SPECIAL DEFENSE EDITION

THE RADIO AMATEUR'S HANDBOOK

A Manual for Radio Training Courses



PUBLISHED BY THE
AMERICAN RADIO RELAY LEAGUE

$1.00

Pvt. Vernon H. Eidsen ASN-16146841

SPECIAL DEFENSE EDITION

NINETEEN-FORTY-TWO

Property of Martha Shirley

THE RADIO AMATEUR'S HANDBOOK



BY THE HEADQUARTERS STAFF OF THE
AMERICAN RADIO RELAY LEAGUE

PUBLISHED BY THE AMERICAN RADIO RELAY LEAGUE,
INCORPORATED, WEST HARTFORD, CONNECTICUT

Other Publications BY THE AMERICAN RADIO RELAY LEAGUE

QST

¶ America's oldest radio magazine and the accepted journal of the amateur, *QST* reports each month in direct and analytical style the rapid developments in amateur radio. Its subscription price is $2.50 per year in U. S. and Possessions.

The Radio Amateur's Handbook

¶ You are reading the Special Defense Edition. The standard edition, in addition to telling the things needed for the comprehensive understanding of amateur radio, contains chapters on the design, construction and operation of many pieces of radio equipment and on operating procedure. Paper bound, $1.00; $1.25 outside continental U. S. A. Buckram bound, $2.50. Also available in Spanish, paper bound, $1.50.

The Radio Amateur's License Manual

¶ This 40-page booklet gives the necessary information, including typical or paraphrased examination questions and answers, to enable one to pass the government examination for amateur operator's license. Price 25c.

The A.R.R.L. Antenna Book

¶ A comprehensive manual on the design and construction of many types of antennas. 144 pages in this format, profusely illustrated. Price 50c.

Hints and Kinks

¶ A compilation of hundreds of good radio ideas, savers of time and money for the experimenter. 128 pages in this format. Price 50c.

How to Become A Radio Amateur

¶ This publication is recognized as the standard elementary guide for the prospective radio amateur. 48 pages, just revised. Price 25c.

Lightning Calculators

¶ Circular slide rules, on 8½ x 11-in. bases, of special cardboard, with satisfactory accuracy. Two types:

Type A solves problems in frequency, wavelength, inductance and capacity. Price $1.

Type B makes computations involving voltage, current and resistance. Price $1.

THE AMERICAN RADIO RELAY LEAGUE, INC.

WEST HARTFORD, CONNECTICUT

Foreword

SINCE 1926 the American Radio Relay League has published *The Radio Amateur's Handbook*, primarily as a manual for the tens of thousands of practitioners of amateur-radio communication all over the world. The League is a nonprofit society of such radio specialists and one of its functions has been to make reliable literature on the art available at modest cost. With most amateur stations closed down because of the war, and with the whole nation engaged in the great struggle, the *Handbook* now has a new mission to perform. With many hundreds of radio training classes in formation in the universities, colleges, high schools and community centers of the country, this special revision of the current edition has been prepared in the belief that it can serve as a more valuable but inexpensive text for defense training courses.

Devoted to a fast-moving and progressive science, the *Handbook* has required sweeping and virtually continuous modification throughout its life. Its annual rewriting is a major task in the family life of the headquarters group of the League at West Hartford, where most of the technically-skilled specialists of the staff participate in the work under the general technical editorship of George Grammer, technical editor of *QST*. Some measure of the book's acceptance is to be found in the fact that there have been nineteen previous editions, with thirty-four pressings and total sales of well over three-quarters of a million copies.

In this special edition for training courses we have retained everything from the current standard edition that seems useful to the task in hand but have eliminated those portions that would serve only to distract or encumber the student. We have omitted everything that treated particularly of this League, the advertising that usually accompanies the *Handbook*, and the chapters that ordinarily deal with the home construction of specific pieces of apparatus for amateur stations. In place of the latter we have prepared a profusely-illustrated new chapter calculated to acquaint the student with the appearance and the circuit diagrams of representative types of apparatus. A valuable new chapter gives fundamental instruction in the solution of formulas and the reading of graphs, something that has long been needed even in the *Handbook*. And because many defense courses will also incorporate instruction in the international telegraph code, the newly-added material contains a chapter on learning the code and some suggestions for a classroom code table. Our aim throughout has been to write an understandable text for busy, practical people of average education, employing a minimum of mathematics. A major objective has been to provide the answers to the questions that naturally arise in the course of study. The material has been so arranged as to make it readily possible to find what is wanted, a multitude of headings identifying subjects at a glance. Information has been presented concisely but with copious cross-references to permit the background always to accompany the subject under consideration. We have endeavored to employ cross-references in such quantity that no treatment of any subject can be considered "too technical," since the references will eventually lead the reader, if he needs it, to the applying fundamentals themselves. The sequence of presentation has been planned to lend itself to an ordered course of instruction. Necessarily compact (as is any good text), information is deliberately presented without sugarcoating, but every effort has been made to make it understandable and to avoid saying things in such a way that they are intelligible only to those who already know the subject thoroughly!

A word about the reference system: It will be noted that each chapter is divided into serially-numbered sections. The number takes the form of two digits or groups connected by a hyphen. The first figure is the chapter number, the second the section number within the chapter. Cross references in the text take such a form as (§4-7), for example, which means that the subject referred to will be found discussed in Chapter Four, Section 7. Illustrations are numbered serially in each chapter. Thus, Fig. 502 can be readily located as the second illustration in Chapter Five. There is a carefully-prepared index at the end of the reading pages.

We here shall be very happy if this special edition of the *Handbook* can be of as much assistance in the national effort as its predecessors have been to licensed amateurs.

KENNETH B. WARNER
Managing Secretary, A.R.R.L.

WEST HARTFORD, CONN.
February, 1942

SPECIAL DEFENSE EDITION

First Printing, February, 1942.....40,000 Copies
Second Printing, April, 1942.....40,000 Copies
Third Printing, September, 1942.....40,000 Copies

Copyright 1942 by The American Radio Relay League, Inc.
Copyright secured under the Pan-American Convention.
All rights of translation reserved.

PRINTED IN U. S. A. BY
THE RUMFORD PRESS
CONCORD, NEW HAMPSHIRE

Contents

Frontispiece:

Schematic Symbols

Chapter 1 Formulas and Graphs 7

Chapter 2 Electrical and Radio Fundamentals 21

Chapter 3 Vacuum Tubes 42

Chapter 4 Radio-Frequency Power Generation 58

Chapter 5 Radiotelephony 84

Chapter 6 Keying 106

Chapter 7 Receiver Principles and Design 111

Chapter 8 Power Supplies 143

Chapter 9 Wave Propagation 156

Chapter 10 Antenna Systems 162

Chapter 11 Radio Equipment 196

Chapter 12 Measurements and Measuring Equipment 225

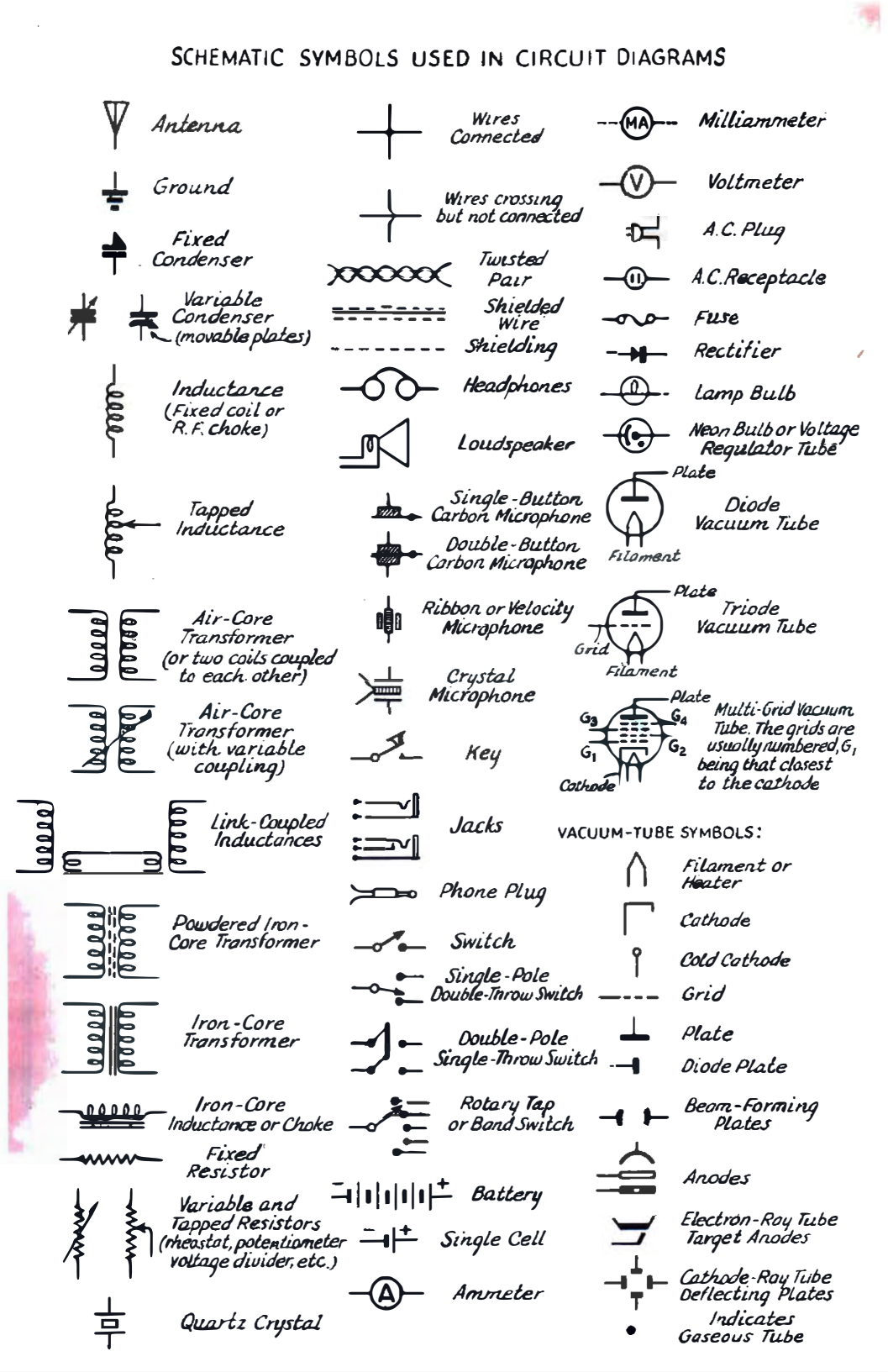
Chapter 13 Workshop Practice 249

Chapter 14 Learning the Code 254

Chapter 15 Miscellaneous Data 268

INDEX 281

SCHEMATIC SYMBOLS USED IN CIRCUIT DIAGRAMS



Formulas and Graphs

IN ORDINARY radio work mathematical applications reduce to the solution of formulas, or equations, and to reading (and sometimes plotting) graphs. Only elementary mathematical knowledge is required, but that knowledge should include a few fundamentals in a number of branches of the subject. This chapter is intended to cover those fundamentals in sufficient detail to give the reader an understanding of principles, as well as to demonstrate the methods of arriving at solutions to the formulas he will find in later chapters. Familiarity with elementary arithmetic is assumed, but a review of decimals and the method of extracting square root is included for the benefit of those whose daily mathematical experience is limited to the very simple arithmetical operations which most people find sufficient for non-technical pursuits.

1-1 DECIMALS

Notation — Our system of writing numbers is based on multiples of ten, with the position of the digit indicating the multiple to be used. Fundamentally we have only the digits 1, 2, 3, 4, 5, 6, 7, 8, 9 and 0; all our numbers are based on the positions of these digits in relation to others in the number. Any number is actually a sum. Thus the number 4826 means the sum of

6	units
2	tens
8	hundreds (10×10)
4	thousands ($10 \times 10 \times 10$)

which we write 4826

Each time we move one place to the left the digit in that place is multiplied by an additional factor of 10.

Since this system lends itself so readily to calculations, it is very convenient to extend it to numbers less than one — in other words, to include fractions. Looking at the number 4826 from a slightly different viewpoint, it is evident that we divide by ten each time we move one place to the right. There is no reason why this process cannot be continued indefinitely. We need some method of distinguishing the units column, and this is accomplished by placing a period or *decimal point* to its right, thus marking off the whole numbers from the fractions. Continuing the division, the multiple for the first place to the right of the decimal point (first decimal place) is 1 divided by 10, or 1/10, the second place is 1/10 divided by 10, or 1/100, the third place is 1/100 divided by 10, or 1/1000, and so on. By converting fractions to make their denominators 10 or multiples of 10 we can fit them neatly into the system.

Forming decimals — Fractions can be converted by multiplying both the numerator and denominator by a number which causes the resulting denominator to be a multiple of ten. For example, the fraction ½ can be converted as follows:

$$\frac{1}{2} \times \frac{5}{5} = \frac{5}{10}$$

which is then written 0.5. The zero is used to show that there are no whole numbers and to add emphasis to the fact that the decimal point is present — the decimal point is small and sometimes is overlooked. A fraction with a larger denominator would be converted to the next larger multiple of ten:

$$\frac{2}{25} \times \frac{4}{4} = \frac{8}{100} = 0.08$$

and similarly for still larger denominators. The zero is placed between the decimal point and the 8 to bring the latter into the "hundredths" column.

Multiplying the numerator and denominator of a fraction by the same number does not change the ratio of numerator to denominator and hence does not change the value of the fraction, so we can for instance, do the following:

$$\frac{5}{10} \times \frac{10}{10} = \frac{50}{100} = 0.50$$

$$\text{or } \frac{5}{10} \times \frac{100}{100} = \frac{500}{1000} = 0.500$$

That is, zeros may be added to or subtracted from columns on the right-hand side of a decimal fraction without changing the value of the decimal. We can just as readily put zeros to the right of the decimal point in a number which has no fraction; 237 and 237.0000 are exactly the same number.

The multiplier by which a fraction can be converted to a decimal can be found by dividing the denominator into a multiple of 10. Thus in the example

$$\frac{2}{25} \times \frac{4}{4} = \frac{8}{100} = 0.08$$

the multiplier is 100 ÷ 25, or 4. The 4 is then immediately multiplied by 2 to give 8, and then the denominator (100) is dropped by writing the result 0.08. This process can be shortened to the very simple one of dividing the numerator of the fraction by its denominator. To do this it is first necessary to add decimal places after the 2 so that we can perform the division by familiar methods. Thus we have 2.00 divided by 25, which for purposes of division can be considered the same as dividing 200 by 25. By short division,

```
   .08
25)2.00
```

Note that the decimal point in the quotient is placed directly above the decimal point in the dividend, and that the place intervening before the first digit is filled by a zero.

Mixed numbers, such as $6\frac{3}{4}$, can be changed to decimals either by making an improper fraction (which in this case would be $\frac{27}{4}$) and then dividing as above, or by changing the fraction alone into a decimal and simply placing the integer (an integer is a whole number) to the left of the decimal point. Thus $\frac{3}{4}=0.75$; adding the 6 gives 6.75.

Rules for placing the decimal point — In addition and subtraction of numbers containing decimals the numbers are set down in a column with the decimal points aligned vertically. Thus to add 24.85, 218.003, and 4.6, the procedure would be

```
  24.85
 218.003
   4.6
 -------
 247.453
```

the addition being carried out as though the decimal point were not there.

Subtraction is handled similarly; for example, suppose 31.028 is to be subracted from 286.2:

```
  286.200
-  31.028
 --------
  255.172
```

In this case addition of the zeros in the first number is helpful, to avoid subtraction from "unfilled" places. In both addition and subtraction the decimal point in the answer is placed directly below the others.

When two decimal numbers are multiplied, the number of decimal places in the result is equal to the sum of the decimal places in the original numbers. Thus,

```
  23.65
 × 4.8
 -------
 18920
  9460
 -------
 113.520
```

The decimal places in the result are counted off from the right, including any terminating zeros that may arise as a result of the multiplication. It should not be hard to see the reason for this rule, remembering that the decimal portions of the numbers are really fractions. Without decimals, the fractional parts of the numbers in the example would be 65/100 and 8/10, respectively, and the normal process of multiplication would be to convert both mixed numbers to improper fractions, one having a denominator of 100 and the other 10. On multiplication the resulting denominator would be 1000, or three decimal places.

Division reverses the rule for multiplication. The number of decimal places in the quotient is equal to the number in the dividend minus the number of places in the divisor. Thus in dividing 163.122 by 52.62 we have

```
52.62)163.122(3.1
   15786
   -----
    5262
    5262
    ----
```

Another example, 22.5 divided by 0.15:

```
0.15)22.50(150
   15
   --
    75
    75
    --
     0
```

illustrates the case where the number of decimal places in the divisor is greater than that in the dividend. There is a simple rule covering such cases: When the number of decimal places in the divisor is greater than that in the dividend, add zeros to the right of the dividend until both have the same number of places, when the quotient will come out as a whole number.

In most cases division can be simplified by using the following form:

```
          3.1
      ---------
52.62)163.122
      15786
      -----
       5262
       5262
       ----
```

Starting from the decimal point in the dividend, count off to the right the number of decimal places in the divisor, place the decimal point for the quotient directly above and after this place, then proceed as in ordinary division. If there are no decimal places in the divisor, the decimal point in the quotient goes directly above that in the dividend. This form has the advantage that the division can be continued conveniently in case the quotient is not "even." For example, to divide 23.84 by 13.3:

```
          1.7 9 2 +
133 ) 23.8 4 0 0
      1 3 3
      -----
      1 0 5 4
        9 3 1
      -------
        1 2 3 0
        1 1 9 7
      ---------
            3 3 0
            2 6 6
          -------
              6 4
```

Zeros can be added to the dividend until the answer is carried out to any desired number of decimal places.

1-2 POWERS AND ROOTS

Powers — The power of a number is the number of times it is used as a *factor*, or multiplier. Thus the number 3 can be considered to be the product of 3×1 and, having been used once as a factor, 3 is said to be the *first power* of itself. If we multiply 3 by 3 we have used 3 twice as a factor, hence the product, 9, is called the *second power* of 3. Multiplying by 3 again gives 27, which is the *third power;* once more gives 81, the *fourth power;* and so on. A number is said to be raised to a certain power when it is used that number of times as a factor or multiplier.

When a number is to be raised to a certain power it is indicated by writing the number of the power as a superscript to the right of the number itself; thus 3 raised to the 4th power is written 3^4. The number which denotes the power is called an *exponent.* The second power is frequently called the "square," because if the number represents the length of a line, multiplying by itself is equivalent to finding the area of a square having sides of that length. Similarly, the third power is called the "cube," since the multiplication would give the volume of a cube having edges equal in length to the number.

Roots — The converse of a power is called the *root* of a number. It is that number which when raised to the required power would be the number given. For example, 3 is called the third (or cube) root of 27, because 3 raised to the third power is 27. Likewise, 4 is the second or square root of 16, because 4^2 (4 raised to the second power) is 16. The fourth root of 16 is 2, because 2 raised to the fourth power equals 16, and so on. When a root is to be found the *radical sign* $\sqrt{}$ is used, with the order or number of the root (called the *index*) written in the opening of the "V." For example, the 4th root of 16 is written

$$\sqrt[4]{16}$$

When the second or square root is to be taken, the index usually is omitted; the order of the root in this case is understood to be 2.

Square root — The numerical operation of extracting square root is frequently necessary in solving problems. The method can best be described by means of an example. Suppose the square root of 528.367 is to be found. Set down the number and point off groups of two places, starting with the decimal point and going both to the right and left:

$$\sqrt{5'28.36'70}$$

Note that a zero has to be added to the decimal to make two figures in the last group; this is necessary on the right-hand side but not on the left. Now find the largest integer whose square is contained in the first group, which in this case is the single digit 5. The largest number whose square is contained in 5 is 2, the square of which is 4 (the square of 3 would be 9, which is greater than 5). Write this root above the 5 and write the square below the 5, then subtract:

```
      2
√5'28.36'70
  4
  1
```

Now bring down the next group of numbers alongside the remainder, 1, and draw a vertical line to the left of the 1:

```
      2
√5'28.36'70
  4
 |128
```

Double the root (2) already found and place it to the left:

```
      2
√5'28.36'70
  4
4 |128
```

Cover the last number (8) and see how many times the divisor, 4, will go into the remaining number, 12. The answer is obviously 3. Place the three in the root above the 28 group and also write it alongside the divisor, 4. Then multiply the complete divisor, 43, by 3 and write the result under the 128:

```
      2 3
√5'28.36'70
  4
43 |128
   129
```

At this point it is necessary to subtract again, but unfortunately the subtrahend is just a bit too large. If 3 gives a product which is too large, obviously the next step is to try 2, so we

substitute 2 for 3 and try again, getting a smaller number which can be subtracted:

```
    2 2
  √5'2 8.3 6'7 0
    4
42  | 1 2 8
      8 4
      4 4
```

Bring down the next group, 36, and repeat the procedure — double the root already found (22) and find the largest multiplier which can be used:

```
    2 2  9
  √5'2 8.3 6'7 0
    4
42  | 1 2 8
      8 4
449 | 4 4 3 6
      4 0 4 1
        3 9 5
```

Continue until all places are filled above the two-digit groups in the original number:

```
    2 2 . 9  8
  √5'2 8.3 6'7 0
    4
42  | 1 2 8
      8 4
449 | 4 4 3 6
      4 0 4 1
4588| 3 9 5 7 0
      3 6 7 0 4
        2 8 6 6
```

Notice that the decimal point in the root is placed directly above the decimal point in the original number. If desired, the solution can be carried out to additional decimal places by adding groups of two zeros to the right of the existing decimal, then proceeding just as before.

Should any of the divisors be too large to go into the remainder, a zero is placed in the answer at that point, just as in the similar case in ordinary division. This is illustrated in the following example:

```
      5
  √2 5'7 0.4 9
    2 5
10  |   7 0
```

On covering up the 0 in 70 it is found that 10 is too large to be contained in 7, hence a zero is put in the second place in the answer and the process continued as before:

```
      5  0
  √2 5'7 0.4 9
    2 5
100 |   7 0 4 9
```

an additional group (49) being brought down for the next division.

●1-3 FORMULAS

Literal notation — The laws of electricity and electrical circuits are most conveniently expressed by formulas in which the various quantities are denoted by arbitrary symbols, generally letters of the alphabet. This is called *literal notation*, and is simply a compact method of writing. To cite an example, Ohm's Law for direct currents, a fundamental law in electrical work, can be expressed as follows: "The current flowing in a circuit is equal to the impressed electromotive force divided by the resistance, using consistent units." In algebraic formula this is written

$$I = \frac{E}{R}$$

in which the letter I stands for current, E for electromotive force and R for resistance. Expressing the law in this way has the additional advantage that anyone familiar with the elementary processes of algebra can see at a glance the further relations:

$$E = RI$$

$$\text{and } R = \frac{E}{I}$$

either of which would have to be learned by rote or else reasoned out from the ordinary-language statement.

"Solving a formula" merely means substituting the proper numbers for the literal quantities whose values are known, and then performing the indicated arithmetical operations to find the unknown quantity.

Signs of operation — The ordinary signs of arithmetic ($+$, $-$, $\times$, $\div$, $\sqrt{\ }$, etc.) are used. However, division is nearly always indicated by writing the dividend over the divisor in the form of a fraction, as $\frac{E}{R}$ in Ohm's Law given above; this means E divided by R. An alternative method of writing the same thing is E/R, where the diagonal bar has the same meaning as the horizontal fraction bar. Multiplication can be indicated by the usual sign ($\times$) and also by a dot (.), but the sign is frequently left out altogether, the two quantities to be multiplied simply being written side by side. Thus, RI means "R multiplied by I." However, the multiplication sign must be used when ordinary numbers appear together in a formula; for instance, 31 multiplied by 42 is written 31×42; the abbreviated method cannot be used in this case because neither 3142 nor 31.42 is the same thing as 31×42. But when an actual number and a literal number are multiplied together they can be written with-

out the sign: $15E$, for example, means 15 times E.

In an expression such as $15E$ the number 15 is called a *coefficient* of E. A coefficient is a factor, either literal or numerical, whose value is fixed. The symbol k is frequently used in electrical formulas where a coefficient, or "constant," is specified. When numerical coefficients are multiplied the ordinary numerical operation is usually performed; thus if we multiply $4R$ by 3, the product is written $12R$, instead of $4 \times 3R$. The latter or algebraic form is used only when multiplication is by a literal quantity; thus, $I \times 3R$ would be written $3IR$. It is usual to assemble the literal factors together and to put the numerical factor first.

Brackets — Bracketed symbols indicate a group that is to be treated as a unit. Thus if we write $(a + b)$ we mean that the bracketed quantity must be treated as though it were all one symbol. This is necessary to distinguish between expressions of the type $ab - c$ and $a(b - c)$. The former means "Multiply a by b and then subtract c from the product," while the latter means "Subtract c from b and then multiply the difference by a." The two are not the same thing, as substitution of any numbers you may choose for a, b and c will readily show. Similarly, $(a - b)^2$ means "Multiply $(a - b)$ by $(a - b)$" as distinguished from $a - b^2$, which means "Multiply b by itself and subtract the product from a."

Fundamental rules — In general, a series of additions can be performed in any order, since the same sum results no matter which number is put first or last. For example, $11 + 6 + 8$ gives the same total as $6 + 8 + 11$ or $8 + 11 + 6$, etc. The same thing is true of multiplication; $5 \times 3 \times 2 = 2 \times 5 \times 3 = 3 \times 5 \times 2$, etc. However, in subtraction and division this is not the case in ordinary arithmetic: $12 - 9$ is not the same thing as $9 - 12$, nor is $42 \div 7$ the same as $7 \div 42$. By considering subtraction as *algebraic addition* of a negative number this restriction can be overcome, likewise with division by considering every divisor as a fraction of the form $1/n$, where n is the divisor, and then using it as a factor. Thus, $12 - 9 = -9 + 12$, the plus and minus signs being considered to be properties of the numbers. A number having no sign prefixed is always understood to be positive. Similarly, $42 \div 7 = 42 \times \frac{1}{7}$, which is the same as $\frac{1}{7} \times 42$. Therefore, by reducing subtraction to algebraic addition, and reducing division to multiplication by a fraction, we can say that, in general, a sequence of the *same* operations can be performed in any convenient order.

When algebraic addition and multiplication are both indicated, as they frequently are in a single formula, the order of operations is very important. Thus if we have $ab - c$, in which a, b, and c have the values 4, 8 and 3, respectively, substitution would give $4 \times 8 - 3$. This gives 29 as an answer, which is not the same as if we first subtracted the 3 from the 8 and then multiplied the difference by 4. The general rule is that the operations indicated in an equation must be performed in sequence from left to right. It is important to note that in a literal expression factors may not be separated by addition or subtraction. Thus in the expression $a + bc$, when numerical values are substituted the multiplication must be performed first. Using the previous values, $a + bc$ does not mean $4 + 8 \times 3$, which is 36, but $4 + (8 \times 3)$, which is 28. In other words, a series of factors is equivalent to a bracketed quantity, when numerical values are substituted for the letter symbols.

Powers and roots — In some formulas it is necessary to raise a series of factors to a certain power or to take a specified root of a group of factors. If we want to square the expression abc we have $(abc)^2$. This means that the whole expression is multiplied by itself; that is, $abc \times abc$, or, dropping the multiplication sign, $abcabc$. Since the order in which the multiplication is performed does not matter, this can be rewritten $aabbcc$. But $aa = a^2$, $bb = b^2$, and $cc = c^2$; therefore $(abc)^2 = a^2b^2c^2$. In other words, the square of a series of factors is equal to the product of the squares of each factor, and similarly for higher powers. Identical reasoning will show that the same rule applies for roots: $\sqrt{abc} = \sqrt{a} \times \sqrt{b} \times \sqrt{c}$, or $\sqrt{a}\sqrt{b}\sqrt{c}$.

In expressions calling for raising the algebraic sum of two or more numbers to a given power, or for extracting a given root, no such simple rule applies. For instance, $(4 + 3)^2$ is equal to 7^2, which is 49. It is not equal to $4^2 + 3^2$, which is the sum of 16 and 9, or 25. (Expressions of this kind should be multiplied out when they are encountered.) Usually, in ordinary radio formulas the actual values will be substituted and it will be possible to take the sum before finding the power or root. This simplifies the computations.

Bracketed quantities — One other type of operation needs to be learned. When a sum is multiplied by a number, as in the expression $a(b + \frac{1}{c} - d)$ each part (called a *term*) of the sum must be multiplied by the factor. In the example this gives

$$a(b + \frac{1}{c} - d) = ab + \frac{a}{c} - ad$$

Notice that the signs of the terms in the right-

hand part of the equation are the same as in the original bracketed quantity (both a and b are positive, since their signs are not written). This is always the case when the multiplier is positive. If the multiplier has a negative sign, as in the expression $a - b(c + d - e)$, all the terms inside the brackets must be reversed when the multiplication is performed and the brackets eliminated. Thus the resulting expression would be $a - bc - bd + be$.

If the bracketed quantity is to be added to another quantity, as $a + (b - c + d)$, removing the brackets requires no changes in the signs, so

$$a + (b - c + d) = a + b - c + d$$

But if the bracketed quantity is preceded by a minus sign, all signs inside the brackets must be reversed when the brackets are removed:

$$a - (b - c + d) = a - b + c - d$$

These rules are easily illustrated by means of numerical examples. In each case the distributed form is shown first, followed by the method of first *collecting terms* (performing the additions) inside the brackets.

Multiplication by a positive number:

$$5 \times (8 - 4 + 1) = 40 - 20 + 5 = 25$$
$$5 \times (5) = 25$$

Multiplication by a negative number:

$$12 - 3(4 - 2) = 12 - 12 + 6 = 6$$
$$12 - 3(2) = 12 - 6 = 6$$

Addition of a bracketed quantity:

$$10 + (8 + 2 - 3) = 10 + 8 + 2 - 3 = 17$$
$$10 + 7 = 17$$

Subtraction of a bracketed quantity:

$$10 - (8 + 2 - 3) = 10 - 8 - 2 + 3 = 3$$
$$10 - 7 = 3$$

Transposition — In general, a satisfactory solution of a formula cannot be obtained unless the numerical values of all but one of the literal quantities are known. Taking Ohm's Law again as an example, the relation

$$I = \frac{E}{R}$$

will give the current, I, when known values of electromotive force (voltage) and resistance can be substituted for E and R in the formula. Using ordinary units, if E should be 80 volts and R 40 ohms, the current would be $80 \div 40$, or 2 amperes. However, should only the voltage be known, we could learn nothing about the current that might flow, since every value of resistance would give a different value of current. Therefore, it must be possible to substitute known values for every quantity in the formula or equation except one, which is called the "unknown." Usually, *any* one of the literal quantities may be the unknown, regardless of its position in the equation, and the equation can be *transposed* so that the unknown quantity appears on one side (usually to the left of the equality sign) and all the known quantities on the other side. It is convenient to arrange an equation in this way before actually substituting the numerical values for the known quantities.

Transposition is based on the principle that performing the same operation on each side of an equation does not alter the equality. Thus if we add the same number to both sides of an equation the resulting sums also are equal. A numerical example will quickly illustrate the point. Suppose we have the following equation:

$$5 \times 3 = 11 + 4$$

and add 6 to each side:

$$(5 \times 3) + 6 = 11 + 4 + 6$$
$$15 + 6 = 21$$

The same quantity subtracted from each side of an equation also leaves both remainders equal; similarly, multiplying both sides by the same number or dividing both sides by the same number does not change the fact of equality. Likewise, raising both sides to the same power or taking the same root on both sides leaves the resulting quantities equal. Whatever the operation, it must be performed similarly on both sides and on the *whole* of each side. For example, if we wish to square both sides of the equation $a = b + c$, the right-hand side must be considered as a unit so that the resulting equation is $a^2 = (b + c)^2$. As we have seen before, this is not the same thing as $b + c^2$ or even $b^2 + c^2$. Careless errors can be avoided by keeping this point in mind.

An equation such as $I = E/R$ can be transposed very readily. Suppose we know the current and resistance and want to find E. The object is to get E alone on one side of the equation and get all the other quantities on the other side. To get E by itself we must eliminate R on that side of the equation. Since E is divided by R, we can eliminate the R in the denominator if we multiply E/R by R, since a number divided by itself is 1. The common expression is that the R's "cancel out." However, to make the resulting expression an equality we must multiply *both* sides of the equation by R, so we have

$$I \times R = \frac{E}{R} \times R$$

The R's on the right-hand side cancel, leaving us with $IR = E$, or, since we usually put the quantity we want to find on the left-hand side,

$E = IR$. From this we can readily find R in terms of the other two quantities by dividing both sides of the equation by I or (which is the same thing) by multiplying both sides by $1/I$:

$$E \times \frac{1}{I} = IR \times \frac{1}{I}$$

In this case the I's on the right-hand side cancel, leaving us

$$R = \frac{E}{I}$$

In more complicated equations it may be necessary to perform several operations of this nature in sequence before the desired arrangement of the equation is secured. For example, the equation for the resonant frequency of a tuned circuit is

$$f = \frac{1}{2\pi\sqrt{LC}}$$

where f is the frequency, L is the inductance and C the capacity in the circuit. (π, the ratio of the circumference to the diameter of a circle, has its usual value of 3.1416 . . .) Suppose we want to find C in terms of the other quantities. First we need to bring C out of the denominator, since we do not want the result in fractional form. To do this, multiply both sides by $2\pi\sqrt{LC}$. This gives

$$f \times 2\pi\sqrt{LC} = \frac{1}{2\pi\sqrt{LC}} \times 2\pi\sqrt{LC}$$

$$\text{or } 2\pi f\sqrt{LC} = 1$$

The order in which the factors 2, π and f are written together does not really matter, but that shown is customary. We next need to get rid of the radical, which we can do by squaring both sides of the equation:

$$(2\pi f\sqrt{LC})^2 = 1^2$$
$$\text{or } (2\pi f)^2 LC = 1$$

since 1^2 is still 1. The last step is to get C alone on one side of the equation, which is accomplished by dividing both sides by all the remaining factors in the left-hand side of the above expression:

$$\frac{(2\pi f)^2 LC}{(2\pi f)^2 L} = \frac{1}{(2\pi f)^2 L}$$

when the $(2\pi f)^2 L$'s on the left cancel out, leaving

$$C = \frac{1}{(2\pi f)^2 L}$$

It is useful to know one more thing in connection with transposition. If we have an equation of the form

$$\frac{a}{b} = \frac{c}{d}$$

and multiply both sides by $\frac{bd}{ac}$, we have

$$\frac{abd}{abc} = \frac{bdc}{adc}$$

The a's and b's on the left cancel out, as do the d's and c's on the right, leaving

$$\frac{d}{c} = \frac{b}{a} \text{ or } \frac{b}{a} = \frac{d}{c}$$

That is, if both sides of an equation are inverted, the resulting expression is an equality. Any equation can be considered to be in this form, even though neither side is expressly fractional, because any whole number can be considered to be a fraction of the form $n/1$ (dividing a number by 1 does not change its value). Such a number inverted becomes $1/n$.

Inversion can be used to advantage in transposing a formula such as the following

$$R = \frac{1}{\frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3} \cdots\cdots}$$

which is the rule for finding the net resistance of resistances connected in parallel. The series of dots indicates that similar terms ($\frac{1}{R_n}$ where n indicates the last number of a consecutive series) may be added until there is one term for each resistor in the actual group considered in the problem. Suppose we have two resistors in parallel and want to add a third to make the total resistance, R, have a specified value. We need then to transpose the equation to give R_3, the third resistor, in terms of the first two, R_1 and R_2, and the total resistance, R. By the rule for inversion,

$$\frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3} = \frac{1}{R}$$

Subtracting $\frac{1}{R_1} + \frac{1}{R_2}$ from both sides we have remaining

$$\frac{1}{R_3} = \frac{1}{R} - \left(\frac{1}{R_1} + \frac{1}{R_2}\right) = \frac{1}{R} - \frac{1}{R_1} - \frac{1}{R_2}$$

Inverting again,

$$R_3 = \frac{1}{\frac{1}{R} - \frac{1}{R_1} - \frac{1}{R_2}}$$

Observe that the whole expression on the right hand side must be inverted, not simply the individual fractions.

Hints for solving formulas — In practical use of formulas the quantity which must be found may or may not appear alone on one side of the equation as given. If it does not, the equation must be transposed to bring the unknown on one side and all the known quantites on the other.

When the formula is in suitable form, substitute the known values for the appropriate

letters, inserting the proper signs of operation. Where two or more factors appear together, write the multiplication sign between them when the figures are substituted. Enclose factors in brackets when their product is to be added to or subtracted from a preceding quantity, and perform the necessary numerical operations to reduce bracketed quantities to a single number *first*.

When the equation has several terms, perform the operations necessary to make each term a single number before adding or subtracting.

These points are illustrated by the following example, using a formula for finding the output voltage of a power supply:

$$E_o = 0.9E_t - \frac{(I_b + I_L)(R_1 + R_2)}{1000} - E_r$$

Since the meanings of the symbols have no particular bearing on the present discussion, we may simply assign the following values for the known quantities: $E_t = 750$; $I_b = 25$; $I_L = 100$; $R_1 = 75$; $R_2 = 125$; $E_r = 15$. Substituting, we have:

$$E_o = (0.9 \times 750) - \frac{(25 + 100)(75 + 125)}{1000} - 15$$

Collecting terms:

$$E_o = 675 - \frac{125 \times 200}{1000} - 15$$
$$= 675 - 25 - 15$$
$$E_o = 635$$

When fractions appear in the denominator of a formula it is usually best to convert them to decimals. For example, the formula for three resistances in parallel previously mentioned:

$$R = \frac{1}{\frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3}}$$

is more easily solved by decimals than by the method of finding a common multiple for the fractions in the denominator and then adding. Thus, suppose that the values given for R_1, R_2 and R_3 are 500, 250 and 100, respectively. Substituting gives

$$R = \frac{1}{\frac{1}{500} + \frac{1}{250} + \frac{1}{100}}$$

The corresponding decimals are 0.002, 0.004, and 0.01. Substituting again:

$$R = \frac{1}{0.002 + 0.004 + 0.01} = \frac{1}{0.016}$$

Performing the division,

$$R = 62.5$$

• 1-4 LAWS OF EXPONENTS

Multiplication and division — As we have seen (§ 1-2), the power to which a number is raised is the number of times it is used as a factor. The number a alone means the first power of a, or a^1, but the exponent 1 is seldom written since a number having no exponent is always understood to be its first power.

Now if we multiply a^2 by a, we really have aaa, since $a^2 = aa$. Also if we multiply a^2 by a^2 we have $aaaa$, or a^4. Similarly, $a^3 \times a^2$ (ordinarily written a^3a^2) is equal to $aaaaa$, or a^5. In other words, when two powers of the same number are multiplied together, the product is the number raised to a power which is equal to the sum of the exponents. Thus, $aa^3 = a^4$; $a^3a^4 = a^7$; $b^4b = b^5$, and so on. It is necessary to remember that a number with no exponent written always has the exponent 1.

Suppose now we have a^3/a. This is the same as saying

$$\frac{aaa}{a}$$

in which one a in the numerator cancels one in the denominator, leaving aa, or a^2. Or, $b^3/b^2 = bbb/bb$; the two b's in the denominator cancel two in the numerator, leaving b as the answer. When a power of a number is divided by another power of the same number, the quotient is the number raised to a power equal to the exponent of the dividend minus the exponent of the divisor. Thus, $a^4/a = a^3$; $b^5/b^2 = b^3$; $b^6/b^3 = b^3$, etc. This fact opens the possibility of writing a divisor in a different way. Thus, $a^4/a = a^3$ can be written as a multiplication in which the exponent of the divisor is negative: $a^4a^{-1} = a^3$. Addition of the negative exponent is the same as subtraction of a positive exponent. Other examples are: $b^5b^{-2} = b^3$; $a^2a^{-1} = a$, and so on.

Consequently, a negative exponent indicates division by the same number with a positive exponent. In other words, $a^{-3} = 1/a^3$; $b^{-5} = 1/b^5$; $a^{-1} = 1/a$, etc. This form is frequently used in formulas where it is desired to avoid writing the equation as a fraction.

When a number (other than 0) is divided by itself the quotient is 1; that is, $a/a = 1$. Since $a = a^1$ and $1/a = a^{-1}$, a/a can be written $a^1a^{-1} = 1$, or simply $aa^{-1} = 1$. By the law of exponents this also can be written $aa^{-1} = a^0$. Similarly, $a^3a^{-3} = a^0$; $b^2b^{-2} = b^0$, etc. Dividing a number by itself is equivalent to raising the number to the 0th power; therefore the exponent 0 means that the number reduces to 1.

Significant figures; order of magnitude — When numerical values are very small or very large it is common practice to make two factors out of the number, one of which is the "significant" part and the other a power of ten, the "order of magnitude." For example,

the number 568,000,000 would be separated into the two factors 568 and 1,000,000 and written 568×10^6; 293,000 would be separated (factored) into 293 and 1000 and written 293×10^3; 0.0016 would be factored to 16 and 1/10,000 (0.0001) and written 16×10^{-4}; 0.000004 would be factored to 4 and 1/1,000,000 (0.000001) and written 4×10^{-6}, and so on. The following table shows the powers of 10 commonly used as factors in writing numbers in such form:

10^6	=	1,000,000
10^5	=	100,000
10^4	=	10,000
10^3	=	1,000
10^2	=	100
10^{-1}	=	0.1
10^{-2}	=	0.01
10^{-3}	=	0.001
10^{-4}	=	0.0001
10^{-5}	=	0.00001
10^{-6}	=	0.000001

The factor 10, when used, is of course simply written as 10.

In practical work it is sufficient in nearly all cases to obtain results only to three significant figures. Subsequent figures can be dropped or rounded off, retaining only the appropriate power of ten to give the order of magnitude.

Powers and roots — To square a number such as a^3, we use it twice as a factor, thus: $a^3a^3 = a^6$. The operation usually would be indicated by writing it in the form $(a^3)^2$, indicating that the quantity inside the brackets is to be squared. Similarly, $(a^2)^3 = a^2\,a^2\,a^2 = a^6$; $(b^4)^2 = b^4\,b^4 = b^8$, and so on. The exponent in the result is equal to the product of the exponents in an expression such as $(a^2)^4 = a^8$. It follows that in taking a *root* of such an expression as $\sqrt[3]{a^9}$ we should divide the exponent of the number under the radical sign by the index of the root to find the exponent of the result; $\sqrt[3]{a^9} = a^3$; $\sqrt{a^4} = a^2$; $\sqrt[4]{a^4} = a$; $\sqrt[3]{a^6} = a^2$, etc. Since division of exponents is indicated in such cases, we can write an expression like $\sqrt[3]{a^9}$ as $a^{\frac{9}{3}}$, which equals a^3; $\sqrt{a^4}$ can be written $a^{\frac{4}{2}}$, equalling a^2; $\sqrt[4]{a^4} = a^{\frac{4}{4}} = a$, $\sqrt[3]{a^6} = a^{\frac{6}{3}} = a^2$, and so on. An expression such as $a^{\frac{4}{3}}$ also can be written $(a^4)^{\frac{1}{3}}$, since multiplication of the exponents gives $a^{\frac{4}{3}}$, and similarly for the other examples. In other words, a fractional exponent of the form $1/n$ means the same thing as $\sqrt[n]{\ }$.

● 1-5 LOGARITHMS

Logarithms — The laws of exponents (§ 1-4) show that, when we deal with one basic number, the processes of multiplication and division reduce to addition and subtraction of exponents, and the processes of raising to a power and taking roots reduce to multiplication and division of exponents. Advantage can be taken of this fact to simplify numerical operations, provided all ordinary numbers can be expressed as exponents of some number chosen as the *base*. Such exponents are called *logarithms*, and in the common system of logarithms the base chosen is the number 10. The logarithm of a number is the exponent of the power to which the base 10 must be raised to produce the given number.

Since $10^1 = 10$, the logarithm of 10 is 1; since $10^3 = 1000$, the logarithm of 1000 is 3, and so on. Likewise, the logarithm of 1 is 0 (written $\log 1 = 0$), since $10^0 = 1$; $\log 0.1 = -1$, since $10^{-1} = 0.1$, etc. Some of the logs of the powers of 10 are as follows:

log	10,000	= 4
log	1000	= 3
log	100	= 2
log	10	= 1
log	1	= 0
log	0.1	= −1
log	0.01	= −2
log	0.001	= −3
log	0.0001	= −4

The logarithms of all numbers between 1 and 10 must have values lying between 0 and 1, since $\log 1 = 0$ and $\log 10 = 1$. They are, for convenience in calculation, usually written as decimals. For example, $10^{\frac{1}{2}}$ (or its equivalent $10^{0.5}$) represents $\sqrt{10}$, the value of which, to two decimal places, is 3.16, therefore $\log 3.16 = 0.5$. Similarly, $\sqrt[3]{10} = 10^{0.333}$, the value of the number being 2.15, therefore $\log 2.15 = 0.333$; $\sqrt[6]{10^2} = 10^{0.666}$; the number is 4.64, therefore $\log 4.64 = 0.666$. The logarithms of the integers from 1 to 10 are given, to four decimal places, in the following table:

log	1	= 0
log	2	= 0.3010
log	3	= 0.4771
log	4	= 0.6021
log	5	= 0.6990
log	6	= 0.7782
log	7	= 0.8451
log	8	= 0.9031
log	9	= 0.9542
log	10	= 1.0000

Despite the fact that the table above is extremely limited in scope, it will serve to illustrate the use of logarithms.

Fundamental relations — Remembering the laws of exponents, and that logarithms are actually exponents of powers of the base, we

can express the rules for their use by the equations:

$$\log (a \times b) = \log a + \log b$$

$$\log \frac{a}{b} = \log a - \log b$$

$$\log a^n = n \times \log a$$

$$\log a^{\frac{1}{n}} = \frac{\log a}{n}$$

Suppose we multiply 2 by 4. Using the first rule and substituting 2 for a and 4 for b,

$$\begin{aligned}\log (2 \times 4) &= \log 2 + \log 4 \\ &= 0.3010 + 0.6021 \\ &= 0.9031\end{aligned}$$

This is the logarithm of the product, and we find from the table that 0.9031 is the log of 8. Therefore $2 \times 4 = 8$.

Divide 10 by 5, using the second rule: Substituting 10 for a and 5 for b,

$$\begin{aligned}\log (10 \div 5) &= \log 10 - \log 5 \\ &= 1.0000 - 0.6990 \\ &= 0.3010\end{aligned}$$

This is the logarithm of 2, therefore the quotient is 2.

Raise 2 to the third power. Using the third rule and substituting 2 for a and 3 for n,

$$\begin{aligned}\log 2^3 &= 3 \times \log 2 \\ &= 3 \times 0.3010 \\ &= 0.9030\end{aligned}$$

Within the limit of accuracy, this is the log of 8, therefore the cube of 2 is 8.

Take the square root of 9. Using the fourth rule and substituting 9 for a and 2 for n,

$$\begin{aligned}\log 9^{\frac{1}{2}} &= \frac{\log 9}{2} \\ &= 0.9542 \div 2 \\ &= 0.4771\end{aligned}$$

This is the logarithm of 3, therefore the square root of 9 is 3.

The characteristic — Since we simply add the exponents when multiplying two powers of the same base, it is unnecessary to determine the logarithms of any numbers other than those between 1 and 10. Every number between 10 and 100 is 10 times some number between 1 and 10, every number between 100 and 1000 is 10^2 times some number between 1 and 10, and so on. Similarly, every number between 1 and 0.1 is 1/10 of (or 10^{-1} times) some number between 1 and 10; every number between 0.1 and 0.01 is 10^{-2} times some number between 1 and 10, etc. For example, the logarithm of 600, which factors to 6×10^2, is equal to the sum of the logarithms of 6 and 10^2, or 0.7782 + 2, which equals 2.7782. Similarly, the logarithm of 6000 is 3.7782, log 60 = 1.7782, log 0.6 = 0.7782 − 1, log 0.06 = 0.7782 − 2 (written $\bar{2}$.7782) and so on. The decimal part of the logarithm is called the *mantissa* and the integral part is called the *characteristic*. When the characteristic is negative it is customary to add 10 to it and then indicate that 10 is to be subtracted from the whole logarithm. Thus $\bar{2}$.7782 would be written 8.7782 − 10, $\bar{1}$.7782 would be written 9.7782 − 10, etc. This avoids having a negative characteristic with a positive mantissa. It would be possible, of course, to subtract directly; thus the algebraic sum of 0.7782 and − 1 is − 0.2218, but this form would be less useful, because a negative exponent indicates that the same number with a positive exponent is to be divided into 1. Thus $10^{-0.2218}$ means $\frac{1}{10^{0.2218}}$ or $\frac{1}{1.666}$. Performing the division gives 0.6 as the required number, but the same result is more conveniently obtained by keeping the mantissa positive and using the negative characteristic simply to indicate the number of decimal places in the result. Thus 0.7782 − 1 is the logarithm of 6 × 0.1, or 0.6. The number of zeros after the decimal point and before the first digit is one less than the characteristic when the characteristic is negative. To place the decimal point when the characteristic is positive, point off from the left one more place than the number in the characteristic.

Logarithms are assembled in tables for ready reference. Only the mantissas are given; in any particular problem the characteristic must be supplied by taking the appropriate power of 10 as described above. A four-place table is given in the Appendix; four-place tables give results which are accurate to the third figure and approximately so to the fourth figure. This usually is high enough accuracy in radio calculations. Fig. 101 shows the relationship between the numbers from 1 to 10 and their logarithms, and the drawing can be used as a three-place table by reading the logarithm opposite the given number. The scale of numbers is called a *logarithmic scale*, and is very frequently used in graphs, as discussed in later sections.

● 1.6 FUNCTIONS

Variables — One quantity is said to be a *function* of another quantity (called the *independent variable*) when the value given the latter determines the value of the former. Thus the area of a square is a function of the length of one side, since assigning a value to the length of the side immediately determines the area. Similarly, the current flowing through an electrical circuit of given characteristics is a function of the applied electomotive force or voltage. That is, if

$$I = \frac{E}{R}$$

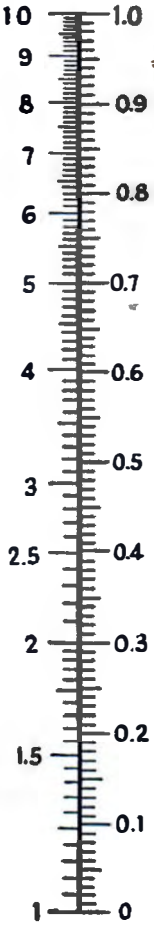


Fig. 101 — The right hand scale gives the logarithms of the numbers shown on the left hand scale.

I is a function of E, when R is constant. We can say the same thing in other ways: *I varies with E*, or *I is proportional to E*. In this case I is *directly* proportional to E, meaning that if E is changed by a certain percentage, I will change by the same percentage and in the same direction (that is, if E is made greater, I will also be greater).

If in the same formula we hold E constant and vary R, we say that I is a function of R. In this case I is said to be *inversely* proportional to R, since if R is made larger I becomes smaller, and vice versa.

Power functions — Other functions may have different modes of proportionality. For example, the area of a circle is

$$A = \pi r^2$$

If π is a constant, the area, A, will be proportional to the square of the radius, r. Speaking generally, in a formula of this kind the value of the function, or dependent variable, is proportional to a *power* (in the example, the second power) of the independent variable. Such a function is called a *power function*. Direct and inverse proportions are really special cases of power functions where the exponents are 1 and -1, respectively. The characteristic feature of a power function is that when the independent variable is increased in a constant ratio the dependent variable also increases in a constant (although not necessarily the same) ratio. For example, when A is proportional to r^2, A will become 4 times as great each time the radius is doubled.

Exponential and logarithmic functions — Another type of function is one in which an increase in equal-value steps in one variable is accompanied by equal-ratio steps in the other variable. This is called an *exponential function*. Thus if we have a series of numbers, a, a^2, a^3, a^4, a^5, the exponent increases by 1 in each step, but the number itself increases by the factor a each time. The difference between this type of function and the power function can be illustrated by the following table:

	Dependent Variable	
Independent Variable	*Power Function*	*Exponential Function*
1	1	1
2	4	2
3	9	8
4	16	16
5	25	32
6	36	64
7	49	128
8	64	256
9	81	512
10	100	1024

using 2 as the factor for the dependent variable in each case. That is, the power function is proportional to the square (exponent = 2) of the independent variable, and the exponential function is doubled (powers of 2) for each equal-value step in the independent variable. If we call the independent variable x and the dependent variable y, the formulas for the two cases are as follows:

Power function: $y = x^2$
Exponential function: $y = 2^x$

From laws of exponents (§ 1-4) and the discussion on logarithms (§ 1-5) it is evident that the exponential formula can also be written in logarithmic form:

$$x = \log_2 y$$

where the subscript 2 indicates that the logarithmic base is the number 2. In ordinary formulas the base 10 would be used; converting

from one base to another is a simple process, but need not concern us here.

An important case of a logarithmic function in radio work is the loudness of signals. To the ear, equal steps in loudness are caused by equal-ratio steps in sound power. A just-detectable increase (or decrease) in loudness is called a "decibel," and the relationship between steps of loudness and power ratios is given by the formula:

$$\text{Decibels} = 10 \times \log \frac{P_1}{P_2}$$

where P_1 and P_2 are the first and second power levels, respectively. The logarithmic base is 10, which is always the case when no other base is definitely specified.

• 1-7 GRAPHS

Coördinates — A graph is a pictorial means of expressing the relationship between a function and independent variable. It shows at a glance the value of the function for a continuous series of values of the independent variable over any range of values of the latter that may be desired. The values are shown on *coördinate systems*, the most common of which is the system of rectangular coördinates, so called because it consists of equally spaced series of lines at right angles to each other. A graph of the relationship between E and I in the formula $I = E/R$, when R is constant, is shown in Fig. 102. The horizontal lines are called *abscissas* and the vertical lines *ordinates*. The horizontal base line is called the "X axis" and the vertical base line (at the left) is called the "Y axis." These names are used because it is customary to use the symbol y for the function and x for the independent variable, and to plot the values of the function on the ordinates and the values of the independent variable on the abscissas.

Plotting — Graphs are plotted by assuming suitable values of the independent variable at relatively small intervals, then solving the equation for the value of the function for each value of the independent variable. Each pair of values represents a point on the graph, the position of the point being determined by the spot where the ordinate and abscissa representing those particular values cross. Thus in Fig. 102 the X axis is marked off in terms of voltage and the Y axis in current. With R constant at 100 ohms, we can make a table of the values of I as follows, using 10-volt steps for E:

When $E = 0$ volts, $I = 0$ amperes
$E = 10$ " $I = 0.1$ "
$E = 20$ " $I = 0.2$ "
$E = 30$ " $I = 0.3$ "
$E = 40$ " $I = 0.4$ "
$E = 50$ " $I = 0.5$ "

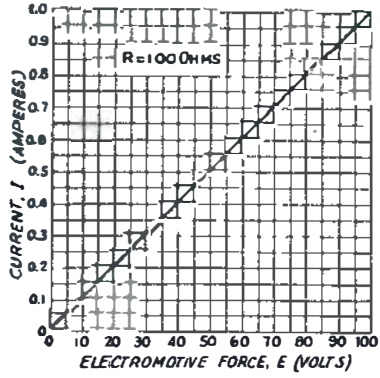


Fig. 102

and so on, all values being found from the formula $I = E/R$. The first point obviously lies on the intersection of the axis at 0 (this point is called the *origin*). The next point, $E = 10$, $I = 0.1$, is plotted by moving out along the X axis to the value 10, then moving up on the ordinate at that point to the value 0.1. The second point is plotted by locating $E = 20$ and then moving up on that ordinate to the value 0.2. The remaining points are plotted similarly. When the series of points has been plotted, a smooth line is drawn joining all of them together; this line is called a *curve*, even though it may be perfectly straight as in the illustration.

It can be seen from Fig. 102 that the graph of I as a function of E is a straight line. A function which is directly proportional to the independent variable always gives a straight-line graph on rectangular coördinates.

Scales — Scales of the type shown in Fig. 102, in which equal segments on a given scale have the same value no matter where they are taken, are called *linear*. Thus the length on the X axis between 0 and 10 (10 volts difference) is the same as between 40 and 50 (again 10 volts difference). The same length of line always represents 10 volts. This type of scale is useful for many kinds of graphs, but in some cases the logarithmic scale is better because parts of some curves are very steep when plotted on a linear scale. That is, a small change in the value of the independent variable causes a large change in the function, which makes it difficult to read the graph accurately. In general, it is advisable to choose scales so that the plot of the function will be a straight line, or nearly straight line, making an angle of about 45 degrees with either axis. This gives maximum readability.

Logarithmic scales for both abscissae and ordinates will give a straight-line curve for any power function. Thus in the formula $P = I^2R$,

which gives the relationship between power, current and resistance, P is a function of I^2 when R is constant. Using 10 ohms for R and selecting values for I, we tabulate as follows:

When $I = 1$, $P = 10$
$I = 2$, $P = 40$
$I = 3$, $P = 90$
$I = 4$, $P = 160$
$I = 5$, $P = 250$, etc.

Plotting these values on logarithmic scales (graph paper of this type is called "logarithmic" paper, or sometimes "log-log" paper) gives a curve of the type shown in Fig. 103. Note that a zero point cannot be plotted, since a logarithmic scale never reaches zero. This type of scale also is useful for plotting functions which are inversely proportional to the independent variable, such as in the formula $I = E/R$, when E is held constant and R is varied. In this case I is a function of $1/R$.

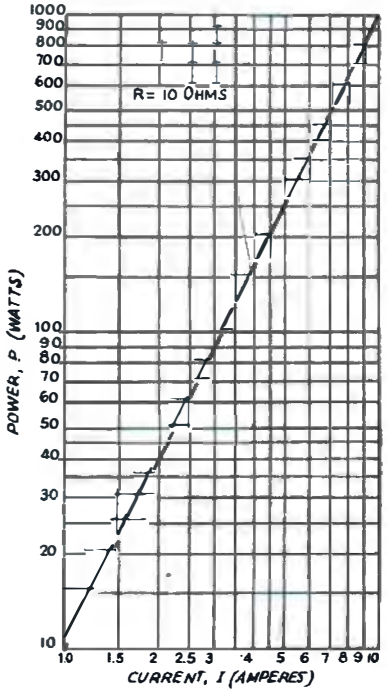


Fig. 103

Exponential functions are best plotted on semi-logarithmic paper, or paper having one scale linear and the coördinate scale logarithmic. The graph of such a function will be a straight line when the exponential variable is plotted to a linear scale and the linear variable to a logarithmic scale. When the formula is in logarithmic form the logs are plotted on the linear scale. Fig. 104 illustrates the graph of the formula for the relationship between decibels and power ratios, the latter being the independent variable and the former the function. It is seen that the curve is a straight line.

In plotting graphs it is desirable to mark off the scales in units which will make the whole graph approximately square, when the "end" values are selected. Thus, suppose we are interested in I as a function of E, with R constant at 100 ohms, for all values of E between 0 and 100 volts. The end values of E are 0 and 100, and the end values of I, found from the formula, are 0 and 1 ampere. Using ordinary rectangular paper having 20 divisions to the inch, with every tenth line heavy, it is convenient to make each small division represent 1 volt on the X axis, and each small division represent 0.01 ampere on the Y axis. The graph will then be square, as shown in Fig. 102. (To avoid difficulty in reproduction, the "units" lines are omitted in the figure, only the "fives" lines being shown.)

If a large range of values has to be shown, logarithmic scales are to be preferred to linear scales. If we extended the graph of Fig. 102 to include all values of I when E varies from 1 to 1000 volts, it is evident that to maintain the same accuracy of reading, particularly in the range shown in Fig. 102, it would be necessary to make the axes 10 times as long. If we kept the graph the same size and simply reduced the scales to accommodate the 10-times larger range of values, considerable percentage error would arise in reading the curve for values below 100 volts. This difficulty can be over-

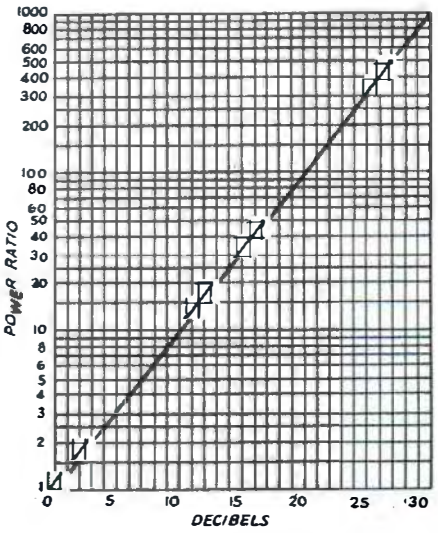


Fig. 104

come by using logarithmic paper, since with logarithmic scales the same percentage accuracy can be obtained no matter what the order of magnitude of the quantities. In the specific problem considered, we could use "three cycle" paper (one cycle is the scale from 1 to 10, or 10 to 100, or 100 to 1000, etc.) for all values of E between 1 volt and 1000 volts, as shown in Fig. 105. The range could be extended indefinitely in either direction by using more

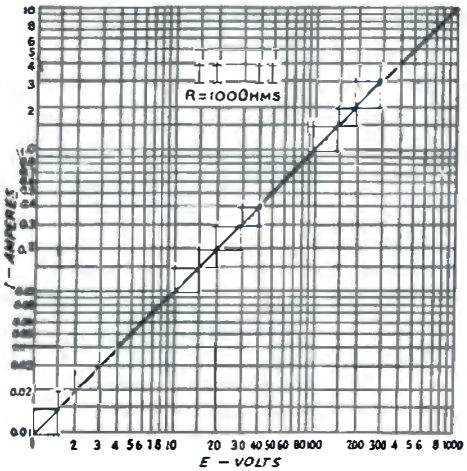


Fig. 105

cycles or changing both scales in the same ratio of some power of 10. As stated before, we cannot reach zero on this type of scale, but in many cases this is no particular handicap.

Interpolation — On graph paper the coördinates must be spaced at definite intervals, and if we are interested in values lying between two coördinate lines it is necessary to *interpolate*. Interpolation is simply estimating the value from the position of the point between two lines. If the scale is linear, half the distance between two lines will add to the value of the lower line half the difference between the assigned values of the two lines. That is, if a point lies half-way between two ordinates marked "4" and "5", the difference between the two values is 1, hence the half-way position indicates 0.5. Adding this to the value of the lower ordinate gives 4.5 as the value at that point. It is usually possible to estimate to one-tenth of a unit when the coördinate lines are spaced ⅛ inch or more apart. If they are closer together it may be possible to estimate accurately only to ¼ unit.

Interpolation on logarithmic scales is a little more difficult, but can be learned by studying how the values vary compared to a linear scale, using a scale such as Fig. 101 for comparison. When in doubt, estimate the position of the point on a linear basis and then measure off a corresponding fractional length on an enlarged logarithmic scale. This procedure will permit fairly accurate interpolation.

Polar coördinates — When it is necessary to show graphically the variation in a quantity whose value is a function of its direction with respect to a fixed reference line, the function can be plotted on polar coördinate paper. Fig. 106 shows such a graph. Polar coördinates consist of a set of radial lines representing units of angle, and a series of concentric circles, both coördinates originating from a common point (the origin).

The polar coördinate graph is especially useful for showing the directional characteristic of an antenna system. The plot in Fig. 106 is the theoretical directional characteristic or "pattern" for a straight, horizontal antenna when its length is considerably less than a half wavelength. If the antenna wire lies in the direction indicated on the graph, the relative field strength, or relative intensity of radiation, will vary with the direction with respect to the wire as shown, assuming that the field strength is observed at the same distance in any direction. Note that the scale in such a case is purely relative, and the maximum strength, which occurs in the direction at right angles to the length of the wire, is arbitrarily assigned a value of 1. If we measure the field strength directly off the end of the wire it will be zero;

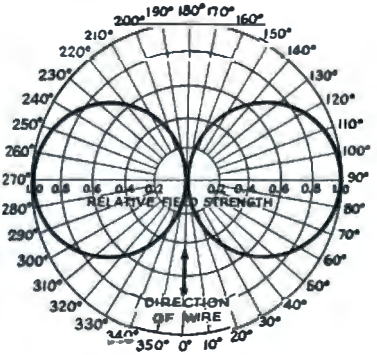


Fig. 106

at 45 degrees it will be 0.71, or 71%, of its value at right angles to the wire, and so on.

If the field strength is uniform in all directions with respect to the antenna, the graph of the function will simply be a circle concentric with the origin.

CHAPTER TWO

Electrical and Radio Fundamentals

● 2-1 FUNDAMENTALS OF A RADIO SYSTEM

THE BASIS of radio communication is the transmission of electromagnetic waves through space. The production of suitable waves constitutes radio transmission, and their detection, or conversion at a distant point into the intelligence put into them at the originating point, is radio reception. There are several distinct processes involved in the complete chain. At the transmitting point, it is necessary first to generate power in such form that when it is applied to an appropriate radiator, called the antenna, it will be sent off into space in electromagnetic waves. The message to be conveyed must be superimposed on that power by suitable means, a process called modulation.

As the waves spread outward from the transmitter they rapidly become weaker, so at the receiving point an antenna is again used to abstract as much energy as possible from them as they pass. The wave energy is transformed into an electric current which is then amplified, or increased in amplitude, to a suitable value. Then the modulation is changed back into the form it originally had at the transmitter. Thus the message becomes intelligible.

Since all these processes are performed by electrical means, a knowledge of the basic principles of electricity is necessary to understand them. These essential principles are the subject of the present chapter.

● 2-2 THE NATURE OF ELECTRICITY

Electrons — All matter — solids, liquids and gases — is made up of fundamental units called *molecules.* The molecule, the smallest subdivision of a substance retaining all its characteristic properties, is constructed of *atoms* of the elements comprising the substance.

Atoms in turn are made up of particles, or charges, of electricity, and atoms differ from each other chiefly in the number and arrangement of these charges. The atom has a nucleus containing both positive and negative charges, with the positive predominating so that the nature of the nucleus is positive. The charges in the nucleus are closely bound together. Exterior to the nucleus are negative charges — electrons — some of which are not so closely bound and can be made to leave the vicinity of the nucleus without too much urging. These electrons whirl around the nucleus like the planets around the sun, and their orbits are not random paths but geometrically-regular ones determined by the charges on the nucleus and the number of electrons. Ordinarily the atom is electrically neutral, the outer negative electrons balancing the positive nucleus, but when something disturbs this balance electrical activity becomes evident, and it is the study of what happens in this unbalanced condition that makes up electrical theory.

Insulators and Conductors — Materials which will readily give up an electron are called conductors, while those in which all the electrons are firmly bound in the atom are called insulators. Most metals are good conductors, as are also acid or salt solutions. Among the insulators are such substances as wood, hard rubber, bakelite, quartz, glass, porcelain, textiles, and many other non-metallic materials.

Resistance — No substance is a perfect conductor — a "perfect" conductor would be one in which an electron could be detached from the atom without the expenditure of energy — and there is also no such thing as a perfect insulator. The measure of the difficulty in moving an electron by electrical means is called resistance. Good conductors have low resistance, good insulators very high resistance. Between the two are materials which are neither good conductors nor good insulators, but they are nonetheless useful since there is often need for intermediate values of resistance in electrical circuits.

Circuits — A circuit is simply a complete path along which electrons can transmit their charges. There will normally be a source of energy (a battery, for instance) and a *load* or portion of the circuit where the current is made to do useful work. There must be an unbroken path through which the electrons can transmit their charges, with the source of energy acting as an electron pump and sending them around the circuit. The circuit is said to be *open* when no charges can move, due to a break in the path. It is *closed* when no break exists — when switches are closed and all connections are properly made.

● 2-3 STATIC ELECTRICITY

The electric charge — Many materials that have a high resistance can be made to acquire

a charge (surplus or deficiency of electrons) by mechanical means such as friction. The familiar crackling when a hard-rubber comb is run through hair on a dry winter day is an example of an electric charge generated by friction. Objects can have either a surplus or a deficiency of electrons — it is called a *negative* charge if there is a surplus of electrons; a *positive* charge if there is a lack of them. As with all things in nature, there must always be a balance, and for every negative charge there will be found a similar positive charge, since each electron that leaves an atom to form a negative charge leaves the rest of the atom with a positive charge. The kind of charge is called *polarity*, a negative charge constituting a negative *pole*, a positive charge being a positive pole.

Attraction and repulsion — Unlike charges (one positive, one negative) exert an attraction on each other. This can be demonstrated by

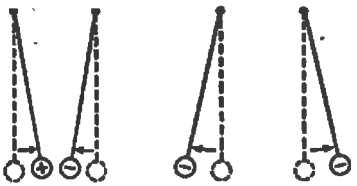


Fig. 201 — Attraction and repulsion of charged objects, as shown by the pith-ball experiment.

giving equal but opposite charges to two very light objects of insulating material (pith balls are used in the classical experiment) and suspending them near each other. They will be drawn toward each other, and if they touch the charges will neutralize, leaving both objects without charge. Charges of the same type, however, repel each other, and a similar experiment with like-charged objects will show them tending to swing apart.

Electrostatic field — From the foregoing it is evident that an electric charge can exert a force through the space surrounding the charged object. The region in which this force is exerted is considered to be pervaded by the *electrostatic field*, this concept of a field being adopted to explain the "action at a distance" of the charge. The field is assumed to consist of *lines of force* originating on the charge and spreading in all directions. The number of lines of force per unit area is a measure of the intensity of the field.

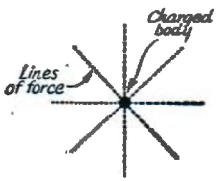


Fig. 202 — Lines of force from a charged object extend outward radially. Although only two dimensions are shown, the field extends in all directions from the charge, and the field should be visualized in three dimensions.

Potential difference — If two objects are charged differently, a *potential difference* is said to exist between them, and this difference is measured by an electrical unit called the *volt*. The greater the potential difference, the higher (numerically) the voltage. This potential difference or voltage exerts an electrical pressure or *force* as explained above, and for this reason it is often called *electromotive force* or, simply, *e.m.f.* It is not necessary to have unlike charges to have a difference of potential; both, for instance, may be negative so long as one charge is more intense than the other. From the viewpoint of the stronger charge, the weaker one appears to be positive in such a case, since it has a smaller number of excess electrons; in other words, its *relative polarity* is positive. The greater the potential difference the more intense is the electrostatic field between the two charged objects.

Capacity — If two metal plates are separated a short distance by a high-resistance material, such as glass, mica, oil or air, it will be found that the two plates can be given a charge by connecting them to a source of potential difference. Such a device is called a *condenser*, and the insulating material between the metal plates is called the *dielectric*. The potential difference, or voltage, of the charge will be equal to that of the source. The *quantity* of the charge will depend upon the voltage of the charging source and the *capacity* of the condenser. The value of capacity of a condenser is a constant depending upon the physical dimensions, increasing with the area of the plates and the thinness and *dielectric constant* of the insulating material in between. The dielectric constant of air is 1, while for other insulating materials it is usually higher. Glass, for instance, has a dielectric constant of about 4; this means, simply, that if glass is substituted for air as the dielectric in an otherwise identical condenser, the capacity of the condenser will be four times as great.

Capacity is measured in *farads*, a unit much too large for practical purposes, and in radio work the terms *microfarad* (abbreviated *μfd.*) and *micro-microfarad* (*μμfd.*) are used. The microfarad is one-millionth of a farad, and the micro-microfarad is one-millionth of a microfarad.

The electrical energy in a charged condenser is considered to be stored in much the same way that mechanical energy is stored in a stretched spring or rubber band. Whereas the mechanical energy in the spring can be stored

because of the elasticity of the material, the electrical energy in a condenser is stored in the electrostatic field between the plates.

Condensers — The construction of a condenser is determined by the use for which it is intended. Where the capacity must be continuously adjustable, as in tuning radio circuits, sets of interleaved metal vanes are used with air as the dielectric. In high capacity units

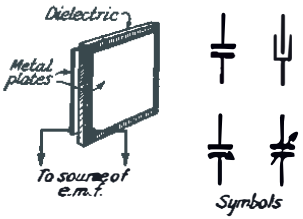


Fig. 203 — A simple type of condenser, consisting of two metal plates with dielectric material between. The diagrammatic symbols for condensers are shown at the right. The two at the top indicate condensers of fixed capacity, the two below, condensers whose capacity is variable. The symbols in the left hand column are more commonly used.

where adjustment is not required, the dielectric may be thin paper or mica. The choice of a dielectric and its thickness is determined by the capacity desired, the voltage for which the condenser is intended and, in many cases, by the losses in the dielectric, since the electrical stress caused by the electrostatic field is accompanied by consumption of energy which appears as a heating effect in the dielectric.

2-4 THE ELECTRIC CURRENT

Conduction — If a difference of potential exists across the ends of a conductor (by connecting the conductor — usually a wire — to a battery or generator or other source of voltage) there will be a continuous drift of electrons from atom to atom, and an electrical *current* is said to be flowing. The individual electrons do not streak from one end of the conductor to the other but the action is rather like a "bucket brigade" where, instead of firemen handing buckets down the line, atoms pass electrons down the line of the conductor. The current, or total effect of the electron drift, travels quite fast, close to the speed of light, but the electrons themselves move only a short distance.

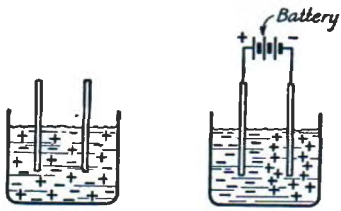


Fig. 204 — Electrolytic conduction. When an e.m.f. is applied to the electrodes, negative ions are attracted to the positively charged plate and positive ions to the negatively charged plate. The battery is indicated by its customary symbol.

The current is measured in *amperes*, and a current of one ampere represents nearly 10^{19} (ten million, million, million) electrons flowing past a point in one second. On more familiar ground, the current which flows through an ordinary 60-watt lamp is approximately one-half ampere.

Gaseous conduction (ionization) — All conduction does not necessarily take place in solid conductors. If a glass tube is fitted with metal plates at each end, and filled with a gas or even ordinary air (which is a mixture of gases) at reduced pressure, an electric current may be passed *through the gas* if a high enough voltage is applied across the metal terminals. When the voltage is applied across the tube, the positively charged plate attracts a few electrons, which acquire considerable velocity because of the electric charge and the fact that the reduced pressure in the tube (less gas) permits the electrons to travel farther before colliding with a gas atom. When one does collide with an atom, it knocks off an outer electron of the gas atom and this electron also joins the procession towards the positive plate, knocking off more electrons from other atoms as it goes. The atoms that have had an electron or two knocked off are no longer true atoms but *ions*, and since they have a positive charge (due to the electron deficiency) they are called "positive ions." These positive ions, being heavier than the electrons, travel more slowly towards the negative plate, where they acquire electrons and become neutral atoms again. The net result is a flow of electrons, and hence of current, from negative plate (called the *cathode*) to positive plate (*anode*).

Current flow in liquids — A very large number of chemical compounds have the peculiar characteristic that when they are put into solution the component parts become ionized. For example, common table salt or sodium chloride, each molecule of which is made up of one atom of sodium and one of chlorine, will, when put into water, break down into a sodium ion (positive, with one electron deficient) and a chlorine ion (negative, with one excess electron). This can only occur so long as the salt is in solution — take away the water and the ions are recombined into the neutral sodium chloride. This spontaneous *disassociation* in solution is another form of ionization, and if two wires with a difference of potential across them are placed in the solution, the negative wire will attract the positive sodium ions and the positive wire will attract the negative chlorine ions, and a current will

flow through the solution. When the ions reach the wires the electron surplus or deficiency will be remedied, and a neutral atom will be formed. The energy supplied by the source of potential difference is used to move the ions through the liquid and to supply or remove electrons. This type of current flow is called *electrolytic conduction*.

Batteries — All batteries depend upon chemical action for the generation of a potential difference across their terminals. The common *dry cell* (which will not work when *completely* dry) depends upon zinc ions (the metal case of a dry cell is the zinc plate) with a positive charge going into solution and leaving the zinc plate strongly negative. The electrical energy is derived from the chemical energy, and in time the zinc will be used up or worn away. However, in lead *storage batteries*, such as are used in automobiles for starting, the electrical energy is stored by chemical means and entails no destruction of the battery materials. The water that must be replaced from time to time is lost by evaporation.

It might be pointed out here that the term "battery" is used correctly only when speaking of more than one cell — a single cell is not a battery, but two or more connected together become a battery.

Current flow in vacuum — If a suitable metallic conductor, such as tungsten or oxide-coated or thoriated tungsten, is heated to a high temperature in a vacuum, electrons will be emitted from the surface. The electrons are

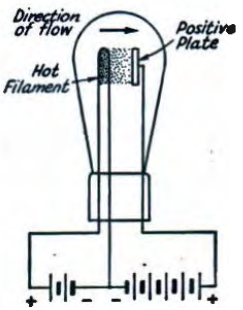


Fig. 205 — Illustrating conduction by thermionic emission of electrons in a vacuum tube. One battery is used *only* to heat the filament to a temperature where it will emit electrons. The other battery places a positive potential on the plate, with respect to the filament, and the electrons are attracted to the plate. The flow of electrons completes the electrical path, and current flows in the plate circuit.

freed from this *filament* or *cathode* because it has been heated to a temperature that activates them sufficiently to allow them to break away from the surface. The process is called *thermionic electron emission*. Now if a metal plate is placed in the vacuum tube and given a positive charge by connecting a battery between plate and cathode, this plate or *anode* will attract a number of the electrons that surround the cathode. The passage of the electrons from cathode to anode constitutes an electric current. All thermionic vacuum tubes depend for their operation on the emission of electrons from a hot cathode.

Direction of current flow — Use was being made of electricity for a long time before its electronic nature was understood, and while it is now clear that current flow is a drift of negative electrical charges or electrons toward a positive potential, in the era preceding the electron theory it was *assumed* that the current flowed from the point of higher positive potential to a point of lower (i.e., less positive or more negative) potential. While this assumption turned out to be wholly wrong, it is still customary to speak of current as flowing "from positive to negative" in many applications. The practice often causes confusion, but this distinction between "current" flow and "electron" flow often must be taken into account. If electron flow is specifically mentioned there is of course no doubt as to the meaning, but when the direction of *current* flow is specified it may be taken, by convention, as being opposite to the true direction.

● 2-5 ELECTROMAGNETISM

The magnetic field — The power that a bar or horseshoe magnet possesses of attracting small pieces of iron to itself is known to everyone. As in the case of electrostatic attraction (§ 2-3) the concept of a field of magnetic force is adopted to explain the magnetic action. The field is made up of *lines of magnetic force*, the number of which per unit area determine the strength of the field.

A moving electron generates a magnetic field of exactly the same nature as that existing about a permanent magnet. Since a moving electron, or group of electrons moving together, constitutes an electric current, it follows that the flow of current is accompanied by the creation of a magnetic field.

