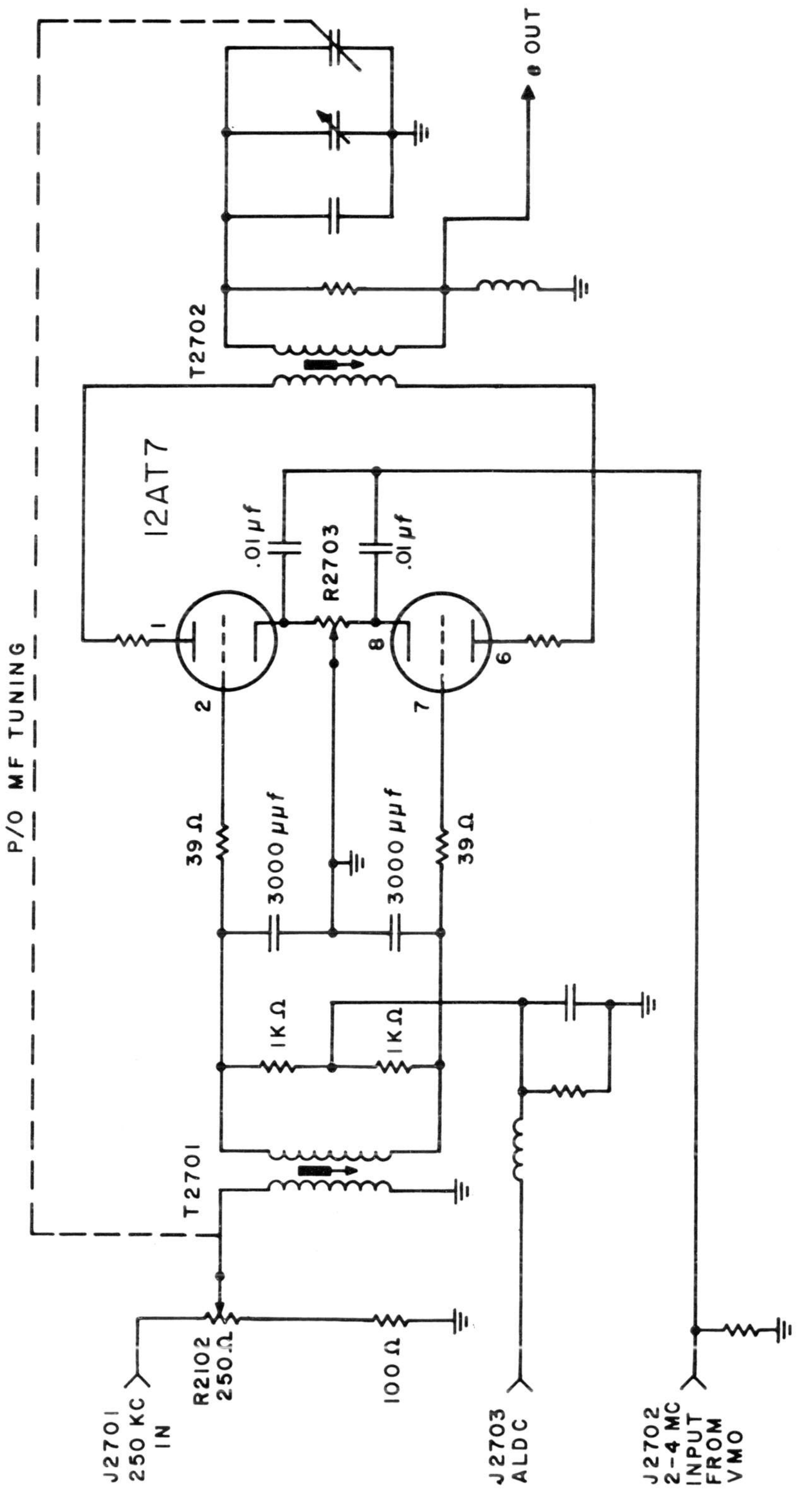


Figure 7-5. Medium Frequency Balanced Modulator employed in Model CHG-2 Frequency Amplifier (AM-2505A/URA-31).

MEDIUM FREQUENCY BALANCED MODULATOR EMPLOYED IN THE MODEL CHG-2 FREQUENCY AMPLIFIER (AM-2505A/URA-31) See Figure 7-5.

a) the intelligence signal, at a nominal 250 KC, arrives from the sideband exciter at J-2701. This is applied to T-2701 via potentiometer R-2102 which is geared to the medium frequency tuning control to change the input amplitude automatically with frequency.

b) the secondary of T-2701 converts the single ended input to a push pull signal; this is applied to the control grids, pins 2, 7, of the balanced modulator tube, a 12AT7. The 39 ohm resistors in the control grid circuit act as parasitic suppressors.

c) the "carrier" injection frequency, in the range 2-4 mcs, arrives at J-2702. This is applied to the cathodes of the balanced modulator tube in parallel via .01 uf capacitors.

d) the circuit incorporates an ALDC jack, J-2703, which connects to a center tap in the secondary of T-2701, via a filter circuit.
The ALDC input comes from a subsequent linear IPA or PA stage. Whenever the peak envelope power in one of these stages exceeds a predetermined value, as determined by the ALDC control, a negative voltage appears at J-2703. This biases both halves of the balanced modulator by an equal amount, reducing the overall gain proportionately.

e) the secondary of T-2702 is tuned to the "difference" frequency; therefore, the signal appearing here is in the range 1.75 - 3.75 mcs. A variable capacitor, ganged to the medium frequency tuning control, provides for continuous coverage within this range. Another tuned RF stage follows the balanced modulator, insuring the required selectivity.

f) balance adjust potentiometer R-2703 is adjusted for minimum "carrier" output, at 2 - 4 mcs.

MEDIUM FREQUENCY BALANCED MODULATOR EMPLOYED IN THE MODEL SBE-3 EXCITER (AN/URA-28). See Figure 7-6.

a) T-127 is a double tuned 250 KC RF transformer; 22 K ohm swamping resistors in both primary and secondary assure the proper bandpass. The 250 KC intelligence signal arrives from the plate of a previous 250 KC amplifier stage. This signal may be an USB, LSB, DSB, ISB signal with any degree of carrier insertion, or it may be a 250 KC carrier signal alone.

b) the push pull outputs at the signal frequency are tapped off the secondary of T-127 for proper matching of impedances. These are applied to the control grids of both halves of the balanced modulator tube, a 12AT7.

c) the "carrier" injection frequency, in the range 2-4 mcs, is applied to both control grids in parallel, via 47 $\mu\mu$f capacitors.

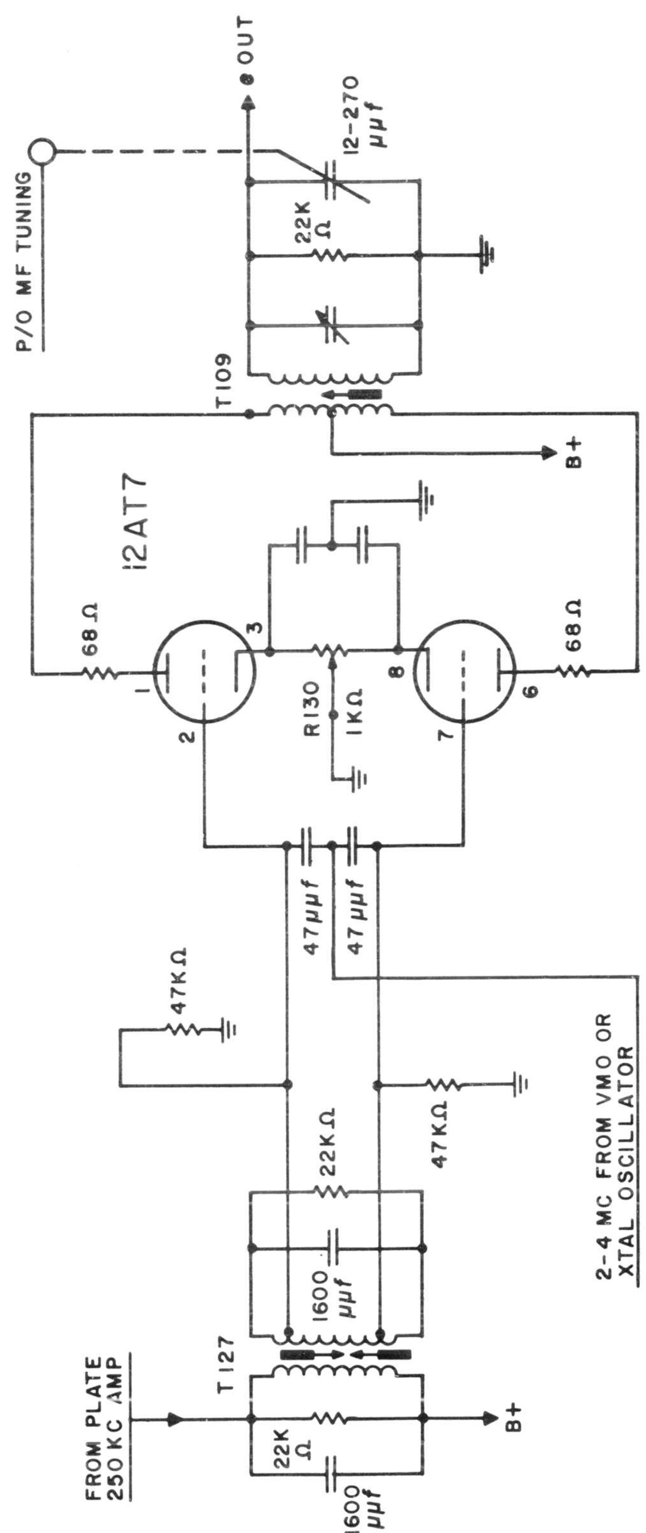


Figure 7-6. Medium Frequency Balanced Modulator employed in Model SBE-3 Exciter (AN/URA-28).

d) R-130, the "carrier balance adjust" assures proper carrier balance by controlling the cathode bias on both tubes.

e) the push pull output at the plates is combined in the primary of T-109. Both plate circuits have associated 68 ohm resistors, which act as parasitic suppressors.

f) the secondary of T-109 is single ended; a tuning capacitor allows tuning through the sum and difference frequencies, since both of these are required in this particular circuit.

g) another tuned RF stage follows the balanced modulator stage, to assure the required selectivity.

HIGH FREQUENCY BALANCED MODULATOR EMPLOYED IN THE MODEL CHG-2 FREQUENCY AMPLIFIER (AM-2505A/URA-31) See Figure 7-7.

a) this is an unusual but effective circuit; both the signal and the injection frequencies are applied in push pull; the tube plates are in a parallel connection, attached to a single ended RF tank circuit. Since the two signals applied cancel in the plate circuit, only the sum and difference frequencies appear in the output.

b) the signal is applied simultaneously to the cathode of the left hand tube and the control grid of the right hand tube. The injection frequency is applied simultaneously to the cathode of the right hand tube and the control grid of the left hand tube. Thus, both inputs are effectively applied in push pull.

c) the appropriate plate tank circuit is switched in by means of a bandswitch control. The balanced modulator circuit is followed by three stages of tuned RF amplification, all operated as linearly as possible, and gang tuned. The selectivity of this cascade arrangement is such that only the desired signals appear in the output.

d) Note the symmetry of the circuit; components in both halves of the circuit are equal and should be matched; as far as is practicable. Final balance is achieved with R-2722, in the cathode circuit.

e) depending on the band of frequencies desired, the balanced modulator output is tuned to either the sum or difference frequencies, because the signal frequencies arrive sometimes inverted and sometimes uninverted. The signal switching arrangements are too complicated to describe here. If further information is required, consult the TMC Lesson Plans for the Model SBG Sideband Generator, Lesson #8.

BALANCED MODULATOR EMPLOYED IN THE "ALIGNMENT GENERATOR" CIRCUITS OF THE TMC MODEL DDR-5 SERIES RECEIVERS

a) this circuit utilizes a type 7360 beam deflection tube, the outline of which is shown in Figure 7-8.

b) the type 7360 is a miniature 9 pin tube. A simple electron "gun", consisting of a filament, cathode, and two control grids creates, con-

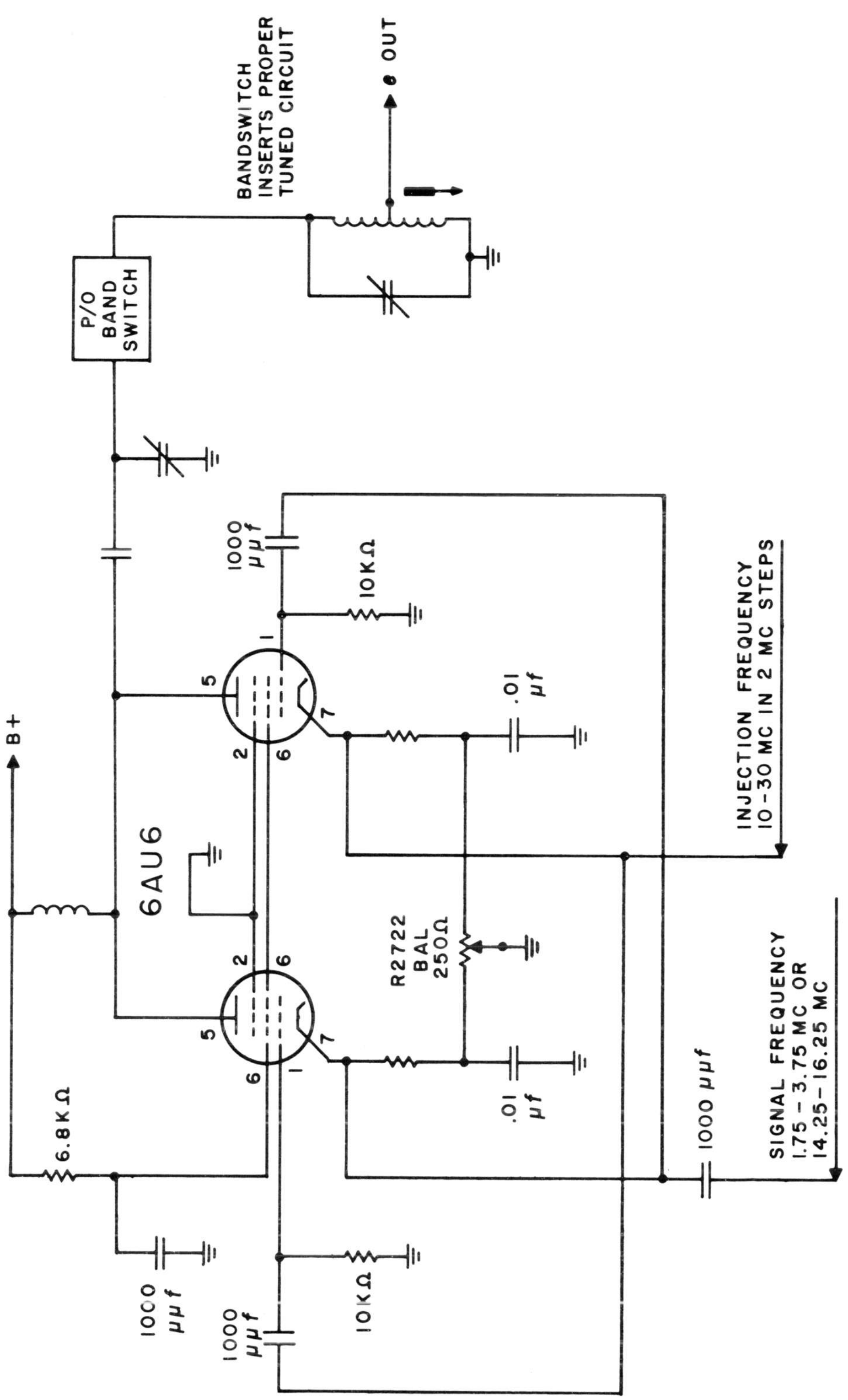


Figure 7-7. High Frequency Balanced Modulator employed in the Model CHG-2 Frequency Amplifier AM-2505A/URA-31.

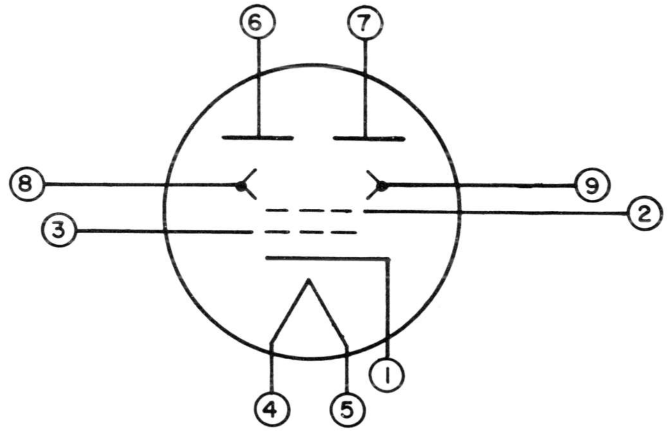


Figure 7-8.

trols and directs an electron beam toward two conventional, independent, plate electrodes. Two deflection plates are also included and these act on the electron beam in much the same manner as the deflection plates used in a cathode ray tube. The instantaneous voltages on the two control grids determine the *total* plate current; the voltages on the deflection plates determine the portion of this current collected by each plate. Generally, grid #2, (pin 2) , is used as an accelerating and isolation anode in much the same manner as the screen grid of a tetrode or pentode tube.

c) consider the elementary circuit shown in Figure 7-9, below:

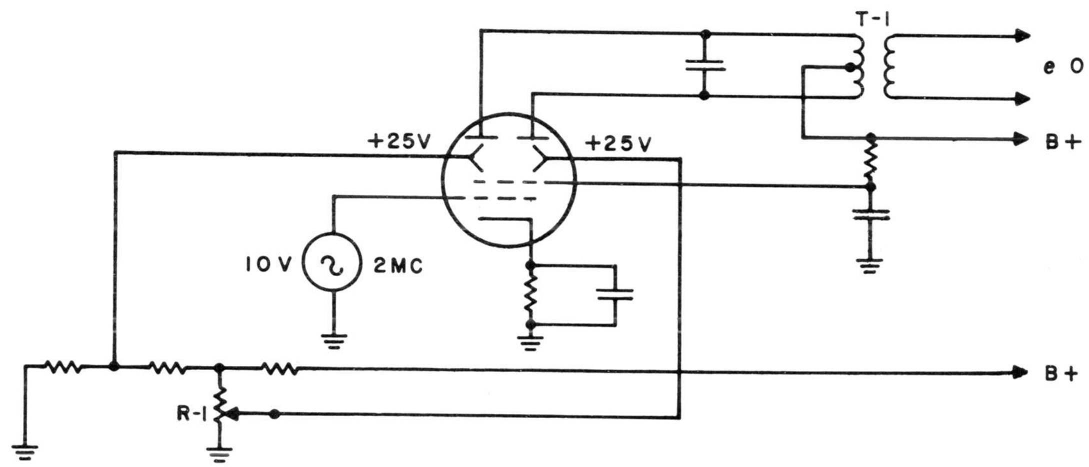


Figure 7-9.

an RF signal at 2 mc is applied to the control grid. A fixed voltage divider places about plus 25 volts on the left deflection electrode. A variable voltage divider allows adjustment of the voltage on the right hand deflection electrode to the same value. With R-1 adjusted for balance, there is no signal output since each plate collects an identical share of the total beam current, and these currents are in phase opposition in the primary of output transformer T-1.

d) should, however, a 250 KC signal of proper amplitude be applied to either deflection electrode, the output is a double sideband signal, containing two discrete frequencies, at 1.75 mc and 2.25 mcs. Either frequency may be chosen by means of suitable tuned circuits or filters. In the TMC Model DDR-5, only the 1.75 mc signal is passed.

e) the complete circuit, as used in the Model DDR-5, is shown in Figure 7-10, below:

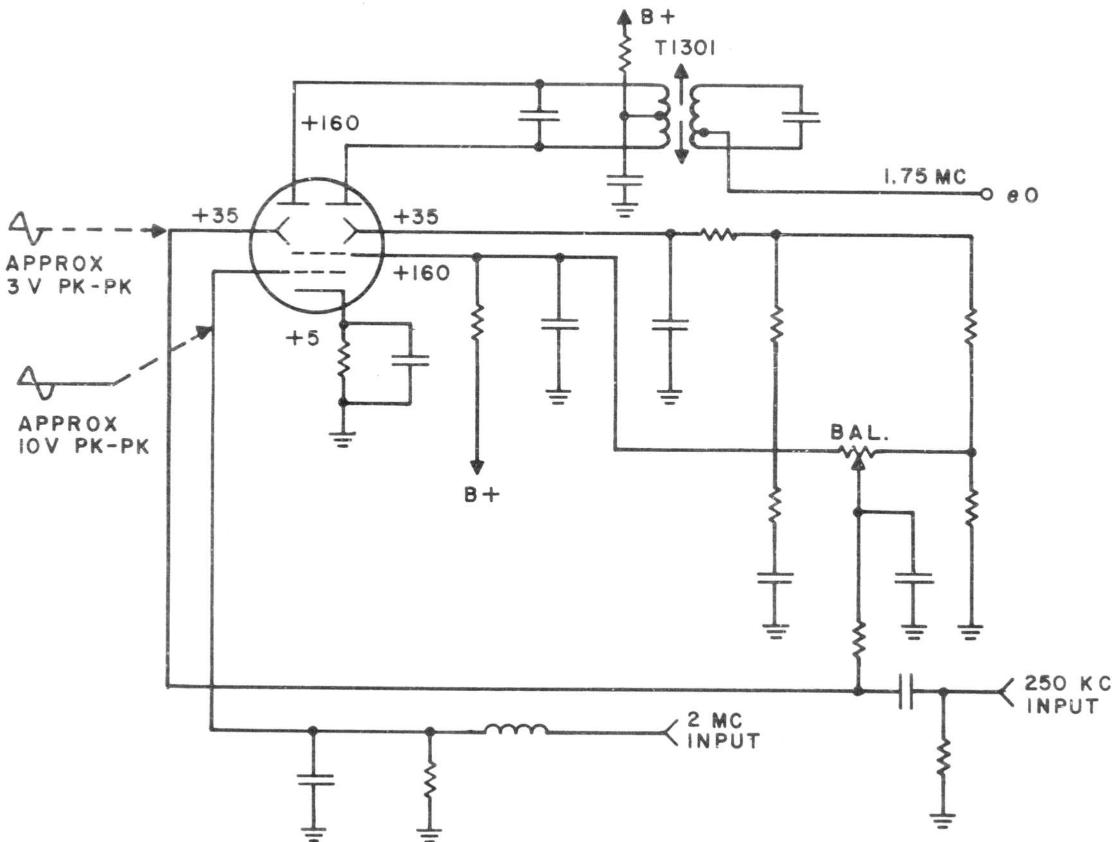


Figure 7-10.

(1) T-1301 is a double tuned transformer which passes only the 1.75 mc difference frequency.

(2) the 2 mc signal is applied to the control grid.

(3) the 250 KC signal is applied to the left hand deflection electrode.

(4) the "screen" grid is returned to B Plus via a normal dropping-bypassing network. This voltage is the source of the reduced deflection electrode potentials.

(5) the balance potentiometer is adjusted for minimum 2 mc output. this will occur when the deflection electrode potentials are equal.

(6) the circuit is used to generate a constant, accurate 1.75 mc signal for use in a special "alignment generator" circuit. The 2 mc input and the 250 KC input is locked to an accurate 1 mc standard; when T-1301 is tuned to pass 1.75 mcs, the accuracy of the output frequency is assured.

f) the type 7360 tube will be found in many new balanced modulator, balanced mixer and product detector circuits. When used as a balanced modulator in a SSB generating system, followed by a selective filter, 70 DB carrier suppression is possible.
Care must be taken to shield the tube from stray magnetic fields, as these will influence the electron beam.

CHAPTER 8

LINEAR POWER AMPLIFIERS

8-1 Introductory Note

In this chapter, linear power amplifiers will be discussed; typical circuits representing current practice will be used as vehicles; power control, overload and interlock circuits associated with linear power amplifiers will be investigated. In addition, PI, and PI-L output tuning networks will be examined in some detail.

8-2 Fundamental Considerations Involving Linear Power Amplifiers for SSB Operation

a) the final power amplifier of an SSB transmitter receives an RF signal at a relatively low power level and increases the power level to that required for transmission. The signal has been processed; that is, it has been translated from the audio spectrum to the RF spectrum; it has had the carrier frequency completely suppressed, and, perhaps, completely or partially re-inserted. It has been passed through balanced modulators, crystal filters, sideband selector switches, notch filters, and linear voltage amplifiers. It now has the configuration desired for transmission. Assuming that the lower level stages are

aligned and operating properly, the signal contains insignificant distortion when it reaches the linear power amplifier.

b) a LINEAR AMPLIFIER is defined as an amplifying device in which the amplified output voltage is proportional to the exciting or driving voltage. This definition implies that no distortion is taking place. Thus, if a linear voltage amplifier has a gain of 30 and three discrete tones at 1, 2 and 3 volts amplitude are applied, the output will contain three discrete tones at 30, 60 and 90 volt amplitudes, respectively. The output will be a replica of the input and no additional outputs will be generated. In a linear power amplifier, power gain, rather than voltage gain, is the criterion. As with any practical device, the ideal conditions are never attained.

c) linear voltage amplifiers can approach the ideal because:
   (1) they are used at relatively low signal levels.
   (2) they are operated over the most linear portions of their characteristic curves.
   (3) their design requirements are not "greedy"; that is, the stages are operated conservatively. The maximum output voltage swing is generally kept to 1 tenth or less than the DC plate voltage to obtain good signal to distortion ratios.

d) linear power amplifiers do not approach the ideal; however, the present state of the art provides excellent results and improvements continue to be made. New tubes have been and continue to be developed solely for high power linear operation. With current practice, using a standard two tone test, odd order intermodulation products (the most objectionable form of distortion in linear power amplifiers) may be down as much as 40 DB from peak envelope power.

e) no discussion per se of linear voltage amplifiers will be undertaken in this text because the classic body of knowledge regarding this subject is well known and understood. The remainder of this chapter will concern itself with linear power amplifiers and associated devices.

CONSIDERATIONS INVOLVING THE CHOICE OF VACUUM TUBES FOR LINEAR POWER AMPLIFIERS

a) to be classified as a good linear power amplifier, a vacuum tube should have the following characteristics:
   (1) HIGH POWER GAIN. The relatively low level of the signal arriving at the input must be raised to the maximum practicable extent without introducing appreciable distortion.
   (2) LOW CONTROL GRID PLATE CAPACITANCE. This is an important factor in determining whether or not neutralization will be required.
   (3) HIGH EFFICIENCY. In high power applications, the percentage of efficiency becomes important in the light of the cost of electricity, the requirements of the power supply, cooling, heat dissipation, and the required value of the driving signal.
   (4) GOOD LINEARITY. The signal arriving at the input con-

tains negligible distortion. The linear power amplifier must not introduce appreciable odd order distortion products since these cannot be eliminated by any means once they are generated.

b) triodes, tetrodes and pentodes are in general use as linear power amplifiers. In general, current TMC practice utilizes beam pentodes as linear power amplifiers in low level exciter units, and air cooled triodes and tetrodes for higher power applications. Triodes are generally used in the highest power applications.

GENERAL CHARACTERISTICS OF TRIODES, TETRODES AND PENTODES

a) Triodes:
triodes are characterized by low gain, low noise and distortion, simplicity of required circuitry and attendant low cost. Because of the relatively large value of control grid to plate capacity, neutralization is almost invariably required, unless the grounded grid configuration is employed. The amplitude of the grid driving signal depends on the circuit configuration and whether a "zero bias" or a "low mu" tube is used. In general, a large value of driving signal is required.

b) Brief Discussion of the "Miller Effect":
in a triode vacuum tube, a capacity is reflected from the plate circuit to the input grid circuit; this capacity is in parallel with the already existing circuit parameters, and must be taken into account when total grid circuit capacity is calculated. This capacity, which we shall call Cm, is equal to:

$C_{GK} + C_{GP} \quad (A + 1) \quad \text{Cos } \theta$ where:

C_{GK} is the control grid — cathode capacity.

C_{GP} is the control grid — plate capacity.

A is the gain of the stage.

θ is the phase angle between the plate voltage and the voltage across the load impedance.

θ is zero degrees when the plate load is resistive.
θ is negative when the plate load is capacitive.
θ is positive when the plate load is inductive.

Cos θ is positive when the load is resistive, and positive when the load is capacitive or inductive. Thus, Cm, the reflected capacity, always has a positive sign.

An input resistance may also be introduced in parallel with the input capacity. This resistance, which we shall call Rm, is equal to:

$$-\frac{\frac{1}{w\,C_{GP}}}{A \text{ Sin } \theta} = -\frac{XC_{GP}}{A \text{ Sin } \theta}$$

where C_{GP}, A and θ have the same meanings as presented for the input capacity, and $w = 2\pi f$.

Sin θ is positive when θ is positive;Sin θ is negative when θ is negative; thus, the input resistance, Rm, may have a positive or negative sign.

As far as the input capacity alone is concerned:

(1) its value changes with gain.
(2) its value changes with the phase angle.
(3) its value is theoretically independent of frequency, since f does not appear in the formula for Cm.

As far as the input resistance alone is concerned:

(1) its value changes with phase angle, gain, and frequency.
(2) its value is infinite when the load is resistive; that is, when θ is zero degrees.

$$R_{IN} = \frac{-XC_{GP}}{A \text{ Sin } \theta} \quad \begin{matrix}(\text{Sin } \theta = 0 \text{ degrees})\\(\text{Sin } \theta = 0)\end{matrix}$$

$$R_{IN} = \frac{-XC_{GP}}{A \times 0}$$

when θ is zero degrees, then, this reflected infinite resistance, R_M, shunting the existing input grid resistance, has no effect. Only the reflected capacity, $C_M = C_{GK} + C_{GP} (A + 1) \text{ Cos } \theta$ appears in the grid circuit.

(3) when the load is capacitive, θ is negative, and Sin θ is negative. Let —N equal Sin θ, which is negative.

$$R_{IN} = \frac{-XC_{GP}}{A \text{ Sin } \theta} = \frac{-XC_{GP}}{A\ (-N)}$$

R_{IN}, then, is positive.

(4) when the load is inductive, θ is positive, and Sin θ is positive.

$$R_{IN} = \frac{-XC_{GP}}{A \text{ Sin } \theta}$$

R_{IN}, then, is negative.
when the load is inductive, and when the reflected negative resistance R_M is less than the aggregate existing grid circuit resistance, energy is transformed from the plate circuit to the grid circuit and oscillations occur.

To Summarize:

(1) because of the Miller effect, a triode tube must be neutralized to reduce the effect of the plate to grid capacity on the circuit.
(2) without neutralization, and with improper tuning, a negative resistance may be transferred from the plate to the grid circuit, causing oscillation.
(3) the changing value of the reflected capacitance with changing gain and phase angle may cause serious de-tuning in the grid circuit.

c) Tetrodes:
tetrodes in general have higher gain than triodes; they also generate more noise. Neutralization may or may not be required, but the control grid to plate capacity is not a serious problem, because the screen grid electrode, at effective RF ground, acts as an electrostatic shield between the control grid and the plate. Neutralization is often employed when the tube is driven for full power output, to obtain optimum signal to distortion ratios. Tetrode tubes generally require "stiff" screen voltage supply regulation; this is tantamount to having a very low screen supply impedance. With good screen regulation, the screen voltage changes negligibly with large changes in the screen current due to signal variations. Stiff screen regulation is necessary because the screen voltage has a marked effect on the dynamic characteristic of the tube.

d) Pentodes:
pentodes have the highest gain, and generate the most noise. They are difficult to design for very high power linear applications. Pentodes have very little control grid - plate capacity, because of the increased shielding between plate and grid afforded by the suppressor electrode; hence, neutralization is seldom required. Pentode circuits are not as simple as triode or tetrode circuits but, in relatively low power applications, they provide excellent service.

CLASSES OF VACUUM TUBE OPERATION

NOTE: this subparagraph is in the nature of a brief review. Readers of this work should already be well acquainted with the classes of amplifier operation.

a) Class A Operation:
under conditions of Class A operation, plate current flows for 360 degrees of the input cycle. The average plate current, I_O, remains substantially constant as the signal current swings above and below this amount. The operating point, which establishes I_O, is adjusted to fall in the central portion of the linear plate current characteristic, and the signal swings are kept within the linear range. As might be expected, the percent efficiency is low, about 30%. This is acceptable, because Class A operation results in negligible distortion. In addition, Class A operation is used only at low power levels.

b) Class B Operation:
under conditions of Class B operation, plate current flows in pulses for about 180 degrees of the input cycle. The tube is biased to about cutoff and the signal amplitude is adjusted so that the maximum positive grid voltage swing is in the neighborhood of 0 volts. Grid current is avoided when low distortion is desired. Despite the fact that the plate current flows in pulses, the "flywheel" effect of the RF tank circuit in the plate circuit produces sinusoidal voltage excursions at the output. Class B operation is often employed with push-pull amplifiers. Efficiencies range from about 50% to 70%.

c) Class C Operation:
under Class C conditions, plate current flows for less than 180 degrees of the input cycle. The bias is adjusted to a value well beyond cutoff under static conditions. The driving signal must be high in amplitude; this signal causes the plate current to traverse the entire range of its characteristic from cutoff to saturation. The driving signal causes the control grid to become positive with respect to the cathode for an angle θ_G, which is smaller than θ_P, the angle of plate current flow. The plate current pulses are much steeper than those found in Class B amplifiers; these pulses "kick" a high Q RF tank circuit, which produces positive and negative excursions at the output. Due to the distorted shape of the plate current pulses, the output may be rich in harmonics; it may be necessary to eliminate these with appropriate tuned circuits and filters. Efficiencies range from about 70% to 85%. Class C amplifiers do not respond to low level input signals; since the static bias is well beyond cutoff, a large value of driving signal is required. Class C amplifiers are not, by themselves, classed as linear amplifiers.

d) Class AB Operation:
under Class AB conditions, the tube is operated such that plate current flows for more than 180 degrees but less than 360 degrees. Grid current may or may not be drawn. The absence or presence of grid current is indicated as a subscript, 1 or 2, respectively. The Class AB amplifier generally has a higher idling current than the Class B amplifier. As might be expected, efficiencies fall between the Class A and Class B values: (about 50% maximum).

e) linear power amplifiers in commercial service are usually operated under Class A or AB_1 conditions, even though applications will be found for the other modes. Because of low efficiency, Class A operation is usually limited to amplifiers with low output power on the order of a few watts. The low efficiency can be tolerated because of the low signal level and the absence of distortion. Class AB_1 operation is employed at the higher power levels. Readers will doubtless find references to Class B and even Class C linears, but these circuits generally require circuit modifications beyond the classic configurations. The Class A and AB_1 modes are generally conceded to offer the cleanest signal with the simplest circuitry.

8-3 Circuit Configurations

a) circuit configurations for linear power amplifiers are of two types:
(1) the grounded cathode circuit.
(2) the grounded grid circuit.

b) both configurations correspond to the well known grounded grid and grounded cathode circuits of conventional operation.

c) An example of a Grounded Cathode Linear Power Amplifier:
(1) Figure 8-1 shows the final power amplifier stage of a typical sideband exciter. This stage is fed by two cascaded stages of linear voltage amplification. The tube employed is a type 6146 power pentode.