Conversely, when a conductor is moved through a magnetic field (or the field is moved past the conductor) electrons in the conductor are forced to move, producing a current. *An electric current generates a magnetic field about it and, conversely, an electric current is generated by a magnetic field moving (or changing) past the conductor.*

When a conductor carrying a current is placed in a magnetic field, a force is exerted on the wire which tends to move it in a direction determined by the relative directions of the flux lines of the external field and that set up by the current flow in the wire. This is a corollary of the fact that a current is induced

in a wire moving in a magnetic field, and is the principle used in the electric motor.

Magnetomotive force — When the conductor is a wire, the lines of force are in the form of concentric circles around the conductor and lie in planes at right angles to the axis of the conductor. The magnetic field constituted by these lines of force exists only when current is flowing through the wire. When the current

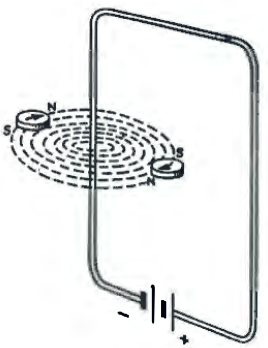


Fig. 206 — Whenever current passes through a wire, a magnetic field exists around the wire. Its direction can be traced by means of a small compass.

is started through the wire, we may visualize the magnetic field as coming into being and sweeping outward from the axis of the wire. On cessation of current flow, the field collapses toward the wire and disappears. *Thus energy is alternately stored in the field and returned to the wire.* When a conductor is wound into the form of a coil of many turns, the magnetic field becomes stronger because there are more lines of force. The force is expressed in terms of *magneto-motive force* (*m.m.f.*) which depends on the number of turns of wire, the size of the coil and the amount of current flowing through it. The same magnetizing effect can be secured with a great many turns and a weak current or with few turns and a strong current. If 10 amperes

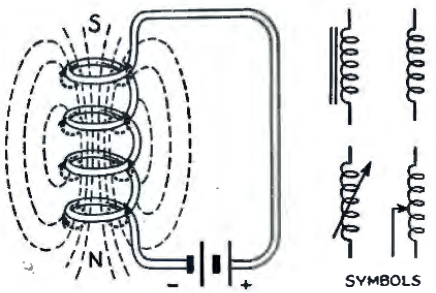


Fig. 207 — When the conducting wire is coiled, the individual magnetic fields of each turn are in such a direction as to produce a field similar to that of a bar magnet. The schematic symbols for inductance are shown at the right. The symbols at the left in the top row indicates an iron-core inductance; at right, air core. Variable inductances are shown in the bottom row.

flow in one turn of wire, the magnetizing effect is 10 *ampere-turns*. Should one ampere flow in 10 turns of wire, the magnetizing effect is also 10 ampere-turns.

Inductance — When a source of voltage is connected across a coil, the current does not immediately reach its final fixed value. The reason for this is that, as the current starts to flow through the coil, the magnetic field around the coil builds up, and as the field changes it induces a voltage back in the coil. The current caused by this induced voltage is always in the opposite direction to the current originally passed through the coil. Therefore, because of this property of *self-induction*, the coil tends constantly to oppose any change in the current flowing through it, and it takes an appreciable amount of time for the current to reach its normal value through the coil. The effect can be visualized as electrical inertia. After the current has come to a steady value, the self-inductance has no effect, and the current is only limited by the resistance of the wire in the coil.

The inductance of a coil is measured in *henrys* or, when smaller units are more convenient, the *millihenry* (one thousandth of a henry) or *microhenry* (one-millionth of a henry). The inductance of a coil depends on several factors, chief of which are the number of turns, the cross-sectional area of the coil, and the material in the center of the coil, or core. A core of magnetic material will greatly increase the inductance of a coil, just as certain dielectrics greatly increase the capacity of a condenser (§ 2-3). Even a straight wire has inductance, although small compared to that of a coil.

The inductance of a straight wire of given length is less as the diameter of the wire is increased. In general, a conductor of large cross-sectional area, or large surface, will have less inductance than one of small area but having the same length.

Magnetic circuits and units — Unlike electrostatic lines of force, magnetic lines of force must always be *closed*, forming circles or loops, so that the complete magnetic path of the lines of force must be considered in computing the effect of a magnetic core material on the inductance of a coil. The measure of the number of magnetic lines of force set up in a closed magnetic path or circuit through a given material for a specified applied m.m.f. is called the *magnetic permeability* of the material. It is expressed as a ratio to the number of lines set up by the same coil with the same applied m.m.f. with air as the core material, air therefore being assigned a permeability of unity. If the magnetic circuit is partly through a magnetic material and partly through a non-magnetic material (as in the case of a coil wound on a straight bar of iron, where part of the magnetic

path must be through air) the permeabilities of both mediums must be taken into account.

Permeability corresponds to conductivity in conductors, and its reciprocal, *reluctance*, corresponds to resistance. *Magnetic flux density*, or lines of force per unit area, is in the magnetic circuit equivalent to current in the electrical circuit, while the magnetomotive force is analogous to electromotive force or voltage.

• 2-6 FUNDAMENTAL RELATIONS

Ohm's law — The current in a conductor is determined by two things, the voltage across the conductor and the resistance of the conductor. The unit of resistance is the *ohm*, and, by definition, an e.m.f. of one volt will cause a current of one ampere to flow through a resistance of one ohm. Since the three quantities are interdependent, if we know the values of any two we can easily determine the third by the simple relation known as Ohm's Law. When I is the current in amperes, E is the electromotive force in volts and R is the circuit resistance in ohms, the formulas of Ohm's Law are:

$$R = \frac{E}{I} \qquad I = \frac{E}{R} \qquad E = IR$$

The *resistance* of the circuit can therefore be found *by dividing the voltage by the current:* the *current* can be found *by dividing the voltage by the resistance:* the *electromotive force or e.m.f.* is equal to the *product of the resistance and the current.*

The resistance of any metallic conductor depends upon the material and its temperature, its cross-sectional area and the length of the conductor. When resistance is deliberately added to a circuit, as is often done to adjust voltages or limit current flow, the resistance is usually lumped in a single unit and the unit is called a *resistor*.

Heating effect and power — When current passes through a conductor there is some molecular friction, and this friction generates heat. The heat generated is dependent only upon the current in the conductor, the resistance of the conductor and the time during which the current flows. The power used in heating (which may be considered sometimes as an undesired power loss) can be determined by substitution in the following equations:

$$P = EI,$$
$$\text{or } P = I^2R,$$
$$\text{or } P = \frac{E^2}{R}$$

P being the power in watts, E the e.m.f. in volts, and I the current in amperes.

It will be noted that if the current in a resistor and the resistance value are known, we can readily find the power. Or if the voltage across a resistance and the current through it are known, the product of volts and amperes will give the power. Knowing the value of a resistor (ohms) and the applied voltage across it, the power dissipated is given by the last formula.

Likewise, when the power and resistance in a circuit are known, the voltage and current can be calculated by the following equations derived from the power formulas given above:

$$E = \sqrt{PR}$$
$$I = \sqrt{\frac{P}{R}}$$

Units — Besides the fundamental units — volt, ampere, watt — fractional and multiple units frequently are convenient. Thus a *milliampere* is 1/1000 ampere and a *microampere* is 1/1,000,000 ampere. *Millivolt* and *microvolt* are corresponding fractional units of the volt. The *kilovolt* also is a frequently used unit; it is equal to 1000 volts. Resistance is frequently expressed in *megohms* (1 megohm = 1,000,000 ohms) and sometimes in *kilohms* (1000 ohms). Other units for power are the *microwatt*, *milliwatt*, and *kilowatt*, having equivalent meanings to those above. The *watt-hour* and *kilowatt-hour* are energy units, representing the total energy consumed when it is delivered at a given power rate for a given period of time; the numerical values are equal to the product of power and time in the units named.

Unless otherwise specified, formulas are always given in terms of the fundamental units, so that fractional or multiple units must first be converted to the fundamental units before an equation can be used.

Resistances in series and parallel — Resistors may be connected in series, in parallel or in series-parallel, as shown in Fig. 208. When two or more resistors are connected in series, the total resistance of the group is

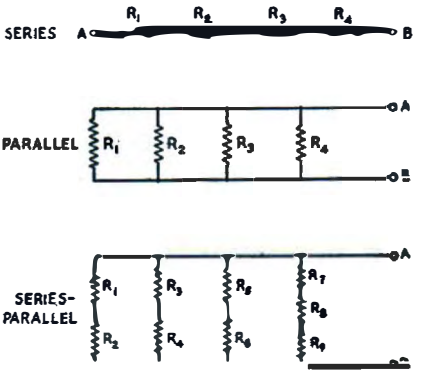


Fig. 208— Diagrams of series, parallel and series-parallel resistance connections.

higher than that of any of the units. Should two or more resistors be connected in parallel, the total resistance is decreased. Fig. 208 and the following formulas show how the value of a bank of resistors in series, parallel or series-parallel may be computed, the total resistance being that which appears between A and B in each case.

Resistances in series:

$$\textit{Total resistance} = R_1 + R_2 + R_3 + R_4$$

Resistances in parallel:

$$\textit{Total resistance} = \frac{1}{\frac{1}{R_1}+\frac{1}{R_2}+\frac{1}{R_3}+\frac{1}{R_4}}$$

Or, in the case of only two resistances in parallel,

$$\textit{Total resistance} = \frac{R_1 R_2}{R_1 + R_2}$$

Resistances in series-parallel:

$$\textit{Total resistance} = \frac{1}{\frac{1}{R_1+R_2}+\frac{1}{R_3+R_4}+\frac{1}{R_5+R_6}+\frac{1}{R_7+R_8+R_9}}$$

This means that in series-parallel circuits the various groups of series resistors should first be added, then each group treated as a single resistor, so that the formula for resistances in parallel can be used.

Voltage dividers and potentiometers — Since the same current flows through resistors connected in series, it follows from Ohm's Law that the voltage (termed *voltage drop*) across each resistor of a series-connected group is proportional to its resistance. Thus in Fig. 209-A the voltage E_1 across R_1 is equal to the applied voltage E multiplied by the ratio of R_1 to the total resistance, or

$$E_1 = \frac{R_1}{R_1 + R_2 + R_3} \cdot E$$

Similarly, the voltage E_2 is equal to

$$\frac{R_1 + R_2}{R_1 + R_2 + R_3} \cdot E$$

Such an arrangement is called a *voltage divider*. When current is drawn from the divider at the various tap points the above relations are no longer strictly true, since the same current does not flow in all parts of the divider. Design data for such cases are given in § 8-10.

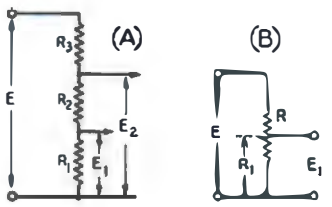


Fig. 209 — The voltage divider or potentiometer.

A similar arrangement is shown in Fig. 209-B, where the total resistance R is equipped with a sliding tap for fine adjustment. Such a resistor is frequently called a *potentiometer*, although the word is not used in its original sense.

Inductances in series and parallel — The formulas for the total inductance of a group of separate inductances connected in series, parallel, or series parallel are exactly the same as those given in the previous paragraph for resistances, provided only that the magnetic fields about the coils are not permitted to interact with each other.

Condensers in series and parallel — The total capacity of a group of condensers connected in series, parallel or series parallel can be computed by formulas similar to those used for resistances and inductances, but with the series and parallel formulas interchanged. Thus, for condensers in parallel,

$$\text{Total capacity} = C_1 + C_2 + C_3 + C_4, \text{ etc.}$$

For condensers in series,

$$\text{Total capacity} = \frac{1}{\frac{1}{C_1}+\frac{1}{C_2}+\frac{1}{C_3}+\frac{1}{C_4}}$$

or for two condensers in series

$$\text{Total capacity} = \frac{C_1 C_2}{C_1 + C_2}$$

With condensers in series parallel, first compute the resultant capacity of the condensers in series in each parallel branch, then add the capacities so found for the various branches.

Time constant — When a condenser and resistor are connected in series with a source of e.m.f. such as a battery the initial flow of current into the condenser is limited by the resistance, so that a longer period of time is required to complete the charging of the condenser than would be the case without the resistor. Likewise, when the condenser is discharged through a resistance, a measurable period of time is taken for the current flow to reach a negligible value. In the case of either charge or discharge the time required is proportional to the capacity and resistance, the product of which is called the *time constant* of the circuit. If C is in farads and R in ohms, or C in microfarads and R in megohms, this product gives the time in seconds required for the voltage across a discharging condenser to drop to $1/e$ or approximately 37% of its original value. (The constant e is the base of the natural series of logarithms.)

A circuit containing inductance and re-

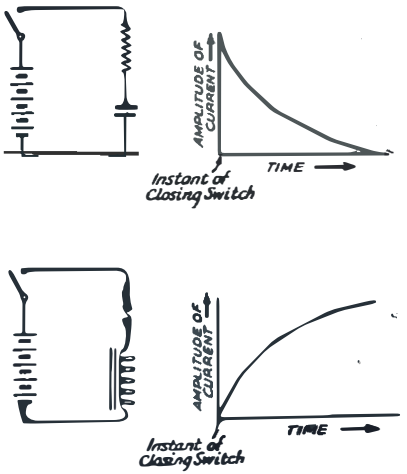


Fig. 210 — Showing how the current in a circuit combining resistance with inductance or capacity takes a finite period of time to reach its steady-state value.

sistance also has a time constant, for similar reasons. The time constant of an inductive circuit is equal to L/R, and when L is in henrys and R in ohms gives the time in seconds required for the current to reach $1-1/e$ or approximately 63% of its final steady value when a constant voltage is applied to the circuit.

Measuring instruments — Instruments for measuring d.c. current and voltage make use of the force acting on a coil carrying current in a magnetic field (§ 2-5), produced by a permanent magnet, to move a pointer along a calibrated scale. All such instruments are therefore current operated, the current required for full-scale deflection of the pointer varying from several milliamperes to a few microamperes according to the sensitivity required. If the instrument is to read high currents, it is *shunted* (paralleled) by a low resistance through which most of the current flows, leaving only enough flowing through the instrument to give a full scale deflection corresponding to the total current flowing through both meter and shunt. An instrument which reads microamperes is called a *microammeter* or *galvanometer;* one calibrated in milliamperes is called a *milliammeter;* one calibrated in amperes is an *ammeter*. A *voltmeter* is simply a milliammeter with a high resistance in series so that the current will be limited to a suitable value when the instrument is connected across a voltage source; it is calibrated in terms of the voltage which must appear across the terminals to cause a given value of current to flow. The series resistance is called a *multiplier*. A *wattmeter* is a combination voltmeter and ammeter in which the pointer deflection is proportional to the power in the circuit.

An ammeter or milliammeter is connected in series with the circuit in which current is being measured, so that the current flows through the instrument. A voltmeter is connected in parallel with the circuit.

• 2-7 ALTERNATING CURRENT

Description — In self-induction the induced voltage always opposes the voltage causing the original current flow (§ 2-5). Similarly, if a closed wire is placed in an expanding magnetic field, the current induced in the wire by the changing field will flow in such a direction that the magnetic field set up in turn by this induced current opposes the field which originally caused it. Now if the original field is caused to collapse (moving toward the wire instead of outward from it) the induced current will change its direction so that its field again will be in opposition to the original field. If the primary field regularly builds up and collapses the current will change direction correspondingly; in other words, it is an *alternating current*. Since current is only caused to flow by a *changing* magnetic field, it is easy to see why alternating currents are widely used; they are a natural result of the application of the principle of induction.

The simplest form of alternating current (or voltage) is shown graphically in Fig. 211. This chart shows that the current starts at zero value, builds up to a maximum in one direction, comes back down to zero, builds up to a maximum in the opposite direction and comes back to zero. The curve followed is described mathematically as a *sine curve;* its wavelike nature causes it to be known as a *sine wave*.

Frequency — The complete wave shown in Fig. 211 is called a *cycle*, or *period*. Each half of the cycle, during which the current is flowing in one direction, although its strength is varying, is known as an *alternation*. The number of cycles the wave goes through each second of time is called the *frequency* of the current. Frequencies vary from a few cycles per second for power line alternating currents to many

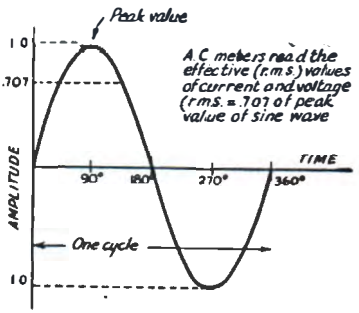


Fig. 211 — Representing sine-wave alternating current or voltage.

millions per second in radio circuits. For convenience, two other units, the *kilocycle* (1000 cycles) and the *megacycle* (1,000,000 cycles) also are used. The abbreviations for these are kc. and Mc., respectively.

Electrical degrees — If we take a fixed point on the periphery of a revolving wheel, we find that at the end of each revolution, or cycle, the point has come back to its original starting place. Its position at any instant can be expressed in terms of the angle between two lines, one drawn from the center of the wheel to the point at the instant of time considered, the other drawn from the wheel center to the starting point. In making one complete revolution the point has travelled through 360 degrees, a half revolution 180 degrees, a quarter revolution 90 degrees, and so on. The periodic wave of alternating current may be treated similarly, one complete cycle equalling one revolution or 360 degrees, one alternation (half cycle) 180 degrees, and so on. With the cycle divided up in this way, the sine curve simply means that the value of current at any instant is proportional to the sine of the angle which corresponds to the particular fraction of the cycle considered.

The concept of angle is universally used in alternating currents. Generally, it is expressed in the fundamental form, using the radian rather than the degree as a unit, whence a cycle is equal to 2π radians, or a half cycle to π radians. The expression $2\pi f$, for which the symbol ω is often used, simply means electrical degrees per cycle times frequency, and is called the *angular velocity*. It gives the total number of electrical radians passed through by a current of given frequency in one second.

Waveform, harmonics — The sine wave is not only the simplest but in many respects the most desirable waveform. Many other waveforms are met with in practice, however, and they may differ considerably from the simple sine case. It is possible to show by analysis that any such waveform can be resolved into a number of components of differing frequencies and amplitudes, but related in frequency in such a way that all are integer multiples of the lowest frequency present. The lowest frequency is called the *fundamental*, and the multiple frequencies are called *harmonics*. Thus a wave may consist of fundamental, 3rd, 5th, and 7th harmonics, meaning, if the fundamental frequency is say 100 cycles, that frequencies of 300, 500 and 700 cycles also are present in the wave.

Effective, peak and average values — It is evident that both the voltage and current are swinging continuously between their positive maximum and negative maximum values, and it might be wondered how one can speak of so many amperes of alternating current when the value is changing continuously. The problem is simplified in practical work by considering that *an alternating current has an effective value of one ampere when it produces heat at the same average rate as one ampere of continuous direct current flowing through a given resistor.* This effective value is the square root of the mean of all the instantaneous current values squared. For the sine-wave form,

$$E_{\text{eff}} = \sqrt{\tfrac{1}{2}E^2_{\text{max}}}$$

For this reason, the effective value of an alternating current, or voltage, is also known as the *root-mean-square* or *r.m.s.* value. Hence, the effective value is the square root of $\frac{1}{2}$ or 0.707 of the maximum value — practically considered 70% of the maximum value.

Another important value, involved where alternating current is rectified to direct current, is the *average*. This is simply the average of all instantaneous values in the wave, and for a sine wave is equal to 0.636 of the maximum (or peak) value of either current or voltage. The three terms *maximum* (or *peak*), *effective* (or *r.m.s.*) and *average* are encountered frequently in radio work. For the sine form they are related to each other as follows:

$$\begin{aligned} E_{\text{max}} &= E_{\text{eff}} \times 1.414 = E_{\text{ave}} \times 1.57 \\ E_{\text{eff}} &= E_{\text{max}} \times 0.707 = E_{\text{ave}} \times 1.11 \\ E_{\text{ave}} &= E_{\text{max}} \times 0.636 = E_{\text{eff}} \times 0.9 \end{aligned}$$

The relationships for current are the same as those given above for voltage.

Phase — It has been mentioned that in a circuit containing inductance, the rise of current is delayed by the effect of electrical inertia presented by the inductance (§ 2-5). Both increases and decreases of current are similarly

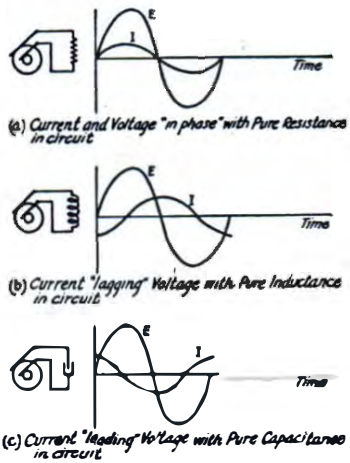


Fig. 212 — Phase relationships between voltage and current in resistive and reactive circuits. The symbol at the left represents a generator.

delayed. It is also true that a current must flow into a condenser before its elements can be charged and so provide a voltage difference between its terminals. Because of these facts, we say that a current "lags" behind the voltage in a circuit which has a preponderance of inductance and that the current "leads" the voltage in a circuit where capacity predominates. Fig. 212 shows three possible conditions in an alternating current circuit. In the first, when the load is a pure resistance, both voltage and current rise to the maximum values simultaneously. In this case the voltage and current are said to be *in phase*. In the second instance, the existence of inductance in the circuit has caused the current to lag behind the voltage. In the diagram, the current is lagging one quarter cycle behind the voltage. The current is therefore said to be 90 degrees *out of phase* with the voltage. In the third example, with a capacitive load, the voltage is lagging one quarter cycle behind the current. The *phase difference* is again 90 degrees. These, of course, are theoretical examples in which it is assumed that the inductance and the condenser have no resistance. Actually, the angle of lag or lead (*phase angle*) depends on the amounts of inductance, capacity and resistanee in the circuit.

The phase relationships between two currents (or two voltages) of the same frequency are defined in the same way. When two such currents are combined the resultant is a single current of the same frequency, but having an instantaneous amplitude equal to the algebraic sum of the amplitudes of the two components at the same instant. The amplitude of the resultant current hence is determined by the phase relationship between the two currents before combination. Thus if the two currents are exactly in phase, the maximum value of the resultant will be the numerical sum of the maximum values of the individual currents; if they are 180 degrees out of phase, one reaches its positive maximum at the instant the other reaches its negative maximum, hence the resultant current is the difference between the two. In the latter case, if the two currents have the same amplitude the resultant current is zero.

The a.c. spectrum — Alternating currents of different frequencies have different properties and are useful in many varieties of ways. For the transmission of power to light lamps, run motors, and perform familiar everyday tasks by electrical means, low frequencies are most suitable. Frequencies of 25, 50 and 60 cycles are in common use, the latter being most widespread. The range of frequencies between about 30 and 15,000 cycles is known as the *audio-frequency* range, because when frequencies of this order are converted from a.c. into air vibrations, as by a loudspeaker or telephone receiver, they are distinguishable as sounds, having a tone pitch proportional to the frequency. Frequencies between 15,000 cycles (15 kilocycles) and about 1,000,000,000 cycles (1000 megacycles) are used for radio communication, because with frequencies of this order it is possible to convert electrical energy into radio waves. The latter frequency is about the highest it is possible to generate at present, but does not necessarily represent the highest frequency that could be used for radio work.

The a.c. spectrum is divided into the following approximate classifications for convenience in reference:

15–15,000 cycles	Audio frequencies
15–100 kilocycles	Low radio frequencies
100–1500 kilocycles	Medium radio frequencies
1.5–6 megacycles	Medium high frequencies
6–30 megacycles	High frequencies
Above 30 megacycles	Ultra-high frequencies

● 2-8 OHM'S LAW FOR ALTERNATING CURRENTS

Resistance — Since current and voltage are always in phase through a resistance, the instantaneous relations are equivalent to those in direct-current circuits, and since by definition the units of current and voltage for a.c. are made equal to those for d.c. in resistive circuits, the various formulas expressing Ohm's Law for d.c. circuits apply without any change for a.c. circuits containing resistance only, or for purely resistive parts of complex a.c. circuits. The formulas are given in § 2-6.

Reactance — In an a.c. circuit containing inductance or capacity, the current and voltage are not in phase (§ 2-7) so that Ohm's Law cannot be applied directly. The current is not limited by resistance, as in d.c. circuits, but by a quantity called *reactance*, which expresses the opposing effect of the voltage of self-induction (§ 2-5), in the case of an inductance, and the accumulation of charge in the case of a condenser. In circuits containing only reactance no energy is consumed, since the energy put into an inductance or capacity in one part of the cycle is stored in the electromagnetic or electrostatic field and is returned to the circuit in another part of the cycle. Thus in a purely reactive circuit it is possible to have both high voltage and high current without the consumption of any power. Of course in practice there is always some resistance in the wire of an inductance, or heating of the dielectric of a condenser, so that some energy may be lost, but it is usually negligible in well-designed components.

Reaetanee is expressed in ohms, the same unit as for resistance, since with a given reactance at a given frequency the current that

will flow is proportional to the applied voltage. Hence,

$$X = \frac{E}{I}$$

for a purely reactive circuit. X is the symbol for reactance.

In circuits containing both resistance and reactance the values of each cannot be added directly because of the different phase relations.

Inductive reactance — The greater the inductance of a coil, the greater is the effect of self-induction (§ 2-5), or the opposition to a change in the value of current, hence the higher the reactance. Also, the higher the frequency the greater the reactance, since the greater the rate of change of current the more opposition the coil offers to the change. Hence, inductive reactance is proportional to inductance and frequency, or

$$X_L = 2\pi fL$$

It will be recognized here that angular velocity, $2\pi f$ (§ 2-7), expresses the rapidity with which the current changes.

The fundamental units (ohms, cycles, henrys) must be used in the above equation, or appropriate factors inserted in case other units are employed. If inductance is in millihenrys, frequency should be in kilocycles; if inductance is in microhenrys, frequency should be in megacycles, to bring the answer in ohms.

Capacitive reactance — When a condenser is used in an a.c. circuit it is rapidly charged and discharged as the a.c. voltage rises and falls and reverses in polarity. This repetition of charge and discharge constitutes the flow of alternating current through the condenser. Since for a given voltage the energy stored in the condenser is fixed by its capacity (§ 2-3) it is obvious that the total amount of energy stored in the condenser (and subsequently restored to the circuit) in one second will be greater when the condenser is charged many times per second than when it is charged only a few times. Hence the current flow will be proportional to the frequency and to the capacity of the condenser, or conversely the reactance will be inversely proportional to the frequency and the capacity. Therefore

$$X_C = \frac{1}{2\pi fC}$$

where $2\pi f$ again is the angular velocity or the rapidity with which the current changes. When f is in cycles per second and C in farads, X_C will be in ohms. If C is in microfarads, f must be expressed in megacycles to bring the resistance in ohms.

Impedance — In circuits containing inductive reactance the current lags the voltage while with capacity reactance the current leads (§ 2-7). Hence the effects of inductive and capacitive reactance are opposite in sense, or, as it is commonly expressed, inductive and capacitive reactances cancel each other. In series circuits having both inductive and capacitive reactance the net reactance is the difference between the two, and the current will either lead or lag depending upon which is larger, capacitive or inductive reactance. Inductive reactance is considered positive and capacitive reactance negative, so that

$$X = X_L - X_C$$

The combined effect of resistance and reactance is termed *impedance*. The symbol for impedance is Z and, for a series circuit, it is computed from the formula:

$$Z = \sqrt{R^2 + X^2}$$

where R is the resistance and X is the reactance. The terms Z, R and X are all expressed in ohms. Ohm's Law for alternating current circuits then becomes

$$I = \frac{E}{Z}; \quad Z = \frac{E}{I}; \quad E = IZ$$

The phase angle depends upon the relative amounts of resistance and reactance, becoming more nearly zero (current and voltage in phase) when reactance is small compared to resistance, and more nearly 90 degrees when resistance is small compared to reactance.

Power factor — The power dissipated in an a.c. circuit containing both resistance and reactance is consumed entirely in the resistance, hence is equal to I^2R. However, the reactance is also effective in determining the current or voltage in the circuit, even though it consumes no energy. Hence the product of volts times amperes (which gives the power consumed in d.c. circuits) for the whole circuit may be several times the actual power used up. The ratio of power dissipated (watts) to the *volt-ampere* product is called the power factor of the circuit, or

$$\text{Power factor} = \frac{\text{Watts}}{\text{Volt-amperes}}$$

Distributed capacity and inductance — It should not be thought that the reactance of coils becomes infinitely high as the frequency is increased to a high value and, likewise, that the reactance of condensers becomes infinitely low at high frequencies. All coils have some capacity between turns, and the reactance of this capacity can become low enough at some high frequencies to tend to cancel the high reactance of the coil. Likewise, the leads and plates of condensers will have considerable inductance at very high frequencies, which will tend to offset the capacitive reactance of the condenser itself. For these reasons, coils for high-frequency work must be designed to have

low "distributed" capacity, and condensers must be made with short, heavy leads to have low inductance.

Units and instruments — The units used in a.c. circuits may be divided or multiplied to give convenient numerical values to different orders of magnitude, just as in d.c. circuits (§ 2-6). Because the rapidly reversing current is accompanied by similar reversals in magnetic field, instruments used for measurement of d.c. (§ 2-6) will not operate on a.c. At low frequencies suitable instruments can be constructed by making the current produce both magnetic fields, one by means of a fixed coil and the other by the moving coil. Such instruments are used for measurement of either current or voltage. At radio frequencies this type of instrument is inaccurate because of distributed capacity and other effects, and the only reliable type of direct-reading instrument is the *thermocouple ammeter* or milliammeter. This is a power-operated device consisting of a resistance wire, heated by the flow of r.f. current through it, to which is attached a thermocouple, or pair of wires of dissimilar metals joined together and possessing the property of developing a small d.c. voltage between the terminals when heated. This voltage, which is proportional to the heat applied to the couple, is used to operate a d.c. instrument of ordinary design.

● 2-9 THE TRANSFORMER

Principles — If two coils of wire are wound on a laminated iron core and one of the coils is connected to a source of alternating current, it will be found that there is an alternating voltage across the terminals of the other coil of wire, and an alternating current will flow through a conductor connecting the two terminals. The alternating current in the first coil, or *primary*, causes a changing magnetic field in the iron core, and this changing magnetic field induces an alternating current in the second coil, or *secondary*. This is simply an application of the principle of induction (§ 2-5) with the induced voltage being caused by a varying magnetic field set up by a current flowing in a separate winding instead of the same coil.

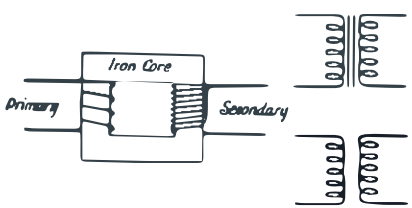


Fig. 213 — The transformer. Power is transferred from one coil to the other by means of the magnetic field. The upper symbol at the right indicates an iron-core transformer, the lower one an air-core transformer.

Voltage and turn ratio — For a given varying magnetic field, the voltage induced in a coil in the field will be proportional to the number of turns on the coil. Since the two coils of a transformer are in the same field, it follows that the induced voltages will be proportional to the number of turns on each coil. In the case of the primary, or coil connected to the source of power, the induced voltage is practically equal to, and opposes, the applied voltage. Hence, the secondary induced voltage is very nearly equal to the voltage applied to the primary, multiplied by the ratio of the number of turns on the secondary to the number of turns on the primary.

Voltage and current relations — A transformer cannot deliver more power to its secondary load than it takes from the primary source of power, since to do so would be to violate the principle of conservation of energy. Hence we find that transforming a given voltage to a new value causes an inverse transformation in the current delivered to the load as compared to that taken from the line. For example, a transformer with a secondary-to-primary voltage ratio of 5 will have a current ratio of ⅕, which means that the primary current will be five times the secondary current. A voltage ratio of less than unity gives a corresponding increase in secondary-to-primary current ratio. Actually these ratios are not exact, since the transformer will have some losses both in the wire and in the iron core, and this additional loss appears as power taken by the primary which is not available for the secondary load. The *efficiency*, or ratio of power delivered to the load to power taken from the line, of small transformers may vary between 60% and 90%, depending upon the design.

Impedance ratio — In a properly designed iron-core transformer practically all the magnetic lines of force cut both primary and secondary coils, hence the relationship between secondary current and primary current described in the preceding paragraph. The only reactance present is that due to "leakage," or magnetic flux lines which cut one coil but not the other. Since the leakage reactance is small, a transformer having a resistive load on its secondary will also "look like" a practically resistive load to the power line which supplies its primary. The impedance is equal to E/I (§ 2-8) and, neglecting losses, if n is the secondary-to-primary turn ratio, then

$$\frac{E_p}{I_p} = \frac{\frac{E_s}{n}}{nI_s} = \frac{E_s}{n^2 I_s} \text{ or } n^2 \frac{E_p}{I_p} = \frac{E_s}{I_s}$$

That is, the impedance (E_p/I_p) presented by

the primary to the line (called the *reflected impedance* or *reflected load*) is equal to the secondary load impedance (E_s/I_s) divided by the square of the secondary-to-primary turns ratio. The *impedance ratio*, or ratio of secondary load impedance to impedance presented by the primary to the line, is therefore equal to the square of the turn ratio. This relation is very frequently used in radio circuits.

Impedance matching — Many devices require a specific value of load resistance (or impedance) for optimum operation. The resistance of the actual load which is to dissipate the power may differ widely from this value, hence the transformer with its impedance-transforming properties is frequently called upon to change the actual load to the desired value. This is called *impedance matching*. From the preceding paragraph,

$$\frac{N_s}{N_p} = \sqrt{\frac{Z_s}{Z_p}}$$

where N_s/N_p is the required secondary-to-primary turn ratio, Z_s is the impedance of the actual load, and Z_p the impedance required for optimum operation of the device delivering the power.

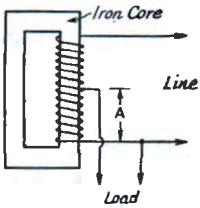


Fig. 214 — The autotransformer. Line and load currents in the common winding (A) flow in opposite directions so that the resultant current is the difference between them.

The autotransformer — The transformer principle can be utilized with only one winding instead of two, as shown in Fig. 214; the principles just discussed apply equally well. The *autotransformer* has the advantage that the line and load currents in the common section are out of phase, hence this portion of the winding carries less current than the remainder of the coil. This advantage is not very marked unless the primary and secondary voltages do not differ very greatly, while it is frequently disadvantageous to have a direct connection between primary and secondary circuits. For these reasons its application is usually limited to boosting or reducing the line voltage for voltage correction or similar purposes.

• 2-10 RESONANT CIRCUITS

Principle of resonance — It has been shown (§ 2-8) that the inductive reactance of a coil and the capacitive reactance of a condenser are oppositely affected by frequency. In any combination of inductance and capacitance, therefore, there is one particular frequency for which the inductive and capacitive reactances are equal and, since these two reactances cancel each other, the net reactance becomes zero, leaving only the resistance of the circuit to impede the flow of current. The frequency at which this occurs is known as the *resonant frequency* of the circuit and the circuit is said to be *in resonance* at that frequency or *tuned* to that frequency.

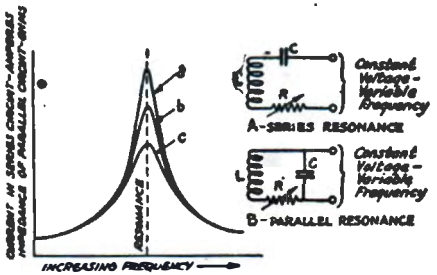


Fig. 215 — Characteristics of series-resonant and parallel-resonant circuits.

Series circuits — The resonant frequency of a simple circuit containing inductance and capacity in series is given by

$$f = \frac{1}{2\pi\sqrt{LC}} \times 10^6$$

where

f is the frequency in kilocycles per second
2π is 6.28
L is the inductance in microhenrys (μh.)
C is the capacitance in micro-microfarads ($\mu\mu$fd.)

The resistance that may be present does not enter into the formula for resonant frequency.

With a constant-voltage alternating current applied as shown in A of Fig. 215 the current flowing through such a circuit will be maximum at the resonant frequency. The magnitude of the current will be determined by the resistance in the circuit. The curves of Fig. 215 illustrate this, curve *a* being for low resistance and curves *b* and *c* being for greater resistances.

Parallel circuits — The parallel resonant circuit is illustrated in B of Fig. 215. This also contains inductance, capacitance and resistance in series, but the voltage is applied in parallel with the combination instead of in series with it as in A. Here we are primarily interested in the characteristics of the circuit as viewed from its terminals, especially in the *parallel impedance* it offers. The variation of parallel impedance of a parallel resonant circuit with frequency is illustrated by the same curves of Fig. 215 that show the variation in current with frequency for the series-resonant circuit. The parallel impedance is maximum at resonance and increases as the series resistance is made smaller.

In the case of parallel circuits, resonance may be defined in three ways: the condition which gives maximum impedance, that which gives maximum power factor, or (as in series circuits) when the inductive and capacitive reactances are equal. If the resistance is low, the resonant frequencies obtained on the three bases are practically identical. This condition usually is satisfied in radio work, so that the resonant frequency of a parallel circuit is generally computed by the series-resonance formula given above.

Resistance at high frequencies — At radio frequencies the resistance of a conductor may be considerably higher than its resistance to direct current or low-frequency a.c. This is because the magnetic field set up inside the wire tends to force the current to flow in the outer part of the wire, an effect which increases with frequency. At high radio frequencies this *skin effect* is so pronounced that practically all the current flows very near the surface of the conductor, thereby in effect reducing the cross-sectional area and hence increasing the resistance. For this reason low resistance can be achieved only by using conductors with large surface area, but since the inner part of the conductor does not carry current, thin tubing will serve just as well as solid wire of the same diameter.

A similar effect takes place in coils for radio frequencies, where the magnetic fields cause a concentration of current in certain parts of the conductors, again causing an effective decrease in the conductor size and raising the resistance. These effects, plus the effects of stray currents caused by distributed capacity (§ 2-8), raise the effective resistance of a coil to many times the d.c. resistance of the wire.

Sharpness of resonance — The resonance curves become "flatter" for frequencies near resonance frequency, as shown in Fig. 215, as the internal series resistance is increased, but are of the same shape for all resistances at frequencies farther removed from resonance frequency. The relative sharpness of the resonance curve near resonance frequency is a measure of the *sharpness of tuning or selectivity* (ability to discriminate between voltages of different frequencies) in such circuits. This is an important consideration in tuned circuits used for radio work.

Flywheel effect; Q — A resonant circuit may be compared to a flywheel in its behavior. Just as such a wheel will continue to revolve after it is no longer driven, so also will oscillations of electrical energy continue in a resonant circuit after the source of power is removed. The flywheel continues to revolve because of its stored mechanical energy; current flow continues in a resonant circuit by virtue of the energy stored in the magnetic and electric fields of the condenser and coil. The sharpness of resonance, which is directly related to the flywheel effect, is determined by the ratio of energy stored to energy dissipated, hence is proportional to the reactance in the circuit and inversely proportional to the resistance. This ratio of stored energy to dissipated energy is called the *Q* of the circuit.

In resonant circuits at frequencies below about 28 Mc. the resistance is practically wholly in the coil; condenser resistance may be neglected. Consequently the *Q* of the circuit as a whole is determined by the *Q* of the coil, or its ratio of reactance to resistance. Coils for frequencies below the ultra-high frequency region may have *Q's* ranging from 100 to several hundred, depending upon their size and construction.

Damping, decrement — The rate at which current dies down in amplitude in a resonant circuit after the source of power has been removed is called the *decrement* or *damping* of the circuit. A circuit with high decrement (low *Q*) is said to be highly damped; one with low decrement (high *Q*) is lightly damped.

Voltage rise — When a voltage of the resonant frequency is inserted in series in a resonant circuit, the voltage which appears across either the coil or condenser is considerably higher than the applied voltage. This is because the current in the circuit is limited only by the resistance and hence may have a relatively high value; however, the same current flows through the high reactances of the coil and condenser and consequently causes large voltage drops (§ 2-8). As explained above, the reactances, and hence the voltages, are opposite in phase so that the net voltage around the circuit is only that applied. The ratio of the reactive voltage to the applied voltage is proportional to the ratio of reactance to resistance, which is the *Q* of the circuit. Hence the voltage across either the coil or condenser is equal to *Q* times the voltage inserted in series with the circuit.

Parallel-resonant circuit impedance — The parallel-resonant circuit offers pure resistance (its *resonant impedance*) between its

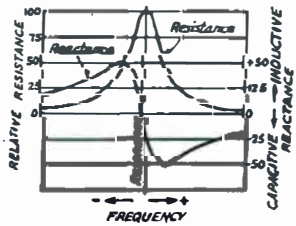


Fig. 216 — The impedance of a parallel-resonant circuit separated into its reactance and resistance components. The parallel resistance is equal to the parallel impedance at resonance.

terminals at the resonant frequency. If the internal or series resistance of the coil is low so that the impedance of the inductance branch is practically the same as its reactance, the current through the coil is equal to the applied voltage divided by the reactance (§ 2-8). The current through the condenser also is equal to E/X. Since the two reactances are equal at resonance the two currents also are equal, and since they are nearly 180 degrees out of phase (§ 2-7) they cancel each other in the external circuit, or line. The small current which flows in the line results from the fact that the resistance in the inductance causes the phase angle to be slightly less than 90 degrees in this branch so that complete cancellation cannot take place. The impedance ($Z = E/I$) is high because the line current is small, and is resistive because the current is practically in phase with the applied voltage. At frequencies off resonance the current increases through the branch having the lower reactance (and vice versa) so that the circuit becomes reactive, and the resistive component of the impedance decreases as shown in Fig. 216.

If the circuit Q is about 10 or more the parallel impedance at resonance is given by the formula

$$Z_r = X^2/R = XQ$$

where X is the reactance of either the coil or the condenser and R is the internal resistance.

Q of loaded circuits — In many applications, particularly in receiving, the only resistance present in the resonant circuit is that of the circuit itself. Hence the coil is designed to have as high Q as possible. Since, within limits, increasing the number of turns raises the reactance faster than it raises the resistance, coils for such purposes are made with relatively large inductance for the frequency under consideration.

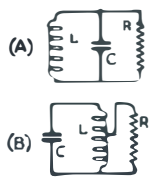


Fig. 217 — The equivalent circuit of a resonant circuit delivering power to a load. The resistor R represents the load resistance. At B the load is shown tapped across part of L, which by transformer action is equivalent to using a higher value of load resistance across the whole circuit.

When the circuit delivers energy to a load, as in the case of resonant circuits used in transmitters, the energy consumed in the circuit itself is usually negligible compared with that consumed by the load. The equivalent of such a circuit can be represented as shown in Fig. 217-A where the parallel resistor represents the load to which power is delivered. Since $Z = XQ$, the Q of a circuit loaded with a resistive impedance Z is (neglecting internal resistance)

$$Q = \frac{Z}{X}$$

Hence for a given parallel impedance, the effective Q of the circuit including the load is proportional to $1/X$, or inversely proportional to the reactance of either the coil or the condenser. A circuit loaded with a relatively low resistance (a few thousand ohms) must therefore have a large capacity and relatively small inductance to have reasonably high Q.

The effect of a load of given resistance on the Q of the circuit also can be changed by connecting the load across only part of the circuit. The most common method of accomplishing this is by tapping the load across part of the coil, as shown in Fig. 217-B. The smaller the portion of the coil across which the load is tapped the less the loading on the circuit; in other words, tapping the load "down" is equivalent to connecting a higher value of load resistance across the whole circuit. This is similar in principle to impedance transformation with an iron-core transformer (§ 2-9). However, in the high-frequency resonant circuit the impedance ratio does not vary exactly as the square of the turn ratio because all the magnetic flux lines do not cut every turn of the coil. A desired reflected impedance usually must be obtained by experimental adjustment.

L/C ratio — For a given frequency the product of L and C must always be the same, but it is evident that L can be large and C small, L small and C large, etc. The relation between the two for a fixed frequency is called the *L/C ratio*. A *high-C* circuit is one which has more capacity than "normal" for the frequency; a *low-C* circuit one which has less than normal capacity. These terms depend to a considerable extent upon the particular application considered, and have no exact numerical meaning.

Piezo-electricity — Properly-ground crystals of quartz and some other materials show a mechanical strain when subjected to an electric charge and, conversely, will show a difference in potential between two faces when subjected to mechanical stress. This characteristic is called the *piezo-electric* effect. A properly-ground quartz crystal is a mechanical vibrator electrically equivalent to a series-resonant circuit of very high Q, and can be used for many of the purposes for which ordinary resonant circuits are used.

●2-11 COUPLED CIRCUITS

Energy transfer; loading — Two circuits are said to be *coupled* when energy can be transferred from one to the other. The circuit delivering energy is called the primary circuit;

that receiving energy is called the secondary circuit. The energy may be practically all dissipated in the secondary circuit itself, as in receiver circuits, or the secondary may simply act as a medium through which the energy is transferred to a load resistance where it does work. In the latter case the coupled circuits

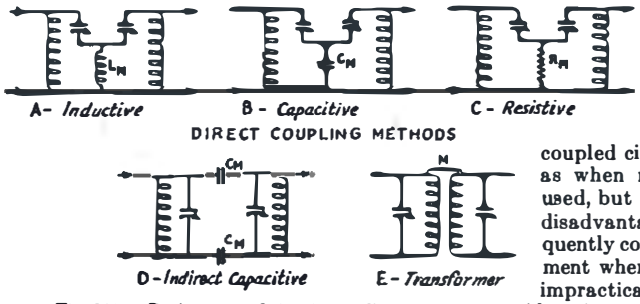


Fig. 218 — Basic types of circuit coupling.

may act as a radio-frequency impedance matching device (§ 2-9) where the matching may be accomplished by adjusting the loading on the secondary (§ 2-10) and by varying the coupling between the primary and secondary.

Coupling by a common circuit element — The three variations of this type of coupling (often called *direct coupling*) are shown at A, B and C of Fig. 218, utilizing common inductance, capacity, and resistance, respectively. Current circulating in one *LC* branch flows through the common element (L_m, C_m, or R_m) and the voltage developed across this element causes current to flow in the other *LC* branch. The degree of coupling between the two circuits is greater as the reactance (or resistance) of the common element is increased in comparison to the remaining reactances in the two branches.

The circuit at D shows electrostatic coupling between two resonant circuits. The coupling increases as the capacity of C_m is made greater (reactance of C_m is decreased).

Inductive coupling — Fig. 218-E illustrates inductive coupling, or coupling by means of the magnetic field. A circuit of this type resembles the iron-core transformer (§ 2-9) but because only a small percentage of the flux lines set up by one coil cut the turns of the other coil the simple relationships between turn ratio, voltage ratio, and impedance ratio in the iron-core transformer do not hold. The interlinking of the lines of force emanating from one coil with the turns of the other is measured by the *mutual inductance* between the two coils. Its value is determined by the self-inductance of each of the two coils and their position with respect to each other. The mutual inductance increases as the two coils are brought closer to each other with their axes coinciding.

Link coupling — A variation of inductive coupling called *link coupling* is shown in Fig. 219. This gives the effect of inductive coupling between two coils which may be so separated that they have no mutual inductance; the link may be considered simply as a means of providing the mutual inductance. Because mutual inductance between coil and link is involved at each end of the link, the total coupling between two link-coupled circuits cannot be made as great as when normal inductive coupling is used, but in practice this is usually not disadvantageous. Link coupling is frequently convenient in the design of equipment where inductive coupling would be impracticable because of constructional considerations.

The link coils generally have few turns compared to the resonant-circuit coils, since the coefficient of coupling (see next paragraph) is relatively independent of the number of turns on either coil.

Coefficient of coupling — The degree of coupling between two coils is a function of their mutual inductance and self-inductances:

$$k = \frac{M}{\sqrt{L_1 L_2}}$$

where k is called the *coefficient of coupling*. It is often expressed as a percentage. The coefficient

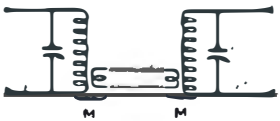


Fig. 219 — Link coupling. Mutual inductance at both ends of the link is equivalent to mutual inductance between the two tuned circuits.

of coupling cannot be greater than 1, and generally is much smaller in resonant circuits.

Critical coupling — When there is little coupling between two circuits tuned to the same frequency (*loose coupling*) each behaves much as though the other were not present. As the coupling is increased, each circuit loads the other because of the energy transfer, an effect which is equivalent to increasing the series resistance in each circuit (or reducing its parallel impedance). Hence the sharpness of resonance, or selectivity, is decreased. At *critical coupling*, maximum energy is transferred from one circuit to the other, and the overall resonance curve shows a single fairly broad peak. At still closer coupling (*tight coupling*) the energy transfer will drop off and

the overall resonance curve will show two peaks, one on either side of the frequency to which the circuits are tuned. The tighter the coupling the greater the frequency separation of the two peaks.

Critical coupling is a function of the *Q's* of the two circuits taken independently. A higher coefficient of coupling is required to reach critical coupling when the *Q's* are low; if the *Q's* are high, as in receiving applications, a coupling coefficient of a few percent may give critical coupling.

Effect of circuit Q — With loaded circuits it is not impossible for the *Q* to reach such low values that critical coupling cannot be obtained even with the highest practicable coefficient of coupling (coils as physically close as possible). In such case the only way to secure sufficient coupling is to increase the *Q* of one or both of the coupled circuits. This can be done either by decreasing the *L/C* ratio or by tapping the load down on the secondary coil (§ 2-10). One or the other of these methods often must be used in link coupling, because the maximum coefficient of coupling between two coils seldom runs higher than 50% or 60%, and the net coefficient is approximately equal to the products of the coefficients at each end of the link. If the load resistance is known beforehand, the circuits may be designed for a *Q* in the vicinity of 10 or so with assurance that sufficient coupling will be available; if unknown, the proper *Q's* can be determined by experiment.

Coupled resistance and reactance — If the two circuits are tuned to the same frequency, their effect on each other is resistive. For example, a loaded and resonant secondary will cause an apparent increase in the series resistance of the primary circuit (representing the energy dissipated in the load) which in turn causes the parallel impedance of the primary to decrease. It is by this means that the parallel impedance of the primary can, by adjustment of secondary loading and coupling, be adjusted to the optimum value for the device furnishing the power (§ 2-9).

Should the secondary circuit be slightly off tune it will have a reactive as well as resistive component, and the reactance is likewise coupled into the primary circuit. Since this in turn throws the primary off tune, the latter must be retuned to bring it back to resonance. The reflected reactance may be either inductive or capacitive. This effect occurs frequently in transmitters, where with certain types of coupling (link coupling, for instance) there may be a small amount of residual reactance in the secondary circuit.

Shielding — It is frequently necessary to *prevent* coupling between two circuits which for constructional reasons must be physically near each other. Capacitive coupling may readily be prevented by enclosing one or both of the circuits in grounded low-resistance metallic containers called *shields*. The electrostatic field from the circuit components does not penetrate the shield because the lines of force are short-circuited (§ 2-3). In many cases a metallic plate, called a *baffle shield*, inserted between two components may suffice to prevent electrostatic coupling between them, since very little of the field tends to bend around such a shield if it is large enough to make the components invisible to each other.

Similar metallic shielding is used at radio frequencies to prevent magnetic coupling. In this case the magnetic field induces a current (*eddy current*) in the shield which in turn sets up its own magnetic field opposing the original field (§ 2-5). The induced current is proportional to the frequency and also to the conductivity of the shield, hence the shielding effect increases with frequency and the conductivity and thickness of the shielding material. A closed shield is required for good magnetic shielding; in some cases separate shields, one about each coil, may be required. The baffle

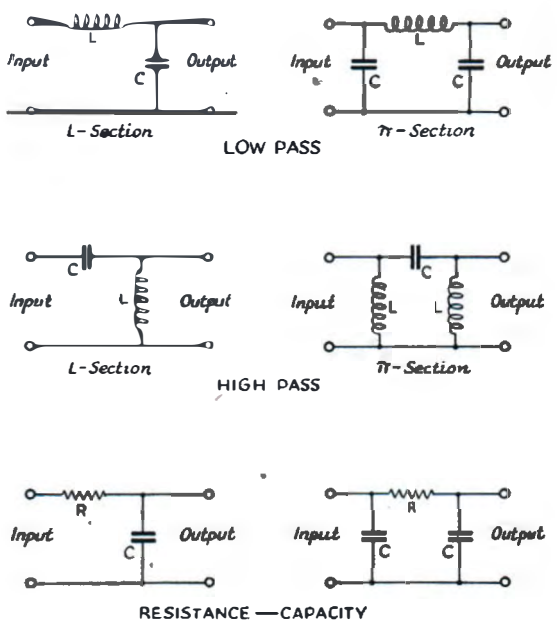


Fig. 220 — Simple filter circuits.

shield is rather ineffective for magnetic shielding, although it will give partial shielding if placed at right angles to the axes of, as well as between, the two coils to be shielded from each other.

Cancellation of part of the field of the coil reduces its inductance, and since some energy is dissipated in the shield, the effective resistance of the coil is raised as well, hence the coil Q is reduced. The effect of shielding on coil Q and inductance becomes less as the distance between the coil and shield is increased. The losses also decrease with an increase in the conductivity of the shield material. Copper and aluminum are satisfactory materials. The Q and inductance will not be greatly reduced if the spacing between the sides of the coil and the shield is at least half the coil diameter, and is not less than the coil diameter at the ends of the coil.

At audio frequencies the shielding container is made of magnetic material, preferably of high permeability (§ 2-5) to short-circuit the external flux about the coil to be shielded. A non-magnetic shield is quite ineffective at these low frequencies because the induced current is small.

Filters — By suitable choice of circuit elements, a coupling system may be designed to pass without undue attenuation all frequencies below and reject all frequencies above a certain value called the *cut-off frequency*. Such a coupling system is called a *filter*, and in the above case is known as a *low-pass filter*. If frequencies above the cut-off frequency are passed and those below attenuated, the filter is a *high-pass filter*. Simple filter circuits of both types are shown in Fig. 220. The fundamental circuit from which more complex filters are constructed is the *L-section*. Fig. 220 also shows *π-section* filters, constructed from the basic L-section and frequently encountered in both low-frequency and r.f. circuits. The proportions of L and C for proper operation of the filters depend upon the load resistance connected across the output terminals, L being larger and C smaller as the load resistance is increased.

A *band-pass* filter is one designed to pass without attenuation all frequencies between two selected cut-off frequencies and to attenuate all frequencies outside these limits. The group of frequencies passed through the filter is called the *pass-band*. Two resonant circuits with greater than critical coupling represent a common form of band-pass filter.

The *resistance-capacity filter* shown in Fig. 220 is used where both d.c. and a.c. are flowing through the circuit and it is desired to provide greater attenuation for the alternating current than the direct current. It is usually employed where the direct current has a low value so that the d.c. voltage drop is not excessive, or when a d.c. voltage drop actually is required. The time constant (§ 2-6) of the filter must be large compared to the time of one cycle of the lowest frequency to be attenuated. In determining the time constant, the resistance of the load must be included as well as that in the filter itself.

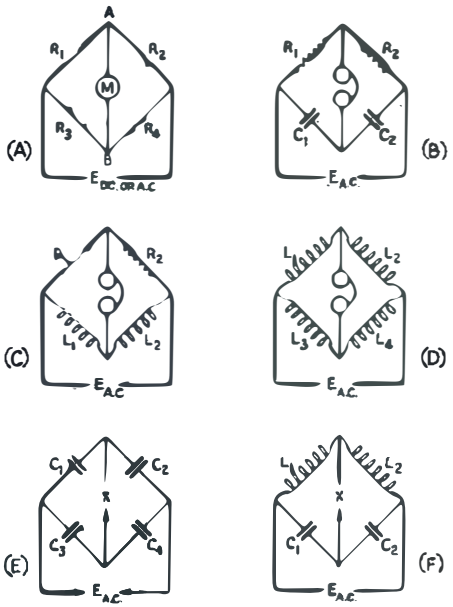


Fig. 221 — Bridge circuits utilizing resistance, inductance and capacity, alone and in combination.

Bridge circuits — A *bridge circuit* is a device primarily used in making measurements of resistance, reactance or impedance (§ 2-8), although it has other applications in radio circuits. The fundamental form of bridge is shown in Fig. 221-A. It consists of four resistances (called *arms*) connected in series-parallel to a source of voltage E, with a sensitive galvanometer M connected between the junctions of the series-connected pairs. When the equation

$$\frac{R_1}{R_2} = \frac{R_3}{R_4}$$

is satisfied no current flows through M because no potential difference exists between points A and B since the drop across R_2 equals that across R_4, and the drop across R_1 equals that across R_3. Under these conditions the bridge is said to be *balanced*. If R_3 is an unknown resistance and R_4 is a variable known resistance, R_3 can be found from the following equation, after R_4 has been adjusted to balance the bridge (*null* indication on M):

$$R_3 = \frac{R_1}{R_2} R_4$$

R_1 and R_2 are known as the *ratio arms* of the bridge; the ratio of their resistances is usually adjustable (frequently in steps of 1, 10, 100, etc.) so that a single variable resistor R_4 can serve as a standard for measuring widely different values of unknown resistance.

Bridges can be similarly formed with condensers, inductances, and combinations of resistance with either. Typical simple arrangements are shown in Fig. 221. For measurements with alternating currents the bridge must not introduce phase shifts which would destroy the balance, hence similar impedances should be used in each branch, as shown in Fig. 221, and the Q's of the coils and condensers should be the same. When bridges are used at audio frequencies a telephone headset is a good null indicator. The bridges at E and F are commonly-used r.f. neutralizing circuits (§ 4-7); the voltage from the source E_{ac} is balanced out at X.

• 2-12 LINEAR CIRCUITS

Standing waves — If an electrical impulse is started along a wire it will travel at approximately the speed of light until it reaches the end. If the end of the wire is open circuited, the impulse will be *reflected* at this point and travel back again. When a high-frequency alternating voltage is applied to the wire a current will flow toward the open end, and reflection will occur continuously. If the wire is long enough so that time comparable to a half cycle or more is required for current to travel to the open end, the phase relations between the reflected current and outgoing current will vary along the wire, and at one point the two currents will be 180 degrees out of phase and at another in phase, with intermediate values between. Assuming negligible losses, this means that the resultant current will vary in amplitude from zero to a maximum value along the wire. Such a variation is called a *standing wave*. The voltage along the wire also goes through standing waves, but reaches its maximum values where the current is minimum, and vice versa.

Frequency and Wavelength — It is possible to describe the constants of such line circuits in terms of inductance and capacitance, or inductance and capacitance per unit length, but it is more convenient to give them simply in terms of fundamental resonant frequency or of length. In the case of a straight-wire circuit, *length is inversely proportional to lowest resonant frequency.* Since the velocity is 300,000 kilometers (186,000 miles) per second, the *wavelength* is

$$\lambda = \frac{300{,}000}{f_{kc.}}$$

where λ is the wavelength in meters and $f_{kc.}$ is the frequency in kilocycles. The lowest frequency at which the wire or line will be resonant is known as its *fundamental* frequency or wavelength. It is common to describe lines (or antennas, which have similar current and voltage distribution) as *half-wave*, *quarter-wave*, etc., for a certain frequency ("half-wave 7000-kc. antenna," for instance).

Wavelength is also used interchangeably with frequency in describing not only antennas but also for tuned circuits, complete transmitters, receivers, etc. Thus the terms "high-frequency receiver" and "short-wave receiver," or "75-meter fundamental antenna" and "4000-kilocycle fundamental antenna," are synonymous.

Harmonic resonance — Although a coil-condenser combination having lumped constants (capacitance and inductance) resonates at only one frequency, circuits such as antennas containing distributed constants resonate readily at frequencies which are very nearly integral multiples of the fundamental frequency. These frequencies are therefore in *harmonic* relationship to the fundamental frequency and, hence, are referred to as *harmonics* (§ 2-7). In radio practice the fundamental itself is called the *first harmonic*, the frequency

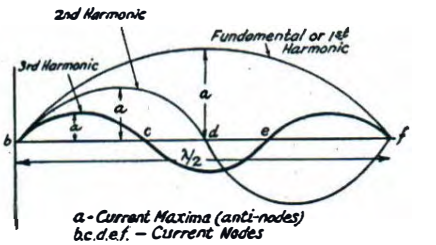


Fig. 222 — **Standing-wave current distribution on a wire operating as an oscillatory circuit at its fundamental, second harmonic and third harmonic frequencies.**

twice the fundamental is called the *second harmonic*, and so on.

Fig. 222 illustrates the distribution of current on a wire for fundamental, second and third harmonic excitation. There is one point of maximum current with fundamental operation, two when operation is at the second harmonic and three at the third harmonic; the number of current maxima corresponds to the order of the harmonic and the number of standing waves on the wire. As noted in the figure, the points of maximum current are called *anti-nodes* (also known as "loops") and the points of zero current are called *nodes.*

Radiation resistance — Since a line circuit has distributed inductance and capacity, current flow causes storage of energy in magnetic and electrostatic fields (§ 2-3, 2-5). At low frequencies practically all the energy so stored is returned to the wire during another part of the cycle (§ 2-8) but above 15,000 cycles or so (radio frequency) some escapes — is radiated — in the form of electromagnetic waves. Energy radiated by a line or antenna is equivalent to energy dissipated as in a resistor. The value of this equivalent resistance is known as *radiation resistance*.

Resonant line circuits — The effective resistance of a resonant straight wire such as an antenna is considerable, because of the power radiated. The resonance curve of such a straight-line circuit is quite broad; in other words, its Q is relatively low. However, by folding the line, as suggested by Fig. 223, the fields about the adjacent sections largely cancel each other and very little radiation takes place. The radiation resistance is greatly reduced and the line-type circuit can be made to have a very sharp resonance curve or high Q.

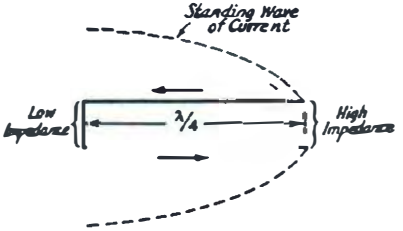


Fig. 223 — Standing wave and instantaneous current (arrows) conditions of a folded resonant-line circuit.

A circuit of this type will have a standing wave on it, as shown by the dash-line of Fig. 223, with the instantaneous current flow in each wire opposite in direction to the flow in the other, as indicated by the arrows on the diagram. This opposite current flow accounts for the cancellation of radiation, since the fields about the two wires oppose each other. Furthermore, the impedance across the open ends of the line will be very high, thousands of ohms, while the impedance across the line near the closed end will be very low.

A folded line may be made in the form of two concentric conductors, as shown in Fig. 224. The *concentric line* has even lower radiation resistance than the folded wire line, since the outer conductor acts as a shield. Standing waves exist, but are confined to the outside of the inner conductor and the inside of the outer conductor, since skin effect prevents the currents from penetrating to the other sides. Thus such a line will have no radio-frequency potentials on its exposed surfaces. Because of

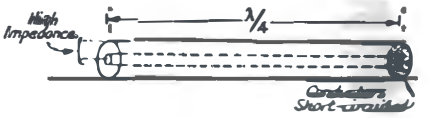


Fig. 224 — A concentric line resonant circuit.

the low radiation resistance and relatively large conducting surfaces, such lines can be made to have much higher Q's than are attainable with coils and condensers. They are most applicable at ultra-high frequencies (very short wavelengths) (§ 2-7) where dimensions are small.

• 2-13 CIRCUITS WITH SUPERIMPOSED CURRENTS

Combined a.c. and d.c. — There are many practical instances of simultaneous flow of alternating and direct current in a circuit. When this occurs there is a *pulsating* current and it is said that an alternating current is *superimposed* on a direct current. As shown in Fig. 225, the maximum value is equal to the d.c. value plus the a.c. maximum, while the minimum value (on the negative a.c. peak) is the difference between the d.c. and the maximum a.c. values. The average value (§ 2-7) of the current is simply equal to the direct-current component alone. The effective value (§ 2-7) of the combination is equal to the square root of the sum of the effective a.c. squared and the d.c. squared:

$$I = \sqrt{I_{ac}^2 + I_{dc}^2}$$

where I_{ac} is the effective value of the a.c. component, I is the effective value of the combination and I_{dc} is the average (d.c.) value of the combination.

Beats — If two or more alternating currents of different frequencies are present in a normal circuit, they have no particular effect upon one another and, for this reason, can be separated again at any time by the proper selective circuits. However, if two (or more) alternating currents of different frequencies are present in an element having unilateral or one-way current flow properties, not only will the two original frequencies be present in the output but also currents having frequencies equal to the sum, and difference, of the original fre-

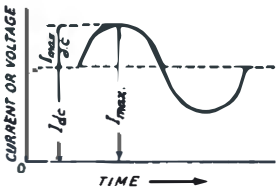


Fig. 225 — Pulsating current composed of alternating current superimposed on direct current.

quencies. These sum and difference frequencies are called the *beat* frequencies. For example, if frequencies of 2000 and 3000 kc. are present in a normal circuit, only those two frequencies exist, but if they are passed through a unilateral-element (such as a properly-adjusted vacuum tube) there will be present in the output not only the two original frequencies of 2000 and 3000 kc. but also currents of 1000 (3000 − 2000) and 5000 (3000 + 2000) kc. Suitable circuits can select the desired beat frequency.

By-passing — In combined circuits it is frequently necessary to provide a low-impedance path for a.c. around, for instance, a source of d.c. voltage. This can be done by using a *by-pass condenser*, which will not pass direct current but will readily permit the flow of alternating current. The capacity of the condenser should be of such value that its reactance is low (of the order of 1/10th or less) compared to the a.c. impedance of the device being by-passed. The lower the reactance, the better is the a.c. confined to the desired path.

Similarly, alternating current can be prevented from flowing through a direct-current circuit to which it may be connected by inserting an inductance of high reactance (called a *choke coil*) between the two circuits. This will permit the d.c. to flow without hindrance, since the resistance of the choke coil may be made quite low, but will effectively prevent the a.c. from flowing where it is not wanted.

If both r.f. and low-frequency (audio or power frequencies) currents are present in a circuit, they may be confined to desired paths by similar means, since an inductance of high reactance for radio frequency will have negligible reactance at low frequencies, while a condenser of low reactance at radio frequencies will have high reactance at low frequencies.

Grounds — The term "ground" is frequently met in discussions of circuits, and normally means the voltage reference point in the circuit. There may or may not be an actual connection to earth, but it is understood that a point in the circuit said to be at *ground potential* could be connected to earth without disturbing the operation of the circuit in any way. In direct-current circuits the negative side is generally grounded. The ground symbol in circuit diagrams is used for convenience in indicating common connections between various parts of the circuit, and with respect to actual ground usually has the meaning indicated above.

Vacuum Tubes

● 3-1 DIODES

Rectification — Practically all of the vacuum tubes used in radio work depend upon thermionic conduction (§ 2-4) for their operation. The simplest type of vacuum tube is that shown in Fig. 301. It has two elements, cathode and plate, and is called a *diode*. The cathode is heated by the "A" battery and emits electrons which flow to the plate when the plate is at a positive potential with respect to the cathode. Because of the nature of thermionic conduction, the tube is a conductor in one direction only. If a source of alternating voltage is connected between the cathode and plate, then electrons will flow only on the positive half-cycles of alternating voltage; there will be no electron flow during the half cycle when the plate is negative. Thus the tube can be used as a *rectifier*, to change alternating current to pulsating direct current. This alternating current can be anything from the 60-cycle kind to the highest radio frequencies.

Characteristic curves — The performance of the tube can be reduced to easily-understood terms by making use of *tube characteristic curves*. A typical characteristic curve for a diode is shown at the right in Fig. 301. It shows the current flowing between plate and cathode with different d.c. voltages applied between the elements. The curve of Fig. 301 shows that, with fixed cathode temperature, the plate current increases as the voltage between cathode and plate is raised. For an actual tube the values of plate current and plate voltage would be plotted along their respective axes.

The power consumed in the tube is the product of the plate voltage multiplied by the plate current, just as in any d.c. circuit. In a vacuum tube this power is dissipated in heat developed in the plate and radiated to the bulb.

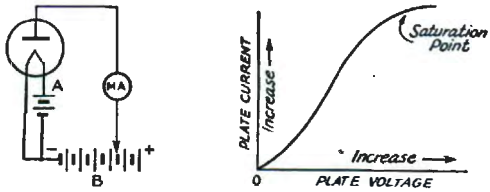


Fig. 301 — The diode or two-element tube and a typical characteristic curve.

Space charge — With the cathode temperature fixed, the total number of electrons emitted is always the same regardless of the plate voltage. Fig. 301 shows, however, that less plate current will flow at low plate voltages than when the plate voltage is large. With low plate voltage only those electrons nearest the plate are attracted to the plate. The electrons in the space near the cathode, being themselves negatively charged, tend to repel the similarly-charged electrons leaving the cathode surface and cause them to fall back on the cathode. This is called the *space charge* effect. As the plate voltage is raised, more and more electrons are attracted to the plate until finally the space charge effect is completely overcome and all the electrons emitted by the cathode are attracted to the plate, and a further increase in plate voltage can cause no increase in plate current. This is called the *saturation point*.

● 3-2 TRIODES

Grid control — If a third element, called the *control grid*, or simply the *grid*, is inserted between the cathode and plate of the diode the space-charge effect can be controlled. The tube then becomes a *triode* (three-element tube) and is useful for more things than rectification. The grid is usually in the form of an open spiral or mesh of fine wire. With the grid connected externally to the cathode and with a steady voltage from a d.c. supply applied between the cathode and plate (the positive of the "B" supply is always connected to the plate), there will be a constant flow of electrons from cathode to plate, through the openings of the grid, much as in the diode. But if a source of variable voltage is connected between the grid and cathode there will be a variation in the flow of electrons from cathode to plate (a variation in plate current) as the voltage on the grid changes about a mean value. When the grid is made less negative (more positive) with respect to the cathode, the space charge is partially neutralized and there will be an increase in plate current; when the grid is made more negative with respect to the cathode, the space charge is reinforced and there will be a decrease in plate current.

Amplification — The grid thus acts as a

valve to control the flow of plate current, and it is found that it has a much greater effect on plate current flow than does the plate voltage; that is, a small change in grid voltage is just as effective as a large change in plate voltage in bringing about a change in plate current. When a resistance or impedance (*load*) is connected in series in the plate circuit, the voltage drop across it, which is a function of the plate current through it, can therefore be changed by varying the grid voltage as well as by giving the plate voltage a new value. Thus a small change in grid voltage will cause a large change in voltage drop across the impedance; in other words, the grid voltage is *amplified* in the plate circuit.

So long as the grid has a negative potential with respect to the cathode, electrons emitted by the cathode are repelled (§ 2-3) with the result that no current flows to the grid. Hence under these conditions the grid consumes no power. However, when the grid becomes positive with respect to the cathode, electrons are attracted to it and a current flows to the grid; when this *grid current* flows power is dissipated in the grid circuit.

Characteristics — The measure of the amplification of which a tube is capable is known as its *amplification factor*, designated by μ (mu). Mu is the ratio of plate-voltage change required for a given change in plate current to the grid-voltage change necessary to produce the same change in plate current. Another important characteristic is the *plate resistance*, designated r_p. It is the ratio, for a fixed grid voltage, of a small plate voltage change to the plate current change it effects. It is expressed in *ohms*. Still another important characteristic used in describing the properties of a tube is *grid-plate transconductance*, or *mutual conductance*, designated by g_m and defined as the rate of change of plate current with respect to a change in grid voltage. The mutual conductance is a rough indication of the design merit of the tube. It is expressed in *micromhos* (the *mho* is the unit of conductance and is equal to $1/R$) and is the ratio of amplification factor to plate resistance, multiplied by one million. These tube characteristics are interrelated and are dependent primarily on the tube structure.

Static and dynamic curves — The operation of a vacuum tube amplifier is graphically represented in elementary form in Fig. 302. The sloping line represents the variation in plate current obtained at a constant plate voltage with grid voltages ranging from a value sufficiently negative to reduce the plate current to zero, to a value slightly positive. Grid voltage is specified with reference to the cathode or filament. Notable facts about this curve are that it is essentially a straight line (is *linear*)

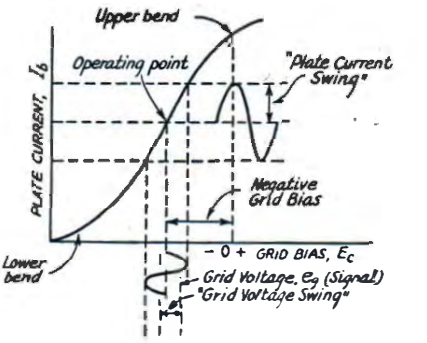


Fig. 302 — **Operating characteristics of a vacuum-tube amplifier. Class-A amplifier operation is depicted.**

over the middle section and that it bends towards the bottom (near *cut-off*) and near the top (*saturation*). In other words, the variation in plate current is directly proportional to the variation in grid voltage over the region between the two bends. With a fixed grid voltage (*bias*) of proper value the plate current can be set at any desired value.

Tube characteristics of the type shown in Fig. 302 may be of either the *static* or *dynamic* type. Static characteristics show the plate current that will flow at specific grid and plate voltages in the absence of any output device in the plate circuit for transferring the plate current variation to an external circuit, while the dynamic characteristic shows the behavior of the same quantities when there is a load in the plate circuit, and thus represents the actual operation of the tube as an amplifier.

Interelectrode capacities — Any pair of elements in a tube forms a miniature condenser (§ 2-3), and although the capacities of these condensers may be only a few micromicrofarads or less, they must frequently be taken into account in vacuum-tube circuits. The capacity from grid to plate (*grid-plate capacity*) has an important effect in many applications. In triodes, the other capacities are the *grid-cathode* and *plate-cathode*. In multi-element tubes (§ 3-5) similar capacities exist between these and other electrodes. With screen-grid tubes, the terms "input" and "output" capacity mean, respectively, the capacity measured from grid to all other elements connected together, and from plate to all other elements connected together. The same terms are used with triodes but are not so easily defined since the effective capacities existing depend upon the operating conditions (§ 3-3).

Tube ratings — Specifications of suitable operating voltages and currents are called *tube ratings*. Ratings include proper values for filament or heater voltage and current, plate voltage and current, and similar operating

specifications for other elements. An important rating in power tubes is the *maximum safe plate dissipation*, which is the maximum power which can be dissipated continuously in heat on the plate (§ 3-1).

•3-3 AMPLIFICATION

Circuits — Besides the vacuum tube, a complete amplifier includes a means for introducing the signal or *exciting voltage* into the grid circuit, a means (the *load*) for taking power or amplified voltage from the plate circuit, and sources of supply for bias voltage, power for heating the cathode, and d.c. power for the plate circuit. A representative circuit for audio-frequency amplification is shown in Fig. 303. The signal is introduced into the grid circuit in series with the bias voltage by means of transformer T_1, and T_2 serves as a means for transferring the amplified signal from the plate circuit to the load. Battery supplies are indicated for simplicity.

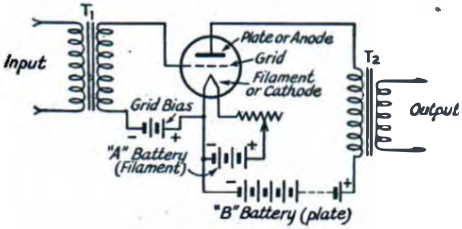


Fig. 303 — A typical audio-frequency amplifier using a triode tube.

A single amplifier such as is shown in Fig. 303 is called an *amplifier stage*, and several such stages may be used in *cascade*, the output of one stage being fed to the grid circuit of the next, to provide large amounts of overall amplification.

Load impedance — The load connected in the output or plate circuit of the tube is called the *load impedance* or *load resistance*, and designated R_p. It is the device in which the power output of the tube is consumed. In some types of amplifiers the load is an actual resistance, but in most cases it is a resistive impedance; that is, it shows resistance for a.c. but not for d.c. This type of load can be obtained with a resonant circuit (§ 2-10) or by coupling through a transformer to a power-consuming device (§ 2-9). The impedance load has the advantage that there is no drop in d.c. plate voltage across the load as there would be in the case of the resistor, since the latter has the same resistance for d.c. as for a.c.

In general, there will be one value of R_p which will give optimum results for a given type of tube and set of operating voltages; its value also depends upon the type of service for which the amplifier is designed. If the impedance of the actual device used is considerably different from the optimum load impedance, the tube and output device can be coupled through a transformer having a turns ratio such that the impedance reflected into the plate circuit of the tube is the optimum value (§ 2-9).

Operating point and grid bias — As indicated in Fig. 302, the relationship between varying grid voltage and plate current will be determined by the grid bias (§ 3-2), which sets the *operating point* on the characteristic curve. The choice of operating point depends upon the type of service in which the tube is to be used.

Distortion — With negative grid bias as shown in Fig. 302 the operating point comes in the middle of the linear region. If an alternating voltage (*signal*) is now applied to the grid in series with the grid bias, the grid voltage swings more and less negative about the mean bias voltage value and the plate current swings up (positive) and down (negative) about the mean plate current value. This is equivalent to an alternating current superimposed on the steady plate current. At this operating point it is evident that the plate current wave shapes (§ 2-7) are identical reproductions of the grid voltage wave shapes and will remain so as long as the grid voltage amplitude does not reach values sufficient to run into the lower- or upper-bend regions of the curve. If this occurs the output waves will be flattened or *distorted*. If the operating point is set towards the bottom or the top of the curve there will also be distortion of the output wave shapes because part or all of the lower or upper half-cycles will be cut off.

Whenever the bias is adjusted so that the tube works over a non-linear portion of its characteristic curve, distortion will take place and the output wave-form will not duplicate the wave-form of the voltage introduced at the grid. This characteristic of *non-linearity* of an amplifier is useful in some applications but is an undesirable feature in others. The distortion will take the form of harmonics added to the original wave (§ 2-7). If the exciting signal is a sine wave, the output wave, when distortion is present, will consist of the fundamental plus harmonics.

Another type of distortion, known as *frequency distortion*, occurs when the amplification varies with the frequency of the a.c. voltage applied to the grid circuit of the amplifier. It is not necessarily accompanied by harmonic distortion. It can be shown by a *frequency response curve*, or graph in which the relative amplification is plotted against frequency over the frequency range of interest.

Voltage amplification — The ratio of the alternating output voltage derived from the plate circuit to the alternating voltage applied to the grid circuit is called the *voltage amplification* of the amplifier. A *voltage amplifier* is one in which this ratio is the primary consideration, rather than the power which may be taken from the output circuit. The load resistance for voltage amplification must be high to give a large voltage across its terminals.

Power amplification — The ratio of output power to a.c. power consumed in the grid circuit (*driving power*) is called the *power amplification* of the amplifier. A *power amplifier* is one designed to deliver power to a load circuit, the voltage amplification being incidental. The power amplification ratio may be practically infinite in certain types of amplifiers. The load impedance for power amplification is selected either to give maximum power with minimum distortion or to give a desired value of plate efficiency.

Plate efficiency — The ratio of output power to d.c. input power to the plate (plate current multiplied by plate voltage) is called the *plate efficiency* of the amplifier. Plate efficiency is generally low in amplifiers designed primarily for minimum distortion, but may be made quite high when distortion is permissible.

Power sensitivity — This is the ratio of output power to alternating grid voltage, and is ordinarily used in connection with amplifiers operating in such a way that no power is consumed in the grid circuit. The same term also is used frequently in connection with radio-frequency power amplifiers, but in this case has the same meaning as power amplification ratio, defined above.

Phase relations in plate and grid circuits — When the exciting voltage on the grid has its maximum positive instantaneous value the plate current also is maximum (§ 3-2) so that the voltage drop across the impedance connected in the plate circuit likewise has its greatest value. The actual instantaneous voltage between plate and cathode is therefore minimum at the same instant, because it is equal to the d.c. supply voltage (which is unvarying) minus the voltage drop across the load impedance. Since the decrease in instantaneous plate voltage is negative in sense, this means that the alternating plate voltage is 180 degrees out of phase with the alternating grid voltage. Thus there is a *phase reversal* through an amplifier tube.

Input capacity — When an alternating voltage is applied between grid and cathode of the amplifier tube an alternating current flows in the small condenser formed by these elements (§ 3-2), just as it would in any other condenser (§ 2-8). Similarly, an alternating current also flows in the condenser formed by the grid and plate, but since the instantaneous voltage between these two elements is considerably larger than the signal voltage when the tube is amplifying, the current in the grid-plate capacity is likewise larger than it would be were no amplification taking place. Looked at from the grid circuit, the increased current is equivalent to an increase in input capacity of the tube, and the *effective* input capacity may be many times that which would be expected from consideration of the interelectrode capacities alone. The effective input capacity is proportional to the actual grid-plate capacity and to the voltage amplification.

Feedback — Some of the amplified energy in the plate circuit can be coupled back into the grid circuit to be re-amplified, this process being called *feedback*. If the voltage induced in the grid circuit is in phase with the grid signal voltage, the feedback is called *positive*, and the resultant voltage is larger and hence the amplification is increased. Positive feedback, usually called *regeneration*, can effectively increase the amplification of a stage many times. If the fed-back voltage is opposite in phase to the exciting voltage, the feedback is called *negative* and, since the resultant grid voltage is smaller, the amplification is decreased. Negative feedback is sometimes called *degeneration*.

Positive feedback is accompanied by a tendency to give maximum amplification at only one frequency, even though the input and output circuits may not otherwise be resonant. It therefore increases the selectivity of the amplifier and hence is used chiefly where high gain and sharpness of resonance are both wanted.

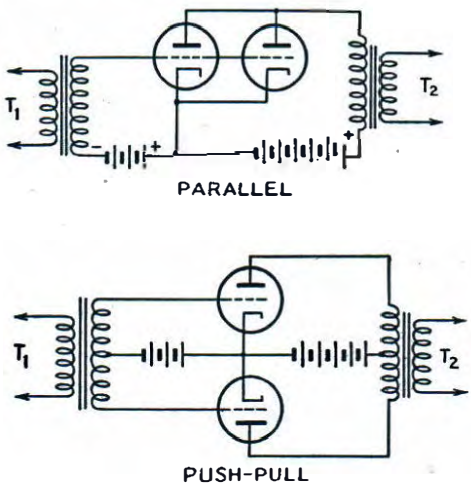


Fig. 304 — Parallel and push-pull amplifier connections.

Negative feedback has the opposite characteristics. It tends to widen the frequency range of the amplifier, even with resonant input and output circuits. It also reduces distortion and makes the amplifier tube more tolerant of changes in load impedance. Hence it is used where low-distortion, wide frequency range amplification is wanted, as in some audio circuits, even though amplification is sacrificed.

Parallel operation — When it is necessary to obtain more power output than one tube is capable of giving, without going to a larger tube structure, two or more tubes may be connected in *parallel*, in which case the similar elements in all tubes are connected together as shown in Fig. 304. The power output will then be in proportion to the number of tubes used; the exciting voltage required, however, is the same as for one tube.

If the amplifier operates in such a way as to consume power in the grid circuit, the grid power required also is in proportion to the number of tubes used.

Push-pull operation — An increase in power output can be secured by connecting two tubes in *push-pull*, the grids and plates of the two tubes being connected to opposite ends of the circuit as shown in Fig. 304. A "balanced" circuit, in which the cathode returns are made to the midpoint of the input and output devices, is necessary with push-pull operation. At any instant the ends of the secondary winding of the input transformer, T_1, will be at opposite potentials with respect to the cathode connection, so the grid of one tube is swung positive at the same instant that the grid of the other is swung negative. Hence, in any push-pull-connected stage the voltages and currents of one tube are out of phase with those of the other tube. The plate current of one tube therefore is rising while the plate current of the other is falling, hence the name "push-pull." In push-pull operation the even-harmonic (second, fourth, etc.) distortion is cancelled in the symmetrical plate circuit, so that for the same output the distortion will be less than with parallel operation.

The exciting voltage measured between the two grids must be twice that required for one tube. If the grids consume power, the driving power for the stage is twice that taken by either tube alone.

The decibel — The ratio of the power levels at two points in a circuit such as an amplifier can be expressed in terms of a unit called the *decibel*, abbreviated *db*. The number of decibels is 10 times the logarithm of the power ratio, or

$$db. = 10 \log \frac{P_1}{P_2}$$

The decibel is a particularly useful unit because it is logarithmic, and thus corresponds to the response of the human ear to sounds of varying loudness. One decibel is approximately the power ratio required to make a just noticeable difference in sound intensity. Within wide limits, changing the power by a given ratio produces the same apparent change in loudness regardless of the power level; thus if the power is doubled the increase is 3 db., or three steps of intensity; if it is doubled again, the increase is again 3 db., or three further distinguishable steps. Successive amplications expressed in decibels can be added to obtain the overall amplification.

A power loss also can be expressed in decibels. A decrease in power is indicated by a minus sign (e.g., -7 db.), and increase in power by a plus sign (e.g., $+4$ db.). Negative and positive quantities can be added numerically. Zero db. indicates the reference power level, or a power ratio of 1.

Applications of amplification — The major uses of vacuum tube amplifiers in radio work are to amplify at audio and radio frequencies (§ 2-7). The audio-frequency amplifier is generally used to amplify without discrimination at all frequencies in a wide range (say from 100 to 3000 cycles for voice communication), and is therefore associated with non-resonant or untuned circuits which offer a uniform load over the desired range. The radio-frequency amplifier, on the other hand, is generally used to amplify selectively at a single radio frequency, or over a small band of frequencies at most, and is therefore associated with resonant circuits tunable to the desired frequency.

An audio-frequency amplifier may be considered a *broad-band amplifier*; most radio-frequency amplifiers are relatively narrow-band affairs.

In audio circuits, the power tube or output tube in the last stage usually is designed to deliver a considerable amount of audio power, while requiring but negligible power from the input or exciting signal. To get the alternating voltage (*grid swing*) required for the grid of such a tube voltage amplifiers are used, employing tubes of high μ which will greatly increase the voltage amplitude of the signal. Voltage amplifiers are used in the radio-frequency stages of receivers as well as in audio amplifiers; power amplifiers are used in r.f. stages of transmitters.

● 3-4 CLASSES OF AMPLIFIERS

Reason for classification — It is convenient to divide amplifiers into groups according to the work they are intended to perform, as related to the operating conditions necessary to accomplish the purpose. This makes identi-

fication easy and obviates the necessity for giving a detailed description of the operation when *specific* operating data are not required.

Class A — An amplifier operated as shown in Fig. 302 in which the output wave shape is a faithful reproduction of the input wave shape, is known as a *Class-A* amplifier.

As generally used, the grid of a Class-A amplifier never is driven positive with respect to the cathode by the exciting signal, and never is driven so far negative that plate-current cut-off is reached. The plate current is constant both with and without grid excitation. The chief characteristics of the Class-A amplifier are low distortion, low power output for a given size of tube, and a high power-amplification ratio. The plate efficiency (§ 3-3) is relatively low, being in the vicinity of 20 to 35 percent at full output, depending upon the design of the tube and the operating conditions.

Class-A amplifiers of the power type find application as output amplifiers in audio systems. Class-A voltage amplifiers are found in the stages preceding the power stage in such applications, and as radio-frequency amplifiers in receivers.

Class B— The *Class-B* amplifier is primarily one in which the output current, or alternating component of the plate current, is proportional to the amplitude of the exciting grid voltage. Since power is proportional to the square of the current, the power output of a Class-B amplifier is proportional to the square of the exciting grid voltage.

The distinguishing operating condition in Class-B service is that the grid bias is set so that the plate current is relatively low without excitation; the exciting signal amplitude is such that the entire linear portion of the tube's characteristic is used. Fig. 305 illustrates Class-B operation with the tube biased practically to cut-off. In this operating condition plate current flows only during the positive half-cycle of excitation voltage. No plate current flows during the negative swing of the excitation voltage. The shape of the plate current pulse is essentially the same as that of the positive swing of the signal voltage. Since the plate current is driven up toward the saturation point, it is usually necessary for the grid to be driven positive with respect to the cathode during part of the grid swing. Grid current flows, therefore, and the driving source must furnish power to supply the grid losses.

Class-B amplifiers are characterized by medium power output, medium plate efficiency (50% to 60% at maximum signal) and a moderate ratio of power amplification. At radio frequencies they are used as *linear amplifiers* to raise the output power level in radiotelephone transmitters after modulation has taken place.

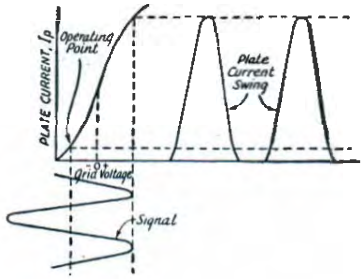


Fig. 305 — Operation of the Class-B amplifier.

For audio-frequency amplification two tubes must be used. The second tube, working alternately with the first, must be included so that both halves of the cycle will be present in the output. A typical method of arranging the tubes and circuit to this end is shown in Fig. 306. The signal is fed to a transformer T_1, whose secondary is divided into two equal parts, with the tube grids connected to the outer terminals and the grid bias fed in at the center. A transformer T_2 with a similarly-divided primary is connected to the plates of the tubes. When the signal swing in the upper half of T_1 is positive, Tube No. 1 draws plate current while Tube No. 2 is idle; when the lower half of T_1 becomes positive, Tube No. 2 draws plate current while Tube No. 1 is idle. The corresponding voltages induced in the halves of the primary of T_2 combine in the secondary to produce an amplified reproduction of the signal wave-shape with negligible distortion. The Class-B amplifier is capable of delivering much more power for a given tube size than a Class-A amplifier.

Class AB — The similarity between Fig. 306 and the ordinary push-pull amplifier circuit (§ 3-3) will be noted. Actually the circuits are the same, the difference being in the method of operation. If the bias is adjusted so that the tubes draw a moderate value of plate current the amplifier will operate Class A at low signal voltages and more nearly Class B at high signal voltages. An amplifier so operated is called *Class AB*. The advantages of this method are low distortion at moderate signal levels and higher efficiency at high levels, so that relatively small tubes can be used to good advantage in audio power amplifiers.

A further distinction can be made between amplifiers which draw grid current and those which do not. The *Class-AB*$_1$ amplifier draws no grid current and thus consumes no power from the driving source; the *Class-AB*$_2$ amplifier draws grid current at higher signal levels and power must therefore be supplied to its grid circuit.

Class C — The *Class-C* amplifier is one operated so that the alternating component of

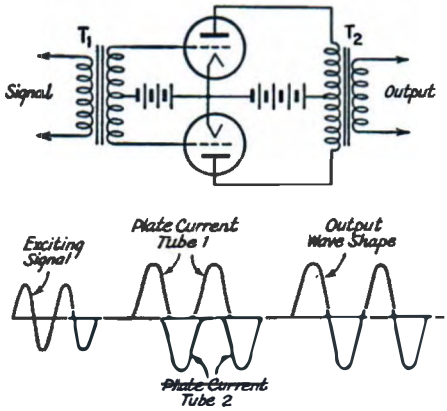


Fig. 306 — The Class-B audio amplifier, showing how the outputs of the two tubes are combined to give distortionless amplification.

the plate current is directly proportional to the plate voltage. The output power is therefore proportional to the square of the plate voltage. Other characteristics inherent to Class-C operation are high plate efficiency, high power output, and a relatively low power-amplification ratio.

The grid bias for a Class-C amplifier is ordinarily set at approximately twice the value required for plate current cut-off without grid excitation. As a result, plate current flows during only a fraction of the positive excitation cycle. The exciting signal should be of sufficient amplitude to drive the plate current to the saturation point, as shown in Fig. 307. Since the grid must be driven far into the positive region to cause saturation, considerable numbers of electrons are attracted to the grid at the peak of the cycle, robbing the plate of some that it would normally attract. This causes the droop at the upper bend of the characteristic, and also causes the plate current pulse to be indented at the top, as shown. Although the output wave-form is badly distorted, at radio frequencies the distortion is largely eliminated by the flywheel effect of the tuned output circuit.

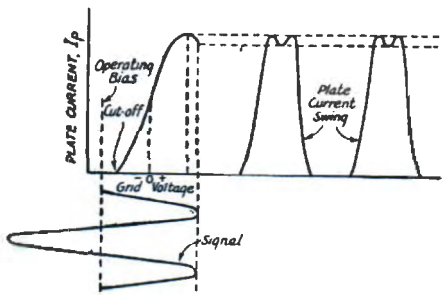


Fig. 307 — Class-C amplifier operation.

Although requiring considerable driving power because of the relatively large grid swing and grid-current flow, the high plate efficiency (ordinarily 70-75%) of the Class-C amplifier makes it an effective generator of radio-frequency power.

● 3-5 MULTIELEMENT AND SPECIAL-PURPOSE TUBES

Radio-frequency amplification — In a radio-frequency amplifier the input (grid) and output (plate) circuits must be tuned to the same frequency for maximum amplification and selectivity. If a triode tube is used in such an arrangement the feedback through the grid-plate capacity will sustain oscillation at radio frequencies (§ 3-7) so that the circuit becomes an oscillator rather than an amplifier. Although special circuits can be used to overcome oscillation, it is preferable to use a tube in which such feedback is negligible. Such a tube can be made by inserting a second grid to act as an electrostatic shield between the control grid and plate and thus reduce the grid-plate capacity to a negligible value. The addition of the extra grid, called the *screen grid* or screen, makes the tube a *tetrode*, or four-element tube.

The tetrode — The screen grid increases the amplification factor and plate resistance of the tube to values much higher than are attainable in triodes of practicable construction, although the mutual conductance is about the same as that of an equivalent triode. The screen grid is ordinarily operated at a lower positive potential than the plate, and is bypassed back to the cathode so that it has essentially the same a.c. potential as the cathode.

Another type of tetrode, in which the electrostatic shielding provided by the second grid is purely incidental, is built for audio power output work. The second grid accelerates the flow of electrons from cathode to plate, and the structure has a higher power sensitivity (§ 3-3) than is possible with triodes.

Secondary emission — When an electron travelling at appreciable velocity through a tube strikes the plate it dislodges other electrons which "splash" from the plate into the interelement space. This phenomenon is called *secondary emission*. In the triode, ordinarily operated with the grid negative with respect to cathode, these secondary electrons are repelled back into the plate and cause no disturbance. In the screen-grid tube, however, the positively charged screen grid attracts the secondary electrons, causing a reverse current to flow between screen and plate. The effect is particularly marked when the plate and screen potentials are nearly equal, which may be the

case during part of the a.c. cycle when the instantaneous plate current is large and the plate voltage low (§ 3-3).

The pentode — To overcome the effects of secondary emission a third grid, called the *suppressor grid*, can be inserted between the screen and plate. This grid is connected directly to the cathode and repels the relatively low-velocity secondary electrons back to the plate without obstructing to any appreciable extent the regular plate-current flow. Larger undistorted outputs therefore

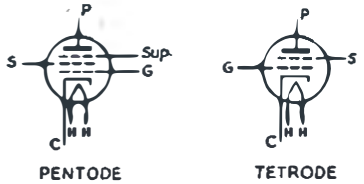


Fig. 308 — Symbols for pentode and tetrode tubes. H, heater; C, cathode; G, control grid; P, plate; S, screen grid; Sup., suppressor grid.

can be secured from the *pentode*, or five-element tube.

Pentode-type screen-grid tubes are used as radio-frequency voltage amplifiers, and in addition can be used as audio-frequency voltage amplifiers to give high voltage gain per stage. Pentode tubes also are suitable as audio-frequency power amplifiers, having greater plate efficiency and power sensitivity than triodes.

Beam tubes — A "beam" type tube is a tetrode incorporating a structure which forms the electrons travelling to the plate into concentrated paths, resulting in higher plate efficiency and power sensitivity. Suitable design also overcomes the effects of secondary emission without the necessity for a suppressor grid. Tubes constructed on the beam principle are used in receivers as both r.f. and audio amplifiers, and are built in larger sizes for transmitting circuits.

Variable-mu and sharp cut-off tubes — Receiving screen-grid tetrodes and pentodes for radio-frequency voltage amplification are made in two types, known as *sharp cut-off* and *variable-μ* or "super-control" types. In the sharp cut-off type the amplification factor is practically constant regardless of grid bias, while in the variable-μ type the amplification factor decreases as the negative bias is increased. The purpose of this design is to permit the tube to handle large signal voltages without distortion in circuits in which grid-bias control is used to vary the amplification.

Multipurpose types — A number of combination types of tubes have been constructed to perform multiple functions, particularly in receiver circuits. Among the simplest are full-wave rectifiers, combining two diodes in one envelope, and twin triodes, consisting of two triodes in one bulb for Class-B audio amplification. More complex types include duplex-diode triodes, duplex-diode pentodes, converters and mixers (for superheterodyne receivers), combination power tubes and rectifiers, and so on. In many cases the tube structure can be identified by the name, and all the types are basically the same as the simpler element combinations already described.

Mercury-vapor rectifiers — The power lost in a diode rectifier (§ 3-1) for a given plate current will be lessened if it is possible to decrease the plate-cathode voltage at which the current is obtained. If a small amount of mercury is put in the tube, the mercury will vaporize when the cathode is heated and, further, will ionize (§ 2-4) when plate voltage is applied. This neutralizes the space charge and reduces the plate-cathode voltage drop to a practically constant value of about 15 volts regardless of the value of plate current. Since this drop is much smaller than can be attained with purely thermionic conduction, there is less power loss in the rectifier. The constant voltage drop also is an advantage. Mercury-vapor tubes are widely used in power rectifiers.

Grid-control rectifiers — If a grid is inserted in a mercury-vapor rectifier it is found that with sufficient negative grid bias it is possible to prevent plate current from flowing if the bias is present before plate voltage is applied. However, if the bias is lowered to the point where plate current can flow, the mercury vapor will ionize and the grid loses control of plate current since the space charge is neutralized. It can assume control again only after the plate voltage is disconnected. The same phenomenon also occurs in triodes filled with other gases which ionize at low pressure. Grid-control rectifiers find considerable application in many circuits where "electronic switching" is desirable.

• 3-6 COMMON ELEMENTS IN VACUUM-TUBE CIRCUITS

Types of cathodes — Cathodes are of two types, directly and indirectly heated. Directly-heated cathodes or filaments used in receiving tubes are of the oxide-coated type, consisting of a wire or ribbon of tungsten coated with certain rare metals and earths which form an oxide capable of emitting large numbers of electrons with comparatively little cathode-heating power.

When directly-heated cathodes are operated on alternating current, the cyclic variation of current causes electrostatic and magnetic effects which vary the plate current of the tube at supply-frequency rate and thus produce

hum in the output. Hum from this source is eliminated by the indirectly-heated cathode, consisting of a thin metal sleeve or thimble, coated with electron-emitting material, enclosing a tungsten wire which acts as a heater. The heater brings the cathode thimble to the proper temperature to cause electron emission. This type of cathode is also known as the equipotential cathode, since all parts are at the same potential.

Methods of obtaining grid bias — Grid bias may be obtained from a source of voltage especially provided for that purpose, as a battery or other type of d.c. power supply. This is indicated in Fig. 309-A. A second method is shown at B, utilizing a *cathode resistor;* plate current flowing through the resistor causes a voltage drop which, with the connections shown, has the right polarity to bias the grid negatively with respect to the cathode. The value of the resistor is determined by the bias required and the plate current which flows at that value of bias, as found from the tube characteristic curves; with the voltage and current known, the resistance can be determined by Ohm's Law (§ 2-6):

$$R_c = \frac{E \times 1000}{I_c}$$

where R_c = cathode bias resistor in ohms
E = desired bias voltage
I_c = total d.c. cathode current in milliamperes.

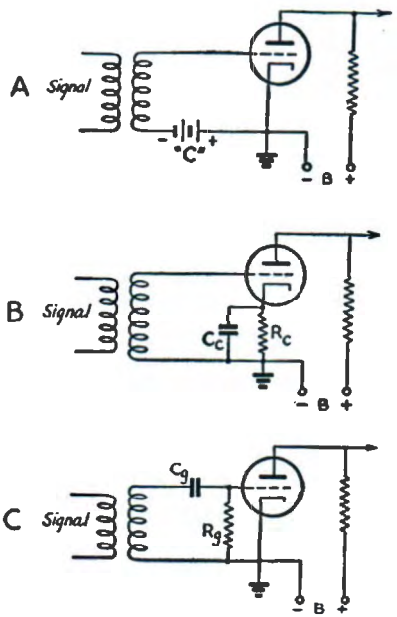


Fig. 309 — Methods of obtaining grid bias. A, fixed bias; B, cathode bias; C, grid-leak bias.

Screen- and suppressor-grid currents should be included with the plate current in multielement tubes to obtain the total cathode current, and also the control-grid current if the control grid is driven positive during operation. The a.c. component of plate current flowing through the cathode resistor will cause a voltage drop which gives negative feedback into the grid circuit (§ 3-3) so to prevent this the resistor usually is by-passed (§ 2-13), C_c being the *cathode by-pass condenser.*

A third method is by use of a *grid leak,* R_g in Fig. 309-C. This requires that the exciting voltage be positive with respect to the cathode during part of the cycle so that grid current will flow. The flow of grid current through the grid leak causes a voltage drop across the resistor which gives the grid a negative bias. The time constant (§ 2-6) of the grid leak and grid condenser should be large in comparison to the time of one cycle of the exciting voltage so that the grid bias will be substantially constant and will not follow the variations in a.c. grid voltage. For grid-leak bias,

$$R_g = \frac{E \times 1000}{I_g}$$

where R_g = grid-leak resistance in ohms
E = desired bias voltage
I_g = d.c. grid current in milliamperes.

When two tubes are operated in push-pull or parallel and use a common cathode- or grid-leak resistor, the value of resistance becomes one-half what it would be for one tube.

Cathode circuits; filament center tap — When a filament-type cathode is heated by a.c. the hum introduced can be minimized by making the two ends of the filament have equal and opposite potentials with respect to a center point, usually grounded (§ 2-13), to which the grid and plate return circuits are connected. The filament transformer winding is frequently *center-tapped* for this purpose, as shown in Fig. 310-A. The same result can be secured with an untapped winding by substituting a center-tapped resistor of 10 to 50 ohms as at B. The by-pass condensers, C_1 and C_2, are used in radio-frequency circuits to avoid having the r.f. current flow through the transformer or resistor, either of which may have considerable reactance at radio frequencies.

The filament supply for tubes with indirectly-heated cathodes sometimes is center-tapped for the same purpose; although frequently one side of the filament supply, and hence one terminal of the tube heater, is simply grounded.

• 3-7 OSCILLATORS

Self-oscillation — If in an amplifier with positive feedback the feedback or regeneration

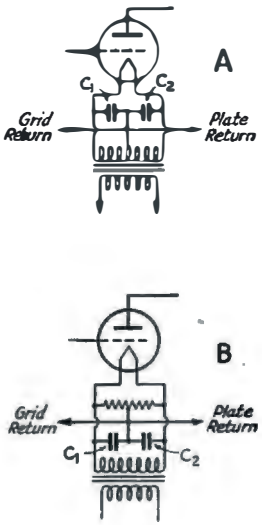


Fig. 310 — Filament center-tap connections.

(§ 3-3) is increased to a critical value, the tube will generate a continuous alternating current. This phenomenon, called *oscillation*, occurs when the power transferred between plate and grid circuits becomes large enough to overcome the circuit losses and the tube provides its own grid excitation. The power consumed is of course taken from the d.c. plate supply.

It is not necessary to apply external excitation to such a circuit, since any random variation in current, even though minute, will rapidly be amplified up to the proper value to cause oscillation. The frequency of oscillation will be that at which losses are least which, in the case of the resonant circuits usually associated with oscillators, is very nearly the resonant frequency of the circuit.

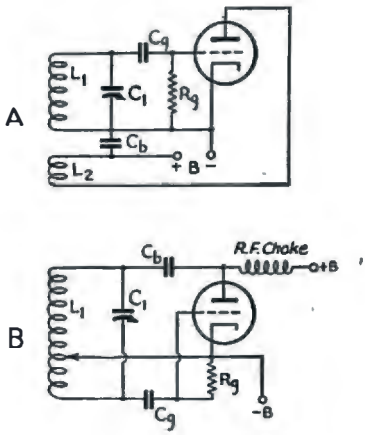


Fig. 311 — Oscillator circuits with magnetic feedback. A, tickler circuit; B, Hartley circuit.

Magnetic feedback — One form of feedback is by electromagnetic coupling between plate (output) and grid (input) circuits. Two representative circuits of this type are shown in Fig. 311. That at A is called the *tickler* circuit. The amplified current flowing in the "tickler," L_2, induces a voltage in L_1 in the proper phase when the coils are connected as shown and wound in the same direction. The feedback can be adjusted by adjusting the coupling between L_1 and L_2.

The *Hartley* circuit, B, is similar in principle. There is only one coil, but it is divided so that part of it is in the plate circuit and part in the grid circuit. The magnetic coupling between the two sections of the coil provides the feedback, which can be adjusted by moving the tap on the coil.

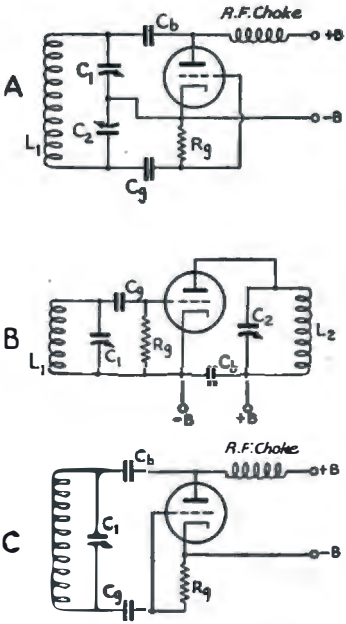


Fig. 312 — Oscillator circuits with capacity feedback. A, Colpitts circuit; B, tuned-plate tuned-grid circuit; C, ultraudion circuit.

Capacity feedback — The feedback can also be obtained through capacity coupling, as shown in Fig. 312. At A, the *Colpitts* circuit, the voltage across the resonant circuit is divided, by means of the series condensers, into two parts. The instantaneous voltages at the ends of the circuit are opposite in polarity with respect to the cathode, hence in the right phase to sustain oscillation.

The tuned-grid tuned-plate circuit at B utilizes the grid-plate capacity of the tube to provide feedback coupling. There should be no magnetic coupling between the two tuned-circuit coils. Feedback can be adjusted by varying the tuning of either the grid or plate circuit. The circuit with the higher Q (§ 2-10) determines the frequency of oscillation, although the two circuits must be tuned approximately to the same frequency for oscillations to occur.

The *ultraudion* circuit at C is equivalent to the Colpitts, with the voltage division for oscillation brought about through the grid-to-filament and plate-to-filament capacities of the tube. In this and in the Colpitts circuit the feedback can be controlled by varying the ratio of the two capacities. In the ultraudion circuit this can be done by connecting a small variable condenser between grid and cathode.

Crystal oscillators — Since a properly-cut quartz crystal is equivalent to a high-Q tuned circuit (§ 2-10) it may be substituted for a conventional circuit in an oscillator to control the frequency of oscillation. A simple crystal oscillator circuit is shown in Fig. 313. It will be

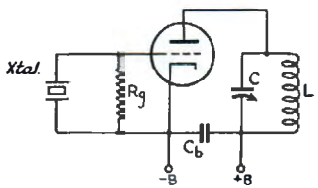


Fig. 313 — Simple crystal oscillator circuit.

recognized as the tuned-plate-tuned-grid circuit with the crystal substituted for the resonant circuit in the grid. Many variations of this fundamental circuit are used in practice.

Series and parallel feed — A circuit such as the tickler circuit of Fig. 311-A is said to be *series fed* because the source of plate voltage and the r.f. plate circuit (the tickler coil) are connected in series, hence the d.c. plate current flows through the coil to the plate. A by-pass (§ 2-13) condenser, C_b, must be connected across the plate supply to shunt the radio-frequency current around the source of power. Other examples of series plate feed are shown in Figs. 312-B and 313.

In some cases the source of plate power must be connected in parallel with the tuned circuit in order to provide a path for direct current to the plate. This is illustrated by the Hartley circuit of Fig. 311-B where it would be impossible to feed the plate current through the coil because there is a direct connection between the coil and cathode. Hence the voltage is applied to the plate through a radio-frequency choke which prevents the r.f. current from flowing to the plate supply and thus short-circuiting the oscillator. The blocking condenser C_b provides a low-impedance path for radio-frequency current flow but is an open circuit for direct current (§ 2-13). Other examples of parallel feed are shown in Figs. 312-A and 312-C.

Values of chokes, by-pass and blocking condensers are determined by the considerations outlined in § 2-13.

Excitation and bias — The excitation voltage required depends upon the characteristics of the tube and the losses in the circuit, including the power consumed in the load. In practically all oscillators the grid is driven positive during part of the cycle, so that power is consumed in the grid circuit (§ 3-2). This power must be supplied by the plate circuit. With insufficient excitation the tube will not oscillate; with too-high excitation the *grid losses*, or power consumed in the grid circuit, will be excessive.

Oscillators are almost always grid-leak biased (§ 3-6), which not only takes advantage of the grid-current flow but also gives better operation since the bias adjusts itself to the excitation voltage available.

Tank circuit — The resonant circuit associated with the oscillator is generally called the *tank circuit*. This name derives from the storage of energy associated with a resonant circuit of reasonably high Q (§ 2-10). It is applied to any resonant circuit in transmitting applications, whether used in an oscillator or amplifier.

Power output — The *power output* of an oscillator is the useful a.c. power consumed in a load connected to the oscillator. The load may be coupled by any of the means described in § 2-11.

Plate efficiency — The *plate efficiency* (§ 3-3) of an oscillator depends upon the load resistance, excitation, and other operating conditions, and usually is in the vicinity of 50%. It is not as high as in the case of an amplifier, since the oscillator must supply its own grid losses, which are usually 10% to 20% of the output power.

Frequency stability — The frequency stability of an oscillator is its ability to maintain constant frequency in the presence of variable operating conditions. The more important factors which may cause a change in frequency are (1) plate voltage, (2) temperature, (3) loading, (4) mechanical variations of circuit elements. Plate-voltage variations will cause a corresponding instantaneous shift in frequency; this type of frequency shift is called *dynamic* instability. Temperature changes will

cause tube elements to expand or contract slightly, thus causing variations in the interelectrode capacities (§ 3-2), and since these are unavoidably part of the tuned circuit the frequency will change correspondingly. Temperature changes in the coil or condenser will change the inductance and capacity slightly, again causing a shift in the resonant frequency. Both these temperature effects are relatively slow in operation, and the frequency change caused by them is called *drift*. Load variations act in much the same way as plate voltage variations except when there is a temperature change in the load, when drift also may be present. Mechanical variations, usually caused by vibration, cause changes in inductance and/or capacity which in turn cause the frequency to "wobble" in step with the vibration.

Dynamic instability can be reduced by using a tuned circuit of high effective Q which means, since the tube and load represent a relatively low resistance in parallel with the circuit, that a low L/C ratio ("high-C") must be used (§ 2-10), and that the circuit should be lightly loaded. Dynamic stability also can be improved by using a high value of grid leak, which gives high grid bias and raises the effective resistance of the tube as seen by the tank circuit, and by using relatively high plate voltage and low plate current, which accomplishes the same result. Drift can be minimized by using low d.c. input (for the size of tube), by using coils of large wire to prevent undue temperature rise, and by providing good ventilation to carry off heat rapidly. A low L/C ratio in the tank circuit also helps because the interelectrode capacity variations have proportionately less effect on the frequency when shunted by a large condenser. Special temperature-compensated components also can be used. Mechanical instability can be prevented by using well-designed components and insulating the oscillator from mechanical vibration.

Negative-resistance oscillators — If a resonant circuit were completely free from losses a current once started would continue indefinitely; that is, sustained oscillations would occur. This condition can be simulated in practice by cancelling the actual resistance in the circuit by inserting an equal or greater amount of *negative resistance*. Negative resistance is exhibited by any device showing an increase of current when the applied voltage is decreased, or vice versa.

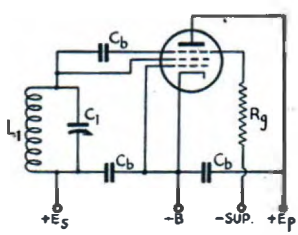


Fig. 314 — Negative-resistance oscillator. This circuit, known as the "transitron," requires that the screen be operated at a higher d.c. potential than the plate of the tube.

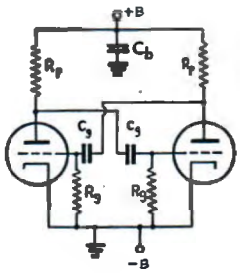


Fig. 315 — The multivibrator circuit, or relaxation oscillator.

The vacuum tube can be made to show negative resistance by a number of arrangements of electrode potentials. One circuit is shown in Fig. 314. Negative resistance is produced by virtue of the fact that as the suppressor grid of a pentode is given more negative bias, electrons normally passing through to the plate are turned back to the screen, thus increasing the screen current, and reversing normal tube action (§ 3-2). The negative resistance so produced is sufficiently low so that ordinary tuned circuits will oscillate readily at frequencies up to 15 Mc. or so.

The multivibrator — The type of oscillator circuit shown in Fig. 315 is known as the *multivibrator*, or *relaxation* oscillator. Two tubes are used with resistance coupling, the output of one tube being fed to the input circuit of the other. The frequency of oscillation is determined by the time constants (§ 2-6) of the resistance-capacity combinations. The principle of oscillation is the same as in the feedback circuits already described, the second tube being necessary to obtain the proper phase relationship (§ 3-3) for oscillation when the energy is fed back.

The multivibrator is a very unstable oscillator, and for this reason its frequency readily can be controlled by a small signal of steady frequency introduced into the circuit. This phenomenon is called *locking*. Its output waveshape is highly distorted, hence has high harmonic content (§ 2-7). A useful feature is that the multivibrator will lock with a frequency corresponding to one of its higher harmonics (the tenth harmonic is frequently used) and it can therefore be used as a *frequency divider*.

● 3-3 CATHODE-RAY TUBES

Principles — The cathode-ray tube is a vacuum tube in which the electrons emitted from a hot cathode are accelerated to give them considerable velocity, formed into a beam, and allowed to strike a special translucent screen which *fluoresces*, or gives off light at the point where the beam strikes. A narrow beam of moving electrons is similar to a wire carrying current (§ 2-4) and is accompanied by electrostatic and electromagnetic fields. Hence it can be deflected (have its direction changed) by application of external electrostatic or magnetic fields which exert a force on the beam in the same way as similar fields do on charged bodies or on wires carrying current (§ 2-3, 2-5). Since the beam consists only of moving electrons, its weight and inertia are negligibly small, hence it can be deflected easily and without any appreciable time lag. For this reason it can be made to follow instantly the variations in fields which are changing periodically at very high radio frequencies.

Electron gun — The electrode arrangement which forms the electrons into a beam is called the *electron gun*. In the simple tube structure shown in Fig. 316, the gun consists of the cathode, grid, and anodes Nos. 1 and 2. The intensity of the electron beam is regulated by the grid in the same way as in an ordinary tube (§ 3-2). Anode No. 1 is operated at a positive potential with respect to the cathode, thus accellerating the electrons which pass through the grid, and is provided with small apertures through which the electron stream passes and is concentrated into a narrow beam. This anode is also known as the *focusing electrode*. Anode No. 2 is operated at a high positive potential with respect to cathode and further increases the velocity of the electrons in the beam. The electron velocity and sharpness of the beam are determined by the relative voltages on the electrodes. In some tubes a second grid is inserted between the control grid and anode No. 1 to provide additional accelleration of the electrons.

Methods of deflection — The gun alone simply produces a small spot on the screen, but when the beam is deflected by either magnetic or electrostatic fields the spot moves across the screen in proportion to the force exerted. When the motion is sufficiently rapid, retentivity of vision makes the path of the moving spot (*trace*) appear to be a continuous line.

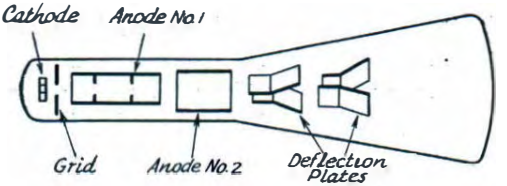


Fig. 316 — Arrangement of elements in the cathode-ray tube with electrostatic beam deflection.

Electrostatic deflection, generally used in the smaller tubes, is produced by *deflection plates*. Two sets of plates are placed at right angles to each other, as indicated in Fig. 316. The fields are created by applying suitable voltages between the two plates of each pair. Usually one plate of each pair is connected to anode No. 2 to establish the polarities (§ 2-3) of the fields with respect to the beam and to each other.

Tubes intended for magnetic deflection have the same type of gun, but have no deflection plates. Instead, the deflecting fields are set up by means of coils corresponding to the plates in tubes having electrostatic deflection. The coils are external to the tube but are mounted close to the glass envelope in the same relative positions occupied by the electrostatic deflection plates.

The beam deflection caused by a given change in the field intensity is called the *deflection sensitivity*. With electrostatic-deflection tubes it is usually expressed in millimeters per volt, which gives the linear movement of the spot on the screen as a function of the voltage applied to a set of deflecting plates. Values range from about 0.1 to 0.6 mm/volt, depending upon the tube construction and gun electrode voltages. The sensitivity is decreased by an increase in anode No. 2 voltage.

Fluorescent screens — The fluorescent screen materials used have varying characteristics according to the type of work for which the tube is intended. The spot color is usually green, white, yellow or blue, depending upon the screen material. The *persistence* of the screen is the time duration of the afterglow which exists when the excitation of the electron beam is removed. Screens are classified as long-, medium- and short-persistence. Small tubes for oscilloscope work are usually provided with medium-persistence screens having greenish fluorescence.

Tube circuits — A representative cathode-ray tube circuit with electrostatic deflection is shown in Fig. 317. One plate of each pair of deflecting plates is connected to anode No. 2. Since the voltages required are normally rather high, the positive terminal of the supply is usually grounded (§ 2-13) so that the common deflection plates will be at ground potential. This places the cathode and other elements at high potentials above ground, hence these elements must be well insulated. The various electrode voltages are obtained from a voltage divider (§ 2-6) across the high-voltage d.c. supply. R_3 is a variable divider or "potentiometer" for adjusting the negative bias on the control grid and thereby varying the beam cur-

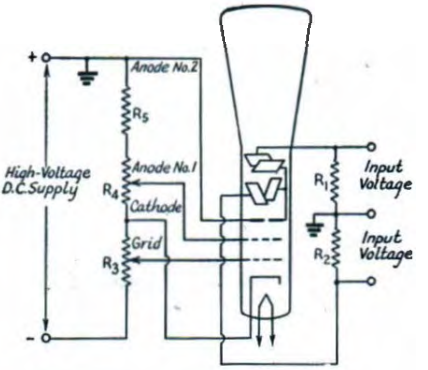


Fig. 317 — Cathode-ray tube circuit. Typical values for a three-inch (screen-diameter) tube such as the type 906:

R_1, R_2 — 1 to 10 megohms.
R_3 — 20,000-ohm potentiometer.
R_4 — 0.2-megohm potentiometer.
R_5 — 0.5 megohm.
The high-voltage supply should furnish about 1300 volts d.c.

rent; it is called the *intensity* or *brightness* control. The *focus*, or sharpness of the luminous spot formed on the screen by the beam, is controlled by R_4, which changes the ratio of anode No. 2 to anode No. 1 voltage. The focusing and intensity controls interlock to some extent, and the sharpest focus is obtained by keeping the beam current low.

Deflecting voltages for the plates are applied to the terminals marked "input voltage," R_1 and R_2 being high resistances (1 megohm or more) to drain off any accumulation of charge on the deflecting plates. Usually some provision is made to place an adjustable d.c. voltage on each set of plates so that the spot can be "centered" when stray electrostatic or magnetic fields are present; the adjustable voltage simply is set to neutralize such fields.

The tube is mounted so that one set of plates produces a horizontal line when a varying voltage is applied to it, while the other set of plates produces a vertical line under similar conditions. They are called, respectively, the "horizontal" and "vertical" plates, but which set of actual plates produces which line is simply a matter of how the tube is mounted. It is usually necessary to provide a mounting which can be rotated to some extent so that the lines will actually be horizontal and vertical.

Power supply — The d.c. voltage required for operation of the tube may vary from 500 volts for the miniature type (1-inch diameter screen) to several thousand for the larger tubes. The current, however, is very small, so that the power required is likewise small. Because of the small current requirements a rectified a.c. supply with half-wave rectification (§ 8-3) and a single 0.5 to 2-μfd. condenser as a filter (§ 8-5) is satisfactory.

● 3-9 THE OSCILLOSCOPE

Description — An oscilloscope is essentially a cathode-ray tube in the basic circuit of Fig. 317, but with provision for supplying a suitable deflection voltage on one set of plates, ordinarily those giving horizontal deflection. The deflection voltage is called the *sweep*. Oscilloscopes are frequently also equipped with vacuum-tube amplifiers for increasing the amplitude of small a.c. voltages to values suitable for application to the deflecting plates. These amplifiers are ordinarily limited to operation in the audio-frequency range, and hence cannot be used at radio frequencies.

Formation of patterns — When periodically-varying voltages are applied to the two sets of deflecting plates the path traced by the fluorescent spot forms a *pattern* which is stationary so long as the amplitude and phase relationships of the voltages remain unchanged. Fig. 318 shows how such patterns are formed. The horizontal sweep voltage is assumed to have the "sawtooth" waveshape indicated; with no voltage applied to the vertical plates, the trace would simply sweep from left to right across the screen along the horizontal axis X–X′ until the instant H is reached, when it reverses direction and returns to the starting point. The sine-wave voltage applied to the vertical plates would similarly trace a line along the axis Y–Y′ in the absence of any deflecting voltage on the horizontal plates. How-

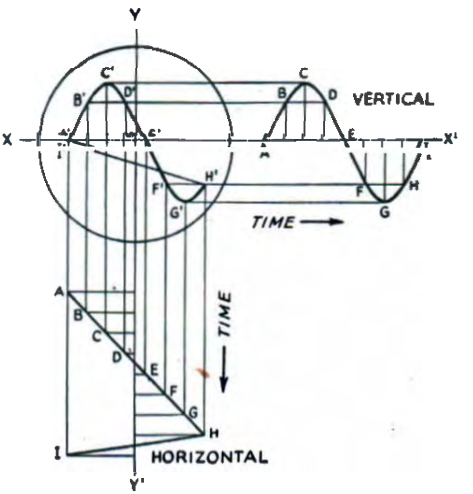


Fig. 318 — Showing the formation of the pattern from the horizontal and vertical sweep voltages.

ever, when both voltages are present the position of the spot at any instant depends upon the voltages on both sets of plates at that instant. Thus at time B the horizontal voltage has moved the spot a short distance to the right and the vertical voltage has similarly moved it upward, so that it reaches the actual position B′ on the screen. The resulting trace is easily followed from the other indicated positions, which are taken at equal time intervals.

Types of sweeps — A horizontal sweep-voltage waveshape such as that shown in Fig. 318 is called a *linear sweep*, because the deflection in the horizontal direction is directly proportional to time. If the sweep were perfect, the "fly-back" time, or time taken for the spot to return from the end (H) to the beginning (I or A) of the horizontal trace would be zero, so that the line HI would be perpendicular to the axis Y–Y′. Although the fly-back time cannot be made zero in practicable sweep-voltage generators, it can be made quite small in comparison to the time of the desired trace AH, at least at most frequencies within the audio range. The fly-back time is somewhat exaggerated in Fig. 318 to show its effect on the pattern. The line H′I′ is called the *return trace;* with a linear sweep it is less brilliant than the pattern because the spot is moving much more rapidly during the fly-back time than during the time of the main trace. If the fly-back time is short enough the return trace will be invisible.

The linear sweep has the advantage that it shows the shape of the wave applied to the vertical plates in the same way in which it is usually represented graphically (§ 2-7). By making the sweep time equal to a multiple of the time of one cycle of the a.c. voltage applied to the vertical plates, several cycles of the vertical or signal voltage will appear in the pattern. The shape of only the last cycle to appear will be affected by the fly-back in such a case. Although the linear sweep is generally most useful, other waveshapes may be desirable for certain purposes. The shape of the pattern obviously will depend upon the shape of the horizontal sweep voltage. If the horizontal sweep is sinusoidal, the main and return sweeps each occupy the same time, and the spot moves faster horizontally in the center of the pattern than it does at the ends. If two sinusoidal voltages of the same frequency are applied to both sets of plates, the resulting pattern may be a straight line, an ellipse or a circle, depending upon the amplitude and phase relationships. If the frequencies are harmonically related (§ 2-7) a stationary pattern will result, but if one frequency is not an exact harmonic of the other the pattern will show continuous motion. This is also the case when a linear sweep circuit is used; the sweep frequency and the frequency under observation must be harmonically related or the pattern will not be stationary.

Sweep circuits — A sinusoidal sweep is easiest to obtain, since it is possible to apply a.c. voltage from the power line directly or through a suitable transformer to the horizontal plates. A variable voltage divider can be used to regulate the width of the horizontal trace.

A typical circuit for a linear sweep is shown in Fig. 319. The tube is a *gas triode* or grid-control rectifier (§ 3-5). The breakdown voltage, or plate voltage at which the tube ionizes and starts conducting, is determined by the grid bias. When plate voltage is applied, the voltage across C_1 rises, as it acquires a charge through R_1, until the breakdown voltage is reached, when the condenser discharges rapidly through the comparatively low plate-cathode resistance of the tube. When the voltage drops to a value too low to maintain plate-current flow, the ionization is extinguished and C_1 once more charges through R_1. If R_1 is large enough, the voltage across C_1 rises linearly with time up to the breakdown point. This voltage is used for the sweep, being coupled to the cathode-ray tube or to an amplifier through C_2. The fly-back is the time required for discharge through the tube, and to keep it small the resistance during discharge must be as low as possible.

To obtain a stationary pattern, the "sawtooth" frequency can be controlled by varying C_1 and R_1, and by introducing some of the voltage to be observed (on the vertical plates) into the grid circuit of the tube. This voltage "triggers" the tube into operation in synchronism with the signal frequency. Synchronizing will occur even though the signal frequency is a multiple of the sweep frequency,

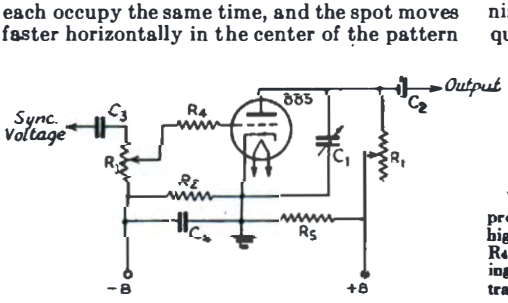


Fig. 319 — A linear sweep-oscillator circuit.

C_1 — 0.001 to 0.25 μfd.
C_2 — 0.5 μfd. C_3 — 0.1 μfd.
C_4 — 25 μfd., 25-volt electrolytic.
R_1 — 0.3 to 1.5 megohms.
R_2 — 2000 ohms. R_4 — 25,000 ohms.
R_3 — 0.25 megohm. R_5 — 0.1 megohm.

"B" supply should deliver 300 volts. C_1 and R_1 are proportioned to give suitable sweep frequency; the higher the time constant (§ 2-6) the lower the frequency. R_4 is a protective resistor to limit grid-current flow during the deionizing period, when positive ions are attracted to the negative grid.

provided the circuit constants and the amplitude of the synchronizing voltage are properly adjusted.

The voltage output of the type of circuit shown in Fig. 319 is limited because the charging rate of the condenser is linear only on that portion of the logarithmic charging curve (§ 2-6) which is practically a straight line. A linear charging rate over a longer period of time can be secured by substituting a current-limiting device, such as a properly-adjusted vacuum tube, for R_1.

Amplifiers — The usefulness of the oscilloscope is enhanced by providing amplifiers for both the horizontal and vertical sweep voltages, thereby insuring that sufficient voltage will be available at the deflection plates to give a pattern of suitable size. With small oscilloscope tubes (3-inch and smaller screens) the voltage required for a deflection of one inch varies from about 30 to 100 volts, depending upon the anode voltages, so that an amplifier tube capable of an undistorted peak output voltage of 100 or so is necessary. (With such an amplifier the voltage difference, or total voltage "swing", between the positive and negative peaks is 200 volts.) A resistance-coupled voltage amplifier (§ 3-3) having a pentode tube is ordinarily used because of the high stage gain obtainable, and the amplifier should be designed to have good frequency response over as wide a range of audio frequencies as possible (§ 5-9). Since a voltage gain of 100 to 150 or more is readily obtainable, full deflection of the beam can be secured with an input of one volt or less with such an amplifier.

Constructional considerations — An oscilloscope should be housed in a metal cabinet, both to shield the tube from stray electromagnetic and electrostatic fields which might deflect the beam, and also to protect the operator from the high voltages associated with the tube. It is good practice to provide an interlock switch which automatically disconnects the high-voltage supply when the cabinet is opened for servicing or other reasons.

CHAPTER FOUR

Radio-Frequency Power Generation

• 4-1 TRANSMITTER REQUIREMENTS

General Requirements — The power output of a transmitter must be as stable in frequency and as free from spurious radiations as the state of the art permits. The steady r.f. output, called the *carrier* (§ 5-1), must be free from amplitude variations attributable to ripple from the plate power supply (§ 8-4) or other causes, its frequency should be unaffected by variations in supply voltages or inadvertent changes in circuit constants, and there should be no radiations on other than the intended frequency. The degree to which these requirements can be met depends upon the operating frequency.

Design principles — The design of the transmitter depends on the output frequency, the required power output, and the type of operation (c.w. telegraphy or 'phone). For c.w. operation at low power on medium-high frequencies (up to 7 Mc. or so) a simple crystal oscillator circuit can meet the requirements satisfactorily. However, the stable power output which can be taken from an oscillator is limited, so that for higher power the oscillator is used simply as a frequency-controlling element, the power being raised to the desired level by means of amplifiers. The requisite frequency stability can be obtained only when the oscillator is operated on relatively low frequencies, so that for output frequencies up to about 60 Mc. it is necessary to increase the oscillator frequency by multiplication (harmonic generation — § 3-3), which is usually done at fairly low power levels and before the final amplification. An amplifier which delivers power on the frequency applied to its grid circuit is known as a *straight amplifier;* one which gives harmonic output is known as a *frequency multiplier.* An amplifier used principally to isolate the frequency-controlling oscillator from the effects of changes in load or other variations in following amplifier stages is called a *buffer amplifier.* A complete transmitter therefore may consist of an oscillator followed by one or more buffer amplifiers, frequency multipliers, and straight amplifiers, the number being determined by the output frequency and power in relation to the oscillator frequency and power. The last amplifier is called the *final amplifier,* and the stages up to the last comprise the *exciter.* Transmitters are usually designed to work in a number of frequency bands, so that means for changing the frequency of resonant circuits in harmonic steps usually is provided, generally by means of plug-in inductances.

The general method of designing a transmitter is to decide upon the power output and the highest output frequency required, and also the number of bands in which the transmitter is to operate. The latter usually will determine the oscillator frequency, since it is general practice to set the oscillator on the lowest frequency band to be used. The oscillator frequency is seldom higher than 7 Mc. except in some portable installations where tubes and power must be conserved. A suitable tube (or pair of tubes) should be selected for the final amplifier and the grid driving power required determined from the tube manufacturer's data. This sets the power required from the preceding stage. From this point the same process is followed back to the oscillator, including frequency multiplication wherever necessary. The selection of a suitable tube complement requires knowledge of the operating characteristics of the various types of amplifiers and oscillators. These are discussed in the following sections.

At 112 Mc. and above the ordinary methods of transmitter design become rather cumbersome, although it is possible to use them with proper choice of tubes and other components. However, in this ultra-high-frequency (§ 2-7) region the requirements imposed are less severe, since the limited transmission range (§ 9-5) mitigates the interference conditions that determine the requirements on the long-distance lower frequencies. Hence simple oscillator transmitters are widely used.

Vacuum tubes — The type of tube used in the transmitter has an important effect on the circuit design. Tubes of high power sensitivity (§ 3-3) such as pentodes and beam tetrodes, give larger power amplification ratios per stage than do triodes, hence fewer tubes and stages may be used to obtain the same output power. On the other hand, triodes have certain operating advantages such as simpler power supply circuits and relatively simpler adjustment for modulation (§ 5-3), and in addition are considerably less expensive for the same power output rating. Consequently it is usually more

economical to use triodes as output amplifiers even though an extra low-power amplifier stage may be necessary.

At frequencies in the region of 56 Mc. and above it is necessary to select tubes designed particularly for operation at ultra-high frequencies, since tubes built primarily for the lower frequencies may work poorly or not at all.

●4-2 SELF-CONTROLLED OSCILLATORS

Advantages and disadvantages — The chief advantage of a self-controlled oscillator is that the frequency of oscillation is determined by the constants of the tuned circuit, and hence readily can be set to any desired value. However, extreme care in design and adjustment are essential to secure satisfactory frequency stability (§ 3-7). Since frequency stability is generally poorer as the load on the oscillator is increased, the self-controlled oscillator should be used purely to control frequency and not for the purpose of obtaining appreciable power output, in transmitters intended for working below 60 Mc.

Oscillator circuits — The inherent stability of all of the oscillator circuits described in § 3-7 is about the same, since stability is more a function of choice of proper circuit values and adjustment than of the method by which feedback is obtained. However, some circuits are more convenient to use than others, particularly from the standpoint of feedback adjustment, mechanical considerations (whether the tuning condenser rotor plates can be grounded or not, etc.), and uniform output over a considerable frequency range. All simple circuits suffer from the fact that the power output must be taken from the frequency-determining tank circuit, so that aside from the effect of loading on frequency stability, the following amplifier stage also can react on the oscillator in such a way as to change the frequency.

The electron-coupled oscillator — The effects of loading and coupling to the next stage can be greatly reduced by use of the *electron-coupled* circuit, in which a screen-grid tube (§ 3-5) is so connected that its screen grid is used as a plate, in connection with the control grid and cathode, in an ordinary triode oscillator circuit. The screen is operated at ground r. f. potential (§ 2-13) to act as a shield between the actual plate and the cathode and control grid; the latter two elements must therefore be above ground potential. The output is taken from the plate circuit. Under these conditions the capacity coupling (§ 2-11) between the plate and other ungrounded tube elements is quite small, hence the output power is secured almost entirely by variations

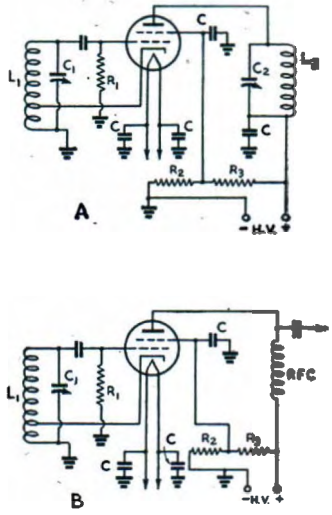


Fig. 401 — Electron-coupled oscillator circuit. For maximum stability the grid leak, R_1, should be 100,000 ohms or more. The grid condenser should be of the order of 100 $\mu\mu$fd., other fixed condensers from 0.002 μfd. to 0.1 μfd. Proper values for R_2 and R_3 may be determined from § 8-10. For maximum isolation between oscillator and output circuits, the tube should have extremely low grid-plate capacity.

in the plate current caused by the varying potentials on the grid and cathode. Since in a screen-grid tube the plate voltage has a relatively small effect on the plate current, the reaction on the oscillator frequency for different conditions of loading is small.

It is generally most convenient to use a Hartley (§ 3-7) circuit in the frequency-determining part of the oscillator. This is shown in Fig. 401, where L_1C_1 is the oscillator tank circuit. The screen is grounded for r.f. through a by-pass condenser (§ 2-13) but has the usual d.c. potential. The cathode connection is made to a tap on the tank coil to provide feedback. In the plate circuit a resonant circuit, L_2C_2, can be connected as shown at A; it may be tuned either to the oscillation frequency or to one of its harmonics. Untuned output coupling is shown at B; with this method the output voltage and power are considerably lower than with a tuned plate circuit, but better isolation between oscillator and amplifier is secured.

If the oscillator tube is a pentode with an external suppressor connection the suppressor grid should be grounded, not connected to cathode. This provides additional internal shielding and further isolates the plate from the frequency-determining circuit.

Factors influencing stability — The causes of frequency instability and the necessary remedial steps have been discussed in § 3-7.

These apply to all oscillators. In addition, in the electron-coupled oscillator the ratio of plate to screen voltage has an important effect on the stability with changes in supply voltage; the optimum ratio is generally of the order of 3:1 but should be determined experimentally for each case. Since the cathode is above ground potential, means should be taken to reduce the effects of heater-to-cathode capacitance or leakage, which by allowing a small a.c. voltage from the heater supply to develop between cathode and ground may cause modulation (§ 4-1) at the supply frequency. This effect, which is usually appreciable only at 14 Mc. and higher, may be reduced by by-passing of the heaters as indicated in Fig. 401 or by operating the heater at the same r.f. potential as the cathode. The latter may be accomplished by the wiring arrangement shown in Fig. 402.

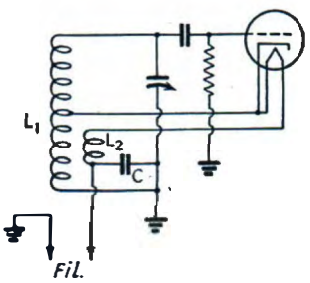


Fig. 402 — Method of operating the heater at cathode r.f. potential in an electron-coupled oscillator. L_2 should have the same number of turns as the part of L_1 between ground and the cathode tap, and should be closely coupled to L_1 (preferably interwound). By-pass condenser C should be 0.01 to 0.1 μfd.

Tank circuit Q — The most important single factor in determining frequency stability is the Q of the oscillator tank circuit. The effective Q must be as high as possible for best stability. Since oscillation is accompanied by grid-current flow, the grid-cathode circuit constitutes a resistance of load of appreciable proportions, the effective resistance being low enough to be the determining factor in establishing the effective parallel impedance of the tank circuit. Consequently, if the ends of the tank are connected to plate and grid, as is usual, a high effective Q can be obtained only by decreasing the L/C ratio and making the inherent resistance in the tank as low as possible. The tank resistance can be decreased by using low-loss insulation on condensers and coils, and by winding the coil with large wire. With ordinary construction the optimum tank capacity is of the order of 500 to 1000 μμfd. at a frequency of 3.5 Mc.

The effective circuit Q can be raised by increasing the resistance of the grid circuit and thus decreasing the loading. This can be accomplished by reducing the oscillator grid current, by using minimum feedback to maintain stable oscillation and by using a high value of grid-leak resistance.

A high-Q tank circuit can also be obtained with a higher L/C ratio by "tapping down" the tube connections on the tank (§ 2-10). This is advantageous in that a coil with higher inherent Q can be used; also, the circulating r.f. current in the tank circuit is reduced so that drift from coil heating is decreased. However, the circuit is complicated to some extent and the taps may cause parasitic oscillations to be set up (§ 4-10).

Plate supply — Since the oscillator frequency will be affected to some extent by changes in plate supply voltage, it is necessary that the latter be free from ripple (§ 8-4) which would cause frequency variations at the ripple-frequency rate (*frequency modulation*). It is also advantageous to use a voltage-stabilized power supply (§ 8-8). Since the oscillator is usually operated at low voltage and current, gaseous regulator tubes are quite suitable.

Power level — The self-controlled oscillator should be designed purely for frequency control and not to give appreciable power output, hence small tubes of the receiving type may be used. The power input is ordinarily not more than a watt or two, subsequent buffer amplifiers being used to increase the power to the desired level. The use of receiving tubes is advantageous mechanically, since the small elements are less susceptible to vibration and are usually securely braced to the envelope.

Oscillator adjustment — The adjustment of an oscillator consists principally in observing the design principles outlined in the preceding paragraphs. Frequency stability should be checked with the aid of a stable receiver, or an auxiliary crystal oscillator may be used as a standard for checking dynamic stability and drift, the self-controlled oscillator being adjusted to approximately the same frequency so that an audio-frequency beat (§ 2-13) can be obtained. If it is possible to vary the oscillator plate voltage (an adjustable resistor of 50,000 or 100,000 ohms in series with the plate supply lead will give considerable variation) the change in frequency with change in plate voltage may be observed and the operating conditions varied until minimum frequency shift results. The principal factors affecting dynamic stability will be the tank circuit L/C ratio, the grid-leak resistance, and the amount of feedback. In the electron-coupled circuit the latter may be adjusted by changing the position of the cathode tap on the tank coil; this adjustment is quite important in its effect on the frequency stability.

Drift may be checked by allowing the oscillator to operate continuously from a cold start, the frequency change being observed at regular intervals. Drift may be minimized by using less than the rated power input to the plate of the tube, by construction which prevents tube heat from reaching the tank circuit elements, and by use of large wire in the tank coil to reduce temperature rise from internal heating.

In the electron-coupled oscillator having a tuned plate circuit (Fig. 401-A) resonance at the fundamental and harmonic frequencies of the oscillator portion of the tube will be indicated by a dip in plate current as the plate tank condenser is varied. This dip should be rather marked at the fundamental, but will be less so on harmonic frequencies.

●4-3 PIEZO-ELECTRIC CRYSTALS

Characteristics — Piezo-electric crystals (§ 2-10) are universally used for controlling the frequency of transmitting oscillators because the extremely high Q of the crystal and the necessarily loose coupling between it and the oscillator tube make the frequency stability of a crystal-controlled oscillator very high. Active plates may be cut from a raw crystal at various angles to its electrical, mechanical and optical axes, resulting in differing characteristics as to thickness, frequency-temperature coefficient, power-handling capabilities, etc. The commonly used cuts are designated as X, Y, AT, V, and LD.

The ability to adhere closely to a known frequency is the outstanding characteristic of a crystal oscillator. This is also its disadvantage, in that the oscillator frequency can be changed appreciably only by using a number of crystals.

Frequency-thickness ratio — Crystals used for transmitting purposes are so cut that the thickness of the crystal is the frequency-determining factor, the length and width of the plate being of relatively minor importance. For a given crystal cut, the ratio between thickness and frequency is a constant; that is,

$$F = \frac{k}{t}$$

where F is the frequency in megacycles and t is the thickness of the crystal in thousandths of an inch. For the X-cut, $k = 112.6$; for the Y-cut, $k = 77.0$; for the AT-cut, $k = 66.2$.

At frequencies above the 7-Mc. region the crystal becomes very thin and correspondingly fragile, so that crystals are seldom manufactured for operation much above this frequency. Direct crystal control on 14 and 28 Mc. is secured by use of "harmonic" crystals, which are ground to be active oscillators when excited at the third harmonic of the frequency represented by their thickness.

Temperature coefficient of frequency — The resonant frequency of a crystal will vary with its temperature, to an extent depending upon the type of cut. The frequency-temperature coefficient is usually expressed in cycles frequency change per megacycle, per degree Centigrade temperature change, and may be either positive (increasing frequency with increasing temperature) or negative (decreasing frequency with increasing temperature). X-cut crystals have a negative coefficient of 15 to 25 cycles/megacycle/degree C. The coefficient of Y-cut crystals may vary from − 20 cycles/megacycle/degree C. to + 100 cycles/megacycle/degree C. The AT, V and LD cuts have very low coefficients. Y-cut crystals frequently "jump" to another frequency when the temperature is changed rather than gradually changing frequency as the nominal coefficient might indicate, and hence are rather unreliable under temperature variations.

The temperature of a crystal depends not only on the temperature of its surroundings but also on the power it must dissipate while oscillating, since power dissipation causes heating (§ 2-6, 2-8). Consequently the crystal temperature may be considerably above that of the surrounding air when the oscillator is in operation. To minimize heating and frequency drift (§ 3-7) the power used in the crystal must be kept to a minimum.

Power limitations — If the crystal is made to oscillate too strongly, as when it is used in an oscillator circuit with high plate voltage and excessive feedback, the amplitude of the mechanical vibration will become great enough to crack or puncture the quartz. An indication of the vibration amplitude can be obtained by connecting an r.f. current indicating device of suitable range in series with the crystal. Safe r.f. crystal currents range from 50 to 200 milliamperes, depending upon the type of cut. A flashlight bulb or dial light of equivalent current rating makes a good current indicator. By choosing a bulb of lower rating than the current specified by the manufacturer as safe for the particular type of crystal used, the bulb will serve as a fuse, burning out before a current dangerous to the crystal is reached. The 60-ma. and 100-ma. bulbs are frequently used for this purpose. High crystal current is accompanied by increased power dissipation and heating, so that the frequency change also is greatest when the crystal is overloaded.

Crystal mountings — To make use of the crystal, it must be mounted between two metal electrodes. There are two types of mountings, one having a small air-gap between the top plate and the crystal and the other maintaining both plates in contact with the crystal. It is es-

sential that the surfaces of the metal plates in contact with the crystal be perfectly flat. In the air-gap type of holder, the frequency of oscillation depends to some extent upon the size of the gap. This property can be used to advantage with most low-drift crystals so that by using a holder having a top plate with closely adjustable spacing a controllable frequency variation can be obtained. A 3.5-Mc. crystal will oscillate without very great variation in power output over a range of about 5 kc. X- and Y-cut crystals are not generally suitable for this type of operation because they have a tendency to "jump" in frequency with different air gaps.

A holder having a heavy metal bottom plate with a large surface exposed to the air is advantageous in radiating quickly the heat generated in the crystal and thereby reducing temperature effects. Different plate sizes, pressures, etc., will cause slight changes in frequency, so that if a crystal is being ground to an exact frequency it should be tested in the holder and with the same oscillator circuit with which it will be used in the transmitter.

• 4-4 CRYSTAL OSCILLATORS

Triode oscillators — The triode crystal oscillator circuit (§ 3-7) is shown in Fig. 403. The limit of plate voltage that can be used without endangering the crystal is about 250 volts. With the r.f. crystal current limited to a safe value of about 100 ma., the power output obtainable is about 5 watts. The oscillation frequency is dependent to some extent on the plate tank tuning because of the change in input capacity with changes in effective amplification (§ 3-3).

Tetrode and pentode oscillators — Since the power output of a crystal oscillator is limited by the permissible r.f. crystal current (§ 4-3), it is advantageous to use an oscillator tube of high power sensitivity (§ 3-3), such as a pentode or beam tetrode (§ 3-5). Thus for a given crystal voltage or current more power output may be obtained than with the triode oscillator, or for a given output the crystal voltage will be lower, thereby reducing crystal heating. In addition, tank circuit tuning and loading react less on the crystal frequency because of the lower grid-plate capacity (§ 3-3).

Fig. 404 shows a typical pentode or tetrode oscillator circuit. The pentode and tetrode tubes designed for audio power work are excellent crystal-oscillator tubes. The screen voltage is generally of the order of half the plate voltage for optimum operation. Small tubes rated at 250 volts for audio work may be operated with 300 volts on the plate and 100–125 on the screen as crystal oscillators. The screen is at ground potential for r.f. and has no part in the operation of the circuit other than to set the operating characteristics of the tube. The larger beam tubes may be operated at 400 to 500 volts on the plate and 250 on the screen for maximum output.

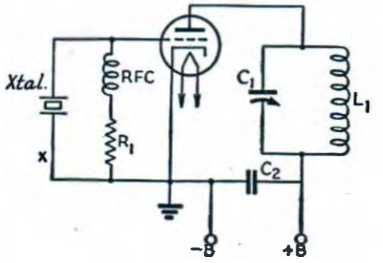


Fig. 403 — Triode crystal oscillator. The tank condenser C_1 may be a 100-$\mu\mu$fd. variable, with L_1 proportioned so that the tank will tune to the crystal frequency. C_2 should be 0.001 μfd. or larger. The grid leak, R_1, will vary with the type of tube; high-μ types take lower values, 2500 to 10,000 ohms, while medium and low-μ types take values of 10,000 to 25,000 ohms. Flashlight bulb or r.f. milliammeter (§ 4-3) may be inserted at X.

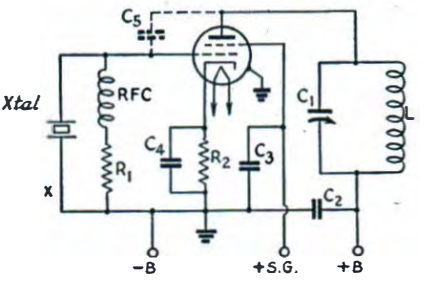


Fig. 404 — Tetrode or pentode crystal oscillator. Typical values: C_1, 100 $\mu\mu$fd. with L wound to suit frequency; C_2, C_3, 0.001 μfd. or larger; C_4, 0.01 μfd.; R_1, 10,000 to 50,000 ohms, best value being determined by trial for the plate voltage and operating conditions chosen; R_2, 250 to 400 ohms. R_2 and C_4 may be omitted, connecting cathode directly to ground, if plate voltage is limited to 250 volts. C_5 (if needed) may be formed by two metal plates about ½ inch square spaced about ¼ inch. If the tube has a suppressor grid, it should be grounded. X indicates point where flashlight bulb may be inserted (§ 4-3).

Pentode oscillators operating at 250 to 300 volts will give 4 or 5 watts output under normal conditions. The beam types 6L6 and 807 will give 15 watts or more at maximum plate voltage.

The grid-plate capacity may be too low to give sufficient feedback, particularly at the lower frequencies, in which case a feedback condenser, C_5, may be required. Its capacity should be the lowest value which will give stable oscillation.

Circuit constants — Typical values for grid-leak resistance and by-pass condenser values are given in Figs. 403 and 404. Since the

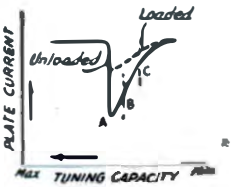


Fig. 405 — D.c. plate current vs. plate tuning capacity with the triode, tetrode or pentode crystal oscillator.

crystal is the frequency-determining element, the *Q* of the plate tank circuit has a relatively minor effect on the oscillator frequency. A *Q* of 12 (§ 4-8) is satisfactory for average conditions, but departure from this figure will not greatly affect the performance of the oscillator.

Adjustment of crystal oscillators — The tuning characteristics and procedure to be followed in tuning are essentially the same for triode, tetrode or pentode crystal oscillators. Using a plate milliammeter as an indicator of oscillation (a 0–100 ma. d.c. meter will have ample range for all low-power oscillators), the plate current will be found to be steady when the circuit is in the non-oscillating state, but will dip when the plate condenser is tuned through resonance at the crystal frequency. Fig. 405 is typical of the behavior of plate current as the tank condenser capacity is varied. An r.f. indicator, such as a small neon bulb touched to the plate end of the tank coil, will show maximum at point *A*. However, when the oscillator is delivering power to a load it is best to operate in the region *B-C*, since the oscillator will be more stable and there is less likelihood that a slight change in loading will throw the circuit out of oscillation, which is likely to happen when operation is too near the critical point, *A*. The crystal current is lower in the *B-C* region.

When power is taken from the oscillator, the dip in plate current is less pronounced, as indicated by the dotted curve. The greater the power output the smaller the dip in plate current. If the load is made too great, oscillations will not start. Loading is adjusted by varying the coupling to the load circuit (§ 2-11).

The greater the loading, the smaller the voltage fed back to the grid circuit for excitation purposes. This means that the r.f. voltage across the crystal also will be reduced, hence there is less crystal heating when the oscillator is delivering power than when operating unloaded.

Failure of a crystal circuit to oscillate may be caused by any of the following:

1. Dirty, chipped or fractured crystal
2. Imperfect or unclean holder surfaces
3. Too tight coupling to load
4. Plate tank circuit not tuning correctly
5. Insufficient feedback capacity

Pierce oscillator — This circuit is shown in Fig. 406. It is equivalent to the ultraudion circuit (§ 3-7) with the crystal replacing the tuned circuit. The output of the Pierce oscillator is relatively small, although it has the advantage that no tuning controls are required. The circuit requires capacitive coupling to a following stage. The amount of feedback is determined by the condenser C_2. To sustain oscillation the net reactance (§ 2-8) of the plate-cathode circuit must be capacitive; this condition is met so long as the inductance of the r.f. choke, together with the inductance of any coils associated with the input circuit

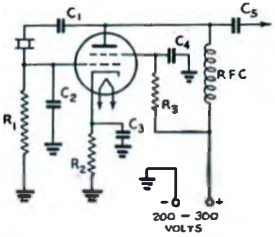


Fig. 406 — Pierce oscillator circuit. Tubes such as the 6C5 and 6F6 are suitable, operating at plate voltages not exceeding 300 to prevent crystal fracture. When a triode is used, R_3 and C_4 are omitted. R_1 should be 25,000 to 50,000 ohms. 1000 ohms is recommended for R_2. R_3 is the screen voltage dropping resistance (75,000 ohms for the 6F6). C_1 may have any value between 0.001 and 0.01 μfd. C_3 and C_4 should be 0.01 μfd. C_2, the regeneration capacity, must be determined by experiment; usual values are between 50 and 150 μμfd. The capacity of C_5, usually 100 μμfd., should be adjusted so that the oscillator is not overloaded.

of the following stage and the tube and stray capacities, forms a circuit tuned to a lower frequency than that of the crystal.

● 4-5 HARMONIC-GENERATING CRYSTAL OSCILLATORS

Tri-tet oscillator — The Tri-tet oscillator circuit is shown in Fig. 407. In this circuit the screen grid is operated at ground potential and the cathode at an r.f. potential above ground. The screen-grid acts as the anode of a triode crystal oscillator, while the plate or output circuit is tuned to the oscillator frequency or, for harmonic output, to a multiple of it.

Besides harmonic output, the Tri-tet circuit has the "buffering" feature of electron-coupling between crystal and output circuits (§ 4-2). This makes the crystal frequency less susceptible to changes in loading or tuning and hence improves the stability.

If the output circuit is to be tuned to the same frequency as the crystal, a tube having low grid plate capacity (§ 3-2, 3-5) must be used, otherwise there may be excessive feedback and danger of fracturing the crystal.

The cathode tank circuit, L_1C_1, is not tuned to the frequency of the crystal, but to a con-

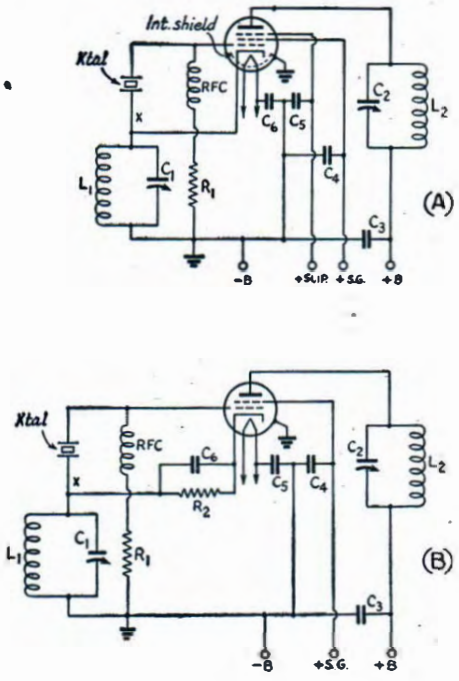


Fig. 407 — Tri-tet oscillator circuit, using pentodes (A) or beam tetrodes (B). C_1 and C_2, 200-$\mu\mu$fd. variable; C_3, C_4, C_5, C_6, 0.001 to 0.01 μfd., not critical; R_1, 20,000 to 100,000 ohms; R_2, 400 ohms for 400- or 500-volt operation.

Following specifications for cathode coils, L_1, are based on a coil diameter of 1½ inches and length 1 inch; turns should be spaced evenly to fill the required length. For 1.75-Mc. crystal, 32 turns; 3.5 Mc., 10 turns, 7 Mc., 6 turns. The screen should be operated at 250 volts or less. Audio beam tetrodes such as the 6L6 and 6L6G should be used only for second-harmonic output. Flashlight bulb may be inserted at X (§ 4-3).

The L-C ratio in the plate tank, L_2C_2, should be adjusted so that the capacity in use is 75 to 100 $\mu\mu$fd. for fundamental output and about 25 $\mu\mu$fd. for second harmonic output.

siderably higher frequency. Recommended values for L_1 are given under the diagram. C_1 should be set as near minimum capacity as is consistent with good output. This reduces the crystal voltage.

With pentode-type tubes having separate suppressor connections, the suppressor may be connected directly to ground or may be operated at about 50 volts positive. The latter method will give somewhat higher output than with the suppressor connected to ground.

With transmitting pentodes or beam tubes operated at 500 volts on the plate an output of 15 watts can be obtained on the fundamental and very nearly as much on the second harmonic.

Grid-plate oscillator — In the grid-plate oscillator, Fig. 408, the crystal is connected between grid and ground and the cathode tuned circuit C_2RFC is tuned to a lower frequency than that of the crystal. This circuit gives high output on the fundamental crystal frequency with low crystal current. The output on even harmonics (2nd, 4th, etc.) is not as great as that obtainable with the Tri-tet, but the output on odd harmonics (3rd, 5th, etc.) is appreciably better.

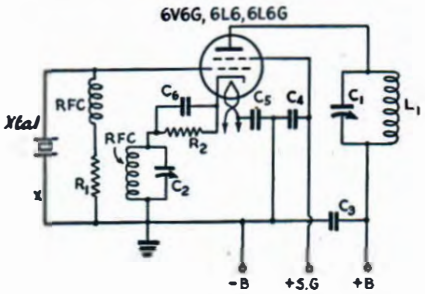


Fig. 408 — Grid-plate crystal oscillator circuit. In the cathode circuit, RFC is a 2.5-mh. r.f. choke. Other constants are the same as in Fig. 506. X indicates point where crystal-current indicator may be inserted (§ 4-3).

If harmonic output is not needed, C_2 may be a fixed capacity of 100 $\mu\mu$fd. The cathode coil, RFC, may be a 2.5-mh. choke, since the inductance is not critical.

Output power of 15 to 20 watts may be obtained at the crystal fundamental with a tube such as the 6L6G at plate and screen voltages of 400 and 250, respectively.