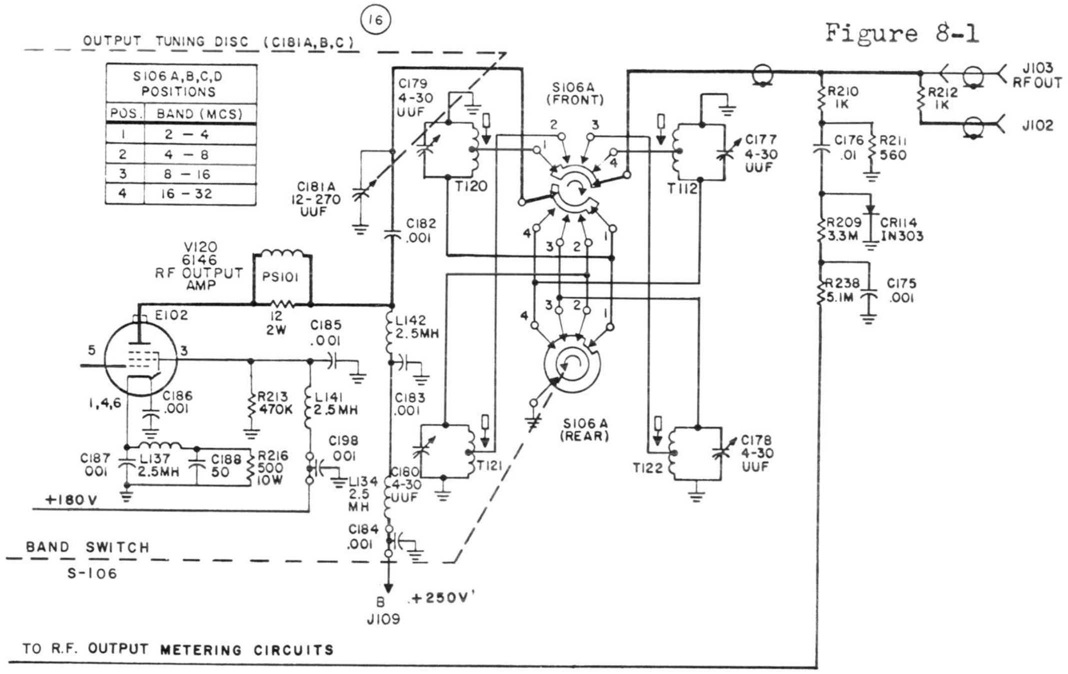


Figure 8-1.

(2) the RF signal from the voltage amplifier, in the frequency range 2-32 mcs, is applied to the control grid, pin 5.

(3) the approximate RF voltage amplitudes of the signal at this point, for maximum output (3 watts) are given below for each of the four bands:

Band #1:	2.0	mcs	6.0	volts
	4.0	mcs	5.5	volts
Band #2:	4.0	mcs	8.0	volts
	8.0	mcs	6.5	volts
Band #3:	8.0	mcs	9.0	volts
	16.0	mcs	7.5	volts
Band #4:	16.0	mcs	10.0	volts
	32.0	mcs	9.0	volts

(4) the static cathode voltage at pins 1,4,6, is found to be about 30 volts. This would indicate:

(a) that the total, no signal plate and screen current is 60 milliamperes.

(b) that the stage is being operated Class A with the bias much greater than the amplitude of the applied signal. The control grid does not, at any time, become positive with respect to the cathode. The peak to peak value of the maximum signal amplitude, (10 volts), is 28 volts.

(5) elaborate cathode, screen and plate bypassing is employed to assure stability and to prevent RF from entering the power supply. A parasitic suppressor, PS-101, is incorporated in the plate lead.

(6) as with most power amplifiers, the average voltage gain of the stage is comparatively low (about 8) but power gain is the criterion here.

(7) a section of the bandswitch, S-106A FRONT, transfers the plate signal via coupling capacitor C-182, to one of four tuned tank circuits. A section of the main tuning capacitor, C-181A, forms most of the tank capacity. The tank circuit contains an adjustable inductance and a trimmer capacitor to permit proper alignment and tracking.

(8) another section of the bandswitch, S-106A REAR, shorts out the three unused tank circuits to prevent unwanted oscillations and to prevent them from absorbing RF energy.

(9) the RF signal is taken from the tap on the inductance of the tank circuit in use, and is applied to:

(a) the RF OUTPUT jack, J-103.

(b) the RF MONITOR jack, J-102, via a 1 K ohm dropping and isolating resistor, R-212.

(c) to a simple voltage divider — rectifier — filter circuit, the output of which is a negative DC voltage. The amplitude of this DC voltage is proportional to the amplitude of the RF signal. This voltage is applied to an RF metering circuit.

(10) the RMS value of the RF voltage with a single tone applied is 14 volts, which corresponds to 3 watts dissipated in a 70 ohm resistive load connected to the output jack, J-103.

(11) electrode potentials are listed below:

(a) Plate cap: approximately 250 volts.

(b) Screen grid: approximately 180 volts.

(c) Cathode: approximately 30 volts.

(d) Control grid to Ground: 0 volts.

(e) Control grid to Cathode: —30 volts.

(12) this circuit is typical of a Class A linear power amplifier operated at a relatively low level. Under other conditions of operation, the type 6146 tube is capable of up to 100 watts output; here, the stage is very conservatively rated to obtain optimum signal to distortion ratios.

d) An Example of a Grounded Grid Linear Power Amplifier:

Note:
Figure 8-2 shows the amplifier circuit. Figures 8-3 through 8-8 show significant overload and power control circuits associated with this amplifier.

(1) the circuit of Figure 8-2 shows a Class AB_1 linear power amplifier employing a type 4CX-5000A air cooled tetrode. The circuit is stripped to its bare essentials; no power control, bias, power supply or regulator circuits are shown. This circuit represents the output stage of a 10 KW PEP linear RF amplifier for general service.

(2) the control grid is returned to ground via the bias supply. The bias is variable, and may be adjusted to provide the optimum operating point. Normally, the bias is adjusted to provide .5 amperes of plate current, with the system in the OPERATE condition, and screen voltage applied. The PA BIAS ADJUST control, not shown, allows adjustment of the bias between about —200 and —300 volts. The bias is monitored on a front panel PA BIAS meter.

(3) due to elaborate RF bypassing arrangements, the control grid is at effective RF ground.

(4) a 10 $\mu\mu$f, 17 KV capacitor, C-929, is connected between the plate and the control grid. This capacitor effects approximately 10 DB of feedback, to improve stability and to reduce distortion.

(5) the screen grid, elaborately bypassed, is returned, via the screen current metering and overload circuit, to:

(a) plus 1200 volts in the OPERATE condition.

(b) plus 600 volts in the TUNE condition.

the screen supply is stiffly regulated. Provision is made to turn the screen voltage on and off independently of the plate; but, because of certain power control arrangements, it is impossible to apply screen voltage before plate voltage has been connected. This prevents the screen from drawing excessive current.

(6) the input RF signal from the IPA stage is applied to the filament (cathode) at J-901.

(7) an RC voltage divider network, C-941, C-942, R-908 and R-910 is incorporated in the cathode circuit. This divider samples the RF drive from the IPA, and makes it available at J-902 for monitoring purposes.

(8) filament voltage is applied from transformer T-801. The secondary is rated at 8.5 volts, 75 amperes, centertapped. The voltage is applied via RF coil L-915. This coil has two individual windings. coaxially mounted. The secondary of T-801 is grounded at the center tap via the plate current meter and the plate overload relay coil (this is shown in Figure 8-4). This point is also at effective RF ground, due to the ground between C-946 and C-947.

(9) Bifilar coil L-915 provides each filament lead with an RF choke. This coil has two main functions:

(a) it provides an impedance for the IPA stage to "work into".

(b) it assists in keeping RF out of the power supply.

(10) neutralization, per se, of this amplifier is not required; the control grid is at RF ground. In addition, the screen grid also acts to reduce the plate to control grid capacitance.

This is a principal advantage of a grounded grid configuration.

(11) C-909, a 3 $\mu\mu$f 17 KV vacuum capacitor, delivers a sample of the RF voltage at the plate of the amplifier tube to:

(a) a PA MONITOR voltage divider and jack.

(b) the PA PLATE VOLTAGE (RF) metering circuit.

(c) the ALDC circuits.

(12) the PA PLATE RF metering circuits incorporate a simple diode rectifier, CR-901, and low pass filter circuits. The sample of the RF voltage delivered by C-909 is further reduced by coupling capacitor C-905. On the negative excursions of the RF signal, diode CR-901 conducts, developing a negative voltage at the anode. This negative voltage is filtered and applied to the PA PLATE RF meter, M-1003, which is calibrated to read RF volts.

(13) the PA MONITOR voltage divider is similar to the IPA MONITOR voltage divider employed in the cathode circuit. A sample of the plate RF is made available at J-900 for monitoring and testing purposes.

(14) the ALDC ADJUST potentiometer is a front panel control. One side of this pot is returned to a positive voltage; the wiper of this control varies the bias on ALDC diode CR-900. The sample of the RF voltage at the plate of the tube is further reduced by capacitor voltage divider C-907, C-904. When the peak value of the RF signal exceeds the bias on CR-900, this diode conducts, developing a negative voltage, which is filtered and applied to ALDC switch S-1003. When ALDC is utilized, this negative voltage is returned to a previous stage, to control the signal amplitude.

(15) the RF signal at the plate is applied to a PI network consisting of TUNE capacitor C-927, the portion of inductances L-902 and L-903 inserted by the bandswitch, and LOAD capacitor C-928. The plate supply voltage, 7500 volts, is applied at the junction of these capacitors through extensive filtering circuits. Capacitor C-930 places the junction of the TUNE and LOAD capacitor at effective RF ground. A simplified sketch of the arrangement is shown in Figure 8-2A.

PI output networks will be discussed in detail later in this chapter; for now, suffice it to say that the network acts as a resonant plate tank and assists in matching the impedance of the tube to the load impedance.

(16) the output circuits may be connected for either a 600 ohm balanced load or a 70 ohm unbalanced load. Figures 8-2B and 8-2C show these connections.

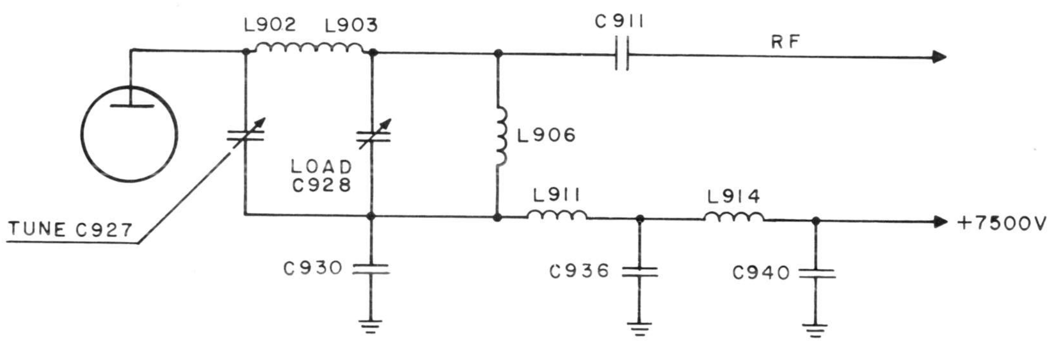


Figure 8-2A.

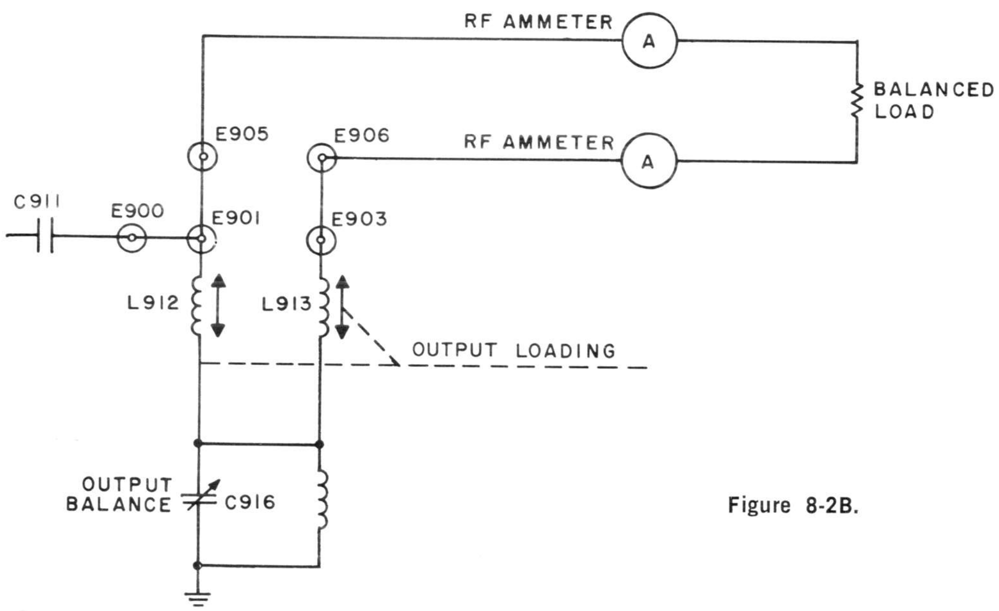
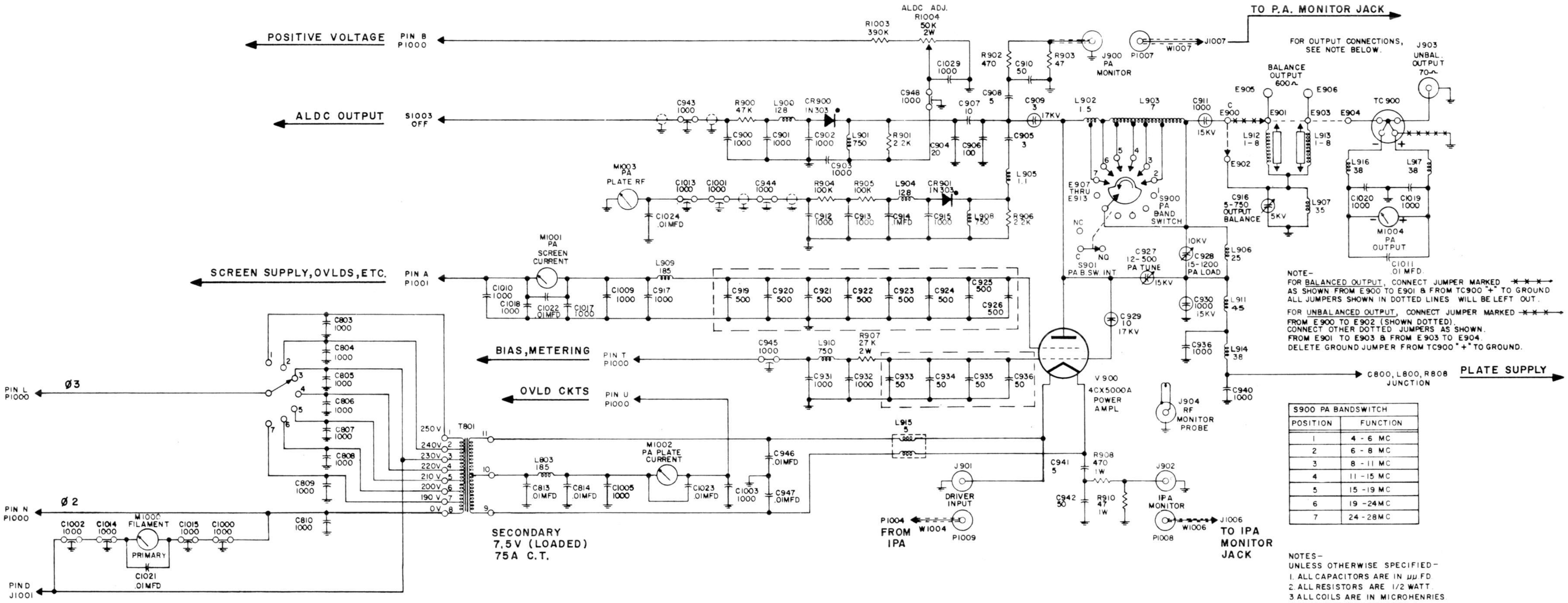


Figure 8-2B.

Refer to Figure 8-2B: (Balanced Output)
the RF output from C-911 is applied to L-912 which is mutually coupled to L-913. L-912 and L-913 are inductances, variable from 1 to 8 uh by means of ferrite slugs ganged to the OUTPUT LOADING control. Actually, L-912, C-916 and L-913 form a "T" network. Currents are caused to flow in and out of terminals E-905, E-906 180 degrees out of phase. The balanced output terminals connect to external RF ammeters, in order that the current in each leg of the balanced transmission line may be monitored. C-916, the OUTPUT BALANCE control, is adjusted for equal currents in the transmission line.
Refer to Figure 8-2C: (Unbalanced Output)

the RF from C-911 is applied to an "L" network consisting of the OUTPUT BALANCE capacitor, C-916, and the two ferrite slugged coils, L-912 and L-913 in parallel. The output is



NOTE-
FOR BALANCED OUTPUT, CONNECT JUMPER MARKED -×-×- AS SHOWN FROM E900 TO E901 & FROM TC900 "+" TO GROUND. ALL JUMPERS SHOWN IN DOTTED LINES WILL BE LEFT OUT.
FOR UNBALANCED OUTPUT, CONNECT JUMPER MARKED -×-×- FROM E900 TO E902 (SHOWN DOTTED). CONNECT OTHER DOTTED JUMPERS AS SHOWN. FROM E901 TO E903 & FROM E903 TO E904. DELETE GROUND JUMPER FROM TC900 "+" TO GROUND.

S900 PA BANDSWITCH

POSITION	FUNCTION
1	4 - 6 MC
2	6 - 8 MC
3	8 - 11 MC
4	11 - 15 MC
5	15 - 19 MC
6	19 - 24 MC
7	24 - 28 MC

NOTES-
UNLESS OTHERWISE SPECIFIED-
1. ALL CAPACITORS ARE IN μμFD
2. ALL RESISTORS ARE 1/2 WATT
3. ALL COILS ARE IN MICROHENRIES
4. S900 SHOWN IN POSITION 1
5. S901 IS P/O INTERLOCK SYSTEM SEE MASTER SCHEM (CK-403) FOR CONNECTIONS.
6. THE 800 & 1000 SYMBOL SERIES SHOWN ARE NOT P/O P.A.

applied to unbalanced output coaxial jack J-903. Thermocouple TC-900 samples the RF heating at this point, and developes a potential, which is applied to PA OUTPUT meter M-1004, which is calibrated in RF amperes. 8.4 amperes into a resistive 70 ohm load connected at J-903 represents an average power of 5 KW. The positive terminal of the thermocouple is grounded when the balanced output connections are made.

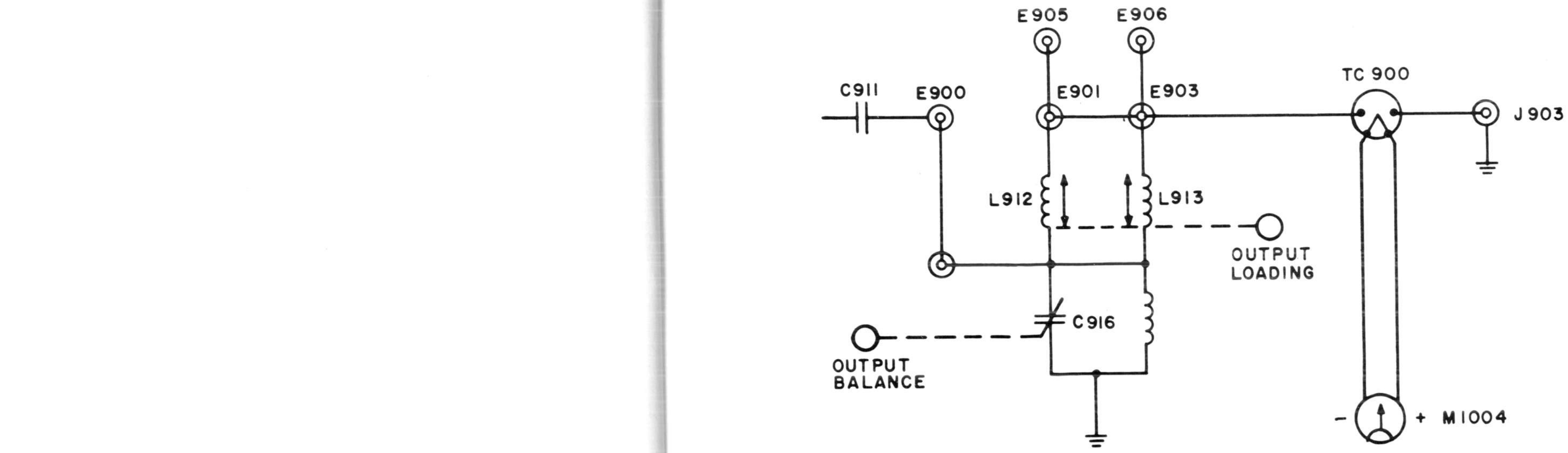


Figure 8-2C.

8-4 Power Control, Safety and Overload Circuits

a) the number and complexity of power control, safety and overload circuits depends on the power of the transmitter, its size, and the general service for which it is intended. A "Ham" transmitter of 30 watts power may require no particular safety or overload circuits; a 40 KW device may contain safety, power control and overload circuits equal in number and complexity to the transmitter circuits themselves.

b) power control circuits are required to turn on and off certain voltages, often in a definite sequence. For example: a time delay is usually incorporated in a transmitter using mercury vapor rectifiers; a definite time must elapse between the application of the filament voltage and the application of the stepped up primary voltage to the plates.

c) safety circuits are necessary to shut down certain devices and remove power when the safety of personnel or equipment is threatened. For example: a "personnel" interlock might shut down the high voltage circuits whenever a door or drawer on the transmitter is opened. An "equipment" interlock might shut down the plate and screen voltages of a tube when the system cooling the tube fails.

d) overload circuits are necessary to shut down certain circuits when the power, voltage, current or heat becomes excessive. For example: an overload circuit might remove plate voltage when the plate current exceeds a set value.

e) certain control devices in transmitters incorporate two or more features of power control, safety or overload. For example: a safety relay might, after performing its function, actuate a power control relay.

f) figure 8-3 through 8-8 illustrate certain power control, safety and overload features of the transmitter illustrated in Figure 8-2. This group of sketches by no means illustrates all of the control and safety features of the transmitter; this is merely a selection of circuits designed to illustrate the basic aspects of safety and power control.

PA BIAS RELAY (Refer to Figure 8-3)

This relay, symbol K-700, is grounded at terminal #1 and receives, at its other terminal, something less than —300 volts from the PA bias supply. A 15 K ohm resistor, R-700, limits the relay current. The relay is shown in the "NO BIAS" condition. In this condition, the PA bias supply is not furnishing the —300 volts required. One set of contacts causes a PA BIAS indicator, I-700, to be illuminated. Three sets of parallel connected contacts remain open, interrupting a series interlock chain which must be closed before the high voltage can be turned on. Thus, the high voltage cannot be brought up until the PA bias is applied.

When the PA bias circuit is operating, furnishing the required —300 volts, a bias, variable from about —200 to —300 volts is applied to the control grid of the final power amplifier tube. The bias may be adjusted by means of R-703, the PA BIAS ADJUST; the value of bias is indicated on PA BIAS meter M-3001. With the bias applied, K-700 energizes, causing the PA BIAS indicator to go out, and closing this one link in the series interlock chain.

Should the bias be lost at any time, the series interlock chain is opened, and the high voltage circuits are disabled. Loss of bias is indicated by the lighting of the PA BIAS lamp.

PA PLATE OVERLOAD RELAY (Refer to Figure 8-4)

This relay, K-701, has two solenoids: an OVERLOAD solenoid and a RESET solenoid. The overload solenoid causes the relay to "trip" when the current exceeds 2 amperes. This opens the series interlock chain, disabling the high voltage circuits. The reset coil is connected to a RESET switch and a source of voltage. When the RESET switch is depressed, the overload relay armature is returned to its former position. The high voltage circuit breaker must be thrown to the ON position to re-apply high voltage.

The relay is shown in the RESET position. In this position, the PA OVERLOAD indicator, I-701, is not lighted, and three paralleled sets of contacts are closing this link in the series interlock chain to enable the high voltage circuits. Total final tube current is flowing from ground,

through the parallel combination of the overload relay and R-704, R-705, through the plate current meter, M-1002, to the filament (cathode) of the final power amplifier tube. Normal plate current fully loaded is about 1.75 amperes. When this current exceeds two amperes, the overload solenoid trips the relay; the PLATE OVERLOAD indicator lights, and the series interlock chain is opened, causing the high voltage to be removed.

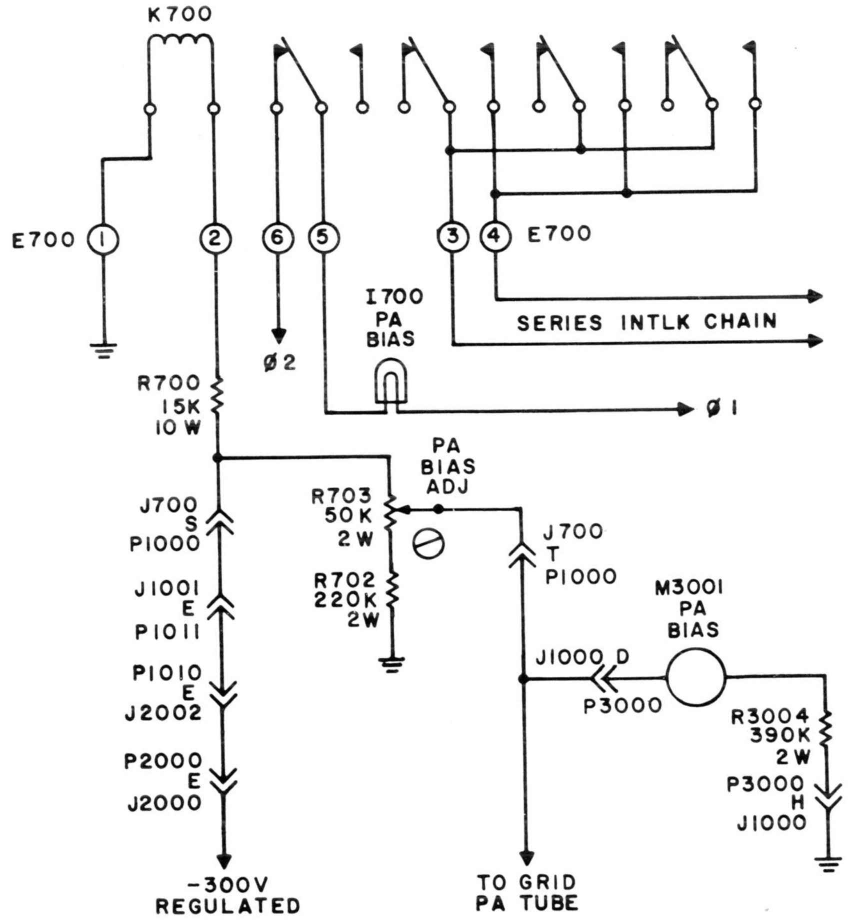


Figure 8-3. PA bias relay.

PA SCREEN ON OFF RELAY (Refer to Figure 8-5)

K-703, the PA SCREEN ON OFF relay, is shown de-energized in the OFF position. S-1005, the PA SCREEN ON OFF switch on the transmitter front panel, applies power to this relay via F-703, the REAR FAN fuse. The rear fan cools a zener diode regulator board, which regulates the screen voltage. With the rear fan fuse blown, it is impossible to apply screen voltage. With S-1005 in the ON position, K-703 energizes, applying 600 or 1200 volts to the final screen grid, via the PA SCREEN OVERLOAD circuit. Thus, the screen may be turned on and off independently of the plate, but plate voltage must be applied before screen voltage is available.

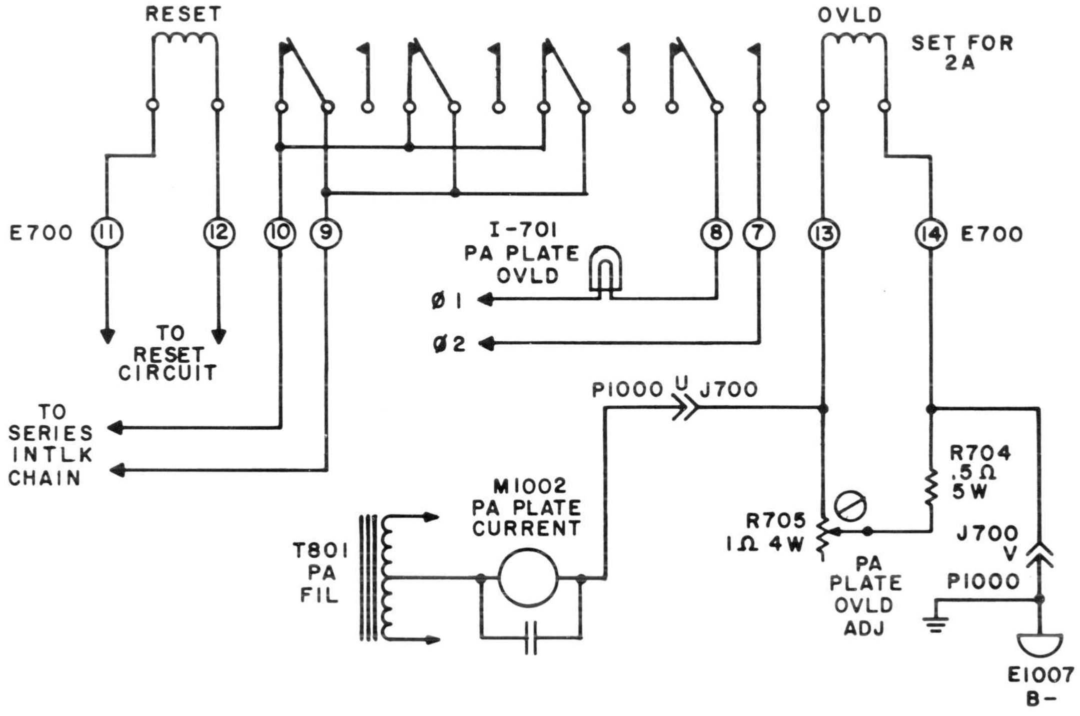


Figure 8-4. PA plate overload relay shown in reset position.

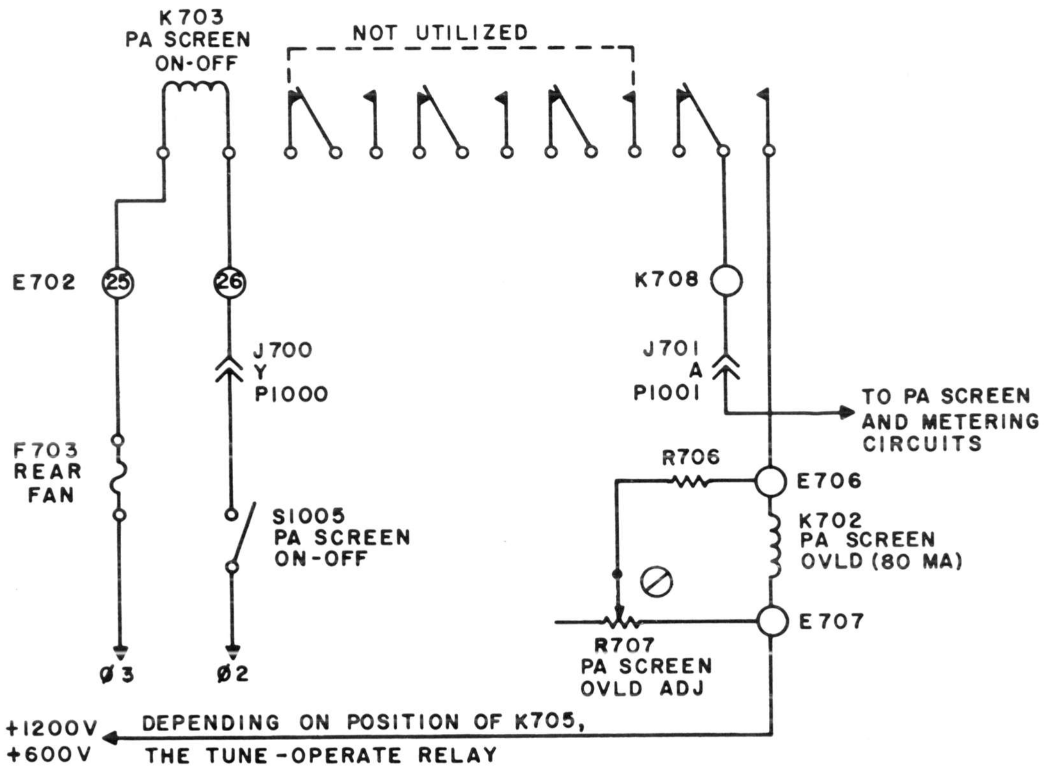


Figure 8-5. PA screen on off relay shown in off position

PA SCREEN OVERLOAD RELAY (Refer to Figure 8-6)

K-702, the PA SCREEN OVERLOAD relay, has two solenoids; one for overload and one for reset. The relay is shown in the reset position. Three sets of paralleled contacts are closing this link in the series interlock chain, and I-702, the PA SCREEN OVERLOAD indicator, is extinguished. Current flows from the PA screen grid, through the PA screen current meter, the PA screen ON OFF relay contacts, the parallel circuit consisting of K-702, R-706 and R-707, to either a 600 or 1200 volt source.

R-706 is adjusted so that, with 80 milliamperes from the screen grid, K-702 overload armature trips. This lights the PA SCREEN OVERLOAD lamp I-702, and opens the series interlock chain, disabling the high voltage circuits. The reset solenoid permits resetting the relay. After resetting, the high voltage circuit breaker must be thrown ON to re-apply the high voltage.

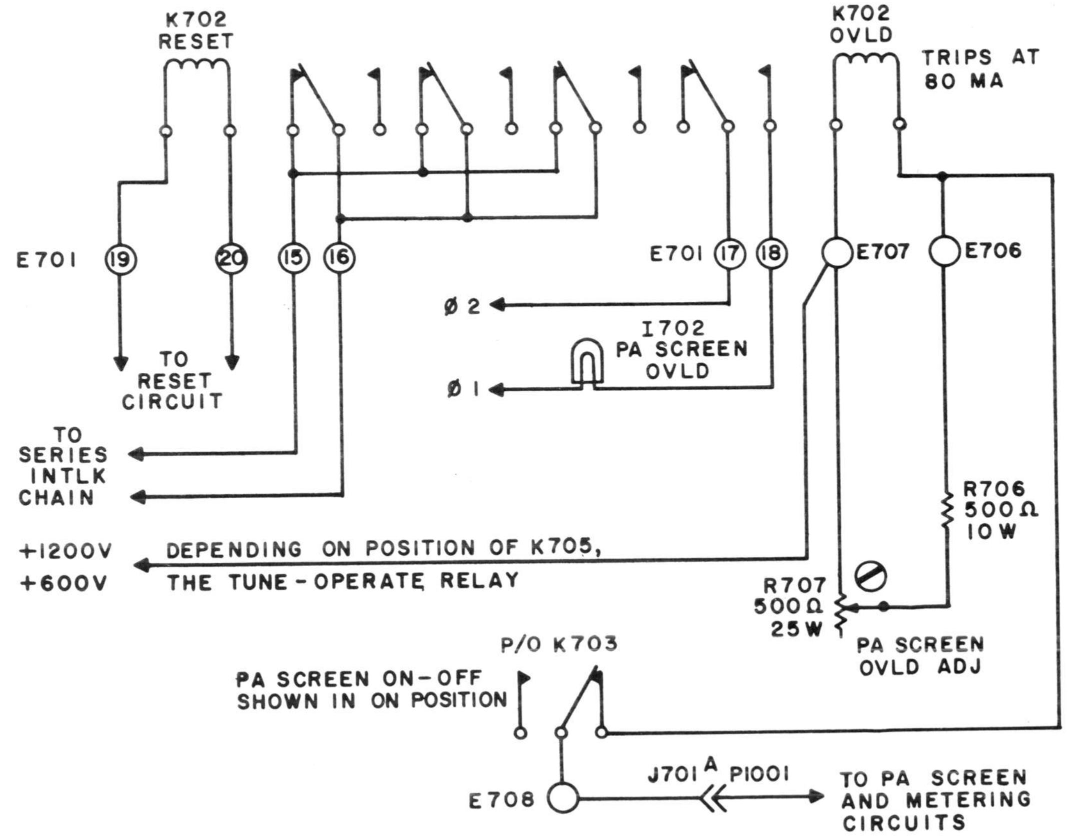


Figure 8-6. PA screen overload relay shown in reset position.

DIODE PROTECT RELAY (Refer to Figure 8-7)

Terminal board TB-800 is a zener diode regulator board, cooled by the REAR FAN. This is the board referred to previously. The diodes act to regulate the screen voltage to 1200 volts in OPERATE and 600 volts in the TUNE, condition. An inverse current flows from ground, through the parallel combination of the DIODE PROTECT OVERLOAD solenoid and R-701. Should the current exceed 80 ma., the overload solenoid trips, opening the series inerlock chain and removing high voltage. This prevents "avalanche" breakdown of the diodes. A reset solenoid is also incorporated.

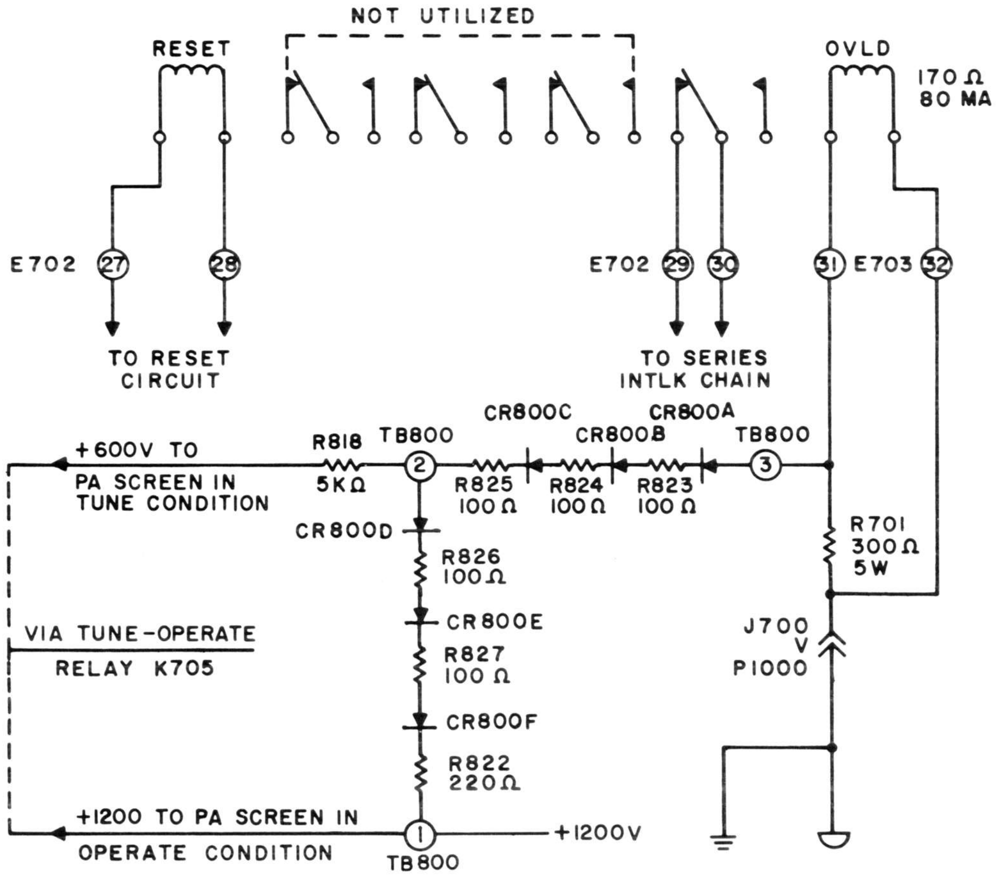


Figure 8-7. Diode protect relay shown in reset position.

OVERLOAD RESET CIRCUITS (Refer to Figure 8-8)

Figure 8-8 shows the RESET solenoids of several overload relays. When S-1000, the RESET button on the transmitter front panel is depressed, the tripped overload solenoids are reset. Note that, if the REAR FAN fuse F-703 is blown, it is impossible to reset the coils and high voltage cannot be applied. The REAR FAN cools the zener diode regulator board referred to previously.

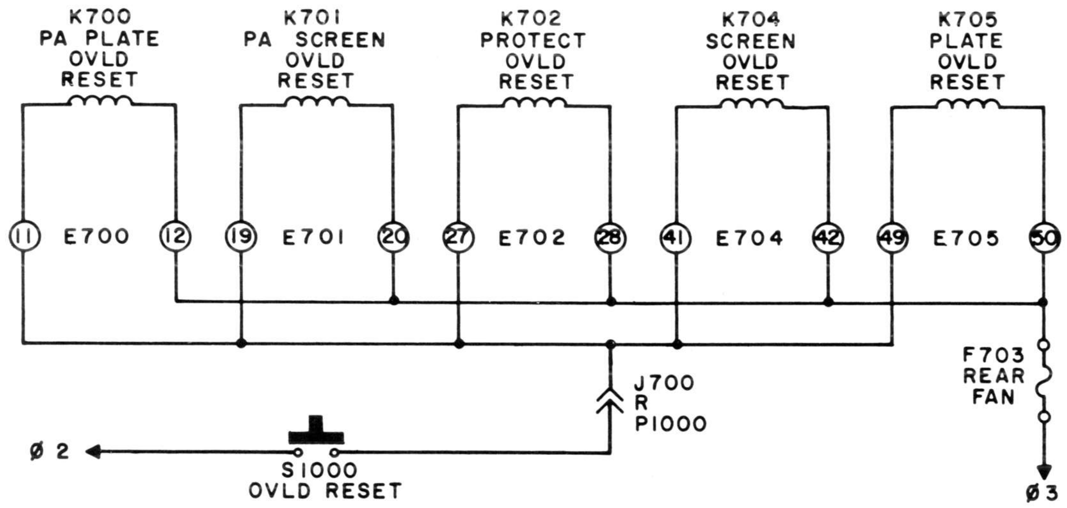


Figure 8-8. Overload reset circuits.

8-5 PI Output Networks for Transmitters

THE "MAXIMUM TRANSFER OF POWER" THEOREM

a) the "Maximum Transfer of Power" theorem states that a *maximum transfer* of power will be realized when the impedance of a load is equal to the impedance of the source supplying the load.

b) the statement in paragraph (a) above is greatly simplified; actually, the power transfer theorem states: "THE POWER TRANSFERRED TO A NETWORK, FROM A GENERATOR OF FINITE INTERNAL IMPEDANCE, IS THE MAXIMUM POSSIBLE WHEN THE INPUT IMPEDANCE OF THE NETWORK IS THE CONJUGATE OF THE GENERATOR IMPEDANCE."
This implies that any reactive component of generator impedance is neutralized by a reactance of a different kind at the input to the load. Since we will concern ourselves with resistive generator impedances, the statement of paragraph (a) will be quite correct.

c) consider the circuit shown in Figure 8-9:

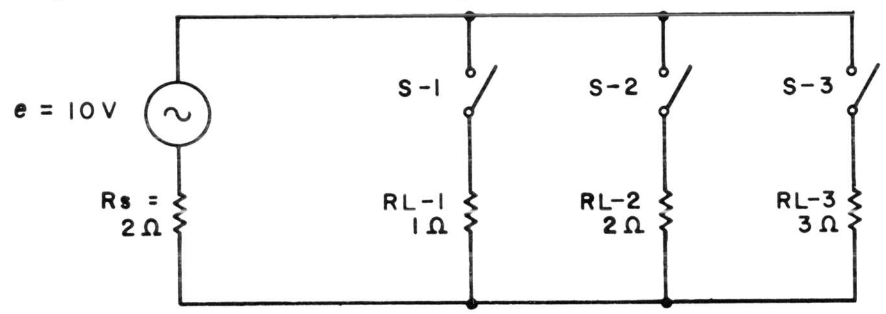


Figure 8-9.

(1) if S-1 is closed, the load resistance is 1 ohm.

$$i = \frac{e}{R_S + R_L} = \frac{10}{2+1} = \frac{10}{3} = 3.333 \text{ AMPS.}$$

total power dissipated is: $EI = 10 \times 3.333 = 33.33$ W.

load power is: $i^2R_L = (3.333)^2 \times 1 = 11.1$ W

power lost in the generator is: $i^2R_S = 22.2$ W

total power is: $P_{GEN} + P_{LOAD} = 22.2 + 11.1 = 33.3$ W

(2) if S-1 is opened and S-2 is closed:

$$i = \frac{E}{R_S + R_L} = \frac{10}{2+2} = \frac{10}{4} = 2.5 \text{ AMPS}$$

total power is: $EI = 10 \times 2.5 = 25$ W

load power is: $i^2R_L = (2.5)^2 2 = 12.5$ W

source power is: $i^2R_S = 12.5$ W

total power is: $P_S + P_L = 25$ W

(3) if S-2 is opened and S-3 is closed:

$$i = \frac{E}{R_S + R_L} = \frac{10}{2+3} = \frac{10}{5} = 2 \text{ AMPS}$$

total power is: $EI = 10 \times 2 = 20$ W

load power is: $i^2R_L = 2^2 \times 3 = 12$ W

source power is: $i^2R_S = 2^2 \times 2 = 8$ W

total power is: $P_S + P_L = 20$ W

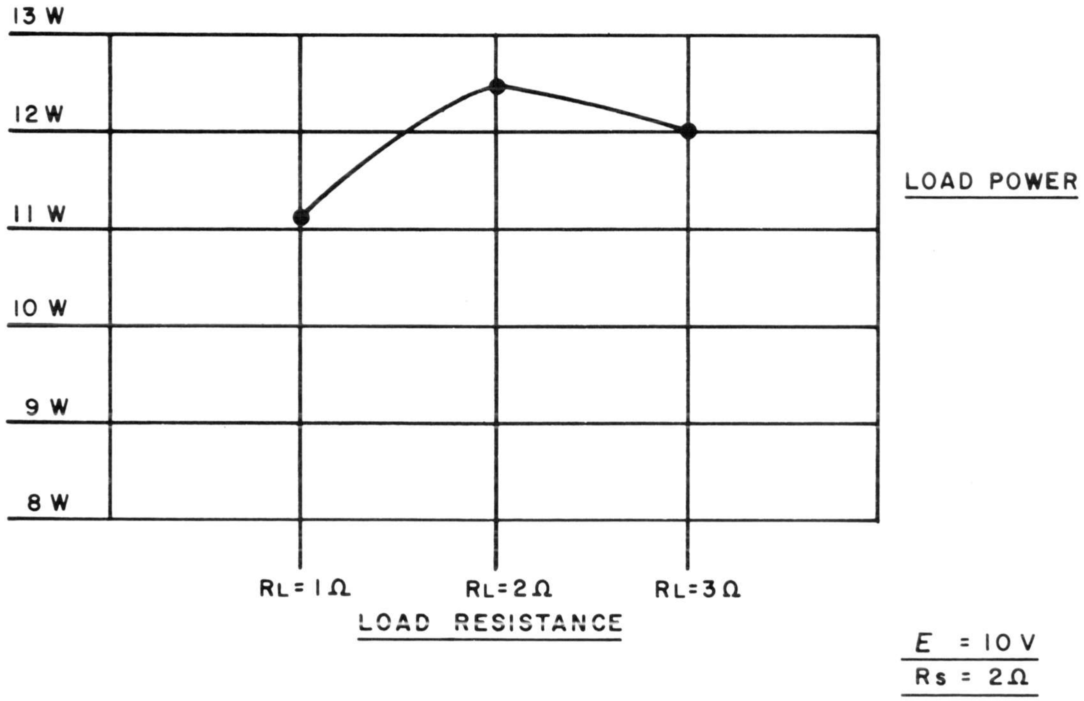


Figure 8-10.

d) in Figure 8-10, load power is plotted against load impedance, for a constant value of voltage and impedance at the source. Note that, when the resistance of the load is equal to the resistance of the source, the power transferred from source to load is maximum.

THE BASIC PROBLEMS OF MATCHING THE OUTPUT OF A POWER AMPLIFIER TO AN ANTENNA, DUMMY LOAD OR TRANSMISSION LINE

a) the final linear power amplifier will present to the output circuits a certain plate impedance; this impedance may be subject to change; it will be mainly resistive. The value of this impedance will depend principally on the tube type and the circuit configuration; for a grounded grid tetrode, it will be in the neighborhood of from 1 K ohm to 2500 ohms. For our examples, a resistive plate impedance of 1500 ohms will be specified.

b) the antenna, transmission line or load into which the transmitter output is to be "dumped" may present an impedance in the range between 50 ohms and 800 ohms. This impedance may be resistive, capacitive, or inductive. It may be fixed or variable. It may be changed when the transmitter frequency is changed. For our examples, a resistive load of 70 ohms will be specified.

c) for maximum transfer of power, then, a network must be connected between plate and load which will cause the plate to "see" 1500 ohms and the load to "see" 70 ohms.

d) in addition to power transfer, the network between plate and antenna:
   (1) must be capable of operation over a wide range of frequencies, if a multi-frequency transmitter is employed.
   (2) must be capable of filtering and harmonic suppression, if a minimum of output circuitry is desired.
   (3) must pass equally well all intelligence frequencies in the pass-band under consideration.
   (4) must have a "Q" of at least 12 if advantage is to be taken of the "flywheel effect".

THE BASIC PI NETWORK

a) figure 8-11 shows a basic PI network, connected between a source with an impedance of 1500 ohms, and a load resistance of 70 ohms. The source and its impedance, of course, represents the final vacuum tube in the transmitter, and the load resistor of 70 ohms represents the antenna, dummy load or transmission line into which the transmitter is working.

b) the circuit of figure 8-11 will be analyzed with conventional AC theory and vector algebra. First, the parallel combination of C-2, R_L will be transformed into an equivalent circuit, consisting of a resistor and a capacitor in series.

c) it can be shown that a series circuit and a parallel circuit are equiva-

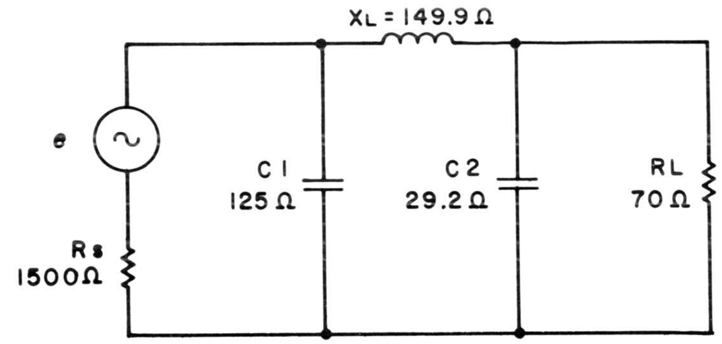


Figure 8-11.

lent, with the same Q and phase angle (same kind of reactance) when:

Q (series) = Q (parallel)

$$\text{where: Q series} = \frac{X_{SER}}{R_{SER}} \text{ and Q Parallel} = \frac{R_{PAR}}{X_{PAR}}$$

and:

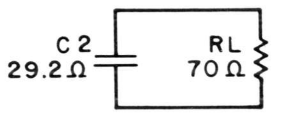
$$X \text{ series} = Q \cdot R \text{ series}$$

$$X \text{ parallel} = \frac{R \text{ parallel}}{Q}$$

$$R \text{ series} = \frac{R \text{ parallel}}{Q^2 + 1}$$

$$R \text{ parallel} = R_{SER}\,(Q^2 + 1)$$

Then:



$$Q = \frac{R_{PAR}}{X_{PAR}} = \frac{70}{29.2} = 2.4$$

$$\text{Then: R series} = \frac{R \text{ parallel}}{Q^2 + 1} = \frac{70}{6.76} = 10.35 \text{ ohms}$$

And: X series = Q × R series = 2.4 × 10.35 = 24.9 Ω

Thus: C-2 and R_L in parallel are exactly equivalent to a 10.35 ohm resistance in series with a capacitive reactance of 24.9 ohms. The two circuits are shown in Figure 8-12 below:

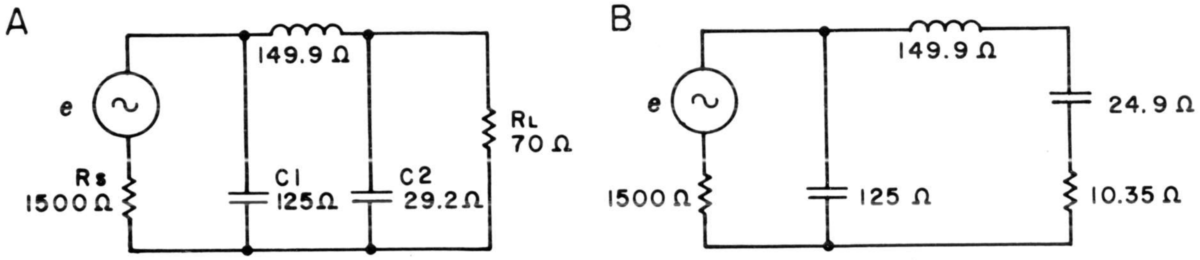


Figure 8-12.

In the right hand branch of Figure 8-12B, the inductive reactance will be partially neutralized by the capacitive reactance; in other words: the impedance of the right hand branch of figure 8-12B is:

$$R + j(X_L - X_C) =$$

$$10.35 + j(149.9 - 24.0) = 10.35 + j125\ \Omega$$

The new equivalent circuit is shown in Figure 8-13:

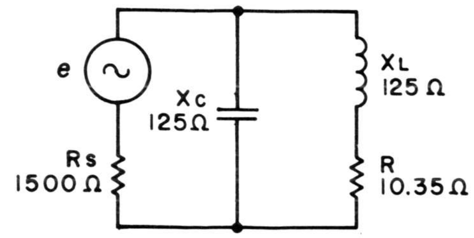


Figure 8-13.

Obviously, this is a parallel resonant circuit; inductive reactance equals capactive reactance; the Q of the circuit is:

$$Q = \frac{X_L}{R} = \frac{125}{10.35} = \cdot 12$$

the circuit, then, fulfills the requirements of a "high Q" parallel resonant circuit. The impedance will be resistive, and its value can be found from many formulas: among these,

$$Z_T = QX_L = 1500 \text{ ohms.}$$

Thus, the generator with a resistive impedance of 1500 ohms, is looking into a resistive load of 1500 ohms. The requirements for maximum transfer of power have been met.

d) assume that the generator voltage of Figure 8-14 is 1500 volts.

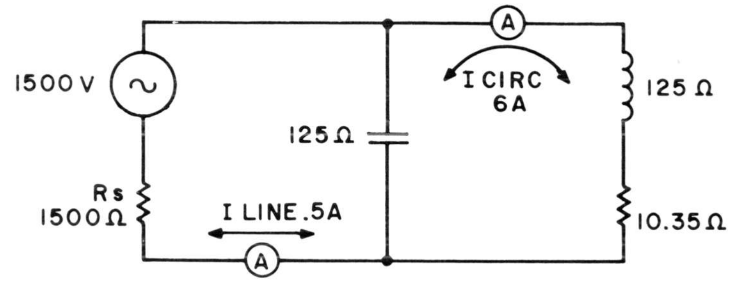


Figure 8-14.

$$I \text{ line} = \frac{e}{R_S + Z_T} = \frac{1500}{3000} = .5 \text{ amperes}$$

the tank circulating current, I circ, = Q I line = 12 × .5 = 6 amperes. Power consumed in the generator impedance will be:

$(I_{LINE})^2$ Rs $= (.5)^2 \times 1500 = .25 \times 1500 = 375$ watts.

Power consumed in the tank circuit resistance will be:

$(I_{CIRC})^2$ R $= 6^2 \times 10.35 = 36 \times 10.35 = 375$ watts.

thus: all of the requirements for maximum transfer of power have been fulfilled. The impedance of the source has been matched to the load impedance; the load, as far as the generator is concerned, is a resistance of 1500 ohms. The Q is sufficient to sustain the flywheel effect; a PI network, then, can be used in the plate circuit of a power amplifier for tuning and loading the transmitter.

e) suppose that this network must be used to cover a range of frequencies between 2 and 28 mcs. The input and output capacitors can be made variable, and the inductance can be tapped, and ganged to a bandswitch. It is only necessary to determine the minimum values of capacity and inductance, then the maximum values required.

2 MCS:

$$L = \frac{X_L}{2\pi f} = \frac{149.9}{6.28 \times 2 \times 10^6} = \frac{149.9}{12.56 \times 10^6} = 11.91\ \mu h$$

$$C_1 = \frac{159 \times 10^{-3}}{X_{C1}\ f} = \frac{159 \times 10^{-3}}{1.25 \times 10^2 \times 2 \times 10^6}$$

$$= \frac{159 \times 10^{-3}}{2.5 \times 10^8} = 63.6 \times 10^{-11} = 636\ \mu\mu f$$

$$C_2 = \frac{159 \times 10^{-3}}{X_{C2}\ f} = \frac{159 \times 10^{-3}}{29.2 \times 2 \times 10^6} = \frac{159 \times 10^{-3}}{58.4 \times 10^6}$$

$$= 2.73 \times 10^{-9} = 2730\ \mu\mu f$$

28 MCS:

$$L = \frac{X_L}{2\pi f} = \frac{149.9}{6.28 \times 2.8 \times 10^7} = 8.52 \times 10^{-7} = .852\ \mu h$$

$$C_1 = \frac{159 \times 10^{-3}}{X_{C1}\ f} = \frac{159 \times 10^{-3}}{1.25 \times 10^2 \times 2.8 \times 10^7}$$

$$= 45.4 \times 10^{-12} = 45.4\ \mu\mu f$$

$$C_2 = \frac{159 \times 10^{-3}}{X_{C2}\ f} = \frac{159 \times 10^{-3}}{2.92 \times 10^1 \times 2.8 \times 10^7}$$

$$= 19.45 \times 10^{-11} = 194.5\ \mu\mu f$$

DESIGN OF BASIC PI NETWORKS

a) *Problem:*

Design a PI network to match a power amplifier with an output impedance of 2000 ohms to a 70 ohm resistive load. A tank circuit Q of 15 is desired. Specify the value of C-1, C-2, and L at a frequency of 10 mcs.

b) Norton's Theorem:

WITH RESPECT TO ITS OUTPUT TERMINALS, ANY COMPLICATED LINEAR NETWORK CONTAINING EMF SOURCES IS EQUIVALENT TO A SOURCE OF CONSTANT CURRENT ACTING IN PARALLEL WITH AN IMPEDANCE, THE CONSTANT CURRENT BEING EQUAL TO THAT TRAVERSING THE OUTPUT TERMINALS WHEN SHORT CIRCUITED AND THE IMPEDANCE BEING EQUAL TO THE OUTPUT IMPEDANCE OF THE NETWORK.

c) in plain English, the theorem states that a voltage source in a linear network with a series internal impedance may be transformed into a current source with impedances in parallel.

(1) Figure 8-15 shows the constant voltage equivalent circuit for a vacuum tube amplifier and its load:

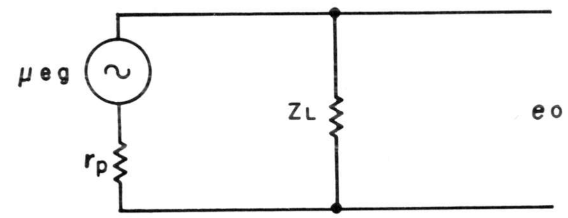


Figure 8-15.

(2) the current in the circuit is:

$$\frac{\mu eg}{rp + Z_L}$$

(3) the output voltage of the circuit is:

$$\frac{\mu eg \cdot Z_L}{rp + Z_L}$$

(4) if both numerator and denominator are divided by rp, the value of the equation is unchanged:

$$E_O = \frac{\frac{\mu eg}{rp}\ Z_L}{\frac{rp + Z_L}{rp}}$$

$\frac{u}{rp}$ = gm; therefore,

$$E_O = \frac{gm\ \ eg\ \ rp\ Z_L}{rp + Z_L}$$

(5) this would provide a new equivalent circuit, as shown in Figure 8-16 below:

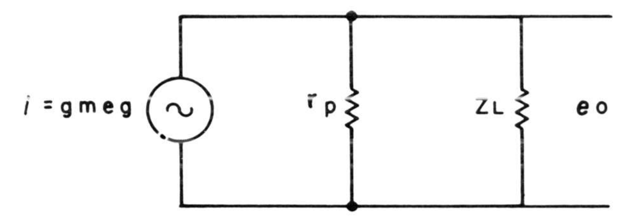


Figure 8-16.

d) the vacuum tube, then, and the PI network may be drawn as shown in Figure 8-17 below:

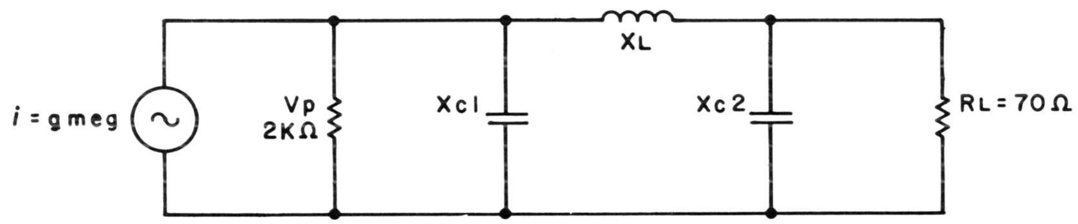


Figure 8-17.

e) the problem now involves matching a 2000 ohm resistance to a 70 ohm resistance. The specified "Q" of the final tank is 15.

$$X_{PAR} = \frac{R_{PAR}}{Q} = \frac{2000}{15} = 133.3 \text{ ohms.}$$

The value of X_{C1} then, is 133.3 ohms.

f) the equivalent series resistance, (or virtual resistance) of the final tank is:

$$R_{SER} = \frac{R_{PAR}}{Q^2 + 1} = \frac{2000}{226} = 8.84$$

this is the virtual resistance, or series resistance, of the final tank circuit; this is the same resistance described in Figure 8-12B.

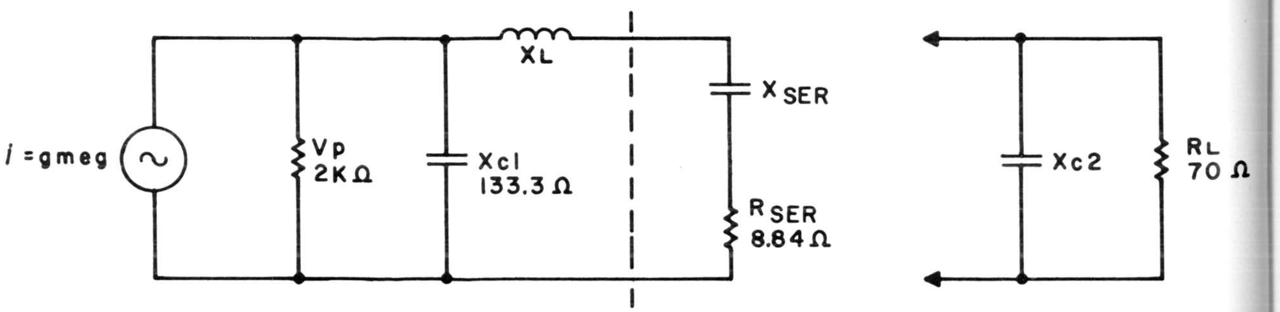


Figure 8-18.

g) we have yet to determine X_{SER}, X_L and X_{C2}.

h) previously, we transformed a parallel RC circuit into an equivalent series circuit. We will now do the same operation in reverse.

(1) if $R_{SER} = \frac{R_{PAR}}{Q^2 + 1}$ then $Q = \sqrt{\frac{R_{PAR}}{R_{SER}} - 1}$

$$Q = \sqrt{\frac{70}{8.84} - 1} = \sqrt{7.92 - 1} = \sqrt{6.92} = 2.63$$

(2) $X_{PAR} = \frac{R_{PAR}}{Q} = \frac{70}{2.63} = 26.6$ ohms.

(3) $X_{SER} = Q \cdot R_{SER} = 2.63 \times 8.84 = 23.3$ ohms.

(4) $X_L = X_{C1} + X_{SER} = 133.3 + 23.3 = 156.6$ ohms.

i) the complete network, with the equivalent series circuit, is shown in Figure 8-19 below:

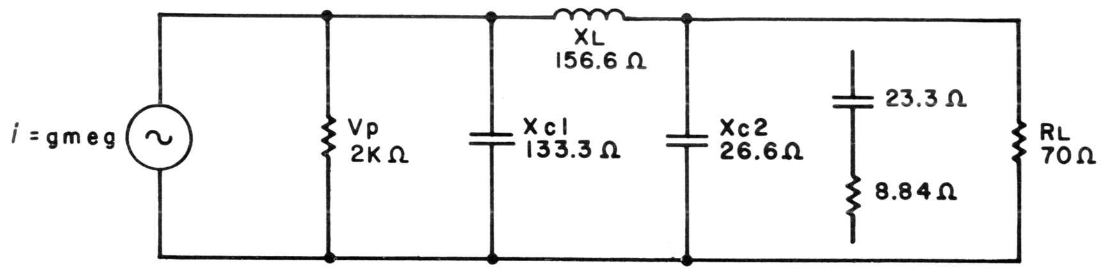


Figure 8-19.

if the equivalent series circuit is inserted, the circuit of figure 8-20 is obtained:

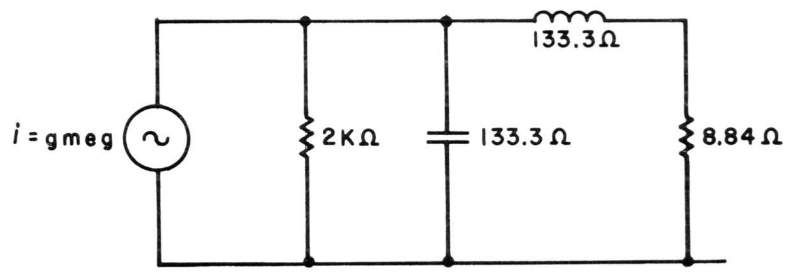


Figure 8-20.

$$Q_{TANK} = \frac{X_L}{R} = \frac{133.3}{8.84} = 15$$

$$Z_{TANK} = Q \cdot X_L = 15 \times 133.3 = 2000\Omega$$

At 10 MCS:

$X_{C1} = 133.3\Omega$

$X_{C2} = 26.6\Omega$

$X_L = 156.6\Omega$

$$C_1 = \frac{159 \times 10^{-3}}{f\, X_{C1}} = \frac{159 \times 10^{-3}}{1 \times 10^7 \times 1.333 \times 10^2}$$

$$= \frac{159 \times 10^{-3}}{1.333 \times 10^9} = 119\ \mu\mu f$$

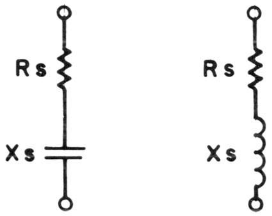
$$C_2 = \frac{159 \times 10^{-3}}{f\, X_{C2}} = \frac{159 \times 10^{-3}}{1 \times 10^7 \times 26.6} = \frac{159 \times 10^{-3}}{26.6 \times 10^7}$$

$$= 597\ \mu\mu f$$

$$L = \frac{X_L}{2\pi f} = \frac{156.6}{6.28 \times 1 \times 10^7} = \frac{156.6}{6.28 \times 10^7}$$
$$= 24.9 \times 10^{-7} = 2.49\ \mu h$$

DERIVATION OF THE FORMULAE USED IN THE DETERMINATION OF EQUIVALENT SERIES AND PARALLEL CIRCUITS

a) the absolute value of the impedance in a series circuit of the type shown below is:

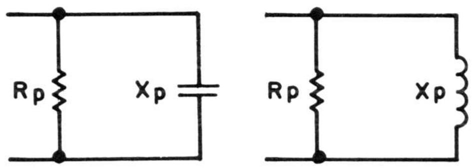


$$Z = \sqrt{R_s^2 + X_s^2} \qquad (1)$$

b) the "Q" of such a series circuit is:

$$Q = \frac{X_S}{R_S} \qquad (2)$$

c) the absolute value of the impedance in a parallel circuit of the type shown below is:



$$= \frac{R_P X_P}{\sqrt{R_P^2 + X_P^2}} \qquad (3)$$

d) the "Q" of such a parallel circuit is:

$$Q = \frac{R_P}{X_P} \qquad (4)$$

e) when a series circuit and a parallel circuit are exactly equivalent, both have the same impedance, Q, phase angle and reactance sign; if equations (1) and (3) are placed equal to each other, and equations (2) and (4) are used to substitute for X_P and X_S, the following resultants are obtained:

$$\frac{R_P}{R_S} = Q^2 + 1$$

$$Q = \sqrt{\frac{R_P}{R_S} - 1}$$

$$R_S = \frac{R_P}{Q^2 + 1}$$

$$R_P = R_S\,(Q^2 + 1)$$

Steps in the derivation:

$$\sqrt{R_s^2 + X_s^2} = \frac{R_P X_P}{\sqrt{R_P^2 + X_P^2}} \qquad (1)$$

$$\sqrt{R_s^2 + Q^2 R_s^2} = \frac{\dfrac{R_P^2}{Q}}{\sqrt{R_P^2 + \dfrac{R_P^2}{Q^2}}} \qquad (2)$$

$$\sqrt{R_s^2 + Q^2 R_s^2} = \frac{\dfrac{R_P^2}{Q}}{\sqrt{\dfrac{Q^2 R_P^2 + R_P^2}{Q^2}}} \qquad (3)$$

$$R_s^2(Q^2 + 1) = \frac{R_P^4}{Q^2} \times \frac{Q^2}{R_P^2(Q^2 + 1)} \qquad (4)$$

$$R_s^2(Q^2 + 1) = \frac{R_P^2}{(Q^2 + 1)} \qquad (5)$$

$$R_s^2(Q^2 + 1)^2 = R_P^2 \qquad (6)$$

$$Q^2 + 1 = \frac{R_P}{R_S} \qquad (7)$$

$$Q^2 = \frac{R_P}{R_S} - 1 \qquad (8)$$

$$Q = \sqrt{\frac{R_P}{R_S} - 1} \qquad (9)$$

CHAPTER 9

FILTERS FOR SINGLE SIDEBAND OPERATION

9-1 Introductory Note

Both Single Sideband transmitters and receivers require highly selective bandpass filters; these filters are generally designed, in commercial installations, for operation at frequencies from approximately 100 KC to 500 KC, although there are notable exceptions to this generalization. For example: the TMC Model SBE-2 Sideband Exciter employs sideband filters for operation at 17 KC.