Tuning and adjustment — The tuning procedure for the Tri-tet oscillator is as follows: With the cathode tank condenser at about three-quarters scale, turn the plate tank condenser until there is a sharp dip in plate current, indicating that the plate circuit is in resonance. The crystal should be oscillating continuously regardless of the setting of the plate condenser. Set the plate condenser so that plate current is minimum. The load circuit may then be coupled and adjusted so that the oscillator delivers power. The minimum plate current will rise; it may be necessary to retune the plate condenser when the load is coupled to bring the plate current to a new minimum. Fig. 409 shows the typical behavior of plate current with plate-condenser tuning.

After the plate circuit is adjusted and the oscillator is delivering power, the cathode condenser should be readjusted to obtain optimum power output. The setting should be as far toward the low-capacity end of the scale as is consistent with good output; it may, in

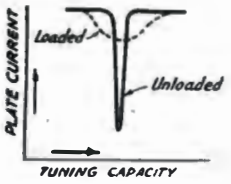


Fig. 409 — D.c. plate current vs. plate tuning capacity with the Tri-tet oscillator.

fact, be desirable to sacrifice a little output if so doing reduces the current through the crystal and thus reduces heating.

For harmonic output the plate tank circuit is tuned to the harmonic instead of the fundamental of the crystal frequency. A plate-current dip will occur at the harmonic. If the cathode condenser is adjusted for maximum output at the harmonic, this adjustment will usually serve for the fundamental as well. The crystal should be checked for evidence of excessive heating, the most effective remedy for which is to lower the plate and/or screen voltage, or to reduce the loading. With this circuit maximum r.f. voltage across the crystal is developed at maximum load so crystal heating should be checked with the load coupled.

When a fixed cathode condenser is used in the grid-plate oscillator the plate tank circuit is simply resonated, as indicated by the plate-current dip, to the fundamental or a harmonic of the output frequency, loading being adjusted to give optimum power output. If the variable cathode condenser is used, it should be set to give, by observation, the maximum power output consistent with safe crystal current. The variable condenser is chiefly useful in increasing the output on the third and higher harmonics; for fundamental operation the cathode capacity is not critical and the fixed condenser may be used.

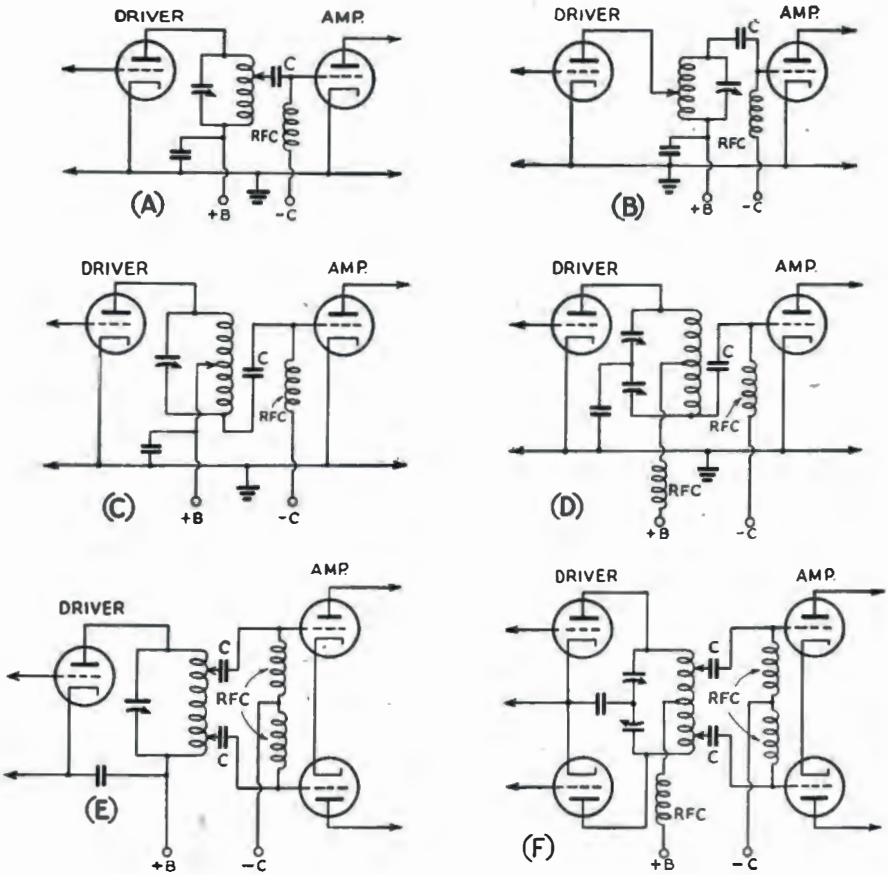


Fig. 410 — Direct- or capacity-coupled driver and amplifier stages. Coupling condenser capacity may be from 50 μμfd. to 0.002 μfd., not critical except when tapping the coils for control of excitation is not possible. Parallel plate feed to the driver and series grid feed to the amplifier may be substituted in any of the circuits (§ 3-7).

● 4-6 INTERSTAGE COUPLING

Requirements — The purpose of the interstage coupling system is to transfer, with as little energy loss as possible, the power developed in the plate circuit of one tube (the *driver*) to the grid circuit of the following amplifier tube or frequency multiplier. The circuits in practical use are based on the fundamental coupling arrangements described in § 2-11. In the process of power transfer, impedance transformation (§ 2-9) also is frequently necessary so that the proper exciting voltage and current will be available at the grid of the driven tube.

Capacity coupling — Fig. 410 shows several types of capacitive coupling. In each case, *C* is the coupling condenser. The coupling condenser serves also as a blocking condenser (§ 2-13) to isolate the d.c. plate voltage of the driver from the grid of the amplifier. The circuits of C and D are preferable when a balanced circuit is used in the output of the driver; instead of both tubes being in parallel across one side, the output capacity of the driver tube and the input capacity of the amplifier are across opposite sides of the tank circuit, thereby preserving a better circuit balance. The circuits of E and F are designed for coupling to a push-pull stage.

In A, B, E and F, excitation is adjusted by moving the tap on the coil to provide an optimum impedance match. In E and F, the two grid taps should be maintained equidistant from the center-tap on the coil.

While capacitive coupling is simplest from the viewpoint of construction, it has certain disadvantages. The input capacity of the amplifier is shunted across at least a portion of the driver tank coil. When added to the output capacity of the driver tube, this additional capacity may be sufficient, in many cases, to prevent use of a desirable L/C ratio in circuits for frequencies above about 7 Mc.

Link coupling — At the higher frequencies it is advantageous in reducing the effects of tube capacities on the L/C ratio to use separate tank circuits for the driver plate and amplifier grid, coupling the two circuits by means of a link (§ 2-11). This method of coupling also has some constructional advantages, in that separate parts of the transmitter may be constructed as separate units without the necessity for running long leads at high r.f. potential.

Circuits for link coupling are shown in Fig. 411. The coupling ordinarily is by a turn or two of wire closely coupled to the tank inductance at a point of low r.f. potential such as the center of the coil of a balanced tank circuit, or the "ground" end of the coil in a single-ended circuit. The link line usually consists of two closely-spaced parallel wires; occasionally the wires are twisted together, but this usually causes undue losses at high frequencies.

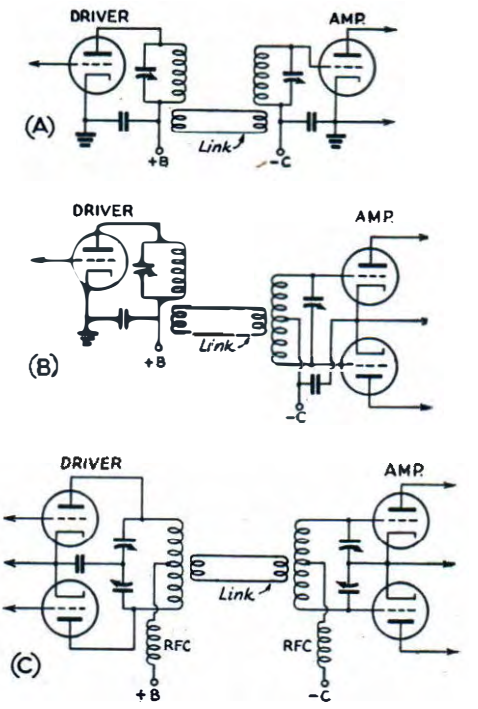


Fig. 411 — Link coupling between driver and amplifier.

It is advisable to have some means of varying the coupling between link and tank coils. The link coil may be arranged to be swung in relation to the tank coil or, when it consists of a large turn around the outside of the tank coil, can be split into two parts which can be pulled apart or closed somewhat in the fashion of a pair of calipers. If the tank coils are wound on forms, the link may be wound close to the main coil.

With fixed coils, some adjustment of coupling can usually be obtained by varying the number of turns on the link. In general the proper number of turns for the link must be found by experiment.

● 4-7 R.F. POWER AMPLIFIER CIRCUITS

Tetrode and pentode amplifiers — When the input and output circuits of an r.f. amplifier tube are tuned to the same frequency, it will oscillate as a tuned-grid tuned-plate oscillator unless some means is provided to eliminate the effects of feedback through the plate-to-grid capacity of the tube (§ 3-5). In all transmitting r.f. tetrodes and pentodes, this capacity is reduced to a satisfactory degree by the internal shielding between grid and plate provided by the screen. Tetrodes and pentodes

designed for audio use (such as the 6L6, 6V6, 6F6, etc.) are not sufficiently well screened for use as r.f. amplifiers without employing additional means for nullifying the effect of the grid-plate capacity.

Typical circuits of tetrode and pentode r.f. amplifiers are shown in Fig. 412. The high power sensitivity (§ 3-3) of pentodes and tetrodes, however, makes them prone to self-oscillate with very small values of feedback voltage, so that particular care must be used to prevent feedback by means external to the tube itself. This calls for adequate isolation of plate and grid tank circuits to prevent undesired magnetic or capacity coupling between them. The requisite isolation can he secured by keeping the circuits well separated and mounting the coils so that magnetic coupling is minimized, or by shielding (§ 2-11).

Triode amplifiers — The feedback through the grid-plate capacity of a triode cannot be eliminated in the tube itself, and therefore special circuit means, called *neutralization*, must be used to prevent oscillation. A properly-neutralized triode amplifier then behaves as though it were operating at very low frequencies where the grid-plate capacity feedback is negligible (§ 3-3).

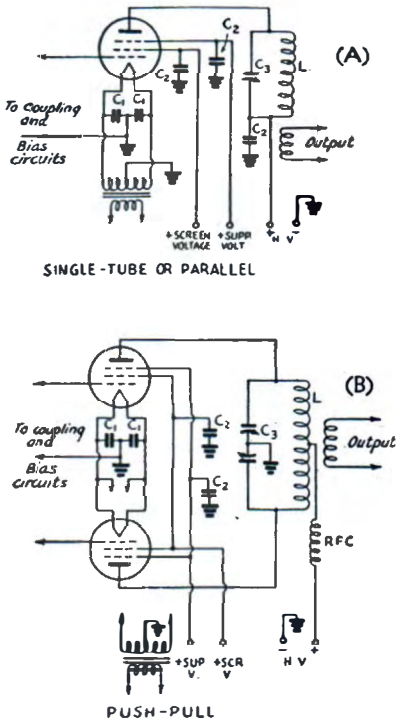


Fig. 412 — Tetrode-pentode r.f. amplifier circuits. C_1 — 0.01 μfd.; C_2 — 0.001 μfd. or larger; C_3-L — see § 4-8.

In circuits for tetrodes, the suppressor-grid connection and by-pass condenser are omitted.

Neutralization — Neutralization amounts to taking some of the radio-frequency current from the output or input circuit of the amplifier and introducing it into the other circuit in such a way that it effectively cancels the current flowing through the grid-plate capacity of the tube, thus rendering it impossible for the tube to supply its own excitation. For complete neutralization it is necessary that the two currents be opposite in phase (§ 2-7) and equal in amplitude.

The out-of-phase current (or voltage) can be obtained quite readily by using a balanced tank circuit in either grid or plate, taking the neutralizing voltage from the end of the tank opposite that to which the grid or plate is connected. The amplitude of the neutralizing voltage can be regulated by means of a small condenser, the *neutralizing condenser*, having the same order of capacity as the grid-plate capacity of the tube. Circuits in which the neutralizing voltage is obtained from a balanced grid tank and fed to the plate through the neutralizing condenser are termed *grid-neutralized* circuits, while if the neutralizing voltage is obtained from a balanced plate tank and fed to the grid of the tube, the circuit is *plate-neutralized.*

Plate-neutralized circuits — The circuits for plate neutralization are shown in Fig. 413 at A, B and C. In A, voltage induced in the extension of the tank coil is fed back to the grid through the neutralizing condenser C_n to balance the voltage appearing between grid and plate. In this circuit the capacity required at C_n increases as the tank coil extension is made smaller; in general, neutralization is satisfactory over only a small range of frequencies since the coupling between the two sections of the tank coil will vary with the amount of capacity in use at C.

In B the tank coil is center-tapped to give equal voltages on either side of the center tap, the tank condenser being across the whole coil. The neutralizing capacity is approximately equal to the grid-plate capacity of the tube in this case. A disadvantage of the circuit, when used with the single tank condenser shown, is that the rotor of the condenser is above ground potential and hence small capacity changes caused by bringing the hand near the tuning control (*hand capacity*) cause detuning. In general, neutralization is complete at only one frequency since the plate-cathode capacity of the tube is across only half the tank coil; also, it is difficult to secure an exact center-tap. Both these cause unbalance which in turn causes the voltages across the two halves of the coil to differ when the frequency is changed.

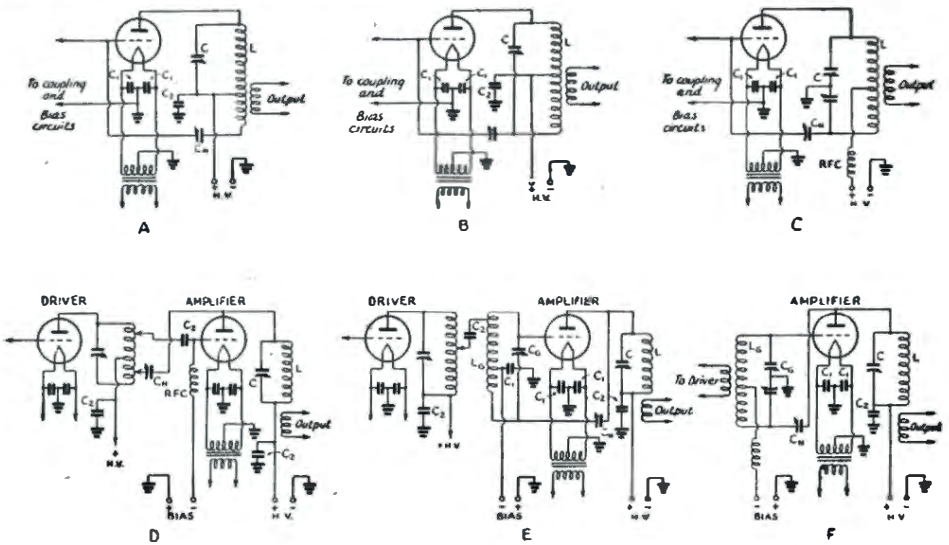


Fig. 413 — Triode amplifier circuits. Plate neutralization is shown in A, B and C; D, E and F show types of grid neutralization. Either capacitive or link coupling may be used with circuits of A, B or C.
C-L — See § 4-8.
C_G-L_G — Grid tank circuit.
C_N — Neutralizing condensers.
C_1 — 0.01 μfd.
C_2 — 0.001 μfd. or larger.

The circuit of C also uses a center-tapped tank circuit, the voltage division being secured by use of a balanced (split-stator) tank condenser, the two condenser sections being identical. C_N is approximately equal to the grid-plate capacity of the tube. In this circuit the upper section of the tank condenser is in parallel with the output capacity of the tube, hence the circuit can be completely neutralized at only one setting of the tank condenser unless a compensating capacity (Fig. 414) is connected across the lower section. In practice, if the capacity in use in the tank circuit is large compared to the plate-cathode capacity the unbalancing effect is not serious.

Grid-neutralized circuits — Typical circuits employing grid neutralization are shown in Fig. 413 at D, E and F. The principle of balancing out the feed-back voltage is the same as in plate neutralization. However, in these circuits the fed-back voltage may either be in phase or out of phase with the excitation voltage on the grid side of the input tank circuit (and the opposite on the other side) depending upon whether the tank is divided by means of a balanced condenser or a tapped coil. Circuits such as those at D and E neutralized by ordinary procedure (described below) will be regenerative when the plate voltage is applied; the circuit at F will be degenerative. In addition, the normal unbalancing effects described in the preceding paragraph are present, so that grid neutralising is less satisfactory than the plate method.

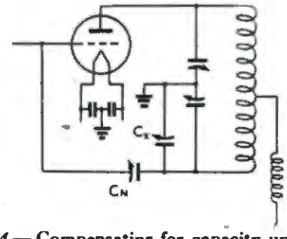


Fig. 414 — Compensating for capacity unbalance in the single-tube neutralizing circuit. C_x, the balancing capacity, should be variable and should have a maximum capacity somewhat larger than the output capacity of the tube. It is adjusted to minimize shift in neutralizing capacity at C_N as the frequency is changed.

Inductive neutralization — With this type of neutralization inductive coupling between the grid and plate circuits is provided in such a way that the voltage induced in the grid coil by magnetic coupling from the plate coil opposes the voltage fed back through the grid-plate capacity of the tube. A representative circuit arrangement, using a coupling link to provide the mutual inductance (§ 2-11) is shown in Fig. 415. Ordinary inductive coupling between the two coils also could be used, but is less convenient. Inductive neutralization is complete at only one frequency, since the effective mutual inductance changes to some extent with tuning, but is useful in cases where the grid-plate capacity of the tube being neutralized is

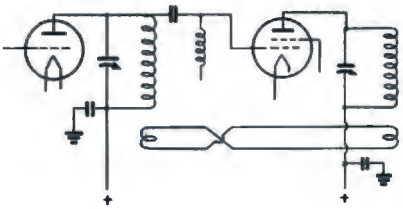


Fig. 415 — Inductive neutralizing circuit. The link coils should have one or two turns and should be coupled to the grounded ends of the tank coils. Neutralization is adjusted by moving the link coils in relation to the tank coils. Reversal of connections to one of the coils may be required to obtain the proper phasing.

very small and suitable circuit balance cannot be obtained with circuits using neutralizing condensers.

Push-pull neutralization — With push-pull circuits two neutralizing condensers are used as shown in Fig. 416. In these circuits the grid-plate capacities of the tubes and the neutralizing capacities form a capacity bridge (§ 2-11) which is independent of the grid and plate tank circuits. The neutralizing capacities are approximately the same as the tube grid-plate capacities. With electrically similar tubes and symmetrical construction (stray capacities to ground equal on both sides of the circuit) the neutralization is complete and independent of frequency. A circuit using a balanced condenser, as at B, is preferred since it is an aid in obtaining good circuit balance.

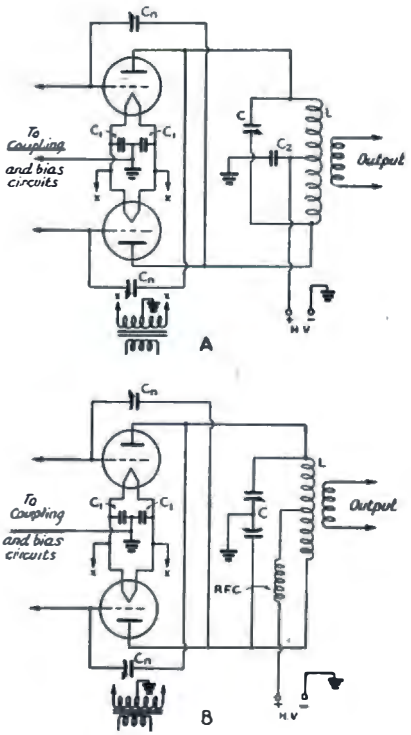


Fig. 416 — Push-pull triode amplifier circuits with "cross-neutralization." Either capacitive or link coupling may be used.
C-L — See § 4-8.
C_n — Neutralizing condensers.
C_1 — 0.01 μfd.
C_2 — 0.001 μfd. or larger.

Frequency effects — The effects of slight dissymmetry in a neutralized circuit become more important as the frequency is raised, and may be sufficient at ultra-high frequencies (or even lower) to prevent good neutralization. At these frequencies the inductances and stray capacities of even short leads become important elements in the circuit, while input loading effects (§ 7-6) may make it impossible to get proper phasing, particularly in single-tube circuits. In such cases the use of a push-pull amplifier, with its general freedom from the effects of dissymmetry, is not only much to be preferred but may be the only type of circuit which can be satisfactorily neutralized.

Neutralizing condensers — In most cases the neutralizing voltage will be equal to the r.f. voltage between the plate and grid of the tube so that for perfect balance the capacity required in the neutralizing condenser theoretically will be equal to the grid-plate capacity. If, in the circuits having tapped tank coils, the tap is more than half the total number of turns from the plate end of the coil, the required neutralizing capacity will increase approximately in proportion to the relative number of turns in the two sections of the coil.

With tubes having grid and plate connections brought out through the bulb, a condenser having at about half-scale or less a capacity equal to the grid-plate capacity of the tube should be chosen. If the grid and plate leads are brought through a common base, the capacity needed is greater because the tube socket and its associated wiring adds some capacity to the actual inter-element capacities.

When two or more tubes are connected in parallel, the neutralizing capacity required will be in proportion to the number of tubes.

The voltage rating of neutralizing condensers must at least equal the r.f. voltage across the condenser plus the sum of the d.c. plate voltage and the grid-bias voltage.

Neutralizing procedure — The procedure in neutralizing is essentially the same for all tubes and circuits. The filament of the tube should be lighted and the excitation from the preceding stage should be fed to the grid circuit. There should be no plate voltage on the amplifier.

The grid-circuit milliammeter makes a good

neutralizing indicator. If the circuit is not completely neutralized, tuning of the plate tank circuit through resonance will change the tuning of the grid circuit and affect its loading, causing a change in the rectified d.c. grid current. The setting of the neutralizing condenser which leaves the grid current unaffected as the plate tank is tuned through resonance is the correct one. If the circuit is out of neutralization, the grid current will drop perceptibly as the plate tank is tuned through resonance. As the point of neutralization is approached, by adjusting the neutralizing capacity in small steps, the dip in grid current as the plate condenser is swung through resonance will become less and less pronounced until, at exact neutralization, there will be no dip at all. Further change of the neutralizing capacity in the same direction will bring the grid-current dip back. The neutralizing condenser should always be adjusted with a screwdriver of insulating material to avoid hand-capacity effects.

Adjustment of the neutralizing condenser may affect the tuning of the grid tank or driver plate tank, so both circuits should be retuned each time a change is made in neutralizing capacity. In neutralizing a push-pull amplifier, the neutralizing condensers should be adjusted together, step by step, keeping their capacities as equal as possible.

With single-ended circuits having split-stator neutralizing, the behavior of the grid meter will depend somewhat upon the type of tube used. If the tube output capacity is not great enough to upset the balance, the action of the meter will be the same as in other circuits. With high-capacity tubes, however, the meter usually will show a gradual rise and fall as the plate tank is tuned through resonance, reaching a maximum right at resonance when the circuit is properly neutralized.

When an amplifier is not neutralized, a neon bulb touched to the plate of the amplifier tube or to the plate side of the tuning condenser will glow when the tank circuit is tuned through resonance, providing the driver has sufficient power. The glow will disappear when the amplifier is neutralized.

However, touching the neon bulb to such an ungrounded point in the circuit may introduce enough stray capacity to unbalance the circuit slightly, thus upsetting the neutralizing.

A flashlight bulb connected in series with a single-turn loop of wire 2½ or 3 inches in diameter, with the loop coupled to the tank coil, will also serve as a neutralizing indicator. Capacitive unbalance can be avoided by coupling the loop to the low-potential part of the tank coil.

Incomplete neutralization — If a setting of the neutralizing condenser can be found which gives minimum r.f. current in the plate tank circuit without completely eliminating it, there may be magnetic or capacity coupling between the input and output circuits external to the tube itself. Short leads in neutralizing circuits are highly desirable, and the input and output inductances should be so placed with respect to each other that magnetic coupling is minimized. Usually this requires that the axes of the coils should be at right angles to each other. In some cases it may be necessary to shield the input and output circuits from each other. Magnetic coupling can be detected by disconnecting the plate tank from the remainder of the circuit and testing for r.f. in it (by means of the flashlight lamp and loop) as the tank condenser is tuned through resonance. The driver stage must be operating, of course.

With single-ended amplifiers there are many stray capacities left uncompensated for in the neutralizing process. With large tubes, especially those having relatively high interelectrode capacities, these commonly neglected stray capacities can prevent perfect neutralization. Symmetrical arrangement of a push-pull amplifier is about the only way to obtain practically perfect balance throughout the amplifier.

The neutralization of tubes with extremely low grid-plate capacity, such as the 6L6, is often difficult, since it frequently happens that the wiring itself will introduce sufficient capacity between the right points to "overneutralize" the grid-plate capacity. The use of a neutralizing condenser only aggravates the condition. Inductive or link neutralization as shown in Fig. 415 has been used successfully with such tubes.

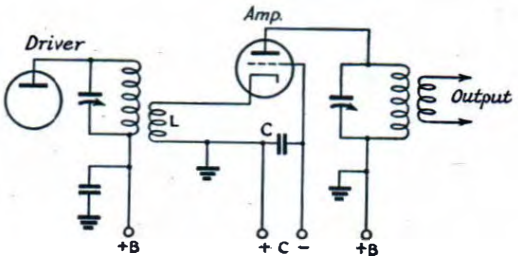


Fig. 417 — Inverted amplifier. The number of turns at L should be adjusted by experiment to give optimum grid excitation to the amplifier. By-pass condenser C may be 0.001 μfd. or larger.

The inverted amplifier — The circuit of Fig. 417 avoids the necessity for neutralization by operating the control grid of the tube at ground potential, thus making it serve as a shield between the input and output circuits. It is particularly useful with tubes of low grid-plate capacity which are difficult to neutralize by ordinary methods. Excitation is applied between grid and cathode through the coupling coil L; since this coil is common to both the plate and grid circuits the amplifier is degenerative with the circuit constants normally used, hence more excitation voltage and power are required for a given output than is the case with a neutralized amplifier. The tube used must have low plate-cathode capacity (of the order of 1 $\mu\mu$fd. or less) since larger values will give sufficient feedback to permit it to oscillate, the circuit then becoming the ultraudion (§ 3-7). Tubes having sufficiently low plate-cathode capacity (audio pentodes, for example) can be used without danger of oscillation at frequencies up to 30 Mc. or so.

● 4-8 POWER AMPLIFIER OPERATION

Efficiency — An r.f. power amplifier is usually operated Class-C (§ 3-4) to obtain a reasonably-high value of plate efficiency (§ 3-3). The higher the plate efficiency the higher the power input that can be applied to the tube without exceeding the plate dissipation rating (§ 3-2), up to the limits of other tube ratings (plate voltage and plate current). Plate efficiencies of the order of 75% are readily obtainable at frequencies up to the 30–60 megacycle region. The *overall* efficiency of the amplifier will be lower by the percentage of power lost in the tank and coupling circuits, so that the actual efficiency is less than the plate efficiency.

Operating angle — The operating angle is the proportionate part of the exciting grid-voltage cycle (§ 2-7) during which plate current flows, as shown in Fig. 418. For Class-C operation it is usually in the vicinity of 120–150 degrees which, with other operating considerations, results in an optimum relationship between plate efficiency and grid driving power.

Load impedance — The load impedance (§ 3-3) for an r.f. power amplifier is adjusted, by tuning the plate tank circuit to resonance, to represent a pure resistance at the operating frequency (§ 2-10). Its value, which is usually in the neighborhood of a few thousand ohms, is adjusted by varying the loading on the tank circuit, closer coupling to the load giving lower values of load resistance and vice versa (§ 2-11). The load may be either the grid circuit of a following stage or the antenna circuit.

For highest efficiency the value of load resistance should be relatively high, but if only limited excitation voltage is available greater power output will be secured by using a lower value of load resistance. The latter adjustment is accompanied by a decrease in plate efficiency. The optimum load resistance is that which, for the maximum permissible peak

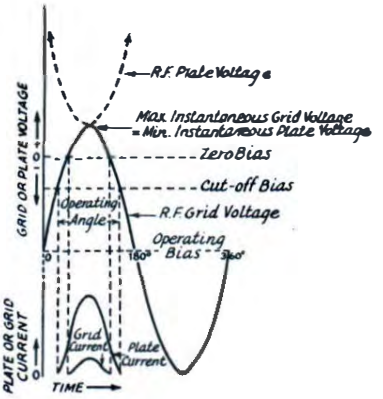


Fig. 418 — **Instantaneous voltages and currents in a Class-C amplifier operating under optimum conditions.**

plate current, causes the minimum instantaneous plate voltage (Fig. 418) to be equal to the maximum instantaneous grid voltage required to cause the peak plate current to flow; this gives the optimum ratio of plate efficiency to required grid driving power.

R.f. grid voltage and grid bias — For most tubes optimum operating conditions result when the minimum instantaneous plate voltage is 10% to 20% of the d.c. plate voltage, so that the maximum instantaneous positive grid voltage must be approximately the same figure. Since plate current starts flowing when the instantaneous voltage reaches the cut-off value (§ 3-2), the d.c. grid voltage must be considerably higher than cut-off to confine the operating angle to 150 degrees or less (with grid bias at cut-off the angle would be 180 degrees). For an angle of 120 degrees the r.f. grid voltage must reach 50% of its peak value (§ 2-7) at the cut-off point. The corresponding figure for an angle of 150 degrees is 25%. Hence the operating bias required is the cut-off value plus 25% to 50% of the peak r.f. grid voltage. These relations are shown in Fig. 418. The grid bias should be at least twice cut-off if the amplifier is to be plate modulated so that the operating angle will not be less than 180 degrees when the plate voltage rises to twice the steady d.c. value (§ 5-3). Because of their relatively high amplification factors, with most modern tubes Class-C operation requires considerably more than twice cut-off bias to make the operating angle fall in the region mentioned above.

Suitable operating conditions are usually given in the data accompanying the type of tube used.

Grid bias may be secured either from a bias source (*fixed bias*), a grid leak (§ 3-6) of suitable value, or from a combination of both. When a bias supply is used, its voltage regulation should be taken into consideration (§ 8-9).

Driving power — As indicated in Fig. 418, grid current flows only during a small portion of the peak of the r.f. grid voltage cycle. The power consumed in the grid circuit is therefore approximately equal to the peak r.f. grid voltage multiplied by the average rectified grid current as read by a d.c. milliammeter. The peak r.f. grid voltage, if not included in the tube manufacturer's operating data, can be estimated roughly by adding 10% to 20% of the plate voltage to the operating grid bias, assuming the operating conditions are as described above.

At frequencies up to 30 Mc. or so the grid losses are practically entirely those resulting from grid-current flow. At ultra-high frequencies, however, dielectric losses in the glass envelope and base materials become appreciable, together with losses caused by transit-time effects (§ 7-6), and may necessitate supplying several times the driving power indicated above. At any frequency, the driving stage should be capable of a power output two to three times the power it is expected the grid circuit of the amplifier will consume. This is necessary because losses in the tank and coupling circuits must also be supplied, and also to provide reasonably good regulation of the r.f. grid voltage. Good voltage regulation (see § 8-1 for general definition) insures that the waveform of the excitation voltage will not be distorted because of the changing load on the driver during the r.f. cycle.

Grid impedance — During most of the r.f. grid voltage cycle, no grid current flows, as indicated in Fig. 418, hence the grid impedance is infinite. During the peak of the cycle, however, the impedance may drop to very low values (of the order of 1000 ohms) depending upon the type of tube. Both the minimum and average values of grid impedance depend to a considerable extent on the amplification factor of the tube, being lower with tubes having large amplification factors.

The average grid impedance is equal to E^2/P, where E is the r.m.s. (§ 2-7) value of r.f. grid voltage and P the grid driving power. Under optimum operating conditions values of average grid impedance ranging from 2000 ohms for high-μ tubes to four or five times as much for low-μ types are representative. Values in the vicinity of 4000 to 5000 ohms are typical of modern triodes with amplification factors of 20 to 30.

Because of the large change in impedance during the cycle it is necessary that the tank circuit associated with the amplifier grid have fairly high Q so that the voltage regulation over the cycle will be good. The requisite Q may be obtained by adjusting the L/C ratio or by tapping the grid circuit across only part of the tank (§ 4-6).

Tank circuit Q — Besides serving as a means for transforming the actual load resistance to the required value of plate load impedance for the tube, the plate tank circuit also should suppress the harmonics present in the tube output as a result of the non-sinusoidal plate current (§ 2-7, 3-3). For satisfactory harmonic suppression a Q of 12 or more (with the circuit fully loaded) is desirable. A Q of this order is also helpful from the standpoint of securing adequate coupling to the load or antenna circuit (§ 2-11). The proper Q can be obtained by suitable selection of L/C ratio in relation to the optimum plate load resistance for the tube (§ 2-10).

For a Class-C amplifier operated under optimum conditions as described above, the plate load impedance is approximately proportional to the ratio of d.c. plate voltage to d.c. plate current. For a given effective Q, the tank capacity required at a given frequency will be inversely proportional to the parallel resistance (§ 2-10), so that it will also be inversely proportional to the plate-voltage/plate-current ratio. The capacity required on various amateur bands for a Q of 12 is shown in Fig. 419 as a function of this ratio. The capacity given is for single-ended tank circuits as shown in Fig. 420 at A and B. When a balanced tank circuit is used, the total tank capacity required is reduced to ¼ this value because the tube is connected across only half the circuit (§ 2-9). Thus if the plate-voltage/plate-current ratio calls for a capacity of 200 $\mu\mu$fd. in a single-ended circuit at the desired frequency, only 50 $\mu\mu$fd. would be needed in a balanced circuit. If a split-stator or balanced tank condenser is used, each section should have a capacity of 100 $\mu\mu$fd., the total capacity of the two in series being 50 $\mu\mu$fd. These are "in use" capacities, not simply the rated maximum capacity of the condenser. Larger values may be used with an increase in the effective Q.

To reduce energy loss in the tank circuit the inherent Q of the coil and condenser should be high. Since transmitting coils usually have Q's ranging from 100 to several hundred, the tank transfer efficiency is generally 90% or more. An unduly large C/L ratio is not advisable since it will result in large circulating r.f. tank current and hence relatively large losses in the tank, with a consequent reduction in the power available for the load.

Tank constants — When the capacity nec-

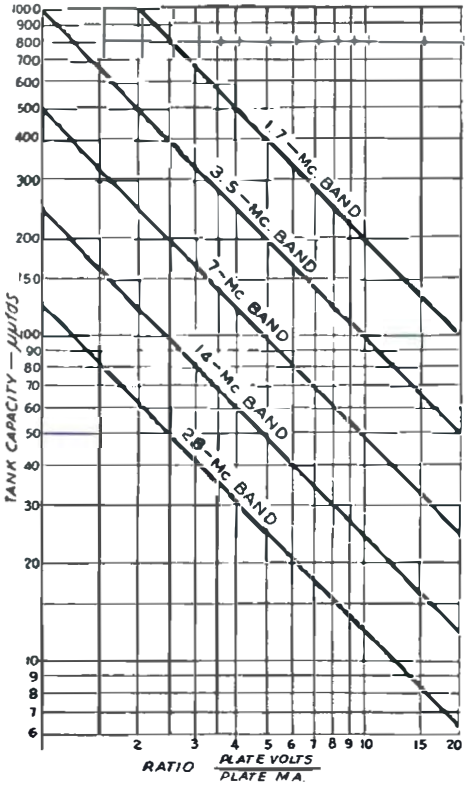


Fig. 419 — Chart showing tank-capacities required for "Q" of 12 with various ratios of plate voltage to plate current for various frequencies. In circuits F, G, H (Fig. 420), the capacities shown in the graph may be divided by four. In circuits C, D, E, I, J and K, the capacity of each section of the split-stator condenser may be one-half that shown by the graph. Values given by the graph should be used for circuits A and B.

essary for a *Q* of 12 has been determined from Fig. 419, the inductance required to resonate at the given frequency can be found by means of the formula in § 2-10. Alternatively, the required number of turns on coils of various construction can be found from the charts of Figs. 421 and 422.

Fig. 421 is for coils wound on receiving-type forms having a diameter of 1½ inches and ceramic forms having a diameter of 1¾ inches and winding length of 3 inches. Such coils would be suitable for oscillator and buffer stages where the power is not over 50 watts. In all cases the number of turns given must be wound to fit the length indicated and the turns should be evenly spaced.

Fig. 422 gives data on coils wound on transmitting-type ceramic forms. In the case of the smallest form, extra curves are given for double-spacing (winding turns in alternate grooves). This is sometimes advisable in the case of 14- and 28-Mc. coils when only a few turns are required. In all other cases it is assumed that the specified number of turns is wound in the grooves without any additional spacing.

Ratings of components — The peak voltage to be expected between the plates of a tank condenser depends upon the arrangement of the tank circuit as well as the d.c. plate voltage. Peak voltage may be determined from Fig. 420, which shows all of the commonly used tank-circuit arrangements. These estimates assume

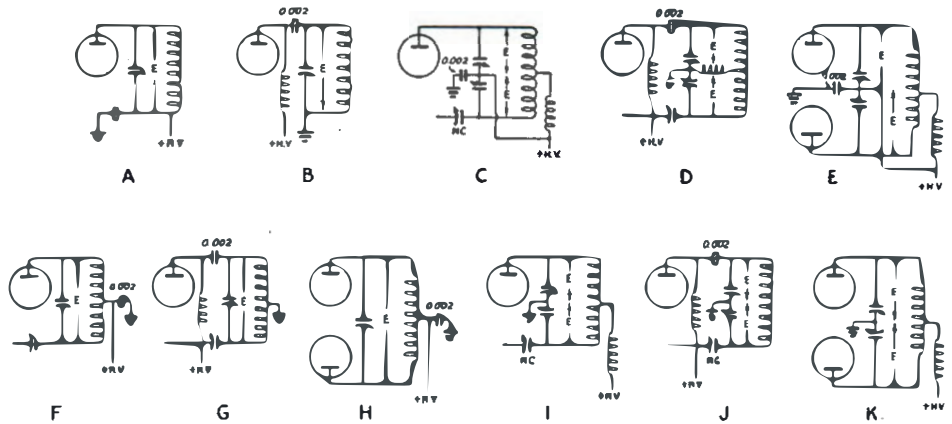


Fig. 420 — In circuits A, B, C, D and E, the peak voltage E will be approximately equal to the d.c. plate voltage applied for c.w. or twice this value for phone. In circuits F, G, H, I, J and K, E will be twice the d.c. plate voltage for c.w. or 4 times the plate voltage for phone. Circuit is assumed to be fully loaded. Tubes in parallel in any of the circuits will not affect the peak voltage. Circuits A, C, E, F, G and H require that the tank condenser be insulated from chassis or ground and be provided with a suitably insulated shaft coupling for tuning.

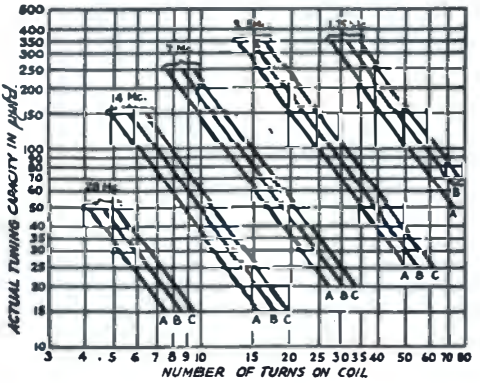


Fig. 421 — Coil-winding data for receiving-type forms, diameter 1½ inches. Curve A — winding length, 1 inch; Curve B — winding length, 1½ inches; Curve C — winding length, 2 inches. Curve C is also suitable for coils wound on 1¾-inch diameter ceramic forms with 3 inches of winding length.

that the amplifier is fully loaded; the voltage will rise considerably should the amplifier be operated without load. The figures include a reasonable factor of safety.

The condenser plate spacing required to withstand any particular voltage will vary with the construction. Most manufacturers specify peak voltage ratings for their condensers.

Plate or screen by-pass condensers of 0.001 μfd. should be satisfactory for frequencies as low as 1.7 Mc. Cathode-resistor and filament by-passes in r.f. circuits should be not less than 0.01 μfd. Condensers should have voltage ratings 25 to 50% greater than the maximum d.c. or a.c. voltage across them.

Interstage coupling condensers should have voltage ratings 50% to 100% greater than the sum of the driver plate and amplifier biasing voltages.

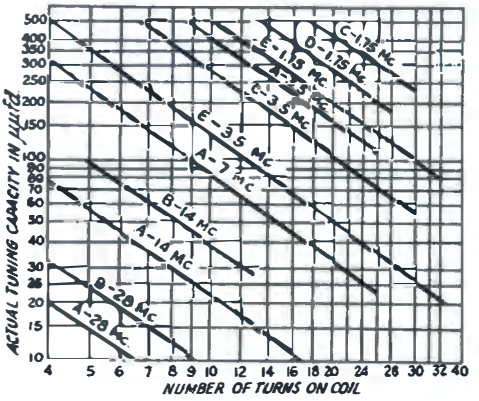


Fig. 422 — Coil-winding data for ceramic transmitting-type forms. Curve A — ceramic form 2½-inch effective diameter, 26 grooves, 7 per inch; Curve B — same as A, but with turns wound in alternate grooves; Curve C — ceramic form 2⅞-inch effective diameter, 32 grooves, 7.1 turns per inch, app.; Curve D — ceramic form 4-inch effective diameter, 28 grooves, 5.85 turns per inch, app.; Curve E — ceramic form 5-inch effective diameter, 26 grooves, 7 per inch. Coils may be wound with No. 12 or No. 14 wire.

● 4-9 ADJUSTMENT OF POWER AMPLIFIERS

Excitation — The effectiveness of adjustments to the coupling between the driver plate and amplifier grid circuits can be gauged by the relative values of amplifier rectified grid current and driver plate current, the object being to obtain maximum grid current with minimum driver loading. The amplifier grid circuit represents the load on the driver, and the average grid impedance must be transformed to the proper value for optimum driver operation (§ 4-8).

With capacity coupling, either the driver plate or amplifier grid must be tapped down on the driver tank coil as shown in Fig. 410 at A and B unless the grid impedance is approximately the right value for the driver plate load, when it will be satisfactory to connect both elements to the end of the tank. If the grid impedance is lower than the required driver plate load, Fig. 410-A is used; if higher, Fig. 410-B. In either case the coupling which gives the desired grid current with minimum driver loading should be determined experimentally by moving the tap. Should both plate and grid be connected to the end of the circuit it is sometimes possible to control the loading, when the grid impedance is low, by varying the capacity of the coupling condenser, *C*, but this method is not altogether satisfactory since it is simply an expedient to prevent driver overloading without giving suitable impedance matching.

In push-pull circuits the method of adjustment is the same, except that the taps should be kept symmetrically located with respect to the center of the tank circuit.

With link coupling, Fig. 411, the object of adjustment is the same. The two tanks are first tuned to resonance, as indicated by maximum grid current, and the coupling adjusted by means of the links (§ 4-6), to give maximum grid current with minimum driver plate current. This will usually suffice to load the driver to its rated output provided the driver plate and amplifier grid tank circuits have reasonable values of *Q*. If the *Q* of one or both of the circuits is too low, it may not be possible to load the driver fully with any adjustment of link turns or coupling at either tank. In such a case the *Q*s of the tank circuits must be increased to the point where adequate coupling is secured. If the driver plate tank is designed to have a *Q* of 12, the difficulty almost invariably is in the

amplifier grid tank. Its Q can be increased to a suitable value by adjustment of the L/C ratio or by tapping the load across part of the coil (§ 2-10).

Whatever the type of coupling, a preliminary adjustment should be made with the proper bias voltage and/or grid leak, but with the amplifier plate voltage off; then the amplifier should be carefully neutralized. After neutralization, the driver-amplifier coupling should be readjusted for optimum power transfer, after which plate voltage may be applied and the amplifier plate circuit adjusted to resonance and coupled to its load. Under actual operating conditions the grid current decreases below the value obtained without plate voltage on the amplifier and the effective grid impedance rises, hence the final adjustment is to recheck the coupling to take care of this shift.

With recommended bias, the grid current obtained before plate voltage is applied to the amplifier should be 25% to 30% higher than the value required for operating conditions. If this value is not obtained, and the driver plate input is up to rated value, the reason may be either improper matching of the amplifier grid to the driver plate or simply insufficient power output from the driver to take care of all losses. Driver operating voltages should be checked to assure they are up to rated values. If batteries are used for bias and are not strictly fresh, they should be replaced, since batteries which have been in use for some time often develop high internal resistance which effectively acts as additional grid-leak resistance. If a rectified a.c. bias supply is used, the bleeder or voltage-divider resistances should be checked to make certain that low grid current is not caused by greater grid-circuit resistance than is recommended. In this connection it is helpful to measure the actual bias when grid current is flowing, by means of a high-resistance d.c. voltmeter. There is also the possibility of loss of filament emission of the amplifier tube either from prolonged service or from operating the filament under or over the rated voltage.

Plate tuning — In preliminary tuning, it is desirable to use low plate voltage to avoid possible damage to the tube. With excitation and plate voltage applied, rotate the plate tank condenser until the plate current dips, then set the condenser at the minimum plate-current point (resonance). When the resonance point has been found, the plate voltage may be increased to its normal value.

With adequate excitation, the off-resonance plate current of a triode amplifier may be two or more times the normal operating value. With screen-grid tubes, the off-resonance plate current may not be much higher than the normal operating value since the plate current is principally determined by the screen rather than the plate voltage.

With reasonably efficient operating conditions, the minimum plate current with the amplifier unloaded will be a small fraction of

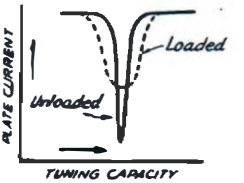


Fig. 423 — Typical behavior of d.c. plate current with tuning of an amplifier.

the rated plate current for the tube (usually a fifth or less) since with no load the parallel impedance of the tank circuit is high. If the excitation is low, the "dip" will not be very marked, but with adequate excitation the plate current at resonance without loading will be just high enough so that the d.c. plate power input supplies all the losses in the tube and circuit. As an indication of probable efficiency, the minimum plate current value should not be taken too seriously, because without load the Q of the circuit is high and the tank current relatively large. When the amplifier is delivering power to a load, the circulating current drops considerably and the tank losses correspondingly decrease. High minimum unloaded plate current is chiefly encountered at 28 Mc. and above, where tank losses are higher and the tank L/C ratio is usually lower than normal because of irreducible tube capacities. The effect is particularly noticeable with screen-grid tubes which have relatively high output capacity. Because of the decrease in tank r.f. current with loading, however, the actual efficiency under load is reasonably good.

With the load (antenna or following amplifier grid circuit) connected, the coupling between plate tank and load should be adjusted to make the tube take rated plate current, keeping the tank always tuned to resonance. As the output coupling is increased, the minimum plate current will also increase about as shown in Fig. 423. Simultaneously, the tuning becomes less sharp, because of the increase in effective resistance of the tank. If the load circuit simulates a resistance, the resonance setting of the tank condenser will be practically unchanged with loading; this is generally the case since the load circuit itself usually is also tuned to resonance. A reactive load (such as an antenna or feeder system which is not tuned exactly to resonance) may cause the tank condenser setting to change appreciably with loading since reactance as well as resistance is coupled into the tank (§ 2-11).

Power output — As a check on the operation of an amplifier, its power output may be

measured by the use of a load of known resistance coupled to the amplifier output as shown in Fig. 424. At A a thermoammeter *M* and non-inductive (ordinary wire-wound resistors are not satisfactory) resistance *R* are connected across a coil of a few turns coupled to the amplifier tank coil. The higher the resistance of *R*, the greater the number of turns required in the coupling coil. A resistor used in this way is generally called a "dummy antenna," since its use permits the transmitter to be adjusted without actually radiating power. The loading may readily be adjusted by varying the coupling between the two coils, so that the amplifier draws rated plate current when tuned to resonance. The power output is then calculated from Ohm's Law:

$$P \text{ (watts)} = I^2R$$

where *I* is the current indicated by the thermoammeter and *R* is the resistance of the non-inductive resistor *R*. Special resistance units are available for this purpose ranging from 73 to 600 ohms (simulating antenna and transmission-line impedances) at power ratings up to 100 watts. For higher powers, the units may be connected in series-parallel. The meter scale required for any expected value of power output may also be determined from Ohm's Law:

$$I = \sqrt{\frac{P}{R}}$$

Incandescent light bulbs can be used to replace the special resistor and thermoammeter.

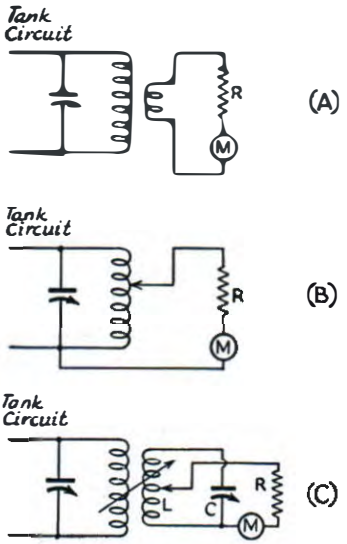


Fig. 424 — "Dummy antenna" circuits for checking power output and making operating adjustments without applying power to the actual antenna.

The lamp should be equipped with a pair of leads, preferably soldered to the terminals on the lamp base. The coupling should be varied until the greatest brilliance is obtained for a given plate input. In using lamps as dummy antennas, a size corresponding to the expected power output should be selected so that the lamp will operate near its normal brilliancy. Then when the adjustments have been completed an approximation of the power output can be obtained by comparing the brightness of the lamp with the brightness of one of similar power rating in a 115-volt socket.

The circuit of Fig. 424-B is for resistors or lamps of relatively high resistance. In using this circuit, care should be taken to avoid accidental contact with the plate tank when the power is on. This danger is avoided by circuit C, in which a separate tank circuit, *LC*, tuned to the operating frequency, is coupled to the plate tank circuit. The loading is adjusted by varying the number of turns across which the dummy antenna is connected on *L* and by changing the coupling between the two coils. With push-pull amplifiers, the dummy antenna should be tapped equally on either side of the center of the tank, when Fig. 424-B is used.

Harmonic suppression — The most important step to take in elimination of harmonic radiation (§ 4-8, 2-12) is to use an output tank circuit having a *Q* of 12 or more. Beyond this, it is desirable to avoid any considerable amount of over-excitation of a Class-C amplifier, since excitation in excess of that required for normal Class-C operation further distorts the plate-current pulse and increases the harmonic content in the output of the amplifier even though the proper tank *Q* is used. If the antenna system will accept harmonic frequencies they will be radiated when present, consequently the antenna coupling system preferably should be selected with harmonic transfer in mind (§ 10-6).

Harmonic content can be reduced to some extent by preventing distortion of the r.f. grid voltage waveshape. This can be done by using a grid tank circuit with high effective *Q*. Link coupling between the driver and final amplifier are helpful, since the two tank circuits provide more attenuation than one at the harmonic frequencies. However, the advantages of link coupling in this respect may be nullified unless the *Q* of the grid tank is high enough to give good voltage regulation and thus prevent distortion in the grid circuit.

The stray capacity between the antenna coupling coil and the tank coil may be sufficient to couple harmonic energy into the antenna system. This coupling may be eliminated by the use of electrostatic shielding (*Faraday shield*) between the two coils. Fig. 425 shows the construction of such a shield, while Fig. 426 illustrates the manner in which it is installed.

The construction shown in Fig. 425 is used to prevent current flow in the shield, which would occur if the wires formed closed circuits since the shield is in the magnetic field of the tank coil. Should this occur there would be magnetic shielding as well as electrostatic; in addition, there would be an undesirable power loss in the shield.

Improper operation — Inexact neutralization or stray coupling between plate and grid circuits may result in regeneration. This effect is most evident with low excitation, when the amplifier will show a sudden increase in output when the plate tank circuit is tuned slightly to the high-frequency side of resonance. It is accompanied by a pronounced increase in grid current.

Self-oscillation is apt to occur with tubes of high power sensitivity such as the r.f. pentodes

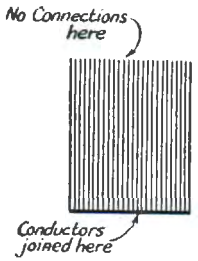


Fig. 425 — **The Faraday shield. It is made of parallel conductors, insulated from each other except at one end where all are joined. Stiff wire or small diameter rod may be used, spaced about the diameter of the wire or rod.**

and tetrodes. In event of either regeneration or oscillation, circuit components should be arranged so that those in the plate circuit are well isolated from those of the grid circuit. Plate and grid leads should be made as short as possible and the screen should be by-passed as close to the socket terminal as possible. A cylindrical shield surrounding the lower portion of the tube up to the lower edge of the plate is sometimes required.

"Double resonance" or two tuning spots on the plate tank condenser, one giving minimum plate current and the other maximum power output, may occur when the tank circuit Q is too low (§ 2-10). A similar effect also occurs at times with screen-grid amplifiers when the screen voltage regulation (§ 8-1) is poor, as when the screen is supplied through a dropping resistor. The screen voltage decreases with an increase in plate current, because the screen current increases under the same conditions. Thus the minimum plate current point causes the screen voltage, and hence power output, to be less than when a slightly higher plate current is drawn.

A phenomenon known as "grid emission" may occur when the amplifier tube is operated at higher than rated power dissipation on either the plate or grid. It is particularly likely to occur with tubes having oxide-coated cathodes such as the indirectly-heated types. It is caused by the grid reaching a temperature high enough to cause electron emission (§ 2-4). The electrons so emitted are attracted to the plate, further increasing the power input and heating, so that grid emission is characterized by gradually increasing plate current and heat which eventually will ruin the tube if the power is not removed. Grid emission can be prevented by operating the tube within its ratings.

• 4-10 PARASITIC OSCILLATIONS

Description — If the circuit conditions in an oscillator or amplifier are such that self-oscillation at some frequency other than that desired exist, the spurious oscillation is termed *parasitic*. The energy required to maintain a parasitic oscillation is wasted so far as useful output is concerned, hence an oscillator or amplifier having parasitics will operate at reduced efficiency. In addition, its behavior at the operating frequency often will be erratic. Parasitic oscillations may be higher or lower in frequency than the operating frequency of the amplifier.

The parasitic oscillation usually starts the instant plate voltage is applied or, when the amplifier is biased beyond cut-off, at the instant excitation is applied. In the latter case, the oscillation frequently will be self-sustaining after the excitation has been removed. At other times the oscillation may not be self-sustaining, becoming active only in the presence of excitation. It may be apparent only by the production of abnormal key clicks (§ 6-1) over a wide frequency range or by the presence of similarly wide-spread spurious side-bands (§ 5-2) with 'phone modulation.

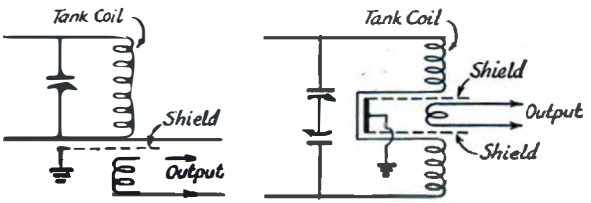


Fig. 426 — **Methods of using the Faraday shield. Two are required with a push-pull or balanced tank circuit. The shield should be somewhat larger than the diameters of the coupled coils, and should be inserted between them so each is completely unexposed to the other.**

Low-frequency parasitics — Parasitic oscillations at low frequencies (usually 500 kc. or less) are of the tuned-plate tuned-grid type, the tuned circuits being formed by r.f. chokes and associated by-pass and coupling condensers, with the regular tank tuning condensers having only a minor effect on the oscillation. The operating-frequency tank coil has negligible inductance for such low frequencies and may be short-circuited without affecting the oscillations. The oscillations do not occur when no r.f. chokes are used, hence whenever possible in series-fed circuits such chokes should be omitted. With single-ended amplifiers it is usually possible to arrange the circuit so that either the grid or plate circuit needs no choke. In push-pull stages where chokes must be used in both plate and grid circuits, it is helpful to connect an unby-passed grid leak from the choke to the bias supply or ground, thus placing the resistance in the parasitic circuit and tending to prevent oscillation. When the driver plate circuit has parallel feed and the amplifier grid circuit series feed (§ 3-7) this type of oscillation cannot occur so long as no choke is used in the series grid circuit, since the grid is grounded through the tank coil for the parasitic frequency.

Parasitics near operating frequency — In circuits utilizing a tap on the plate tank coil to establish a ground for a balanced neutralizing circuit, such as Fig. 413-B, a parasitic oscillation may be set up if the amplifier grid is tapped down on the grid (or driver plate) tank circuit for adjustment of driver-amplifier coupling (§ 4-6). In this case the turns between grid and ground, and between plate and ground, form with the stray and other capacities present a t.p.t.g. circuit (§ 3-7) which oscillates at a frequency somewhat higher than the nominal operating frequency. Such an oscillation can be prevented by dispensing with the taps in either the plate or grid circuit. Balancing the plate circuit by means of a split-stator condenser, as in Fig. 413-C, is recommended.

Ultra-high frequency parasitics — Parasitics in the u.h.f. region are likely to occur with any amplifier having a balanced tank circuit, particularly when associated with neutralizing connections. The parasitic circuit may be either of the t.p.t.g. or ultraudion type, and is formed by the leads connecting the various components.

The frequency of such oscillations may be determined by connecting a tuned circuit in series with the grid lead to the tube. A variable condenser (50 or 100 $\mu\mu$fd.) may be used in conjunction with three or four self-supporting turns of heavy wire wound in a coil an inch or so in diameter. With the amplifier oscillating at the parasitic frequency, the condenser is slowly tuned through its range until oscillations cease. In case this point is not found on first trial, the turns of the coil may be spread apart or a turn removed and the process repeated. While this may not be the simplest cure in all cases, the use of such a tuned circuit as a trap is an almost certain remedy, if the frequency can be determined, and introduces little if any loss at the operating frequency.

An alternative cure which is feasible when the oscillation is of the t.p.t.g. type is to detune the parasitic circuit in either the plate or grid circuit. Since this type of oscillation occurs most frequently with push-pull amplifiers, it may often be cured by making the grid and plate leads to their respective tank circuits of considerably different length. Similar considerations apply to neutralizing connections in push-pull circuits. The extra wire length may be coiled up in the form of a so-called "choke," which in this case is simply additional inductance for detuning the parasitic circuit.

Testing for parasitic oscillations — An amplifier always should be tested for parasitic oscillations before being considered ready for service. The preferable method is first to neutralize the amplifier, then apply sufficient fixed bias to permit a moderate value of plate current to flow without excitation. (The plate current should not be large enough to cause the power input to exceed the rated plate dissipation of the tube.) If the amplifier is free from self-starting parasitics the plate current will remain steady as the tank condensers are varied in capacity; also, there will be no grid current and a neon bulb touched either to the plate or grid will show no glow. Care must be used not to let the hand come in contact with any metal parts of the transmitter in using the neon bulb.

If any of these effects are present the frequency of the parasitic must first be determined. If r.f. chokes are used in both the plate and grid circuits one of them should be short-circuited to determine if the oscillation is at a low frequency; if so, it may be eliminated by the methods outlined above. If the test indicates that the parasitic is not a low-frequency oscillation, the grid trap described above should be tried for the u.h.f. type. The type which occurs near the operating frequency will not occur unless the plate and grid tank coils are both tapped, hence may be eliminated from consideration if this is not the case in the circuit used. When it is possible for such an oscillation to be present, its existence can be detected very readily by moving the grid tap to include the whole tank circuit, when the oscillation will cease.

Some indication of the frequency of the parasitic can be obtained from the color of the glow in the neon bulb. Usually it will be yellow-

ish with low-frequency oscillations and violet with u.h.f. oscillations.

If the amplifier is stable under the conditions described above, excitation should be applied and then removed to ascertain if a self-sustaining oscillation is set up with excitation. If the plate current does not return to the previous value when the excitation is cut off, the same tests should be applied to determine the parasitic frequency.

As a final test, the transmitter should be put on the air and a nearby receiver tuned over as wide a frequency range as possible to locate any off-frequency signals associated with the radiation. Parasitics usually can be recognized by their poor stability, as contrasted to the normal transmitter harmonics, which will have the same stability as the fundamental signal as well as the usual harmonic frequency relationship. Harmonics should be quite weak compared to the fundamental frequency, whereas parasitic oscillations may have considerable strength.

• 4-11 FREQUENCY MULTIPLICATION

Circuits — A frequency multiplier is an amplifier having its plate tank circuit tuned to a multiple (harmonic) of the frequency applied to its grid. The difference between a straight amplifier (§ 4-1) and a frequency multiplier is in the way in which it is operated rather than in the circuit. However, since the grid and plate tank circuits are tuned to different frequencies a triode frequency multiplier will not self-oscillate, hence does not need neutralization. A typical circuit arrangement is shown in Fig. 427-A. For screen-grid multipliers the circuit is the same as in Fig. 412-A. Under usual conditions the plate efficiency of a frequency multiplier drops off rapidly with an increase in the number of times the frequency is multiplied. For this reason most multipliers are used as *frequency doublers*, giving second harmonic output.

A special circuit for frequency doubling ("push-push" doubler) is shown in Fig. 427-B. The grids of the tubes are in push-pull and the plates in parallel, thus the plate tank circuit receives two pulses of plate current for each cycle of excitation frequency. The circuit is similar in principle to the full-wave rectifier (§ 8-3) where the ripple frequency is twice the applied frequency.

Push-pull amplifiers are suitable for frequency multiplication at odd harmonics but are unsuited to doubling or other even-harmonic multiplication because the even harmonics are largely balanced out in the tank circuit (§ 3-3).

Operating conditions and circuit constants — To obtain good efficiency the operating angle at the harmonic frequency must be 180 degrees or less, preferably in the vicinity of 150–120 degrees (§ 4-8). In a doubler this means that plate current should flow during only half this angle of fundamental frequency. Consequently the r.f. grid voltage, operating bias, and grid driving power must be increased considerably beyond the values obtaining for normal Class-C amplification. For comparable plate efficiency the bias will ordinarily be four to five times the normal Class-C bias, and the r.f. grid voltage must be considerably larger to drive the tube to the same peak plate current. Since the plate and grid current pulses under these conditions have the same peak amplitudes but only half the time duration as in a straight amplifier, the average d.c. values should be one-half those for normal Class-C operation. That is, a tube operated in this way will have the same plate efficiency as a Class-C amplifier, but can be operated at only half the plate input so that the output power also is halved. The driving power required is usually about twice that for straight-through amplification with the same plate efficiency.

Greater output can be secured by using a larger operating angle (lower grid bias) or lower plate load resistance to increase the plate

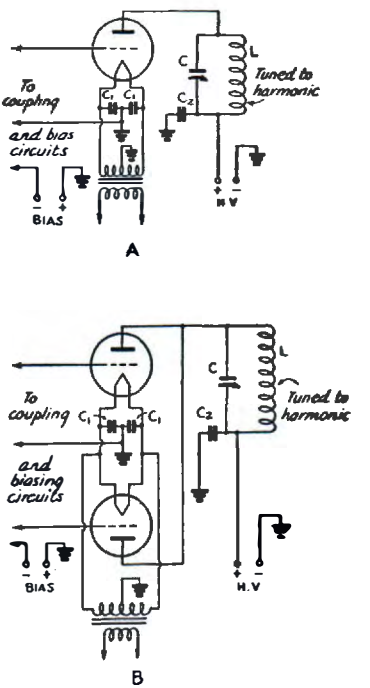


Fig. 427 — Frequency-multiplying circuits. A is for triodes, used either singly or in parallel. The push-push doubler is shown at B. Any type of coupling may be used between the grid circuit and the driver. C_1 should be 0.01 μfd. or larger; C_2, 0.001 μfd. or larger.

current, but this is accompanied by a decrease in efficiency. Since operation as described above is below the maximum plate dissipation rating of the tube, the decrease in efficiency can usually be tolerated in the interests of securing somewhat more power output. Ordinarily the efficiency is 40% to 50%.

The tank circuit should have reasonably high Q (12 is satisfactory) to give good output voltage regulation (§ 4-9) since a plate-current pulse occurs only once every two cycles of output frequency. A low-Q circuit (high L/C ratio) is helpful chiefly when the operating angle is greater than 180 degrees at the second harmonic. Such a tank circuit will have relatively high impedance to the considerable fundamental-frequency component of plate current which is present with large operating angles, and thus aid in reducing the average d.c. plate current.

The grid impedance of a frequency multiplier is considerably higher than that of a straight amplifier because of the high bias voltage. The average impedance can be calculated as previously described (§ 4-8). The L/C ratio of the grid tank circuit may be higher, therefore, for a given Q. It is often advantageous to use a fairly high ratio since a large r.f. voltage must be developed between grid and cathode, so long as it is not made too high (Q too low) to permit adequate coupling between the grid tank circuit and the driver stage. In some cases it may be necessary to step up the driver output voltage to obtain sufficient r.f. grid voltage for the doubler; this may be done by tapping the driver plate on its tank circuit, when capacity coupling is used, or by similar tapping or use of a higher C/L ratio in the driver plate tank when the stages are link-coupled (§ 4-6).

Tubes for frequency multiplication — There is no essential difference between tubes of various characteristics in their performance as frequency doublers. Tubes having high amplification factors will require somewhat less bias for equivalent operation, but the grid driving power needed is almost independent of the μ, assuming tubes of otherwise similar construction and characteristics. Pentodes and tetrodes having high power sensitivity will, as in normal amplifier operation, require less driving power than triodes for efficient doubling, although more power will be needed than for straight amplification.

● 4-12 ULTRA-HIGH-FREQUENCY OSCILLATORS

Linear circuits — At ultra-high frequencies tube interelectrode capacities become of increasing importance, so that eventually the shortest possible straight wire connection between elements, in conjunction with internal leads and capacities, represents the highest possible frequency to which the tube can be tuned. The tube usually will not oscillate up to this limit because of dielectric losses in the seals and other loading effects (§ 7-6). With most small tubes of ordinary construction the upper limit of oscillation is in the region of 150 Mc.; for higher frequencies it is necessary to use special u.h.f. tubes having low interelectrode capacities and low internal lead inductance. Only a few types are capable of developing more than a few watts at frequencies of 300 Mc. and higher.

Although ordinary coil and condenser tank circuits can be used at frequencies as high as 112 Mc., the Q of such circuits is low at ultra-high frequencies because of increased losses, so that both stability and efficiency are poor. For this reason special tank circuits of the linear type (§ 2-12) are preferable. These may be any multiple of a quarter wave in length, the Q increasing with the number of quarter waves. The quarter-wave line is generally used, however, because of the considerable space required for longer lines. At 112 Mc. it is also possible to build high-Q tank circuits with lumped constants, not in the form of ordinary coils and condensers but with large conducting surfaces to reduce resistance to the lowest possible value.

The oscillator circuits used are the same in principle as on the lower frequencies (§ 3-7) although frequently modified considerably to compensate for inherent capacities and inductances which are negligible at lower frequencies.

Two-conductor lines — The quarter-wave two-conductor open line is equivalent to a resonant circuit (§ 2-12) and can be used as the tank circuit (§ 3-7) in an oscillator. It should be used as a balanced circuit to avoid unequal currents in the two conductors and consequent loss of Q because of radiation.

A typical oscillator circuit of the ultraudion type is shown in Fig. 428. The resonant line is usually constructed of copper tubing to give a large conducting surface and hence reduce resistance, and also to make a mechanically-stable circuit and thus minimize the effects of vibration on the oscillator frequency. The line should be approximately a quarter wavelength long, although the resonant frequency will decrease somewhat when the tube with its internal capacities is connected across it so that a somewhat shorter length is ordinarily sufficient. The frequency can be changed by means of the shorting bar, which can be moved along the line to change its effective length.

The tube elements preferably should be tapped down on the line as shown to reduce the loading effect and thus prevent an undue decrease in Q. In general, these taps should be as close to the shorted end of the line as is con-

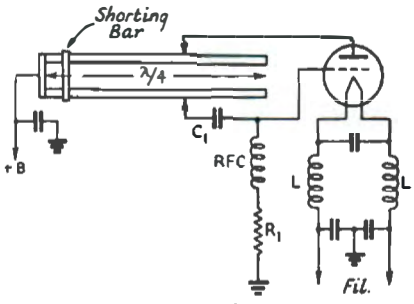


Fig. 428 — Single-tube line oscillator. The grid condenser, C_1, may be 50 $\mu\mu$fd.; grid leak, R_1, 5000 to 50,000 ohms depending upon the type of tube. The choke, RFC, will in general consist of relatively few turns (20 to 50) wound to a diameter of ¼ inch, although dimensions will change considerably with the frequency. By-pass condensers should be small in size to reduce lead inductance; 500 $\mu\mu$fd. is a satisfactory value.

sistent with reliable operation and satisfactory power output, since the frequency stability will be better under these conditions.

The coils (L) in the filament circuit are frequently required at 112 Mc. and higher to compensate for the effects of the inductance of connecting leads, which in many cases are long enough to cause an appreciable phase shift (§ 2-7) which reduces the oscillator efficiency. The effective length of the filament circuit to the points of connection to the lines should be approximately ½ wavelength to bring the filament to the same potential as the shorted ends of the lines. The proper inductance must be determined by experiment, the coils being adjusted until optimum stability and power output are obtained.

The oscillation frequency may also be adjusted by connecting a low-capacity variable condenser across the open end of the line. The added capacity makes it necessary to shorten the line considerably for a given frequency, however, and this together with the additional loss in the condenser causes a marked decrease in the Q of the line. These effects will be less if the condenser is connected down on the line rather than at the open end. Tapping down also gives a greater band-spread tuning effect (§ 7-7).

Push-pull oscillators — It is often advantageous to use a push-pull oscillator circuit at ultra-high frequencies, not only as a means to secure more power output than can be obtained from one tube but also because better circuit symmetry is possible with open lines. Fig. 429 shows a typical push-pull circuit of the t.p.t.g. (§ 3-7) type. The grid line is usually operated as the frequency-controlling circuit since it is not associated with the load and hence its Q can be kept high. The same adjustment considerations apply as in the case of the single-tube oscillator described in the preceding paragraph. The grid taps in particular should be tapped down as far as possible, thus improving the frequency stability.

It is also possible to use a linear tank in the grid circuit for frequency control in conjunction with a conventional coil-condenser tank in the plate circuit, where the lower Q does not have so great an effect on the stability.

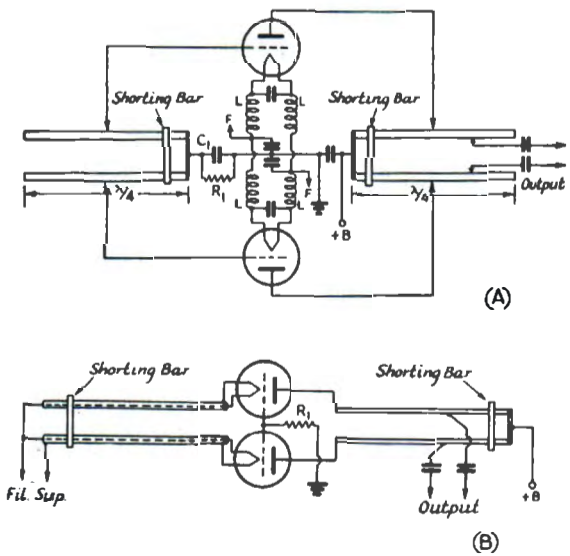


Fig. 429 — Push-pull line oscillator circuits. See Fig. 428 and text for discussion of circuit constants.

Fig. 429-B shows a push-pull oscillator having tuned plate and cathode circuits, using linear tanks for each. The grids are connected together and grounded through the grid leak, R_1; ordinarily no by-pass condenser is needed across R_1. This circuit gives good power output at ultra-high frequencies, but is not especially stable unless the plates are tapped down on the plate tank circuit to avoid too great a reduction in Q. Tapping on the cathode line is not feasible for mechanical reasons, since one filament lead must be brought through the tubing in order to maintain both sides of the filament at the same r.f. potential.

Concentric-line circuits — At frequencies in the neighborhood of 300 Mc. radiation (§ 2-12) from the open line becomes so serious that the Q is greatly reduced. This is because the conductor spacing represents an appreciable fraction of the wavelength.

Consequently at these frequencies the concentric line must be used. In this type the field is confined inside the line so that radiation is negligible; there is a further advantage in that the outside of the line is "cold"; that is, no r.f. potentials develop between points on the outer surface. The concentric line also is advantageous at lower frequencies, but as it is more complicated to construct and length adjustment and tapping both are difficult mechanically, the open lines are generally favored.

The concentric line is usually constructed of copper pipes arranged concentrically and short-circuited at one end. The optimum ratio of inner diameter of the outer conductor to the outer diameter of the inner conductor is 3.6. Taps are usually made on the inner conductor and brought through a hole in the outer conductor to the tube element, as shown in Fig. 430. The tube loads the line in the same way as described in the preceding paragraphs, hence the length is generally shorter than an actual quarter wavelength. The length can be adjusted by a sliding short-circuiting disc at the closed end, a close fit and low-resistance contact being necessary to avoid reduction of the Q. It is also possible to make the inner conductor a pair of close-fitting concentric tubes so that one may be slid in and out of the other to change the effective conductor length.

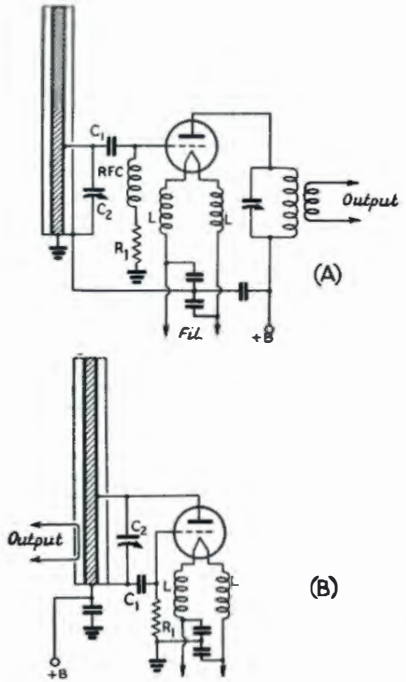


Fig. 430 — Concentric-line oscillator circuits. The line, usually of tubular conductors, is shown in cross-section. See Fig. 428 and text for discussion of circuit constants.

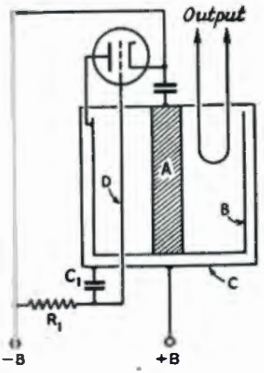


Fig. 431 — High-Q lumped-constant tank circuit in a u.h.f. oscillator. This drawing shows a cross-section of the tank, which is usually built of concentric cylinders. C_1 and R_1 are the grid condenser and leak, respectively; see Fig. 428 for discussion of circuit constants.

The circuit of Fig. 430-A is a t.p.t.g. (§ 3-7) oscillator using the concentric line in the grid circuit for frequency control. An ordinary coil-condenser tank is shown in the plate circuit, but a linear tank may be substituted. The filament inductances have the same function as in the preceding circuits. The ultraudion circuit is shown at B; the same considerations apply. In this case the output is taken from the line inductively by means of the half-turn "hairpin" shown; coupling can be changed to some extent by varying the position of the hairpin. Both circuits may be tuned by means of the small variable condenser C_2, although this condenser may be omitted and the tuning accomplished by changing the line length.

For ease of construction, the concentric line is sometimes modified into a "trough," in which the cross-section of the outer conductor is in the shape of a square "U," one side being left open for tapping and adjustment of the inner conductor. Some radiation takes place with this construction, although not as much as with open lines.

High-Q circuits with lumped constants — To obtain reasonably high effective Q when a low resistance is connected across the tank circuit it is necessary to use a high C/L ratio and a tank of inherently high Q (§ 2-10). At low frequencies the inherent Q of any well-designed circuit will be high enough so that it may be neglected in comparison to the effective Q when loaded, so that no special precautions have to be taken with respect to the resistance of coils and condensers. At ultra-high frequencies these internal resistances are too large to be ignored, and a reduction of the

L/C ratio will not increase the effective Q unless the internal resistance of the tank can be made very small. The reduction in resistance can be brought about by use of large conducting surfaces and elimination of radiation. In such cases the inductance and capacity are generally built as a unit; several arrangements are possible, one being shown in Fig. 431. The tank circuit consists of a rod A (the inductance) inside two concentric cylinders B and C which form a two-plate condenser, one plate being connected to each end of the inductance. The resonant frequency is determined by the length and diameter of A, and the length, diameter and spacing of B and C. The oscillator shown uses the tickler circuit (§ 3-7) with the feedback coil in the grid circuit; this inductance is the wire D in the diagram. Output is taken from the tank circuit by means of the hairpin coupling coil. The tank circuit may also be used in the ultraudion circuit, replacing the concentric line in Fig. 430-B. A variable condenser may be connected across the tank for tuning, if desired, although the Q may be reduced if a considerable portion of the tank r.f. current flows through it.

This type of circuit actually has lumped constants only when the length is small (10% or less) in proportion to the wavelength. At greater lengths it tends to act as a linear circuit, eventually evolving into the concentric line.

Radiotelephony

● 5-1 MODULATION

The carrier — The steady radio-frequency power generated by transmitting circuits cannot alone result in the transmission of an intelligible message to a receiving point. It serves only as a "carrier" for the message; the intelligence is conveyed by *modulation* (a change) of the carrier. In radiotelephony this modulation reproduces electrically the sounds it is intended to convey.

Sound and alternating currents — Sounds are caused by vibrations of air particles. The pitch of the sound depends upon the rate of vibration; the more rapid the vibration the higher the pitch. Most sounds consist of complex combinations of vibrations of differing rates or frequencies; the human voice, for instance, generates frequencies from about 100 per second to several thousand per second. The problem of transmitting speech by radio is therefore one of varying the r.f. carrier in a way which corresponds to the air-particle vibrations. The first step in doing this is to change the sound vibrations into alternating electrical currents of the same frequency and relative intensity; the electromechanical device which achieves this translation is the *microphone*. These currents may then be amplified and used to modulate the normally-steady r.f. output of the transmitter.

Methods of modulation — The carrier may be made to vary in accordance with the speech current by using the current to change the phase (§ 2-7) frequency or amplitude of the carrier. *Amplitude modulation* is by far the most common system, and is used exclusively on all frequencies below the ultra-high-frequency region (§ 2-7). *Frequency modulation*, which has special characteristics which make its use desirable under certain conditions, is used to a considerable extent on ultra-high frequencies. *Phase modulation*, which is closely related to frequency modulation, has had little or no direct application in practical communication.