In this chapter, the general construction of these filters will be discussed; terms associated with the response characteristics of filters will be explained.

9-2 Problems in the Design of Filters

The design of sideband filters is a compromise, as is usually the case in electronics. The basic problem involves passing a relatively narrow band of frequencies, and completely attenuating all other frequencies. Many factors in addition to the bandpass must be taken into account; among these are:

a) the input impedance of the filter.
b) the output impedance of the filter.
c) the insertion loss of the filter.
d) the passband ripple.
e) the shape factor.
f) the relative difficulty of designing a filter for the particular bandpass selected.
g) the bounceback.
h) the spurious response that may be encountered.
i) the required physical size and complexity.
j) the cost.
k) the amount of phase shift introduced by the filter.

9-3 Sideband Filter Classifications

Sideband filters may be classified in three general ways:

a) in accordance with the manner in which the filter is physically constructed.
b) in accordance with the shape of the response characteristic; that is, whether the filter is symmetrical or asymmetrical.
c) in accordance with the ratio of the bandwidth to the center frequency; that is, whether the filter has a narrow bandpass or a wide bandpass.

CLASSIFICATION IN ACCORDANCE WITH PHYSICAL CONSTRUCTION

a) *LC Filters:*
these are filters whose elements are made up of lumped and distributed inductance, capacitance, and resistance.
b) *Mechanical Filters:*
these are filters whose elements are mechanically resonant. Such filters receive electrical energy, convert it to mechanical energy and then cause the process to be reversed, to deliver electrical energy at the output.
c) *Crystal Filters:*
these are filters whose elements include, in addition to other parameters, one or more piezoelectric crystals.

CLASSIFICATION IN ACCORDANCE WITH THE SHAPE OF THE RESPONSE CHARACTERISTIC

a) a symmetrical filter is one whose response characteristic is a mirror image on either side of the center frequency.
b) an asymmetrical filter is one whose response characteristic has a different shape on either side of the center frequency.

CLASSIFICATION IN ACCORDANCE WITH THE RATIO OF BANDWIDTH TO CENTER FREQUENCY

a) filters may be classified as "narrow" or "wide" band, in accordance with the ratio of bandwidth to center frequency.

b) as far as crystal filters are concerned:
   (1) narrow band filters can be designed with bandwidths as narrow as .005% of center frequency, and up to .7%.
   (2) wide band filters can be designed with bandwidths from about .6% to about 10% of center frequency.

c) as far as mechanical filters are concerned, bandwidths from about 500 cycles to about 35 KC can be manufactured in the 100 KC - 500 KC range. This represents, at 250 KC, a ratio of bandwidth to center frequency of from .2% to 14%.

d) as far as LC filters are concerned, the center frequency is relatively low (about 20 KC), and may be as high as 50 KC. The Q of these filters cannot approach that of mechanical and crystal units; the bandpass of LC filters is usually considered to be wide. For example, a bandwidth of 4 KC at a center frequency of 20 KC is a ratio of 20%.

9-4 LC Filters

LC filters are finding limited application as sideband filters in modern commercial use. Effective LC filters can be constructed to work well at frequencies in the range 15 KC - 50 KC; however, this is still, for all practical purposes, in the audio range. Additional translation circuits are required to raise the initially generated sideband intelligence to a higher place in the spectrum.

The TMC Model SBE-2 Sideband Exciter is a typical example; the initial sideband generation is accomplished at 17 KC; the 17 KC balanced modulators are followed by 17 KC LC filters. A second translation raises the intelligence to a frequency in the vicinity of 270 KC. A third translation raises the intelligence to a variable frequency in the range 1.73 - 3.73 mcs. A final translation raises the intelligence to a frequency in the 2 - 32 mc spectrum. This involves four balanced modulator circuits, with attendant injection oscillator circuits.

In TMC's later Model SBE-3, crystal filters rather than LC filters are employed. The initial sideband generation, carrier suppression and filtering is carried out at 250 KC, which greatly simplifies the circuitry by elimination of one balanced moduator-frequency injection circuit.

In addition to the disadvantage of effectiveness at low frequencies only, LC filters are undesirable because of their relatively large size and weight.

An example of a "classic" LC bandpas filter is presented for interested readers. The filter is symmetrical, with a center frequency, fc, of 40 KC, a bandpass of 4 KC, and a characteristic impedance, R_O, of 10 K ohms. The source and load impedances are also 10 K ohms, resistive. The filter is shown in Figure 9-1. The response characteristic is shown in Figure 9-2.

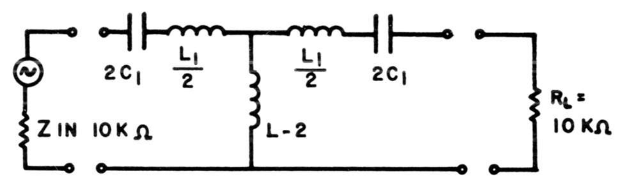


Figure 9-1.

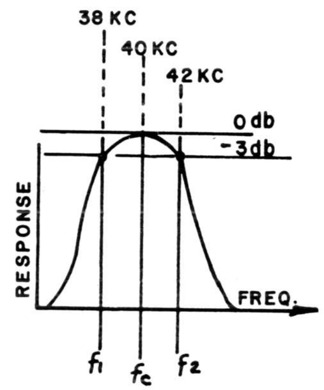


Figure 9-2.

Component values are determined as follows:

$$L_1 = \frac{R_O\, f_1}{\pi f_2 (f_2 - f_1)} = \frac{1 \times 10^4 \times 38 \times 10^3}{3.14 \times 4.2 \times 10^4 \times 4 \times 10^3}$$

$$= \frac{38 \times 10^7}{12.56 \times 4.2 \times 10^7} = .72 \text{ hy}$$

$$\frac{L_1}{2} = \frac{.72}{2} = .36 \text{ hy}$$

$$C_1 = \frac{(f_2 - f_1)}{4\pi\ f_1 f_2\ R_O}$$

$$= \frac{4 \times 10^3}{12.56 \times 3.8 \times 10^4 \times 4.2 \times 10^4 \times 1 \times 10^4}$$

$$= \frac{4 \times 10^3}{2 \times 10^{14}} = 20\ \mu\mu\text{f}$$

$$2C_1 = 40\ \mu\mu\text{f}$$

$$L_2 = \frac{R_O(f_2 + f_1)}{4\pi\ f_1\ f_2} = \frac{1 \times 10^4 \times 80 \times 10^3}{12.56 \times 3.8 \times 10^4 \times 4.2 \times 10^4}$$

$$= \frac{80 \times 10^7}{2 \times 10^{10}} = 40 \times 10^{-3} = 40 \text{ mhy}$$

Calculation of the Reactances at center frequency, 40 KC:

$$\frac{X_{L1}}{2} = 2\pi f\ \frac{L_1}{2} = 6.28 \times 4 \times 10^4 \times 3.6 \times 10^{-1} = 90.5 \text{ K}\Omega$$

$$2X_{C1} = \frac{1}{2\pi f\ 2_{C1}} = \frac{159 \times 10^{-3}}{4 \times 10^4 \times 4 \times 10^{-11}}$$

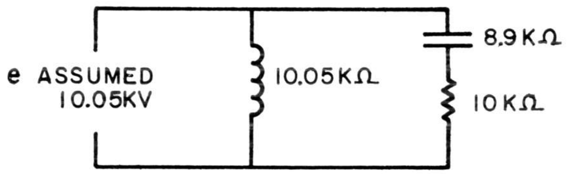
$$= 9.94 \times 10^4 = 99.4 \text{ K}\Omega$$

$$X_{L2} = 2\pi f\, L_2 = 6.28 \times 4 \times 10^4 \times 4 \times 10^{-2} = 100.5 \times 10^2 = 10.05 \text{ K}\Omega$$

Simplification of the Circuit to the right of L-2:

$$Z = (R - jX_C) = (10\ K\Omega - j\ 8.9\ K\Omega) = 13.4\ K\Omega\ \angle -41.7^\circ$$

Determining the total impedance of the combination just simplified, in parallel with L_2. The assumed voltage method will be used.



$Z = 13.4\ K\Omega\ \angle -41.7^\circ$

$$\text{Current in } L_2 = \frac{10.05\ KV\ \angle 0^\circ}{10.05\ K\Omega\ \angle 90^\circ} = 1a\ \angle -90^\circ = 0 - j\ 1a$$

$$\text{Current in RC Branch} = \frac{10.05\ KV\ \angle 0^\circ}{13.4\ K\Omega\ \angle -41.7^\circ} = .75a\ \angle 41.7^\circ = .56 + j\ .5a$$

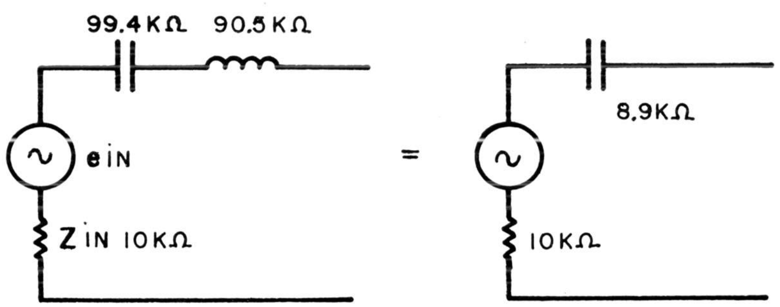
$$\begin{array}{lrl} \text{Adding} & 0 & -j\ 1 \\ \text{Vectorially} & .56 & +j\ .5 \\ \hline \text{Total Current} = & .56 & -j\ .5 = .75a\ \angle -41.7^\circ \end{array}$$

$$\text{Total Impedance} = \frac{10.05\ KV\ \angle 0^\circ}{.75a\ \angle -41.7^\circ} = 13.4\ K\Omega\ \angle 41.7^\circ$$

$13.4\ K\Omega\ \angle 41.7^\circ = 10K\Omega + j\ 8.9\ K =$ 

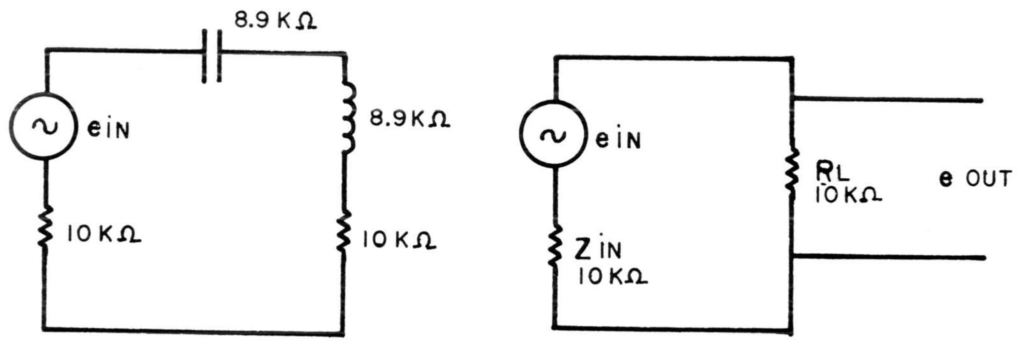
Note that the action of L_2 has transformed the circuit to the right, with an impedance of $13.4\ K\Omega\ \angle -41.7^\circ$ into an impedance of $13.4\ K\Omega\ \angle +41.7^\circ$ that is, into a conjugate impedance.

Simplification of the circuit to the left of L_2:



The circuit to the left of L_2 is exactly equivalent to the circuit to the right of L_2; therefore, the simplified circuits are the same.

Combining the simplified circuits:



Thus, at the center frequency, 40 KC, the circuit is purely resistive; the output voltage is half the input voltage; the impedance of the source equals the load impedance, and a maximum transfer of power is accomplished.

If the input voltage is 10 volts, the output voltage is 5 volts.

The filter will now be analyzed at the low frqeuency, f-1, which is 38 KC.

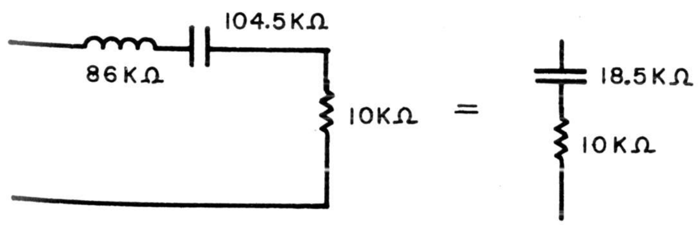
Determination of reactances at 38 KC:

$$\frac{X_{L1}}{2} = 2\pi f\ \frac{L_1}{2} = 6.28 \times 3.8 \times 10^4 \times 3.6 \times 10^{-1} = 86 \times 10^3$$
$$= 86\ K\Omega$$

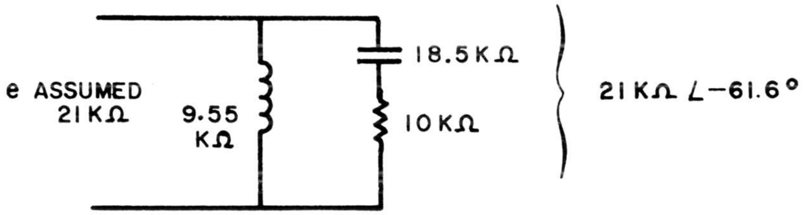
$$X2_{C1} = \frac{1}{2\pi f\ 2_{C1}} = \frac{159 \times 10^{-3}}{3.8 \times 10^4 \times 4 \times 10^{-11}}$$
$$= 10.45 \times 10^4 = 104.5\ K\Omega$$

$$X_{L2} = 2\pi f\ L_2 = 6.28 \times 3.8 \times 10^4 \times 4 \times 10^{-2} = 95.5 \times 10^2$$
$$= 9.55\ K\Omega$$

Simplification of the circuit to the right of L_2:



Determining the total impedance of the combination just simplified, in parallel with L_2. The assumed voltage method will be used.



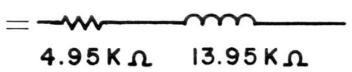
$$\text{Current in } L_2 = \frac{21.0 \text{ KV } \angle 0^\circ}{9.55 \text{ K}\Omega \ \angle 90^\circ} = 2.2a \angle -90^\circ = 0 - j\ 2.22a$$

$$\text{Current in RC Combination} = \frac{21.0 \text{ KV } \angle 0^\circ}{21 \text{ K}\Omega \ \angle -61.6^\circ} = 1a \angle 61.6^\circ = .475 + j\ .88a$$

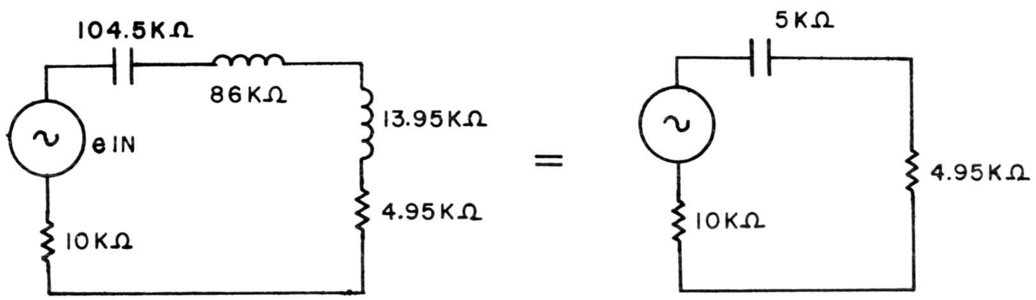
$$\begin{array}{lrcl} \text{Adding} & 0 & -j & 2.22 \\ \text{Vectorially} & .475 & +j & .88 \\ \hline \text{Total Current} = & .475 & -j & 1.34\ a = 1.42 \angle -70.5^\circ a \end{array}$$

$$\text{Total Impedance} = \frac{21 \text{ KV } \angle 0^\circ}{1.42a \ \angle -70^\circ} = 14.8 \ \angle +70.5^\circ \text{ K}\Omega$$

$$14.8 \text{ K}\Omega \ \angle 70.5^\circ = \underset{4.95 \text{ K}\Omega}{4.95 \text{ K}\Omega} + j\ \underset{13.95 \text{ K}\Omega}{13.95 \text{ K}\Omega} =$$



Simplification of the entire circuit:



$$Z_T = R - j\ X_C = 14.95 \text{ K}\Omega - j\ 5 \text{ K}\Omega = 15.75 \text{ K}\Omega \ \angle -18.5^\circ$$

$$I_t = \frac{e_{IN}}{Z_T} = \frac{10V \ \angle 0^\circ}{15.75 \text{ K}\Omega \angle -18.5^\circ} = .635 \text{ ma} \angle 18.5^\circ$$

$$e_{OUT} = it\ R_2 = .635 \text{ ma } \angle 18.5^\circ \times 4.95 \text{ K}\Omega \ \angle 0^\circ = 3.14 \text{ V}$$

At 40 KC, the center frequency, e_{IN} was 10 volts and e_{OUT} was 5 volts. At 38 KC, the low frequency, e_{IN} is 10 volts and e_{OUT} is 3.14 volts. The drop in db from center frequency to the low frequency is:

$$\text{db} = 20 \log \frac{e}{e} \ \therefore \ \text{db} = 20 \log \frac{5.00}{3.14}$$

$$\text{db} = 20 \log 1.59$$

$$\text{db} = 20 \times .202 = 4.040 \text{ db}$$

The filter will now be analyzed at the high frequency, which is 42 KC. Determination of reactances at 42 KC:

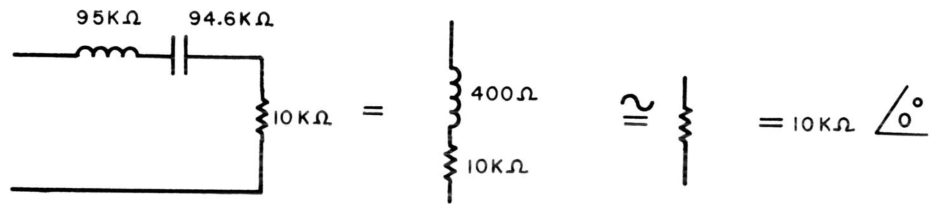
$$\frac{X_{L1}}{2} = 6.28 \times 4.2 \times 10^4 \times 3.6 \times 10^{-1} = 95 \times 10 = 95 \text{ K}\Omega$$

$$X2_{C1} = \frac{159 \times 10^{-3}}{4.2 \times 10^4 \times 4 \times 10^{-11}} = \frac{159 \times 10^{-3}}{16.8 \times 10^{7-}}$$

$$= 9.46 \times 10^4 = 94.6 \text{ K}\Omega$$

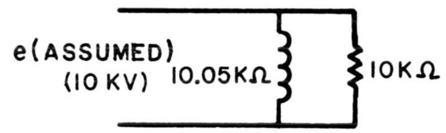
$$X_{L2} = 6.28 \times 4.2 \times 10^4 \times 4 \times 10^{-2} = 105.5 \times 10^2 = .10.55 \text{ K}\Omega$$

Simplification of the circuit to the right of L-2:



400 ohms of inductive reactance is so small in comparison with 10 K ohms of resistance that the latter is considered to be the total impedance.

Determining the total impedance of the combination just simplified, in parallel with L-2: the assumed voltage method will be used.



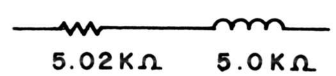
$$I_L = \frac{10 \text{ KV } \angle 0^\circ}{10.05 \text{ K}\Omega \angle 90^\circ} = .994a \angle -90^\circ = 0 - j\ .994$$

$$I_R = \frac{10 \text{ KV} \angle 0^\circ}{10 \text{ K}\Omega \angle 0^\circ} = 1a \ \angle 0^\circ \quad 1 + j \ \ 0$$

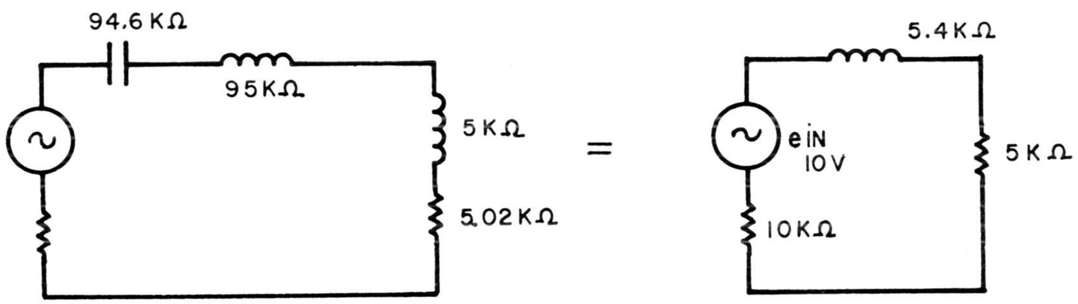
$$I_t = 1 - j\ .994a = 1.41a \angle -44.8^\circ$$

$$Z_T = \frac{10 \text{ KV} \angle 0^\circ}{1.41a \angle -44.8^\circ} = 7.08 \text{ K}\Omega \ \angle 44.8^\circ = 5.02 \text{ K}\Omega + J\ 5.0 \text{ K}\Omega$$

$$Z_T =$$



Anaylzing the complete simplified circuit:



$$Z_T = R + j\,X_L = 15\ \text{K}\Omega + j\ 5.4\ \text{K} = 15.9\ \text{K}\Omega\angle 19.8^\circ$$

$$I_t = \frac{10\ \text{V}\angle 0^\circ}{15.9\ \text{K}\Omega\ \angle 19.8^\circ} = .628\ \text{ma}\angle -19.8^\circ$$

$$e_{\text{OUT}} = I_t \times R_L = .628\ \text{ma}\angle -19.8^\circ \times 5\ \text{K}\Omega\angle 0^\circ = 3.14\ \text{V}$$

This is the same output voltage obtained at the low frequency; it therefore represents a 4 db drop from the output voltage at the center frequency.

The "shape factor" of a filter refers to its response characteristics at two different levels of attenuation; usually, the shape factor refers to the bandwidth at the 3 db points and the bandwidth at the 40, 50 or 60 db points. Figure 9-3 illustrates this point.

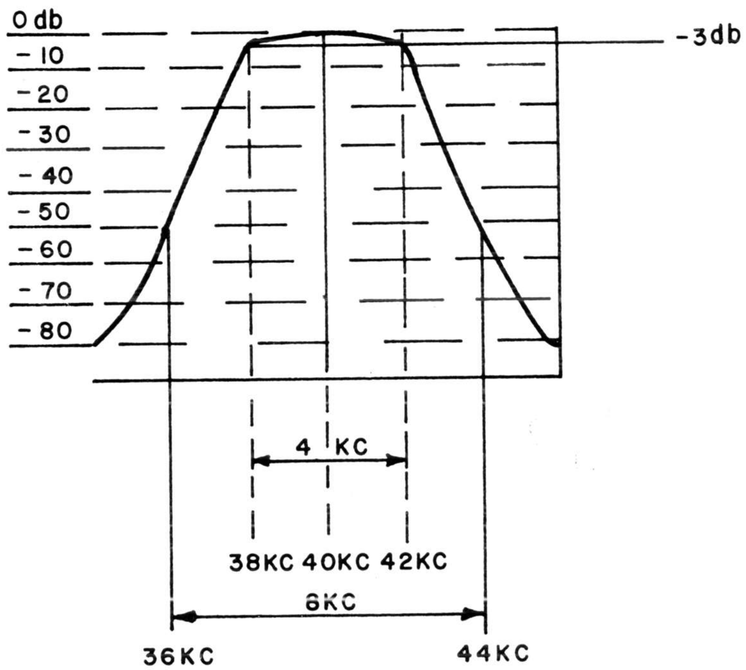


Figure 9-3.

For a perfect, ideal response characteristic, the shape factor should be 1.0; that is, the bandwidth at the 3 db points should equal the bandwidth at any other point of attenuation. The shape factor can be improved by addition of similar filter sections; for an LC filter, this increases the weight, size, and cost.

When complete carrier suppression is desired, the shape factor should be as close to unity as possible to obtain. Referring to Figure 9-3: if the carrier were at 37 KC, it would be down only 30 db.

9-5 Mechanical or Magnetostrictive Filters

Mechanical filters offer many advantages over LC filters; they are compact, have excellent rejection characteristics, and are comparatively rugged. The Q's obtainable are much, much greater than can be obtained with LC filters. In addition, they are physically smaller.

THE MAGNETOSTRICTIVE EFFECT

Certain materials exhibit a change in dimension when subjected to a magnetic field. The change of dimension may result in lengthening or shortening, depending on the type of material and the strength of the magnetic field. Nickel shows a marked magnetostrictive effiect; this substance is commonly found in Sonar transducers and mechanical filters using the magnetostrictive effect. A Nickel rod contracts in a magnetic field; the change in length is very nearly proportional to the magnetic field strength over a large range of magnetic density. It should be noted that Nickel contracts in a magnetic field without regard to the direction of the field.

Figure 9-4 shows a nickel rod upon which is wound a coil. The voltage source is an alternator having a sinusoidal output. Figure 9-5 shows a plot of the change of length of the rod with changes in the alternating voltage applied.

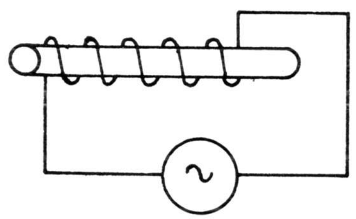
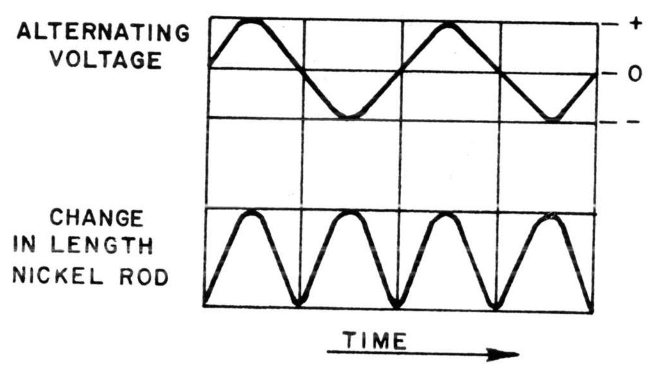


Figure 9-4.

Figure 9.5.



Figures 9-4 and 9-5 show that, because Nickel contracts in a magnetic field regardless of the direction of the field, the frequency of the induced rod vibration is twice that of the frequency of the applied voltage. This situation may or may not be desired.

When it is required that the frequency of mechanical vibration equal the frequency of the applied voltage, a polarizing voltage may be employed. This voltage will be of such magnitude that it produces a current greater than the peak alternating current. The polarizing voltage and the alternating voltage are applied to the coil in series, as shown in Figure 9-6, below. The same effect may be obtained by mounting "bias" magnets to create a static field, the strength of which is greater than the alternating field. This is shown in Figure 9-7. Figure 9-8 shows the resultant change in length of the Nickel rod, under conditions of polarization.

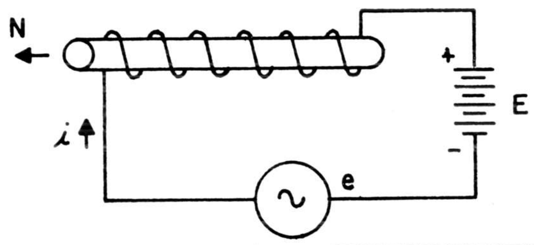


Figure 9-7.

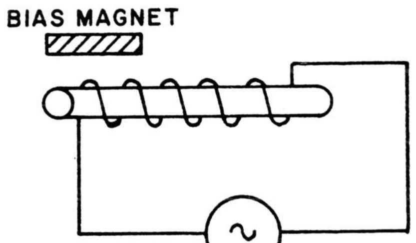


Figure 9-6.

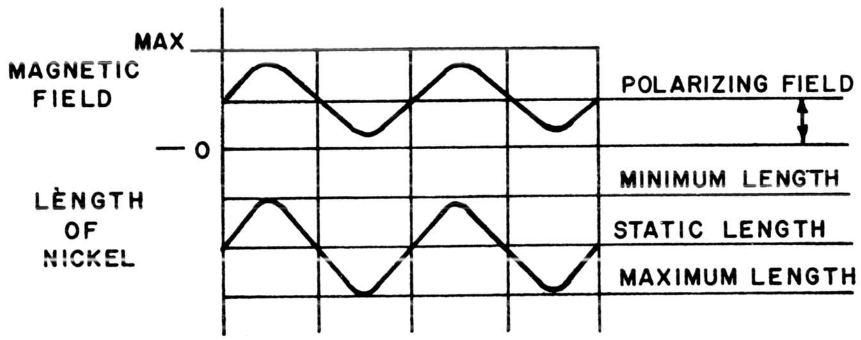


Figure 9-8.

AN ELEMENTARY MECHANICAL FILTER

Figure 9-9 below shows an elementary mechanical filter. The filter consists of an input transducer, nickel discs, connector rods, external capacitors, and an output transducer.

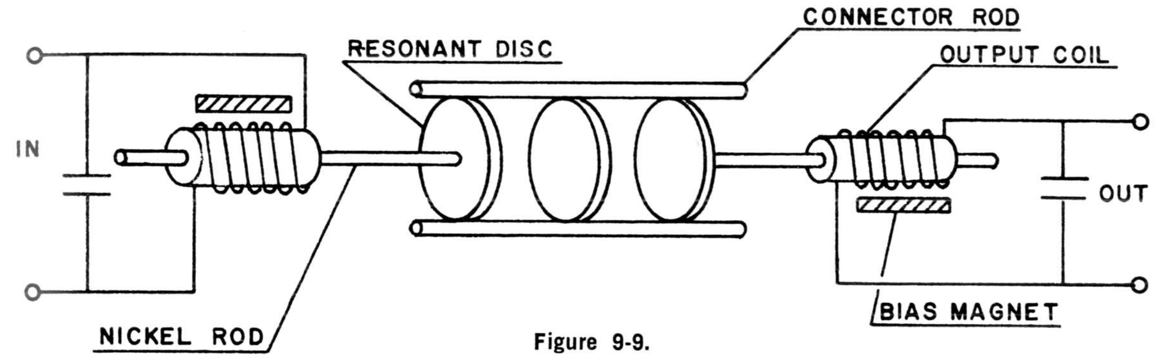


Figure 9-9.

The incoming signal is applied to the input transducer; this is composed of a coil within the filter and an external capacitor, made to resonate with the coil at the center frequency of the passband.

The capacitor may be connected for series or parallel resonance; Figure 9-9 shows it connected for the latter. The resonant current moves the transducer rod; this transmits motion to the resonant discs and the coupling rods. Each disc acts as a series resonant circuit; hence, the number of discs is important in determining the selectivity and shape factor. The area of the coupling rods primarily determines the bandpass. A terminal disc vibrates the output transducer rod, which induces, by means of the "generator" effect, a current in the output transducer coil. This coil is also resonated with an external capacitor to the center frequency of the passband.

The equivalent circuit of the mechanical filter is shown in Figure 9-10. C-1, L-1 represents a resonant disc; C-2 represents the coupling rods.

Note the resemblance of Figure 9-10 to Figure 9-1. They have the same general configuration, except that the mechanical filter has more sections for better shape factor.

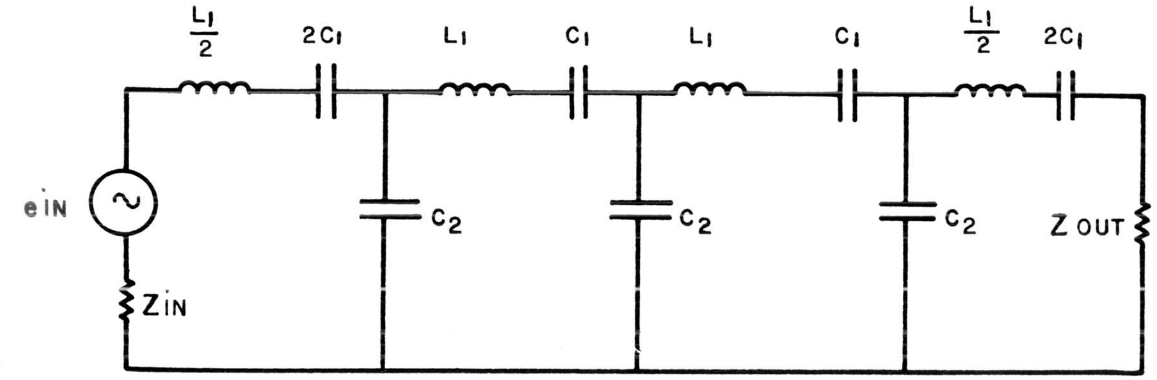


Figure 9-10.

Some important facts about mechanical filters.

a) the filters are symmetrical.

b) the input and output impedances are almost always resistive, because of the input and output resonant circuits. (1000 to 50,000 ohms).

c) the shape factor can be improved by adding more discs. The present limit for the number of discs is 9.

d) the bandpass can be changed by alternating the area of the coupling rods, or changing their number.

e) in the frequency range from 100 KC to 500 KC, bandwidths from about .5 KC to 35 KC are practical.

f) insertion losses vary from 2 to 16 db.

9-6 Crystal Filters

Crystal filters are used to good effect in SSB systems to obtain narrow bandwidths. Such filters may be designed for either symmetrical or asymmetrical response. As previously stated, crystal filters can be manufactured with bandwidths as narrow as .005% and as wide as 10% of the center frequency. An asymmetrical crystal filter is particularly effective following a balanced modulator in that additional suppression of the carrier may be obtained.

THE QUARTZ CRYSTAL AS A RESONANT ELEMENT

The equivalent circuit of a quartz crystal alone, without a holder, is shown in Figure 9-11. The component values represent typical orders of magnitude.

Note: the calculations that followed were performed on a slide rule, and decimals were rounded off. The results, therefore, must be regarded as approximations.

The circuit of Figure 9-11 is seen to be a simple series LCR circuit, the resonant frequency of which is:

$$F_R = \frac{1}{2\pi\sqrt{L_1 C_1}} = \frac{159 \times 10^{-3}}{\sqrt{12 \times 10^{-14}}} = \frac{159 \times 10^{3-}}{3.464 \times 10^{-7}}$$

$$= 459 \text{ KCS}$$

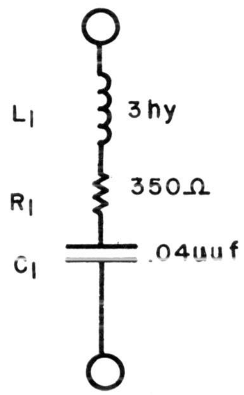


Figure 9-11.

The circuit has an extremely high "Q"; this may appear strange, in view of the relatively high value of series resistance; nevertheless, the ratio of inductive reactance to resistance is also high.

$$Q = \frac{X_{L1}}{R_1} = \frac{\sqrt{\frac{L_1}{C_1}}}{R_1} = \frac{\sqrt{75 \times 10^{12}}}{3.5 \times 10^2} = \frac{8.66 \times 10^6}{3.5 \times 10^2} = 24{,}900$$

A quartz crystal, then, has a point of series resonance and a very high "Q." A crystal, however, must be connected a practical circuit with some type of holder. The equivalent circuit of a crystal in its holder is shown in Figure 9-12 below. The order of magnitude of C-2 is typical.