● 5-2 AMPLITUDE MODULATION

Carrier requirements — For proper amplitude modulation, the carrier should be completely free from inherent amplitude variations such as might be caused by insufficient filtering of a rectified-a.c. power supply (§ 8-4). It is also essential that the carrier *frequency* be entirely unaffected by the application of modulation. If modulating the amplitude of the carrier also causes a change in the carrier frequency, the signal wobbles back and forth with the modulation, introducing distortion and widening the channel taken by the signal. This causes unnecessary interference to other transmissions. In practice, this undesirable frequency modulation is prevented by applying the modulation to an r.f. amplifier stage which is isolated from the frequency-controlling oscillator by a "buffer" amplifier. Amplitude modulation of an oscillator is almost always accompanied by frequency modulation. It is permitted on ultra-high frequencies above 112 Mc. because the problem of interference is less acute than on lower frequencies.

Percentage of modulation — In the amplitude-modulation system the audible output at the receiver depends entirely upon the amount of variation — termed *depth of modulation* — in the carrier wave and not upon the strength of the carrier alone. It is therefore desirable to obtain the largest permissible variations in the carrier wave. This condition is reached when the carrier amplitude during modulation is at times reduced to zero and at other times increased to twice its unmodulated value. Such a wave is said to be *fully modulated*, or *100% modulated*. Any desired degree of modulation can be expressed as a percentage, using the unmodulated carrier as a base. Fig. 501 shows at A an unmodulated carrier wave; at B the same wave modulated 50%, and at C the wave with 100% modulation, using a sine-wave (§ 2-7) modulating signal. The outline of the modulated r.f. wave is called the *modulation envelope*.

The percentage modulation can be found by dividing either Y or Z by X and multiplying the result by 100. If the modulating signal is not symmetrical, the larger of the two (Y or Z) should be used.

Power in modulated wave — The amplitude values correspond to current or voltage, so that the drawings may be taken to represent instantaneous values of either. Since power varies as the square of either the current or voltage (so long as the resistance in the circuit is unchanged), at the peak of the modulation up-swing the instantaneous power in the wave

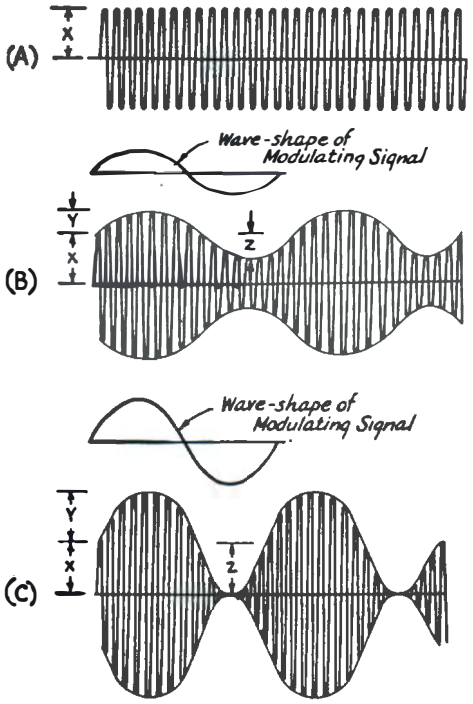


Fig. 501 — Graphical representation of (A) unmodulated carrier wave, (B) wave modulated 50%, (C) wave modulated 100%.

of Fig. 501-C is four times the unmodulated carrier power. At the peak of the down-swing the power is zero since the amplitude is zero. With a sine-wave modulating signal, the *average* power in a 100%-modulated wave is one and one-half times the unmodulated carrier power: that is, the power output of the transmitter increases 50% with 100% modulation.

Linearity — Up to the limit of 100% modulation, the amplitude of the carrier should follow faithfully the amplitude variations of the modulating signal. When the modulated r.f. amplifier is incapable of meeting this condition it is said to be *non-linear*. The amplifier may not, for instance, be capable of quadrupling its power output at the peak of 100% modulation. A non-linear modulated amplifier causes distortion of the modulation envelope.

Modulation characteristic — A graph showing the relationship between r.f. amplitude and instantaneous modulating voltage is called the *modulation characteristic* of the modulated amplifier. This graph should be a straight line (linear) between the limits of zero and twice carrier amplitude. Curvature of the line between these limits indicates non-linearity.

Modulation capability — The *modulation capability* of the transmitter is the maximum percentage of modulation that is possible without objectionable distortion from non-linearity. The maximum capability is, of course, 100%. The modulation capability should be as high as possible so that the most effective signal can be transmitted for a given carrier power.

Overmodulation — If the carrier is modulated more than 100%, a condition such as is shown in Fig. 502 occurs. Not only does the peak amplitude exceed twice the carrier amplitude, but there may actually be a considerable period during which the output is entirely cut off. The modulated wave is therefore distorted (§ 3-3) with the result that harmonics of the audio modulating frequency appear. The carrier should never be modulated more than 100%.

Sidebands — The combining of the audio frequency with the r.f. carrier is essentially a heterodyne process and therefore gives rise to beat frequencies equal to the sum and difference of the a.f. and r.f. frequencies involved (§ 2-13). Therefore, for each audio frequency appearing in the modulating signal two new radio frequencies appear, one equal to the carrier frequency plus the audio frequency, the other equal to the carrier minus the audio frequency. These new frequencies are called *side frequencies*, since they appear on each side of the carrier, and the groups of side frequencies representing a band or group of modulation frequencies are called *sidebands*. Hence a modulated signal occupies a group of radio frequencies, or *channel*, rather than a single frequency as in the case of the unmodulated carrier. The *channel width* is twice the highest modulation frequency. To accommodate the largest number of transmitters in a

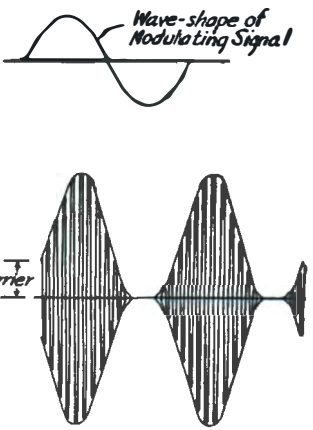


Fig. 502 — An overmodulated wave.

given part of the r.f. spectrum it is apparent that the channel width should be as small as possible, but on the other hand it is necessary, for speech of reasonably good quality, to use modulating frequencies up to about 3000 or 4000 cycles. This calls for a channel width of 6 to 8 kc.

Spurious side bands — Besides the normal side bands required by speech frequencies, unwanted side bands may be generated by the transmitter. These usually lie outside the normally-required channel and hence cause it to be wider without increasing the useful modulation. By increasing the channel width these spurious side bands cause unnecessary interference to other transmitters. The quality of transmission is also adversely affected when spurious side bands are generated.

The chief causes of spurious side bands are harmonic distortion in the audio system, overmodulation, unnecessary frequency modulation, and lack of linearity in the modulated r.f. system.

Types of amplitude modulation — The most widely used type of amplitude modulation system is that in which the modulating signal is applied in the plate circuit of a radio-frequency power amplifier (*plate modulation*). In a second type the audio signal is applied to a control-grid circuit (*grid-bias modulation*). A third system involves variation of both plate voltage and grid bias and is called *cathode modulation*.

●5-3 PLATE MODULATION

Transformer coupling — In Fig. 503 is shown the most widely-used system of plate modulation. A balanced (push-pull Class-A, Class-AB or Class-B) modulator is transformer-coupled to the plate circuit of the modulated r.f. amplifier. The audio-frequency power generated in the modulator plate circuit is combined with the d.c. power in the modulated-amplifier plate circuit by transfer through the coupling transformer, *T*. For 100% modulation the audio-frequency output of the modulator and the turns ratio of the coupling transformer must be such that the voltage at the plate of the modulated amplifier varies between zero and twice the d.c. operating plate voltage, thus causing corresponding variations in the amplitude of the r.f. output.

Modulator power — The average power output of the modulated stage must increase 50% for 100% modulation (§ 5-2), so that the modulator must supply audio power equal to 50% of the d.c. plate input to the modulated r.f. stage. For example, if the d.c. plate power input to the r.f. stage is 100 watts, the sine-wave audio power output of the modulator must be 50 watts.

Modulating impedance, linearity — The *modulating impedance* or load resistance presented to the modulator by the modulated r.f. amplifier, is equal to

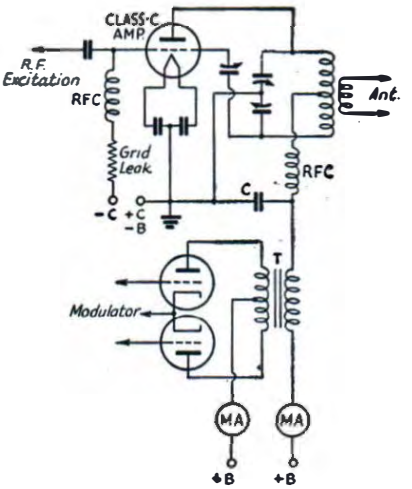


Fig. 503 — Plate modulation of a Class-C r.f. amplifier. The plate by-pass condenser, C, in the r.f. stage should have high reactance at audio frequencies. A value of 0.002 μfd. or less is usually satisfactory.

$$\frac{E_b}{I_p} \times 1000$$

where E_b is the d.c. plate voltage and I_p the d.c. plate current in milliamperes, both measured without modulation.

Since the power output of the r.f. amplifier must vary as the square of the plate voltage (r.f. voltage proportional to applied plate voltage) in order for the modulation to be linear, the amplifier must operate Class-C (§ 3-4). The linearity depends upon having sufficient grid excitation, proper bias, and adjustment of circuit constants to the proper values (§ 4-8).

Power in speech waves — The complex waveform of a speech sound translated into alternating current does not contain as much power, on the average, as there is in a pure tone or sine wave of the same peak (§ 2-7) amplitude. That is, with speech waveforms the ratio of peak to average amplitude is higher than in the sine wave. For this reason, the previous statement that the power output of the transmitter increases 50% with 100% modulation, while true for tone modulation, is not true for speech. On the average, speech waveforms will contain only about half as much power as a sine wave, both having the same peak amplitude. The average power output of the transmitter therefore increases only about 25% with 100% speech modulation. However, the *instantaneous* power output must quadruple on the peak of 100% modulation

(§ 5-2) regardless of the modulating waveform. Therefore the peak capacity of the transmitter must be the same for any type of modulating signal.

Adjustment of plate-modulated amplifiers — The general operating conditions for Class-C operation have been described (§ 3-4, 4-8). The grid bias and grid current required for plate modulation are usually given in the operating data supplied by the tube manufacturer; in general, the bias should be such as to give an operating angle (§ 4-8) of about 120 degrees at carrier plate voltage, and the excitation should be sufficient to maintain the plate efficiency constant when the plate voltage is varied over the range from zero to twice the d.c. plate voltage applied to the amplifier. For best linearity, the grid bias should be obtained partly from a fixed source of about the cut-off value supplemented by grid-leak bias to supply the remainder of the required operating bias.

The maximum permissible d.c. plate power input for 100% modulation is twice the sine-wave audio-frequency power output of the modulator. This input is obtained by varying the loading on the amplifier (keeping its tank circuit tuned to resonance) until the product of d.c. plate voltage and plate current is the desired power. The modulating impedance under these conditions will be the proper value for the modulator if the proper output transformer turn ratio (§ 2-9) is used.

Neutralization, when triodes are used, should be as nearly perfect as possible, since regeneration may cause non-linearity. The amplifier also should be free from parasitic oscillations (§ 4-10).

Although the *effective* value (§ 2-7) of power input increases with modulation, as described above, the *average* plate input to a plate-modulated amplifier does not change, since each increase in plate voltage and plate current is balanced by an equivalent decrease in voltage and current. Consequently the d.c. plate current to a properly-modulated amplifier is constant with or without modulation.

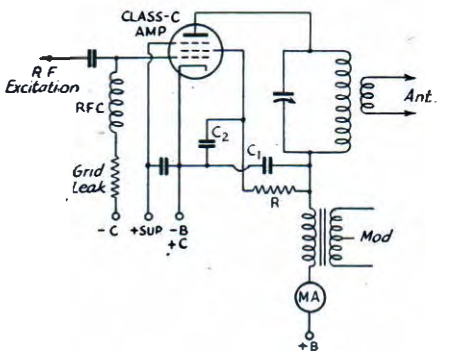


Fig. 504 — Plate and screen modulation of a pentode Class-C r.f. amplifier. Plate and screen by-pass condensers, C_1 and C_2, should have high reactance at audio frequencies (0.002 μfd. or less).

Screen-grid amplifiers — Screen-grid tubes of the pentode or beam tetrode type can be used as Class-C plate-modulated amplifiers provided the modulation is applied to both the plate and screen grid. The method of feeding the screen grid with the necessary d.c. and modulation voltage is shown in Fig. 504. The dropping resistor, *R*, should be of the proper value to apply normal d.c. voltage to the screen under steady carrier conditions. Its value can be calculated by taking the difference between plate and screen voltages and dividing it by the rated screen current.

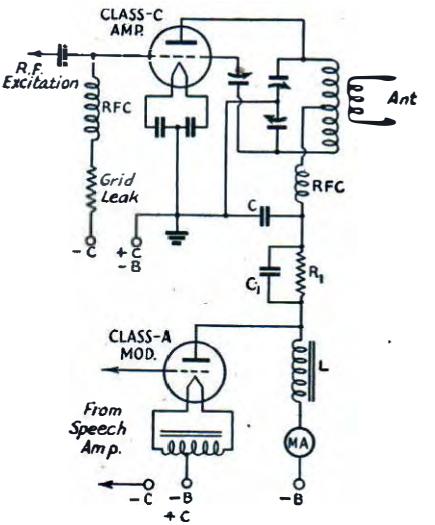


Fig. 505 — Choke-coupled plate modulation.

The modulating impedance is found by dividing the d.c. plate voltage by the sum of the plate and screen currents. The plate voltage multiplied by the sum of the two currents is the power input figure which is used as the basis for determining the audio power required from the modulator.

Choke coupling — In Fig. 505 is shown the circuit of the choke-coupled system of plate modulation. The plate power for the modulator tube and modulated amplifier is furnished from a common source through the modulation choke, *L*, which has high impedance for audio frequencies. The modulator operates as a power amplifier with the plate circuit of the r.f. amplifier as its load, the audio output of the modulator being superimposed on the d.c. power supplied to the amplifier. For 100% modulation the audio voltage applied to the r.f. amplifier plate circuit across the choke, *L*,

must have a peak value equal to the d.c. voltage on the modulated amplifier. To obtain this without distortion, the r.f. amplifier must be operated at a d.c. plate voltage less than the modulator plate voltage, the extent of the voltage difference being determined by the type of modulator tube used. The necessary drop in voltage is provided by the resistor R_1, which is by-passed for audio frequencies by the condenser C_1.

This type of modulation is seldom used except in very low-power portable sets, because a single-tube Class-A (§ 3-4) modulator is required. The output of a Class-A modulator is very low compared to that obtainable from a pair of tubes of the same size operated Class-B, hence only a small amount of r.f. power can be modulated.

● 5-4 GRID-BIAS MODULATION

Circuit — Fig. 506 is the diagram of a typical arrangement for grid-bias modulation. In this system the secondary of an audio-frequency output transformer, the primary of which is connected in the plate circuit of the modulator tube, is connected in series with the grid-bias supply for the modulated amplifier. The audio voltage thus introduced varies the grid bias and thus the power output of the r.f. stage, when suitable operating conditions are chosen. The r.f. stage is operated as a Class-C amplifier, with the d.c. grid bias considerably beyond cut-off.

Operating principles — In this system the plate voltage is constant, and the increase in power output with modulation is obtained by making the plate current and plate efficiency vary with the modulating signal. For 100% modulation, both plate current and efficiency must, at the peak of the modulation up-swing, be twice their carrier values so that the peak power will be four times the carrier power. Since the peak efficiency in practicable circuits is of the order of 70% to 80%, the carrier efficiency ordinarily cannot exceed about 35% to 40%. For a given size of r.f. tube the carrier output is about one-fourth the carrier obtainable from the same tube plate-modulated. The grid bias, r.f. excitation, plate loading and audio voltage in series with the grid must be adjusted to give a linear modulation characteristic.

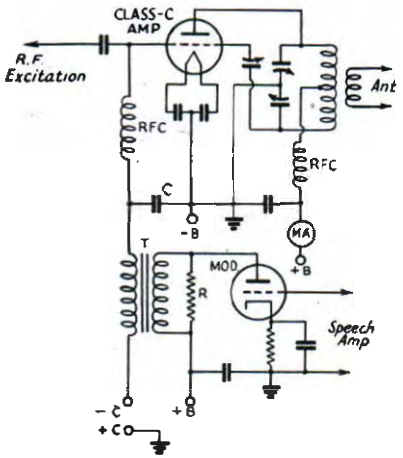


Fig. 506 — Grid-bias modulation of a Class-C amplifier. The r.f. grid by-pass condenser, C, should have high reactance at audio frequencies (0.002 μfd. or less in usual cases).

Modulator power — Since the increase in average carrier power with modulation is secured by varying the plate efficiency and d.c. plate input of the amplifier, the modulator need only supply such power losses as may be occasioned by connecting it in the grid circuit. These are quite small, hence a modulator capable of only a few watts output will suffice for transmitters of considerable power. The load on the modulator varies over the audio-frequency cycle as the rectified grid current of the modulated amplifier changes, hence the modulator should have good voltage regulation (§ 5-6).

Grid-bias source — The change in bias voltage with modulation causes the rectified grid current of the amplifier also to vary, the r.f. excitation being fixed. If the bias source has appreciable resistance, the change in grid current also will cause a change in bias in a direction opposite to that caused by the modulation. It is therefore necessary to use a grid-bias source having low resistance so that these bias variations will be negligible. Battery bias is satisfactory. If a rectified a.c. bias supply is used the type having regulated output (§ 8-9) should be used. Grid-leak bias for a grid-modulated amplifier is unsatisfactory and its use should not be attempted.

Driver regulation — The load on the driving stage varies with modulation, and a linear modulation characteristic may not be obtained if the r.f. voltage from the driver does not stay constant with changes in load. Driver regulation (ability to maintain constant output voltage with changes in load) may be improved by using a driving stage having two or three times the power output necessary for excitation of the amplifier (this is somewhat less than the power required for ordinary Class-C operation), and by dissipating the extra power in a constant load such as a resistor. The load variations are thereby reduced in proportion to the total load.

Adjustment of grid-bias modulated amplifiers — This type of amplifier should be adjusted with the aid of an oscilloscope to obtain optimum operating conditions. The

oscilloscope should be connected as described in § 5-10, the wedge pattern being preferable. A tone source for modulating the transmitter will be convenient. The fixed grid bias should be two or three times the cut-off value (§ 3-2). The d.c. input to the amplifier, assuming 33% carrier efficiency, will be 1½ times the plate dissipation rating of the tube or tubes used in the modulated stage, and the plate current for this input (in milliamperes, $1000P/E$, where P is the power and E the d.c. plate voltage) determined. Apply r.f. excitation and, without modulation, adjust the plate loading (keeping the plate tank circuit tuned to resonance) to give the required plate current. Next, apply modulation and increase in the modulating signal until the modulation characteristic shows curvature (§ 5-10). This will probably occur well below 100% modulation, indicating that the plate efficiency is too high. Increase the plate loading and reduce the excitation to maintain the same plate current, apply modulation and check the characteristic again. Continue this process until the characteristic is linear from the axis to twice the carrier amplitude. It is advantageous to use the maximum permissible plate voltage on the tube, since it is usually easier to obtain a more linear characteristic with high plate voltage and low current (carrier conditions) rather than with relatively low plate voltage and high plate current.

The amplifier can be adjusted without an oscilloscope by determining the plate current as described above, then setting the bias to the cut-off value (or slightly beyond) for the d.c. plate voltage used and applying maximum excitation. Adjust the plate loading, keeping the tank circuit at resonance, until the amplifier draws twice the carrier plate current, and note the antenna current. Decrease the excitation until the output and plate current just start to drop, then increase the bias, leaving the excitation and plate loading unchanged, until the plate current drops to the proper carrier value. The antenna current should be just half the previous value; if it is larger, try somewhat more loading and less excitation; if smaller, less loading and more excitation. Repeat until the antenna current drops to half its maximum value when the plate current is biased down to the carrier value. Under these conditions the amplifier should modulate properly, provided the plate supply has good voltage regulation (§ 8-1) so that the plate voltage is practically the same at both values of plate current during the initial testing.

The d.c. plate current should be substantially constant with or without modulation (§ 5-3).

Suppressor modulation — The circuit arrangement for suppressor-grid modulation of a pentode tube is shown in Fig. 507. The operating principles are the same as for grid-bias modulation. However, the r.f. excitation and modulating signals are applied to separate grids, which gives the system a simpler operating technique, since best adjustment for proper excitation requirements and proper modulating circuit requirements are more or less independent. The carrier plate efficiency is approximately the same as for grid-bias modulation, and the modulator power requirements are similarly small. With tubes having suitable suppressor-grid characteristics, linear modulation up to practically 100% can be obtained with negligible distortion.

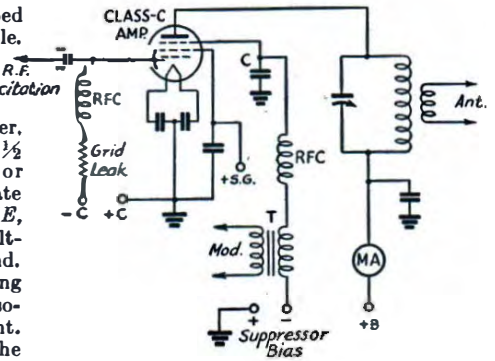


Fig. 507 — Suppressor-grid modulation of a pentode r.f. amplifier. The suppressor r.f. by-pass condenser, C, should be 0.002 μfd. or less.

The method of adjustment is essentially the same as that described in the preceding paragraph. Apply normal excitation and bias to the control grid and, with the suppressor bias at zero or the positive value recommended for c.w. telegraph operation with the particular tube used, adjust the plate loading to obtain twice the carrier plate current (on the basis of 33% carrier efficiency). Then apply sufficient negative bias to the suppressor to bring the plate current to the carrier value, leaving the loading unchanged. Simultaneously, the antenna current also should drop to half its maximum value. The amplifier is then ready for modulation. Should the plate current not follow the antenna current in the same proportion when the suppressor bias is made negative, the loading and excitation should be readjusted to make them coincide.

• 5-5 CATHODE MODULATION

Circuit — The fundamental circuit for cathode or "center-tap" modulation is shown in Fig. 508. This type of modulation is a combination of the plate- and grid-bias methods, and permits a carrier efficiency midway between the two. The audio power is introduced

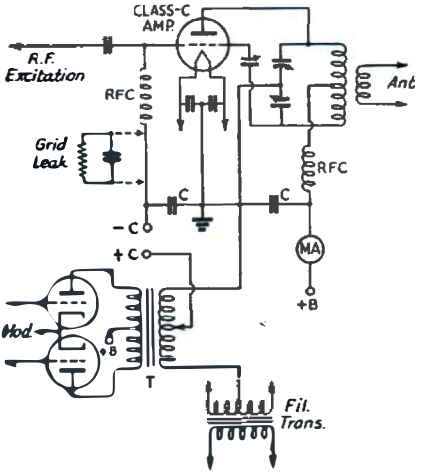


Fig. 508 — Cathode modulation of a Class-C r.f. amplifier. The grid and plate by-pass condensers, C, should be 0.002 μfd. or less (bigh reactance at audio frequencies).

in the cathode circuit, and both grid bias and plate voltage vary during modulation.

The cathode circuit of the modulated stage must be independent of other stages in the transmitter; that is, when filament-type tubes are modulated they must be supplied from a separate filament transformer. The filament by-pass condensers should not be larger than about 0.002 μfd., to avoid by-passing the audio.

Operating principles — Because part of the modulation is by the grid-bias method, the plate efficiency of the modulated amplifier must vary during modulation. The carrier efficiency therefore must be lower than the efficiency at the modulation peak. The required reduction in carrier efficiency depends upon the proportion of grid modulation to plate modulation; the higher the percentage of plate modulation the higher the permissible carrier efficiency, and vice versa. The audio power required from the modulator also varies with the percentage of plate modulation, being greater as this percentage is increased.

The way in which the various quantities vary is illustrated by the curves of Fig. 509. In these curves, the performance of the cathode-modulated r.f. amplifier is plotted in terms of the tube ratings for plate-modulated telephony, with the percentage of plate modulation as a base. As the percentage of plate modulation is decreased, it is assumed that the grid-bias modulation is increased to make the overall percentage of modulation reach 100%. The limiting condition, 100% plate modulation and no grid-bias modulation, is at the right (A); pure grid-bias modulation is represented by the left-hand ordinate (B and C).

As an example, assume that 40% plate modulation is to be used. Then the modulated r.f. amplifier must be adjusted for a carrier plate efficiency of 56%, the permissible plate input will be 65% of the ratings of the same tube with pure plate modulation, the power output will be 48% of the rated output of the tube with plate modulation, and the audio power required from the modulator will be 20% of the d.c. input to the modulated amplifier.

Modulating impedance — The modulating impedance of a cathode-modulated amplifier is approximately equal to

$$m\frac{E_b}{I_b}$$

where m is the percentage of plate modulation expressed as a decimal, E_b is the plate voltage, and I_b the plate current of the modulated r.f. amplifier. This figure for the modulating impedance is used in the same way as the corresponding figure for pure plate modulation in determining the proper modulator operating conditions (§ 5-6).

Conditions for linearity — R.f. excitation requirements for the cathode-modulated amplifier are midway between those for plate modulation and grid-bias modulation. More excitation is required as the percentage of plate modulation is increased. Grid bias should be considerably beyond cut-off; fixed bias from a supply having good voltage regulation (§ 8-9) is preferred, especially when the percentage of plate modulation is small and the amplifier is operating more nearly like a grid-bias modulated stage. At the higher percent-

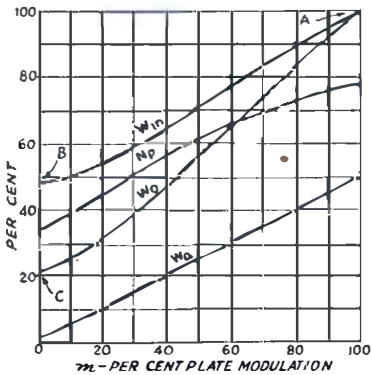


Fig. 509 — Cathode modulation performance curves, in terms of percentage of plate modulation against per cent of Class-C telephony tube ratings.

W_{in} — D.c. plate input watts in per cent of plate-modulation rating.
W_o — Carrier output watts in per cent of plate-modulation rating (based on plate efficiency of 77.5%).
W_a — Audio power in per cent of d.c. watts input.
N_p — Plate efficiency in per cent.

ages of plate modulation a combination of fixed and grid-leak bias can be used since the variation in rectified grid current is smaller. The grid-leak should be by-passed for audio frequencies. The percentage of grid modulation may be regulated by choice of a suitable tap on the modulation transformer secondary.

Adjustment of cathode-modulated amplifiers — In most respects the adjustment procedure is similar to that for grid-bias modulation (§ 5-4). The critical adjustments are those of antenna loading, grid bias, and excitation. The proportion of grid-bias to plate modulation will determine the operating conditions.

Adjustments should be made with the aid of an oscilloscope (§ 5-10). With proper antenna loading and excitation, the normal wedge-shaped pattern will be obtained at 100% modulation. As in the case of grid-bias modulation too-light antenna loading will cause flattening of the up-peaks of modulation (downward modulation), as will also too-high excitation (§ 5-10). The cathode current will be practically constant with or without modulation when the proper operating conditions are reached (§ 5-3).

• 5-6 CLASS-B MODULATORS

Modulator tubes — In the case of plate modulation, the relatively-large audio power needed (§ 5-3) practically dictates the use of a Class-B (§ 3-4) modulator, since the power can be obtained most economically with this type of amplifier. A typical circuit is given in Fig. 510. A pair of tubes must be chosen which is capable of delivering sine-wave audio power equal to half the d.c. input to the modulated Class-C amplifier. It is sometimes convenient to use tubes which will operate at the same plate voltage as that applied to the Class-C stage, since one power supply of adequate current capacity may then suffice for both stages. Available components do not always permit this, however, and better overall performance and economy may frequently result from the use of separate power supplies.

Matching to load — In giving Class-B ratings on power tubes, manufacturers specify the plate-to-plate load impedance (§ 3-3) into which the tubes must operate to deliver the rated audio power output. This load impedance seldom is the same as the modulating impedance (§ 5-3) of the Class-C r.f. stage, so that a match must be brought about by adjusting the turn ratio of the coupling transformer. The required turn ratio, primary to secondary, is

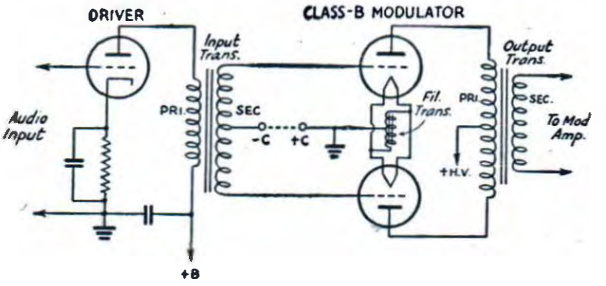


Fig. 510 — Class-B modulator and driver circuit.

$$\sqrt{\frac{Z_p}{Z_m}}$$

where Z_m is the Class-C modulating impedance and Z_p is the plate-to-plate load impedance specified for the Class-B tubes.

Commercial Class-B output transformers usually are rated to work between specified primary and secondary impedances and are designed for specific Class-B tubes. In such a case the turn ratio can be found by substituting the given impedances in the formula above. Many transformers are provided with primary and secondary taps so that various turn ratios can be obtained to meet the requirements of a large number of tube combinations.

Driving power — Class-B amplifiers are driven into the grid-current region, so that power is consumed in the grid circuit (§ 3-3). The preceding stage (*driver*) must be capable of supplying this power at the required peak audio-frequency grid-to-grid voltage. Both these quantities are given in the manufacturer's tube ratings. The grids of the Class-B tubes represent a variable load resistance over the audio-frequency cycle, since the grid current does not increase directly with the grid voltage. To prevent distortion, therefore, it is necessary to have a driving source which has good *regulation* — that is, which will maintain the waveform of the signal without distortion even though the load varies. This can be brought about by using a driver capable of delivering two or three times the actual power consumed by the Class-B grids, and by using an input coupling transformer having a turn ratio giving the largest step-down in voltage, between the driver plate or plates and Class-B grids, that will permit obtaining the specified grid-to-grid a.f. voltage.

Driver coupling — A Class-A or Class-AB (§ 3-4) driver is used to excite a Class-B stage. Tubes for the driver preferably should be triodes having low plate resistance, since these will have the best regulation. Having chosen a tube or tubes with ample power out-

put from tube data sheets, the peak output voltage will be, approximately,

$$E_o = 1.4\sqrt{PR}$$

where P is the power output and R the load resistance. The input transformer ratio, primary to secondary, will be

$$\frac{E_o}{E_g}$$

where E_o is as given above and E_g is the peak grid-to-grid voltage required by the modulator tubes.

Commercial transformers usually are designed for specific driver-modulator combinations, and usually are adjusted to give as good driver regulation as the conditions will permit.

Grid bias — Modern Class-B audio tubes are intended for operation without fixed bias. This lessens the variable grid-circuit loading effect and eliminates the need for a grid-bias supply.

When a grid-bias supply is required, it must have low internal resistance so that the flow of grid current with excitation of the Class-B tubes does not cause a continual shift in the actual grid bias and thus cause distortion. Batteries or a regulated bias supply (§ 8-9) should be used.

Plate supply — The plate supply for a Class-B modulator should be sufficiently well filtered (§ 8-3) to prevent hum modulation of the r.f. stage (§ 5-2). An additional requirement is that the output condenser of the supply should have low reactance (§ 2-8) at 100 cycles or less compared to the load into which each tube is working, which is ¼ the plate-to-plate load resistance. A 4-μfd. output condenser with a 1000-volt supply, or a 2-μfd. condenser with a 2000-volt supply, usually will be satisfactory. With other plate voltages, condenser values should be in inverse proportion to the plate voltage.

Overexcitation — When a Class-B amplifier is overdriven in an attempt to secure more than the rated power, distortion in the output waveshape increases rapidly. The high-frequency harmonics which result from the distortion (§ 3-3) modulate the transmitter, producing spurious sidebands (§ 5-2) which readily can cause serious interference over a band of frequencies several times the channel width required for speech. This may happen even though the transmitter is not being overmodulated, as in the case where the modulator is incapable of delivering the power required to modulate the transmitter fully, or when the Class-C amplifier is not adjusted to give the proper modulating impedance (§ 5-3).

The tubes used in the Class-B modulator should be capable of somewhat more than the power output nominally required (50% of the d.c. input to the modulated amplifier) to take care of losses in the output transformer. These usually run from 10% to 20% of the tube output. In addition, the Class-C amplifier should be adjusted to give the proper modulating impedance and the correct output transformer turn ratio should be used. Such high-frequency harmonics as may be generated in these circumstances can be reduced by connecting condensers across the primary and secondary of the output transformer (about 0.002 μfd. in the average case) to form, with the transformer leakage inductance (§ 2-9) a low pass filter (§ 2-11) which cuts off just above the maximum audio frequency required for speech transmission (about 4000 cycles). The condenser voltage ratings should be adequate for the peak a.f. voltages appearing across them.

Operation without load — Excitation should never be applied to a Class-B modulator until after the Class-C amplifier is turned on and is drawing the proper plate current to present the rated load to the modulator. With no load to absorb the power, the primary impedance of the transformer rises to a high value and excessive audio voltages are developed across it — frequently high enough to break down the transformer insulation. If the modulator is to be tested separately from the transmitter a load resistance of the same value as the modulating impedance, and capable of dissipating the full power output of the modulator, should be connected across the transformer secondary.

• 5-7 LOW-LEVEL MODULATORS

Selection of tubes — Modulators for grid-bias and suppressor modulation usually can be small audio power output tubes, since the audio power required is quite small. A triode such as the 2A3 is preferable because of its low plate resistance, but pentodes will work satisfactorily.

Matching to load — Since the ordinary Class-A receiving power tube will develop about 200 to 250 peak volts in its plate circuit, which is ample for most low-level modulator applications, a 1:1 coupling transformer is generally used. If more voltage is required, a step-up ratio must be provided in the transformer. It is usual practice to load the primary of the output coupling transformer with a resistance equal to or slightly higher than the rated load resistance for the tube in order to stabilize the voltage output and thus improve the regulation. This is indicated in Figs. 506 and 507.

• 5-8 MICROPHONES

Sensitivity — The *sensitivity* of a microphone is its electrical output for a given speech intensity input. Sensitivity varies greatly with

microphones of different basic types, and also varies between different models of the same type. The output is also greatly dependent on the character of the individual voice and the distance of the speaker's lips from the microphone, decreasing approximately as the square of the distance. It also may be affected by reverberation in the room. Hence, only approximate values based on averages of "normal" speaking voices can be attempted. The values given in the following paragraphs are based on close talking; that is, with the microphone six inches or less from the speaker's lips.

Frequency response — The *frequency response* of a microphone is its relative ability to convert sounds of different frequencies into alternating current. With fixed sound intensity at the microphone, the electrical output may vary considerably as the sound frequency is varied. For intelligible speech transmission only a limited frequency range is necessary, and natural-sounding speech can be obtained if the output of the microphone does not vary more than a few decibels (§ 3-3) over a range of about 100 cycles to 3000 or 4000 cycles. When the variation in decibels is small between two frequency limits, the microphone is said to be *flat* between those limits.

Carbon microphones — Fig. 511 shows connections for single- and double-button carbon microphones, with a variable potentiometer included in each circuit for adjusting the button current to the correct value as specified with each microphone. The single-button microphone consists of a metal diaphragm placed against an insulating cup containing loosely-packed carbon granules (*microphone button*). Current from a battery flows through the granules, the diaphragm being one connection and the metal back-plate the other. The primary of a transformer is connected in series with the battery and microphone. As the diaphragm vibrates its pressure on the granules alternately increases and decreases, causing a corresponding increase and decrease of current flow through the circuit, since the pressure changes the resistance of the mass of granules. The change in current flowing through the transformer primary causes an alternating voltage, of corresponding frequency and intensity, to be set up in the transformer secondary (§ 2-9). The double-button type operates similarly, but with two buttons in push-pull.

Good quality single-button carbon microphones give outputs ranging from 0.1 to 0.3 volt across 50 to 100 ohms; that is, across the primary winding of the microphone transformer. With the step-up of the transformer, a peak voltage of between 3 and 10 volts across 100,000 ohms or so can be assumed available at the grid of the first tube. These microphones are usually operated with a button current of 50 to 100 ma.

The sensitivity of good-quality double-button microphones is considerably less, ranging from 0.02 volt to 0.07 volt across 200 ohms. With this type microphone and the usual push-pull input transformer, a peak voltage of 0.4 to 0.5 volt across 100,000 ohms or so can be assumed available at the first speech amplifier grid. The button current with this type microphone ranges from 5 to 50 ma. per button.

Crystal microphones — The input circuit for a piezo-electric or crystal type microphone is shown in Fig. 511-E. The element in this type consists of a pair of Rochelle salts crystals cemented together, with plated electrodes. In the more sensitive types the crystal is mechanically coupled to a diaphragm. Sound waves actuating the diaphragm cause the crystal to vibrate mechanically and, by piezo-electric action (§ 2-10), to generate a corresponding alternating voltage between the electrodes, which are connected to the grid circuit of a vacuum tube amplifier as shown. The crystal type requires no separate source of current or voltage.

Although the sensitivity of crystal microphones varies with different models, an output

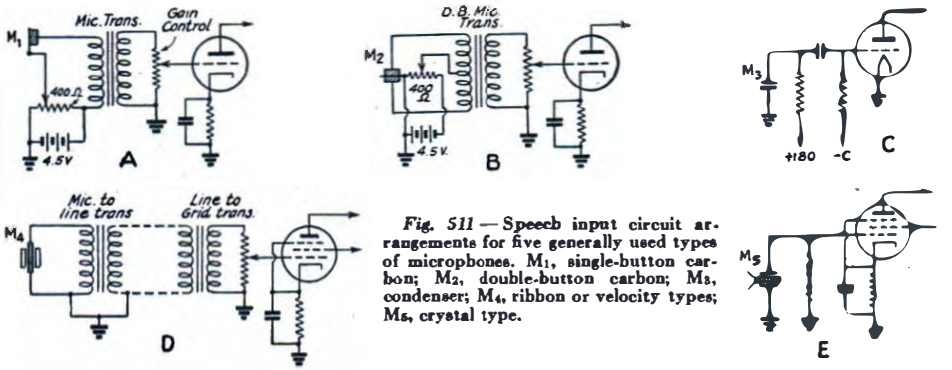


Fig. 511 — Speech input circuit arrangements for five generally used types of microphones. M_1, single-button carbon; M_2, double-button carbon; M_3, condenser; M_4, ribbon or velocity types; M_5, crystal type.

of 0.01 to 0.03 volt is representative for communication types. The sensitivity is affected by the length of the cable connecting to the first amplifier stage; the above figure is for lengths of 6 or 7 feet. The frequency characteristic is unaffected by the cable but the load resistance (amplifier grid resistor) does affect it, the lower frequencies being attenuated as the shunt resistance becomes less. Grid resistor values of 1 megohm and higher should be used, 5 megohms being a customary figure.

Condenser microphones — The condenser microphone of Fig. 511-C consists of a two-plate capacity with one plate stationary and the other, separated from the first by about a thousandth of an inch, a thin metal membrane serving as a diaphragm. This condenser is connected in series with a resistor and d.c. voltage source. When the diaphragm vibrates the change in capacity causes a small charging current to flow through the circuit. The resulting audio voltage which appears across the resistor is fed to the tube grid through the coupling condenser.

The output of condenser microphones varies with different models, the high-quality type being about one-hundredth to one-fiftieth as sensitive as the double-button carbon microphone. The first amplifier tube must be built into the microphone since the capacity of a connecting cable would impair both output and frequency range.

Velocity and dynamic microphones — In a velocity or ribbon microphone, the element acted upon by the sound waves is a thin corrugated metallic ribbon suspended between the poles of a magnet. When made to vibrate the ribbon cuts the lines of force between the poles in first one direction and then the other, thus generating an alternating voltage.

The sensitivity of the velocity microphone, with a suitable coupling transformer, is about 0.03 to 0.05 volt.

The dynamic microphone is similar to the ribbon type in principle, but the ribbon is replaced by a coil attached to a diaphragm. The coil provides several turns of wire cutting the magnetic field, and thus gives greater sensitivity. A small permanent-magnet loud-speaker makes a practical dynamic microphone.

●5-9 THE SPEECH AMPLIFIER

Description — The function of the speech amplifier is to build up the weak microphone voltage to a value sufficient to excite the modulator to the required output. It may have from one to several stages. The last stage nearly always must deliver a certain amount of audio power, especially when it is used to excite a Class-B modulator. Speech amplifiers for grid-bias modulation usually end in a power stage which also functions as the modulator.

The speech amplifier is frequently built as a separate unit from the modulator, and in such a case may be provided with a step-down transformer designed to work into a low impedance, such as 200 or 500 ohms (tube-to-line transformer). When this is done, a step-up input transformer intended to work between the same impedance and the modulator grids (line-to-grid transformer) is provided in the modulator circuit. The line connecting the two transformers may be made any convenient length.

General design considerations — The last stage of the speech amplifier must be selected on the basis of the power output required from it; for instance, the power necessary to drive a Class-B modulator (§ 5-6). It may be either single-ended or push-pull (§ 3-3) the latter being generally preferable because of higher power output and lower harmonic distortion. Push-pull amplifiers may be either Class A, Class AB_1 or Class AB_2 (§ 3-4) as the power requirements dictate. If a Class A or AB_1 amplifier is used, the preceding stages may all be voltage amplifiers, but when a Class-AB_2 amplifier is used the stage immediately preceding it must be capable of furnishing the power consumed by its grids at full output. The requirements in this case are much the same as those which must be met by a driver for a Class-B stage (§ 5-6), but the actual power needed is considerably smaller and usually can be supplied by one or two small receiving triodes. Any lower-level stages are invariably worked as purely voltage amplifiers.

The minimum amplification which must be provided ahead of the last stage is equal to the peak audio-frequency grid voltage required by the last stage for full output (peak grid-to-grid-voltage in the case of a push-pull stage) divided by the output voltage of the microphone or secondary of the microphone transformer if one is used (§ 5-8). The peak a.f. grid voltage required by the output tube or tubes is equal to the d.c. grid bias in the case of a single-tube Class A amplifier, and approximately twice the grid bias for a push-pull Class-A stage. The requisite information for Class AB_1 and AB_2 amplifiers can be obtained from the manufacturer's data on the type considered. If the gain is not obtainable in one stage, several stages must be used in cascade. When the output stage is operated Class AB_2, due allowance must be made for the fact that the next-to-the-last stage must deliver power as well as voltage. In such cases suitable driver combinations are usually recommended by manufacturers of tubes and interstage transformers. The coupling transformer must be designed especially for the purpose.

The total gain provided by a multi-stage

amplifier is equal to the product of the individual stage gains. For example, when three stages are used, the first having a gain of 100, the second 20 and the third 15, the total gain is 100 × 20 × 15, or 30,000. It is good practice to provide two or three times the minimum required gain in designing the speech amplifier. This will insure having ample gain available to cope with varying conditions.

When the gain must be fairly high, as when a crystal microphone is used, the speech amplifier frequently has four stages, including the power output stage. The first is generally a pentode because of the high gain attainable with this type of tube. The second and third stages are usually triodes, the third frequently having two tubes in push-pull when it drives a Class AB_2 output stage. Two pentode stages are seldom used consecutively because of the difficulty of getting stable operation when the gain per stage is high. With carbon microphones less amplification is needed, hence the pentode first stage usually is omitted, one or two triode stages being ample to obtain full output from the power stage.

Stage gain and voltage output — In voltage amplifiers, the *stage gain* is the ratio of a.c. output voltage to a.c. voltage applied to the grid. It will vary with the applied audio frequency, but for speech work the variation should be small over the range 100–4000 cycles. This condition is easily met in practice.

The output voltage is the maximum value which can be taken from the plate circuit without distortion. It is usually expressed in terms of the peak value of the a.c. wave (§ 2-7) since this value is independent of the waveform. The peak output voltage usually is of interest only when the stage drives a power amplifier, since only in this case is the stage called upon to work near its maximum capabilities. Low-level stages are very seldom worked near full capacity, hence harmonic distortion is negligible and the voltage gain of the stage is the primary consideration.

Resistance coupling — Resistance coupling is generally used in voltage amplifier stages. It is relatively inexpensive, good frequency response can be secured, and there is little danger of hum pick-up from stray magnetic fields associated with heater wiring. It is the only type of coupling suitable for the output circuits of pentodes and high-μ triodes, since with audio-frequency transformers a sufficiently high load impedance (§ 3-3) cannot be obtained without considerable frequency distortion. Typical resistance-coupled circuits are given in Fig. 512.

The frequency response of the amplifier will be determined by the circuit constants, particularly C_3R_4, the coupling condenser and resistor to the following stage, and C_1R_1, the cathode bias resistor and by-pass condenser. For adequate amplification at low frequencies the time constant (§ 2-6) of both these CR combinations should be large. Depending upon the type of tube used in the next stage, R_4 may vary from 50,000 ohms (with power tubes such as the 2A3 or 6F6) to 1 megohm; it is advantageous to use the highest value recommended for the type of tube used since this gives greatest low-frequency response with a given size of coupling condenser, C_3. A capacity of 0.1 μfd. at C_3 will provide ample coupling at low frequencies with any ordinarily-used tube, load resistance (R_3) and next-stage grid resistance (R_4).

The reactance (§ 2-8) of C_1 must be small compared to the resistance of R_1 for good low-frequency response. While with values of R_1 in the vicinity of 10,000 ohms, more or less, a condenser of 1 μfd. will suffice, it is more common practice to use 5- or 10-μfd. low-voltage electrolytic condensers for the purpose, since they are inexpensive and provide ample bypassing. A value of 10 μfd. is usually sufficient with values of R_1 as low as 500 ohms.

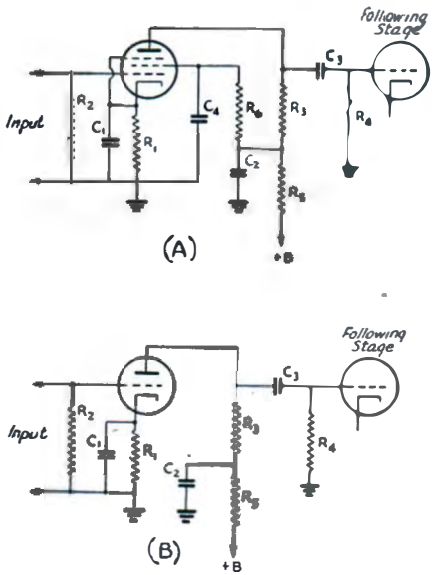


Fig. 512 — Resistance-coupled voltage amplifier circuits. A, pentode; B, triode. Designations are as follows:
C_1 — Cathode by-pass condenser.
C_2 — Plate by-pass condenser.
C_3 — Output coupling condenser (blocking condenser).
C_4 — Screen by-pass condenser.
R_1 — Cathode resistor.
R_2 — Grid resistor.
R_3 — Plate resistor.
R_4 — Next-stage grid resistor.
R_5 — Plate decoupling resistor.
R_6 — Screen resistor.
Values for commonly-used tubes are given in Table I.

For maximum voltage gain the resistance at R_3 should be as high as possible without causing too great a drop in voltage at the plate of the tube. Values range from 50,000 ohms to 0.5 megohm, the smaller figure being used with triodes having comparatively low plate resistance. The value of R_1 depends upon R_3, which principally determines the plate current; in general, the grid bias is somewhat smaller than in circuits having low-resistance output devices (such as a transformer) because of the lower voltage effective at the plate of the tube. This is also true of the screen voltage, for similar reasons, and values of the screen resistor, R_6, may vary from 0.25 to 2 megohms. A screen by-pass (C_4) of 0.1 μfd. will be adequate in all cases.

Table I shows typical values for some of the more popular tube types used in speech amplifiers. The stage gain and peak undistorted output voltage also are given. Other operating conditions are of course possible. The value of the grid resistor, R_2, does not affect any of these quantities, but should not exceed the maximum value recommended by the manufacturer for the particular type of tube used.

The resistance-capacity filter (§ 2-11) formed by C_2R_5 is called a *decoupling circuit*. It isolates the stage from the power supply so that unwanted coupling between this and other stages through the output impedance of the power supply is eliminated. Such coupling is a frequent cause of low-frequency oscillation (*motorboating*) in multistage resistance-coupled amplifiers.

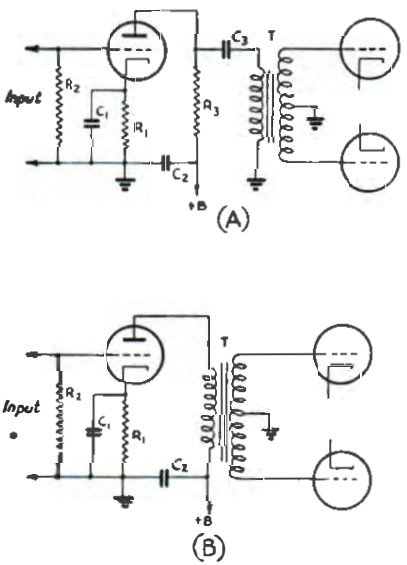


Fig. 513 — Transformer-coupled amplifier circuits for driving a push-pull amplifier. *A*, resistance-transformer coupling; *B*, transformer coupling. Designations correspond to those of Fig. 512. In *A*, values can be taken from Table I. In *B*, the cathode resistor is calculated from the rated plate current and grid bias as given for the particular type of tube used (§ 3-6).

Transformer coupling — Transformer coupling between stages is ordinarily used only when power is to be transferred (in such a case resistance coupling is very inefficient) or when it is necessary to couple between a single-ended and a push-pull stage. Triodes having an amplification factor of 20 or less are used in transformer-coupled voltage amplifiers.

Representative circuits for single-ended to push-pull are shown in Fig. 513. That at A uses a combination of resistance and transformer coupling and may be used for exciting the grids of a Class-A or AB_1 following stage. The resistance coupling is used to keep the d.c. plate current from flowing through the transformer primary and thereby prevent a reduction in primary inductance below its no-current value (§ 8-4). This improves the low-frequency response. With triodes ordinarily used (6C5, 6J5, etc.) the gain is equal to that with resistance coupling (typical values in Table I) multiplied by the secondary-to-primary turn ratio of the transformer. This ratio is generally 2:1

In B the transformer primary is in series with the plate of the tube and thus must carry the tube plate current. When the following amplifier operates without grid current, the voltage gain of the stage is practically equal to the μ of the tube multiplied by the transformer ratio. This circuit is also suitable for transferring power (within the capabilities of the tube) as in the case of a following Class-AB_2 stage used as a driver for a Class-B modulator.

Gain control — The overall gain of the amplifier may be changed to suit the output level of the microphone, which will vary with voice intensity and distance of the speaker from the microphone, by varying the proportion of a.c. voltage applied to the grid of one of the stages. This is done by means of an adjustable voltage divider (§ 2-6), commonly called a "potentiometer" or "volume control," as shown in Fig. 514. The actual voltage applied between grid and cathode will be very nearly equal to the ratio of the resistance between AB to the total resistance AC, multiplied by the a.c. voltage which appears across AC. The gain control is usually also the grid resistor for the amplifier stage with which it is associated.

The gain control potentiometer should be near the input end of the amplifier so that there will be no danger that stages ahead of the gain control will overload. With carbon microphones the gain control may be placed directly across the microphone transformer secondary, but with other types the gain con-

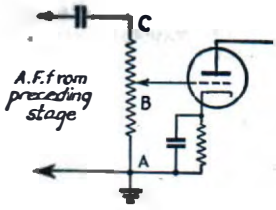


Fig. 514 — Gain control circuit.

trol usually will affect the frequency response of the microphone when connected directly across it. The control is therefore usually placed in the grid circuit of the second stage.

Phase inversion — Push-pull output may be secured with resistance coupling by using an extra tube as shown in Fig. 515. There is a phase shift of 180 degrees through any normally-operating resistance-coupled stage (§ 3-3) and the extra tube is used purely to provide this phase shift without additional gain. The outputs of the two tubes are then added to give push-pull excitation to the next amplifier.

In Fig. 515, V_1 is the regular amplifier, connected in normal fashion to the grid of one of the push-pull tubes. The next-stage grid resistor is tapped so that part of the output voltage is fed to the grid of the phase inverter, V_2. This tube then amplifies the signal and applies it in reverse phase to the grid of the second push-pull tube. Two similar tubes should be used at V_1 and V_2, with identical plate resistors and output coupling condensers. The tap on R_4 is adjusted to make V_1 and V_2 give equal voltage outputs so that balanced excitation is applied to the grids of the following stage.

The cathode resistor, R_6, commonly is left un-bypassed since this tends to help balance the circuit. Double-triode tubes are frequently used as phase inverters.

TABLE I—TYPICAL VOLTAGE AMPLIFIER DATA

Tube Type	R_3, *megohms*	R_5, *megohms*	R_1, *ohms*	*Peak Output Volts*	*Voltage Gain*
6C5	0.1	—	6000	88	13
6J5	0.1	—	3000	64	14
6F5, 6SF5	0.25	—	3000	54	63
6J7	0.25	1.2	1200	104	140
6SJ7	0.25	1.0	900	88	167
	0.5	2.0	1300	64	200

Other values (Fig. 512): C_1, 10 μfd. (low-voltage electrolytic); C_2, 8-μfd. electrolytic; C_3, C_4, 0.1-μfd. paper; R_2, 0.1 to 1 megohm; R_4, 0.5 megohm; R_5, 10,000 to 50,000 ohms. Data are based on a plate-supply voltage of 300; lower values will reduce the undistorted peak output voltage in proportion, but will not materially affect the voltage gain.

Output limiting — It is desirable to modulate as heavily as possible without overmodulating, yet it is difficult to speak into the microphone at a constant intensity. To maintain reasonably constant output from the modulator in spite of variations in speech intensity, it is possible to use automatic gain control which follows the *average* (not instantaneous) variations in speech amplitude. This is accomplished by rectifying and filtering (§ 8-2, 8-3) some of the audio output and applying the rectified and filtered d.c. to a control electrode in an early stage in the amplifier.

A practical circuit for this purpose is shown in Fig. 516. The rectifier must be connected, through the transformer, to a tube capable of

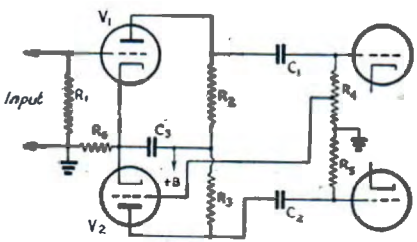


Fig. 515 — Phase inverter circuit for resistance-coupled push-pull output. With a double-triode tube (6N7) the following values are typical:
R_1 — 0.5 megohm.
R_2, R_3 — 0.1 megohm.
R_4, R_5 — 0.5 megohm.
R_6 — 1500 ohms.
C_1, C_2 — 0.1 μfd.
R_4 should be tapped as described in the text. The voltage gain with these constants is 22.

delivering some power output (a small part of the output of the power stage may be used) or else a separate amplifier for the rectifier circuit alone may have its grid connected in parallel with that of the last voltage amplifier. Resistor R_4 in series with R_5 across the plate supply provides variable bias on the rectifier plates so that the limiting action can be delayed until a desired microphone input level is reached. R_2, R_3, C_2, C_3, and C_4 form the filter, (§ 2-11) and the output of the rectifier is connected to the suppressor grid of the pentode first stage of the speech amplifier.

A step-down transformer giving about 50 volts when its primary is connected to the output circuit should be used. A half-wave rectifier can be used instead of the full-wave circuit shown, although satisfactory filtering is more difficult.

Noise — It is important that the noise level in a speech amplifier be low compared to the level of the desired signal. Noise in the speech amplifier is chiefly hum, which may be the result of insufficient power-supply filtering or may be introduced into the grid circuit of a

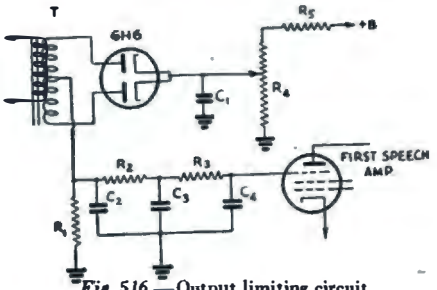


Fig. 516 — Output limiting circuit.
C_1, C_2, C_3, C_4 — 0.1-μfd. paper.
R_1, R_2, R_3 — 0.25 megohm.
R_4 — 25,000-ohm potentiometer.
R_5 — 0.1 megohm.
T — See text.

tube by magnetic or electrostatic means from heater wiring. The plate voltage for the amplifier should be free from ripple (§ 8-4), particularly the voltage applied to the low-level stages. A two-section condenser-input filter (§ 8-5) is usually satisfactory. The decoupling circuits mentioned in the preceding paragraphs also are helpful in reducing plate-supply hum.

Hum from heater wiring may be reduced by keeping the wiring well away from ungrounded components or wiring, particularly in the vicinity of the grid of the first tube. Complete shielding of the microphone jack is advisable, and when tubes with grid caps instead of the single-ended types are used the caps and the exposed wiring to them should be shielded. Heater wiring preferably should run in the corners of a metal chassis to reduce the magnetic field. A ground should be made either on one side of the heater circuit or to the center-tap of the heater winding. The shells of metal tubes should be grounded; glass tubes require separate shields, especially when used in low-level stages. Heater connections to the tube sockets should be kept as far as possible from the plate and grid prongs, and the heater wiring to the sockets should be kept close to the chassis. A connection to a good ground (such as a cold water pipe) also is advisable. The speech amplifier always should be constructed on a metal chassis.

When the power supply is mounted on the same chassis with the speech amplifier, the power transformer and filter chokes should be well separated from audio transformers in the amplifier proper, to reduce magnetic coupling.

●5-10 CHECKING 'PHONE TRANSMITTER OPERATION

Modulation percentage — The most reliable method of determining percentage of modulation is by means of the cathode-ray oscilloscope (§ 3-9). The oscilloscope gives a direct picture of the modulated output of the transmitter, and the waveform errors inherent in other types of measurements are eliminated.

Two types of oscilloscope patterns may be obtained, known as the "wave envelope" and "trapezoid." The former shows the shape of the modulation envelope (§ 5-2) directly, while the latter in effect plots the modulation characteristic (§ 5-2) of the modulated stage on the cathode-ray tube screen. To obtain the wave-envelope pattern the oscilloscope must have a horizontal sweep circuit. The trapezoid pattern requires only the oscilloscope, the sweep circuit being supplied by the transmitter itself. Fig. 517 shows methods of connecting the oscilloscope to the transmitter for both types of patterns. The oscilloscope connections for the wave-envelope pattern, Fig. 517-A, are usually simpler than those for the trapezoidal figure. The vertical deflection plates are coupled to the amplifier tank coil or an antenna coil by means of a pickup coil of a few turns connected to the oscilloscope through a twisted-pair line. The position of the pickup coil is varied until a carrier pattern, Fig. 518-B, of suitable height is obtained. The sweep voltage should be adjusted to make the width of the pattern somewhat more than half the diameter of the screen. It is frequently helpful in eliminating r.f. harmonics from the pattern to connect a resonant circuit, tuned

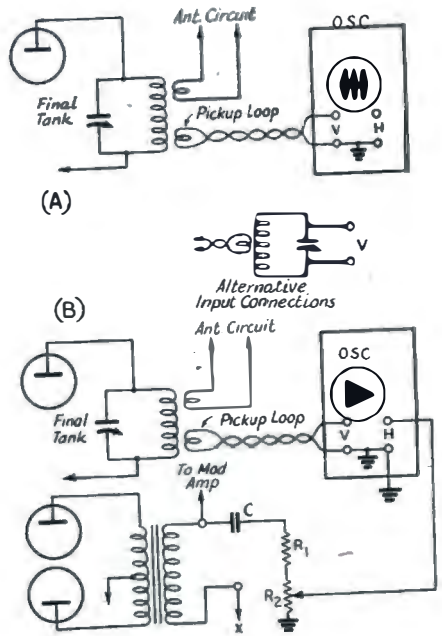


Fig. 517 — Methods of connecting an oscilloscope to the modulated amplifier for checking modulation.

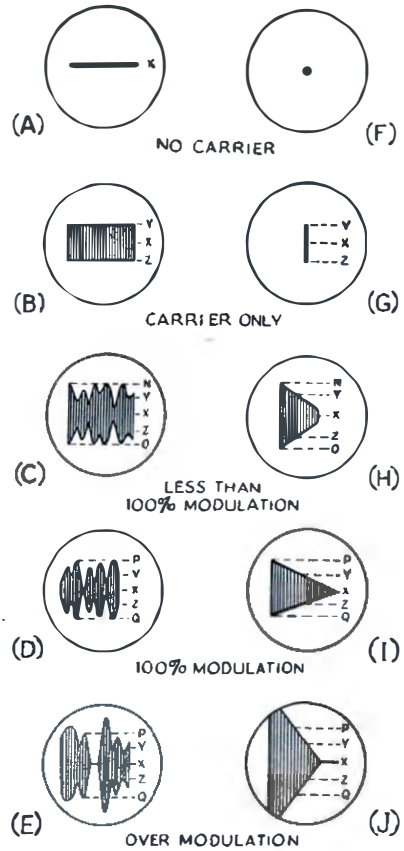


Fig. 518— **Wave-envelope and trapezoidal patterns under different conditions of modulation.**

to the operating frequency, between the vertical deflection plates, using link coupling between this circuit and the transmitter tank circuit.

With the application of voice modulation a rapidly-changing pattern of varying height will be obtained. When the maximum height of this pattern is just twice that of the carrier alone, the wave is being modulated 100% (§ 5-2). This is illustrated by Fig. 518-D, where the point X represents the sweep line (reference line) alone, YZ is the carrier height, and PQ is the maximum height of the modulated wave. If the height is greater than the distance PQ, as illustrated in E, the wave is overmodulated in the upward direction. Overmodulation in the downward direction is indicated by a gap in the pattern at the reference axis, where a single bright line appears on the screen. Overmodulation in either direction may take place even when the modulation in the other direction is less than 100%. Assuming that the modulation is symmetrical, however, any modulation percentage can be measured directly from the screen by measuring the maximum height with modulation and the height of the carrier alone; calling these two heights h_1 and h_2, respectively, the modulation percentage is

$$\frac{h_1 - h_2}{h_2} \times 100$$

Connections for the trapezoidal pattern are shown in Fig. 517-B. The vertical plates are similarly coupled to the transmitter tank circuit through a pick-up loop; the tuned input circuit to the oscilloscope may also be used. The horizontal plates are coupled to the output of the modulator through a voltage divider (§ 2-6) R_1R_2, the latter resistance being variable to permit adjustment of the audio voltage to a suitable value to give a satisfactory horizontal sweep on the screen. R_2 may be a 0.25-megohm volume control resistor. The value of R_1 will depend upon the audio output voltage of the modulator. This voltage is equal to $\sqrt{PR}$, where P is the audio power output of the modulator and R is the modulating impedance of the modulated r.f. amplifier. In the case of grid-bias modulation with a 1:1 output transformer, it will be satisfactory to assume that the a.c. output voltage of the modulator is equal to $0.7E$ for a single tube, or $1.4E$ for a push-pull stage, where E is the d.c. plate voltage on the modulator. If the transformer ratio is other than 1:1, the voltage so calculated should be multiplied by the actual secondary-to-primary turn ratio. The total resistance of R_1 and R_2 in series should be 0.25 megohm for every 150 volts of modulator output; for example, if the modulator output voltage is 600, the total resistance should be four (600/150) times 0.25 megohm, or 1 megohm. Then with 0.25 megohm at R_2, R_1 should be 0.75 megohm. The blocking condenser C should be 0.1 μfd or more and its voltage rating should be greater than the maximum voltage appearing in the circuit. With plate modulation, this is twice the d.c. voltage applied to the plate of the modulated amplifier.

The trapezoidal patterns are shown in Fig. 518 at F to J, each alongside the corresponding wave-envelope pattern. With no signal, only the cathode ray spot appears on the screen. When the unmodulated carrier is applied a vertical line appears, and its length should be adjusted by means of the pickup coil coupling to a convenient value. When the carrier is modulated the wedge-shaped pattern appears; the higher the modulation percentage the wider and more pointed the wedge becomes. At 100% modulation it just makes a point on the axis A at one end and the height PQ at the other end is equal to twice the carrier height YZ.

Overmodulation in the upward direction is indicated by increased height over *PQ*, and in the downward direction by an extension along the axis *X* at the pointed end. The modulation percentage may be found by measuring the modulated and unmodulated carrier heights in the same way as with the wave envelope pattern.

Non-symmetrical waveforms — In voice waveforms the average maximum amplitude in one direction from the axis is frequently greater than in the other direction, although the average *energy* on both sides is the same. For this reason the percentage of modulation in the up direction frequently differs from that in the down direction, and with a given voice and microphone this difference in modulation percentage is usually always in the same direction. Since overmodulation in the downward direction causes more out-of-channel interference than overmodulation upward because of the steeper wavefront (§ 6-1), it is advisable to "phase" the modulation so that the side of the voice waveform having the larger excursions causes the instantaneous carrier power to increase and the smaller excursions to cause a power decrease. This reduces the likelihood of overmodulation on the down peak. The direction of the larger excursions can readily be found by careful observation of the oscilloscope pattern. The phase can be reversed by reversing the connections of one winding of any transformer in the speech amplifier or modulator.

Modulation monitoring — While it is desirable to modulate as fully as possible, 100% modulation should not be exceeded, particularly in the downward direction, because harmonic distortion will be introduced and the channel width increased (§ 5-2), thus causing unnecessary interference to other stations. The oscilloscope may be used to provide a continuous check on the modulation, but simpler indicators may be used for the purpose, once calibrated. A convenient indicator, when a Class-B modulator (§ 5-6) is used, is the plate milliammeter in the Class-B stage, since plate current fluctuates with the voice intensity. Using the oscilloscope, determine the gain-control setting and voice intensity which gives 100% modulation on voice peaks, and simultaneously observe the maximum Class-B plate-milliammeter reading on the peaks. When this maximum reading is obtained, it will suffice in regular operation to adjust the gain so that it is not exceeded.

A sensitive rectifier-type voltmeter (copper oxide type) can also be used for modulating monitoring. It should be connected across the output circuit of an audio driver stage where the power level is a few watts, and similarly calibrated against the oscilloscope to determine the reading which represents 100% modulation.

The plate milliammeter of the modulated r.f. stage may also be used as an indicator of overmodulation. Since the average plate current is constant (§ 5-3, 5-4, 5-5) when the amplifier is linear, the reading will be the same with or without modulation. When the amplifier is overmodulated, especially in the downward direction, the operation is no longer linear and the average plate current will change. A flicker of the pointer may therefore be taken as an indication of overmodulation or non-linearity. However, it is possible that the average plate current will remain constant with considerable overmodulation under some operating conditions, so such an indicator is not wholly reliable unless it has been previously checked against an oscilloscope.

Linearity — The linearity (§ 5-2) of a modulated amplifier may readily be checked with the oscilloscope. The trapezoidal pattern is more easily interpreted than the wave envelope pattern and less auxiliary equipment is required. The connections are the same as for measuring modulation percentage (Fig. 517). If the amplifier is perfectly linear, the sloping sides of the trapezoid will be perfectly straight from the point at the axis up to at least 100% modulation in the upward direction. Nonlinearity will be shown by curvature of the sides. Curvature near the point, extending the point farther along the axis than would occur with straight sides, indicates that the output power does not decrease rapidly enough in this region; it may also be caused by imperfect neutralization (a push-pull amplifier is recommended because better neutralization is possible than with single-ended amplifiers) or r.f. leakage from the exciter through the final stage. The latter condition can be checked by removing the plate voltage from the modulated stage, when the carrier should disappear and only the beam spot remain on the screen (Fig. 518-F). If a small vertical line remains the amplifier should be re-neutralized to eliminate it; if this does not suffice, r.f. is being picked up from lower-power stages either by coupling through the final tank circuit or through the oscilloscope pickup circuit.

Inward curvature at the large end of the pattern is caused by improper operating conditions of the modulated amplifier, usually improper bias or insufficient excitation, or both, with plate modulation. In grid-bias and cathode-modulated systems, the bias, excitation and plate loading are not correctly proportioned when such curvature occurs, usually because the amplifier has been adjusted to have too-high carrier efficiency without modulation (§ 5-4, 5-5).

For the wave-envelope pattern it is necessary to have a linear horizontal sweep circuit in the oscilloscope and a source of sine-wave audio signal (such as an audio oscillator or signal generator) which can be synchronized with the sweep circuit. The linearity can be judged by comparing the wave envelope with a true sine wave. Distortion in the audio circuits will affect the pattern in this case (such distortion has no effect on the trapezoidal pattern, which shows the modulation characteristic of the r.f. amplifier alone), and it is also readily possible to misjudge the shape of the modulation envelope, so that the wave envelope is less useful than the trapezoid for checking linearity of the modulated amplifier.

Fig. 519 shows typical patterns of both types. The cause of the distortion is indicated for grid-bias and suppressor modulation. The patterns at *A*, although not truly linear, are representative of properly-operated grid-bias modulation systems. Better linearity can be obtained with plate modulation of a Class-C amplifier.

Faulty patterns — The drawings of Figs. 518 and 519 show what is normally to be expected in the way of pattern shapes when the oscilloscope is used to check modulation. If the actual patterns differ considerably from those shown, it is probable that the pattern is faulty rather than the transmitter. It is important that only r.f. from the modulated stage be coupled to the oscilloscope, and then only to the vertical plates. The effect of stray r.f. from other stages in the transmitter has been mentioned in the preceding paragraph. If r.f. is also present on the horizontal plates, the pattern will lean to one side instead of being upright. If the oscilloscope cannot be moved to a spot where the unwanted pick-up disappears, a small by-pass condenser (10 $\mu\mu$fd.) should be connected across the horizontal plates as close to the cathode-ray tube as possible. An r.f. choke (2.5 mh. or smaller) may also be connected in series with the ungrounded horizontal plate.

"Folded" trapezoidal patterns occur when the audio sweep voltage is taken from some point in the audio system other than that where the a.f. power is applied to the modulated stage, and are caused by a phase difference between the sweep voltage and the modulating voltage. The connections should always be as shown in Fig. 517-B.

Plate-current shift — As mentioned above, the d.c. plate current of a modulated amplifier will be the same with and without modulation so long as the amplifier operation is perfectly linear and other conditions remain unchanged. This also assumes that the modulator is working within its capabilities. Because there is usually some curvature of the modulation

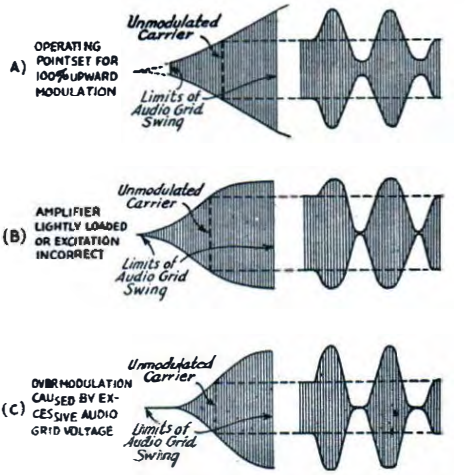


Fig. 519 — Oscilloscope patterns representing proper and improper grid-bias or cathode modulation. The pattern obtained with a correctly adjusted amplifier is shown at A. The other two drawings indicate nonlinear modulation.

characteristic with grid-bias modulation there is normally a slight upward change in plate current of a stage so modulated, but this occurs only at high modulation percentages and is barely detectable under the usual conditions of voice modulation.

With plate modulation, a downward shift in plate current may indicate one or more of the following:

1. Insufficient excitation to the modulated r.f. amplifier.
2. Insufficient grid bias on the modulated stage.
3. Wrong load resistance for Class-C r.f. amplifier.
4. Insufficient output capacity in filter of modulated amplifier plate supply.
5. Heavy overloading of Class-C r.f. amplifier tube or tubes.

Any of the following may cause an upward shift in plate current:

1. Overmodulation (excessive audio power, audio gain too great).
2. Incomplete neutralization of the modulated amplifier.
3. Parasitic oscillation in the modulated amplifier.

When a common plate supply is used for both Class-B (or Class AB) modulator and modulated r.f. amplifier, the plate current of the latter may "kick" downward because of poor power-supply voltage regulation (§ 8-1) with the varying additional load of the modulator on the supply. The same effect may occur with high-power transmitters because of poor

regulation of the a.c. supply mains, even when a separate power supply unit is used for the Class-B modulator. Either condition may be detected by measuring the plate voltage applied to the modulated stage; in addition, poor line regulation may also be detected by a downward shift in filament or line voltage.

With grid-bias modulation, any of the following may be the cause of a plate current shift greater than the normal mentioned above:

Downward kick: Too much r.f. excitation; insufficient operating bias; distortion in modulator or speech amplifier; too-high resistance in bias supply; insufficient output capacity in plate-supply filter to modulated amplifier; amplifier plate circuit not loaded heavily enough; plate-circuit efficiency too high under carrier conditions.

Upward kick: Overmodulation (excessive audio voltage); distortion in audio system; regeneration because of incomplete neutralization; operating grid bias too high.

A downward kick in plate current will accompany an oscilloscope pattern like that of Fig. 519-B; the pattern with an upward kick will look like Fig. 519-A with the shaded portion extending farther to the right and above the carrier, for the "wedge" pattern.

Noise and hum on carrier — These may be detected by listening to the signal on a receiver sufficiently removed from the transmitter to avoid overloading. The hum level should be low compared to the voice at 100% modulation. Hum may come either from the speech amplifier and modulator or from the r.f. section of the transmitter. Hum from the r.f. section can be detected by completely shutting off the modulator; if hum remains when this is done the power-supply filters for one or more of the r.f. stages have insufficient smoothing (§ 8-4). With a hum-free carrier, hum introduced by the modulator can be checked by turning on the modulator but leaving the speech amplifier off; power-supply filtering is the likely source of such hum. If carrier and modulator are both clean, connect the speech amplifier and observe the increase in hum level. If the hum disappears with the gain control at minimum, the hum is being introduced in the stage or stages preceding the gain control. The microphone may also pick up hum, a condition which can be checked by removing the microphone from the circuit but leaving the first speech-amplifier grid circuit otherwise unchanged. A good ground on the microphone and speech system is usually essential to hum-free operation.

Hum can be checked with the oscilloscope, where it appears as modulation on the carrier in the same way as the normal modulation. Usually the percentage is rather small, but if the carrier shows modulation with no speech input hum is the likely cause. The various parts of the transmitter may be checked through as described above.

Spurious sidebands — A superheterodyne receiver having a crystal filter (§ 7-8, 7-11) is needed for checking spurious sidebands outside the normal communication channel (§ 5-2). The r.f. input to the receiver must be kept low enough, by removing the antenna or by adequate separation from the transmitter, to avoid overloading and consequent spurious receiver responses (§ 7-8). With the crystal filter in its sharpest position and the beat oscillator turned on, tune through the region outside the normal channel limits (3 to 4 kc. each side of the carrier) while another person talks into the microphone. Spurious sidebands will be observed as intermittent beat notes coinciding with voice peaks, or in bad cases of distortion or overmodulation as "clicks" or crackles well away from the carrier frequency. Sidebands more than 4 kc. from the carrier should be of negligible strength in a properly modulated 'phone transmitter. The causes are overmodulation or non-linear operation (§ 5-3).

R.f. in speech amplifier — A small amount of r.f. current in the speech amplifier — particularly in the first stage, which is most susceptible to such r.f. pick-up — will cause overloading and distortion in the low-level stages. Frequently also there is a regenerative effect which causes an audio-frequency oscillation or "howl" to be set up in the audio system. In such cases the gain control cannot be advanced very far before the howl builds up, even though the amplifier may be perfectly stable when the r.f. section of the transmitter is not turned on.

Complete shielding of the microphone, microphone cord and speech amplifier are necessary to prevent r.f. pick-up, and a ground connection separate from that to which the transmitter is connected is advisable. Unsymmetrical or capacity coupling to the antenna (single-wire feed, feeders tapped on final tank circuit, etc.) may be responsible in that these systems sometimes cause the transmitter chassis to take an r.f. potentional above ground. Inductive coupling to a two-wire transmission line is advisable. This antenna effect can be checked by disconnecting the antenna and dissipating the power in a dummy antenna (§ 4-9) when it usually will be found that the r.f. feedback disappears. If it does not, the speech amplifier and microphone shielding are at fault.

• 5-11 FREQUENCY MODULATION

Principles — In frequency modulation the carrier amplitude is constant and the output frequency of the transmitter is made to vary about the carrier or mean frequency at a rate

corresponding to the audio frequencies of the speech currents. The extent to which the frequency changes in one direction from the unmodulated or carrier frequency is called the *frequency deviation*. It corresponds to the change of carrier amplitude in the amplitude-modulation system (§ 5-2). Deviation is usually expressed in kilocycles, and is equal to the difference between the carrier frequency and either the highest or lowest frequency reached by the carrier in its excursions with modulation. There is no modulation percentage in the usual sense; with suitable circuit design the deviation may be made as large as desired without encountering any effect equivalent to overmodulation in the amplitude system.

Deviation ratio — The ratio of the maximum frequency deviation to the audio frequency of the modulation is called the *deviation ratio*. Unless otherwise specified, it is taken as the ratio of the maximum frequency deviation to the *highest* audio frequency to be transmitted.

Advantages of f.m. — The chief advantage of frequency modulation over amplitude modulation is noise reduction at the receiver. All electrical noises in the radio spectrum, including those originating in the receiver, are r.f. oscillations which vary in amplitude, this variation causing the noise response in amplitude-modulation receivers. If the receiver does not respond to amplitude variations but only to frequency changes, noise can affect it only by causing a phase shift which appears as frequency modulation on the signal. The effect of such frequency modulation by the noise can be made small by making the frequency change (deviation) in the signal large.

A second advantage is that the power required for modulation is inconsequential, since there is no power variation in the modulated output.

Triangular spectrum — The way in which noise is reduced by a large deviation ratio is illustrated by Fig. 520. In this figure the noise is assumed to be evenly distributed over the channel used, an assumption which is almost always true. It is also assumed that audio frequencies above 4000 cycles (4 kc.) are not necessary to voice communication, and that the audio system in the receiver has no response above this frequency. Then if an amplitude modulation receiver is used and its selectivity is such that there is no attenuation of sidebands (§ 5-2) below 4000 cycles, the noise components of all frequencies within the channel will produce equal response when they beat with a carrier centered in the channel. This is shown by the line *DC*.