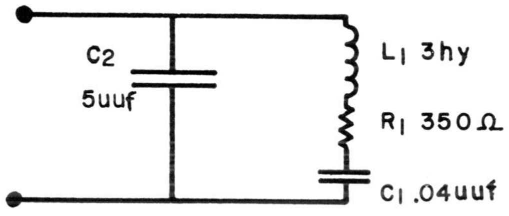


Figure 9-12.

At the series resonant frequency, the reactance of C-2, the holder capacity, is:

$$X_{C2} = \frac{1}{2\pi f_R\, C_2} = \frac{159 \times 10^{-3}}{4.59 \times 10^5 \times 5 \times 10^{-12}} = \frac{159 \times 10^{-3}}{22.95 \times 10^{-7}}$$

$$= 69.3 \text{ K}\Omega$$

At the series resonant frequency the very high reactance of the holder capacity has no effect on the impedance of the series resonant circuit. At a frequency slightly higher than the series resonant frequency, however, the reactance of C-2 will equal the algebraic sum of X_{L1} and X_{C1}. This algebraic sum will be an inductive reactance, and a parallel resonant (anti-resonant) circuit will result. The anti-resonant frequency may be derived as follows:

Derivation of the Anti-Resonant Crystal Frequency:

$$\frac{1}{\omega C_2} = \omega L_1 - \frac{1}{\omega C_1}$$

$$\frac{1}{\omega C_2} = \frac{\omega^2 L_1 C_1 - 1}{\omega C_1}$$

$$\frac{1}{C_2} = \frac{\omega^2 L_1 C_1 - 1}{C_1}$$

$$C_1 = \omega^2 L_1 C_1 C_2 - C_2$$

$$C_1 + C_2 = \omega^2 L_1 C_1 C_2$$

$$\omega^2 = \frac{C_1 + C_2}{L_1C_1C_2}$$

$$fa = \frac{1}{2\pi}\sqrt{\frac{C_1 + C_2}{L_1C_1C_2}}$$

In this particular case:

$$fa = 159 \times 10^{-3}\sqrt{\frac{5.04 \times 10^{-12}}{3 \times 4 \times 10^{-14} \times 5 \times 10^{-12}}} =$$

$$159 \times 10^{-3}\sqrt{\frac{5.04 \times 10^{-12}}{.6 \times 10^{-24}}} =$$

$$159 \times 10^{-3}\sqrt{8.4 \times 10^{12}} =$$

$$159 \times 10^{-3} \times 2.9 \times 10^{6} = 460 \times 10^{3} = 460 \text{ KCS}$$

When a crystal and its holder are connected in shunt or in series with an inductance, a second anti-resonant frequency, f_{A1}, will be found *below* the series resonant frequency. The equivalent circuit of such an arrangement is shown in Figure 9-13, below:

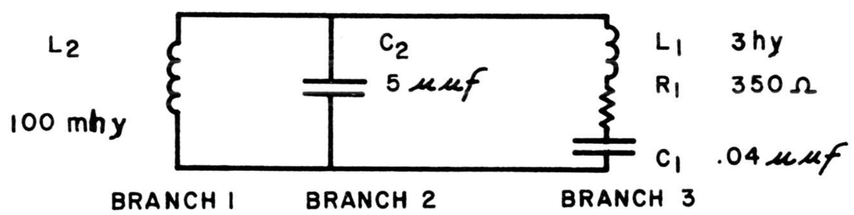


Figure 9-13.

At a frequency below the series resonant point, branches 2 and 3 will be capacitive and branch 1 will be inductive. The second anti-resonant frequency, f_{A1}, may be derived thus:

$$\omega L_2 = \frac{1}{\omega C_2} + \omega L_1 - \frac{1}{\omega C_1}$$

$$\omega L_2 = \frac{1}{\omega C_2} + \frac{\omega^2 L_1 C_1 - 1}{\omega C_1}$$

$$\omega L_2 = \frac{C_1 + W^2 L_1 C_1 C_2 - C_2}{\omega C_1 C_2}$$

$$L_2 C_1 C_2 = C_1 + \omega^2 L_1 C_1 C_2 - C_2$$

$$L^2 C_1 C_2 - C_1 + C_2 = \omega^2 L_1 C_1 C_2$$

$$f_{A1} = \frac{1}{2\pi}\sqrt{\frac{L_2 C_1 C_2 - C_1 + C_2}{L_1 C_1 C_2}}$$

In this particular case:

$$f_{A1} = 159 \times 10^{-3}\sqrt{\frac{496 \times 10^{-14}}{6 \times 10^{-25}}} =$$

$$159 \times 10^{-3}\sqrt{8.26 \times 10^{12}} =$$

$$159 \times 10^{-3} \times 2.88 \times 10^{6} = 458 \text{ KCS}$$

Figure 9-14 shows a plot of effective reactance against frequency for the crystal circuit just examined. Below f_{A1}, the net reactance is inductive. A very high resistive impedance is presented at f_{A1}. Between f_{A1} and f_R the net reactance is capacitive. At f_R the series resonant frequency, the circuit presents a very low resistive impedance. Above f_R, the circuit is again inductive.

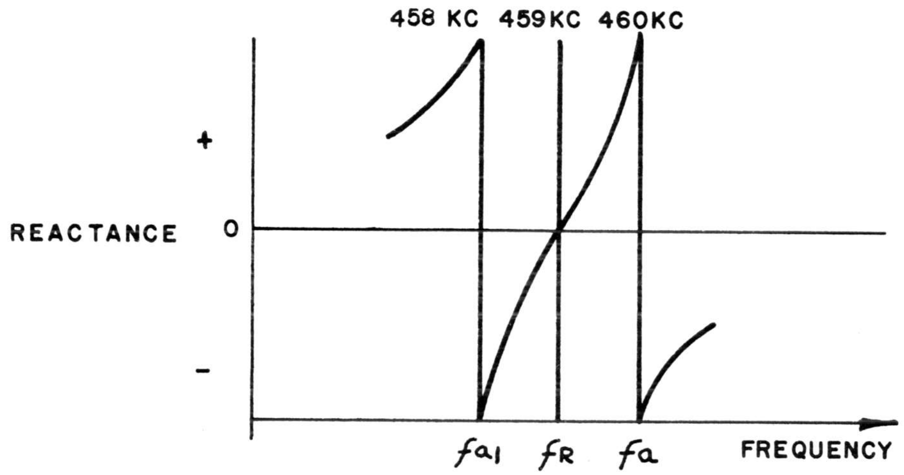


Figure 9-14.

A BASIC SINGLE SECTION CRYSTAL FILTER

Figure 9-15 shows a simple single section lattice network, connected as a crystal filter.

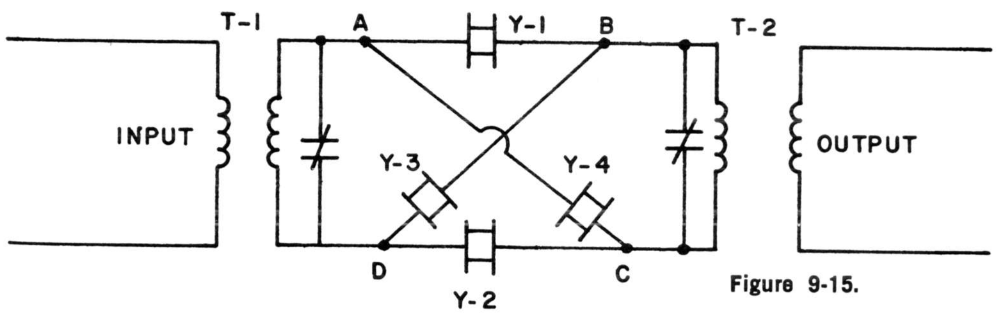


Figure 9-15.

a) T-1 and T-2 are tuned RF transformers, designed to resonate in the vicinity of the bandpass.
b) Y-1 and Y-2 are a carefully matched pair of crystals. The frequencies should be exactly matched, within a few cycles, if possible.
c) Y-3 and Y-4 also form an exactly matched pair of crystals.
d) the frequency difference between Y-1, Y-2 and Y-3, Y-4 determines, to a great degree, the bandpass of the filter.

It can be shown, in a mathematical transformation, that each crystal is effectively shunted with inductance in Figure 9-15; then, each crystal will respond at three frequencies: f_R, f_A and f_{A1}.

It can also be shown that the lattice network is exactly equivalent to a conventional four arm bridge connection; readers should note the letters at the junctions in Figure 9-16, and compare these with the letters in Figure 9-15.

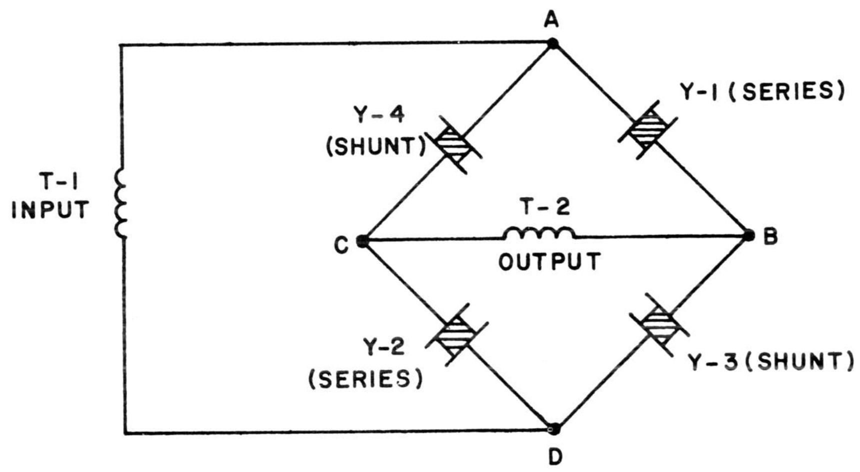


Figure 9-16.

Note also, in Figure 9-16, that T-2 is connected in the position normally regarded as "the output," "null indicator" or "galvanometer" position in a four arm bridge.

The shunt crystals, Y-3, Y-4, are higher in frequency than the series crystals, Y-1, Y-2, by an amount determined by the required bandpass. Note also that series and shunt crystals in the bridge equivalent ciruit are diagonally opposite.

The bandwidth between the series resonant frequency, f_R, and the upper anti-resonant frequency, f_A, depends to a great extent on the ratio of C-1 to C-2, that is, on the ratio of capacity in the crystal alone to the total shunt capacity, which is the holder capacity plus any additional capacity effectively across the holder. Stated as a formula:

$$\frac{2\triangle f}{f_R} = \frac{C_1}{C_2}$$

where: $\triangle f$ is the bandwidth between f_R and f_A. Then:

$$\triangle f = \frac{C_1 \, f_R}{2C_2}$$

Thus, for a given crystal, f_R is fixed, but f_A and f_{A1} may be shifted by careful adjustment of C-2 and L-2. See Figure 9-13.

In Figures 9-15 and 9-16, we will assume that f_R of the series crystal pair, Y-1, Y-2, is 459 KCS, and that the f_R of the shunt pair, Y-3, Y-4, is 460 KCS.

By careful circuit adjustment it will be possible to have:

a) f_{A1} of the shunt pair coincide with f_R of the series pair.

b) f_R of the shunt pair coincide with f_A of the series pair.

Figure 9-17 shows the reactance curves for both sets of crystals superimposed. The resulting reactance components in the bridge arms are also shown, for the frequency ranges involved. In addition, the ideal resultant response characteristic is sketched.

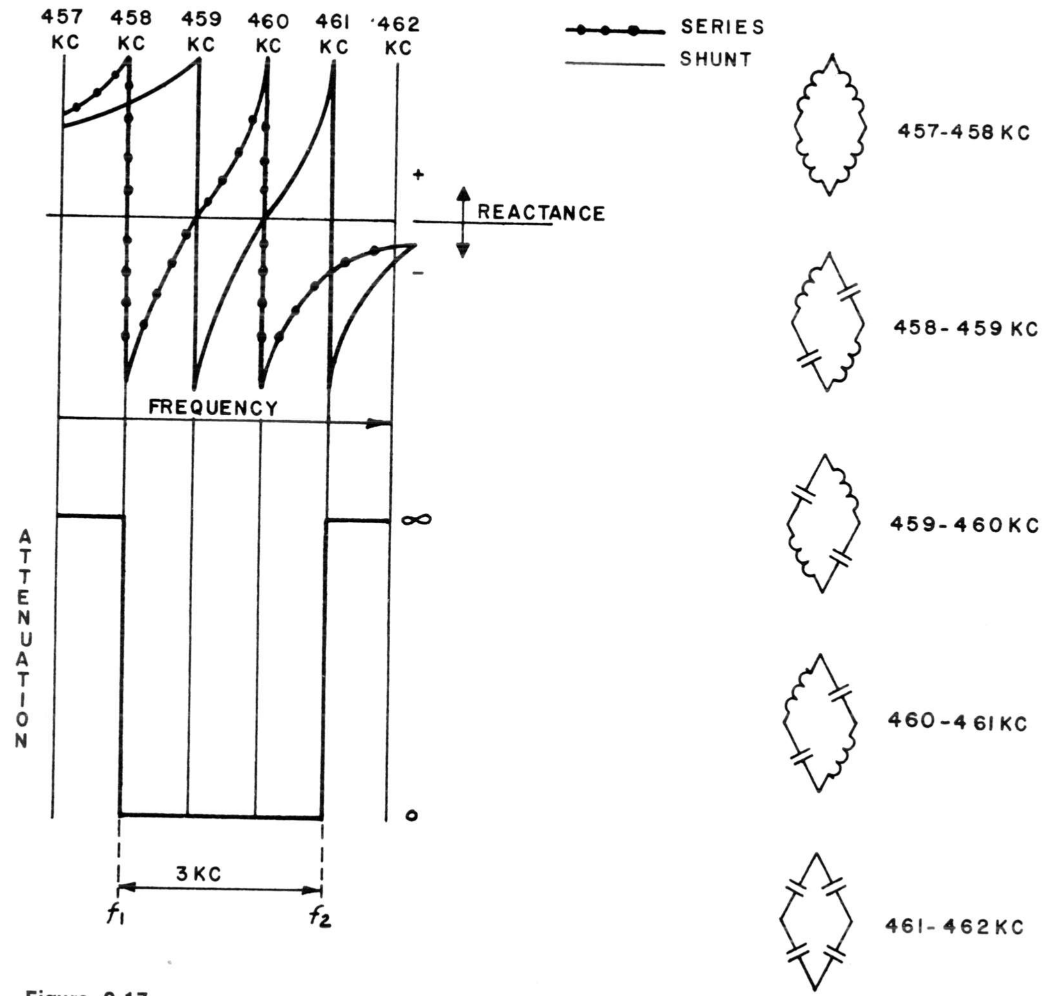


Figure 9-17.

In a bridge circuit containing reactances:

a) the bridge is balanced when the reactances are equal in amplitude and of the same reactance sign. Under these conditions, there is no output and attenuation is maximum.

b) the bridge is unbalanced to the maximum extent when the reactances are equal in amplitude and opposite in sign. Under these conditions, the output is maximum.

It should be noted that even though the response curve shown in Figure 9-17 is ideal, the reactance curves are not; the bridge balance and unbalance will not be perfect, and neither will the response curve. To improve the response curve, addtiional sections are cascaded, in much the same manner as with other types of filters.

Figure 9-18 shows a typical response curve for a multi-section filter superimposed on an ideal response curve.

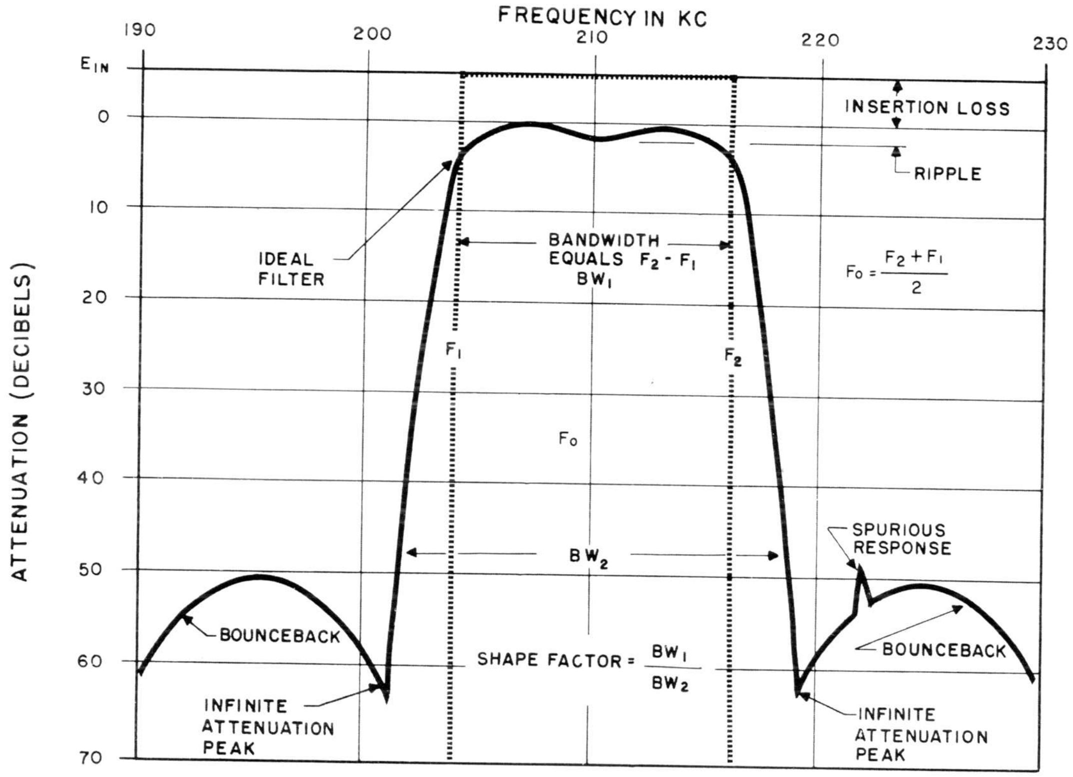


Figure 9-18.

The INSERTION LOSS refers to the difference between the output voltage and the input voltage, in DB. Notice that this may vary over the passband.

The RIPPLE refers to the variation of output voltage over the passband. It is usually expressed as a percentage, expressing the ratio of peak to peak variation of the output voltage between the 3 DB points, and the output voltage at the center frequency.

The BANDWIDTH or BANDPASS is that range of frequencies over which the output voltage does not fall below 3 DB of the voltage at the center frequency.

The SHAPE FACTOR describes the ratio of the bandwidth at the 3 DB points to the bandwidth at some high value of attenuation, usually 40, 50 or 60 DB. The shape factor of an ideal filter is unity. The shape factor of a practical filter is less than unity. In Figure 9-18, the bandwidth at the —3 DB points is about 12 KC; at the —50 DB points, the bandwidth is about 16 KC, the shape factor, then, is .75.

The FREQUENCIES OF INFINITE ATTENUATION occur at points of perfect bridge balance, resulting in maximum attenuation. See Figure 9-17: 457 and 462 KC. Also Figure 9-18, at about 202 KC and 219 KC.

BOUNCEBACK refers to the decrease of attenuation at frequencies outside the passband, away from the infinite attenuation frequencies. Note the bounceback in Figure 9-18; another infinite attenuation is being approached at 190 KCS; additional bounceback will occur below this second infinite attenuation frequency.

A SPURIOUS RESPONSE refers to any unwanted response in or near the passband. In Figure 9-18, note the spurious response at about 222 KC; this is down about 48 DB.

Crystal filters may also be configured for asymmetrical response. A typical asymmetrical response for a multi-section crystal filter is shown in Figure 9-19. This response is designed to pass an upper sideband; the carrier usually appears at the point between the infinite attenuation frequency and the bounceback peak.

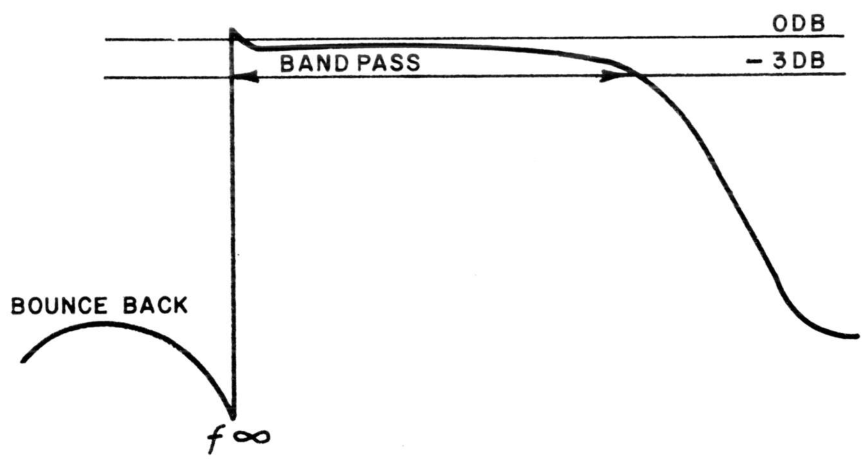


Figure 9-19.

CHAPTER 10

FREQUENCY SYNTHESIZERS

10-1 Dictionary Definition of Synthesis

Source: Webster's 7th Collegiate Dictionary:

SYNTHESIS: the combination or composition of parts or elements so as to form a whole.

SYNTHESIZE: to combine or produce by synthesis.

10-2 General Discussion

Frequency synthesis is the production or creation of a relatively large number of frequencies from a single stable frequency or group of frequencies. The synthesis may be accomplished in various ways. In the strictest sense, a simple harmonic generator may be considered as a frequency synthesizer. Consider Figure 10-1, below:

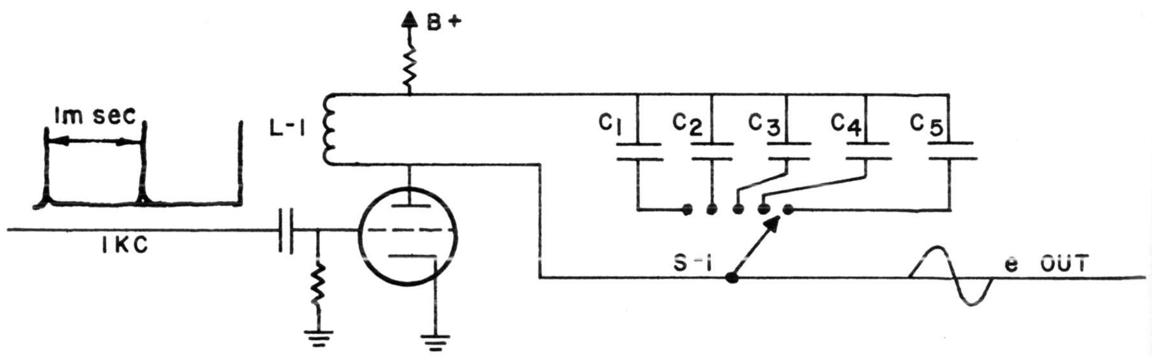


Figure 10-1.

Figure 10-1 shows a simple harmonic generator circuit. A 1 KC pulse, shaped for large harmonic content, is applied to the control grid. The plate circuit is tuned to resonance by L-1 and one of 5 capacitors arranged to resonate with L-1 at the 4th through the 8th harmonics of 1 KC. Thus, five discrete frequencies have been created from a single frequency; the accuracy of these output frequencies will depend on the accuracy of the applied 1 KC pulse.

10-3 A Common Misconception

Transmitters and Receivers are often referred to as "synthesized," when their stability is carefully controlled. This terminology may be in error. Frequency synthesis *is* employed in certain transmitters and receivers to *control* frequency, but the final frequencies under consideration are usually *not* synthesized. To be truly synthesized, the output frequency of a radio transmitter must be the product of frequency synthesis.

For example: in the TMC Model DDR-5 receivers, the local oscillator operates 1.75 mc above the incoming RF signal (2-32 mcs). The local oscillator frequency, in the range 3.75 - 33.75 mcs, is generated by a *conventional type of variable oscillator circuit,* but the *accuracy* and *stability* of the local oscillator is controlled by a frequency synthesizer, which generates 9,999 discrete frequencies in 100 cycle steps, in the range 3.2501 mcs to 4.250 mcs. To indicate that the local oscillator is synthesized, intimates that the local oscillator frequencies, in the range 3.75 - 33.75 mcs, are the product of frequency synthesis, which, in fact, they are not.

Certain early communication systems, (and some may still be operational), *did* synthesize the final output frequencies. This was extremely difficult to accomplish, and it has been found unnecessary in modern techniques. Consider the number of discrete frequencies that would be required to obtain a synthesized transmitter output, in 100 cycle steps, in the frequency range 2 - 32 mcs.

$$\frac{32\text{ mcs} - 2\text{ mcs}}{100} = \frac{30 \times 10^6}{1 \times 10^2} = 30 \times 10^4 = 300{,}000.$$

10-4 Frequency Synthesis and Frequency Accuracy

In Chapter 6 it was shown that, when all of the frequency determining elements in a communication system are synchronized to a single fixed frequency source, it is possible to achieve almost incredible accuracy and stability. Modern frequency synthesizers invariably use a single, stable, carefully controlled standard, upon which can be lavished most of the cost of frequency control. "Standard" frequencies of 1 mc and 100 KC are common. There are many solid state cesium beams and other types of frequency standards on the market today that are capable of delivering stabilites up to 1 part in 10^{11}. For practical communication purposes, 1 part in 10^8 or 10^9 per day stability is sufficient.

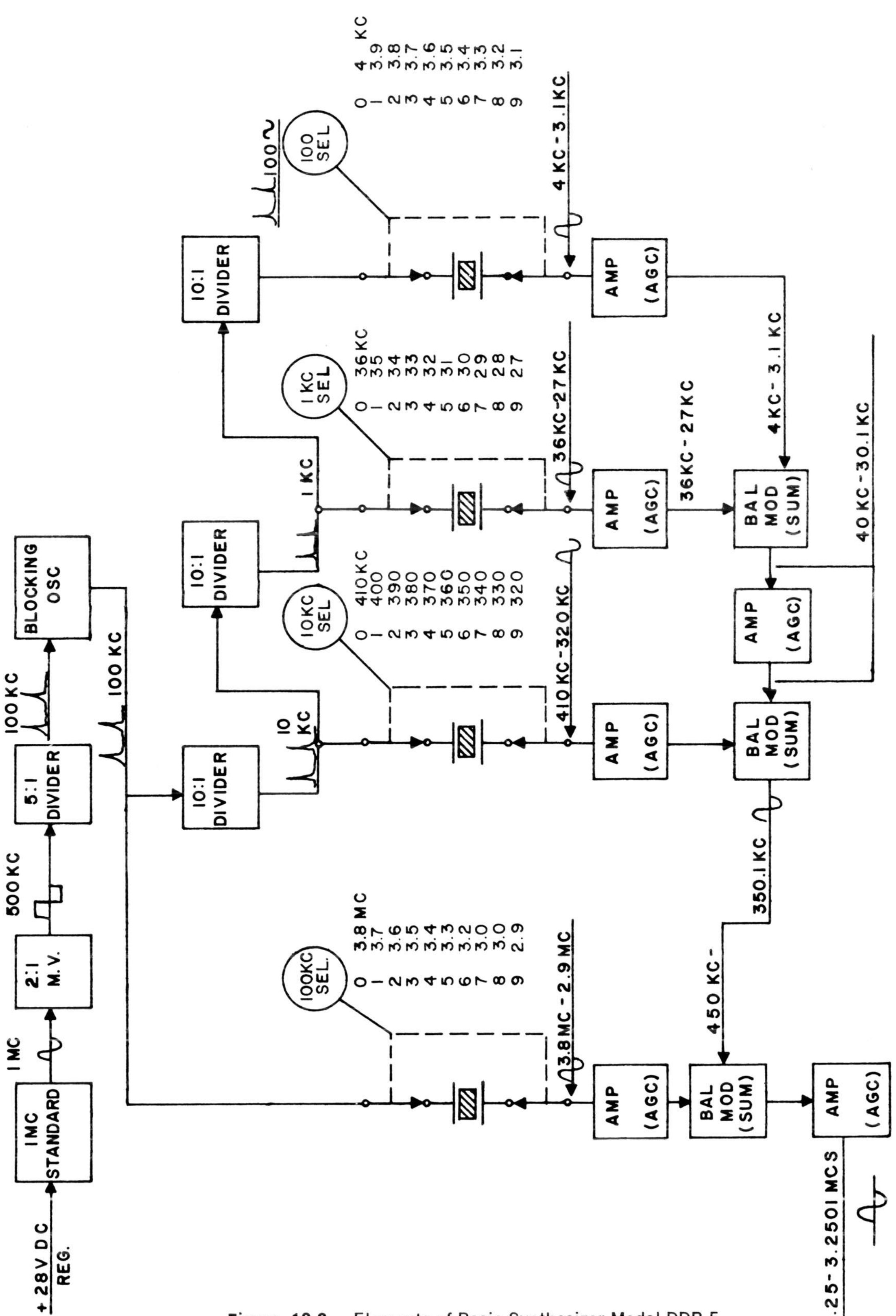


Figure 10-2. Elements of Basic Synthesizer Model DDR-5.

10-5 The Basic Synthesizer of the TMC Model DDR-5 Series Receivers

a) Figure 10-2 shows the elements of the basic synthesizer. The unit has several functions; the primary function is to produce any discrete frequency, in the range 4.25 - 3.2501 mcs, in 100 cycle steps, locked to a primary 1 megacycle standard. The particular frequency selected depends on the settings of four continuously variable front panel controls. The control settings, read on illuminated "NIXIE" indicators, are coded to read the 100 KC, 10 KC, 1 KC and 100 cycle digits of any associated RF frequency in the range 2.0 - 31.9999 mcs. Thus, each indicator illuminates a single digit, 0 through 9.

b) a transistorized, oven controlled 1 mc standard is employed. This plug in unit receives a regulated DC voltage, and delivers a 1 mc sinusoidal output, accurate to 1 part in 10^8 per day.

c) the 1 mc sinusoidal signal is counted down in a driven multivibrator circuit, which produces an output approaching a square wave, at 500 KC, locked to the 1 mc standard.

d) the 500 KC square wave is differentiated. The positive spikes at 500 KC trigger a 5:1 Phantastron divider circuit, which produces a new waveform at 100 KC, locked to the 1 mc standard.

e) the 100 KC waveform is used to trigger a blocking oscillator circuit; the blocking oscillator "regenerates" the 100 KC pulses, to improve the harmonic content.

f) the 100 KC pulses are applied, in sequence to three 10:1 Phantastron divided circuits, to produce, respectively, pulses at 10 KC, 1 KC and 100 cycles. The pulses have large harmonic content.

g) the 100 cycle pulses are applied to one of ten individual quartz crystals, in the range 4 - 3.1 KCS, by means of a 100 cycle selector switch. The switch has ten positions, each of which illuminates a "NIXIE" readout indicator. Thus, in position "0," the 4 KC crystal is inserted; in position "9," the 3.1 KC crystal is connected. The crystals act as narrow band selective filters, passing only the desired harmonics of 100 cycles. Thus, the 40th through the 31st harmonics of 100 cycles are passed in switch positions 0 through 9, respectively. Since the crystal is a high Q resonant circuit, the crystal output is sinusoidal. Note that the 100 cycle selector switch changes the frequency in 100 cycle steps.

h) the 4.0 KC - 3.1 KC output of the 100 cycle selector circuit is amplified, and sent to a balanced modulator. AGC circuits maintain a constant amplitude for each position of the selector switch.

i) the 1 KC pulse is applied, via a 1 KC selector switch with associated 1 KC indicator, to one of ten crystals in the range 36 - 27 KC. Thus, the 36th through the 27th harmonics of 1 KC are generated for switch positions 0 through 9, respectively. The sinusoidal signal is amplified, then applied to a balanced modulator which receives, in addition, the 4 - 3.1 KC output of the 100 cycle selector circuits. The circuits following the balanced modulators are configured to pass only the sum frequencies, which are in the range 40 KC - 30.1 KC. This sum signal is amplified and passed on to yet another balanced modulator. Note that the 1 KC selector switch changes the frequency in 1 KC steps.

j) the 10 KC pulse is applied, via a 10 KC selector switch with associated 10 KC readout indicator, to one of ten crystals in the range 410 - 320 KC. Thus, the 41st through the 32nd harmonics of 10 KC are generated for switch positions 0 through 9, respectively. The sinusoidal signal is amplified, then applied to a balanced modulator which receives, in addition, the signal from the 100 cycle and 1 kilocycle selector circuits, in the range 40 - 30.1 KC. The balanced modulator passes only the sum frequencies, in the range 450 - 350.1 KC. Note that the 10 KC selector switch changes the frequency in 10 KC steps.

k) the 100 KC pulses are applied to one of ten crystals in the range 3.8 - 2.9 mcs, via a 100 KC selector switch with associated readout indicator. Thus, the 38th through the 29th harmonics of 100 KC are generated for switch positions 0 through 9, respectively. The signals are amplified, then applied to the final balanced modulator, where they mix with the frequency generated by the 100 cycle, 1 KC, and 10 KC selector decks. The balanced modulator passes only the sum frequencies, in the range 4.25 - 3.2501 mcs. This signal is amplified; we will not carry it further at this point. Note that the 100 KC selector switch changes the frequency in 100 KC steps.

l) Figure 10-3 shows the front panel of the synthesizer unit. Note that there is a selector switch and readout indicator for the MC, 100 KC, 10 KC, 1 KC and .1 KC digits of any RF frequency in the range 2 - 31.9999 mcs. Figure 10-3 indicates that a receiver frequency of 13.4528 mcs has been selected. The MC switch and readout indicator will be ignored since they have no immediate connection with the basic synthesizer.