In the f.m. receiver the output amplitude is proportional to the frequency deviation, and

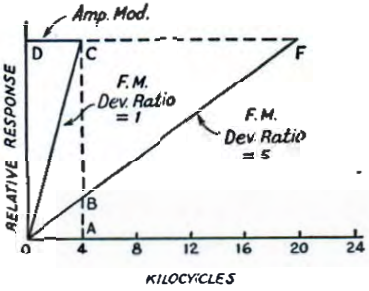


Fig. 520 — **Triangular spectrum of noise response in an f.m. receiver compared with amplitude modulation. Deviation ratios of 1 and 5 are shown.**

noise components in the channel can be considered to frequency-modulate the steady carrier with a deviation proportional to the difference between the actual frequency of the component and the frequency of the carrier, and also to give an audio-frequency beat of the same frequency difference. This leads to a rising response characteristic such as the line *OC*, where the noise amplitude is proportional to the audio beat frequency. The average noise power output is proportional to the square root of the sum of the squares of all the amplitude values (§ 2-7), so that the noise power with f.m. having a deviation ratio of 1 is only ⅓ that with amplitude modulation, or an improvement of 4.75 db.

If the deviation ratio is increased to 5, the noise response is represented by the line *OF*, but only frequencies up to 4000 cycles are reproduced in the output so the audible noise is confined to the triangle *OAB*. These relations only hold when the carrier is strong compared to the noise. With weak signals the signal-to-noise ratio is better with a deviation ratio of 1.

Linearity — A transmitter in which frequency deviation is directly proportional to the amplitude of the modulating signal is said to be *linear*. It is also essential that the carrier amplitude remain constant under modulation, which in turn requires that the transmitter tuned circuits be broad enough to handle without discrimination the range of frequencies transmitted. This requirement is easily met under ordinary conditions.

Sidebands — In frequency modulation there is a series of sidebands on either side of the carrier frequency for each audio-frequency component in the modulation. In addition to the usual sum and difference frequencies (§ 5-2) there are also beats at harmonics of the fundamental modulating frequency, even though the latter may be a pure tone. This occurs because of the necessity for maintaining the proper phase relationships between the carrier and sidebands to keep the power output constant. Hence a frequency-modulated signal

inherently occupies a wider channel than an amplitude-modulated signal, and because of the necessity for conserving space in the usual communication spectrum the use of f.m. is confined to the ultra-high frequencies.

The number of sidebands for a single modulating frequency increases with the frequency deviation. When the deviation ratio is of the order of 5 the sidebands beyond the maximum frequency deviation are usually negligible, so that the channel required is approximately twice the frequency deviation.

• 5-12 METHODS OF FREQUENCY MODULATION

Requirements and methods — At present there are no fixed standards of frequency deviation in amateur work. Since a deviation ratio of 5 is considered high enough in any case, the maximum deviation necessary is 15 to 20 kc. for an upper audio-frequency limit of 3000 or 4000 cycles (§ 5-2), or a channel width of 30 to 40 kc. The permissible deviation is determined by the receiver (§ 7-18), since deviation beyond the limits of the receiver pass-band causes distortion. If the transmitter is designed to be linear (§ 5-11) with a deviation of about 15 kc. it can be used at lower deviation ratio simply by reducing the gain in the speech amplifier, and thereby made to conform to the requirements of the particular receiver in use.

The several possible methods of frequency modulation include mechanical (for instance, varying condenser plate spacing in accordance with voice vibrations), initial phase modulation which is later transformed into frequency modulation, and direct frequency modulation of an oscillator by electrical means. The latter, in the form of the *reactance modulator*, is the simplest system.

The reactance modulator — The reactance modulator is a vacuum tube amplifier connected to the r.f. tank circuit of an oscillator in such a way as to act as a variable inductance or capacity of a value dependent upon the instantaneous a.f. voltage applied to its grid. Fig. 521 is a representative circuit. The control grid circuit of the 6L7 tube is connected across the small capacity C_1, which is in series with the resistor R_1 across the oscillator tank circuit. Any type of oscillator circuit (§ 3-7) may be used. R_1 is large compared to the reactance (§ 2-8) of C_1, so the r.f. current through R_1C_1 will be practically in phase (§ 2-7) with the r.f. voltage appearing at the terminals of the tank circuit. However, the voltage across C_1 will lag the current by 90 degrees (§ 2-8). The r.f. current in the plate circuit of the 6L7 will be in phase with the grid voltage (§ 3-3) and consequently is 90 degrees behind the current through C_1, or 90 degrees behind the r.f. tank voltage. This lagging current is drawn through the oscillator tank, giving the same effect as though an inductance were connected across the tank (in an inductance the current lags the voltage by 90 degrees — § 2-8). The frequency is therefore increased in proportion to the lagging plate current of the modulator. This in turn is determined by the a.f. voltage applied to the No. 3 grid of the 6L7, hence the oscillator frequency varies with the audio signal.

Other circuit arrangements to produce the same effect can be used. It is convenient to use a tube (such as the 6L7) in which the r.f. and a.f. voltages can be applied to separate control grids; however, both voltages may be applied to the same grid with suitable precautions taken to prevent r.f. from flowing in the external audio circuit and vice versa (§ 2-13).

The modulated oscillator is usually operated on a relatively low frequency so that a high order of carrier stability can be secured. Frequency multipliers are used to raise the frequency to the final frequency desired. The

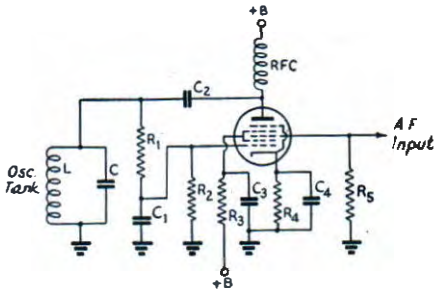


Fig. 521 — Reactance modulator circuit using a 6L7 tube.

C — Oscillator tank capacity.
C_1 — 3–10 μμfd. (See text.)
C_2 — 250 μμfd.
C_3 — 8-μfd. electrolytic (a.f. by-pass) in parallel with 0.01-μfd. paper (r.f. by-pass).
C_4 — 0.01 μfd.
L — Oscillator tank inductance.
R_1 — 50,000 ohms.
R_2 — 0.5 megohm.
R_3 — 30,000 ohms.
R_4 — 300 ohms.
R_5 — 0.5 megohm.

frequency deviation increases with the number of times the initial frequency is multiplied; for instance, if the oscillator is operated on 7 Mc. and the output frequency is to be 112 Mc., an oscillator frequency deviation of 1000 cycles will be raised to 16,000 cycles at the output frequency.

Design considerations — The sensitivity of the modulator (frequency change per unit change in grid voltage) increases when C_1 is made smaller, for a fixed value of R_1, and also increases with an increase in L/C ratio in the oscillator tank circuit. Since the carrier

stability of the oscillator depends on the L/C ratio (§ 3-7) it is desirable to use the highest tank capacity which will permit the desired deviation to be secured while keeping within the limits of linear operation. When the circuit of Fig. 521 is used in connection with a 7-Mc. oscillator, a linear deviation of 2000 cycles above and below the carrier frequency can be secured when the oscillator tank capacity is approximately 200 $\mu\mu$fd. A peak a.f. input of two volts is required for full deviation. At 56 Mc. the maximum deviation would be 8×2000 or 16 kc.

Since a change in any of the voltages on the modulator tube will cause a change in r.f. plate current and consequently a frequency change, it is advisable to use a regulated plate supply for both modulator and oscillator. At the low voltages used (250 volts), the required stabilization can be secured by means of gaseous regulator tubes (§ 8-8).

Speech amplification — The speech amplifier preceding the modulator follows ordinary design (§ 5-9) except that no power is required from it and the a.f. voltage taken by the modulator grid is usually small — not more than 10 or 15 volts even with large modulator tubes. Because of these modest requirements only a few speech-amplifier stages are needed; a two-stage amplifier consisting of a pentode followed by a triode, both resistance coupled, will suffice for crystal microphones (§ 5-8).

R.f. amplifier stages — The frequency multiplier and output stages following the modulated oscillator may be designed and adjusted in accordance with ordinary principles. No special excitation requirements are imposed, since the amplitude of the output is constant. Enough frequency multiplication must be used to give the desired maximum deviation at the final frequency; this depends upon the maximum linear deviation available from the modulator-oscillator. All stages in the transmitter should be tuned to resonance, and careful neutralization of any straight amplifier stages (§ 4-7) is necessary to prevent r.f. phase shifts which might cause distortion.

Checking operation — The two quantities to be checked in the f.m. transmitter are linearity and frequency deviation. With a modulator of the type shown in Fig. 521, both the r.f. and a.f. voltages are small enough to make the operation Class A (§ 3-4) so that the plate current of the modulator is constant so long as operation is over the linear portions of the No. 1 and No. 3 grid characteristics. Hence non-linearity will be indicated by a change in plate current as the a.f. modulating voltage is increased. The distortion will be within acceptable limits with the tube and constants given in Fig. 521 when the plate current does not change more than 5% with signal.

Non-linearity is accompanied by a shift in the carrier frequency, so it can also be checked by means of a selective receiver such as one with a crystal filter (§ 7-11). A tone source is convenient for the test. Set the receiver for high selectivity, switch on the beat oscillator and tune to the oscillator carrier frequency. (The check does not need to be made at the output frequency, and the oscillator frequency usually is more convenient since it will fall within the tuning range of a communications receiver.) Increase the modulating signal until a definite shift in carrier frequency is observed; this indicates the point at which non-linearity starts. The modulating signal should be kept below this level for minimum distortion.

A selective receiver also can be used to check frequency deviation, again at the oscillator frequency. A source of tone of known frequency is required, preferably a continuously variable calibrated audio oscillator or signal generator. Tune in the carrier as described above, using the beat oscillator and high selectivity, and adjust the modulating signal to the maximum level at which linear operation is secured. Starting with the lowest frequency available, slowly raise the tone frequency while listening closely to the carrier beat note. As the tone frequency is raised the beat note will decrease in intensity, disappear entirely at a definite frequency, then come back and increase in intensity as the tone frequency is raised still more. The frequency at which the beat note disappears multiplied by 2.4 is the frequency deviation at that level of modulating signal; for example, if the beat note disappears with an 800-cycle tone the deviation is 2.4×800, or 1920 cycles. The deviation at the output frequency is the oscillator deviation multiplied by the number of times the frequency is multiplied; in this example, if the oscillator is on 7 Mc. and the output on 56 Mc., the final deviation is 1920×8, or 15.36 kc.

The output of the transmitter can be checked for amplitude modulation by observing the antenna current. It should not change from the unmodulated carrier value when the transmitter is modulated. If there is no antenna ammeter in the transmitter, a flashlight lamp and loop can be coupled to the final tank coil to serve as a current indicator. The lamp brilliance should not change with modulation.

Keying

●6-1 KEYING PRINCIPLES AND CHARACTERISTICS

Requirements — The keying of a transmitter can be considered satisfactory if the method employed reduces the power output to zero when the key is open, or "up," and permits full power to reach the antenna when the key is closed, or "down." Furthermore, it should do this without causing keying transients or "clicks," which cause interference with other amateur stations and with local broadcast reception, and it should not affect the frequency of the emitted wave.

Back-wave — From various causes some energy may get through to the antenna during keying spaces. The effect then is as though the dots and dashes were simply louder portions of a continuous carrier; in some cases, in fact, the *back-wave*, or signal heard during the keying spaces, may seem to be almost as loud as the keyed signal. Under these conditions the keying is hard to read. A pronounced back-wave often results when the amplifier stage feeding the antenna is keyed; it may be present because of incomplete neutralization (§ 4-7) of the final stage, allowing some energy to get to the antenna through the grid-plate capacity of the tube, or because of magnetic coupling between antenna coupling coils and one of the low-power stages.

A back-wave also may be radiated if the keying system does not reduce the input to the keyed stage to zero during keying spaces. This trouble will not occur in keying systems which cut off the plate voltage when the key is open, but may be present in grid-blocking systems (§ 6-3) if the blocking voltage is not great enough and, in power supply primary-keyed systems (§ 6-3) if only the final stage power supply primary is keyed.

Keying waveform and sidebands — A c.w. signal can be considered equivalent to any modulated signal (§ 5-1) except that instead of being modulated by sinusoidal waves and their harmonics, it is modulated by a rectangular wave as in Fig. 601-A. If it were modulated by a sinusoidal wave of single frequency, as in Fig. 601-B, the only sidebands would be those equal to the carrier frequency plus and minus the modulation frequency (§ 5-2). A keying speed of 50 words per minute, sending sinusoidal dots, would give sidebands only 20 cycles either side of the carrier. However, when harmonics are present in the modulation, the sidebands will extend out on both sides of the signal as far as the frequency of the highest harmonic. The rectangular wave form contains an infinite number of harmonics of the keying frequency, so a carrier modulated by truly rectangular dots would have sidebands covering the entire spectrum. Actually the high-order harmonics are eliminated because of the selectivity of the tuned circuits (§ 2-10) in the transmitter, but there is still enough energy in the lower harmonics to extend the sidebands considerably. Considered from another viewpoint, whenever a pulse of current has a steep front (or back) high frequencies are certain to be present. If the pulse can be slowed down, or caused to *lag*, through a filter the highest-order harmonics are filtered out.

Key clicks — Because the high-order harmonics exist only when the keying character is started or ended (when the amplitude is building up or dying down) their effects outside the normal communication channel are observed as pulses of very short duration. These pulses are called *key clicks*.

Tests have shown that practically all operators prefer to copy a signal which is "solid" on the "make" end of each dot or dash; i.e., one that does not build up too slowly, but just slowly enough to have a slight click when the key is closed. The same tests indicate that the most pleasing and least difficult signal to copy, particularly at high speeds, is one that has a fairly soft "break" characteristic; i.e., one that has practically no click as the key is opened. A signal with heavy clicks on both

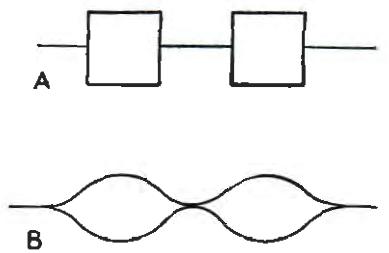


Fig. 601 — Extremes of possible keying waveshapes. A, rectangular characters; B, sine-wave characters.

make and break is difficult to copy at high speeds (and also causes considerable interference), but if it is too "soft" the dots and dashes will tend to run together. It is relatively simple to adjust the keying of a transmitter so that for all normal hand speeds (15 to 40 w.p.m.) the readability will be satisfactory while the keying still will not cause interference to reception of other signals near the frequency of the transmitter.

Break-in keying — Since in code transmission there are definite intervals, between dots and dashes and between words, when no power is being radiated by the transmitter it is possible, with suitable keying methods, to allow the receiver to operate continuously and thus be capable of receiving incoming signals during the keying intervals. This practice facilitates communication because the receiving operator can signal the transmitting operator, by holding down the key of his transmitter, whenever he has failed to copy part of the message and thus obtain a repetition of the missing part without loss of time. This is called *break-in* operation.

Frequency stability — Keying should have no effect upon the output frequency of a properly designed and adjusted transmitter. However, in many instances keying will cause a "chirp," or small frequency change at the instant of closing or opening the key, which makes the signal difficult to read. Multi-stage transmitters keyed in a stage subsequent to the oscillator are usually free from this condition unless the keying causes line-voltage changes which in turn affect the frequency of the oscillator. When the oscillator is keyed for break-in operation special care must be taken to insure that the signal does not have keying chirps.

Selecting the stage to key — It is advantageous from an operating standpoint to design the c.w. transmitter for break-in operation. In ordinary cases this dictates that the oscillator be keyed, since a continuously-running oscillator will create interference in the receiver and thus prevent break-in operation on or near the transmitter frequency. On the other hand, it is easier to avoid a chirpy signal by keying a buffer or amplifier stage. In either case, the tubes following the keyed stage must be provided with sufficient fixed bias to limit the plate currents to safe values when the key is up and they are not being excited (§ 8-9). Complete cut-off reduces the possibility of a back-wave if a stage other than the oscillator is keyed, but the keying waveform is not as well preserved and some clicks can be introduced even though the keyed stage itself produces no clicks. It is a good general rule to bias the tubes to take a key-up plate current equal to about 5% of the normal key-down value.

Keyed power — The power broken by the key is an important consideration, both from the standpoint of safety for the operator and arcing at the key contacts. Keying the oscillator or a low-power stage is favorable in both respects. The use of a keying relay is highly recommended when a high-power circuit is keyed.

• 6-2 KEYING CIRCUITS

Plate-circuit keying — Any stage of the transmitter can be keyed by opening and closing the plate power circuit. Two methods are shown in Fig. 602. In *A* the key is in series with the negative lead from the plate power supply to the keyed stage. It could also be placed in the positive lead, although this is to be avoided whenever possible because the key is necessarily at the plate voltage above ground and there is danger of shock unless a keying relay is used.

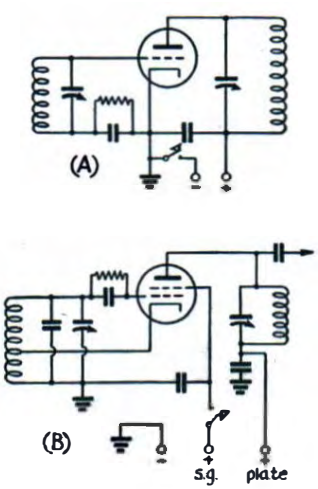


Fig. 602 — A, plate keying; B, screen-grid keying. Oscillator circuits are shown in both cases, but the keying methods also can be used with amplifiers.

Fig. 602-B shows the key in the screen-supply lead of an electron-coupled oscillator. This can be considered to be a variation of plate keying.

The circuits of Figs. 602-A and B respond well to the use of key-click filters, and are particularly suitable for use with crystal and self-controlled oscillators operating at low plate voltage and power input.

Power-supply keying — A variation of plate keying, in which the keying is introduced in the power supply itself rather than between the power supply and transmitter, is illustrated by the diagrams in Fig. 603.

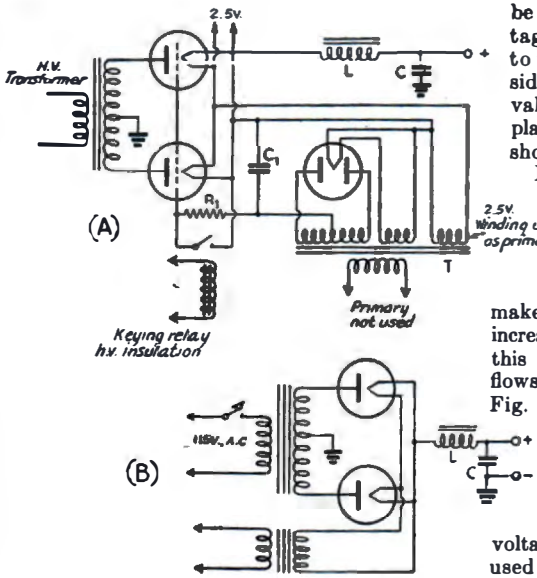


Fig. 603 — Power-supply keying. Grid-control rectifiers are used in A. Transformer T is a small multiple-secondary unit of the type used in receiver power supplies, and is used in conjunction with the full-wave rectifier tube to develop bias voltage for the grids of the high-voltage rectifiers. R_1 limits the load on the bias supply when the keying relay is closed; 50,000 ohms is a suitable value. C_1 may be 0.1 μfd. or larger. L and C constitute the smoothing filter for the high-voltage supply in both circuits. B shows primary keying.

Fig. 603-A shows the use of grid-controlled rectifier tubes (§ 3-5) in the power supply. Keying is accomplished by applying suitable bias to the grids to cut off plate current flow when the key is open, and removing the bias when the key is closed. Since this circuit is used only with high-voltage supplies, a well-insulated keying relay is a necessity. Direct keying of the primary of the plate power transformer for the keyed stage or stages is shown in Fig. 603-B. This and the method at B inherently have a keying lag because of the time constant (§ 2-6) of the smoothing filter. If enough filter is provided to reduce ripple to a low percentage (§ 8-4) the lag (§ 6-1) is too great to permit crisp keying at speeds above about 25 words per minute, although this type of keying is very effective in eliminating key clicks. A single-section filter (§ 8-6) is about the most that can be used for a reasonably-good keying characteristic.

Blocked-grid keying — Keying may be accomplished by applying sufficient negative bias voltage to a control or suppressor grid to cut off plate current flow when the key is open, and by removing this *blocking* bias when the key is closed. The blocking bias voltage must be sufficient to overcome the r.f. grid voltage, in the case where the bias is applied to the control grid, and hence must be considerably higher than the nominal cut-off value for the tube at the operating d.c. plate voltage. The fundamental circuits are shown in Fig. 604.

In both circuits the key is connected in series with a resistor, R_1, which limits the current drain on the blocking-bias source when the key is closed. R_2C_1 is a resistance-capacity filter (§ 2-11) for controlling the lag on make and break of the key circuit. The lag increases as the time constant (§ 2-6) of this circuit is made larger. Since grid current flows through R_2 when the key is closed in Fig. 604-A, additional operating bias is developed, hence somewhat less bias is needed from the regular bias supply. The operating and blocking biases can be obtained from the same supply, if desired, by utilizing suitable taps on a voltage divider (§ 8-10) or if no fixed bias is used R_2 can be the regular grid leak (§ 3-6) for the stage.

With blocked-grid keying a relatively small direct current is broken as compared to other systems, thus sparking at the key is reduced. The keying characteristic (lag) readily can be controlled by choice of values for C_1 and R_2.

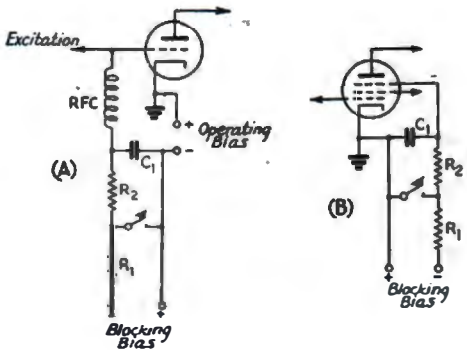


Fig. 604 — Blocked-grid keying. R_1, the current-limiting resistor, should have a value of about 50,000 ohms. C_1 may have a capacity of 0.1 to 1 μfd., depending upon the keying characteristics desired. R_2 is similarly variable, values being of the order of 5000 to 10,000 ohms in most cases.

Cathode keying — Opening the d.c. circuits of both plate and grid simultaneously is called *cathode keying*, or *center-tap* keying with a directly-heated filament-type tube, since in the latter case the key is placed in the filament-transformer center-tap lead. The circuits are shown in Fig. 605.

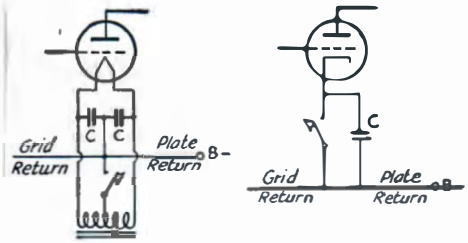


Fig. 605 — Center-tap and cathode keying. Condenser C is an r.f. by-pass condenser having a capacity of 0.001 to 0.01 μfd.

Cathode keying results in less sparking at the key contacts, for the same plate power, as compared with keying in the plate-supply lead. When used with an oscillator it does not respond as readily to key-click filtering (§ 6-3) as does plate keying, but there is little difference in this respect between the two systems when an amplifier is keyed.

● 6-3 KEY-CLICK REDUCTION

R.f. filters — A spark at the key contacts, even though minute, will cause a damped oscillation to be set up in the keying circuit which may modulate the transmitter output or may simply be radiated by the wiring to

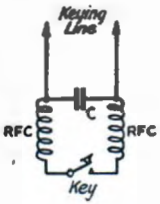


Fig. 606 — R.f. filter for eliminating effects of sparking at key contacts. Values are discussed in the text.

the key. Interference from this source is usually confined to the immediate vicinity of the transmitter, and is similar in nature and effects to the click which is frequently heard in a receiver when an electric light is turned on or off. It can be minimized by isolating the key from the wiring by means of a low-pass filter (§ 2-11), which usually consists of an r.f. choke in each key lead, placed as close as possible to the key, by-passed on the keying-line side by a condenser, as shown in Fig. 606. Suitable values must be determined by experiment. Chokes values may range from 2.5 to 80 millihenrys, and condenser capacities from 0.001 to 0.1 μfd.

This type of filter is required in nearly every keying installation, in addition to the lag circuits discussed in the next paragraph.

Lag circuits — A filter used to give a desired shape to the keying character to eliminate unnecessary sidebands and consequent interference is called a *lag circuit.* In one form, suitable for the circuits of Figs. 602 and 605, it consists of a condenser across the key terminals and an inductance in series with one of the leads. This is shown in Fig. 607. The optimum values of capacity and inductance must be found by experiment, but are not especially critical. If a high-voltage low-current circuit is being keyed, a small condenser and large inductance will be necessary, while if a low-voltage high-current circuit is keyed, the capacity required will be high and the inductance small. For example, a 300-volt 6-ma. circuit will require about 30 henrys and 0.05 μfd., while a 300-volt 50-ma. circuit needs about 1 henry and 0.5 μfd. For any given circuit

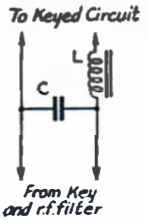


Fig. 607 — Lag circuit for shaping the keying character. Values are discussed in the text.

and fixed values of current and voltage, increasing the inductance will reduce clicks on "make," and increasing the capacity will reduce the clicks on "break."

Blocked-grid keying is adjusted by changing the values of resistors and condensers in the circuit. In Fig. 604, the click on "make" is reduced by increasing the capacity of C_1 and the click on break is reduced by increasing C_1 and/or R_2. The values will vary with the amount of blocking voltage and the grid current. The constants given in Fig. 604 will serve as a first approximation.

Tube keying — A tube keyer is a convenient adjunct to the transmitter because it allows the keying characteristic to be adjusted easily without necessitating condenser and inductance values which may not be readily available. It uses the plate resistance of a tube (or tubes in parallel) to replace the key in a plate or cathode circuit, the keyer tube (or tubes) being keyed by the blocked-grid method (§ 6-2). A typical circuit is shown in Fig. 608. Type 45 tubes are suitable because of their low plate resistance and consequently small voltage drop between plate and cathode. When a tube keyer is used to replace the key in a plate or cathode circuit the power output of the stage will be somewhat reduced because of the voltage drop across the keyer tube, but this can be compensated for by a slight increase in the supply voltage. A tube keyer makes the key itself very safe to handle, since the high resistance in series with the key and blocking voltage prevents shock.

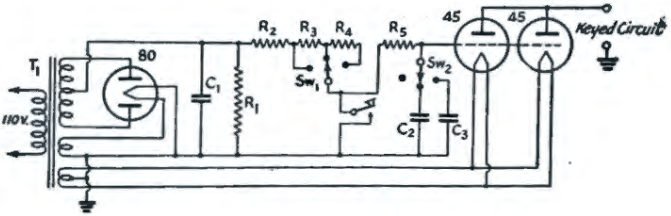


Fig. 608 — Vacuum-tube keyer circuit. The voltage drop across the tubes will be approximately 90 volts with the two type 45 tubes shown when the keyed current is 100 milliamperes. More tubes can be connected in parallel to reduce the drop. Suggested values are as follows:

C_1 — 2 μfd., 600-volt paper.
C_2 — 0.003-μfd. mica.
C_3 — 0.005-μfd. mica.
R_1 — 0.25 megohm.
R_2 — 50,000 ohms, 10-watt.
R_3, R_4 — 5 megohms.
R_5 — 0.5 megohm.
Sw_1, Sw_2 — 3-position 1-circuit rotary switch.
T_1 — 325 volts each side c.t., with 5-volt and 2.5-volt windings.

A wider range of lag adjustment may be obtained by using additional resistors and condensers. Suggested values of capacity, in addition to the above (C_2 and C_3), are 0.001 and 0.002 μfd. Resistors in addition to R_3 could be 2, 2, 3 and 5 megohms.

● 6-4 CHECKING TRANSMITTER KEYING

Clicks — Transmitter keying can be checked by listening to the signal on a superheterodyne receiver. The antenna should be disconnected so that the receiver does not overload and, if necessary, the r.f. gain can be reduced as well. Listening with the beat oscillator and a.v.c. off, the keying should be adjusted so that a slight click is heard as the key is closed, but practically none is heard as the key is released. When the keying constants have been adjusted to meet this condition, the clicks will be about optimum for all normal amateur work. If the clicks are too pronounced, they will cause interference with other amateurs and possibly to nearby broadcast receivers.

Chirps — Keying chirps (instability) may be checked by tuning in the signal or one of its harmonics on the highest frequency range of the receiver and listening with the b.f.o. on and the a.v.c. off. The gain should be sufficient to give moderate signal strength but it should be low enough to preclude the possibility of overloading. Adjust the tuning to give a low-frequency beat note and key the transmitter. Any chirp introduced by the keying adjustment will be readily apparent. By listening to a harmonic, the effect of any instability is magnified by the order of the harmonic and thus made more perceptible.

Oscillator keying — The keying of an amplifier is relatively straightforward and requires no special considerations other than those mentioned, but a few additional precautions are necessary with oscillator keying. Any oscillator, either self-excited or crystal, will key well if it will oscillate at low plate voltages (of the order of one or two volts) and if its change in frequency with plate-voltage changes is negligible. A crystal oscillator will oscillate at low plate voltages if a regenerative type of circuit such as the Tritet or grid-plate (§ 4-5) is used and if an r.f. choke is connected in series with the grid leak to reduce loading on the crystal. Crystal oscillators of this type are generally free from chirp unless there is a relatively large air-gap between the crystal and top plate of the holder, as is the case with a variable-frequency crystal set at the high-frequency end of its range.

Self-controlled oscillators can be made to meet the same requirements by using a high C/L ratio in the tank circuit, low plate and screen currents, and judicious adjustment of the feedback (§ 3-7). A self-controlled oscillator intended to be keyed should be designed for good keying rather than maximum output.

Stages following keying — When a keying filter is being adjusted, the stages following the keyed tube should be made inoperative by removing the plate voltage. This facilitates monitoring the keying without the introduction of additional effects. The following stages should then be added one at a time, checking the keying after each addition. An increase in click intensity (for the same carrier strength) indicates that the clicks are being added in the stages following the one which is keyed. The fixed bias on such stages should be sufficient to reduce the idling plate current (no excitation) to a low value but not to zero. Under these conditions any instability or tendency toward parasitic oscillations, either of which can adversely affect the keying characteristic, usually will evidence itself. It is particularly necessary that the transmitter be free from parasitic oscillations, since they can be the cause of key clicks which do not respond to the methods of treatment outlined in the preceding sections.

CHAPTER SEVEN

Receiver Principles and Design

• 7-1 ELEMENTS OF RECEIVING SYSTEMS

Basic requirements — The purpose of a radio receiving system is to abstract energy from passing radio waves and convert it into a form which conveys the intelligence contained in the signal. It must also be able to select a desired signal and eliminate those not wanted. The fundamental processes involved are amplification and detection.

Detection — The high frequencies used for radio signalling are well beyond the audio-frequency range (§ 2-7) and therefore cannot be used directly to actuate a loudspeaker. Neither can they be used to operate other devices, such as relays, by means of which a message might be transmitted. The process of converting a modulated radio-frequency wave to a usable low frequency, called *detection* or *demodulation*, is essentially that of rectification (§ 3-1). The modulated carrier (§ 5-1) is thereby converted to a unidirectional current the amplitude of which will vary at the same rate as the modulation. These low-frequency variations are readily applied to a headset, loudspeaker, or other form of electro-mechanical device.

Code signals — The dots and dashes of code (c.w.) transmissions are rectified as described, but in themselves can produce no audible tone in a headset or loudspeaker because they are of constant amplitude. For aural reception it is necessary to introduce a second radio frequency, differing from the signal frequency by a suitable audio frequency, into the detector circuit to produce an audible beat (§ 2-13). The frequency difference, and hence the *beat note*, is generally of the order of 500 to 1000 cycles since these tones are within the range of optimum response of both the ear and the headset. If the source of the second radio frequency is a separate oscillator the system is known as *heterodyne* reception; if the detector itself is made to oscillate and produce the second frequency, it is known as an *autodyne* detector.

Amplification — To build up weak signals to usable output level, modern receivers employ considerable amplification — often of the order of hundreds of thousands of times. Amplifiers are used at the frequency of the incoming signal (*r.f. amplifiers*), after detection (*a.f. amplifiers*) and, in the superheterodyne receiver, at one or more intermediate radio frequencies (*i.f. amplifiers*). The r.f. and i.f. amplifiers practically always employ tuned circuits.

Types of receivers — Receivers may vary in complexity from a simple detector with no amplification to multi-tube arrangements having amplification at several different radio frequencies as well as at audio frequency. A regenerative detector (§ 7-14) with or without audio frequency amplification is known as a *regenerative receiver;* if the detector is preceded by one or more tuned radio-frequency amplifier stages the combination is known as a *t.r.f. (tuned radio frequency) receiver.* The *superheterodyne receiver* (§ 7-8) employs r.f. amplification at a fixed intermediate frequency as well as at the frequency of the signal itself, the latter being converted by the heterodyne process to the intermediate frequency.

At ultra-high frequencies the superregenerative detector (§ 7-4), usually with audio amplification, is used in the *superregenerative receiver,*

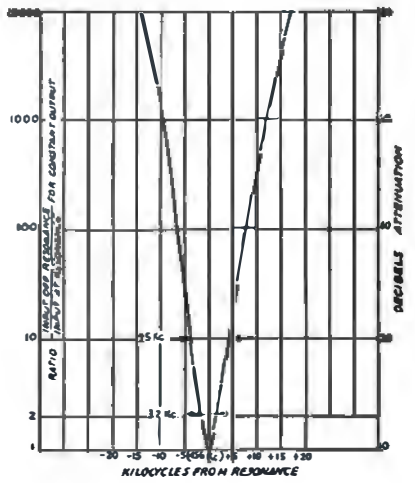


Fig. 701 — Selectivity curve of a modern superheterodyne receiver. The relative response is plotted against deviations above and below the resonance frequency. The scale at the left is in terms of voltage ratios; the corresponding decibel steps (§ 3-3) are shown at the right.

providing high amplification of weak signals with simple circuit arrangements.

● 7-2 RECEIVER CHARACTERISTICS

Sensitivity — Sensitivity is defined as the strength of the signal (usually expressed in microvolts) which must be applied to the input terminals of the receiver to produce a specified audio-frequency power output at the loudspeaker or headset. It is a measure of the amplification or gain.

Signal-to-noise ratio — Every receiver generates some noise of a hiss-like character, and signals weaker than the noise cannot be separated from it no matter how much amplification is used. This relation between noise and a weak signal is expressed by the *signal-to-noise* ratio. It can be defined in various ways, one simple one being to give it as the ratio of signal power output to noise output from the receiver at a specified value of modulated carrier voltage applied to the input terminals.

Since the noise is uniformly distributed over the whole spectrum, its effect will depend upon the selectivity of the receiver, greater selectivity giving smaller noise output and hence a higher signal-to-noise ratio.

Selectivity — Selectivity is the ability of a receiver to discriminate against signals of frequencies differing from that of the desired signal. The overall selectivity will depend upon the selectivity of the individual tuned circuits and the number of such circuits.

The selectivity of a receiver is shown graphically by drawing a curve which gives the ratio of signal strength required at various frequencies off resonance, to the signal strength at resonance, to give constant output. A *resonance curve* of this type (taken on a typical communications-type superheterodyne receiver) is shown in Fig. 701. The *band-width* is the width of the resonance curve (in cycles or kilocycles) at a specified ratio; in Fig. 701, the band-widths are indicated for ratios of 2 and 10.

Selectivity for signals within a few kilocycles of the desired signal frequency is called *adjacent-channel* selectivity, to distinguish it from the discrimination against signals considerably removed from the desired frequency.

Stability — Stability of a receiver is its ability to give constant output, over a period of time, from a signal of constant strength and frequency. Primarily, it means the ability to stay tuned to a given signal, although a receiver which at some settings of its controls has a tendency to break into oscillation, or "howl," is said to be unstable.

The stability of a receiver is affected principally by temperature variations, voltage changes, and constructional features of a mechanical nature.

Fidelity — Fidelity is the relative ability of the receiver to reproduce in its output the modulation (keying, 'phone, etc.) carried by the incoming signal. For exact reproduction, the band-width must be great enough to accommodate the highest modulation frequency, and the relative amplitudes of the various frequency components within the band must not be changed.

● 7-3 DETECTORS

Characteristics — The important characteristics of a detector are its sensitivity, fidelity or linearity, resistance, and signal-handling capability.

Detector *sensitivity* is the ratio of audio-frequency output to radio-frequency input. *Linearity* is a measure of the ability of the detector to reproduce, as an audio frequency, the exact form of the modulation on the incoming signal. The *resistance* or *impedance* of the detector is important in circuit design, since a relatively low resistance means that

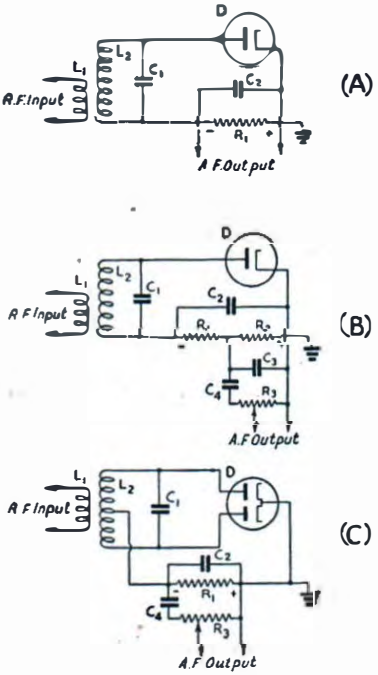


Fig. 702 — Simplified and practical diode detector circuits. A, the elementary half-wave diode detector; B, a practical cricuit, with r.f. filtering and audio output coupling; C, full-wave diode detector, with output coupling indicated. The circuit L_2C_1 is tuned to the signal frequency; typical values for C_2 and R_1 in A and C are 250 $\mu\mu$fd. and 250,000 ohms, respectively; in B, C_2 and C_3 are 100 $\mu\mu$fd. each; R_1, 50,000 ohms; and R_2, 250,000 ohms. C_4 is 0.1 μfd. and R_3, 0.5 to 1 megohm in all three diagrams.

power is consumed in the detector. The *signal-handling capability* means the ability of the detector to accept signals of a specified amplitude without overloading.

Diode detectors — The simplest detector is the diode rectifier. Circuits for both half-wave and full-wave (§ 8-3) diodes are given in Fig. 702. The simplified half-wave circuit at 702-A includes the r.f. tuned circuit L_2C_1, with a coupling coil L_1 from which the r.f. energy is fed to L_2C_1; the diode, D, and the load resistance R_1 and by-pass condenser C_2. The flow of rectified r.f. current through R_1 causes a d.c. voltage to develop across its terminals, and this voltage varies with the modulation on the signal. The − and + signs show the polarity of the voltage. Variation in amplitude of the r.f. signal with modulation causes corresponding variations in the value of the d.c. voltage across R_1. The load resistor, R_1, usually has a rather high value so that a fairly large voltage will develop from a small rectified-current flow.

In the circuit at 702-B, R_1 and C_2 have been divided for the purpose of filtering r.f. from the output circuit (§ 2-11); any r.f. voltage in the output may cause overloading of a succeeding amplifier tube. These audio-frequency variations can be transferred to another circuit through a coupling condenser, C_4 in Fig. 702, to a load resistor R_3, which usually is a "potentiometer" so that the volume can be adjusted to a desired level.

The full-wave diode circuit at 702-C is practically identical in operation to the half-wave circuit, except that both halves of the r.f. cycle are utilized. The full-wave circuit has the advantage that very little r.f. voltage appears across the load resistor, R_1, because the midpoint of L_2 is at the same potential as the cathode or "ground" for r.f.

The reactance of C_2 must be small compared to the resistance of R_1 at the radio frequency being rectified, but at audio frequencies must be relatively large compared to R_1 (§ 2-8, 2-13). This condition is satisfied by the values shown. If the capacity of C_2 is too large, the response at the higher audio frequencies will be low.

Compared with other detectors, the sensitivity of the diode is low. Since the diode consumes power, the Q of the tuned circuit is reduced, bringing about a reduction in selectivity (§ 2-10). The linearity is good, however, and the signal-handling capability is high.

Grid-leak detectors — The grid-leak detector is a combination diode rectifier and audio-frequency amplifier. In the circuit of Fig. 703-A, the grid corresponds to the diode plate, and the rectifying action is exactly the same. The d.c. voltage from rectified current flow through the grid leak, R_1 biases the grid negatively with respect to cathode, and the audio-frequency variations in voltage across R_1 are amplified through the tube just as in a normal a.f. amplifier. In the plate circuit, R_2 is the plate load resistance (§ 3-3) and C_3 a by-pass condenser to eliminate r.f. in the output circuit. C_4 is the output coupling condenser. With a triode, the load resistor R_2 may be replaced by an audio transformer, T, as shown, in which case C_4 is not used.

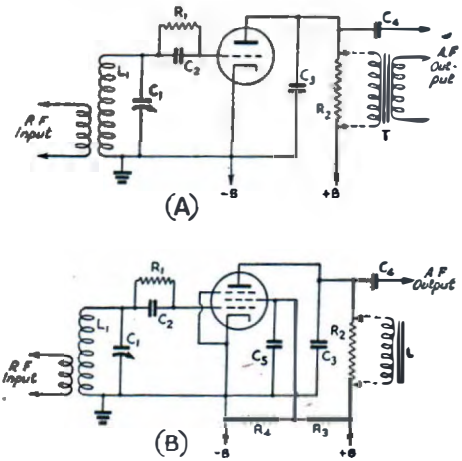


Fig. 703 — Grid-leak detector circuits, A, triode; B, pentode. A tetrode may be used in the circuit of B by neglecting the suppressor-grid connection. Transformer coupling may be substituted for resistance coupling in A, or a high-inductance choke may replace the plate resistor in B. L_1C_1 is a circuit tuned to the signal frequency. The grid leak, R_1, may be connected directly from grid to cathode instead of across the grid condenser as shown. The operation with either connection will be the same. Representative values are:

Component	*Circuit A*	*Circuit B*
C_2	100 to 250 μμfd.	100 to 250 μμfd.
C_3	0.001 to 0.002 μfd.	250 to 500 μμfd.
C_4	0.1 μfd.	0.1 μfd.
C_5		0.5 μfd. or larger.
R_1	1 to 2 megohms.	1 to 5 megohms.
R_2	50,000 ohms.	100,000 to 250,000 ohms.
R_3		50,000 ohms.
R_4		20,000 ohms.
T	Interstage audio transformer.	
L		500-henry choke.

The plate voltage in A should be about 50 volts for best sensitivity. In B the screen voltage should be about 30 volts, plate voltage from 100 to 250.

Since audio amplification is added to rectification, the grid-leak detector has considerably greater sensitivity than the diode. The sensitivity can be further increased by using a screen-grid tube instead of a triode, as at 703-B. The operation is equivalent to that of the triode circuit. C_5, the screen by-pass condenser, should have low reactance (§ 2-8, 2-13) for both radio and audio frequencies. R_3 and R_4 constitute a voltage divider (§ 8-10)

from the plate supply to furnish the proper d.c. voltage to the screen. In both circuits, C_2 must have low r.f. reactance and high a.f. reactance compared to the resistance of R_1; the same consideration applies to C_3 with respect to R_2.

The sensitivity of the grid-leak detector is higher than that of any other type. Like the diode, it "loads" the tuned circuit and reduces its selectivity. The linearity is rather poor, and the signal-handling capability is limited.

Plate detectors — The plate detector is arranged so that rectification of the r.f. signal takes place in the plate circuit of a triode, tetrode or pentode, as contrasted to the grid rectification just described. Sufficient negative bias is applied to the grid to bring the plate

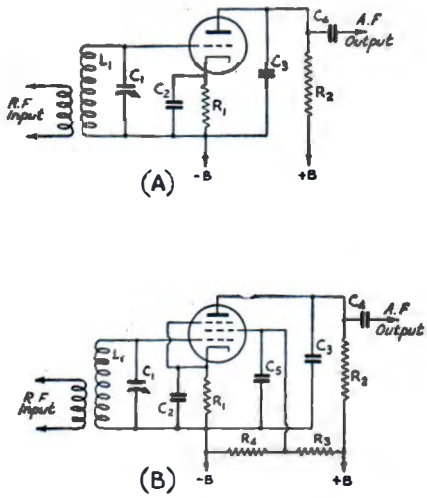


Fig. 704 — Circuits for plate detection. A, triode; B, pentode. L_1C_1 is tuned to the signal frequency. Typical values for other constants are:

Component	Circuit A	Circuit B
C_2	0.5 μfd. or larger.	0.5 μfd. or larger.
C_3	0.001 to 0.002 μfd.	250 to 500 μμfd.
C_4	0.1 μfd.	0.1 μfd.
C_5		0.5 μfd. or larger.
R_1	10,000 to 20,000 ohms.	10,000 to 20,000 ohms.
R_2	50,000 to 100,000 ohms.	100,000 to 250,000 ohms
R_3		50,000 ohms.
R_4		20,000 ohms.

Plate voltages from 100 to 250 volts may be used. Screen voltage in B should be about 30 volts.

current nearly to the cut-off point, so that the application of a signal to the grid circuit causes an increase in average plate current. The average plate current follows the changes in signal amplitude in a fashion similar to the rectified current in a diode detector.

Circuits for triodes and pentodes are given in Fig. 704. C_3 is the plate by-pass condenser, R_1 is the cathode resistor which provides the operating grid bias (§ 3-6), and C_2 is a by-pass, for both radio and audio frequencies, across R_1 (§ 2-13). R_2 is the plate load resistance (§ 3-3) across which a voltage appears as a result of the rectifying action described above. C_4 is the output coupling condenser. In the pentode circuit at B, R_3 and R_4 form a voltage divider to supply the proper potential (about 30 volts) to the screen, and C_5 is a by-pass condenser between the screen and cathode. C_5 must have low reactance for both radio and audio frequencies.

The plate detector is more sensitive than the diode, since there is some amplifying action in the tube, but less so than the grid-leak detector. It will handle considerably larger signals than the grid-leak detector, but is not quite as tolerant in this respect as the diode. Linearity, with the self-biased circuits shown, is good. Up to the overload point, the detector takes no power from the tuned circuit and hence does not affect its Q and selectivity (§ 2-10).

Infinite impedance detector — The circuit of Fig. 705 combines the high signal-handling

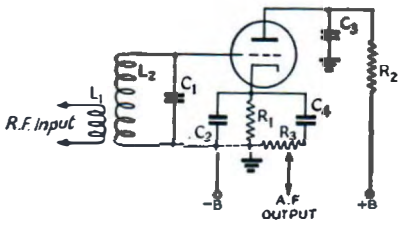


Fig. 705 — The infinite impedance detector. L_2C_1 is tuned to the signal frequency. Typical values for other constants are:
C_2 — 250 μμfd.
C_3 — 0.5 μfd.
C_4 — 0.1 μfd.
R_1 — 0.15 megohm.
R_2 — 25,000 ohms.
R_3 — 0.25-megohm volume control.
A tube having a medium amplification factor (about 20) should be used. The plate voltage should be 250 volts.

capabilities of the diode detector with low distortion (good linearity) and, like the plate detector, does not load the tuned circuit to which it is connected. The circuit resembles that of the plate detector except that the load resistance, R_1, is connected between cathode and ground and is thus common to both grid and plate circuits, giving negative feedback for the audio frequencies. The cathode resistor is by-passed for r.f. (C_1) but not for audio (§ 2-13), while the plate circuit is by-passed

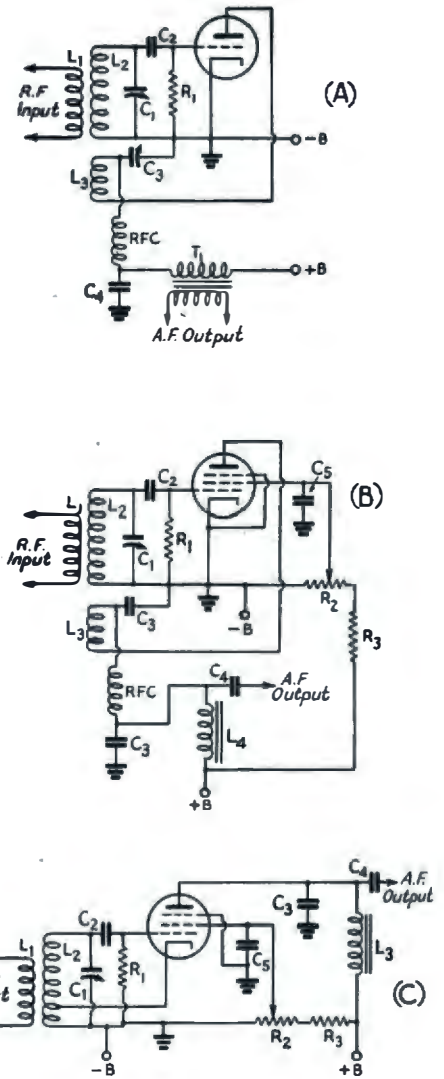


Fig. 706 — Triode and pentode regenerative detector circuits. The circuit L_2C_1 is tuned to the signal frequency. The grid condenser, C_2, should have a value of about 100 $\mu\mu$fd. in all circuits; the grid leak, R_1, may range in value from 1 to 5 megohms. The tickler coil, L_3, will ordinarily have from 10% to 25% of the number of turns on L_2; in C, the cathode tap is about 10% of the number of turns on L_2 above ground. Regeneration control condenser C_3 in A should have a maximum capacity of 100 $\mu\mu$fd. or more; by-pass condensers C_3 in B and C are likewise 100 $\mu\mu$fd. C_5 is ordinarily 1 μfd. or more; R_2, 50,000-ohm potentiometer; R_3, 50,000 to 100,000 ohms. L_4 in B (L_3 in C) is a 500-henry inductance, C_4 is 0.1 μfd. in both circuits. T_1 in A is a conventional audio transformer for coupling from the plate of a tube to a following grid. RFC is 2.5 mh.

In A, the plate voltage should be of the order of 50 volts for best sensitivity. The pentode circuits require about 30 volts on the screen; plate voltage may be from 100 to 250 volts.

to ground for both audio and radio frequencies. R_2 with C_2 forms an RC filter (§ 2-11) to isolate the plate from the "B" supply at audio frequencies.

The plate current is very low at no signal, increasing with signal as in the case of the plate detector. The voltage drop across R_1 similarly increases with signal because of the increased plate current. Because of this and the fact that the initial drop across R_1 is large, the grid cannot be driven positive with respect to the cathode by the signal, hence no grid current can be drawn.

• 7-4 REGENERATIVE DETECTORS

Circuits — By providing controllable r.f. feedback or regeneration (§ 3-3) in a triode or pentode detector circuit, the incoming signal can be amplified many times, thereby greatly increasing the sensitivity of the detector. Regeneration also increases the effective Q of the circuit and hence increases the selectivity (§ 2-10), by virtue of the fact that the maximum regenerative amplification takes place at only the frequency to which the circuit is tuned. The grid-leak type of detector is most suitable for the purpose. Except for the regenerative connection, the circuit values are identical with those previously described for this type of detector, and the same considerations apply.

Fig. 706 shows the circuits of regenerative detectors of various types. The circuit of A is for a triode tube, with a variable by-pass condenser, C_3, in the plate circuit to control regeneration. When the capacity is small the tube does not regenerate, but as it increases toward maximum its reactance (§ 2-8) becomes smaller until a critical value is reached where there is sufficient feed-back to cause oscillation. If L_2 and L_3 are wound end to end in the same direction, the plate connection is to the outside of the plate or "tickler" coil, L_3, when the grid connection is to the outside of the tuned circuit coil, L_2.

The circuit of B is for a screen-grid tube, regeneration being controlled by adjustment of the screen-grid voltage. The tickler, L_3, is in the plate circuit. The portion of the control resistor between the rotating contact and ground is by-passed by a large condenser (0.5 μfd. or more) to filter out scratching noise when the arm is rotated (§ 2-11). The feedback should be adjusted by varying the number of turns on L_3 or the coupling (§ 2-11) between L_2 and L_3 so that the tube just goes into oscillation at a screen voltage of approximately 30 volts.

Circuit C is identical with B in principle of operation, except that the oscillating circuit is of the Hartley type (§ 3-7). Since the screen and plate are in parallel for r.f. in this circuit,

only a small amount of "tickler" — that is, relatively few turns between cathode tap and ground — is required for oscillation.

Adjustment for smooth regeneration — The ideal regeneration control would permit the detector to go into and out of oscillation smoothly, would have no effect on the frequency of oscillation, and would give the same value of regeneration regardless of frequency and the loading on the circuit. In practice the effects of loading, particularly the loading that occurs when the detector circuit is coupled to an antenna, are difficult to overcome. Likewise, the regeneration is affected by the frequency to which the grid circuit is tuned.

In all circuits it is best to wind the tickler at the ground or cathode end of the grid coil, and to use as few turns on the tickler as will allow the detector to oscillate easily over the whole tuning range with the proper plate (and screen, if a pentode) voltage. Should the tube break into oscillation suddenly, making a click, as the regeneration control is advanced, the operation can frequently be made smooth by changing the grid-leak resistance to a higher or lower value. The wrong grid leak plus too-high plate and screen voltage are the most frequent causes of lack of smoothness in going into oscillation.

Antenna coupling — If the detector is coupled to an antenna, slight changes in the antenna constants (as when the wire swings in a breeze) affect the frequency of the oscillations generated, and thereby the beat frequency when c.w. signals are being received. The tighter the antenna coupling is made, the greater will be the feedback required or the higher will be the voltage necessary to make the detector oscillate. The antenna coupling should be the maximum that will allow the detector to go into oscillation smoothly with the correct voltages on the tube. If capacity coupling (§ 2-11) to the grid end of the coil is used, only a very small amount of capacity will be needed to couple to the antenna. Increasing the capacity increases the coupling.

At frequencies where the antenna system is resonant the absorption of energy from the oscillating detector circuit will be greater, with the consequence that more regeneration is needed. In extreme cases it may not be possible to make the detector oscillate with normal voltages, causing so-called "dead spots". The remedy for this is to loosen the antenna coupling to the point which permits normal oscillation and smooth regeneration control.

Body capacity — A regenerative detector occasionally shows a tendency to change frequency slightly as the hand is moved near the dial. This condition (*body capacity*) can be caused by poor design of the receiver or by the antenna, if the detector is coupled to an antenna. If the body capacity is present when the antenna is disconnected, it can be eliminated by better shielding, and sometimes by r.f. filtering of the 'phone leads. Body capacity present only when the antenna is connected is caused by resonance effects in the antenna which tend to cause part of a standing wave (§ 2-12) of r.f. voltage to appear on the ground lead and thus raise the whole detector circuit above ground potential. A good, short ground connection should be made to the receiver and the length of the antenna varied electrically (by adding a small coil or variable condenser in the antenna lead) until the effect is minimized. Loosening the coupling to the antenna circuit also will help reduce body capacity.

Hum — Power-supply frequency hum may be present in a regenerative detector, especially when it is used in an oscillating condition for c.w. reception, even though the plate supply is free from ripple (§ 8-4). It may result from the use of a.c. for the tube heater, but effects of this type are normally troublesome only when the circuit of Fig. 706-C is used, and then only at 14 Mc. and higher frequencies. Connecting one side of the heater supply to ground, or grounding the center-tap of the heater transformer winding, is good practice to reduce hum, and the heater wiring should be kept as far as possible from the r.f. circuits.

House wiring, if of the "open" type, will have a rather extensive electrostatic field which may cause hum if the detector tube, grid lead, grid condenser and leak are not electrostatically shielded. This type of hum is easily recognizable because of its rather high pitch, a result of harmonics (§ 2-7) in the power-supply system. The hum is caused by a species of grid modulation (§ 5-4) because the field causes a small a.c. voltage to develop across the grid leak.

Antenna resonance effects frequently cause a hum, of the same nature as that just described, which is most intense at the various resonance points and hence varies with tuning. For this reason it is called tunable hum. It is prone to occur with a rectified a.c. plate supply (§ 8-1) when a standing wave effect of the type described in the preceding paragraph occurs, and is associated with the non-linearity of the rectifier tube in the plate supply. Elimination of antenna resonance effects and by-passing the rectifier plates to cathode usually will cure it.

Tuning — For c.w. reception, the regeneration control is advanced until the detector breaks into a "hiss," which indicates that the detector is oscillating. Further advancing of the regeneration control after the detector starts oscillating will result in a slight decrease

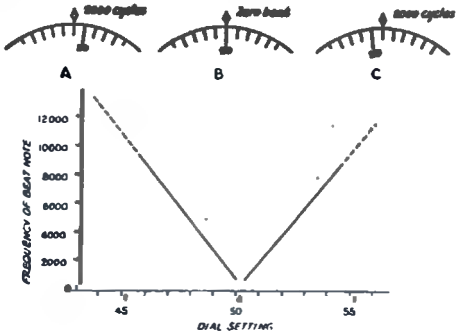


Fig. 707 — As the tuning dial of a receiver is turned past a c.w. signal, the beat note varies from a high tone down through "zero beat" (no audible frequency difference) and back up to a high tone, as shown at A, B and C. The curve is a graphical representation of the action. The beat exists past 8000 or 10,000 cycles but usually is not heard because of the limitations of the audio system of the receiver.

in the strength of the hiss, indicating that the sensitivity is decreasing.

The proper adjustment for the reception of c.w. signals is just after the detector has started to oscillate, when it will be found that c.w. signals can be tuned in and will give a tone with each signal depending on the setting of the tuning control. As the receiver is tuned through a signal, the tone will first be heard as a very high pitch, go down through "zero beat" and then disappear at a high pitch on the other side, as shown in Fig. 707. It will be found that a low-pitched beat-note cannot be obtained with a strong signal because the detector "pulls in" or "blocks," but this condition can be corrected by advancing the regeneration control until the beat-note occurs again. If the regenerative detector is preceded by an r.f. amplifier stage, the blocking can be eliminated by reducing the gain of the r.f. stage. If the detector is coupled to an antenna the blocking condition can be eliminated by advancing the regeneration control or loosening the antenna coupling.

The point just after the receiver starts oscillating is the most sensitive condition for c.w. reception — further advancing of the regeneration control makes the receiver less prone to blocking by strong signals but less capable of receiving weak signals.

If the receiver is in the oscillating condition and a 'phone signal is tuned in, a steady beat-note will result and, while it is possible to listen to 'phone if the receiver can be tuned to exact zero beat, it is more satisfactory to reduce the regeneration to the point just before the receiver goes into oscillation. This is the most sensitive operating point for this type of reception.

Superregeneration — The limit to which ordinary regenerative amplification can be carried is the point at which oscillations commence, since at that point further amplification ceases. The *superregenerative* detector overcomes this limitation by introducing into the detector circuit an alternating voltage of a frequency somewhat above the audible range (of the order of 20 to 100 kilocycles) in such a way as to vary the detector's operating point (§ 3-3). As a consequence of the introduction of this *quench* or *interruption* frequency the detector can oscillate only when the varying operating point is in a region suitable for the production of oscillations. Because the oscillations are constantly being interrupted the regeneration can be greatly increased and the signal will build up to relatively tremendous proportions. The superregenerative circuit is suitable only for the reception of modulated signals, and operates best on ultra-high frequencies where it has found considerable application in simple receivers.

A typical superregenerative circuit for ultra-high frequencies is shown in Fig. 708. The regenerative detector circuit is an ultraudion oscillator (§ 3-7). The quench frequency, obtained from the separate quench oscillator, is

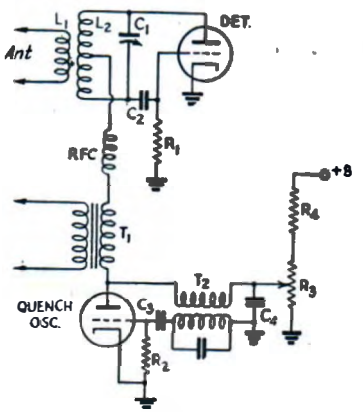


Fig. 708 — Superregenerative detector circuit with separate quench oscillator. L_2C_1 is tuned to the signal frequency. Typical values for other components are as follows:

C_2 — 100 μμfd.
C_3 — 500 μμfd.
C_4 — 0.1 μfd.
R_1 — 5 megohms.
R_2 — 50,000 ohms.
R_3 — 50,000-ohm potentiometer.
R_4 — 50,000 ohms.
T_1 — Audio transformer, plate-to-grid type.
RFC — Radio-frequency choke, constants depending upon frequency of operation. Special low-capacity chokes are required for ultra-high frequencies.

introduced in the plate circuit. Many other circuit arrangements are possible.

If regeneration in an ordinary regenerative circuit is carried sufficiently far, the circuit will break into a low-frequency oscillation simultaneously with that at the operating radio frequency. This low-frequency oscillation has much the same quenching effect as that from a separate oscillator, hence a circuit so operated is called a *self-quenching* superregenerative detector. This type of circuit is more successful at ultra-high than at ordinary communication frequencies. The frequency of the quench oscillation depends upon the feedback and upon the time constant of the grid leak and condenser, the oscillation being a form of "blocking" or "squegging" in which the grid accumulates a strong negative charge which cannot leak off rapidly enough through the grid leak to prevent a relatively slow variation of the operating point.

The superregenerative detector has little selectivity, but discriminates against noise such as that from automobile ignition systems. It also has marked automatic volume control action, since strong signals are amplified to a much smaller extent than weak signals.

Adjustment of superregenerative detectors — Because of the greater amplification, the hiss when the superregenerative detector goes into oscillation is much stronger than with the ordinary regenerative detector. The most sensitive condition is at the point where the hiss first becomes marked. When a signal is tuned in the hiss will disappear to a degree which depends upon the signal strength.

Lack of hiss indicates insufficient feedback at the signal frequency or inadequate quench voltage. Antenna loading effects will cause dead spots similar to those with regenerative detectors and these can be overcome by the same methods. The self-quenching detector may require critical adjustment of the grid leak and grid condenser values for smooth operation, since these determine the frequency and amplitude of the quench voltage.

●7-5 AUDIO-FREQUENCY AMPLIFIERS

General — Audio-frequency amplifiers are used after the detector to increase the power to a level suitable for operating a loud-speaker or, in some cases, a headset. There are seldom more than two stages of a.f. amplification in a receiver, and often only one.

In all except battery-operated receivers, the negative grid bias of audio amplifiers is usually secured from the voltage drop in a cathode resistor (§ 3-6). The cathode resistor must be bypassed by a condenser having low reactance, at the lowest audio frequency to be amplified, compared to the resistance of the cathode resistor (10% or less) (§ 2-8, 2-13). In battery-operated sets, a separate grid-bias battery generally is used.

Headset and voltage amplifiers — The circuits shown in Fig. 709 are typical of those used for voltage amplification and for providing sufficient power for operation of headphones (§ 3-3). Triodes usually are preferred to pentodes because they are better suited to working into an audio transformer or headset, which have impedances of the order of 20,000 ohms.

In these circuits, R_2 is the cathode bias resistor and C_1 the cathode by-pass condenser. R_1, the grid resistor, gives volume control

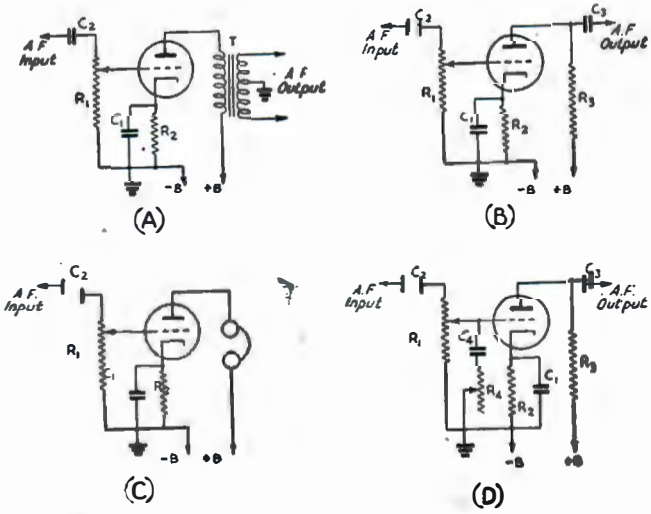


Fig. 709 — Audio amplifier circuits for voltage amplification and headphone output. The tubes are operated as Class-A amplifiers (§ 3-4).

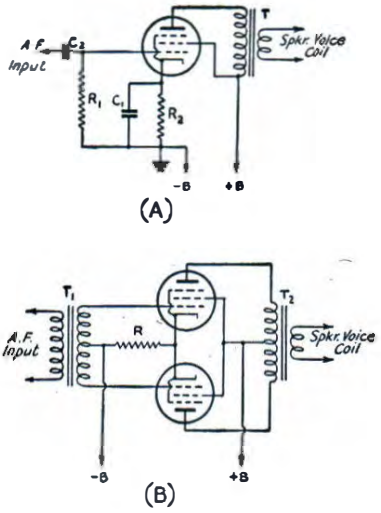


Fig. 710 — Audio power output amplifier circuits. Class A or AB (§ 3-4) amplification is used.

action (§ 5-9). Its value ordinarily is from 0.25 to 1 megohm. C_2 is the input coupling condenser, already discussed under detectors; it is, in fact, identical to C_4 in Figs. 703 and 704 if the amplifier is coupled to a detector.

Power amplifiers — A popular type of power amplifier is the single pentode; the circuit diagram is given in Fig. 710-A. The grid resistor, R_1, may he a potentiometer for volume control as shown at R_1 in Fig. 709. The output transformer T should have a turns ratio (§ 2-9) suitable for the speaker used; most of the small speakers now available are furnished complete with output transformer.

When greater volume is needed a pair of pentodes or tetrodes may be connected in push-pull (§ 3-3) as shown in Fig. 710-B. Transformer coupling to the voltage-amplifier stage is the simplest method of obtaining push-pull input for the amplifier grids. The interstage transformer, T_1, has a center-tapped secondary, with a secondary-to-primary turns ratio of about 2 to 1. An output transformer, T_2, with a center-tapped primary must be used. No by-pass condenser is needed across the cathode resistor, R, since the a.f. current does not flow through the resistor as it does in single-tube circuits.

Tone control — A tone control is a device for changing the frequency response (§ 3-3) of an audio amplifier; usually it is simply a method for reducing high-frequency response. This is helpful in reducing hissing and crackling noises without disturbing the intelligibility of the signal. R_4 and C_4 together in Fig. 709-D form an effective tone control of this type. The maximum effect is secured when R_4 is entirely out of the circuit, leaving C_4 connected between grid and ground. R_4 should be large enough compared to the reactance of C_4 (§ 2-8) so that when it is all in circuit the effect of C_4 on the frequency response is negligible.

• 7-6 RADIO-FREQUENCY AMPLIFIERS

Circuits — Although there are variations in detail, practically all r.f. amplifiers conform to the basic circuit shown in Fig. 711. A screen-grid tube, usually a pentode, is invariably used, since a triode will oscillate when its grid and plate circuits are tuned to the same frequency (§ 3-5). The amplifier operates Class A, without grid current (§ 3-4). The tuned grid circuit, L_1C_1, is coupled through L_2 to the antenna (or, in some cases, to a preceding stage). R_1 and C_2 are the cathode bias resistor and cathode by-pass condenser, C_3 the screen by-pass condenser, and R_2 the screen dropping resistor. L_3 is the primary of the output transformer (§ 2-11), tightly coupled to L_4 which, with C_5, constitutes the tuned circuit feeding the detector or a following amplifier tube. L_1C_1 and L_4C_5 are both tuned to the frequency of the incoming signal.

Shielding — The screen-grid construction prevents feedback (§ 3-3) from plate to grid inside the tube, but in addition it is necessary to prevent transfer of energy from the plate circuit to the grid circuit external to the tube. This is accomplished by enclosing the coils in grounded shielding containers, and by keeping the plate and grid leads well separated. With "single-ended" tubes care in laying out the wiring to obtain the maximum possible physical separation between plate and grid leads is necessary to prevent capacity coupling.

The shield around a coil will reduce the inductance and Q of the coil (§ 2-11) to an extent which depends upon the shielding material and its distance from the coil. Adjustments to the inductance therefore must be made with the shield in place.

By-passing — In addition to shielding, good by-passing (§ 2-13) is imperative. This is not simply a matter of choosing the proper type

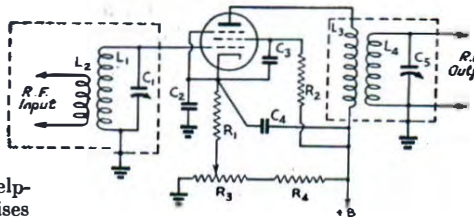


Fig. 711 — The circuit of a tuned radio-frequency amplifier. Circuit values are discussed in the text.

and capacity of by-pass condenser. Short *separate* leads from C_3 and C_4 to cathode or ground are a prime necessity, since at the higher radio frequencies even an inch or two of wire will have enough inductance to provide feedback coupling, and hence cause oscillation, if the wire happens to be common to both the plate and grid circuits.

Gain control — The gain of an r.f. amplifier usually is varied by varying the grid bias. This method is applicable only to variable-μ type tubes (§ 3-5) hence this type usually is found in r.f. amplifiers. In Fig. 711, R_3 and R_4 comprise the gain-control circuit. R_3 is the control resistor (§ 3-6) and R_4 a dropping resistor of such value as to make the voltage across the outside terminals of R_3 about 50 volts (§ 8-10). The gain is maximum with the variable arm all the way to the left (grounded) on R_3 and minimum at the right. R_3 could simply be placed in series with R_1, omitting R_4 entirely, but the range of control is limited when this connection is used.

In a multi-tube receiver, the gain of several stages would be varied simultaneously, a single control sufficing for all. In such a case, the lower ends of the several cathode resistors (R_1) would be connected together and to the movable contact on R_3.

Circuit values — The value of the cathode resistor, R_1, should be calculated for the *minimum* recommended bias for the tube used. The capacities of C_2, C_3 and C_4 must be such that the reactance is low at radio frequencies; this condition is easily met by using 0.01-μfd. condensers at communication frequencies, or 0.001 to 0.002 mica units at ultra-high frequencies up to 112 Mc. R_2 is found by taking the difference between the recommended plate and screen voltages, then substituting this and the rated screen current in Ohm's Law (§ 2-6). R_3 must be selected on the basis of the number of tubes to be controlled; a resistor must be chosen which is capable of carrying, at its low-resistance end, the sum of all the tube currents plus the bleeder current. A resistor of suitable current-carrying capacity being found, the bleeder current necessary to produce a drop through it of about 50 volts can be calculated by Ohm's Law. The same formula will give R_4 using the plate voltage less 50 volts for E and the bleeder current just found for I.

The constants of the tuned circuits will depend upon the frequency range, or band, being covered. A fairly high L/C ratio (§ 2-10) should be used on each band; this is limited, however, by the irreducible minimum capacities. An allowance of 10 to 20 $\mu\mu$fd. should be made for tube and stray capacity, and the minimum capacity of the tuning condenser also must be added.

If the input circuit of the amplifier is connected to an antenna, the coupling coil L_2 should be adjusted to provide critical coupling (§ 2-11) between the antenna and grid circuit. This will give maximum energy transfer. The turns ratio L_1/L_2 will depend upon the frequency, the type of tube used, the Q of the tuned circuit, and the antenna system, and in general is best determined experimentally. The selectivity will increase as the coupling is reduced below this "optimum" value, a consideration which it is well to keep in mind if selectivity is of more importance than maximum gain.

The output circuit coupling depends upon the plate resistance (§ 3-2) of the tube, the input resistance of the succeeding stage, and the Q of the tuned circuit L_4C_5. L_3 is usually coupled as closely as possible to L_4 (this avoids the necessity for an additional tuning condenser across L_3) and the energy transfer is about maximum when L_3 has $\frac{3}{5}$ to $\frac{1}{5}$ as many turns as L_4, with ordinary receiving screen-grid pentodes.

Tube and circuit noise — In any conductor electrons will be moving in random directions simultaneously and, as a result, small irregular voltages are developed across the conductor terminals. The voltage is larger the greater the resistance of the conductor and the higher its temperature. This is known as the *thermal agitation* effect, and it produces a hiss-like noise voltage distributed uniformly throughout the radio-frequency spectrum. The thermal agitation noise voltage appearing across the terminals of a tuned circuit will be the same as in a resistor of a value equal to the parallel impedance of the tuned circuit (§ 2-10) even though the actual circuit resistance is low. Hence the higher the Q of the circuit the greater the thermal agitation noise.

Another component of hiss noise is developed in the tube, because the rain of electrons on the plate is not entirely uniform. Small irregularities caused by gas in the tube also contribute to the effect. Tube noise varies with the type of tube, and is proportional in a general way to the inverse ratio of the mutual conductance (§ 3-2) of the tube to the square root of the plate current.

To obtain the best signal-to-noise ratio, the signal must be made as large as possible at the grid of the tube, which means that the antenna coupling must be adjusted to that end, and also that the Q of the grid tuned circuit must be high. A tube with low inherent noise obviously should be chosen. In an amplifier having good signal-to-noise ratio the thermal agitation noise will be greater than the tube noise. This can easily be checked by grounding the grid through a 0.01-μfd. condenser and observing whether there is a decrease in noise. If there is no change, the tube noise is greatly

predominant, indicating a poor signal-to-noise ratio in the stage. The test is valid only if there is no regeneration in the amplifier. The signal-to-noise ratio will decrease as the frequency is raised because it becomes increasingly difficult to obtain a tuned circuit of high effective Q (§ 7-7).

The first stage of the receiver is the important one from the signal-to-noise ratio standpoint. Noise generated in the second and subsequent stages, while comparable in magnitude to that generated in the first, is masked by the amplified noise and signal from the first stage. After the second stage, further contributions by tubes and circuits to the total noise are inconsequential in any normal receiver.

Tube input resistance — At high frequencies the tube may consume power from the tuned grid circuit even though the grid is not driven positive by the signal. Above 7 Mc. all tubes load the tuned circuit to an extent which depends upon the type of tube. This effect comes about because the time necessary for electrons to travel from the cathode to the grid becomes comparable to the time of one r.f. cycle, and because of a degenerative effect (§ 3-3) of the cathode lead inductance. With certain tube types the input resistance may be as low as a few thousand ohms at 28 Mc. and as low as a few hundred ohms at ultra-high frequencies.

This *input loading* effect is in addition to the normal decrease in the Q of the circuit alone at the higher frequencies because of increased losses in the coil and condenser. Thus the selectivity and gain of the circuit are both adversely affected.

Comparison of tubes — At 7 Mc. and lower frequencies, the signal-to-noise ratio, gain and selectivity of an r.f. amplifier stage are sufficiently high with any of the standard receiving tubes. At 14 Mc. and higher, however, this is no longer true, and the choice of a tube must be based on several conflicting considerations.

Gain is highest with high mutual-conductance pentodes, the 1851 and 1852 being examples of this type. These tubes also develop less noise than any of the others. The input-loading effect is greatest with them, however, so that selectivity is decreased and the tuned-circuit gain is lowered.

Pentodes such as the 6K7, 6J7 and corresponding types in glass have lesser input-loading effects at high frequencies, moderate gain, and relatively-high inherent noise.

The "acorn" and equivalent miniature pentodes are excellent from the input-loading standpoint, the gain is about the same as with the standard types, and the inherent noise is somewhat lower.

Where selectivity is paramount, the acorns are best, standard pentodes second, and the 1851–52 types last. On signal-to-noise ratio the 1851–52 tubes are first, acorns second, and standard pentodes third. The same order holds for overall gain.

At 56 Mc. the standard types are usable, but acorns are capable of better performance because of lesser loading. The 954, 956 and the corresponding types 9001 and 9003 are the only usable types for r.f. amplification at 112 Mc. and higher.

• 7-7 TUNING AND BAND-CHANGING METHODS

Band changing — The resonant circuits which are tuned to the frequency of the incoming signal constitute a special problem in the design of amateur receivers since the amateur frequency assignments consist of groups or bands of frequencies at widely-spaced intervals. The same LC combination cannot be used for, say, 14 Mc. to 3.5 Mc., because of the impracticable maximum-minimum capacity ratio required, and also because the tuning would be excessively critical with such a large frequency range. It is necessary, therefore, to provide a means for changing the circuit constants for various frequency bands. As a matter of convenience, the same tuning condenser usually is retained, but new coils are inserted in the circuit for each band.

There are two favorite methods of changing inductances; one is to use a switch, having an appropriate number of contacts, which connects the desired coil and disconnects the others. The second is to use coils wound on forms with contacts (usually pins) which can be inserted in and removed from a socket.

Band spreading — The tuning range of a given coil and variable condenser will depend

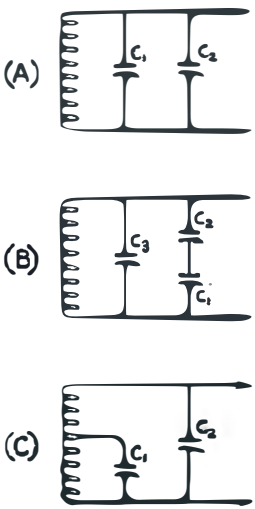


Fig. 712 — Essentials of band-spread tuning systems.

upon the inductance of the coil and the change in tuning capacity. For ease of tuning it is desirable to adjust the tuning range so that practically the whole dial scale is occupied by the band in use. This is called *band-spreading*. Because of the varying widths of the bands, special tuning methods must be devised to give the correct maximum-minimum capacity ratio on each. Several of these are shown in Fig. 712.

In A, a small *band-spread condenser* C_1 (15 to 25 $\mu\mu$fd. maximum capacity) is used in parallel with a condenser, C_2, which is usually large enough (140 to 175 $\mu\mu$fd.) to cover a 2-to-1 frequency range. The setting of C_2 will determine the minimum capacity of the circuit, and the maximum capacity for band-spread tuning will be the maximum capacity of C_1 plus the setting of C_2. The inductance of the coil can be adjusted so that the maximum-minimum ratio will give adequate band-spread. In practicable circuits it is almost impossible because of the non-harmonic relation of the various bands to get full band-spread on all bands with the same pair of condensers, especially when the coils are wound to give continuous frequency coverage on C_2, which is variously called the *band-setting* or *main-tuning* condenser. C_2 must be re-set each time the band is changed.

The method shown at *B* makes use of condensers in series. The tuning condenser, C_1, may have a maximum capacity of 100 $\mu\mu$fd. or more. The minimum capacity is determined principally by the setting of C_3, which usually has low capacity, and the maximum capacity by the setting of C_2, which is of the order of 25 to 50 $\mu\mu$fd. This method is capable of close adjustment to practically any desired degree of band-spread. C_2 and C_3 must be adjusted for each band or else separate pre-adjusted condensers must be switched in.

The circuit at *C* also gives complete spread on each band. C_1, the band-spread condenser, may have any convenient value of capacity; 50 $\mu\mu$fd. is satisfactory. C_2 may be used for continuous frequency coverage ("general coverage") and as a band-setting condenser. The effective maximum-minimum capacity ratio depends upon the capacity of C_2 and the point at which C_1 is tapped on the coil. The nearer the tap to the bottom of the coil, the greater the band-spread, and vice versa. For a given coil and tap, the band-spread will be greater if C_2 is set at larger capacity. C_2 may be mounted in the plug-in coil form and pre-set, if desired. This requires a separate condenser for each band, but eliminates the necessity for re-setting C_2 each time the band is changed.