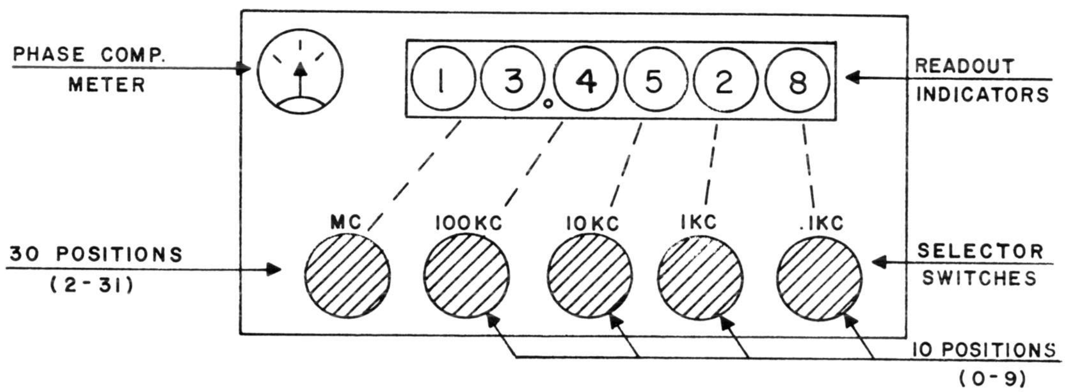


Figure 10-3.

m) reference to Figure 10-2 shows that the basic synthesizer is producing a frequency of 3.7972 mcs, as follows:

100 KC selector:	4	3400 KCS	
10 KC selector:	5	360 KCS	
1 KC selector:	2	34 KCS	
.1 KC selector:	8	3.2 KCS	
		3797.2 KCS	3.7972 mcs

It should be noted that the basic synthesizer will produce this frequency *regardless of the MC setting.*

Additional Examples:

100 KC	10 KC	1 KC	.1 KC	Output frequency
2	3	5	7	4.0154 mcs
0	0	0	0	4.25 mcs
9	9	9	9	3.2501 mcs
7	6	4	0	3.386 mcs
5	5	0	1	3.6999 mcs

10-6 The "High Frequency" Synthesizer of the Model DDR-5 Series Receivers

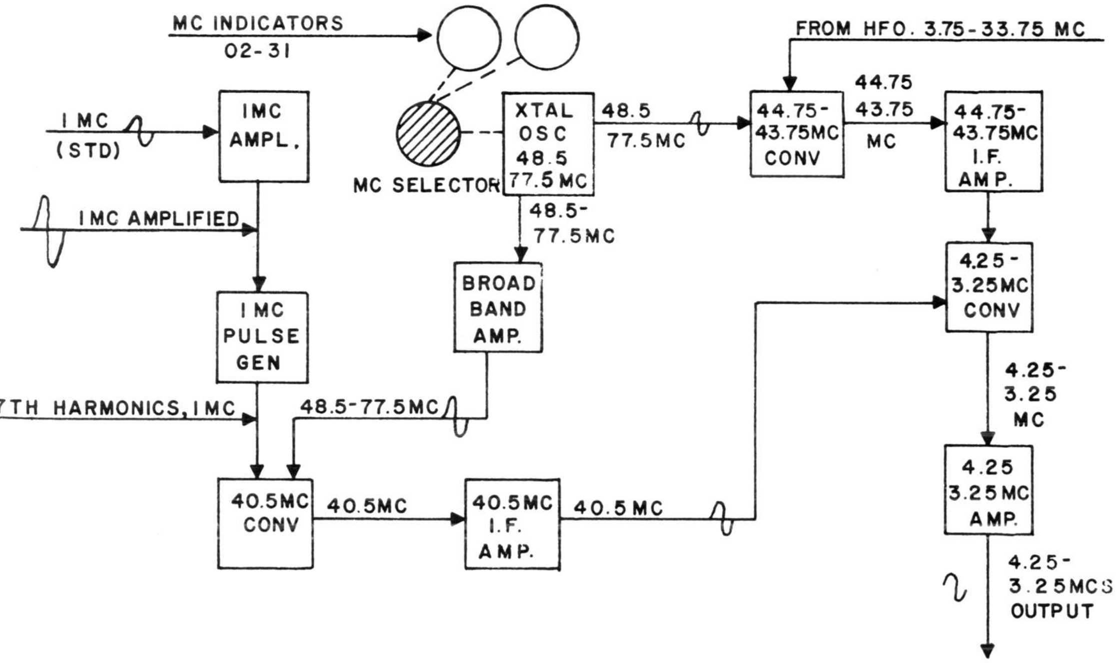


Figure 10-4.

a) Figure 10-4 shows the essential elements of the high frequency synthesizer. The 1 mc signal from the primary standard is greatly amplified in a 1 mc amplifier; the resultant 1 mc signal is used to drive a 1 mc pulse generator. This is a peaking circuit, which produces many harmonics of 1 mc, locked to the 1 mc standard. We are interested in the 8th through the 37th harmonics; these are applied to a 40.5 mc converter circuit.

b) the "MC" selector switch, with associated readout indicator, changes and indicates the frequency of an associated independent crystal oscillator circuit. Thirty crystals, in the range 48.5 - 77.5 mc, in 1 mc steps, are used. With the MC indicator at 02, the crystal oscillator frequency is 48.5 mcs; the frequency increases in steps of 1 mc with the MC selector, until, at position 31, the frequency is 77.5 mcs. The crystal oscillator output is applied to a broad band amplifier; the output of the amplifier is delivered to the 40.5 mc converter circuit.

c) it should be pointed out again that the MC readout indicator is coded to read the MC digit(s) of any RF frequency in the range 02 - 31 mcs. The associated crystal oscillator operates in the range 48.5 - 77.5 mc. To determine the crystal oscillator frequency for any readout position, subtract "2" from the readout, and add 48.5. This will give the crystal oscillator frequency.

Examples:

NIXIE INDICATOR		XTAL OSC.
02	02 — 02 = 0 + 48.5 =	48.5 mc
05	05 — 02 = 3 + 48.5	51.5 mc
12	12 — 02 = 10 + 48.5	58.5 mc
24	24 — 02 = 22 + 48.5	70.5 mc
31	31 — 02 = 29 + 48.5	77.5 mc

d) it should be noted that the crystal oscillator is not locked to the 1 mc standard, and the 30 discrete frequencies may be in error by about plus or minus 3 KCS. It will be shown subsequently that this error is cancelled out in the system.

e) the 40.5 mc converter is followed by a selective 40.5 mc IF strip, which has a very narrow bandpass. The output of this strip is always 40.5 mcs, plus or minus the error of the crystal oscillator circuit. When the crystal frequency is 48.5 mcs, the 8th harmonic of 1 mc causes the converter output to be 40.5 mcs. When the crystal frequency is 77.5 mcs, the 37th harmonic of 1 mc causes the converter output to be 40.5 mcs. The 40.5 mc output of the IF strip is applied to a final 4.25 - 3.2 mc converter.

f) the "local oscillator" or "high frequency oscillator" in the RF circuits of the DDR-5 receiver operates 1.75 mc above the received frequency, 2-32 mcs, to produce a first I.F. of 1.75 mcs.

Thus, the range of the L.O. (HFO) is 3.75 - 33.75 mcs. It is this oscillator that the control synthesizer will lock on frequency, in 100 cycle steps, to the accuracy of the 1 mc standard. A sample of the L.O. (HFO) frequency is applied to a 44.75 - 43.75 mc converter circuit. The second input to this converter arrives from the crystal oscillator, in the range 48.5 - 77.5 mcs.

g) the output of the 44.75 - 43.75 mc converter is always in the range 44.75 - 43.75 mcs. When the receiver is operated between 2 and 3 mcs, the MC selector is at "02" and the crystal oscillator operates at 48.5 mcs. If the TUNE control on the RF head is at 2 mcs, the L.O. frequency is 3.75 mcs.

48.5 mcs — 3.75 mcs equals 44.75 mcs.

If the TUNE control is at 3 mcs, the L.O. frequency is 4.75 mcs.

48.5 mcs — 4.75 mcs equals 43.75 mcs.

This situation will occur over every 1 mc increment because, as the L.O. changes frequency, the MC selector is also changed to the appropriate position.

Example #1

NIXIE READOUT:	08.0500
XTAL OSCILLATOR:	08 — 2 plus 48.5 = 54.5 mcs
L.O. Frequency:	08.0500 + 1.75 = 9.8 mcs.
Output of Converter:	54.5 — 9.8 = 44.7 mcs

Example #2:

NIXIE READOUT:	15.3450
XTAL OSCILLATOR:	15 — 2 plus 48.5 = 61.5 mcs
L.O. Frequency:	15.3450 + 1.75 = 17.0950 mcs
Output of Converter:	61.5 — 17.0950 = 44.4050 mcs

Example #3:

NIXIE READOUT:	31.0000
XTAL OSCILLATOR:	31 — 2 + 48.5 = 77.5 mcs
L.O. Frequency:	31.0000 + 1.75 = 32.75 mcs
Output of Converter:	77.5 — 32.75 = 44.75 mcs

h) the output of the 44.75 - 43.75 mc converter is applied to an IF strip. The output of the IF strip is applied to the final 4.25 - 32.5 mc converter, where it mixes with the constant 40.5 mc signal, developed earlier. The difference is a frequency in the range 4.25 - 3.25 mcs, based on the L.O. frequency and the 1 mc standard. As stated previously, the error of the Crystal oscillator does not appear in the output. The 4.25 - 3.25 mc signal is amplified, and will be taken no further at this point.

i) Summary:

(1) the basic synthesizer produces a signal in the range 4.25 - 3.25 mcs, depending on the settings of the 100 KC, 10 KC, 1 KC and .1 KC selectors. This signal is locked to the 1 mc standard, and can be considered "without error".

(2) the high frequency synthesizer produces a signal in the range 4.25 - 3.25 mcs, based on the 1 mc standard *and* the L.O. frequency. This signal then, contains the error of the local oscillator.

Examples of System Operation:

#1: NIXIE READOUT: 05.0482

Basic Synthesizer:			
100KC	0	3800	KC
10KC	4	370	
1KC	8	28	
.1KC	2	3.8	
		4.2018	mcs

High Frequency Synthesizer:

XTAL OSC: 05 — 2 + 48.5 = 51.5 mcs

L.O.: 05.0482 + 1.75 = 6.7982 mcs

44.75 - 43.75 mc

converter 51.5 — 6.7982 = 44.7018 mcs

4.25 - 3.25 mc

converter 44.7018 — 40.5 = 4.2018 mcs

#2: NIXIE READOUT: 17.0050

Basic Synthesizer:			
100 KC	0	3800	KC
10 KC	0	410	KC
1 KC	5	31	KC
.1 KC	0	4	KC
		4.245	mcs

High Frequency Synthesizer:

XTAL OSC: 17 — 2 + 48.5 = 63.5 mcs

L.O.: 17.0050 + 1.75 = 18.7550 mcs

44.75 - 43.75 mc

converter: 63.5 — 18.7550 = 44.7450 mcs

4.25 - 3.25 mc

converter: 44.7450 — 40.5 = 4.2450 mcs

#3: NIXIE READOUT: 09.9999

Basic Synthesizer:			
100 KC	9	2900	KC
10 KC	9	320	
1 KC	9	27	
.1 KC	9	3.1	
		3.2501	mcs

High Frequency Synthesizer:

XTAL OSC: 09 — 2 + 48.5 = 55.5 mcs

L.O.: 09.9999 + 1.75 = 11.7499 mcs

44.75 - 43.75 mc

converter: 55.5 — 11.7499 = 43.7501 mcs

4.25 - 3.25 mcs

converter: 43.7501 — 40.5 = 3.2501 mcs

j) the high frequency synthesizer has a drift cancelling feature which eliminates the error of the crystal oscillator circuit. Figure 10-5 shows the arrangement. Note that f2, the crystal oscillator frequency, does not appear in the output.

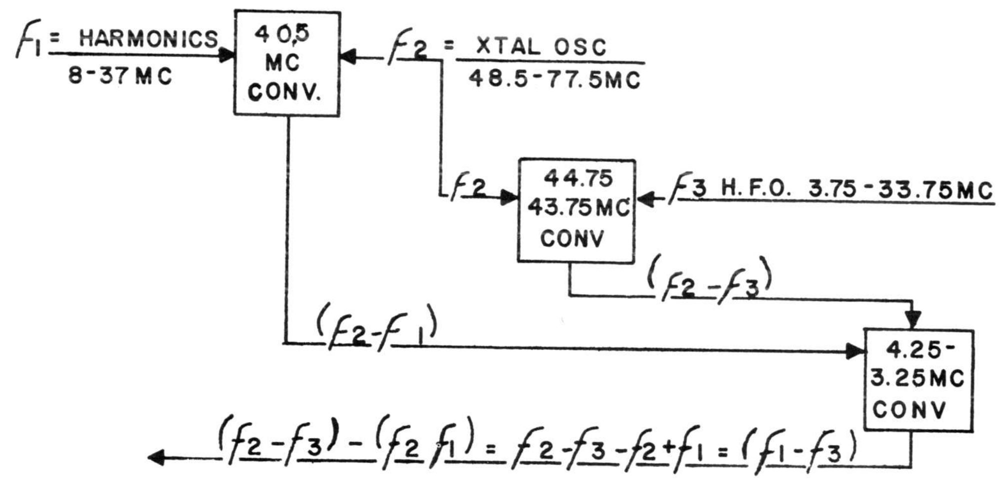


Figure 10-5.

10-7 Control of the L.O. (HFO) by Means of the Basic and High Frequency Synthesizers in the Model DDR-5 Series Receivers

a) The TMC Phase Detector Circuit: See Figure 10-6.
Figure 10-6A shows the basic TMC phase detector circuit. Figure 10-6B shows the output voltage from the phase detector for various phase angles; square waves are used to simplify the discussion, even though the waveforms ordinarily applied are sinusoidal.

e-1 is a standard frequency, derived exclusively from a primary standard. This signal is applied to the phase detector via T-1. It is shown in Figure 10-6B as being always at 0 degrees.

e-2 is a frequency ostensibly identical to e-1, derived from a source subject to error. This signal is applied to the phase detector via coupling capacitor C-1.

A Phase detector compares the signals from two different sources, at very close to the same frequency, and develops a DC voltage, which is used to correct the frequency of the source subject to error.

Assume that e-2 is not operating. When the polarity at the secondary of T-1 is positive, as shown by the polarity dot, diodes CR-1 and CR-2 conduct; this causes a current to flow in R-1, the Phase Detector Balance Adjust. Since the secondary of T-1 is centertapped to ground, R-1 can be adjusted so that, with the diodes conducting, the wiper of R-1 is at ground potential. Thus, under these conditions, any signal from e-2 is grounded.

When the polarity at the secondary of T-1 reverses, the diodes cannot conduct; the wiper of R-1 is at a high impedance with respect to ground, and any signal from e-2 will be developed here.

The correction voltage is taken from the wiper of R-1 and is applied to a low pass filter, R-2, C-2, which filters out the AC variations, and delivers an average DC voltage to the output terminal. This DC correction voltage is sent to a Reactance Modulator or Varicap control circuit, to bring the frequency of e-2 to the point where a zero correction voltage is developed.

Any attempt by e-2 to change frequency results in a correction voltage, which brings e-2 immediately back to the correct frequency. Zero correction voltage is developed when e-1 and e-2 are at the same frequency, plus or minus 90 degrees out of phase.

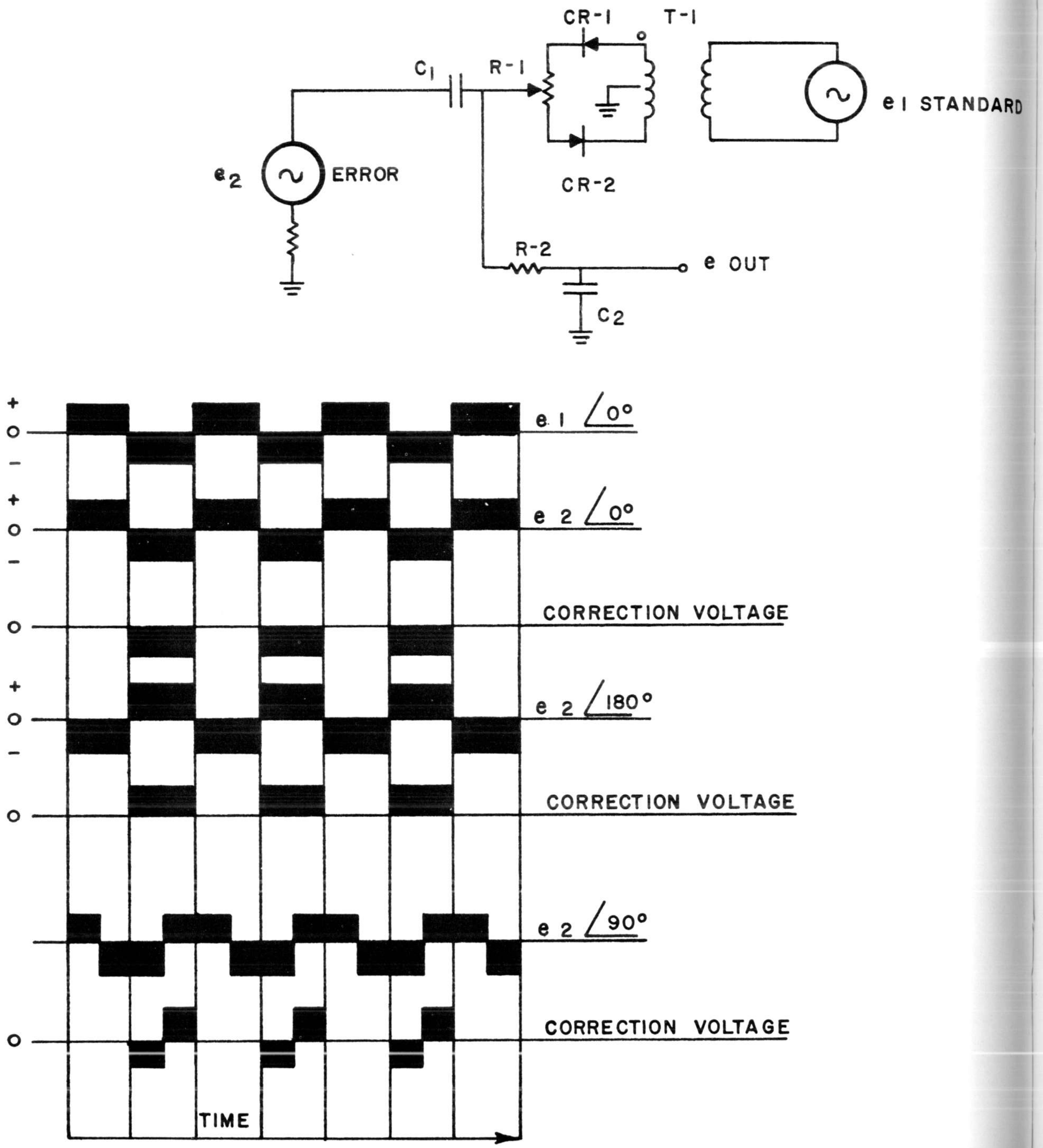


Figure 10-6. The TMC Phase Detector Circuit and Waveforms.

b) Interconnection of the Phase Detector, Basic Synthesizer, and High Frequency Synthesizer: See Figure 10-7.

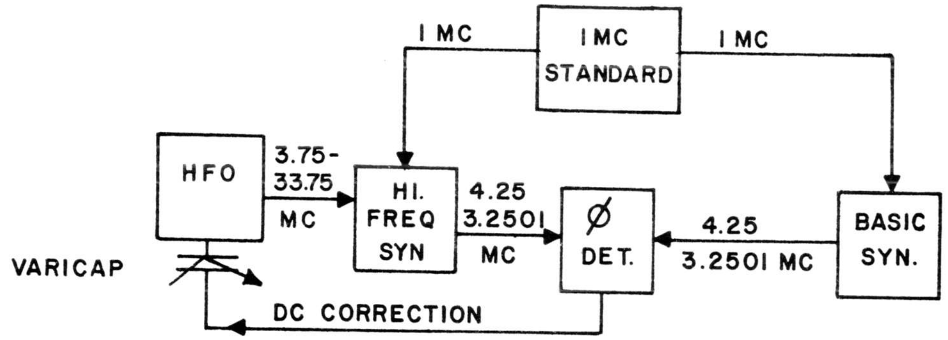


Figure 10-7.

(1) the basic synthesizer delivers to the phase detector a signal in the range 4.25 - 3.2501 mc, locked to the 1 mc standard; the exact frequency depends on the settings of the 100 KC, 10 KC, 1 KC and .1 KC selector switches.

(2) the high frequency synthesizer delivers to the phase detector a frequency very close to the frequency generated by the basic synthesizer; this frequency is based on the 1 mc standard and on the HFO frequency. Thus, it is in error by the amount of error of the HFO.

(3) the phase detector compares the two nominally identical frequencies; this circuit produces a DC correction voltage, which is applied to a Varicap in the HFO circuit. The Varicap is a semiconductor device which changes its capacity in accordance with the DC voltage impressed on it. The correction voltage causes the HFO to produce a frequency exactly equal to that which will cause the output of the high frequency synthesizer to equal the output of the basic synthesizer. When the two synthesizer frequencies are equal and 90 degrees out of phase, the HFO is locked to the 1 mc standard at the correct frequency. Any attempt by the HFO to change frequency causes a correction voltage to be developed, immediately returning the HFO to the correct frequency.

10-8 The High Frequency Synthesizer of the TMC Model SBG Sideband Generator

In this case, a high frequency oscillator produces injection frequencies in the range 8 - 15 mcs; it is required that these frequencies be exact, and locked to a 1 mc standard.

Figure 10-8 shows the essential elements of the system. The bandswitch causes the crystal oscillator to deliver 8 discrete frequencies, in the range 8 - 15 mcs, to a mixer circuit.

The second input to this mixer circuit arrives from a harmonic generator, also operated by the bandswitch. As the bandswitch is rotated, the harmonic generator produces the 6th through 13th harmonics of 1 mc, derived from the 1 mc standard.

When the crystal oscillator is at 8 mcs, the harmonic selector is producing 6 mcs; when the crystal oscillator is at 9 mcs, the harmonic selector is producing 7 mcs; thus, the difference frequency is always 2 mcs. This 2 mc signal, which contains the error of the crystal oscillator, is applied to the wiper of the phase detector balance adjust potentiometer.

The second input to the phase detector is derived from a second harmonic generator, which is locked to the 1 mc standard. Thus, this signal is the "standard", and is considered without error.

The secondary of the phase detector transformer is not grounded; in this case it is returned to a stable voltage divider which puts plus 4 volts at the center tap. This is the "zero correction" voltage. The wiper of the phase detector potentiometer is set at 4 volts, with the 2 mc signal from the mixer disconnected.

The phase detector develops a correction voltage, based on the error of the HFO. This correction voltage is filtered, then applied to a Varicap in the HFO circuit. Thus, the HFO is forced to come to the correct frequency.

The output of the HFO is a stable frequency in the range 8 - 15 mcs, locked to the 1 mc standard. This signal supplies the injection frequencies for the final balanced modulator in the system.

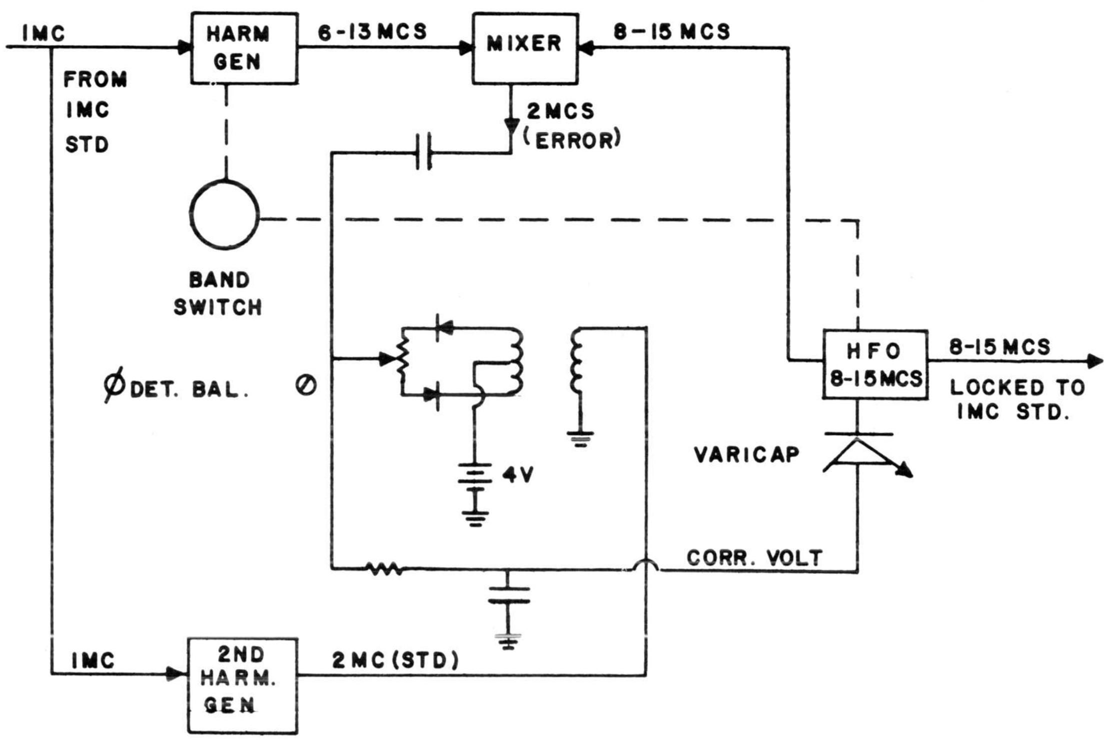


Figure 10-8. High Frequency Synthesizer of the Model SBG-1 Sideband Generator System.

CHAPTER 11

SINGLE SIDEBAND RECEIVERS AND CONVERTERS; AUTOMATIC FREQUENCY CONTROL.

11-1 Introductory Note

In this chapter, SSB receiving systems will be discussed. Various TMC equipments will be used as vehicles to explain the important aspects of SSB reception and automatic frequency control.

11-2 General Considerations

A SSB receiver is not significantly different from a good conventional superheterodyne AM receiver, except for the detector circuit and for the fact that, in the SSB receiver, the "carrier" must be inserted in this detector circuit at a frequency which corresponds almost exactly with the relative position of the carrier in the original spectrum. The AGC circuits of the SSB receiver may also be different, because, in the conventional AM receiver, AGC is usually based on the received carrier. This carrier may be altogether absent in the SSB receiver, up to the detector circuit. The relative complexity of the circuits in a SSB receiver will depend primarily on the service for which the receiver is intended.

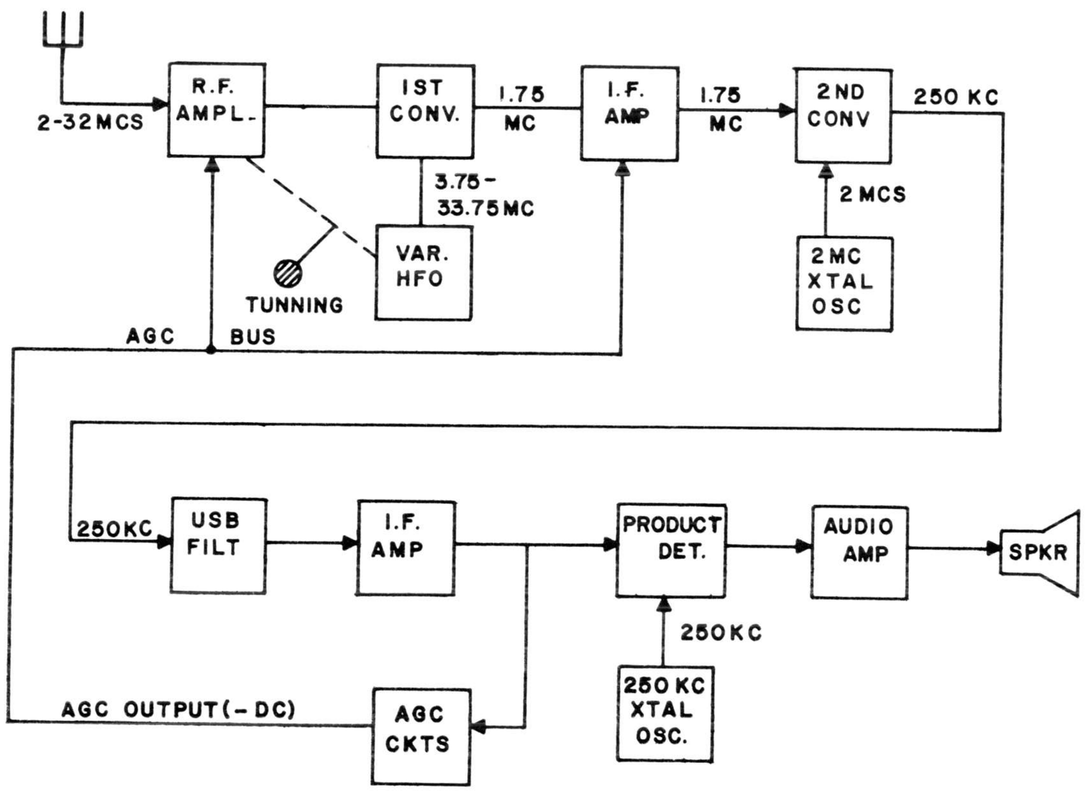


Figure 11-1.

11-3 An Elementary SSB Receiver Without Automatic Frequency Control

a) in figure 11-1, the system is a double conversion superheterodyne receiver, similar to a conventional AM receiver.

b) the number of RF stages is determined by the sensitivity and selectivity requirements. The automatic gain control voltage may be applied to one or more of these stages. The output of the RF amplifier chain is applied to a first converter circuit. Here, the selected RF frequency is mixed with the output of a stable high frequency oscillator, operated 1.75 mcs above the incoming signal.

c) in some receivers, the first mixer injection is supplied by a crystal oscillator and the second converter oscillator is made variable. This aids the stability of the system. In the circuit of Figure 11-1, the first injection is variable, producing a constant first intermediate frequency of 1.75 mcs. It is assumed that this oscillator has been designed for high stability and accuracy.

d) the oscillator injection, signal injection and converter bias voltages are such that intermodulation products from the mixer are negligible. Sideband inversion takes place but this is of no significance, since another inversion takes place in the second converter.

e) the first IF frequency, at 1.75 mcs, with full response, is applied to a second converter after amplification in one or more stages of intermediate amplifiers. The second converter receives a constant, stable 2 mc signal from a crystal oscillator. The output circuits of the second converter are tuned to the difference frequencies. As with the first converter, every effort is made to reduce intermodulation distortion.

f) since this receiver has been designed to accept only the upper sideband frequencies, the second converter is followed by an USB filter, which may be of the LC, mechanical, or crystal type. This filter is responsive to frequencies greater than 250 KC; the exact bandpass will depend on the service for which the receiver is designed. We will assume that the filter passes frequencies from 250,350 cycles to 253,500 cycles. Note that this response does not include the "carrier", at 250 KC.

g) it should be noted that, up to the input of the sideband filter, the only limit to the passband is that imposed by the selectivity of the RF and IF amplifier circuits.

h) the sideband filter is followed by one or more stages of IF amplification, depending on the required gain. The output of the IF amplifier is applied to a product detector. The second input to the product detector is an injection frequency of 250 KC, derived from a stable, accurate crystal oscillator.

i) a product detector is simply a mixer or converter circuit, similar in many respects to the first and second converters, and to the translation converters of any SSB transmitter system. The converter tube is usually a triode or pentode. The injected "carrier" frequency is usually applied to the control grid at a high amplitude with respect to the amplitude of the intelligence, so that conditions of saturation and cutoff are reached on opposite excursions of the injection signal. Because of the mixing action of the two applied signals, many products appear in the plate circuit. The plate circuit is usually tuned by a circuit sensitive only to the difference frequencies; the other products are attenuated in this process. With careful design, the product detector can be made to have insignificant intermodulation distortion; this distortion is more pronounced in conventional AM detectors.

j) the difference frequencies from the product detector are the original audio intelligence frequencies at the transmitter. These are amplified in a conventional audio amplifier.

k) the output of the final IF amplifier may also be applied to an AGC detector circuit, which produces a negative DC voltage which varies in amplitude with the amplitude of the signal at the antenna. This AGC voltage is based on intelligence, since no carrier is present here. The AGC voltage may control the gain of one or more RF and IF amplifier circuits.

11-4 Frequency Translation in the Elementary SSB Receiver

a) let us assume:

(1) that the transmitter is emanating an USB signal. The frequency of the suppressed carrier is 5.0 mcs. The intelligence consists

of two discrete continuously keyed tones at 1 KC and 2 KC. These are transmitted, then, at 5.001 mcs and 5.002 mcs. The transmitter frequencies are correct, and there is no transmitter drift.

(2) that the injection oscillators in the receiver are accurate to plus or minus .005%.

(a) the correct first injection is 5.0 mcs, plus 1.75 mcs, or 6.75 mcs. This frequency may drift, then, plus or minus 337 cycles.

(b) the correct second injection is 2.0 mcs. This frequency may drift, then, plus and minus 100 cycles.

(c) the correct product detector injection is 250 KCS. This frequency may drift,then, plus and minus 12.5 cycles.

b) it is possible, that, at some particular instant,

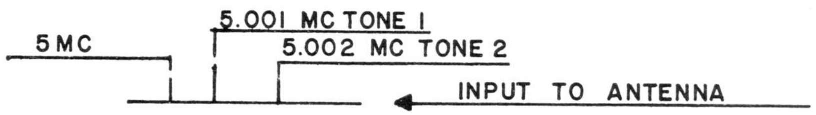
(1) the first injection frequency is 6.7501 mcs.

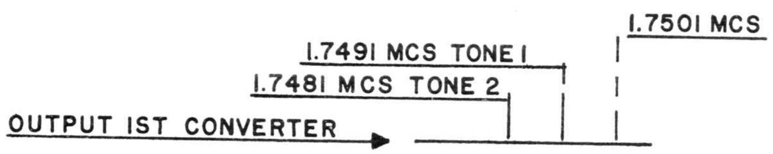
(2) the second injection frequency is 2,000,050 cycles.

(3) the product detector injection frequency is 250,008 cycles.

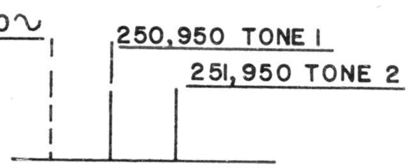
c) the resultant frequency translations at this particular instant are shown below:

(1) first translation:

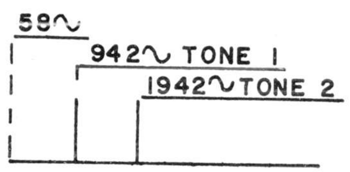




(2) output, second converter:



(3) output, product detector:



d) the two audio output tones are in error: they are exactly 58 cycles below the original audio tones of 1 KC and 2 KC. This situation may or may not be tolerated, as the following discussion will show.

(1) for audio voice work this error is not too large. If a voice spectrum were shifted by 58 cycles, the sound would be unnatural, but the intelligibility would not suffer exceedingly.

(2) if it were required that the two original audio tones be received exactly as transmitted, this situation could not be tolerated.

e) consider the following situation:

by means of multiplexing devices, an upper sideband with a response from 350 cycles to 3500 cycles, is to contain 16 discrete channels of information. Each channel will consist of 2 audio tones, the lower frequency tone corresponding to a teletype SPACE and the upper frequency tone corresponding to a teletype MARK. There is to be a 40 cycle "guard band" between each channel, and at both extremes of the sideband. The total frequency shift of each channel is to be 154 cycles. Figure 11-2 shows the layout of the first three channels.

A glance at Figure 11-2 should indicate the chaos that would result if we attempted to receive such a signal on our elementary single sideband receiver of Figure 11-1. While this receiver would be quite adequate for ordinary voice communication, it would be absolutely useless in a situation such as illustrated in Figure 11-2. To receive a signal of the type depicted in Figure 11-2, our receiver might employ one of three devices:

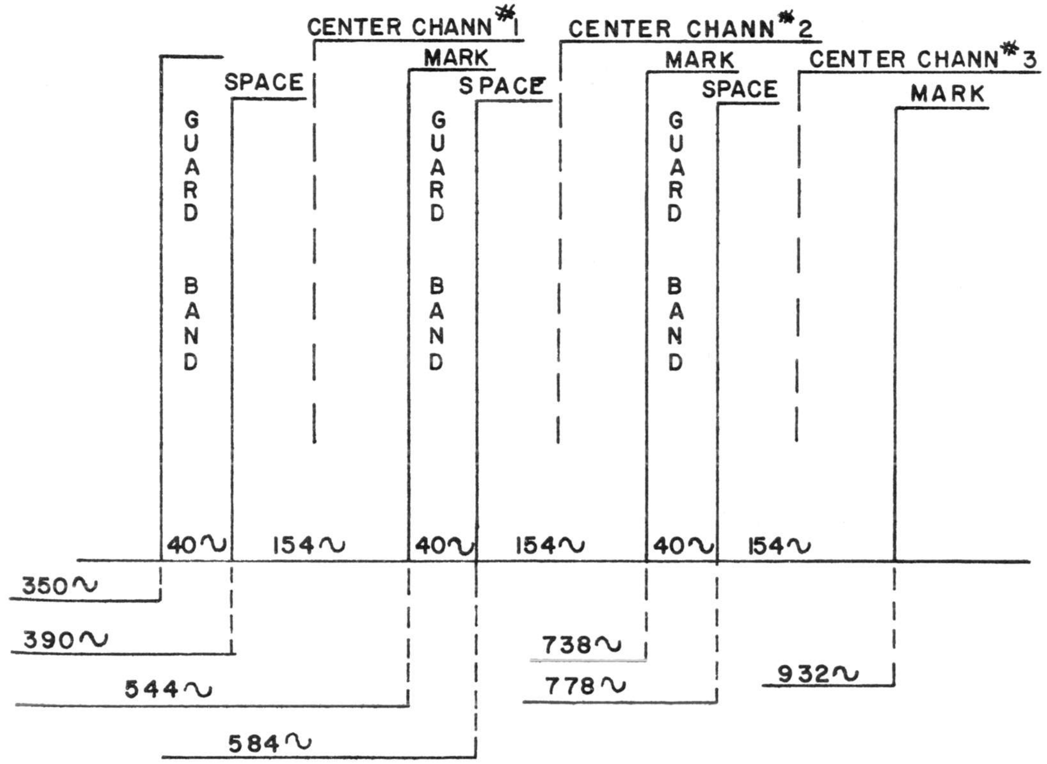


Figure 11-2.

(1) Precision Frequency Control:
this would require that all injection frequencies be derived from sources with incredible degrees of stability, or from a single source or control synthesizer with the same degree of stability. Even so, the receiver would still be at a disadvantage *unless the transmitter were similarly controlled.*

(2) Automatic Frequency Control:
this would require that the transmitter emanate along with the upper sideband, a small "pilot" carrier. This carrier would be restored in the receiver, and used to control the second converter injection and product detector injection frequencies. The first converter injection would be derived from a stable oscillator. An elementary block diagram of such a system is shown in Figure 11-3.

In Figure 11-3 an upper sideband signal with small pilot carrier is shown emanating from a SSB transmitter. This signal is picked up by the receiver, where it is translated, in the first converter, to an IF frequency of 1.75 mcs. As this point, the signal frequency contains the error of the transmitted frequency and the HFO. The signal is again translated in the second converter. The USB filter and IF amplifier pass the intelligence to the product detector; the pilot carrier, in the region of 250 KC, is passed to the automatic frequency control unit, which

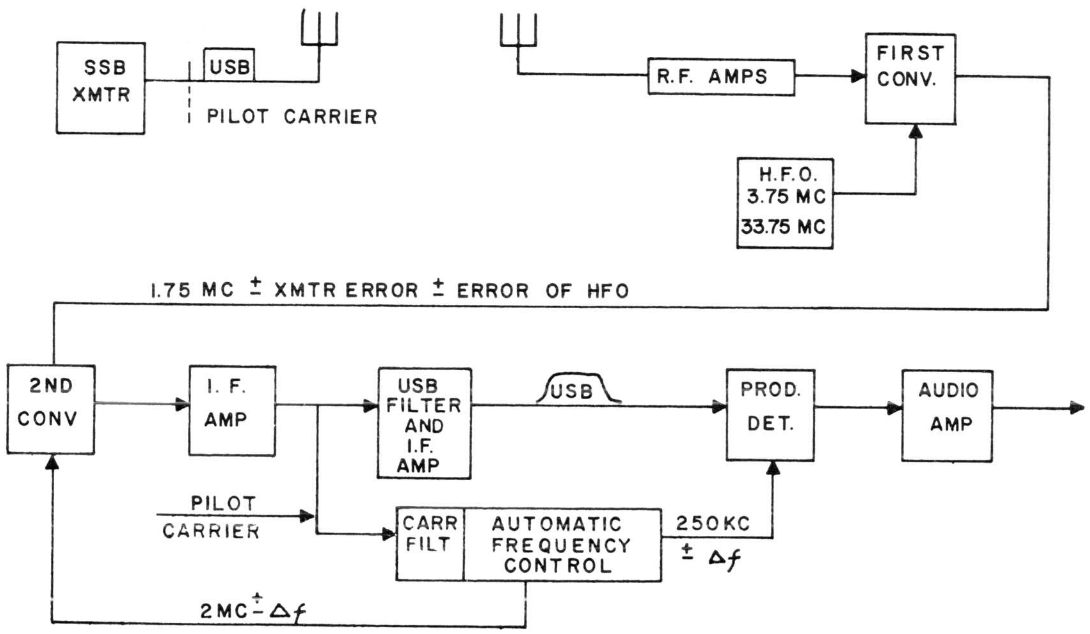


Figure 11-3.

contains an input selective filter. This selective filter passes a small range of frequencies in the immediate vicinity of 250 KC, so that only the carrier affects the AFC circuits. Thus, the AFC unit receives a "reference" frequency which follows the transmitted pilot carrier frequency. The small pilot carrier is restored in the AFC unit, to an amplitude suitable for use as the product detector injection frequency. The product detector injection, then, is identical with the pilot carrier. Thus, as the transmitted intelligence drifts upward or downward in frequency the product detector injection frequency increases and decreases by the same amount, and the audio frequencies out of the product detector maintain their proper place in the spectrum. Other circuits in the AFC unit cause the second converter injection frequency to change in such a manner as to keep the second IF "carrier" frequency close to 250 KC. The only error, then, is the HFO error which can be reduced by careful design. Still, the error might be great enough to prevent reception of a signal of the type illustrated in Figure 11-2.

(3) Precision Frequency Control and Automatic Frequency Control:
this system is similar to that illustrated in Figure 11-3, except that the HFO is precision frequency controlled. The only error, then, is the error of the transmitted frequency, which is compensated for by the automatic frequency control unit. Such a system can be expected to hold down the total residual error to within a cycle or two.

f) it should be noted that, when both the transmitter and the receiver are precision controlled, no automatic frequency control is required.

11-5 Effects of Intermodulation Distortion in SSB Receivers

The importance of reducing intermodulation distortion products in SSB transmitters was discussed in detail in earlier chapters of this work. This type of distortion is equally important in SSB receivers. In a SSB receiver employed for single channel voice transmission, intermodulation distortion is not liable to produce grave results; however, return for a moment to Figure 11-2. Here it is imperative that the tone frequencies *and only the tone frequencies* appear in the audio output. Severe intermodulation distortion with this type of signal would cause havoc.

Consider, now, an USB signal occupying a spectrum from 350 cycles to 7500 cycles. This sideband is subdivided into two "slots", each 3575 cycles wide, including guard bands. The lower slot, from 350 cycles to 3925 cycles, contains 16 FSK tone channels and the upper slot, from 3925 to 7500 cycles, contains a voice channel. Severe intermodulation distortion with such an arrangement would cause spurious frequencies from both channels to spill over into the adjacent channels.

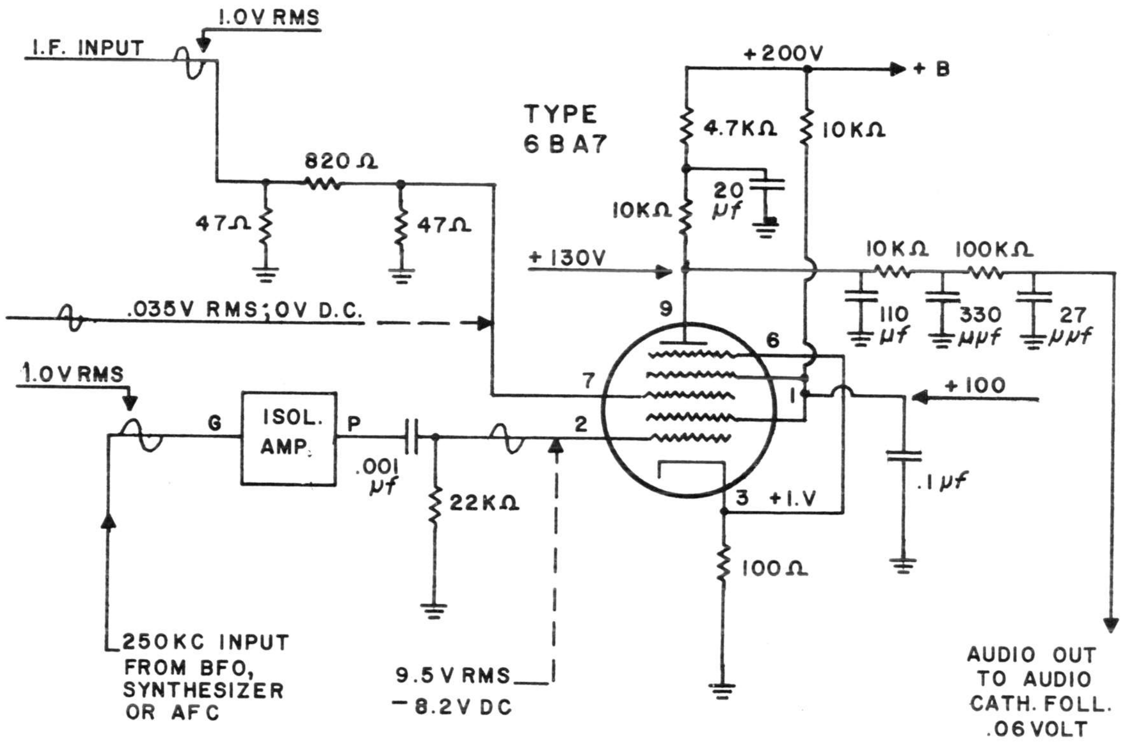


Figure 11-4.

11-6 The Product Detector Circuits of the Model HFA Detector and Audio Unit

a) in figure 11-4, the product detector "carrier" injection is applied, via an isolation amplifier, to the control grid, pin 2. The input to the isolation amplifier arrives, at a level of 1 volt RMS, from:

(1) a beat frequency oscillator in the CW mode. The BFO frequency may be shifted plus or minus 5 KCS about 250 KC, to provide the desired audio tone at the output.

(2) an automatic frequency control unit in AFC ON operation. The nominal 250 KC signal from the AFC unit follows the drift of the received pilot carrier.

(3) a control synthesizer unit in AFC OFF operation. This injection frequency is 250 KCS, locked to a primary 1 mc standard, with an accuracy of 1 part per 100,000,000 per day.

b) the signal at the control grid is 9.5 volts RMS. This represents a peak to peak excusion of 26.6 volts. This signal is "clamped" negatively due to control grid current on positive peaks. The DC voltage from pin 2 to ground is —8.2 volts. Thus, the signal at the control grid swings the tube from cutoff to saturation.

c) the IF signal is applied to pin 7 via a PI attenuator network. With a single frequency in the region of 250 KC at 1.0 volt RMS present at the input to the PI network, pin 7 receives a signal at 35 millivolts RMS. Thus, the "signal" injection is much, much smaller than the "carrier" injection.

d) the two input frequencies are mixed in the tube, but many products appear at the plate. These consist of original frequencies, sums, differences, intermodulation products and harmonics. The plate impedance consists of a low pass RC filter network, which passes only the difference audio frequencies. For the conditions shown in Figure 11-4, a voltage of about 3.0 volts RMS will be measured at the plate, pin 9. This measurement is absolutely insignificant, however, since it does not indicate the value of the one desired tone which appears at the audio output terminals of the low pass filter.

e) the output of the product detector circuit is applied to a cathode follower, the first stage in the audio amplifier chain.

11-7 A Discussion of the TMC Model HFI Intermediate Amplifier Unit

a) the Model HFI-1 is an intermediate frequency amplifier, part of the TMC Model DDR-5 (AN/FRR-60(v)) series receivers.

b) the unit receives a 1.75 mc IF signal, with full response, from an associated RF unit and delivers one or two IF outputs, at 250 KC, to an associated Detector and Audio Amplifier unit. The output response, in the region of 250 KC, is determined by IF Bandwidth selector switches. The local oscillator (HFO) of the associated RF head may be locked, if desired, to a primary 1 mc standard with an accuracy of 1 part per 100,000,000 per day. Thus, if the transmitter being received is precision frequency controlled, the 1.75 mc input is also precision frequency controlled. If the transmitter being received is subject to drift, an automatic frequency control feature may be switched in.

The signal input at 1.75 mc has undergone one conversion in the associated RF unit; hence, the sidebands are inverted at this point.

c) refer to the simplified block diagram, Figure 11-5:

(1) the 1.75 mc first IF frequency, with full response, if applied to the second converter at J-6205. The second converter, V-6201A, also receives a 2 mc injection frequency from an isolation amplifier, V-6201B.

(2) when the AFC switch is in the OFF position, the 2 mc signal arrives, at J-6206, from an associated control synthesizer. Under these conditions, the 2 mc signal is precision frequency controlled.

(3) when the AFC switch is in the ON position, a nominal 2 mc signal arrives, at J-6202, from the associated automatic frequency control unit. Under these conditions, the 2 mc signal is varied as the received "pilot" carrier varies, to maintain the second IF frequency in the immediate region of 250 KC.

(4) since V-6201A is a "difference" converter, the sideband output, at 250 KC, is "right side up." This is applied to an IF amplifier, V-6202.

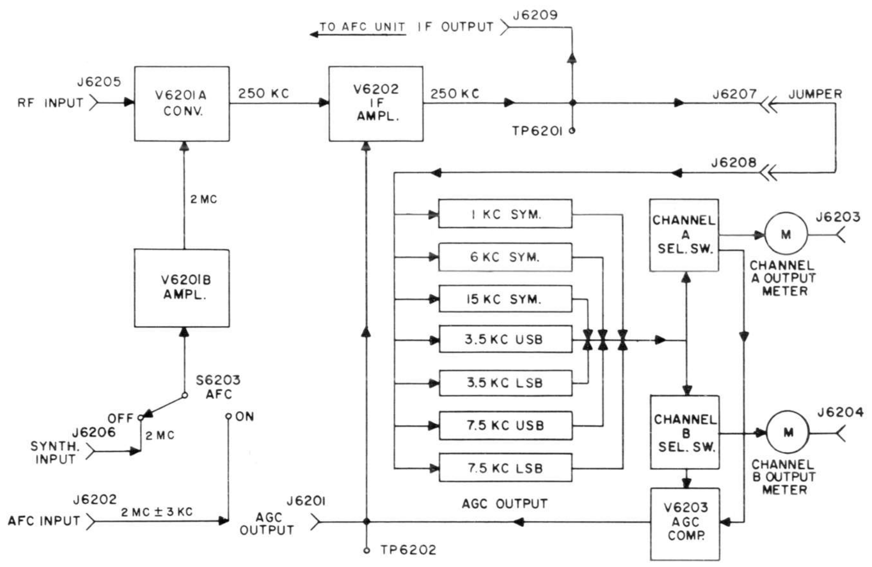


Figure 11-5. Simplified Block Diagram, Model HFI.

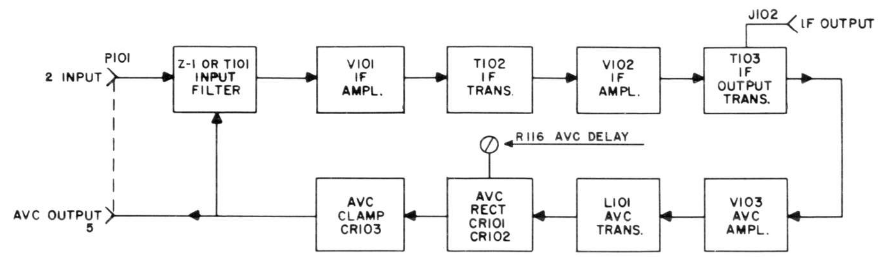


Figure 11-6: Simplified Block Diagram, Individual IF Strip.

(5) the amplified 250 KC signal, still with full response, is applied:

(a) to IF output jack J-6209. This jack connects to the automatic frequency control unit. It is here that the pilot carrier, if present, is delivered to the AFC unit.

(b) to a test point, TP-6201.

(c) to a pair of coaxial jacks, J-6207 and J-6208. An associated Notch Filter, the Model HNF, may be connected here. If the notch filter is not part of the installation, a short jumper connects J-6207 and J-6208.

(d) to the signal input terminals of seven individual IF strips. Figure 11-6 shows the simplified block diagram of one of these strips.

(e) each IF strip receives the 250 KC IF signal with full response. This enters the strip at P-101, terminal 2. This is applied to a crystal sideband filter or, in the case of the 15 KC symmetrical strip, to a double tuned IF transformer. This input filter, (Z-1 or T-101), passes only that portion of the response desired. The strips are almost identical, except for the input filter and the values of swamping resistors.

(f) as indicated by Figure 11-6, each strip consists of an IF amplifier chain, tuned to pass the response selected by the input filter. The IF output is taken off at J-102 on each strip. This is a miniature coaxial connection, which is delivered to the Channel A and Channel B IF Bandwidth Selector switches. A portion of the IF signal is amplified in V-103; this signal is used to develop an AGC voltage, based on the intelligence passed by the strip. R-116, the AVC DELAY adjustment, is set to provide an IF signal of 1 volt RMS out at J-102 for a wide range of input amplitudes at P-101, terminal 2. The AGC voltage is delivered to V-101 on the strip, and to an AGC comparator tube, V-6203, via the IF Bandwidth Selector switches.

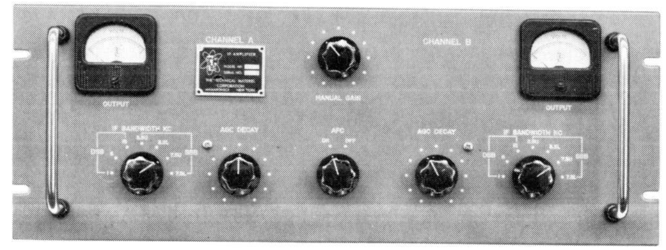


Figure 11-7. Front Panel, Model HFI.

(g) as indicated by Figure 11-5, the AGC and IF outputs of the individual strips are applied via the selector switches, to the AGC comparator and IF output circuits, respectively.

(h) the IF outputs are monitored on front panel meters. These meters are peak indicating devices; thus, for a wide range of input signal amplitudes, these meters will indicate 1.4 volts; the actual RMS output voltage is 1.0 volt. The IF output signals are now designated "A" and "B".

(i) the AGC outputs are compared in the AGC comparator circuit, V-6203. This circuit selects the strongest AGC voltage supplied, and presents this to:

(1) an AGC test point, TP-6202.

(2) the single IF stage on the converter chassis, V-6202.

(3) the RF amplifiers and IF amplifiers in the associated RF unit, which are designated to be supplied with AGC voltage.

(j) the system is designed such that, if the signal input at the antenna varies from about 1 uv to about 100,000 uv, the IF output at J-102 on the strip(s) selected remains substantially at 1.0 volt RMS.

(k) the IF Bandwidth selector switches connect B Plus only to the strip(s) selected.

(l) AGC DECAY controls for each channel control the rate at which the AGC voltage is allowed to change, with changes in the input signal level.

(m) a MANUAL GAIN control is also associated with the AGC circuits. When this control is fully CCW, MANUAL GAIN is switched OFF. As this control is moved slightly CW, a high negative voltage is impressed on the AGC bus, greatly reducing the gain of the IF and RF amplifiers affected. As the control is moved further CW, the negative voltage is reduced, increasing the gain. At some further CW point, the negative voltage impressed by the gain control is less than the normal AGC developed by the strip(s) in use, and again the MANUAL GAIN control is ineffective.

Figure 11-8. View of HFI-1 unit, top, cover removed, unit viewed from front.

d) the two IF outputs are delivered to an associated Detector and Audio unit, where they are converted to audio frequencies.

e) the Audio and Detector unit contains two AM detectors, two Product Detectors, two Beat Frequency Oscillators, and two separate audio amplifier chains. Thus, the system is capable of operation in the AM, CW, SSB and ISB modes.

f) when two such systems are connected for diversity operation, provision is made for selecting the strongest AGC from either system, to control the gain of *both* systems.

g) Figure 11-9, below, shows a simplified block diagram of the interconnection between the RF unit, a control synthesizer, the IF unit, and the Detector and Audio unit in AFC OFF operation. An independent sideband signal is being selected.

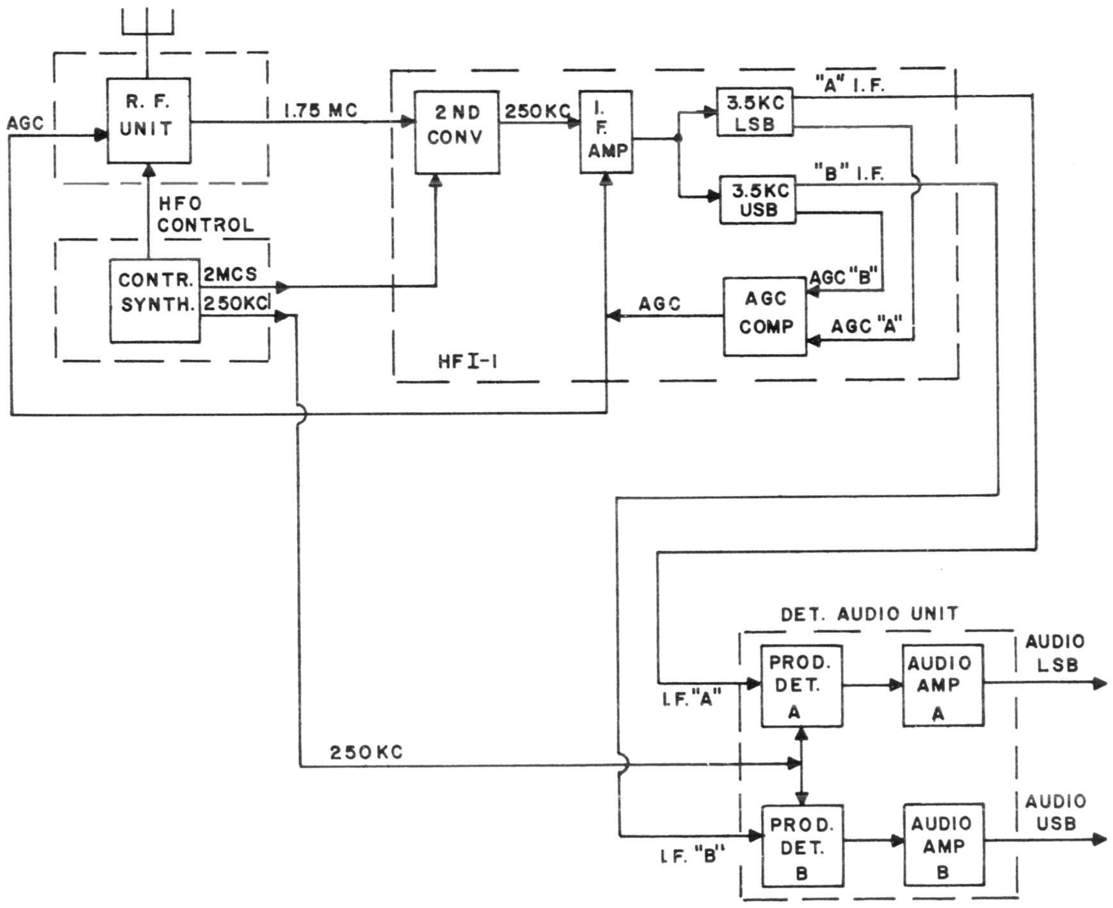


Figure 11-9.

h) assume that the transmitter is precision frequency controlled, that no pilot carrier is being transmitted, that an independent sideband signal is being received at an RF frequency of 10.42 mcs, and that each sideband contains a different voice channel with audio frequencies from approximately 300 to 3000 cycles per second.

(1) the HFO local oscillator in the RF unit is operating at a frequency of 12.7 mcs. The HFO is locked on frequency by a control voltage from the associated synthesizer unit.

(2) the first IF frequency at 1.75 mcs is applied to the second converter, where it is translated to the second IF frequency, 250 KCS. The injection voltage at 2 mcs is supplied by the control synthesizer, hence the second IF is correct. The 250 KC IF signal with full response is applied to two IF strips: Channel A selects the 3.5 KC LSB strip, and Channel B selects the 3.5 KC USB strip.

(3) the AGC outputs of both strips is applied to the AGC comparator, which selects the stronger AGC voltage. This voltage is used to control the gain of certain IF and RF amplifiers.

(4) the "A" and "B" IF outputs of the IF Amplifier are applied to the "A" and "B" product detector circuits in the Detector and Audio unit. The product detector injection at 250 KC is supplied by the control synthesizer; since the IF signal is not subject to drift, and the 250 KC injection frequency is correct, the audio frequencies take their proper places in the spectrum.

(5) the audio output of Channel A represents the voice intelligence in the lower sideband; the audio output of Channel B represents the voice intelligence in the upper sideband.

i) Figure 11-10 shows the interconnection between the RF unit, the control synthesizer, the Model HFI-1 IF Amplifier, the Detector and Audio Unit, and the associated Automatic Frequency Control unit.

It is assumed that:

(1) the transmitter is subject to drift.

(2) a pilot carrier is being transmitted, at a level —20 db below peak sideband power, at a frequency of 10.42 mcs.

(3) an independent sideband transmission is being made; the sidebands contain independent voice channels, as in the previous example.

(4) the AFC ON OFF switch on the Model HFI is in the ON position.

j) the action of the system in AFC ON operation is as follows:

(1) the RF signal at 10.42 mcs, subject to drift, is picked up by the antenna. The local oscillator (HFO) is operating at 12.17 mcs, that is, 1.75 mc above the incoming signal. The control synthesizer sends a control voltage to the RF head to lock the HFO at this frequency. The first IF will drift, however, because the transmitter is drifting.

(2) the injection for the second converter in the IF unit is obtained from the automatic frequency control unit. The TUNING control on the AFC unit is adjusted, to change the second converter injection frequency, to a signal in the region of 2 mcs plus or minus 3 KC. The exact frequency will depend on the frequency of the pilot carrier from the transmitter at this time.

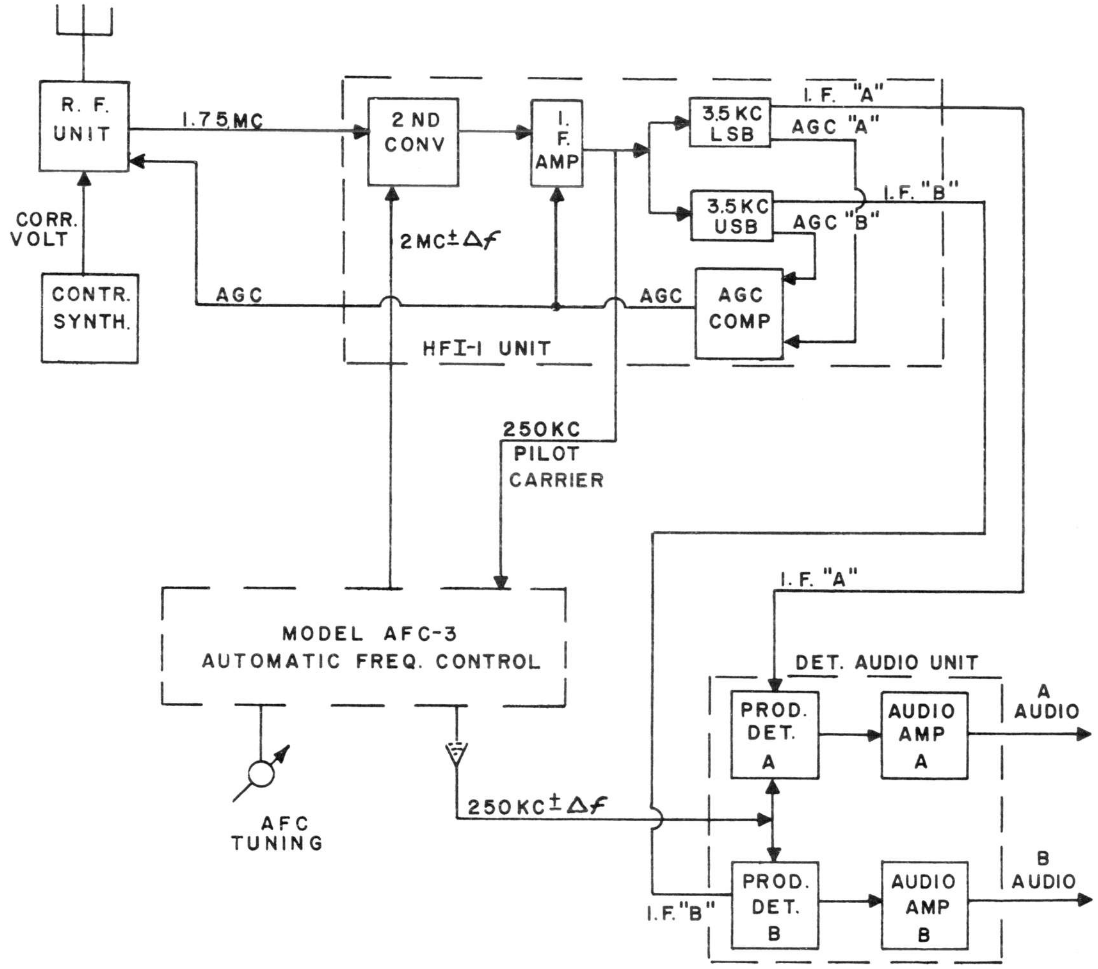


Figure 11-10.

(3) when the TUNING control on the AFC unit is properly set, the second IF frequency is at 250 KC, plus or minus 50 cycles. This signal is sent to the IF amplifier and selected IF strips.

(4) the second IF signal is also sent to the AFC unit; here, it is processed in a selective filter, which passes a narrow band of frequencies, plus and minus 50 cycles about 250 KC. Thus, as the second IF frequency attempts to change, a reference is delivered to the AFC unit, notifying it of the tendency of the pilot carrier to drift.

(5) the automatic frequency control unit performs two primary functions:

(a) it delivers, to the product detectors in the Detector and Audio unit, an injection frequency exactly equal to the pilot carrier frequency at any instant.

(b) it changes the injection frequency to the second converter in such a manner as to keep the second IF frequency always within plus or minus 50 cycles of the desired 250 KC frequency.

Figure 11-12. Front Angle View, Model AFC.

11-8 A Discussion of the Model AFC Automatic Frequency Control

Refer to the Block Diagram, Figure 11-11, and to Figure 11-12, which shows a front angle view of the unit, with all controls and indicators.

a) the second IF frequency, at a nominal 250 KC, with pilot carrier, arrives at J-5000 from the IF amplifier unit.

b) the selective filter Z-5000 passes a narrow band of frequencies plus and minus 50 cycles about 250 KC; this allows the pilot carrier to come through, but strips the remainder of the intelligence.

c) the output of the carrier filter, Z-5000, is applied to a carrier amplifier chain. Three cascade, high gain stages are employed. The gain is high, in order that a pilot carrier, suppressed as much as 30 DB, may be restored. The output of the carrier amplifier chain is applied:

(1) to a front panel carrier level meter circuit, which indicates the relative strength of the reconstructed carrier.

(2) to S-5000, the CARRIER SELECTOR switch. With the switch in the RCC (reconstructed carrier) position as shown, the unit is operating in the RCC mode; the output of the carrier amplifier chain is applied to the product detector circuits of the associated Detector and Audio unit. With S-5000 in the OSC position, the product detector circuits receive the 250 KC signal from the internal 250 KC oscillator and amplifier circuits of the AFC-3.

(3) to a relay amplifier circuit, which operates the FADE relay, K-5000.

(4) to a phase detector circuit.

(5) to an automatic gain control loop, with associated SENSITIVITY control, which maintains the output level of the carrier amplifier chain almost constant for a wide range of the amplitude of the input pilot carrier.

d) the phase detector circuit receives two inputs:

(1) a nominal 250 KC signal from the carrier amplifier chain.

(2) a nominal 250 KC signal from internal 250 KC oscillator and amplifier circuits.

e) the phase detector compares the frequency of the two inputs; the carrier amplifier input is taken as the reference frequency. The phase detector produces a correction voltage, the amplitude and polarity of which depends on the amount and direction of the error between the two 250 KC signals. The correction voltage is applied to the fade relay, K-5000.

f) the fade relay is operated by the relay amplifier, V-5007B, which receives its input from the carrier amplifier circuits. When the amplitude of the carrier is sufficient, the fade relay is de-energized; under these conditions, the FADE INDICATOR is extinguished, and the correction voltage appears at point "A". When the carrier amplitude falls below a certain point, due to carrier "fade", the relay amplifier causes the fade relay to energize. The FADE indicator is illuminated, and the correction voltage does *not* appear at point "A."

g) the correction voltage at point "A" is applied:

(1) to a drift meter circuit; this indicates the amount and direction of carrier drift.

(2) to a drift alarm circuit; the drift alarm is lighted when the drift reaches plus or minus 750 cycles.

(3) to a Varicap circuit of the internal 250 KC oscillator circuit. The correction voltage acts immediately on this circuit to bring the frequency of the internal 250 KC oscillator to the frequency of the pilot carrier.

(4) to a Varicap circuit in the internal 2 mc oscillator — amplifier circuits, via a delay network. This changes the frequency of the 2 mc oscillator circuits in such a manner as to bring the second IF in the IF Amplifier unit back to 250 KC.

h) the delay or memory network delays the 2 mc loop from returning to a neutral position during periods of temporary pilot carrier fade. Under conditions of carrier fade, the FADE relay interrupts the correction voltage from the phase detector. The memory circuit retains the correction voltage stored in it at the time of fade.

i) the "pull in" range of the AFC is about 100 cycles; that is, the second IF must be within plus or minus 50 cycles of 250 KC before the system can be "locked in."

j) the "hold in" range of the AFC is about 2 kcs, for a given setting of the controls.

k) suppose that a drifting signal near the limit of the "hold in" range, should fade. Without the memory circuit, the loop would return to a neutral position, and, when the carrier returned, it would be beyond the "pull in" range, and the system would fall out of synchronization.

l) a RESET button, S-5001, connects the memory circuit to a reference voltage when it is depressed. The reference voltage represents the neutral position of the phase detector circuit, that is, the "zero" correction voltage. The RESET button returns the system to the neutral position, when required.

m) a manual front panel TUNING KCS control allows tuning of the 2 mc oscillator circuits over a plus and minus 3 KC range.

n) in the OSC position of the CARRIER SELECTOR switch, the product detector circuits of the associated Detector and Audio unit receive the 250 KC output of the product detector oscillator circuits of the AFC-3. This is a constant 1 volt RMS output, as opposed to the possibly fading reconstructed carrier.