Ganged tuning — The tuning condensers of the several r.f. circuits may be coupled together mechanically and operated by a single control. This operating convenience involves more complicated construction, both electrically and mechanically. It becomes necessary to make the various circuits *track* — that is, tune to the same frequency at each setting of the tuning control.

True tracking can be obtained only when the inductance, tuning condensers, circuit minimum capacity and maximum capacity are identical in all "ganged" stages. A small *trimmer* or *padding* condenser is connected across the coil so that variations in minimum capacity can be compensated. The fundamental circuit is shown in Fig. 713, where C_1 is the trimmer and C_2 the tuning condenser. The use of the trimmer increases the minimum circuit capacity, but is a necessity for satisfactory tracking. Condensers having maximum capacities of 15 to 30 $\mu\mu$fd. generally are used for the purpose.

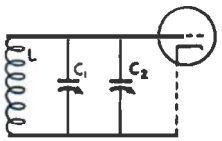


Fig. 713 — Showing the use of a trimmer condenser across the tuned circuit to set the minimum circuit capacity for ganged tuning.

The same methods are applied to band-spread circuits which must be tracked. The circuits are identical with those of Fig. 712, although if both general-coverage and band-spread tuning are to be available, an additional trimmer condenser must be connected across the coil in each circuit shown. If only amateur-band tuning is desired, however, then C_3 in Fig. 712-B, and C_2 in Fig. 712-C serve as trimmers.

The coil inductance can be adjusted by starting with a larger number of turns than necessary, then removing a turn or fraction of a turn at a time until the circuits track satisfactorily. An alternative method of adjusting inductance, providing it is reasonably close to the correct value initially, is to make the coil so that the last turn is variable with respect to the whole coil, or to use a single short-circuited turn the position of which can be varied with respect to the coil. These methods are shown in Fig. 714.

U.H.F. circuits — Tube interelectrode capacities are practically constant for a given tube type regardless of the operating frequency, and the same thing is approximately true of stray circuit capacities. Hence at ultra-high frequencies these capacities become an increasingly larger part of the usable tuning capacity and reasonably-high L/C ratios (§ 2-10) are more difficult to secure as the frequency is raised. Because of this irreducible minimum capacity, standard types of tubes cannot be tuned to frequencies higher than about 200 Mc., even when the inductance in

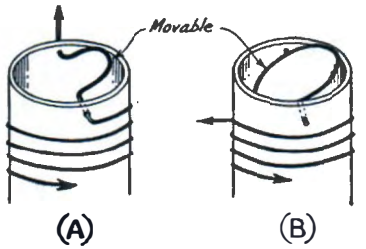


Fig. 714 — Methods of adjusting inductance for ganging. The half turn in A can be moved so that its magnetic field either aids or opposes the field of the coil. The shorted loop in B is not connected to the coil, but operates by induction. It will have no effect on the coil inductance when the plane of the loop is parallel to the axis of the coil, and will give maximum reduction of the coil inductance when perpendicular to the coil axis.

the circuit is simply that of a straight wire between the tube elements.

Along with these capacity effects the input loading (§ 7-6) increases rapidly at ultra-high frequencies so that ordinary tuned circuits have very low effective Q's when connected to the grid circuit of a tube. The effect is still further aggravated by the fact that losses in the tuned circuit itself are higher, causing a still further reduction in Q. For these reasons the frequency limit at which an r.f. amplifier will give any gain is in the vicinity of 60 Mc.,

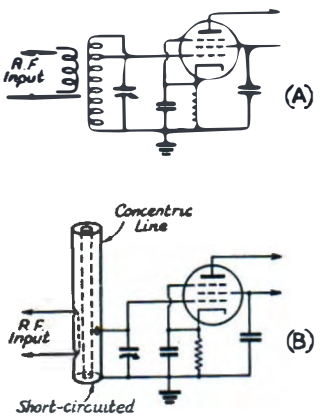


Fig. 715 — Circuits of improved Q for ultra-high frequencies. A, reducing tube loading by tapping down on the resonant circuit; B, use of a concentric-line circuit, with the tube similarly tapped down. The line should be a quarter-wave long, electrically; because of the additional shunt capacity represented by the tube the physical length will be somewhat less than given by the formula (§ 10-5). In general, this reduction in length will be greater the higher the grid tap on the inner conductor. One method of coupling to an antenna or preceding stage is indicated. The coupling turn should be parallel to the axis of the line and insulated from the outer conductor.

with standard tubes, and at higher frequencies there is a loss instead of amplification. The condition can be mitigated somewhat by taking steps to improve the effective Q of the circuit, either by tapping the grid down on the coil as shown in Fig. 715-A or by using a lower L/C ratio (§ 2-10). The Q of the tuned circuit alone can be greatly improved by using a linear circuit (§ 2-12), which when properly constructed will give Q's much higher than those attainable at lower frequencies with conventional coils and condensers. The concentric type of line, Fig. 715-B, is best both from the standpoint of Q and adaptability to non-symmetrical circuits such as are used in receivers. Since the capacity and resistance loading effect of the tube are still present, the Q of such a circuit will be destroyed if the grid-cathode circuit of the tube is connected directly across it, hence tapping down, as shown, is necessary.

Ultra-high frequency amplifiers should employ tubes of the acorn type which have the smallest loading effect as well as low inter-electrode capacities. This is because the smaller loading effect means higher input resistance and hence, for a given loaded Q of the tuned circuit, higher voltage developed between grid and cathode. Thus the amplification of the stage is higher.

A concentric circuit may be tuned by varying the length of the inner conductor (usually by using close-fitting tubes, one sliding inside the other) or by connecting an ordinary tuning condenser across the line. Tapping the condenser down as shown in Fig. 715-B gives a band-spread effect which is advantageous, and in addition helps to keep the Q of the circuit higher than it would be with the condenser connected directly across the open end of the line, since at ultra-high frequencies most condensers have losses which cannot be neglected.

U.h.f. oscillators such as those used in the superregenerative detector usually will work well at frequencies where r.f. amplification is impossible with standard tubes (as in the 112-Mc. band) since tube losses are compensated for by energy taken from the power supply. Ordinary coil and condenser circuits are practicable with such tubes and circuits at 112 Mc., and although not as good as linear circuits are more convenient to construct.

● 7-8 THE SUPERHETERODYNE

Principles — In the *superheterodyne*, or *superhet*, receiver the frequency of the incoming signal is changed to a new radio frequency, the *intermediate frequency* (*i.f.*), then amplified, and finally detected. The frequency is changed by means of the heterodyne process (§ 7-1), the output of an adjustable *local oscil-*

lator (*h.f. oscillator*) being combined with the incoming signal in a *mixer* or *converter* stage (*first detector*) to produce a beat frequency equal to the i.f. Fig. 716 gives the essentials of the superhet in block form. C.w. signals are made audible by heterodyning the signal at the second detector by an oscillator (the *beat frequency oscillator* (*b.f.o.*) or *beat oscillator*), set to differ from the i.f. by a suitable audio frequency.

As a numerical example, assume that an intermediate frequency of 455 kc. is chosen, and that the incoming signal is on 7000 kc. Then the h.f. oscillator frequency may be set to 7455 kc. in order that the beat frequency (7455 minus 7000) will be 455 kc. The h.f. oscillator also could be set to 6545 kc., which will give the same frequency difference. To produce an audible c.w. signal of say 1000 cycles at the second detector, the beat oscillator would be set either to 454 kc. or 456 kc.

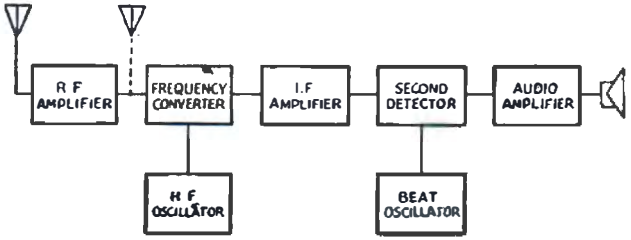


Fig. 716 — The basic superheterodyne arrangement.

Characteristics — The frequency-conversion process permits r.f. amplification at a relatively-low frequency where high selectivity can be obtained, and this selectivity is constant regardless of the signal frequency. Higher gain is also possible at the low frequencies used for intermediate amplification. The separate oscillators can be designed for stability, and since the h.f. oscillator is working at a frequency considerably removed from the signal frequency its stability is practically unaffected by the strength of the incoming signal.

Images — Each h.f. oscillator frequency will cause i.f. response at two signal frequencies, one higher and one lower than the oscillator frequency. If the oscillator is set to 7455 kc. to respond to a 7000-kc. signal, for example, it will also respond to a signal on 7910 kc., which likewise gives a 455-kc. beat. The undesired signal of the two is called the *image*. When the r.f. circuit is tuned to the desired signal frequency, and desired-signal and image voltages of equal magnitude are alternately applied to the circuit, the ratio of desired-signal to image i.f. output is called the *signal-to-image* ratio, or *image ratio*.

The image ratio depends upon the selectivity of the r.f. tuned circuits preceding the mixer tube. Also, the higher the intermediate frequency the higher the image ratio, since raising the i.f. increases the frequency separation between signal and image and thus places the latter farther away from the peak of the resonance curve (§ 2-10) of the signal-frequency circuits.

Other spurious responses — In addition to images, other signals to which the receiver is not ostensibly tuned may be heard. Harmonics of the high-frequency oscillator may beat with signals far removed from the desired frequency to produce output at the intermediate frequency; such spurious responses can be reduced by adequate selectivity before the mixer stage and good shielding to prevent signal pickup by any means other than the antenna. When a strong signal is received, the harmonics (§ 2-7) generated by rectification in the second detector may, by stray coupling, be introduced into the r.f. or mixer circuit and be converted to the intermediate frequency to go through the receiver in the same way as an ordinary signal. These "birdies" appear as a heterodyne beat on the desired signal and are principally bothersome when the incoming signal is not very greatly different from the intermediate frequency. They can be prevented by proper circuit isolation and shielding. Harmonics of the beat oscillator also can be converted and amplified through the receiver in similar fashion; these responses can be reduced by shielding the beat oscillator and operating it at low output level.

The double superhet — At high and ultra-high frequencies it is difficult to secure an adequate image ratio when the intermediate frequency is of the order of 455 kc. To reduce image response the signal frequently is first converted to a rather high intermediate frequency (1500, 5000, or even 10,000 kc.), and then — sometimes after further amplification — reconverted to a lower i.f. where higher adjacent-channel selectivity can be obtained. Such a receiver is called a *double superheterodyne*.

● 7-9 FREQUENCY CONVERTERS

Characteristics — The first detector or mixer resembles an ordinary detector. A cir-

cuit tuned to the intermediate frequency is placed in the plate circuit of the mixer so that the highest possible i.f. voltage will be developed. The signal- and oscillator-frequency voltages appearing in the plate circuit are bypassed to ground since they are not wanted in the output. The i.f. tuned circuit should have low impedance for these frequencies, a condition easily met if they do not approach the intermediate-frequency.

The *conversion efficiency* of the mixer is measured by the ratio of i.f. output voltage from the plate circuit to r.f. signal voltage applied to the grid. High conversion efficiency is obviously desirable. The mixer tube noise also should be low if a good signal-to-noise ratio is wanted, particularly if the mixer is the first tube in the receiver.

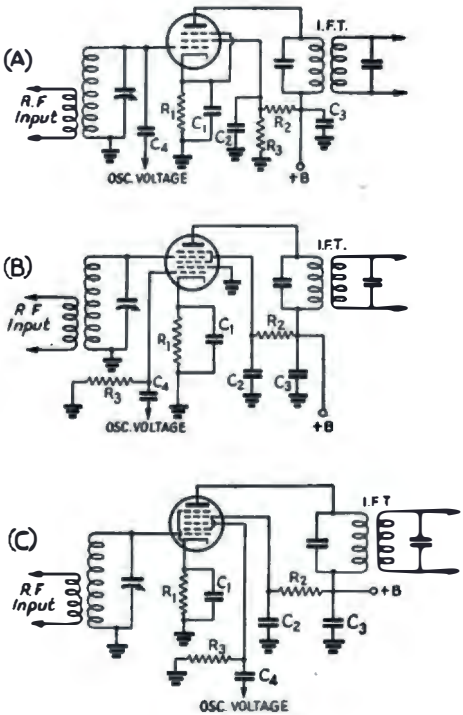


Fig. 717 — Mixer or converter circuits. A, grid injection with a pentode plate detector; B and C, separate injection circuits for converter tubes.

Circuit values are as follows:

	Circuit A	*Circuit B*	*Circuit C*
C_1, C_2, C_3 —	0.01-0.1 μfd.	0.01-0.1 μfd.	0.01-0.1 μfd.
C_4 —	approx. 1 μμfd.	50-100 μμfd.	50-100 μμfd.
R_1 —	10,000 ohms.	300 ohms.	500 ohms.
R_2 —	0.1 megohm.	50,000 ohms.	15,000 ohms.
R_3 —	50,000 ohms.	50,000 ohms.	50,000 ohms.

Plate voltage should be 250 in all three circuits. If an 1851 or 1852 is used in Circuit A, R_1 should be changed to 500 ohms.

The mixer should not require too much r.f. power from the h.f. oscillator, since it may be difficult to supply the power and maintain good oscillator stability (§ 3-7). Also, the conversion efficiency should not depend too critically on the oscillator voltage (that is, a small change in oscillator output should not change the gain appreciably) since it is difficult to maintain constant oscillator output over a wide frequency range.

A change in oscillator frequency caused by tuning of the mixer grid circuit is called *pulling*. If the mixer and oscillator could be completely isolated, mixer tuning would have no effect on the oscillator frequency, but in practice this is a difficult condition to attain. Pulling causes oscillator instability and should be minimized, because the stability of the whole receiver depends critically upon the stability of the h.f. oscillator. The pulling effect decreases with the separation between the signal and h.f. oscillator frequencies, hence is less with high intermediate frequencies and greater with low i.f.'s.

Circuits — Typical frequency-conversion circuits are given in Fig. 717. The variations are chiefly in the way in which the oscillator voltage is introduced. In 717-A, the screen-grid pentode functions as a plate detector; the oscillator is capacity-coupled to the grid of the tube, in parallel with the tuned input circuit. Inductive coupling may be used instead. The conversion gain and input selectivity are generally good so long as the sum of the two voltages (signal and oscillator) impressed on the mixer grid does not exceed the grid bias. It is desirable to make the oscillator voltage as high as possible without exceeding this limitation. The oscillator voltage required is small and the power negligible.

A pentagrid-converter tube is used in the circuit at B. Although intended for combination oscillator-mixer use, this type of tube usually will give more satisfactory performance when used in conjunction with a separate oscillator, the output of which is coupled in as shown. The circuit gives good conversion efficiency, and because of the electron coupling gives desirable isolation between the mixer and oscillator circuits. A small amount of power is required from the oscillator.

Circuit C is for the 6L7 mixer tube. The value of oscillator voltage can vary over a considerable range without affecting the conversion gain. There are no critical adjustments and the oscillator-mixer isolation is good. The oscillator must supply somewhat more power than in B.

A more stable receiver generally results, particularly at the higher frequencies, when separate tubes are used for the mixer and oscillator. The same number of circuit com-

ponents is required whether or not a combination tube is used, so that there is little difference from the cost standpoint.

Tubes for frequency conversion — Any sharp cut-off pentode may be used in the circuit of Fig. 717-A. The 1851 or 1852 give very high conversion gain and an excellent signal-to-noise ratio — comparable, in fact, to the gain and signal-to-noise ratio obtainable with r.f. amplifiers, and in these respects far superior to any other tubes used as mixers. However, this type of tube loads the circuit more (§ 7-6) and thus decreases the selectivity.

The 6K8 is a good tube for the circuit at B; its oscillator plate connection may be ignored. The 6SA7 also is excellent in this circuit, although it has no anode grid (No. 2 grid in the diagram). In addition to these two types, any pentagrid converter tube may be used.

● 7-10 THE HIGH-FREQUENCY OSCILLATOR

Design considerations — Stability of the receiver (§ 7-2) is chiefly dependent upon the stability of the h.f. oscillator, and particular care should be given this part of the receiver. The frequency of oscillation should be insensitive to changes in voltage, loading, and mechanical shock. Thermal effects (slow change in frequency because of tube or circuit heating) should be minimized. These ends can be attained by the use of good insulating materials and good-quality circuit components, by suitable electrical design, and by careful mechanical construction.

In addition, the oscillator must be capable of furnishing sufficient r.f. voltage and power to the particular mixer circuit chosen, at all frequencies within the range of the receiver, and its harmonic output should be as low as possible to reduce spurious response (§ 7-8).

It is desirable to make the L/C ratio in the oscillator tuned circuit as low as possible (high-C) since this results in increased stability (§ 3-7). Particular care should be taken to insure that no part of the oscillator circuit will vibrate mechanically. This calls for short leads and very "solid" mechanical construction. The chassis and panel material should be heavy and rigid enough so that pressure on the tuning dial will not cause torsion and a shift in the frequency. Care in mechanical construction is well repaid by increased frequency stability.

Circuits — Several oscillator circuits are shown in Fig. 718. The point at which output voltage is taken for the mixer is indicated by the "X" or "Y" in each case. A and B will give about the same results, and require only one coil. However, in these two circuits the cathode is above ground potential for r.f., which often is a cause of hum modulation of the oscillator output at 14 Mc. and higher frequencies when 6.3-volt heater tubes are used. Hum is usually not bothersome with 2.5-volt tubes, nor, of course, with tubes which are heated by direct current. The circuit of 718-C overcomes hum with 6.3-volt tubes since the cathode is grounded. The two coils are advantageous in construction since the feedback adjustment (number of turns on L_2) is simple mechanically.

Besides the use of a fairly high C/L ratio in the tuned circuit, it is necessary to adjust the

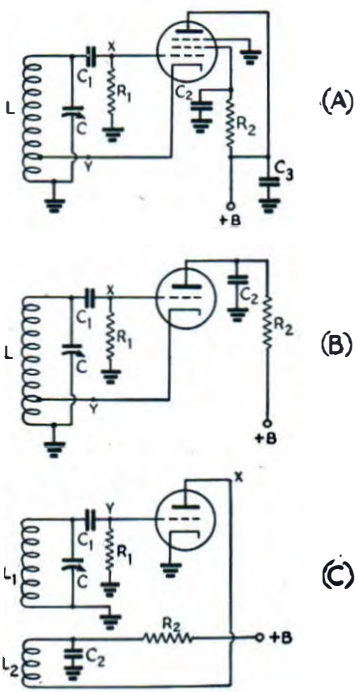


Fig. 718 — High-frequency oscillator circuits. A, screen-grid grounded-plate oscillator; B, triode grounded-plate oscillator; C, triode, tickler circuit. Coupling to mixer may be taken from points X and Y. In A and B, coupling from Y will reduce pulling effects, but gives less voltage than from X; this type of coupling is therefore best adapted to those mixer circuits with small oscillator-voltage requirements.

Typical values are as follows:

	Circuit A	*Circuit B*	*Circuit C*
C1—	100 μμfd.	100 μμfd.	100 μμfd.
C2—	0.1 μfd.	0.1 μfd.	0.1 μfd.
C3—	0.1 μfd.		
R1 —	50,000 ohms.	50,000 ohms.	50,000 ohms.
R2—	50,000 ohms.	10,000 to 25,000 ohms.	10,000 to 25,000 ohms.

The "B" supply voltage should be 250 volts. In circuits B and C, R2 is for the purpose of dropping the supply voltage to 100–150 volts; it may be omitted if this voltage is taken from a voltage divider in the power supply (§ 8–10).

feedback to obtain optimum results. Too much feedback will cause the oscillator to "squeg," or operate at several frequencies simultaneously (§ 7-4); too little feedback will cause the output to be low. In the tapped-coil circuits (A, B) the feedback is increased by moving the tap toward the grid end of the coil; in *C*, by increasing the number of turns on L_2 or by moving L_2 closer to L_1.

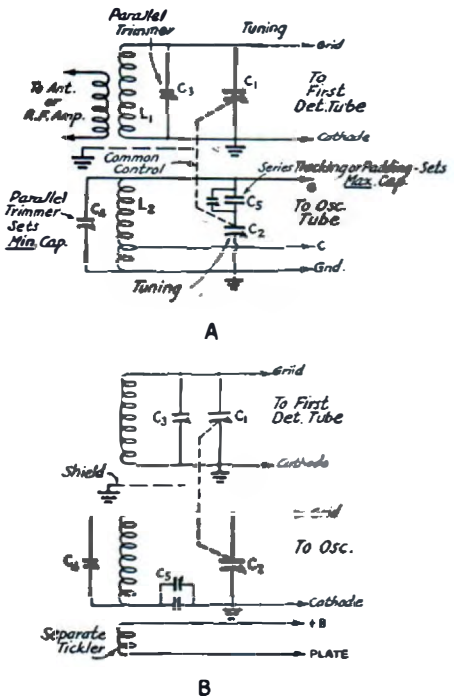


Fig. 719 — Converter circuit tracking methods. Approximate circuit values for 450- to 465-kc. intermediates with tuning ranges of approximately 2.15-to-1, C_1 and C_2 having a maximum of 140 $\mu\mu$fd. and the total minimum capacitance, including C_3 or C_4, being 30 to 35 $\mu\mu$fd.

Tuning Range	L_1	L_2	C_5
1.7–4 Mc.	50 μh.	40 μh.	0.0013 μfd.
3.7–7.5 Mc.	14 μh.	12.2 μh.	0.0022 μfd.
7–15 Mc.	3.5 μh.	3 μh.	0.0045 μfd.
14–30 Mc.	0.8 μh.	0.78 μh.	None used

Approximate values for 450- to 465-kc. i.f. with a 2.5-to-1 tuning range, C_1 and C_2 being 350-$\mu\mu$fd. maximum, minimum capacitance including C_3 and C_4 being 40 to 50 $\mu\mu$fd.

Tuning Range	L_1	L_3	C_5
0.5–1.5 Mc.	240 μh.	130 μh.	425 $\mu\mu$fd.
1.5–4 Mc.	32 μh.	25 μh.	0.00115 μfd.
4–10 Mc.	4.5 μh.	4 μh.	0.0028 μfd.
10–25 Mc.	0.8 μh.	0.75 μh.	None used

To avoid changes in plate voltage which may cause the oscillator frequency to change, it is good practice to regulate the plate supply by means of a gaseous voltage regulator tube (§ 8-8).

Tracking — For ganged tuning there must be a *constant difference* in frequency between the oscillator and mixer circuits. This difference is equal to the intermediate frequency (§ 7-8).

Tracking methods for covering a wide frequency range, suitable for general-coverage receivers, are shown in Fig. 719. The tracking capacity C_5 commonly consists of two condensers in parallel, a fixed one of somewhat less capacity than the value needed and a smaller variable in parallel to allow for adjustment to the exact proper value. In practice, the trimmer C_4 is first set for the high-frequency end of the tuning range and then the tracking condenser is set for the low-frequency end. The tracking capacity becomes larger as the percentage difference between the oscillator and signal frequencies becomes smaller (that is, as the signal frequency becomes higher). Typical circuit values are given in the accompanying table.

In amateur-band receivers tracking is simplified by choosing a band-spread circuit which gives practically straight-line-frequency tuning (equal frequency change for each dial division) and then adjusting the oscillator and mixer tuned circuits so that both cover the same total number of kilocycles. For example, if the i.f. is 455 kc. and the mixer circuit tunes from 7000 to 7300 kc. between two given points on the dial, then the oscillator must tune from 7455 to 7755 kc. between the same two dial readings. With the band-spread arrangement of Fig. 712-C the tuning will be practically straight-line frequency if the capacity actually in use at C_2 is not too small; the same is true of 712-A if C_1 is small compared to C_2.

● 7-11 THE INTERMEDIATE FREQUENCY AMPLIFIER

Choice of frequency — The selection of an intermediate frequency is a compromise between various conflicting factors. The lower the i.f., the higher the selectivity and gain, but a low i.f. brings the image nearer the desired

signal and hence decreases the image ratio (§ 7-8). A low i.f. also increases pulling of the oscillator frequency (§ 7-9). On the other hand, a high i.f. is beneficial to both image ratio and pulling, but the selectivity and gain are lowered. The difference in gain is least important.

An i.f. of the order of 455 kc. gives good selectivity and is satisfactory from the standpoint of image ratio and oscillator pulling at frequencies up to 7 Mc. The image ratio is poor at 14 Mc. when the mixer is connected to the antenna, but adequate when there is a tuned r.f. amplifier between antenna and mixer. At 28 Mc. and the ultra-high frequencies the image ratio is very poor unless several r.f. stages are used. Above 14 Mc. pulling is likely to be bad unless very loose coupling can be used between mixer and oscillator.

With an i.f. of about 1600 kc., satisfactory image ratios can be secured on 14, 28 and 56 Mc., and pulling can be reduced to negligible proportions. However, the i.f. selectivity is considerably lower, so that more tuned circuits must be used to increase the selectivity. For ultra-high frequencies, including 28 Mc., the best solution is to use a double superhet (§ 7-8), choosing one i.f. for image reduction (5 and 10 Mc. are frequently used) and the second for gain and selectivity.

In choosing an i.f. it is wise to avoid frequencies on which there is considerable activity by the various radio services, since such signals may be picked up directly on the i.f. wiring. The frequencies mentioned are fairly free of such interference.

Circuits — I.f. ampifiers usually consist of one or two stages. Two stages at 455 kc. will give all the gain usable, in view of the minimum receiver noise level, and also give suitable selectivity for good-quality 'phone reception (§ 7-2).

A typical circuit arrangement is shown in Fig. 720. A second stage would simply duplicate the circuit of the first. In principle, the i.f. amplifier is the same as the tuned r.f. amplifier (§ 7-6). However, since a fixed frequency is used the primary as well as the secondary of the coupling transformer is tuned, giving higher selectivity than is obtainable with a closely-coupled untuned primary. The cathode resistor, R_1, is connected to a gain control circuit of the type previously described (§ 7-6); usually both stages, if two are used, are controlled by a single variable resistor. The decoupling resistor, R_3 (§ 2-11), helps isolate the amplifier and thus prevent stray feedback. C_2 and R_4 are part of the automatic volume control circuit (§ 7-13); if no a.v.c. is used the lower end of the i.f. transformer secondary is simply connected to ground.

In a two-stage amplifier the screen grids of both stages may be fed from a common supply, either through a resistor (R_2) as shown, the screens being connected in parallel, or from a voltage divider (§ 8-10) across the plate supply. Separate screen dropping resistors are preferable for preventing undesired coupling between stages.

When two stages are used the high gain will tend to cause instability and oscillation, so that good shielding, by-passing and careful circuit arrangement to prevent stray coupling, with exposed r.f. leads well separated, is necessary.

I.F. transformers — The tuned circuits of i.f. amplifiers are built up as transformer units consisting of a shielding container in which the coils and tuning condensers are mounted. Both air-core and powdered-iron-core universal-wound coils are used, the latter having somewhat higher *Q*'s and, hence, greater selectivity and gain per unit.

Variable tuning condensers are of the midget type, air-dielectric condensers being preferable because their capacity is practically unaffected by changes in temperature and humidity. Iron-core transformers may be tuned by varying the inductance (permeability tuning) in which case stability comparable to that of variable air-condenser tuning can be obtained by use of high-stability fixed mica condensers. Such stability is of great importance, since a circuit whose frequency "drifts" with time will eventually be tuned to a different frequency than the other circuits and thereby reduce the gain and selectivity of the amplifier.

Besides the type of i.f. transformer shown in Fig. 720, special units to give desired selectivity characteristics are available. For higher than ordinary adjacent channel selectivity (§ 7-2) *triple-tuned* transformers, with a third tuned circuit inserted between the input and output

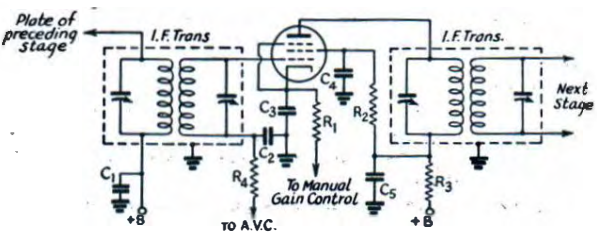


Fig. 720 — Intermediate-frequency amplifier circuit. Typical values are as follows:
C_1 — 0.1 μfd. at 455 kc.; 0.01 μfd. at 1600 kc. and higher.
C_2 — 0.01 μfd.
C_3, C_4, C_5 — 0.1 μfd. at 455 kc.; 0.01 μfd. at 1600 kc. and higher.
R_1 — 300 ohms.
R_2 — 0.1 megohm.
R_3 — 2000 ohms.
R_4 — 0.25 megohm.

windings, are used. The energy is transferred from the input to the output windings via this *tertiary* winding, thus adding its selectivity to the overall selectivity of the transformer. *Variable-selectivity* transformers also can be obtained, these usually being provided with a third (untuned) winding which can be connected to a resistor, thereby loading the tuned circuits and decreasing the *Q* and selectivity (§ 2:-10) to broaden the selectivity curve. The variation in selectivity is brought about by switching the resistor in and out of the circuit. Another method is to vary the coupling between primary and secondary, overcoupling being used to broaden the selectivity curve, undercoupling to sharpen it (§ 2-11).

Selectivity — The overall selectivity of the i.f. amplifier will depend on the frequency and the number of stages. The following figures are indicative of the band-widths to be expected with good-quality transformers, with construction in which regeneration is kept to a minimum:

	Band Width, kc.		
Intermediate Frequency	*3 times down*	*10 times down*	*100 times down*
One stage 455 kc. (air core)...	8.7	17.8	32.3
One stage 455 kc. (iron core)..	4.3	10.3	20.4
Two stage 455 kc. (iron core)..	2.9	6.4	10.8
Two stage 1600 kc.....	11.0	16.6	27.4
Two stage 5000 kc.....	25.8	46.0	100.0

Tubes for I.F. amplifiers — Variable-μ pentodes (§ 3-5) are almost invariably used in i.f. amplifier stages, since grid-bias gain control (§ 7-6) is practically always applied to the i.f. amplifier. Tubes with high plate resistance will have least effect on the selectivity of the amplifier, and those with high mutual conductance will give greatest gain. The choice of i.f. tubes will have practically no effect on the signal-to-noise ratio, since this will have been determined by the preceding mixer and r.f. amplifier (if used).

If single-ended tubes are used, care should be taken to keep the plate and grid leads well separated. With these tubes it is advisable to mount the screen by-pass condenser directly on the bottom of the socket crosswise between the plate and grid pins to provide additional shielding, making sure that the outside foil of the condenser is connected to ground.

Single-signal effect — In heterodyne c.w. reception with a superhet receiver the beat oscillator is set to give a suitable audio-frequency beat note when the incoming signal is converted to the intermediate frequency. For example, the beat oscillator may be set to 456 kc. (the i.f. being 455 kc.) to give a 1000-cycle beat note. Now if an interfering signal appears at 457 kc., it also will be heterodyned by the beat oscillator to produce a 1000-cycle beat. This *audio-frequency image* corresponds to the high-frequency images already discussed (§ 7-8) and can be reduced by providing enough selectivity since the image signal is off the peak of the i.f. resonance curve.

When this is done, tuning through a given signal will show a strong response at the desired beat tone on one side of zero beat only, instead of the two beat notes on either side of zero beat which are characteristic of less selective reception, hence the name "single signal" reception.

The necessary selectivity is difficult to obtain with non-regenerative amplifiers employing ordinary tuned circuits unless a very low intermediate frequency or a large number of circuits is used. In practice it is secured either by regenerative amplification or by the use of a crystal filter.

Regeneration — Regeneration can be used to give a pronounced single-signal effect, particularly when the i.f. is 455 kc. or lower. The resonance curve of an i.f. stage at critical regeneration (just below the oscillating point) is extremely sharp, a band width of 1 kc. at 10 times down and 5 kc. at 100 times down being readily obtainable in one stage. The audio-frequency image of a given signal can thus be reduced by a factor of nearly 100 for a 1000-cycle beat note (image 2000 cycles from resonance).

Regeneration is easily introduced in an i.f. amplifier by providing a small amount of capacity coupling between grid and plate (bringing a short length of wire, connected to the grid, into the vicinity of the plate lead, usually will suffice) and may be controlled by the regular cathode-resistor gain control. When the i.f. is regenerative, it is usually preferable to operate the tube at reduced gain (high bias) and depend upon the regeneration to bring the signal strength back to normal. This prevents overloading on strong signals and thereby increases the effective selectivity.

The higher selectivity with regeneration reduces the response to noise generated in the earlier stages of the receiver, just as in the case of high selectivity produced by other means, and therefore improves the signal-to-noise ratio. The disadvantage is that the regenerative gain varies with the signal strength, being less on strong signals, and the selectivity varies accordingly.

Crystal filters — The most satisfactory method of obtaining high selectivity is by the use of a pieso-electric quartz crystal as a selective filter in the i.f. amplifier (§ 2-10). Compared to a good tuned circuit, the *Q* of such a crystal is extremely high. The dimensions of the crystal are made such that it is resonant at the desired intermediate frequency, and it is then used as a selective coupler between i.f. stages.

Fig. 721 gives a typical crystal-filter reso-

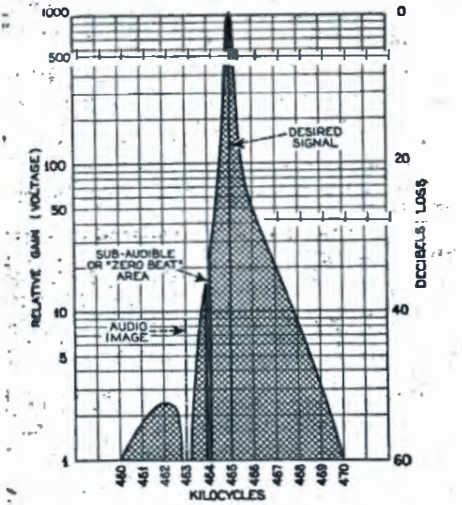


Fig. 721 — Graphical representation of single-signal selectivity. The shaded area indicates the region in which response is obtainable.

nance curve. For single-signal reception the audio-frequency image can be reduced by a factor of 1000 or more. Besides practically eliminating the a.f. image, the high selectivity of the crystal filter provides great discrimination against signals very close to the desired signal in frequency, and, by reducing the band width, reduces the response of the receiver to noise both from sources external to the receiver and in the r.f. stages of the receiver itself.

Crystal-filter circuits; phasing — Several crystal-filter circuits are shown in Fig. 722. Those at A and B are practically identical in performance, although differing in details. The crystal is connected in a bridge circuit (§ 2-11) with the secondary side of T_1, the input transformer, balanced to ground either through a pair of condensers, *C-C*, (A) or by a center-tap on the secondary, L_2 (B). The bridge is completed by the crystal X, and the *phasing condenser*, C_2, which has a maximum capacity somewhat higher than the capacity of the crystal in its holder. When C_2 is set to balance the crystal-holder capacity the resonance curve of the crystal circuit is practically symmetrical; the crystal acts as a series-resonant circuit of very high Q and thus allows signals of the desired frequency to be fed through C_3 to L_3L_4, the output transformer. Without C_2 the holder capacity (with the crystal acting as a dielectric) would by-pass signals of undesired frequencies to the output circuit.

The phasing control has an additional function besides neutralization of the crystal-holder capacity. The holder capacity becomes a part of the crystal circuit and causes it to act as a parallel-tuned resonant circuit at a frequency slightly higher than its series-resonant frequency. Signals at the parallel-resonant frequency are thus prevented from reaching the output circuit. The phasing control, by varying the effect of the holder capacity, permits shifting the parallel-resonant frequency over a considerable range, thus providing adjustable rejection of interfering signals. The effect of rejection is illustrated in Fig. 721, where the audio image is reduced far below the value that would be expected if the resonance curve were symmetrical.

Variable selectivity — In circuits such as A and B, Fig. 722, variable selectivity is obtained by adjustment of the variable input impedance, which is effectively in series with the crystal resonator. This is accomplished by

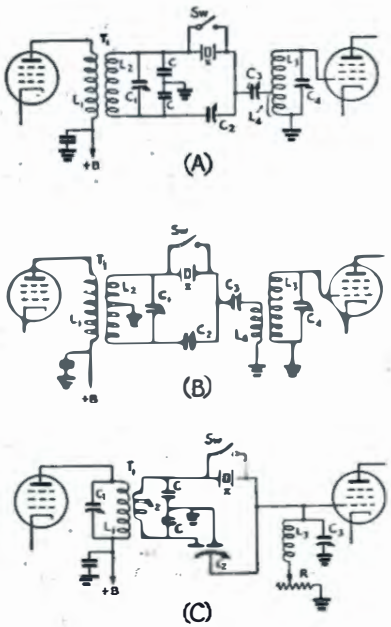


Fig. 722 — Crystal filter circuits of three types. All give variable band-width, with C having the greatest range of selectivity. Their operation is discussed in the text. Suitable circuit values are as follows: Circuit A, T_1, special i.f. input transformer with high-inductance primary, L_1, closely coupled to tuned secondary, L_2; C_1, 50-$\mu\mu$fd. variable; C, each 100-$\mu\mu$fd. fixed (mica); C_2, 10- to 15-$\mu\mu$fd. (max.) variable; C_3, 50-$\mu\mu$fd. trimmer; L_3C_4, i.f. tuned circuit, with L_3 tapped to match crystal-circuit impedance. In Circuit B, T_1 is the same as in Circuit A except that the secondary is center-tapped; C_1 is 100-$\mu\mu$fd. variable; C_3, C_3 and C_4 same as for Circuit A; L_3L_4 is a transformer with primary, L_4, corresponding to tap on L_3 in A. In Circuit C, T_1 is a special i.f. input transformer with tuned primary and low-impedance secondary; C, each 100-$\mu\mu$fd. fixed (mica); C_2, opposed-stator phasing condenser, app. 8 $\mu\mu$fd. maximum capacity each side; L_3C_3, high-Q i.f. tuned circuit; R, 0 to 3000 ohms (selectivity control).

varying C_1 (the *selectivity control*) which tunes the balanced secondary circuit of T_1. When the secondary is tuned to i.f. resonance, the parallel impedance of the L_2C_1 combination is maximum and is purely resistive (§ 2-10). Since the secondary circuit is center-tapped, approximately one-fourth of this resistive impedance is in series with the crystal through C_3 and L_4. This lowers the Q of the crystal circuit and makes its selectivity minimum. At the same time, the voltage applied to the crystal circuit is maximum.

When the input circuit is detuned from the crystal resonant frequency, the resistance component of the input impedance decreases, and so does the total parallel impedance. Accordingly, the selectivity of the crystal circuit becomes higher and the applied voltage falls off. At first the resistance decreases faster than the applied voltage, with the result that at first the c.w. output from the filter *increases* as the selectivity is increased. The output then falls off gradually as the input circuit is detuned farther from resonance and the selectivity becomes still higher.

In the circuits of A and B, Fig. 722, the minimum selectivity is still much greater than that of a normal two-stage 455-kc. amplifier, and it is desirable to provide a wider range of selectivity, particularly for 'phone reception. A circuit which does this is shown at Fig. 722-C. The principle of operation is similar, but a much higher value of resistance can be introduced in the crystal circuit to reduce the selectivity. The output tuned circuit L_3C_3 must have high Q. A compensated condenser is used at C_2 (phasing) to maintain circuit balance, so that the phasing control does not affect the resonant frequency. The output circuit functions as a voltage divider in such a way that the amplitude of the carrier delivered to the next grid does not vary appreciably with the selectivity setting. The variable resistor, R, may consist of a series of separate fixed resistors selected by a tap switch.

• 7-12 THE SECOND DETECTOR AND BEAT OSCILLATOR

Detector circuits — The second detector of a superhet receiver performs the same function as the detector in the simple receiver, but usually operates at a higher input level because of the relatively great r.f. amplification. Therefore the ability to handle large signals without distortion is preferable to high sensitivity. Plate detection is used to some extent, but the diode detector is most popular. It is especially adapted to furnishing automatic gain or volume control (§ 7-13), which gives it an additional advantage. The basic circuits are as described in § 7-3, although in many cases the diode elements are incorporated in a multipurpose tube which also has an amplifier section.

The beat oscillator — Any standard oscillator circuit (§ 3-7) may be used for the beat oscillator. Special beat-oscillator transformers are available, usually consisting of a tapped coil with adjustable tuning; these are most conveniently used with circuits such as those shown at Fig. 718-A and -B, with the output taken from "Y." A variable condenser of about 25 $\mu\mu$fd. capacity often is connected between cathode and ground to provide fine adjustment of the beat frequency. The beat oscillator usually is coupled to the second detector tuned circuit through a fixed condenser of a few $\mu\mu$fd. capacity.

The beat oscillator should be well shielded to prevent coupling to any part of the circuit except the second detector, and to prevent its harmonics from getting into the front end of the receiver and being amplified like regular signals. To this end, the plate voltage should be as low as is consistent with sufficient audio output. If the beat-oscillator output is too low, strong signals will not give a proportionately strong audio response.

A regenerative second detector may be used to give the audio beat note, but since the detector must be detuned from the i.f. the selectivity and signal strength are reduced, while blocking (§ 7-4) is pronounced, because of the high signal level at the second detector.

• 7-13 AUTOMATIC VOLUME CONTROL

Principles — Automatic regulation of the gain of the receiver in inverse proportion to the signal strength is a great advantage, especially in 'phone reception, since it tends to keep the output level of the receiver constant regardless of input signal strength. It is readily accomplished in the superheterodyne by using the average rectified d.c. voltage developed by the received signal across a resistance in a detector circuit (§ 7-3) to vary the bias on the r.f. and i.f. amplifier tubes. Since this voltage is proportional to the average amplitude of the signal, the gain is reduced as the signal strength is greater. The control will be more complete as the number of stages to which the a.v.c. bias is applied is greater. Control of at least two stages is advisable.

Circuits — A typical circuit of a diode-triode type tube used as a combined a.v.c. rectifier, detector and first audio amplifier is shown in Fig. 723. One plate of the diode section of the tube is used for signal detection and the other for a.v.c. rectification. The a.v.c. diode plate is fed from the detector diode through the small coupling condenser, C_3. Negative bias resulting from the flow of rectified carrier current is developed across R_4, the diode

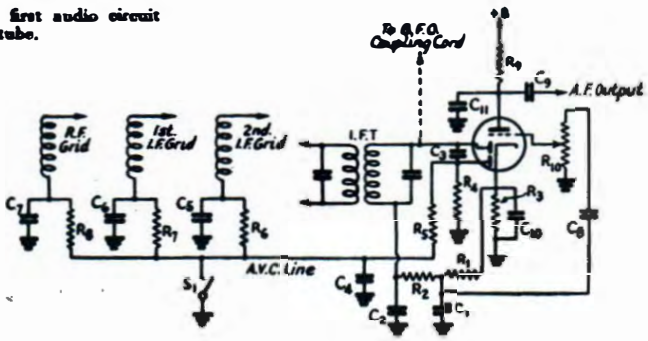


Fig. 723 — Second-detector and first audio circuit with a.v.c., using duo-diode-triode tube.
R_1 — 0.25 megohm.
R_2 — 50,000 to 250,000 ohms.
R_3 — 2000 ohms.
R_4 — 2 to 5 megohms.
R_5 — 0.5 to 1 megohm.
R_6, R_7, R_8 — 0.25 megohm.
R_9 — 0.25 megohm.
R_{10} — 0.5-megohm volume control.
C_1, C_2, C_3 — 100 $\mu\mu$fd.
C_4 — 0.1 μfd.
C_5, C_6, C_7 — 0.01 μfd.
C_8, C_9 — 0.01 to 0.1 μfd.
C_{10} — 5. to 10-μfd. electrolytic.
C_{11} — 250 $\mu\mu$fd.

load resistor. This negative bias is applied to the grids of the controlled stages through the filtering resistors (§ 2-11) R_5, R_6, R_7 and R_8.

It does not matter which of the two diode plates is selected for audio and which for a.v.c. Frequently the two plates are connected together and used as a combined detector-a.v.c. rectifier. This could be done in Fig. 723. The a.v.c. filter and line would connect to the junction of R_2 and C_3, while C_3 and R_4 would be omitted from the circuit.

When S_1 is closed the a.v.c. line is grounded, thereby removing the a.v.c bias from the amplifier stages.

Delayed a.v.c. — In Fig. 723 the audio diode return is made directly to the cathode and the a.v.c. diode return to ground. This places negative bias on the a.v.c. diode equal to the d.c. drop through the cathode resistor (a volt or two) and thus delays the application of a.v.c. voltage to the amplifier grids, since no rectification takes place in the a.v.c. diode circuit until the carrier amplitude is large enough to overcome the bias. Without this delay, the a.v.c. would start working even with a very small signal. This is undesirable because the full amplification of the receiver then cannot be realized on weak signals. In the audio diode circuit this fixed bias would cause distortion and must be avoided, hence the return is made directly to the cathode.

Time constant — The time constant (§ 2-6) of the resistor-condenser combinations in the a.v.c. circuit is an important part of the system. It must be high enough so that the modulation on the signal is completely filtered from the d.c. output, leaving only an average d.c. component which follows the relatively slow carrier variations with fading; audio-frequency variations in the a.v.c. voltage applied to the amplifier grids would reduce the percentage of modulation on the incoming signal and in the practical case would cause frequency distortion. On the other hand, the time constant should not be too great since the a.v.c. then would be unable to follow rapid fading. The values indicated are satisfactory for high-frequency reception.

Signal strength and tuning indicators — A useful accessory to the receiver is an indicator which will show relative signal strength. Not only is it an aid in giving reports, but it also is helpful in aligning the receiver circuits, in conjunction with a test oscillator or other steady signal.

Three types of indicators are shown in Fig. 724. That at A uses an electron-ray tube, several types of which are available. The grid of the triode section is usually connected to the a.v.c. line. The particular type of tube to use will depend upon the voltage available for its grid; where the a.v.c. voltage is relatively large, a remote-cutoff type tube such as the 6G5 or 6N5 should be used in preference to the sharp-cutoff type (6E5).

In B, a milliammeter is connected in series with the d.c. plate lead to one or more r.f. and i.f. tubes whose grids are controlled by a.v.c. Since the plate current of such tubes varies with the strength of the incoming signal, the meter will indicate relative signal intensity and may be calibrated in "S" points. The scale range of the meter should be chosen to fit the number of tubes in use; the maximum plate current of the average remote-cutoff r.f. pentode is from 7 to 10 milliamperes. The shunt resistor R enables setting the plate current to the full-scale value ("zero adjustment"). With this system the ordinary meter reads downwards from full scale with increasing signal strength, which is the reverse of normal pointer movement (clockwise with increasing reading). Special instruments with the zero-current position of the pointer at the right-hand side of the scale are used in commercial receivers.

The system at C uses a 0-1 milliammeter in a bridge circuit arranged so that the meter reading and signal strength increase together. The current through the branch containing R_1 should be approximately equal to the current

through that containing R_2. In some manufactured receivers this is brought about by draining the screen voltage-divider current and the current to the screens of three r.f. pentodes (r.f.

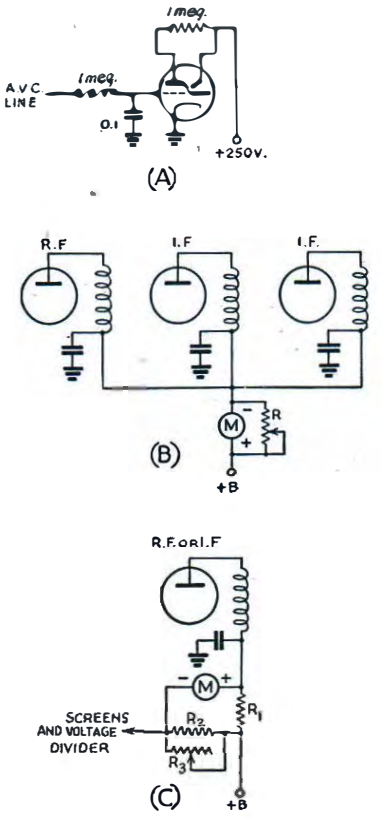


Fig. 724 — Tuning indicator or "S"-meter circuits for superhet receivers. A, electron-ray indicator; B, plate-current meter for tubes on a.v.c.; C, bridge circuit for a.v.c. controlled tube. In B, resistor R should have a maximum resistance several times that of the milliammeter. In C, representative values are: R_1, 250 ohms; R_2, 350 ohms; R_3, 1000-ohm variable.

and i.f. stages) through R_2, the sum of these currents being about equal to the maximum plate current of one a.v.c. controlled tube. Typical values for this type of circuit are given. The sensitivity can be increased by making R_1, R_2 and R_3 larger. The initial setting is made with the manual gain control set near maximum, when R_3 should be adjusted to make the meter reading zero with no signal.

• 7-14 PRESELECTION

Purpose — Preselection is added signal-frequency selectivity before the mixer stage is reached. An r.f. amplifier preceding the mixer is generally called a *preselector*, its purpose, in part at least, being to discriminate in favor of the signal against the image. The preselector may consist of one or more r.f. amplifier stages. When its tuning is ganged with that of the mixer and oscillator, its circuits must track with the mixer circuit.

The circuit is the same as discussed earlier (§ 7-6). An external preselector stage may be used with receivers having inadequate image ratios, in which case it is built as a separate unit, often with a tuned output circuit which gives a further improvement in selectivity. The output circuit usually is link-coupled (§ 2-11) to the receiver.

Signal/noise ratio — An r.f. amplifier will have a better signal-to-noise ratio (§7-2) than a mixer because the gain is higher and because the mixer electrode arrangement results in higher internal tube noise than does the ordinary pentode structure. Hence a preselector is advantageous in increasing the signal-to-noise ratio over that obtainable when the mixer is fed directly from the antenna.

Image suppression — The image ratios (§7-8) obtainable at frequencies up to and including 7 Mc. with a single preselector stage are high enough, when the intermediate frequency is 455 kc., so that for all practical purposes there is no appreciable image response. Average image ratios on 14 Mc. and 28 Mc. are 50–75 and 10–15, respectively. This is the overall selectivity of the r.f. and mixer tuned circuits. A second preselector stage, adding one more tuned circuit, will increase the ratios to several hundred at 14 Mc. and to 30 or 40 at 28 Mc.

On ultra-high frequencies it is impracticable to attempt to secure a good image ratio with a 455-kc. i.f. Good performance in this respect can be secured only by using a high-frequency i.f. or by using a double superhet (§ 7-8) with a high-frequency first i.f.

Regeneration — Regeneration may be used in a preselector stage to increase both gain and selectivity. Since this makes tuning more critical and increases ganging problems, regeneration is seldom used except at 14 Mc. and above where adequate image suppression is difficult to obtain with non-regenerative circuits. The same disadvantages exist as in the case of a regenerative i.f. amplifier (§ 7-11). The effect of regeneration is roughly equivalent to the addition of another non-regenerative preselector stage.

The regeneration may be introduced by the same method used in regenerative i.f. amplifiers (§ 7-11). The manual gain control of the stage will serve as a volume control.

Regeneration does not improve the signal-to-noise ratio, since the tube noise is fed back to the grid circuit along with the signal to add

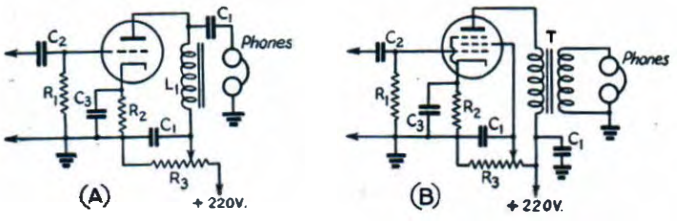


Fig. 725 — Audio output limiting circuits.

C_1 — 0.25 μfd.
C_2 — 0.01 μfd.
C_3 — 5 μfd.
R_1 — 0.5 megohm.
R_2 — 2000 ohms.
R_3 — 50,000-ohm potentiometer.
T — Output transformer.
L_1 — 15-henry choke

to the thermal agitation noise originally present. The latter noise also is amplified.

• 7-15 NOISE REDUCTION

Types of noise — In addition to tube and circuit noise (§ 7-6) much of the noise interference experienced in reception of amateur signals is caused by domestic electrical equipment and automobile ignition systems. The interference is of two types in its effects. The first is of the "hiss" type consisting of overlapping pulses, similar in nature to the receiver noise. It is largely reduced by high selectivity in the receiver, especially for code reception. The second is the "pistol shot" or "machine gun" type, consisting of separated impulses of high amplitude. The "hiss" type of interference is usually caused by commutator sparking in d.c. and series a.c. motors, while the "shot" type results from separated spark discharges (a.c. power leaks, switch and key clicks, ignition sparks, and the like).

Impulse noise — Impulse noise, because of the extremely short duration of the pulses as compared to the time between them, must have high pulse amplitude to give much average energy. Hence noise of this type strong enough to cause much interference generally has an instantaneous amplitude much higher than that of the signal being received. The general principle of devices intended to reduce such noise is that of allowing the signal amplitude to pass through the receiver unaffected, but making the receiver inoperative for amplitudes greater than that of the signal. The greater the amplitude of the pulse compared to its time of duration the more successful the noise-reducing device, since more of the energy in the pulse can be suppressed.

In passing through selective receiver circuits the time duration of the impulses is increased because of the Q or flywheel effect (§ 2-10) of the circuits. Hence the greater the selectivity ahead of the noise-reducing device the more difficult it becomes to secure good noise suppression.

Audio limiting — A considerable degree of noise reduction in code reception can be accomplished by amplitude limiting arrangements applied to the audio output circuit of a receiver. Such limiters also maintain the signal output nearly constant with fading. Diagrams of typical output limiter circuits are shown in Fig. 725. Circuit A employs a triode tube operated at reduced plate voltage (approximately 10 volts) so that it saturates at a low signal level. The arrangement of B has better limiting characteristics. A pentode audio tube is operated at reduced screen voltage (35 volts or so) so that the output power remains practically constant over a grid excitation voltage range of more than 100 to 1. These output limiter systems are simple and adaptable to nearly all receivers. However, they cannot prevent noise peaks from overloading previous circuits.

Second detector circuits — The circuit of Fig. 726 "chops" noise peaks at the second detector of a superhet receiver by means of a biased diode which becomes non-conducting above a predetermined signal level. The audio output of the detector must pass through the diode to the grid of the amplifier tube. The diode would normally be non-conducting with the connections shown were it not for the fact that it is given positive bias from a 30-volt source through the adjustable potentiometer R_3. Resistors R_1 and R_2 must be fairly large in value to prevent loss of audio signal.

The audio signal from the detector can be considered to modulate (§ 5-1) the steady diode current, and conduction will take place so long as the diode plate is positive with respect to the cathode. When the signal is sufficiently large to swing the cathode positive with respect to the plate, however, conduction ceases and that portion of the signal is cut off from the audio amplifier. The point at which cut-off occurs can be selected by adjustment of R_3. By setting R_3 so that the signal just passes through the "valve," noise pulses higher in amplitude than the signal will be cut off. The circuit of Fig. 726-A, using an infinite-impedance detector (§ 7-3) gives a positive voltage on rectification. When the rectified voltage is negative, as from the usual diode detector (§ 7-3) a different circuit arrangement, shown in Fig. 726-B, is required.

An audio signal of about ten volts is required for good limiting action. When a beat oscillator is used for c.w. reception the b.f.o. voltage should be small so that incoming noise will not

have a strong carrier to beat against and produce large audio output.

A second-detector noise limiting circuit which automatically adjusts itself to the received carrier level is shown in Fig. 727. The diode load circuit (§ 7-3) consists of R_6, R_7, R_8 (shunted by the high-resistance audio volume control, R_4) and R_5 in series. The cathode of the 6N7 noise-limiter is tapped on the load resistor at a point such that the average rectified carrier voltage (negative) at its grid is approximately twice the negative voltage at the cathode, both measured with reference to ground. A filter network, R_1C_1, is inserted in the grid circuit so that the audio modulation on the carrier does not reach the grid, hence the grid potential is maintained at substantially the rectified carrier voltage alone. The cathode, however, is free to follow the modulation, and when the modulation is 100% the peak cathode voltage will just equal the steady grid voltage.

At all modulation percentages below 100% the grid is negative with respect to cathode and current cannot flow in the 6N7 plate-cathode circuit. A noise pulse exceeding the peak voltage which represents 100% modulation will, however, make the grid positive with respect to cathode and the relatively-low plate-cathode resistance of the 6N7 shunts the high-resistance audio output circuit, effectively short-circuiting it so that there is practically no response for the duration of the noise peak over the 100% modulation limit.

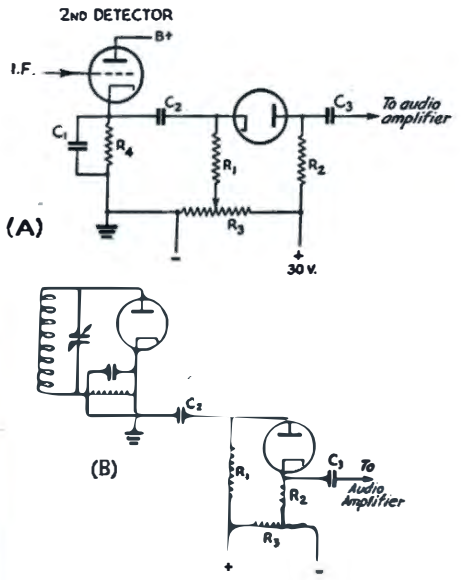


Fig. 726 — The series-valve noise-limiter circuit. A, with an infinite-impedance detector; B, with diode detector. Values are as follows:
R_1 — 0.25 megohm.
R_2 — 50,000 ohms.
R_3 — 10,000-ohm potentiometer.
R_4 — 20,000 to 50,000 ohms.
C_1 — 250 $\mu\mu$fd.
C_2, C_3 — 0.1 μfd.
Diode circuit constants in B are conventional.

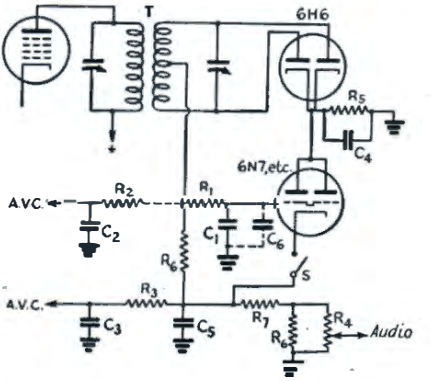


Fig. 727 — Automatic noise limiting circuit for superhet receivers.
T — I.f. transformer with balanced secondary for working into diode rectifier.
R_1, R_2, R_3 — 1 megohm.
R_4 — 1-megohm volume control.
R_5 — 250,000 ohms.
R_6, R_8 — 100,000 ohms.
R_7 — 25,000 ohms.
C_1 — 0.1-μfd. paper.
C_2, C_3 — 0.05-μfd. paper.
C_4, C_5 — 50-$\mu\mu$fd. mica.
C_6 — 0.001-μfd. mica (for r.f. filtering, if needed).
Sw — S.p.s.t. toggle (on-off switch).
The switch should be mounted close to the circuit elements and controlled by an extension shaft if necessary.

R_5 is used to make the noise-limiting tube more sensitive, by applying to the plate an audio voltage out of phase with the cathode voltage so that at the instant the grid goes positive with respect to cathode, the highest positive potential also is applied to the plate, thus further lowering the effective plate-cathode resistance.

I.F. noise silencer — In the circuit shown in Fig. 728 noise pulses are made to decrease the gain of an i.f. stage momentarily and thus silence the receiver for the duration of the pulse. Noise voltage in excess of the desired signal's maximum i.f. voltage is taken off at the grid of the i.f. amplifier, amplified by the noise amplifier stage and rectified by the full-wave diode noise rectifier. The noise circuits are tuned to the i.f. The rectified noise voltage is applied as a pulse of negative bias to the No. 3 grid of the 6L7 used as an i.f. amplifier, wholly or partially disabling this stage for the duration of the individual noise pulse, depending on the amplitude of the noise voltage. The noise amplifier-rectifier circuit is biased so that rectification will not start until noise voltage

exceeds the desired-signal amplitude, by means of the "threshold control," R_2. For reception with automatic volume control, the a.v.c. voltage can be applied to the grid of the noise amplifier to augment this threshold bias. This system of noise silencing gives signal-noise ratio improvement of the order of 30 db. (power ratio of 1000) with heavy ignition interference, raising the signal-noise ratio from −10 db. without the silencer to +20 db. in a typical instance.

Circuit values are normal for i.f. amplifiers (§ 7-11) except as indicated. The noise rectifier transformer T_1 has an untuned secondary closely-coupled to the primary, center-tapped for full-wave rectification. The center-tap rectifier (§ 8-3) is used to reduce the possibility of r.f. feedback into the i.f. amplifier (noise silencer) stage. The time constant (§ 2-6) of the noise rectifier load circuit, $R_1C_1C_2$, must be small to prevent disabling the noise silencer stage for a longer period than the duration of the noise pulse. The radio-frequency choke, *RFC*, must be effective at the intermediate frequency.

Adequate shielding and isolation of the noise amplifier and rectifier circuits from the noise silencer stage must be provided to prevent possible self-oscillation and instability. This circuit is preferably applied to the first i.f. stage of the receiver before the high-selectivity circuits are reached, and is most effective when the signal and noise levels are fairly high (one or two r.f. stages before the mixer) since several volts must be obtained from the noise rectifier for good silencing.

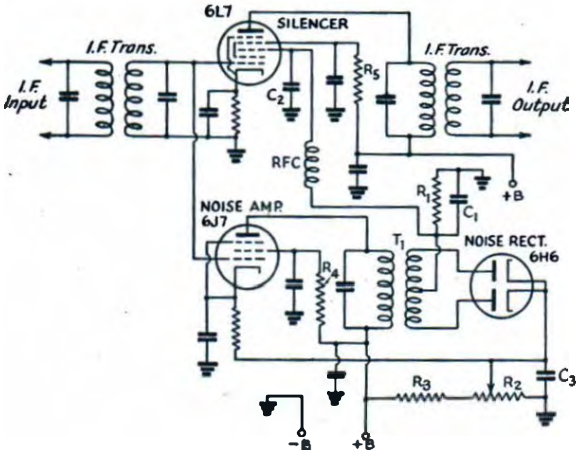


Fig. 728 — I.f. noise-silencing circuit. The "B" supply should be 250 volts.
C_1 — 50–250 $\mu\mu$fd.; use smallest value possible without r.f. feedback.
C_2 — 50 $\mu\mu$fd.
C_3 — 0.1 μfd.
R_1 — 0.1 megohm.
R_2 — 5000-ohm volume control.
R_3 — 20,000 ohms.
R_4, R_5 — 0.1 megohm.
T_1 — Special i.f. transformer for noise rectifier.

• 7-16 OPERATING SUPERHET RECEIVERS

C.w. reception — Proper adjustment of the beat oscillator is to a frequency slightly different from the intermediate frequency (§ 7-8). This adjustment may be made by tuning in a moderately-weak steady carrier, with the beat oscillator turned off, for maximum signal strength as determined by maximum hiss, then turning on the beat oscillator and adjusting its frequency (leaving the receiver tuning alone) to give a suitable beat note. Subsequently the beat oscillator need not be touched except for occasional checking to make certain the frequency has not drifted from the initial setting. The b.f.o. may be set on either the high- or low-frequency side of zero beat.

The use of a.v.c. (§ 7-13) is not generally satisfactory in c.w. reception because the receiver gain rises in the spaces between dots and dashes, giving an increase in noise in the same intervals, and also because the rectified beat oscillator voltage in the second detector circuit also works the a.v.c. circuit. This gives a constant reduction in gain and prevents utilization of the full gain of the receiver. Hence the gain is preferably manually adjusted to give suitable audio-frequency output.

To avoid overloading in the i.f. circuits it is usually best to control the i.f. and r.f. gain and keep the audio gain at a fixed value, rather than to use the a.f. gain control as a volume control and permit the r.f. gain to stay fixed at its highest level.

Tuning with the crystal filter — If the receiver is equipped with a crystal filter the tuning instructions in the preceding paragraph still apply, but more care must be used both in initial adjustment of the beat oscillator and in tuning. The beat oscillator is set as described above, but with the crystal filter in operation and adjusted to its sharpest position, if variable selectivity is available. This initial adjustment should be made with the phasing control

(§ 7-11) in the intermediate position, and after it is completed the beat oscillator should be left set and the receiver tuned to the other side of zero heat (audio-frequency image) on the same carrier to give a beat note of the same tone. This beat will be considerably weaker than the first, and may be "phased out" almost completely by careful adjustment of the phasing control. This is the adjustment for normal operation, and it will be found that one side of zero beat has practically disappeared, leaving the receiver with maximum response on the desired side.

An interfering signal having a beat note differing from that of the a.f. image can similarly be phased out, provided its carrier frequency is not too near the desired carrier.

Depending upon the filter design, maximum selectivity may cause the dots and dashes to lengthen out so that they seem to "run together." This, plus the fact that the tuning is quite critical with extremely high selectivity, may make it desirable to use somewhat less selectivity in regular operating. However, it must be emphasized that to realize the benefits of the crystal filter in reducing interference it is necessary to do *all* tuning with it in the circuit. The selectivity is so high that it is almost impossible to find the desired station quickly should the filter be switched in only when interference is present.

'Phone reception — In reception of 'phone signals the normal procedure is to set the r.f. and i.f. gain at maximum, switch the a.v.c. on, and use the audio gain control for setting the volume. This insures maximum effectiveness of the a.v.c. system in compensating for fading or maintaining constant audio output when either strong or weak signals are tuned in. On occasions a strong signal close to the frequency of a weaker desired station may take control of the a.v.c., in which case the weaker station will practically disappear because of the reduced gain. In such a situation better reception may result if the a.v.c. is switched off, using the manual r.f. gain control to set the gain at a point which prevents "blocking" by the stronger signal.

A crystal filter will do much toward reducing interference in 'phone bands. Although the high selectivity cuts sidebands and thereby reduces the audio output, especially at the higher audio frequencies, it is possible to use quite high selectivity without destroying intelligibility even though the "quality" of the transmission suffers. As in the case of c.w. reception, it is advisable to do all tuning with the filter in circuit when interference is likely to occur. Variable-selectivity filters permit a choice of selectivity which give varying degrees of sideband cutting to suit conditions.

An undesired carrier close in frequency to a desired carrier will heterodyne with it to produce a beat note equal to the frequency difference. Such a heterodyne can be reduced by adjustment of the phasing control when the crystal filter is used. It cannot be prevented in the "straight" superhet having no crystal filter.

A tone control often will be of help in reducing the effects of high-pitched heterodynes, sideband splatter (§ 5-2) and noise, by cutting off the higher audio frequencies. This, like sideband cutting with high selectivity, causes some reduction in naturalness.

Spurious responses — Spurious responses can be recognized without a great deal of difficulty. It is often possible to identify an image by the type of station transmitting, knowing the frequency assignments applying to the frequency to which the receiver is tuned. However, an image also can be recognized by its behavior with tuning. If the signal causes a heterodyne beat note with the desired signal and is actually on the same frequency, the beat note will not change as the receiver is tuned through the signal, but if the interfering signal is an image the beat will vary in pitch as the receiver is tuned. The beat oscillator in the receiver must be off for this test. Using a crystal filter with the beat oscillator on, the image will peak on the opposite side of zero beat to that on which the desired signal peaks.

Harmonic response can be recognized by the "tuning rate," or movement of the tuning dial required to give a specified change in beat note. Signals getting into the i.f. via high-frequency oscillator harmonics will tune more rapidly (less dial movement) through a given change in beat note than signals received by normal means.

Harmonics of the beat oscillator can be recognized by the tuning rate of the beat oscillator pitch control. A smaller movement of the control will suffice for a given change in beat note than is necessary with legitimate signals.

• 7-17 SERVICING SUPERHET RECEIVERS

I.f. alignment — A calibrated signal generator or test oscillator is a practical necessity for initial alignment of an i.f. amplifier. Some means for measuring the output of the receiver also is needed. If the receiver has a tuning meter, its indications will serve for this purpose. Alternatively, if the signal generator is of the modulated type an a.c. output meter (high-resistance voltmeter with copper-oxide rectifier) can be connected across the primary of the output transformer feeding the loudspeaker, or from the plate of the last audio amplifier through a 0.1-μfd. blocking condenser (§ 2-13) to the receiver chassis. The intensity of sound from the loud-speaker can also be judged by ear (with the modulated test oscilla-

tor) if no output meter is available, although this method is not as accurate as those using instruments.

The procedure is as follows: The test oscillator is adjusted to the desired intermediate frequency and the "hot" or ungrounded output lead is clipped on the grid lead of the last i.f. amplifier tube. The grounded lead is connected to the receiver chassis. The trimmer condensers of the transformer feeding the second detector are then adjusted for maximum signal output. The hot lead from the generator is next clipped on the grid of the next to the last i.f. tube and the second from last i.f. transformer brought into alignment by adjusting its trimmers for maximum output. This process is continued, working back from the second detector, until all of the i.f. transformers have been aligned. It will be necessary to reduce the output of the signal generator as more of the i.f. amplifier is brought into use because the increased gain is likely to cause overloading and consequent inaccurate readings. It is desirable in all cases to use the minimum signal strength which gives useful output readings. The i.f. transformer in the plate circuit of the mixer is aligned with the signal-generator lead connected to the mixer grid. Since the tuned circuit feeding the mixer grid may, because it is tuned to a considerably higher frequency, effectively short-circuit the signal-generator output, it may be necessary to disconnect this circuit. With tubes having a top grid connection this can be done by removing the grid cap.

If the tuning indicator is used as an output meter the a.v.c. should be switched on; if the audio output method is used the a.v.c. should be off. The beat oscillator should be off in either case.

If the i.f. amplifier has a crystal filter, the filter should first be switched out and the alignment carried out as above, setting the signal generator as closely as possible to the frequency of the crystal. When completed, the crystal should be switched in and the oscillator frequency varied back and forth over a small range either side of the crystal frequency to find the exact frequency, which will be indicated by a sharp rise in output. Leaving the generator set on the crystal peak, the i.f. trimmers may be realigned for maximum output. The necessary readjustment should be small. The oscillator frequency should be checked frequently to make sure it has not drifted from the crystal peak.

A modulated signal is not of much value for aligning a crystal-filter i.f. amplifier, since the high selectivity cuts sidebands and the results may be inaccurate if the audio output of the receiver is used as a criterion of alignment. Lacking the a.v.c. tuning meter, the transformers may be aligned by ear, using a weak unmodulated signal adjusted to the crystal peak. Switch on the beat oscillator, adjust to a suitable tone, and align the transformers for maximum audio output.

An amplifier which is only slightly out of alignment as a result of normal drift from temperature, humidity or aging effects can be realigned by using any steady signal, such as a local broadcasting signal, in lieu of the test oscillator. Allow the receiver to warm up thoroughly (an hour or so), tune in the signal as usual and "touch up" the i.f. trimmers for maximum output.

R.f. alignment — The object of alignment of the r.f. circuits in a gang-tuned receiver is to secure adequate tracking over each tuning range. The adjustment may be carried out with a test oscillator of suitable frequency range or even on noise or such signal as may be heard. Set the tuning dial at the high-frequency end of the range in use and adjust the h.f. oscillator trimmer condenser for maximum hiss. Next adjust the mixer trimmer condenser for maximum hiss or signal, then the r.f. trimmers. Reset the tuning dial to the low-frequency end of the range and repeat; if the circuits are properly designed no change in trimmer settings should be necessary. Should it be necessary to increase thetrimmer capacity in any circuit, more inductance is needed; if less capacity resonates the circuit, less inductance is required. In the oscillator circuit, the proper frequency range may be secured by adjustment of the tracking condenser capacity (§ 7-10) as well as by inductance adjustment.

Tracking is seldom perfect throughout a tuning range, so that a check of alignment at intermediate points in the range may show it to be slightly off. Normally the gain variation from this cause will be small, however, and it will suffice to bring the circuits into line at both ends of the range. If most reception is in a particular part of the range, such as an amateur band, the circuits may be aligned at that frequency to insure maximum performance, even though the ends of the whole frequency range may be slightly out of alignment.

Oscillation of r.f. or i.f. amplifiers — Oscillation in high-frequency amplifier and mixer circuits is evidenced by squeals or "birdies" as the tuning is varied. It can be caused by poor connections in the common ground circuits, especially to the tuning condenser rotors. Inadequate or defective by-pass condensers in cathode, plate and screen-grid circuits also can cause such oscillation. In some cases it may be advisable to provide a shield between the stators of pre-r.f. amplifier and first-detector ganged tuning condensers, in addition to the usual tube and inter-stage shielding. A metal tube with an ungrounded shell will cause

this trouble. Improper screen-grid voltage, which might result from a shorted or too-low screen-grid series resistor, also could be responsible.

Oscillation in the i.f. circuits is independent of high-frequency tuning and is indicated by a continuous squeal which appears when the gain is advanced with the c.w. beat oscillator on. It can result from similar defects in i.f. amplifier circuits. Inadequate cathode resistor by-pass capacitance is a common cause of such oscillation. Additional by-pass capacitance, of 0.1 to 0.25 μfd. usually will remedy it. Similar treatment can be applied to screen-grid and plate by-passes of i.f. tubes.

Instability — "Birdies" or a mushy hiss occurring with tuning of the high-frequency oscillator may indicate that the oscillator is "squegging" or oscillating simultaneously at high and low frequencies (§ 7-4). This may be caused by a defective tube, too-high oscillator plate or screen-grid voltage, excessive feed back in the oscillator circuit or too-high grid-leak resistance.

A varying beat note in c.w. reception indicates instability in either the h.f. oscillator or beat oscillator, usually the former. The stability of the beat oscillator can be checked by introducing a signal of intermediate frequency (from a test oscillator) into the i.f. amplifier; if the beat note is unstable, the trouble is in the beat oscillator. Poor connections or defective parts are the likely cause. Instability in the high-frequency oscillator may be the result of poor circuit design (§ 7-10), loose connections, defective tubes or circuit components, or poor voltage regulation in the oscillator plate and/or screen supply circuits. Mixer pulling of the oscillator circuit (§ 7-9) also will cause the beat-note to chirp on strong c.w. signals because the oscillator load changes slightly under these conditions.

In 'phone reception with a.v.c., a peculiar type of instability ("motorboating") may appear if the h.f. oscillator frequency is sensitive to changes in plate voltage. As the a.v.c. voltage rises the electrode currents of the controlled tubes decrease, decreasing the load on the power supply and causing the plate voltage on the oscillator to rise. The oscillator frequency changes correspondingly, detuning the circuit and reducing the a.v.c. voltage, thus tending to restore the original conditions. The process then repeats itself at a rate determined by the signal strength and the time constant of the power supply circuits. It is more pronounced with high selectivity, as when a crystal filter is used, and can be cured by designing the oscillator circuit to be relatively insensitive to plate voltage changes and by regulating the voltage applied to the oscillator (§ 7-10).

• 7-18 RECEPTION OF FREQUENCY-MODULATED SIGNALS

F.m. receivers — A frequency-modulation receiver differs in circuit design from one designed for amplitude modulation chiefly in the arrangement used for detecting the signal. Detectors for amplitude-modulated signals do not respond to frequency modulation. It is also necessary, for full realization of the noise-reducing benefits of the f.m. system, that the signal applied to the detector be completely free from amplitude modulation. In practice this is attained by preventing the signal from rising above a given amplitude by means of a limiter (§ 7-15). Since the weakest signal must be amplitude-limited, high gain must be provided ahead of the limiter; the superheterodyne type of circuit is almost invariably used to provide the necessary gain.

The r.f. and i.f. stages in such a superhet are identical in circuit with those in an a.m. receiver. Since the use of f.m. is confined to the ultra-high frequencies (above 28 Mc.) a high intermediate frequency is employed, usually between 4 and 5 Mc. This not only reduces image response but also gives the greater band-width necessary to accommodate wide-band f.m. signals.

Receiver requirements — The primary requirements are sufficient r.f. and i.f. gain to "saturate" the limiter even with a weak signal, sufficient band-width (§ 7-2) to accommodate the full frequency deviation either side of the carrier frequency without undue attenuation at the edges of the band, a limiter circuit which functions properly on both rapid and slow variations in amplitude, and a detector which gives a linear relationship between frequency deviation and *amplitude* output. The audio circuits are the same as in other receivers (§ 7-5) except that it is desirable to cut off the upper audio range by means of a low-pass filter (§ 2-11) because the higher-frequency noise components have the greatest amplitude in an f.m. receiver.

The limiter — Limiter circuits are generally of the plate saturation type (§ 7-15) where low plate and screen voltage are used to limit the plate current flow at high signal amplitudes. Fig. 729-A is a typical circuit. The tube is self-biased (§ 3-6) by a grid leak, R_1, and condenser, C_1. R_2, R_3 and R_4 form a voltage divider (§ 8-8) which puts the desired voltages on the screen and plate. The lower the voltages the lower the signal level at which limiting occurs, but the r.f. output voltage of the limiter also is lower. C_2 and C_3 are the plate and screen by-pass condensers, of conventional value for the intermediate frequency used. The time constant (§ 2-6) of R_1C_1 determines the behavior of the limiter with respect to rapid and slow amplitude variations. For best operation

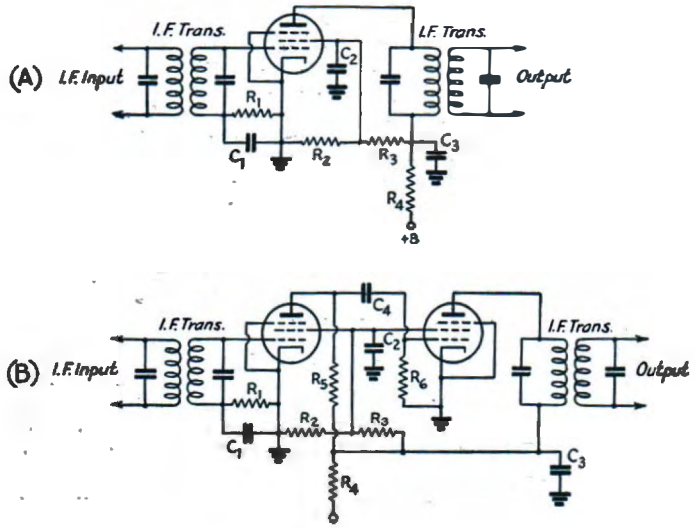


Fig. 729 — F.m. limiter circuits. A, single-tube plate-saturation limiter; B, cascade limiter. Typical values are as follows:

	Circuit A	*Circuit B*
C_1 —	100 $\mu\mu$fd.	100 $\mu\mu$fd.
C_2, C_3 —	0.1 μfd.	0.1 μfd.
C_4 —		250 $\mu\mu$fd.
R_1 —	0.1 megohm.	50,000 ohms.
R_2 —	2000 ohms.	2000 ohms.
R_3 —	50,000 ohms.	50,000 ohms.
R_4 —	0 to 50,000 ohms.	0 to 50,000 ohms.
R_5 —		4000 ohms.
R_6 —		0.2 megohm.