INTRODUCTION TO 4-CHANNEL INDEPENDENT AGC

In the late 1940's, multi-channel tone telegraph transmission consisted of six teletype channels capable of operating at 60 words per minute on 12 tones in one side of a sideband transmitter. The multiple tones contained on one sideband of an SSB transmitter increased the traffic capability of the transmitter up to 6 times that of a single channel RTTY transmitter. Due to the complexity of this type of transmission, the other sideband was used as an audio order wire to facilitate tuning and system alignment. The intelligence level in the multiple tone side was always raised to a higher level than the audio intelligence side so that the important communication path was not impaired.

It is interesting at this time to note that the early models of the sideband transmitters and receivers were operating under the stability of non-oven controlled crystals and some even with stabilized but free-running oscillators. This meant that the transmitters had to operate with pilot carrier transmission with no greater than a 20 db carrier suppression. When circuit conditions got poor, the receiver operator would then ask for 10 db carrier suppression so that his AFC would hold the circuit in. The transmission of power for pilot carrier operation reduces the amount of power that the transmitter could provide for the intelligence channels. Therefore, the highly stabilized and synthesized sideband transmitters came into being.

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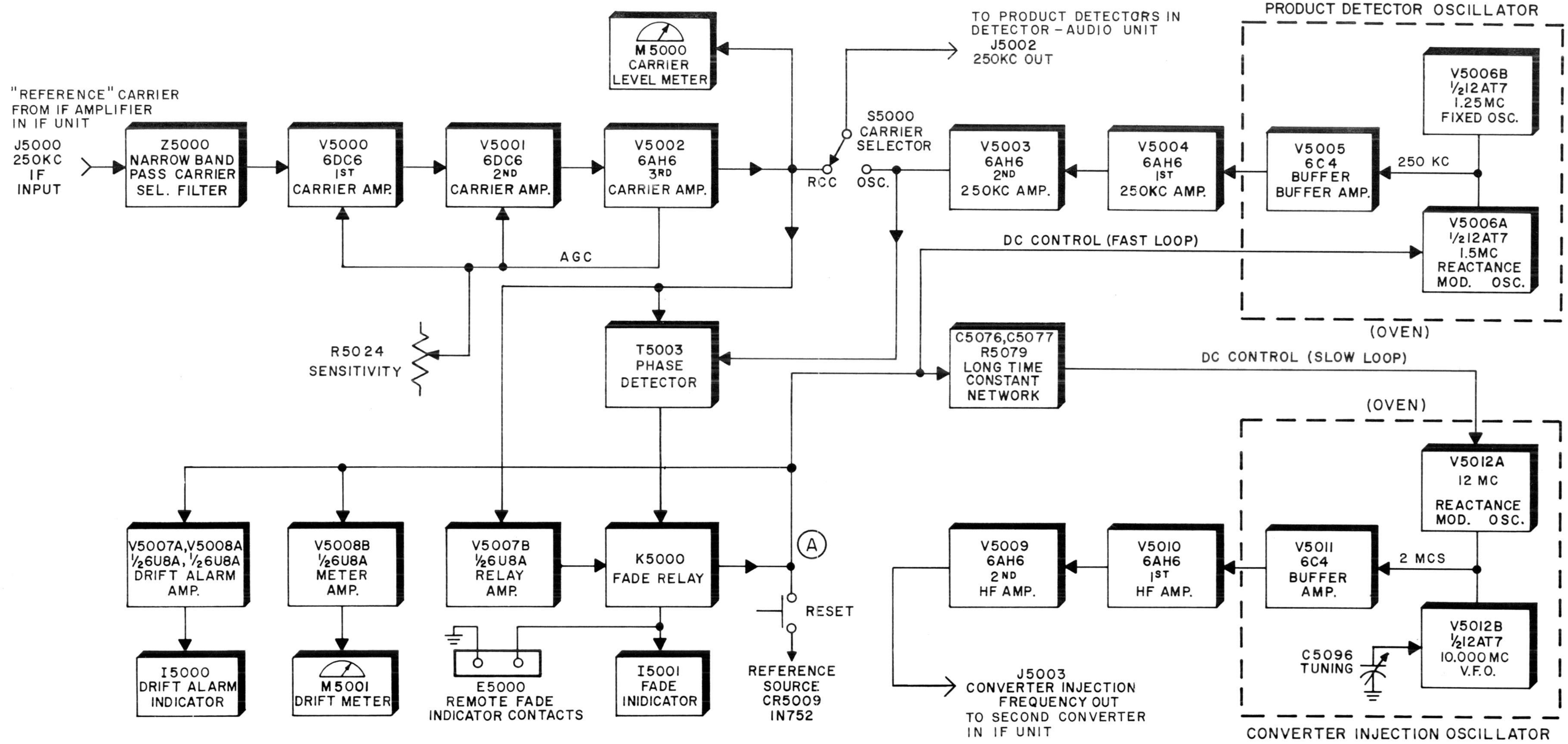

FIGURE II-11 SIMPLIFIED BLOCK DIAGRAM
MODEL AFC-3
AUTOMATIC FREQUENCY CONTROL

The advent of synthesized transmitters and receivers increased circuit reliability and fostered the increase of the number of TTY channels for each sideband. Further, the use of completely suppressed carrier high stability transmitters and receivers evolved the splitting of available frequency spectrum for a single transmitter into two 3 kc voice slots for each sideband of operation, thus enabling a single transmitter to double its intelligence content. This was called Voice Frequency Multiplexing or "Voice Frequency Translation Techniques."

The initial voice frequency multiplexing was accomplished at the audio level. The division of power between the two sidebands, then, was accomplished by increasing the power in one sideband and subordinating the power in the other sideband, taking two channels down simultaneously or increasing two channels simultaneously so that the overall power would not exceed the rating of the transmitter.

This system had major disadvantages, particularly at the receiver end. The AGC action within each IF channel at the receiver can only act upon the average level of power contained within that channel. Therefore, if, at the transmitter, the power in channel A had been increased to a higher level and channel B subordinated, the receive AGC circuit could only act upon the average level contained within A when that was alone on the circuit or the combination of A and B when there was voice appearing in conjunction with the tones.

11-9 A Discussion of TMC Model MSG-1 Independent AGC Receiving System

a) the TMC Model MSG-1 Independent AGC Receiving System is an intermediate frequency receiving adapter which produces four independent audio intelligence channels, each 2775 cycles in width, from a single composite IF signal at a center frequency of 1.75 mcs, with a minimum bandwidth of 12 KCS.

An IF frequency of 455 KCS can also be accomodated, if required. The MSG-1 system was designed primarily for integration with the TMC Model DDR-5 series of receivers, but it can be adapted easily for use with other receiving systems. Single and dual MSG-1 systems are available for normal or diversity operation. Three basic modes of frequency control are possible:

(1) a SYNTHESIZED mode, using injection frequencies derived from the control synthesizer of an associated DDR-5 receiver and a 100.64 KC internal frequency standard in the MSG-1 system.

(2) an INTERNAL mode, using injection frequencies generated by stable oscillators, including the 100.64 KC standard, entirely within the MSG-1 system.

(3) an AUTOMATIC FREQUENCY CONTROL mode, using injection frequencies derived from a transmitted pilot carrier, an associated TMC Model AFC-3 Automatic Frequency Control unit, and the internal 100.64 KC standard in the MSG-1 system.

b) basically, the system accepts a composite four channel IF signal at a center frequency of 1.75 mcs, and produces four independent audio channels, each 2775 cycles wide. In addition, an automatic gain control voltage, based on the intelligence level in each demultiplexed IF channel, is developed and applied to a "four input" AGC comparator circuit. The resultant output of the AGC comparator is an AGC voltage controlled by the strongest of the four independent demultiplexed channels. The AGC voltage is used to control the gain of RF and IF circuits in the associated receiving equipment, and IF circuits in the MSG-1 system. In this manner, intermodulation distortion products, usually a source of trouble in adjacent channel systems, are held down to prescribed limits. The AGC circuits are designed to provide a constant IF output voltage over a wide range of input signal amplitudes; this maintains the audio output reasonably constant for a given setting of the audio level controls.

c) Figure 11-13 shows the Model MSG-1 system. For diversity operation, this system is composed of:

(1) two Model MSA-1 Multiple Sideband Adapters.

(2) two Model MAF-1 Multiple Audio Filters.*

(3) two model MNF-1 Multiple Notch Filters.*

(4) two Model MCG-1 Multiplex Carrier Generators.

*optional items

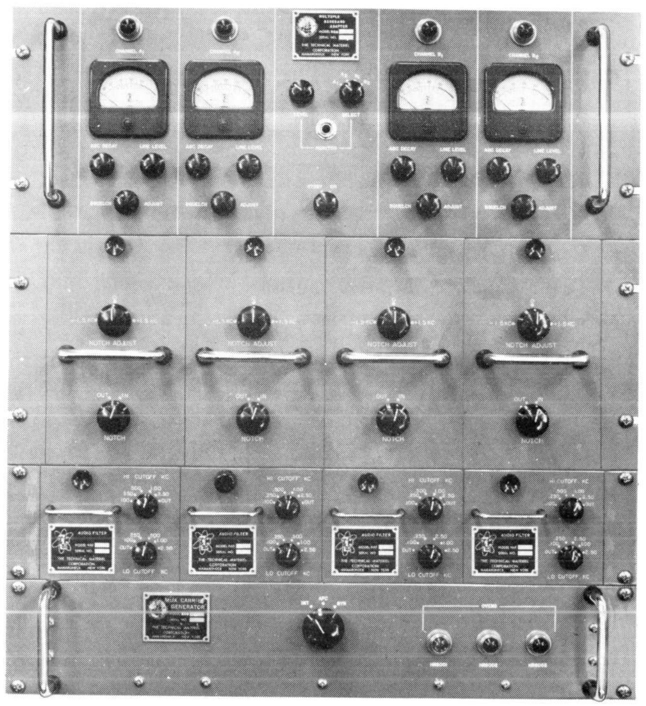

Figure 11-13: Model MSG Independent AGC Receiving System.

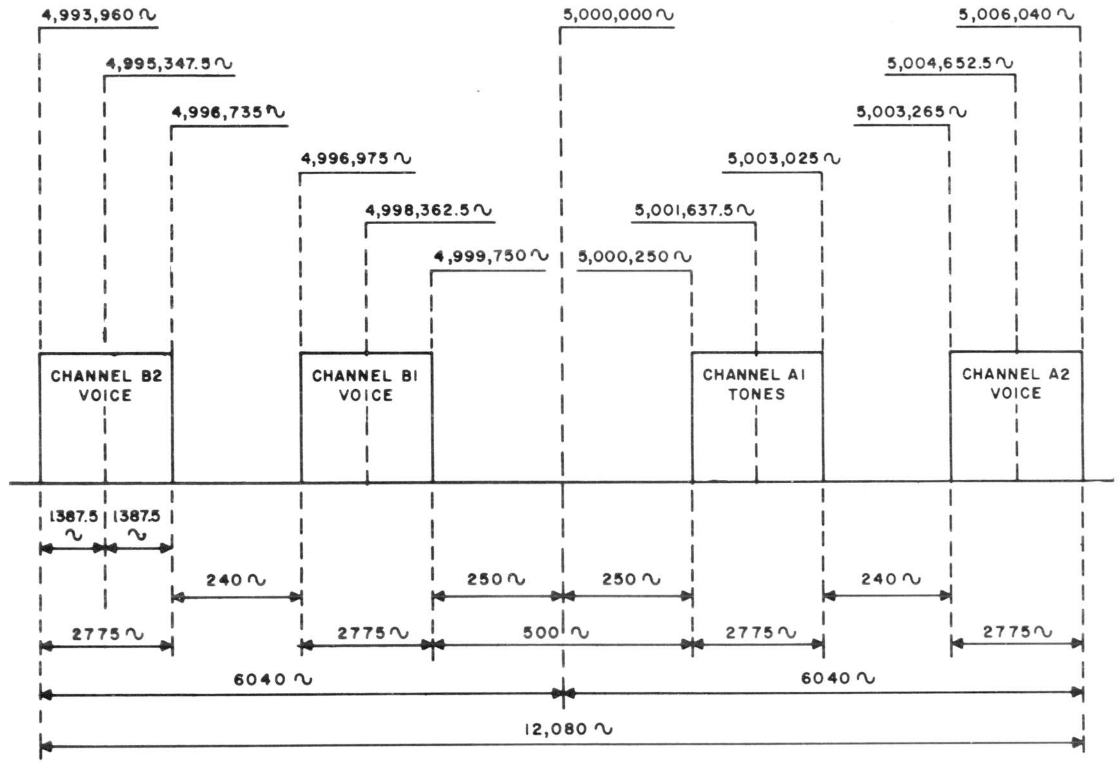

Figure 11-14. Idealized spectrum received four channel signal: 5 mcs. Not to scale.

(5) one Model HFP-1 Power Supply.

(6) one Model MFP-1 Power Supply.

(7) one Model MPS-1 Power Supply.

This system is designed for integration with a Model DDR-5 dual receiver in a diversity arrangement.

d) The remainder of this discussion will concern itself with the operation of a single receiver system.

(1) The Model MCG-1 Multiplex Carrier Generator:

This unit supplies all of the required injection frequencies, as follows:

(a) 250 KC for Channel A_1, B_1, Product Detectors.

(b) 2 mcs for the 1.75 mc to 250 KC converter.

(c) 243.710 KC for the Channel B_2 Product Detector.

(d) 256.290 KC for the Channel A_2 Product Detector.

The MCG-1 also acts as a distribution and switching center for the 2 mc and 250 KC injection frequencies when these originate in either the DDR-5 synthesizer unit Model HFS-1 or in the Model AFC-3 Automatic Frequency Control unit.

A front panel switch labelled "SYN" (synthesized), "INT" (internal), or "AFC" (automatic frequency control), determines the point of origin of the 2 mc and 250 KC injection frequencies. The "SYN" and "AFC" positions of this switch are applicable

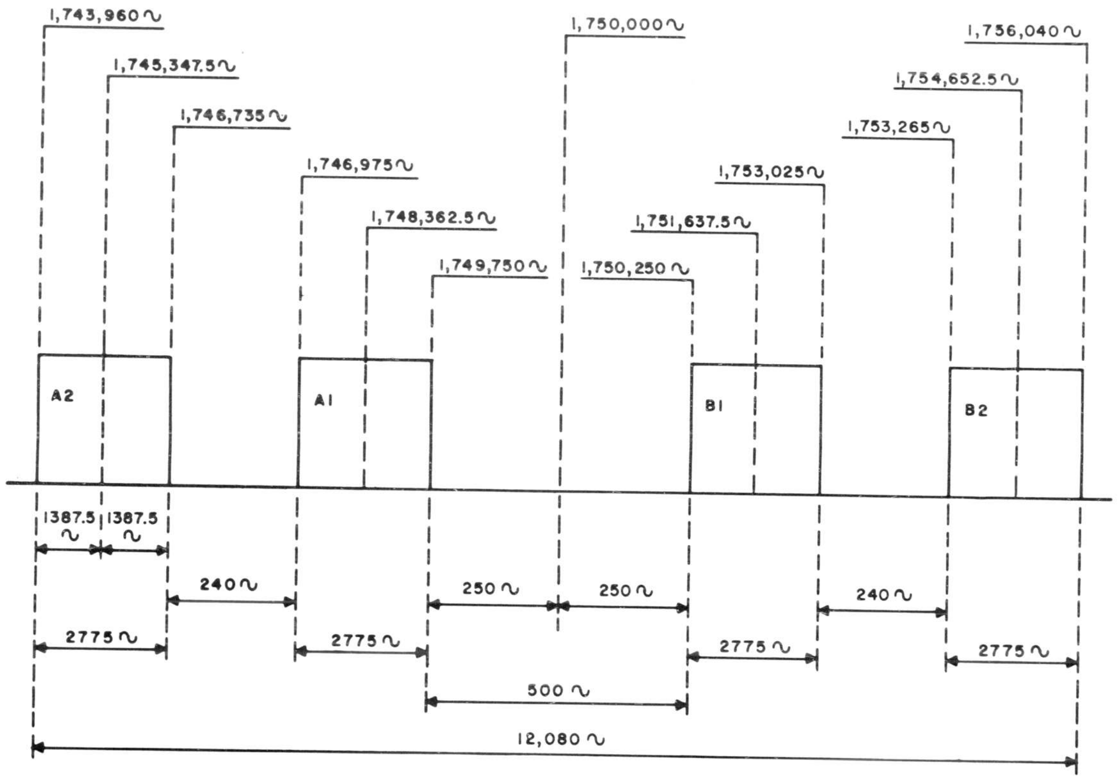

Figure 11-16. Idealized spectrum 1.75 mc IF input to MSA-1. Sidebands inverted.

only to integrated operation with the TMC Model DDR-5 series of receivers.

(2) The Model MSA-1 Multiple Sideband Adapter:

This unit contains a converter, four highly selective IF modules, 4 audio modules (each with a product detector), and AGC circuits.

The MSA-1 receives the composite 1.75 mc four channel IF signal from the associated receiver, plus the three product detector injection frequencies from the Multiplex Carrier Generator, and produces four independent audio channels, each 2775 cycles wide. It also provides an AGC voltage controlled by the strongest of the four demultiplexed IF signals. Additional features such as squelch, AGC decay, audio level and monitor circuits are incorporated.

(3) The Model MNF-1 Multiple Notch Filter:

This unit contains four individual tunable notch filters, one for each demultiplexed IF channel. These units will notch out any single interfering frequency in the 3 KC bandpass of each individual IF channel. The notch filters have unity gain, so that they offer no circuit disruption whether "ON" or "OFF".

(4) The Model MAF-1 Multiple Audio Filter:

This unit contains four individual adjustable passive audio filter units, which may be inserted in series with each audio amplifier channel. Front panel controls allow adjustment of the high and low cutoff frequencies of the audio amplifiers. High and low cutoff frequencies of 100, 250, 500, 1000 and 2500 cycles per second are available.

e) Figure 11-14 shows the idealized spectrum of a four channel RF signal at a carrier frequency of 5 mcs. The following characteristics of the signal should be noted:

(1) the carrier frequency, 5 mcs, is at the center of the spectrum. The carrier may be present, partially suppressed, or attenuated completely.

(2) the total bandwidth is 12,080 cycles.

(3) Channels B_2 — B_1, and A_2 — A_1 are separated by a guard band 240 cycles wide.

(4) Channels A_1 — B_1 are separated by a guard band 500 cycles wide (actually, two guard bands, each 250 cycles in width). The carrier frequency position is in the center of this guard band.

(5) each channel of information is 2775 cycles wide.

(6) each sideband, including guard bands, is 6040 cycles wide.

(7) the signal is not inverted; that is, Channels B_2, B_1, A_1 and A_2 appear in the spectrum from left to right.

(8) Channels A_2 and B_2 may be referred to as the "OUTBOARD" or "TRANSLATED" channels.

(9) Channels A_1 and B_1 may be referred to as the "INBOARD" or "DIRECT" channels.

f) in Figure 11-14, Channels B_2, B_1 and A_2 each contain an independent voice channel. Channel A_1 contains 16 channels of narrow band frequency shift teletype intelligence. Thus, the total signal contains 19 discrete channels of intelligence in a spectrum width normally reserved for a single conventional "AM" signal. The signal shown in Figure 11-14 is constructed by means of sophisticated multiplexing methods at the transmitter site, and is put on the air at a relatively high peak envelope power level by independent sideband transmission techniques.

g) the signal of Figure 11-14 is picked up by the receiving antenna at a distant point, and is conveyed to the input circuits of a Model DDR-5 receiver. (See point "A" of Figure 11-15, which shows the block diagram of the Model MSG-1 system, together with certain units of the Model DDR-5 receiver. Continual reference will be made to Figure 11-15. (Fold out sheet at end of chapter.)

h) to the first converter of the receiver is applied the amplified RF signal and an injection frequency of 6.75 mcs. The 6.75 mc injection

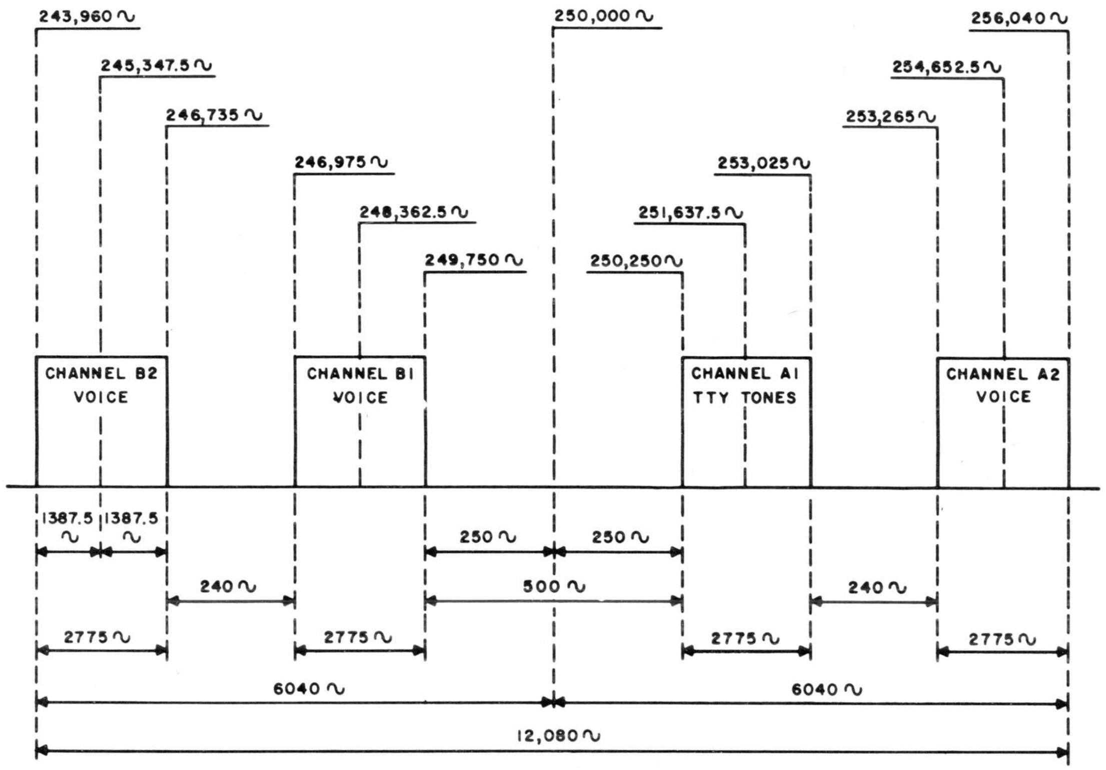

Figure 11-17. Idealized spectrum at 250 KC. Sidebands not inverted.

frequency is supplied by the local oscillator or "HFO", and is held fast to this frequency by the associated control synthesizer.

i) the first converter is tuned to pass the difference between the received RF signal frequency and the local oscillator frequency; this is an intermediate frequency centered at 1.75 mcs. The response of the associated tuned circuits is such that the total 12 KC bandpass is readily accommodated. The resultant spectrum is shown in Figure 11-16. This signal appears at point "B" of Figure 11-15. Note that the spectrum of Figure 11-16 retains all of the essential characteristics of the originally received signal except that sideband inversion has taken place. The signal is applied to the MSA-1 Multiple Sideband Adapter in this condition.

j) in the Model MSA-1, a second conversion and sideband inversion takes place. The 1.75 mc signal is mixed with a 2 mc injection frequency and a new IF at 250 KC is produced. This is the spectrum shown in Figure 11-17; this signal is found at point "C" of Figure 11-15. Note that the new spectrum at 250 KC retains all the characteristics of the originally received signal, and that it is not inverted. In Figure 11-15, the 2 mc injection frequency originates in the receiver synthesizer and is delivered to the MSA-1 from the MCG-1 Multiplex Carrier Generator via the mode switch, which is shown in the "SYN" position.

k) the second converter, V-6501, is followed by four individual IF

preamplifiers, each tuned to pass one of the four IF channels as shown below:

STAGE	*CHANNEL DESIGNATION*	*IF CENTER FREQUENCY*
V-6502	B2	245,347.5 cps
V-6503	B1	248,362.5 cps
V-6504	A1	251,637.5 cps
V-6505	A2	254,652.5 cps

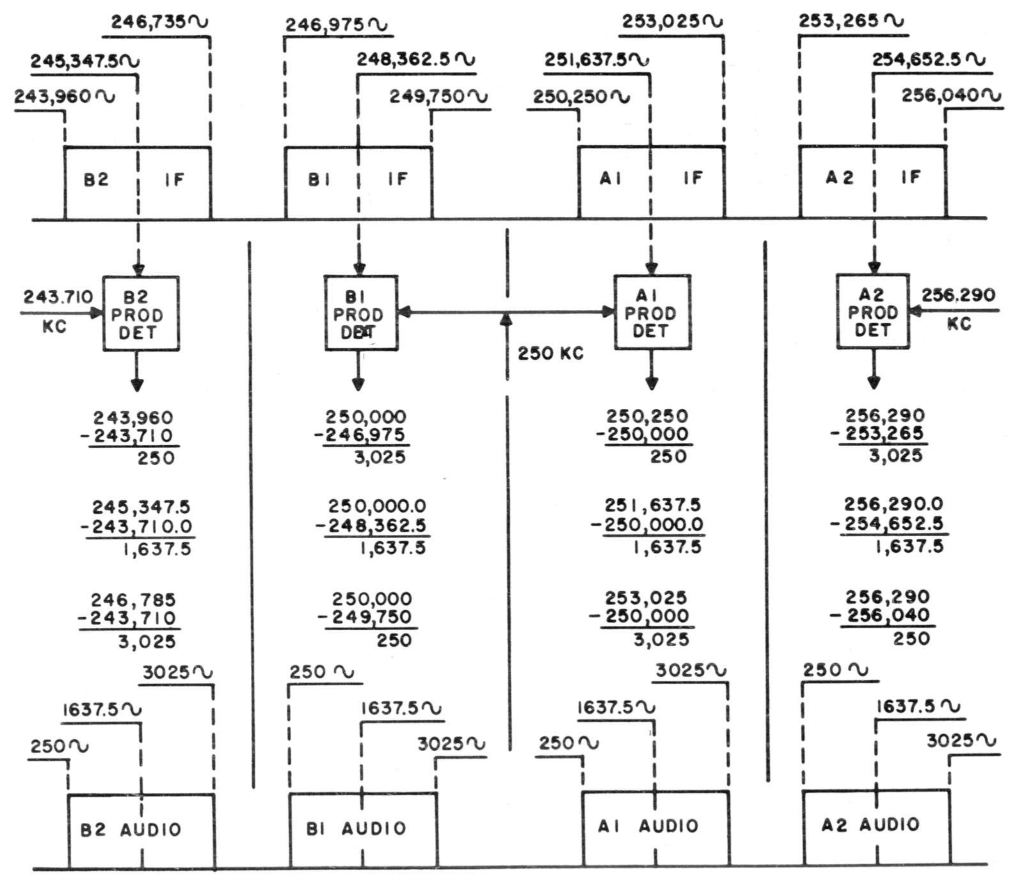

Figure 11-18. Resultant four independent audio channels after demultiplexing in product detectors.

absolute selectivity is not of paramount importance at this point, because subsequent filtering and IF amplification will complete the demultiplexing of the signal at the IF level. At this point, Fig. 11-15, coaxial connections are provided for connection to and from the individual notch filters.

It should be noted that the IF preamplifiers eliminate the carrier frequency, 250 KC, should it happen to be present.

l) in the "AFC" mode, a "reference" carrier must be extracted at a point between the 250 KC converter and the four individual IF preamplifiers; this reference, based on the drift of the transmitter, is sent to the Model AFC-3 unit in order that the AFC-3 may "correct" the 2 mc and 250 KC injection frequencies in response to transmitter frequency drift.

m) the notch filter jumpers are shown shorted, since it will be assumed that the notch filter switches are in the OFF position. Beyond this point, the four individual IF channels are applied to individual IF modules; each IF module contains:
   (1) a selective crystal filter, which passes only the desired IF spectrum.
   (2) two stages of IF amplification.
   (3) an AGC generator circuit which produces a negative DC voltage, the amplitude of which is based on signal strength in the particular channel under consideration.

n) the four independent IF signal outputs are applied to four individual audio modules, each of which contains, essentially:
   (1) a product detector circuit.
   (2) an audio voltage and power amplifier chain.
   (3) squelch circuits.
   (4) level control circuits.

o) the four AGC outputs of the IF modules are applied to a pair of AGC comparators, as shown in Figure 11-15. The AGC voltage, as developed in the IF module, is applied also to the first IF amplifier in the module, and to the appropriate IF preamplifier feeding the module. This AGC voltage is also used in the squelch circuits of the audio modules, but this is not shown in the block diagram.

   The resultant output of the AGC comparator pair is a negative voltage, which is delivered to the "front end" of the associated receiver, and to certain IF circuits. The action of the AGC circuits provides reasonably constant audio output for a wide range of input signal amplitudes, and holds down the adjacent channel intermodulation distortion to a minimum.

p) Each product detector receives its particular IF spectrum, and an injection frequency. The inboard product detectors both receive the 250 KC signal, and Channel B_2 and A_2 product detectors receive the 243.710 KC and 256.290 KC signals, respectively, from the MCG-1 Multiplex Carrier Generator.

   The product detectors are designed to pass the difference frequencies; the resultant audio spectrums are shown in Figure 11-18.

   The individual audio amplifiers have flat response from 50 cps to 10,000 cps, and work into a 600 ohm load. Jumpers are included in each audio amplifier for interconnection from and to the audio filter units.

q) since channel A consists of 16 channels of narrow band frequency shift teletype, further demultiplexing will be required to separate the 16 channels. This operation is accomplished in equipment external to the MSG-1 system.

r) Summary:
   (1) with the MSG-1 system, it is possible to demultiplex a four channel IF signal at 1.75 mcs to produce four independent audio outputs.

   (2) the frequency stability of all converter and product detector injection frequencies is of paramount importance. This is particularly true when adjacent channels are very close together, as in the case of channel A_1, where 16 discrete channels are contained in an audio spectrum 2775 cycles in width. An excessive variation of the product detector injection frequency, for example, could render this portion of the signal useless.
   (3) the optimum frequency stability is obtained when the mode switch on the Multiplex Carrier Generator Model MCG-1 is in the "SYN" position, and the system is integrated with the Model DDR-5 receiver. Under these conditions, the first converter in the receiver, the second converter injection of 2 mcs, and the product detector injection frequencies of 250 KC, 243.710 KC, and 256.290 KC are controlled by a 1 megacycle standard with a stability of 1 part per 100,000,000 per day, and a 100.64 KC standard with a stability of 1 part per 10,000,000 per day. This mode requires that the transmitter at the distant point be similarly precision frequency controlled.
   (4) under certain conditions, particularly when the transmitter at the distant point is not precision frequency controlled, it may be desirable to operate the system in the "AFC" mode. In this case, a TMC Model AFC-3 Automatic Frequency Control corrects the injection frequencies so that the audio output is within 1 cycle of the transmitted intelligence. The transmitter must radiate a certain amount of pilot carrier, which is used as a reference.
   (5) consider the possibilities of this equipment; it is possible, under conditions of optimum frequency control, to receive, in each of the four "slots", 16 or more channels of narrow band frequency shift teletype information at high speed. This would provide at least 64 discrete channels of information in a spectrum about 12 KC wide.
   (6) the MSG-1 equipment can be supplied in a variety of configurations.
      (a) to accept either an IF signal at 1.75 mc, or 455 KC.
      (b) to demultiplex at the audio level with product detector injection frequencies of 250 KC and 250 KC plus and minus 6.29 KC or 6.25 KC, to satisfy any requirement.

Reference Material

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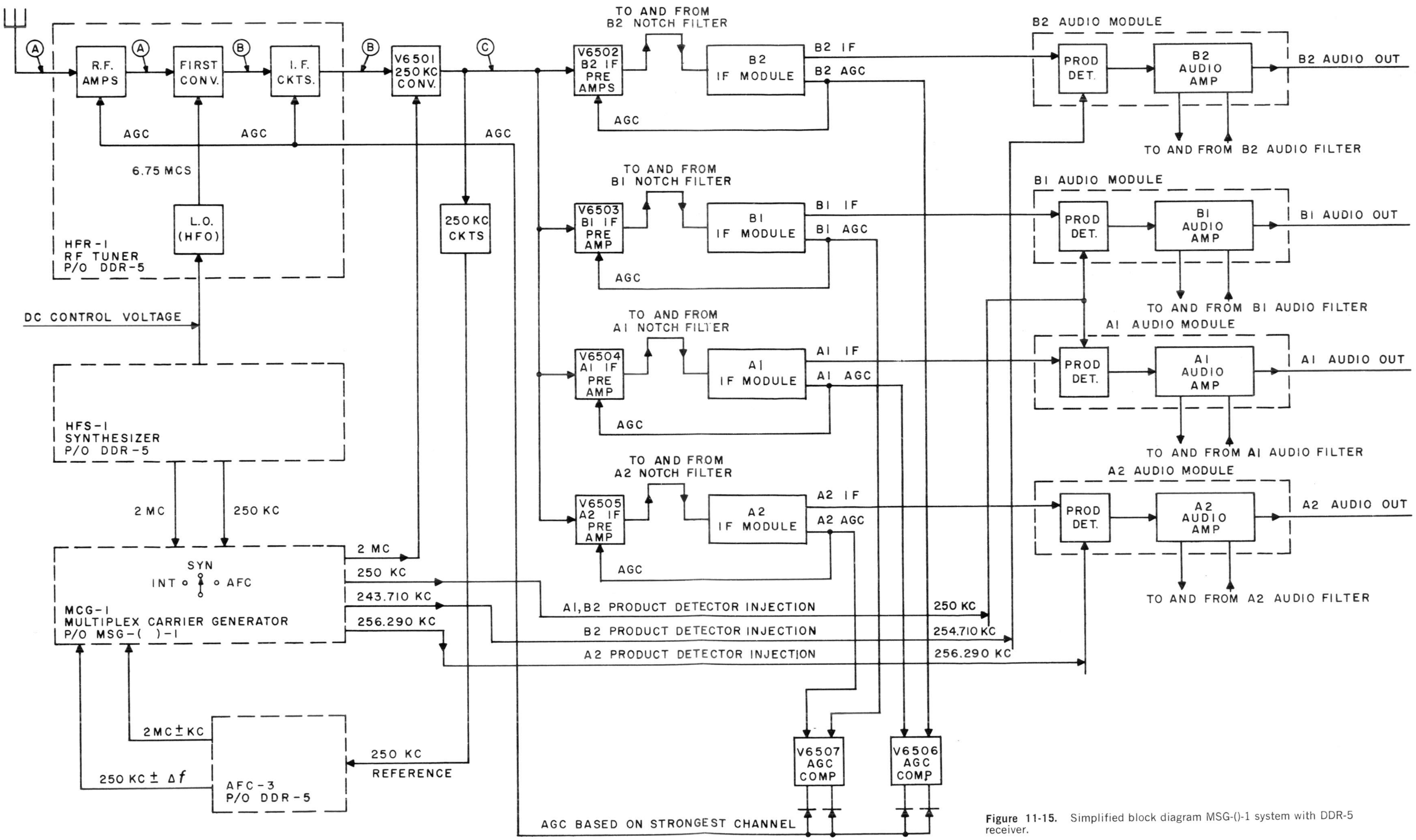

Figure 11-15. Simplified block diagram MSG-()-1 system with DDR-5 receiver.