Plate supply voltage should be 250 volts in each circuit.

on impulse noise (§ 7-15) the time constant should be small, but a small time constant limits the range of signal strengths which the limiter can handle without departing from the constant-output condition. A larger time constant is better in the latter respect but is not so effective for rapid variations, hence a compromise set of constants must be used.

The cascade limiter, Fig. 729-B, overcomes this by making the time constant in the first grid circuit suitable for effective operation on impulse noise and that in the second grid (C_4R_6) optimum for a wide range of input signal strengths. This results, in addition, in more constant output over a very wide range of input signal amplitudes because the voltage at the grid of the second stage is already partially amplitude-limited, thus giving the second stage less work to do. Resistance coupling ($R_5C_4R_6$) between stages is used in preference to transformer coupling for simplicity and to prevent unwanted regeneration, additional gain at this point being unnecessary.

The rectified voltage developed across R_1 in either circuit may be used for a.v.c. (§ 7-13).

Discriminator circuits and operation — The f.m. detector is commonly called a *discriminator*, because of its ability to discriminate between frequency deviations above and those below the carrier frequency. The circuit generally used is shown in Fig. 730-A. A special i.f. coupling transformer is used between the limiter and detector. Its secondary, L_1, is centertapped and is connected back to the plate side of the primary circuit, which is otherwise conventional. C_4 is the tuning condenser. The load circuits of the two diode rectifiers (R_1C_1, R_2C_2) are connected in series; the constants used are of the same order as in ordinary diode detector circuits (§ 7-3). The audio output is taken from across the two load resistances.

The primary and secondary circuits are both adjusted to resonance in the center of the i.f. pass-band. The voltage applied to the rectifiers consists of two components, that induced in the secondary by the inductive coupling, and that fed to the center of the secondary through C_2. The phase relations between the two are such that at resonance the rectified load currents are equal in amplitude but flow in opposite directions through R_1 and R_2, hence the net voltage across the terminals marked "audio output" is zero. When the carrier deviates from resonance, the induced secondary current either lags or leads, depending upon whether the deviation is to the high- or low-frequency side, and this phase shift causes the induced current to combine with that fed through C_2 in such a way that one diode gets more voltage than the other when the frequency is below resonance, while the second diode gets the larger voltage when the frequency is higher

than resonance. The voltage appearing across the output terminals is the difference between the two diode voltages, hence a characteristic like that of Fig. 731 results, where the net rectified output voltage has opposite polarity for frequencies on either side of resonance, and up to a certain point becomes greater in amplitude as the frequency deviation is greater. The straight-line portion of the curve is the useful detector characteristic. The separation between the peaks which mark the ends of the linear portion of the curve depends upon the *Q's* of the primary and secondary circuits and the degree of coupling. The separation becomes greater with low *Q's* and close coupling. It is ordinarily set so that the peaks fall just outside the limits of the pass-band, thus utilising most of the straight portion of the curve. Since the audio output is proportional to the change in d.c. voltage with deviation, it is advantageous from this standpoint to have the peak separation the minimum necessary for a linear characteristic.

A second type of discriminator circuit is shown in Fig. 730-B. Two secondary circuits S_1 and S_2 are used, one tuned above the center frequency of the i.f. pass-band, the other below. They are coupled equally to the primary, which is tuned to the center frequency. As the carrier frequency deviates, the voltages induced in the secondaries will change in amplitude, with the larger voltage appearing across the secondary nearer resonance with the instantaneous frequency. The detection characteristic is similar to that of the first type of discriminator. The peak separation is determined by the *Q's* of the circuits, the coefficient of coupling, and the tuning of the two secondaries. High *Q's* and loose coupling are necessary for close peak separation.

F.m. receiver alignment — Alignment of f.m. receivers up to the limiter is carried out as described in § 7-17. For output measurement, a 0–1 milliammeter or 0–500 microammeter should be connected in series with the limiter grid resistor (R_1 in Fig. 729) at the grounded end; or, if the voltage drop across R_1 is used for a.v.c. and the receiver is provided with a tuning meter (§ 7-13), the tuning meter may be used as an output meter. An accurately calibrated signal generator or test oscillator is desirable, since the i.f. should be aligned to be as symmetrical as possible; that is, the output reading should be the same for any two test oscillator settings the same number of kilocycles above or below resonance. It is not necessary to have uniform response over the whole band to be received, although the output at the edges of the band (limit of deviation (§ 5-11) of the transmitted signals) should not be too low — not less than 25% of the voltage at resonance. In communications work a bandwidth of 30 kc. or less (15 kc. or less deviation) is commonly used.

Output readings should be taken with the test oscillator set at intervals of a few kilocycles either side of resonance until the band limits are reached, and the i.f. trimmers adjusted to give as symmetrical a curve as possible.

After the i.f. (and front end) are aligned the limiter operation should be checked. This can be done by temporarily disconnecting C_3, if the discriminator circuit of Fig. 730-A is used, disconnecting R_1 and C_1 from the upper diode's cathode in the same diagram, and inserting the milliammeter or microammeter in series with R_2 at the grounded end. This converts the discriminator to an ordinary diode rectifier. Varying the signal generator frequency over the channel, with the discriminator transformer adjusted to resonance, should show no change in output (at the band-widths used for communications purposes) as indicated by the rectified current read by the meter. At this

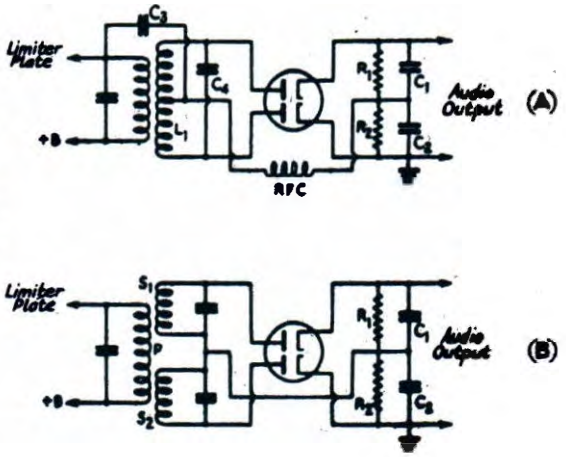


Fig. 730 — F.m. discriminator circuits. In both circuits typical values for C_1 and C_2 are 100 $\mu\mu$fd. each; R_1 and R_2, 0.1 megohm each. C_3 in A is approximately 50 $\mu\mu$fd., depending upon the intermediate frequency; *RFC* should be of the type designed for the i.f. in use (2.5 mh. is satisfactory for i.f.'s of 4 to 5 megacycles). The special three-winding transformer in B is described in the text.

In either circuit the ground may be removed from the lower end of C_2 and moved to the junction of C_1 and C_2 for push-pull audio output.

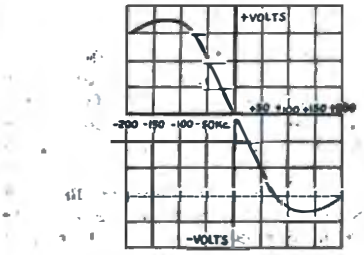


Fig. 731 — Characteristic of a typical f.m. detector. The vertical axis represents the voltage developed across the load resistor as the frequency varies from the exact resonance frequency.

A detector with this characteristic would handle f.m. signals up to a band-width of about 150 kc. over the linear portion of the curve.

point various plate and screen voltages can be tried on the limiter tube or tubes to determine the set of conditions which gives maximum output with adequate limiting (no change in rectified current).

When the limiter has been checked the discriminator connections can be restored, leaving the meter connected in series with R. Provision should be made for reversing the connections to the meter terminals to take care of the reversal in polarity of the net rectified current. Set the signal generator to the center frequency of the band and adjust the discriminator transformer trimmer condensers to resonance, which will be indicated by zero rectified current. Then set the test oscillator at the deviation limit (§ 5-11) on one side of the center frequency and note the meter reading. Reverse the meter terminals and set the test oscillator at the deviation limit on the other side. The two readings should be the same. If they are not, they can be made so by a slight adjustment of the primary trimmer. This will necessitate re-checking the response at resonance to make sure it is still zero. Generally speaking, the secondary trimmer will chiefly affect the zero-response frequency, while the primary trimmer will have most effect on the symmetry of the discriminator peaks. A detector curve having satisfactory linearity can be obtained by cut-and-try adjustment of both trimmers.

Tuning and operation — An f.m. receiver gives greatest noise reduction when the carrier is tuned exactly to the center of the receiver pass-band and to the point of zero response in the discriminator. Because of the decrease in noise, this point is readily recognized. Aside from this no special tuning instructions are necessary. The effectiveness of the receiver will depend almost wholly on how accurately it is aligned.

When an amplitude-modulated signal is tuned in, its modulation practically disappears at exact resonance, only those nonsymmetrical modulation components which may be present being detected. If the signal is to one side or the other of resonance, however, it will be heard and is capable of causing interference to an f.m. signal.

Power Supply

•8-1 POWER SUPPLY REQUIREMENTS

Filament supply — Except for tubes designed for battery operation, the filaments or heaters of vacuum tubes used in both transmitters and receivers are universally operated on alternating current obtained from the power line through a step-down transformer (§ 2-9) delivering a secondary voltage equal to the rated voltage of the tubes used. The transformer should be designed to carry the current taken by the number of tubes which may be connected in parallel (§ 2-6) across it. The filament or heater transformer is generally center-tapped to provide a balanced circuit for eliminating hum (§ 3-6).

For medium- and high-power r.f. stages of transmitters, and for high-power audio stages, it is desirable to use a separate filament transformer for each section of the transmitter, installing the transformer near the tube sockets. This avoids the necessity for abnormally large wires to carry the total filament current for all stages without appreciable filament voltage drop. Maintenance of rated filament voltage is highly important, especially with thoriated-filament tubes, since under- or over-voltage may reduce filament life.

Plate supply — Direct current must be used for the plates of tubes, since any variation in plate current arising from power supply causes will be super-imposed on the signal being received or transmitted, giving an undesirable type of modulation (§ 5-1) if the variations occur at an audio-frequency (§ 2-7) rate. Unvarying direct current is commonly called *pure d.c.* to distinguish it from current which may be unidirectional but of pulsating character. The use of pure direct current on transmitting tubes is required by FCC regulations on frequencies below 60 megacycles.

Sources of plate power — D.c. plate power is usually obtained from rectified and filtered alternating current, but in low-power and portable installations may be secured from batteries. Dry batteries may be used for very low-power portable equipment, but in many cases a storage battery is used as the primary source of power, in conjunction with an interrupter to give pulsating d.c. which is applied to the primary of a step-up transformer (§ 8-10).

Rectified a.c. supplies — Since the power line voltage is ordinarily 115 or 230 volts, a step-up transformer (§ 2-9) must be used to obtain the desired voltage for the plates of the tubes in the equipment. The alternating secondary current is changed to unidirectional current by means of diode rectifier tubes (§ 3-1), then passed through an inductance-capacity filter (§ 2-11) to the load circuit. The *load resistance* in ohms is equal to the d.c. output voltage of the power supply divided by the current in amperes (Ohm's Law, § 2-6).

Voltage regulation — Since there is always some resistance in power supply circuits, and since the filter normally depends to a considerable extent upon the energy storage of inductance and capacity (§ 2-3, 2-5) the output voltage will depend upon the current drain on the supply. The change in output voltage with change in load current is called the *voltage regulation* of the supply. Expressed as a percentage,

$$\% \text{ Regulation} = \frac{100\,(E_1 - E_2)}{E_2}$$

where E_1 is the no-load voltage (no current in the load circuit) and E_2 the full-load voltage (rated current in load circuit).

•8-2 RECTIFIERS

Purpose and ratings — A rectifier is a device which will conduct current in only one direction. The diode tube (§ 3-1) is used almost exclusively for the purpose in d.c. power supplies used with radio equipment. The important characteristics of tubes used as power supply rectifiers are the voltage drop between plate and cathode at rated current, the maximum permissible inverse peak voltage, and the permissible peak plate current.

Voltage drop — Tube voltage drop depends upon the type of tube. In vacuum rectifiers it increases with the current flowing because of space-charge effect (§ 3-1), but can be minimized by using very small spacing between plate and cathode as is done in some rectifiers for receiver power supplies. Mercury-vapor rectifiers (§ 3-5) have a constant drop of about 15 volts regardless of current. This is much smaller than the voltage drops encountered in vacuum rectifiers.

Inverse peak voltage — This is the maximum voltage developed between plate and cathode of the rectifier when the tube is not conducting; i.e., when the plate is negative with respect to the cathode.

Peak plate current — This is the maximum *instantaneous* current flowing through the rectifier. It can never be smaller than the load current in ordinary circuits, and may be several times higher.

Operation of mercury-vapor rectifiers — Because of its constant voltage drop, the mercury vapor rectifier is more susceptible to damage than the vacuum type. With the latter, the increase in voltage drop tends to limit current flow on heavy overloads, but the mercury-vapor rectifier does not have this limiting action and the cathode may be damaged under similar conditions.

In mercury-vapor rectifiers a phenomenon known as "arc-back," or breakdown of the mercury vapor and conduction in the opposite direction to normal, occurs at high inverse peak voltages, hence such tubes always should be operated within their inverse-peak voltage ratings. Arc-back also may occur if the cathode temperature is below normal, therefore the heater or filament voltage should be checked to make sure that the rated voltage is applied. This check should be made at the tube socket to avoid errors caused by drop in the leads from the filament transformer to the tube. For the same reason the cathode should be allowed to come up to its final temperature before plate voltage is applied; the time required for this is of the order of 15 to 30 seconds. When a tube is first installed or is put into service after a long period of idleness, the cathode should be heated for a period of 10 minutes or so before application of plate voltage.

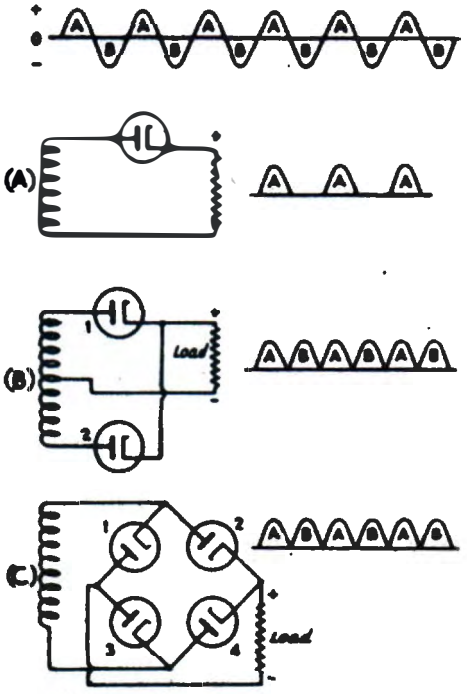


Fig. 801 — Fundamental rectifier circuits.

8-3 RECTIFIER CIRCUITS

Half-wave rectifiers — The simple diode rectifier (§ 3-1) is called a *half-wave rectifier* because it can pass only half of each cycle of alternating current. It is shown in Fig. 801-A. At the top of the figure is a representation of the applied a.c. voltage, with positive and negative alternations (§ 2-7) marked. When the plate is positive with respect to cathode, plate current flows through the load as indicated in the drawing at the right, but when the plate is negative with respect to cathode no current flows. This is indicated by the gaps in the output drawing. The output current is unidirectional, but pulsating.

In this circuit the inverse peak voltage is equal to the maximum transformer voltage, which in the case of a sine wave is 1.41 times the r.m.s. voltage (§ 2-7).

Full-wave center-tap rectifier — Fig. 801-B shows the "full-wave center-tap" rectifier circuit, so called because both halves of the a.c. cycle are rectified and because the transformer secondary winding must consist of two equal parts with a connection brought out from the center. When the upper end of the winding is positive, current can flow through rectifier No. 1 to the load; this current cannot pass through rectifier No. 2 because its cathode is positive with respect to its plate. The circuit is completed through the transformer center-tap. When the polarity reverses, the upper end of the winding is negative and no current can flow through rectifier No. 1, but the lower end is positive and therefore rectifier No. 2 passes current to the load, the return connection again being the center-tap. The resulting wave shape is shown at the right.

Since the two rectifiers are working alternately in this circuit, each half of the transformer secondary must be wound to deliver the full load voltage, hence the total voltage across the transformer terminals is twice that required with the half-wave rectifier. Assuming negligible voltage drop in the particular rectifier which may be conducting at any instant, the inverse peak voltage on the other rectifier is equal to the maximum voltage between the outside terminals of the transformer. In the case of a sine wave this is 1.41 times the total secondary r.m.s. voltage (§ 2-7).

Because energy is delivered to the load at twice the average rate as in the case of a half-wave rectifier, each tube carries only half the load current.

The bridge rectifier — The "bridge" type of full-wave rectifier is shown in Fig. 801-C. Its operation is as follows: When the upper end of the winding is positive, current can flow through No. 2 to the load, but not through No. 1. On the return circuit, current flows through No. 3 by way of the lower end of the transformer winding. When the polarity reverses and the lower end of the winding becomes positive, current flows through No. 4 and the load and through No. 1 by way of the upper side of the transformer. The output wave shape is shown at the right.

The inverse peak voltage is equal to the maximum transformer voltage, or 1.41 times the r.m.s. secondary voltage in the case of a sine wave (§ 2-7). Energy is delivered to the load at the same average rate as in the case of the full-wave center-tap rectifier, so that each *pair* of tubes in series carries half the load current.

● 8-4 FILTERS

Purpose of filter — As shown in Fig. 801, the output of a rectifier is pulsating d.c., which would be unsuitable for most vacuum-tube applications (§ 8-1). A *filter* is used to smooth out the pulsations so that practically unvarying direct current flows through the load circuit. The filter utilizes the energy-storage properties of inductance and capacity (§ 2-3, 2-5) by virtue of which energy stored in electromagnetic and electrostatic fields when the voltage and current are rising is restored to the circuit when the voltage and current fall, thus filling in the "gaps" or "valleys" in the rectified output.

Ripple voltage and frequency — The pulsations in the output of the rectifier can be considered to be caused by an alternating current superimposed on a steady direct current (§ 2-13). Viewed from this standpoint, the filter may be considered to consist of bypass condensers which short-circuit the a.c. while not interfering with the flow of d.c., and *chokes* or inductances which permit d.c. to flow through them but which have high reactance for the a.c. (§ 2-13). The alternating component is called the *ripple*. The effectiveness of the filter may be measured by the *percent ripple*, which is the r.m.s. value of the a.c. ripple voltage expressed as a percentage of the d.c. output voltage. With an effective filter the ripple percentage will be low. Five percent ripple is considered satisfactory for c.w. transmitters, but lower values (of the order of 0.25%) are necessary for hum-free speech transmission and receiver plate supplies.

The ripple frequency depends upon the line frequency and the type of rectifier. In general, it consists of a fundamental plus a series of harmonics (§ 2-7), the latter being relatively unimportant since the fundamental is hardest to smooth out. With a half-wave rectifier the fundamental is equal to the line frequency; with a full-wave rectifier the fundamental is equal to twice the line frequency, or 120 cycles in the case of a 60-cycle supply.

Types of filters — Inductance-capacity filters are of the low-pass type (§ 2-11), using series inductances and shunt capacitances. Practical filters are identified as *condenser-input* and *choke-input*, depending upon whether a capacity or inductance is used as the first element in the filter. Resistance-capacity filters (§ 2-11) are occasionally used in applications, particularly in receivers and speech amplifiers, where the current is very low and the voltage drop in the resistor can be tolerated.

Bleeder resistance — Since the condensers in a filter will retain their charge for a considerable time after power is removed (provided the load circuit is open at the time) it is good practice to connect a resistor across the output of the filter to discharge the condensers when the power supply is not in use. The resistance is usually high enough so that only a relatively small percentage of the total output current is consumed in it during normal operation of the supply.

Components — Filter condensers are made in several different types. Electrolytic condensers are available for voltages up to about 800, and combine high capacity with small size, since the dielectric is an extremely thin film of oxide on aluminum foil. Condensers for higher voltages are usually made with a dielectric of thin paper impregnated with oil. The *working voltage* rating of a condenser is the voltage which it will withstand continuously.

Filter chokes or inductances are wound on iron cores, with a small gap in the core to prevent magnetic saturation of the iron at high currents. When the iron becomes saturated its permeability (§ 2-5) decreases, consequently the inductance also decreases. Despite the air-gap, the inductance of a choke usually varies to some extent with the direct current flowing in the winding, hence it is necessary to specify the inductance at the current which the choke is intended to carry. Its inductance with little or no direct current flowing in the winding may be considerably higher than the load value.

● 8-5 CONDENSER-INPUT FILTERS

Ripple voltage — The conventional condenser-input filter is shown in Fig. 802-A. No simple formulas are available for computing

the ripple voltage, but it will be smaller as both capacity and inductance are made larger. Adequate smoothing for transmitting purposes can be secured by using 4 to 8 μfd. at C_1 and C_2, and 20 to 30 henrys at L_1 with 120-cycle ripple (§ 8-4). A higher ratio of inductance to capacity may be used at higher load resistances (§ 8-1).

For receivers, an additional choke, L_2, and condenser, C_3, of the same approximate values,

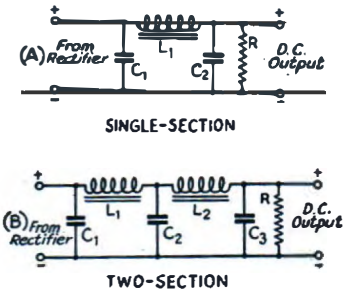


Fig. 802 — Condenser-input filters.

as shown in Fig. 802-B, are used to give additional smoothing. In such supplies the three condensers are generally 8 μfd. each, although the input condenser, C_1, is sometimes reduced to 4 μfd. Inductances of 10 to 20 henrys each will give satisfactory filtering with these capacity values.

For ripple frequencies other than 120 cycles, the inductance and capacity values should be multiplied by the ratio $120/F$, where F is the actual ripple frequency.

The bleeder resistance R should be chosen to draw 10% or less of the rated output current of the supply. Its value is equal to $1000E/I$, where E is the output voltage and I the load current in milliamperes.

Rectifier peak current — The ratio of rectifier peak current to average load current is high with a condenser-input filter. Small rectifier tubes designed for low-voltage supplies (type 80, etc.) generally carry load-current ratings based on the use of condenser-input filters. With rectifiers for higher power, such as the 866/866A, the load current should not exceed about 25% of the rated peak plate current of one tube when a full-wave rectifier is used, or ⅛ the rating with half-wave rectification.

Output voltage — The d.c. output voltage from a condenser-input supply will, with light loads or no load, approach the peak transformer voltage. This is 1.41 times the r.m.s. voltage (§ 2-7) of the transformer secondary in the case of Figs. 801-A and C, or 1.41 times the voltage from center-tap to one end of the secondary in Fig. 801-B. At heavy loads it may decrease to the *average* value of secondary voltage, or about 90% of the r.m.s. voltage or even less. Because of this wide range of output voltage with load current the voltage regulation (§ 8-1) of the condenser-input filter is inherently poor.

The output voltage obtainable from a given supply cannot readily be calculated, since it depends critically upon the load current and filter constants. Under average conditions it will be approximately equal to or somewhat less than the r.m.s. voltage between center-tap and one end of the secondary in the full-wave center-tap rectifier circuit (§ 8-3).

Ratings of components — Because the output voltage may rise to the peak transformer voltage at light loads, the condensers should have a working-voltage rating (§ 8-4) at least this high and preferably somewhat higher as a safety factor. Thus in the case of a center-tap rectifier having a transformer delivering 550 volts each side of the center-tap, the minimum safe condenser voltage rating will be 550 × 1.41, or 775 volts. An 800-volt or preferably a 1000-volt condenser should be used. Filter chokes should have the inductance specified at full-load current, and should have insulation between winding and core adequate to withstand the maximum output voltage.

• 8-6 CHOKE-INPUT FILTERS

Ripple voltage — The circuit of a single-section choke-input filter is shown in Fig. 803-A. For 120-cycle ripple a close approximation of the ripple to be expected at the output of the filter is given by the formula:

$$\left.\begin{matrix}\text{Single}\\ \text{Section}\\ \text{Filter}\end{matrix}\right\} \%\ \text{Ripple} = \frac{100}{LC}$$

where L is in henrys and C in μfd. The product LC must be equal to or greater than 20 to reduce the ripple to 5 per cent or less. This figure represents, in most cases, the economical limit for the single-section filter. Smaller percentages of ripple are usually more economically obtained with the two-section filter of Fig.

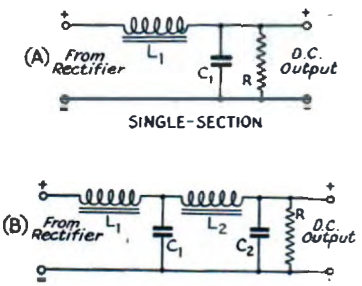


Fig. 803 — Choke-input filters.

803-B. The ripple percentage (120-cycle ripple) with this arrangement is given by the formula:

$$\left.\begin{matrix}\text{Two}\\ \text{Section}\\ \text{Filter}\end{matrix}\right\} \% \text{ Ripple} = \frac{650}{L_1 L_2 (C_1 + C_2)^2}$$

For a ripple of 0.25 per cent or less, the denominator should be 2600 or greater.

The formulas can be used for other ripple frequencies by multiplying each inductance and capacity value in the filter by the ratio $120/F$, where F is the actual ripple frequency.

The distribution of inductance and capacity in the filter will be determined by the value of input-choke inductance required (next paragraph), and the permissible a.c. output impedance. If the supply is intended for use with an audio-frequency amplifier the reactance (§ 2-8) of the last filter condenser should be small (20% or less) compared to the other a.f. resistance or impedance in the circuit, usually the tube plate resistance and load resistance (§ 3-2, 3-3). On the basis of a lower a.f. limit of 100 cycles for speech amplification (§ 5-9), this condition is usually satisfied when the output capacity (last filter capacity) of the filter is 4 to 8 μfd., the higher values being used for the lower tube and load resistances.

The input choke — The rectifier peak current and the supply voltage regulation depend almost entirely upon the inductance of the input choke in relation to the load resistance (§ 8-1). The function of the choke is to raise the ratio of average to peak current (by its energy storage) and to prevent the d.c. output voltage from rising above the average value (§ 2-7) of the a.c. voltage applied to the rectifier. For both purposes its impedance (§ 2-8) to the flow of the a.c. component (§ 8-4) must be high.

The value of input choke inductance which prevents the d.c. output voltage from rising above the average of the rectified a.c. wave is called the *critical inductance*, and for 120-cycle ripple frequency is given by the approximate formula:

$$L_{crit.} = \frac{\text{Load resistance (ohms)}}{1000}$$

For other ripple frequencies, the inductance required will be the above value multiplied by the ratio of 120 to the actual ripple frequency.

With inductance values less than critical the d.c. output voltage will rise because the filter tends to act as a condenser-input filter (§ 8-5). With critical inductance the peak plate current of one tube in a center-tap rectifier will be approximately 10% higher than the d.c. load current taken from the supply.

An inductance of twice the critical value is called the *optimum* value. It gives a further reduction in the ratio of peak to average plate current, and represents the point at which further increase in inductance does not give a corresponding return in improved operating characteristics.

Swinging chokes — The formula for critical inductance indicates that the inductance required varies widely with the load resistance. In the case where there is no load except the bleeder (§ 8-4) on the power supply the critical inductance required is highest; much lower values are satisfactory when the full-load current is being delivered. Since the inductance of a choke tends to rise as the direct current flowing through it is decreased (§ 8-4) it is possible to effect an economy in materials by designing the choke to have a "swinging" characteristic such that it has the required critical inductance value with the bleeder load only, and about the optimum inductance value at full load. Thus in the case where the bleeder resistance is 20,000 ohms and the full-load resistance (including the bleeder) 2500 ohms, a choke which swings from 20 henrys to 5 henrys over the full output-current range will fulfill the requirements.

Resonance — Resonance effects in the series circuit across the output of the rectifier formed by the first choke (L_1) and first filter condenser (C_1) must be avoided, since the ripple voltage would build up to large values (§ 2-10). This is not only the opposite action to that for which the filter is intended, but also may cause excessive rectifier peak currents and abnormally high inverse-peak voltages. For full-wave rectification the ripple frequency will be 120 cycles for a 60-cycle supply (§ 8-4) and resonance will occur when the product of choke inductance in henrys times condenser capacity in microfarads is equal to 1.77. The corresponding figure for 50-cycle supply (100-cycle ripple frequency) is 2.53 and for 25-cycle supply (50-cycle ripple frequency) 13.5. At least twice these products should be used to ensure that no resonance effects will be present.

Output voltage — Provided the input-choke inductance is at least the critical value, the output voltage may be calculated quite closely by the equation:

$$E_o = 0.9E_i - \frac{(I_b + I_L)(R_1 + R_2)}{1000} - E_r$$

where E_o is the output voltage; E_i is the r.m.s. voltage applied to the rectifier (r.m.s. voltage between center-tap and one end of the secondary in the case of the center-tap rectifier); I_b and I_L are the bleeder and load currents, respectively, in milliamperes; R_1 and R_2 are the resistances of the first and second filter chokes; and E_r is the drop between rectifier plate and cathode (§ 8-2). These voltage drops are shown in Fig. 804.

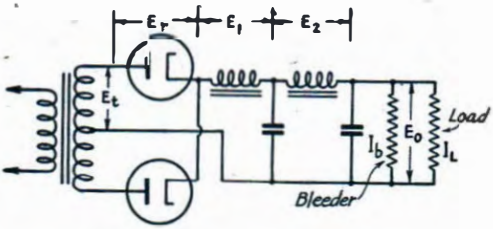


Fig. 804 — Voltage drops in the power supply circuit.

At no load I_L is zero, hence the no-load voltage may be calculated on the basis of bleeder current only. The voltage regulation may be determined from the no-load and full-load voltages (§ 8-1).

Ratings of components — Because of better voltage regulation, filter condensers are subjected to smaller variations in d.c. voltage than in the condenser-input filter (§ 8-5). However, it is advisable to use condensers rated for the peak transformer voltage in case the bleeder resistor should burn out when there is no external load on the power supply, since in this case the voltage will rise to the same maximum value as with a condenser-input filter.

The input choke may be of the swinging type, the required no-load and full-load inductance values being calculated as described above. The second choke (*smoothing choke*) should have constant inductance with varying d.c. load currents. Values of 10 to 20 henrys are ordinarily used. Since chokes are usually placed in the positive leads, the negative being grounded, the windings should be insulated from the core to withstand the full d.c. output voltage of the supply.

• 8-7 THE PLATE TRANSFORMER

Output voltage — The output voltage of the plate transformer depends upon the required d.c. load voltage and the type of rectifier circuit. With condenser-input filters the r.m.s. secondary voltage is usually made equal to or slightly more than the d.c. output voltage, allowing for voltage drops in the rectifier tubes and filter chokes as well as in the transformer itself. The full-wave center-tap rectifier requires a transformer giving this voltage each side of the secondary center-tap (§ 8-3).

With a choke-input filter the required r.m.s. secondary voltage (each side of center-tap for a center-tap rectifier) can be calculated by the equation:

$$E_t = 1.1\left[E_o + \frac{I(R_1 + R_2)}{1000} + E_r\right]$$

where E_o is the required d.c. output voltage, I is the load current (including bleeder current) in milliamperes, R_1 and R_2 are the resistances of the filter chokes, and E_r is the voltage drop in the rectifier. E_t is the full load r.m.s. (§ 2-7) secondary voltage; the open-circuit voltage usually will be 5% to 10% higher.

Volt-ampere rating — The volt-ampere rating (§ 2-8) of the transformer depends upon the type of filter (condenser or choke input). With a condenser-input filter the heating effect in the secondary is higher because of the high ratio of peak to average current, consequently the volt-amperes consumed by the transformer may be several times the watts delivered to the load. With a choke-input filter, provided the input choke has at least the critical inductance (§ 8-6), the secondary volt-amperes can be calculated quite closely by the equation:

$$\text{Sec. V.A.} = 0.00075\ EI$$

where E is the *total* r.m.s. voltage of the secondary (between the outside ends in the case of a center-tapped winding) and I is the d.c. output current in milliamperes (load current plus bleeder current). The primary volt-amperes will be 10% to 20% higher because of transformer losses.

• 8-8 VOLTAGE STABILIZATION

Gaseous regulator tubes — There is frequent need for maintaining the voltage applied to a low-voltage, low-current circuit (such as the oscillator in a superhet receiver or the frequency-controlling oscillator in a transmitter) at a practically constant value regardless of the voltage regulation of the power supply or variations in load current. In such applications gaseous regulator tubes (VR105–30, VR150–30, etc.) can be used to good advantage. The voltage drop across such tubes is constant over a moderately-wide current range. The first number in the tube designation indicates the terminal voltage, the second the maximum permissible tube current.

The fundamental circuit for a gaseous regulator is shown in Fig. 805-A. The tube is connected in series with a *limiting resistor*, R_1, across a source of voltage which must be higher than the *starting* voltage, or voltage required for ionization of the gas in the tube. The starting voltage is about 30% higher than the operating voltage. The load is connected in parallel with the tube. For stable operation a minimum tube current of 5 to 10 milliamperes is required. The maximum permissible current with most types is 30 milliamperes, consequently the load current cannot exceed 20 to 25 milliamperes if the voltage is to be stabilized over a range from zero to maximum load current.

The value of the limiting resistor must lie between that which just permits minimum tube current to flow and that which just passes the maximum permissible tube current when

there is no load current. The latter value is generally used. It is given by the equation

$$R = \frac{1000\ (E_s - E_r)}{I}$$

Where R is the limiting resistance in ohms, E_s the voltage of the source across which tube and resistor are connected, E_r is the rated voltage drop across the regulator tube, and I is the maximum tube current in milliamperes (usually 30 ma.).

Fig. 805-B shows how two tubes may be used in series to give a higher regulated voltage than is obtainable with one, and also to give two values of regulated voltage. The limiting resistor may be calculated as above, using the sum of the voltage drops across the two tubes for E_r. Since the upper tube must carry more current than the lower, the load connected to the low-voltage tap must take small current. The total current taken by the loads on both the high and low taps should not exceed 20 to 25 milliamperes.

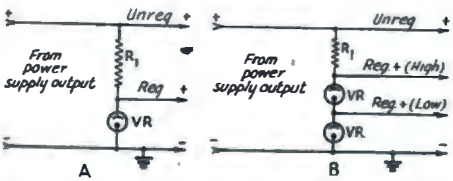


Fig. 805 — **Voltage stabilizing circuits using gaseous regulator tubes.**

Voltage regulation of the order of 1% can be obtained with tubes of this type.

Electronic voltage regulation — A voltage regulator circuit suitable for higher voltages and currents than the gas tubes, and also having the feature that the output voltage can be varied over a rather wide range, is shown in Fig. 806. A high-gain voltage amplifier tube (§ 3-3), usually a sharp-cutoff pentode (§ 3-5) is connected in such a way that a small change in the output voltage of the power supply causes a change in grid bias and thereby a corresponding change in plate current. Its plate current flows through a resistor (R_5) the voltage drop across which is used to bias a second tube — the "regulator" tube — whose plate-cathode circuit is connected in series with the load circuit. The regulator tube therefore functions as an automatically-variable series resistor. Should the output voltage increase slightly, the bias on the control tube becomes more positive, causing the plate current of the control-tube to increase and the drop across R_5 to increase correspondingly. The bias on the regulator tube therefore becomes more negative and the effective resistance of the regulator tube increases, causing the terminal voltage to drop. A decrease in output voltage causes the reverse action. The time lag in the action of the system is negligible and with proper circuit constants, the output voltage can be held within a fraction of a per cent of the desired value throughout the useful range of load currents and over a wide range of supply voltages.

An essential in the system is the use of a constant-voltage bias source for the control tube. The voltage change which appears at the grid of the tube is the *difference* between a fixed negative bias and a positive voltage which is taken from the voltage divider across the output. To get the most effective control, the negative bias must not vary with plate current. The most satisfactory type of bias is a dry battery of 45 to 90 volts, but a gaseous regulator tube (VR75-30) or a neon bulb of the type without the resistor in the base may be used instead. This is indicated in the diagram. If the gas tube or neon bulb is used, a negative-resistance type of oscillation (§ 3-7) may take place at audio frequencies or above, in which case a condenser of 0.1 μfd. or more should be connected across it. A similar condenser between the control tube grid and cathode is also frequently helpful in this respect.

The variable resistor R_3 is used to adjust the bias on the control tube to the proper operating value. It also serves as an output voltage control, setting the value of regulated voltage within the existing operating limits.

The maximum output voltage obtainable is equal to the power supply voltage minus the minimum drop through the regulator tube. This drop is of the order of 50 volts with the

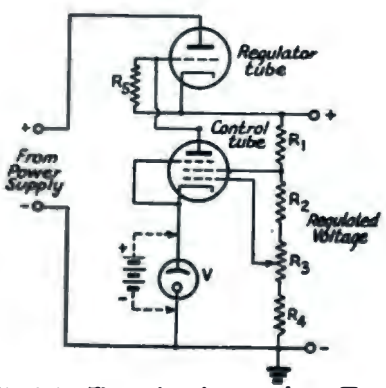


Fig. 806 — **Electronic voltage regulator. The regulator tube is ordinarily a 2A3 or a number of them in parallel, the control tube a 6SJ7 or similar type. The filament transformer for the regulator tube must be insulated for the plate voltage, and cannot supply current to other tubes when a filament-type regulator tube is used. Typical circuit values are as follows: R1, 10,000 ohms; R2, 25,000 ohms; R3, 10,000-ohm potentiometer; R4, 5000 ohms; R5, 0.5 megohm.**

tubes ordinarily used (power triodes having low plate resistance, such as the 2A3). The maximum current is also limited by the regulator tube; 100 milliamperes is a safe value for a 2A3. Two or more regulator tubes may be connected in parallel to increase the current-carrying capacity, no other changes in the circuit being required.

● 8-9 BIAS SUPPLIES

Requirements — A bias supply is not called upon to deliver current to a load circuit but simply to furnish a fixed grid voltage to set the operating point of a tube (§ 3-3). However, in most applications it is nevertheless true that current flows through the bias supply, because such supplies are chiefly used in connection with power amplifiers of the Class-B and Class-C type where grid-current flow is a feature of operation (§ 3-4). In circuit design a bias supply resembles the rectified a.c. plate supply (§ 8-1), having a transformer-rectifier-filter system employing similar circuits. Bias supplies may be classified in two types, those furnishing only *protective* bias, intended to prevent excessive plate current flow in a power tube in case of loss of grid leak bias (§ 3-6) from excitation failure, and those which furnish the actual *operating* bias for the tubes. In the former type voltage regulation (§ 8-1) is relatively unimportant; in the latter it may be of considerable importance.

In general, a bias supply should have well-filtered d.c. output, especially if it furnishes the operating bias for the stage, since ripple voltage may modulate the signal on the grid of the amplifier tube (§ 5-1). Condenser-input filters are generally used, since the regulation of the supply is not a function of the filter. The constants discussed in § 8-5 are applicable.

Voltage regulation — A bias supply must always have a bleeder resistance (§ 8-4) connected across its output terminals to provide a d.c. path from grid to cathode of the tube being biased. Although the grid circuit takes no current from the supply, grid current flows through the bleeder resistor and the voltage across the resistor therefore varies with grid current. This variation in voltage is practically independent of the design of the bias supply unless special voltage-regulating means are used.

Protective bias — This type of bias supply is designed to give an output voltage sufficient to bias the tube to which it is applied to or near the plate-current cut-off point (§ 3-2). A typical circuit is given in Fig. 807. The resistance R_1 is the grid-leak resistor (§ 3-6) for the amplifier tube with which the supply is used, and the normal operating bias is developed by the flow of grid current through this

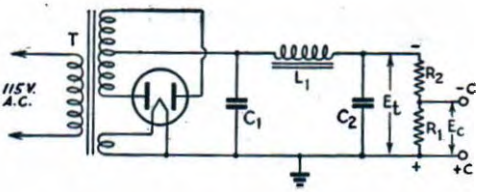


Fig. 807 — Supply for furnishing protective bias to a power amplifier. The transformer T should furnish a peak voltage at least equal to the protective bias required. Other constants are discussed in the text.

resistor. R_2 is connected in series with R_1 across the output of the supply to reduce the voltage across R_1, when there is no grid-current flow, to the cut-off value for the tube being biased. R_2 is given by the formula

$$R_2 = \frac{E_t - E_c}{E_c} \times R_1$$

where E_t is the output voltage of the supply with R_2 and R_1 in series as a load, E_c is the cut-off bias for the tube with which the supply is used, and R_1 is as described above.

When such a supply is used with a Class-C amplifier, the voltage across R_1 from grid-current flow will normally be higher than that from the bias supply itself, since the latter is adjusted to cut-off while the operating bias will be twice cut-off or higher (§ 3-4). In some cases the grid-leak voltage may even exceed the peak output voltage of the transformer (1.41 times half the total secondary voltage, in the circuit shown). The filter condensers in such a bias supply must therefore be rated to stand the maximum operating bias voltage on the Class-C amplifier, if this voltage exceeds the nominal output voltage of the supply.

Voltage stabilization — When the bias supply furnishes operating rather than simply protective bias, the value of bias voltage should be as constant as possible even when the grid current of the biased tube varies. A simple method of improving bias voltage regulation is to make the bleeder resistance low enough so that the current through it from the supply is several times the maximum grid current to be expected. By this means the percentage variation in current is reduced. This method, however, requires that a considerable amount of power be dissipated in the bleeder, which in turn calls for a relatively large power transformer and filter choke.

Bias voltage variation may also be reduced by means of a regulator tube, as shown in Fig. 808. The regulator tube is usually a triode having a plate-current rating adequate to carry the expected grid current. It is cathode-biased (§ 3-6) by the resistor R_1, which is of the order of several hundred thousand ohms or a few megohms so that with no grid current the tube

is biased practically to cut-off. Because of this high resistance, the grid current will flow through the plate resistance of the regulator tube, which is comparatively low, rather than through R_1 and R_2, hence the voltage from the supply across R_1 and the cathode-plate circuit of the regulator tube in series can be considered constant. The bias voltage is equal to the voltage across the tube alone. When grid current flows the voltage across the tube will tend to increase, hence the drop across R_1 decreases, lowering the bias on the regulator and reducing its plate resistance. This in turn reduces the tube voltage drop, and the bias voltage tends to remain constant over a fairly wide range of grid current values.

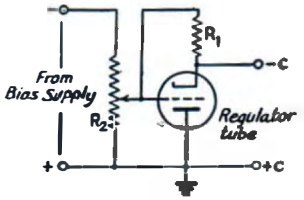


Fig. 808 — Automatic voltage regulator for bias supplies. For best operation the tube used should be one of high mutual conductance (§ 3-2).

At low bias voltages it may be necessary to use a number of tubes in parallel to get sufficient variation of plate resistance for good regulating action. The bias supply must furnish the required bias voltage plus the voltage required to bias the regulator tube to cut-off, considering the output bias voltage as the plate voltage applied to the regulator. The current taken from the bias supply is negligible. R_2 may be tapped to provide a range of bias voltages to meet different tube requirements.

Multi-stage bias supplies — When several power amplifier tubes are to be biased from a single supply, the various bias circuits must be isolated by some means. If the grid currents of all stages should flow through a single bleeder resistor a variation in grid current in one stage would change the bias on all, a condition which would interfere with effective adjustment and operation of the transmitter.

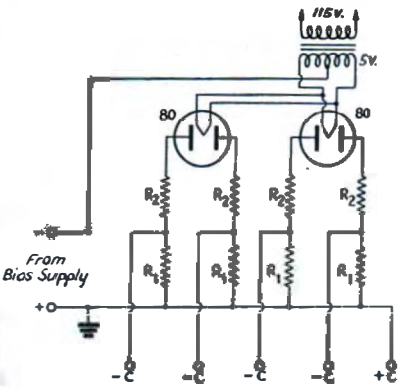


Fig. 809 — Isolating circuit for multiple-stage bias supply.

When protective bias is to be furnished several stages, the circuit arrangement of Fig. 809, using rectifier tubes to isolate the individual grid-leaks of the various stages, may be employed. In the diagram two type 80 rectifiers are used to furnish bias to four stages. Each pair of resistors (R_1R_2) constitutes a separate bleeder across the bias supply. R_1 is the grid-leak for the biased stage; R_2 is a dropping resistor to adjust the voltage across R_1 to the cut-off value (without grid-current flow) for the biased tube. The values of R_1 and R_2 may be calculated as described in the paragraph on protective bias. In this case the bias supply should be designed to have inherently good voltage regulation; i.e., a choke-input filter with appropriate filter and bleeder constants (§ 8-6) should be used, the bleeder being separate from those associated with the rectifier tubes. When the voltage across R_1R_2 rises because of grid-current flow through R_1, the load on the supply will vary (hence the necessity for good voltage regulation in the supply) but there is no interaction of grid currents in the separate bleeders because the rectifiers can pass current in only one direction.

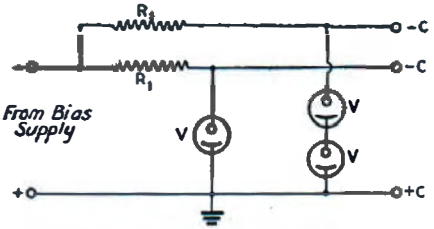


Fig. 810 — Use of gaseous regulator tubes to stabilize bias voltage.

When a single supply is to furnish operating bias for several stages, a separate regulator tube circuit (Fig. 808) may be used for each one. Individual voltages for the various stages may be obtained by appropriate taps on R_2.

Well-regulated bias for several stages may be obtained by the use of gaseous regulator tubes when the voltage and current ratings of the tubes permit their use. This is shown in Fig. 810. A single tube or two or more in series can be used to give the desired bias voltage drop; the bias supply voltage must be high enough to provide starting voltage for the tubes in series. R_1 is the protective resistance (§ 8-8); its value should be calculated for mini-

mum stable tube current. The maximum grid current that can be handled is 20 to 25 milliamperes with available regulator tubes.

8-10 MISCELLANEOUS POWER SUPPLY CIRCUITS

Voltage dividers — A voltage divider is a resistance connected across a source of voltage and tapped at appropriate points from which voltages lower than the terminal voltage may be taken (§ 2-6). Since the voltage at any tap depends upon the current drawn from the tap, the voltage regulation (§ 8-1) of such a divider is inherently poor. Hence a voltage divider is best suited to applications where the currents drawn are constant, or where separate voltage-regulating circuits (§ 8-8) are used to compensate for voltage variations at the taps.

A typical voltage divider arrangement is shown in Fig. 811. The terminal voltage is E, and two taps are provided to give lower voltages E_1 and E_2 at currents I_1 and I_2 respectively. The smaller the resistance between taps in proportion to the total resistance, the smaller the voltage between the taps. In addition to the load currents I_1 and I_2 there is also the bleeder current, I_b. The voltage divider may be the bleeder for the power supply. For convenience, the voltage divider in the figure is considered to be made up of separate resistances, R_1, R_2, R_3, between taps. R_1 carries only the bleeder current, I_b. R_2 carries I_1 in addition to I_b; R_3 carries I_2, I_1 and I_b. For the purpose of calculating the resistances required, a bleeder current I_b must be assumed; generally it is low compared to the total load current (10% or so). Then

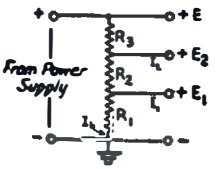


Fig. 811 — Typical voltage-divider circuit.

$$R_1 = \frac{E_1}{I_b}$$

$$R_2 = \frac{E_2 - E_1}{I_b + I_1}$$

$$R_3 = \frac{E - E_2}{I_b + I_1 + I_2}$$

the currents being expressed in amperes.

The method may be extended to any desired number of taps, each resistance section being calculated by Ohm's Law (§ 2-6) using the voltage drop across it and the total current through it. The power dissipated by each section may be calculated by multiplying the same quantities together.

In case it is desired to have the bleeder resistance total to a predetermined value, the same method of calculation may be followed, but different values of bleeder current should be tried until the correct result is found.

Transformerless plate supplies — It is possible to rectify the line voltage directly, without using a step-up power transformer, for certain applications (such as some types of receivers) where the low voltage so obtained is satisfactory. A simple power supply system of this type, using a half-wave rectifier, is shown in Fig. 812. Tubes for this purpose are provided with heaters operating at relatively high voltages (25, 35, 70, or 115 volts) which can be connected across the line in series with other tube filaments and/or a resistor R of suitable value to limit the current to the rated value for the tube heater. The rectifier is often incorporated in the same tube envelope with an audio power amplifier tube.

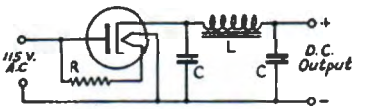


Fig. 812 — Transformerless plate supply with half-wave rectifier.

The half-wave circuit shown has a fundamental ripple frequency equal to the line frequency (§ 8-4) and hence requires more inductance and capacity in the filter for a given ripple percentage (§ 8-5) than the full-wave rectifier. A condenser-input filter is generally used, frequently with a second choke and third condenser (§ 8-5) to provide the necessary smoothing.

A disadvantage of the transformerless circuit is that no ground connection can be used on the power supply unless care is used to insure that the grounded side of the power line is connected to the grounded side of the supply. Receivers using this type of supply are generally grounded through a low capacity (0.05 μfd.) condenser to avoid short-circuiting the line should the line plug be inserted in the socket the wrong way. The input condenser should be at least 16 and preferably 32 μfd. to keep the output voltage high and to improve voltage regulation.

Voltage-doubling circuits — The circuit arrangement of Fig. 813, frequently used in transformerless plate supplies, gives full-wave rectification combined with doubling of the output voltage. This is accomplished by using a double-diode rectifier, one section of which charges C_1 when the line polarity between its plate and cathode is positive while the other

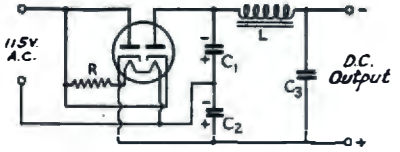


Fig. 813 — Full-wave voltage-doubling transformerless plate supply circuit.

section charges C_2 when the line polarity reverses. Each condenser is thus charged separately to the same d.c. voltage, and they discharge in series into the load circuit. For effective operation of this circuit the capacities of C_1 and C_2 must be at least 16 μfd. each and preferably higher.

The ripple frequency with this circuit is twice the line frequency, since it is a full-wave circuit (§ 8-4). The voltage regulation is inherently poor and depends critically upon the capacities of C_1 and C_2, being better as these capacities are made larger. A typical supply with 16 μfd. each at C_1 and C_2 will have an output voltage of approximately 300 at light loads, dropping to about 210 volts at the rated current of 75 milliamperes.

No direct ground can be used on this supply or on the equipment with which it is used. If an r.f. ground is made through a condenser, the condenser capacity should be small (about 0.05 μfd.) since it is in shunt from plate to cathode of one rectifier. A large capacity (low reactance) would by-pass the rectifier and thereby nullify its operation.

Duplex plate supplies — In some cases it may be advantageous economically to obtain two plate supply voltages from a single power supply, making one or more of the components serve a double purpose. Two circuits of this type are shown in Figs. 814 and 815.

In Fig. 814 a bridge rectifier is used to obtain the full transformer voltage, while a connection is also brought out from the center tap to obtain a second voltage corresponding to half the total transformer secondary voltage. The sum of the currents drawn from the two taps should not exceed the d.c. ratings of the rectifier tubes and transformer. Filter values for each tap should be computed separately (§ 8-6).

Fig. 815 shows how a transformer with multiple secondary taps may be used to obtain both high and low voltages simultaneously. A separate full-wave rectifier is used at each tap. The filter chokes are placed in the common negative lead, but separate filter condensers are required. The sum of the currents drawn from each tap must not exceed the transformer rating and the chokes must be rated to carry the total load current. Each bleeder resistance should have a value in ohms of 1000 times the maximum rated inductance in henrys of the swinging choke, L_1, for best regulation (§ 8-6).

Rectifiers in parallel — Vacuum-type rectifiers may be connected in parallel (plate to plate and cathode to cathode) for higher current carrying capacity. No circuit changes are required.

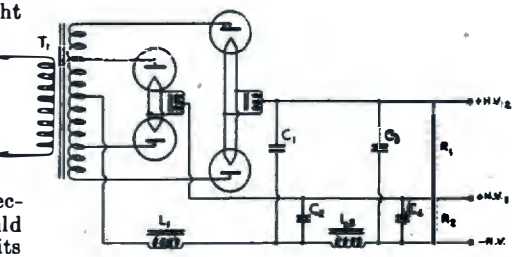


Fig. 815 — Power supply circuit in which a single transformer and set of chokes serve for two different voltages.

When mercury vapor rectifiers are connected in parallel, slight differences in tube characteristics may make one ionize at a slightly lower voltage than the other. Since the ignition voltage is higher than the operating voltage, this means that the first tube to ionize carries the whole load, since the voltage drop is then too low to ignite the second tube. This condition can be prevented by connecting resistors of 50 to 100 ohms in series with each plate as shown in Fig. 816, thereby insuring that a high-enough voltage for ignition will always be available.

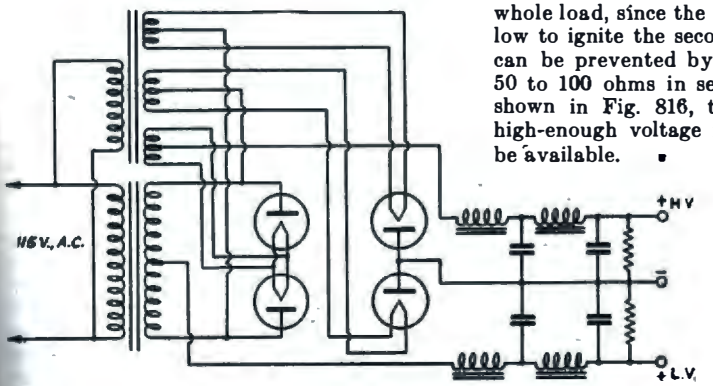


Fig. 814 — Combination bridge and center-tap rectifier to deliver two output voltages with good regulation.

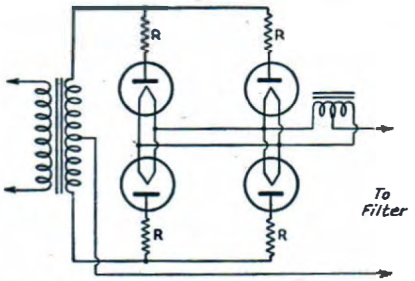


Fig. 816 — Operating mercury-vapor rectifiers in parallel. Resistors marked R should have values between 50 and 100 ohms.

Vibrator power supplies — For portable or mobile work the most common source of power for both filaments and plates is the 6-volt automobile-type storage battery. Filaments may be heated directly from the battery, while plate power is obtained by passing current from the battery through the primary of a suitable transformer, interrupting it at regular intervals to give the changing magnetic field required for inducing a voltage in the secondary (§ 2-5), and rectifying the secondary output. The rectified output is pulsating d.c. which may be filtered by ordinary means (§ 8-5).

Fig. 817 shows two types of circuits used, both with vibrating-reed interrupters (*vibrators*). At A is shown the *non-synchronous* type of vibrator. When the battery circuit is open the reed is midway between the two contacts, touching neither. On closing the battery circuit the magnet coil pulls the reed into contact with the lower point, causing current to flow through the lower half of the transformer primary winding. Simultaneously the magnet coil is short-circuited and the reed swings back, and is carried by inertia into contact with the upper point, causing current to flow through the upper half of the transformer primary. The magnet coil is again energized and the cycle repeats itself, usually at a rate about equivalent to a 60-cycle supply frequency.

The synchronous circuit of Fig. 817-B is provided with an extra pair of contacts which rectify the secondary output of the transformer, thus eliminating the need for a separate rectifier tube. The secondary center-tap furnishes the positive output terminal when the relative polarities of primary and secondary windings are correct. The proper connections may be determined by experiment, reversing the secondary connections if the first trial is wrong.

The buffer condenser, C_2, across the transformer secondary is used to absorb surges which would occur on breaking the current, when the magnetic field collapses practically instantaneously and hence causes a very high voltage to be induced in the secondary (§ 2-5). Its value is usually between 0.005 and 0.03 μfd. and for 250–300 volt supplies should be rated at 1500 to 2000 volts d.c. The proper value is rather critical and should be determined experimentally, the optimum value being that which results in least battery current for a given rectified d.c. output from the supply.

Sparking at the vibrator contacts causes r.f. interference ("hash") when such a supply is used with a receiver. This can be minimized by installing hash filters, consisting of RFC_1 and C_1 in the battery circuit, and RFC_2 with C_3 in the d.c. output circuit. C_1 is usually from 0.5 to 1 μfd., a 50-volt rating being adequate. RFC_1 consists of about 50 turns wound to about half-inch diameter, No. 12 or No. 14 wire being required to carry the rather heavy battery current without undue loss of voltage. C_3 may be of the order of 0.01 to 0.1 μfd., and RFC_2 a 2.5-millihenry choke of ordinary design. Equally as important as the hash filter is thorough shielding of the power supply and its connecting leads, since even a small piece of wire or metal will radiate enough hash to cause interference in a sensitive receiver.

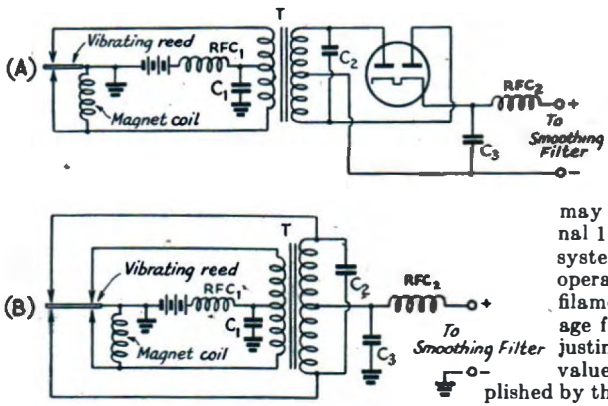


Fig. 817 — Vibrator power supply circuits. Constants and operation are discussed in the text.

Line-voltage adjustment — In some localities the line voltage may vary considerably from the nominal 115 volts as the load on the power system changes. Since it is desirable to operate tube equipment, particularly filaments and heaters, at constant voltage for maximum life, a means of adjusting the line voltage to the rated value is desirable. It can be accomplished by the circuit shown in Fig. 818, utilizing a step-down transformer with a tapped secondary connected as an auto-transformer

(§ 2-9). The secondary should preferably be tapped in steps of two or three volts, and should have sufficient total voltage to compensate for the widest variations encountered. Depending upon the end of the secondary to which the line is connected, the voltage to the load can be made either higher or lower than the line voltage. A secondary winding capable of carrying five amperes or so will be adequate for loads up to 500 volt-amperes on a 115-volt line.



Fig. 8-10—Line-voltage compensation by means of tapped step-down transformer.

CHAPTER NINE

Wave Propagation

● 9-1 RADIO WAVES

Nature of radio waves — Radio waves are electromagnetic waves, consisting of traveling electrostatic and electromagnetic fields so related to each other that the energy is evenly divided between the two, and with the lines of force in the two fields at right angles to each other in a plane perpendicular to the direction of propagation as shown in Fig. 901. Except for the difference in order of wavelength, they have the same nature as light waves, travel with the same speed (300,000,000 meters per second in space), and, similarly to light, can be reflected, refracted and diffracted.

Polarization — The polarization of a radio wave is taken as the direction of the lines of force in the electrostatic field. If the direction of the electrostatic component is perpendicular to the earth, the wave is said to be *vertically polarized*, while if the electrostatic component is parallel to the earth the wave is *horizontally polarized*. The electromagnetic component, being at right-angles to the electrostatic, therefore has its lines of force vertical when the wave is horizontally polarized, and horizontal when the wave is vertically polarized.

Reflection — Radio waves may be reflected from any sharply-defined discontinuity, of suitable characteristics and dimensions, in the medium in which they are propagated. Any good conductor meets this requirement provided its dimensions are at least comparable with the wavelength. The surface of the earth also forms such a discontinuity, and waves are readily reflected from the earth.

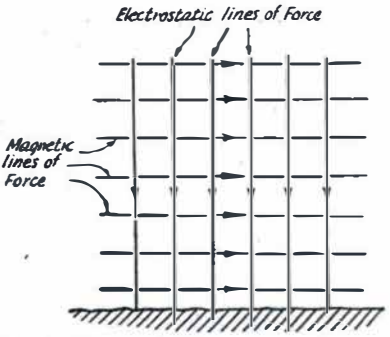


Fig. 901 — Representation of electrostatic and electromagnetic lines of force in a radio wave. Arrows indicate instantaneous directions of the fields for a wave traveling out of the page toward the reader. Reversing the direction of one set of lines would reverse the direction of travel.

Refraction — Refraction of radio waves is similar to the refraction of light; that is, the wave is bent when moving obliquely into a region having a different refractive index from that of the region it leaves. This bending results because the velocity of propagation differs in the two regions, so that the part of the wavefront which enters first travels faster or slower than the part which enters the new region last, causing the wavefront to turn.

Diffraction — When a wave grazes the edge of an object in passing it is bent around the object. This bending is called *diffraction*.

Ground and sky waves — Two types of waves occur, one traveling along the surface of the ground, the other traveling through the atmosphere and having no contact with the ground along most of its path. The former is called the *ground wave*, the latter the *sky wave*. The ground wave dies out rather rapidly, but the sky wave can travel to great distances, especially on high frequencies (short wavelengths).

Field strength — The intensity of the electrostatic field of the wave is called the *field strength* at the point of measurement. It is usually expressed in microvolts per meter, and is equivalent to the voltage induced in a wire one meter long placed with its axis parallel to the direction of polarization.

● 9-2 THE GROUND WAVE

Description — The ground wave is continuously in contact with the surface of the earth and, in cases where the distance of transmission makes the curvature of the earth important, is propagated by means of diffraction, with refraction in the lower atmosphere also having some effect. The ground wave is practically independent of seasonal and day and night effects at high frequencies (above 1500 kc.).

Polarization — A ground wave must be vertically polarized because the electrostatic field of a horizontally-polarized wave would be short-circuited by the ground, which acts as a conductor at the frequencies for which the ground wave is of most interest.

Ground characteristics and losses — The

wave induces a current in the ground in traveling along its surface. If the ground were a perfect conductor there would be no loss of energy, but actual ground has appreciable resistance so that the current flow causes some energy dissipation. This loss must be supplied by the wave, which is correspondingly weakened. Hence the transmitting range depends upon the ground characteristics. Because sea water is a good conductor, the range will be greater over the ocean than over land. The losses increase with frequency, so that the ground wave is rapidly attenuated at high frequencies and above about 2 megacycles is of little importance except in purely local communication.

Range of ground wave — At frequencies in the vicinity of 2 megacycles the ground wave range is of the order of 200 miles over average land and perhaps two or three times as far over sea water, for a medium-power transmitter (500 watts or so) using a good antenna. At higher frequencies the range drops off rapidly, and above 4 megacycles the ground wave is useful only for work over quite short distances.

● 9-3 THE IONOSPHERE

Description — Since a sky wave leaving the transmitting antenna has to travel upward with respect to the earth's surface, it would simply continue out into space if its path were not bent sufficiently to bring it back to the earth. The medium which causes such bending is the *ionosphere*, a region in the upper atmosphere where free ions and electrons exist in sufficient quantity to cause a change in the refractive index. Ultraviolet radiation from the sun is considered to be responsible for the ionization. The ionosphere is not a single region but consists of a series of "layers" which occur at different heights, each layer consisting of a central region of ionization which tapers off in intensity both above and below.

Refraction, absorption, reflection — For a given intensity of ionization the amount of refraction becomes less as the frequency of the wave becomes higher (shorter wavelength). The bending is therefore smaller at high than at low frequencies, and if the frequency is raised to a high-enough value the bending eventually will become too slight to bring the wave back to earth, even when it enters the ionosphere at a very small angle to the "edge" of the ionized zone. At this and higher frequencies long-distance communication becomes impossible.

The greater the intensity of ionization the greater the bending on a given frequency. Thus an increase in ionization increases the maximum frequency which can be bent sufficiently for long-distance communication. The wave loses some energy in the ionosphere, and this energy loss increases with ionization density and the wavelength. Unusually high ionization may cause complete absorption of the wave energy, especially when the ionization is high in the lower regions of the ionosphere and below the lowest normally-useful layer. When the wave is absorbed in the ionosphere it is no more useful for communication than if it had passed through without sufficient bending to bring it back to earth.

In addition to refraction, reflection may take place at the lower boundary of a layer, if that boundary is well-defined; i.e., if there is an appreciable change in ionization within a relatively short interval of distance. For waves approaching the layer at or near the perpendicular, the change in ionization must take place within a difference in height comparable to the wavelength, hence reflection is more apt to occur at longer wavelengths (lower frequencies).

Critical frequency — When the frequency is low enough, a wave sent vertically upward to the ionosphere will be bent sufficiently to return to the transmitting point. The highest frequency at which this occurs, for a given state of the ionosphere, is called the *critical frequency*. It serves as an index for transmission conditions, although it is not the highest *useful* frequency since waves which enter the ionosphere at smaller angles than 90 degrees (vertical) will be bent sufficiently to return to earth. The maximum usable frequency, for waves leaving the earth at very small angles to the horizontal, is in the vicinity of three times the critical frequency.

Besides being directly observable, the critical frequency is of more practical interest than ionization density because it includes the effects of absorption as well as refraction.

Virtual height — Although a layer is a region of considerable depth, it is convenient to assign to it a definite height, called the *virtual height*. The virtual height is the height from

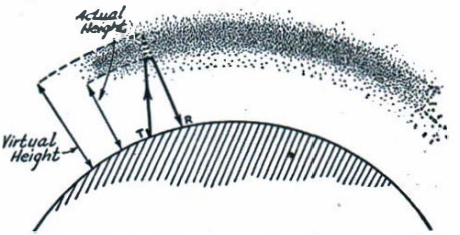


Fig. 902 — **Bending in the ionosphere and method of determining virtual height.**

which a pure reflection would give the same effect as the refraction which actually takes place. This is illustrated in Fig. 902. The wave traveling upward is bent back over a path having appreciable radius of turning, and a measurable interval of time is consumed in the turning process. The virtual height is then the

height of the triangle formed as shown, having equal sides of a total length equivalent to the actual time taken for the wave to travel from T to R.

The E layer — The lowest normally-useful layer is called the E layer. Its average height (maximum ionization) is about 70 miles. The ionization density is greatest around local noon, and the layer is only weakly ionized at night when the radiation from the sun is not present. This is because the air at this height is sufficiently dense so that free ions and electrons very quickly meet and recombine.

The F, F_1 and F_2 layers — The second principal layer is the F, which is at a height of about 175 miles at night. In this region the air is so thin that recombination of ions and electrons takes place very slowly, since the particles can travel relatively great distances before meeting. The ionization decreases after sundown, reaching a minimum just before sunrise. In the daytime the F layer splits into two layers, the F_1 and F_2, at average virtual heights of about 140 miles for the F_1 and around 200 miles for the F_2. These are most highly ionized at about local noon, and merge again at sunset into the F layer.

Seasonal effects — In addition to day and night variations, there are also seasonal changes in the ionosphere as the quantity of radiation received from the sun changes. Thus the E layer has higher critical frequencies in the summer (about 4 Mc., average, in daytime) than in the winter, when the critical frequency is near 3 Mc. The F layer shows little variation, the critical frequency being of the order of 4 to 5 Mc. in the evening. The F_1 layer, which has a critical frequency in the neighborhood of 5 Mc. in summer, usually disappears in winter. The critical frequencies are highest in the F_2 layer in winter (11 to 12 Mc.) and lowest in summer (around 7 Mc.). The virtual height of the F_2 layer is also less in winter (around 185 miles) than in summer (average 250 miles).

In the spring and fall a transition period occurs, and conditions in the ionosphere are more variable at these times of the year.

Sunspot cycles — The critical frequencies mentioned in the preceding paragraph are mean values, since the ionization also varies with the 11-year sunspot cycle, being higher during times of greatest sunspot activity. Critical frequencies are highest during sunspot maxima and lowest during sunspot minima. The E critical frequency does not change greatly, but the F and F_2 critical frequencies change in a ratio of about 2 to 1.

Magnetic storms and other disturbances — Unusual disturbances in the earth's magnetic field (magnetic storms) usually are accompanied by disturbances in the ionosphere, when the layers apparently break up and expand. There is usually also an increase in absorption during such a period. Radio transmission is poor and there is a drop in critical frequencies so that lower frequencies must be used for communication. A storm may last for several days.

Unusually high ionization in the region of the atmosphere below the normal ionosphere may increase absorption to such an extent that sky-wave transmission becomes impossible on high frequencies. The length of such a disturbance may be several hours, with a gradual falling off of transmission conditions at the beginning and an equally gradual building up at the end of the period. *Fadeouts*, similar to the above in effect, are caused by sudden disturbances on the sun. They are characterized by very rapid ionization, with sky-wave transmission disappearing almost instantly, occur only in daylight, and do not last as long as the first type of absorption.

● 9-4 THE SKY WAVE

Wave angle (angle of radiation) — The smaller the angle at which the wave leaves the earth, the smaller the bending required in the ionosphere to bring it back and, in general, the greater the distance between the point where it leaves the earth and that at which it returns (§ 9-3). This is shown in Fig. 903. The vertical angle which the wave makes with a tangent to the earth is called the *wave angle*, or *angle of radiation*, the latter term being used more in connection with transmitting than receiving.

Skip distance — Since more bending is re-

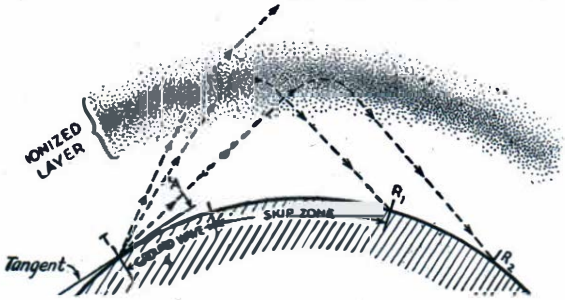


Fig. 903 — Refraction of sky waves, showing critical wave angle and skip zone.

quired to return the wave to earth when the wave angle is high, it is found that at high frequencies the refraction frequently is not great enough to give the required bending unless the wave angle is smaller than a certain angle called the *critical angle*. This is shown in Fig. 903, where wave angles A and lower give useful signals, but waves sent at higher angles travel through the layer and do not return. The distance between T and R_1 is therefore the shortest possible distance over which sky-wave communication can be carried on. The area between the end of the useful ground wave and the beginning of sky-wave reception is called the *skip zone*. The skip distance depends upon the frequency and the state of the ionosphere and is greater the higher the transmitting frequency and the lower the critical frequency (§ 9-3). It also depends upon the height of the layer in which the refraction takes place, the higher layers giving longer distances for the same wave angle. The wave angles at the transmitting and receiving points are usually, although not necessarily, approximately the same for a given wave path.

It is readily possible for the sky wave to pass through the E layer and be refracted back to earth from the F, F_1 or F_2 layers. This is because the critical frequencies are higher in the latter layers, so that a signal too high in frequency to be returned by the E layer can still come back from the F_1, F_2 or F, depending upon the time of day and the conditions existing. Depending upon the wave angle and frequency, it is also possible to have communications via either the E or F_1-F_2 layers on the same frequency.

Multi-hop transmission — On returning to earth the wave can be reflected (§ 9-1) upward and travel again to the ionosphere where refraction once more takes place, again with bending back to the earth. This process, which can be repeated several times, is necessary for transmission over great distances because of the limited heights of the layers and the curvature of the earth, since at the lowest useful wave angles (of the order of a few degrees, waves at smaller angles generally being absorbed rapidly at high frequencies by being in contact with the earth) the maximum one-hop distance is about 1250 miles with refraction from the E layer, and around 2500 miles from the F_2 layer. Ground losses absorb some of the energy from the wave on reflection, the amount of loss varying with the type of ground and being least for reflection from sea water. When the distance permits, it is better to have one hop rather than several, since the multiple reflections introduce losses which are higher than those caused by the ionosphere.

Multi-hop transmission is shown in Fig. 904, two- and three-hop paths being indicated.

Fading — Two or more parts of the wave may follow slightly different paths in traveling to the receiving point, in which case the difference in path lengths will cause a phase difference to exist between the wave components at the receiving antenna. The field strength may therefore have any value between the numerical sum of the components (when they are all in phase) and zero (when there are only two components and they are exactly out of phase). Since the paths change from time to time this causes a variation in signal strength called *fading*. Fading can also result from the combination of single-hop and multi-hop waves, or the combination of a ground wave and sky wave. The latter condition gives rise to an area of severe fading near the limiting distance of the ground wave, better reception being obtained at both shorter and longer distances where one component or the other is considerably stronger. Fading may be rapid or slow, the former type usually resulting from rapidly changing conditions in the ionosphere, the latter occurring when transmission conditions are relatively stable.

●9-5 ULTRA-HIGH-FREQUENCY PROPAGATION

Direct ray — In the ultra-high frequency part of the spectrum (above 30 megacycles) the bending of the waves in the normal ionosphere layers is so slight that the sky wave (§ 9-4) does not ordinarily play any part in communication. The ground-wave (§ 9-2) range also is extremely limited because of high absorption in the ground at these frequencies. Normal u.h.f. transmission is by means of a *direct ray*, or wave traveling directly from the transmitter to the receiver through the atmosphere. Since the energy lost in ground absorption by a wave

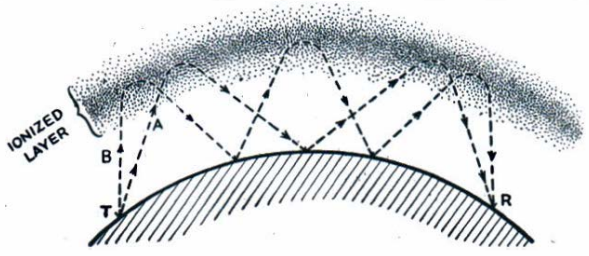


Fig. 904 — Multi-hop transmission.

traveling close to the ground decreases very rapidly with its height in wavelengths above ground, an ultra-high-frequency wave can be relatively close (in physical height) to the ground without suffering the absorption effects which would occur at the same physical heights on longer wavelengths.

Since the wave travels practically in a straight line, the maximum signal strength can be obtained only when there is an unobstructed atmospheric path between the transmitter and receiver. This means that the transmitting and receiving points should be sufficiently high to provide such a path, and on long paths the curvature of the earth must be taken into account as well as the intervening terrain.

Reflected ray—In addition to the direct ray, part of the wave strikes the ground between the transmitter and receiver and is reflected upward at a slight angle to produce a *reflected-ray* component at the receiver. This is shown in Fig. 905. The reflected ray is more or less out of phase with the direct ray, hence the net field strength at the receiving point is less than that of the direct ray alone. The canceling effect of the reflected ray depends upon the heights of the transmitter and receiver above the point of reflection, the ground losses when reflection takes place, and the frequency,



Fig. 905 — Direct and reflected waves in u.h.f. transmission.

decreasing with an increase in any of these factors.

Atmospheric refraction — There is normally some change in the refractive index of the air with height above ground, its nature being such as to cause the waves to bend slightly towards the ground. Where curvature of the earth must be considered, this has the effect of lengthening the distance over which it is possible to transmit a direct ray. It is convenient to consider the effect of this "normal" refraction as equivalent to an increase in the earth's radius in determining the transmitting and receiving heights necessary to provide a clear path for the wave. The equivalent radius, taking refraction into account, is 4/3 the actual radius.

Range vs. height — The height required to provide a clear path ("line of sight") over level ground from an elevated transmitting point to a receiving point on the surface, not including the effect of refraction, is

$$h = \frac{d^2}{1.51}$$

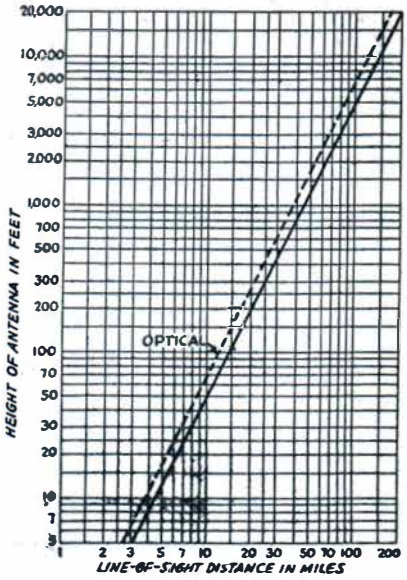


Fig. 906 — Chart for determining line-of-sight distance for u.h.f. transmission. Solid line includes effect of refraction, dotted line is the optical distance.

where h is the height in feet and d the distance in miles. Conversely, the line of sight distance in miles for a given height in feet is equal to $1.23\sqrt{h}$. Taking refraction into account, the latter equation becomes $1.41\sqrt{h}$. The graph of Fig. 906 gives the answer directly when one quantity is known.

When transmitter and receiver are both elevated the maximum direct ray distance to ground level as given by the formulas can be determined separately for each. Adding the two distances so obtained together will give the maximum distance by which they can be separated for direct-ray communication. This is shown in Fig. 907.

Diffraction — At distances beyond the direct-ray path, the wave is diffracted around the curvature of the earth. The diffracted wave is attenuated very rapidly, so that beyond the maximum direct-ray distance the signal strength decreases considerably faster

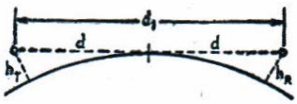


Fig. 907 — Method of determining total line-of-sight distance when both transmitter and receiver are elevated. Since only earth curvature is taken into account in Fig. 906, irregularities in the ground between the transmitting and receiving points must be considered for each actual path.

with distance than it does within the direct-ray path.

● 9-6 TROPOSPHERE REFRACTION

Temperature inversions — The refractive index of the lower atmosphere depends principally upon the temperature, moisture content and pressure. Of the three, only temperature differences cause a large enough change in refractive index to refract ultra-high frequency waves in such a way as to extend the distance range beyond the normal direct-ray and diffracted-wave ranges discussed in the preceding section. This occurs when there is a "temperature inversion," or a layer of warm air over cooler air near the ground. Temperature inversions are relatively frequent in the summer, and usually occur at heights from a few thousand feet to two miles or so above the ground.

Lower atmosphere bending — When there is a sufficiently marked temperature inversion; i.e., a rapid rise of temperature with height, a wave is refracted back to earth in much the same way as in the ionosphere, although the cause of the change in refractive index is different. The amount of bending is small compared to the bending in the ionosphere. Consequently the wave angle (§ 9-4) must be quite low (zero or nearly so), but since the bending takes place at a low altitude it is possible to extend the range of u.h.f. signals to several hundred miles when both transmitter and receiver are well below the line of sight.

Fig. 908 illustrates the conditions existing when the air is "normal" and when a temperature inversion is present. Since the bending is relatively small, it is advantageous to have as much height as possible at both the receiving and transmitting points, even though these heights may be considerably less than those necessary for "line of sight" transmission.

Frequency effects — The amount of bending is greater at longer wavelengths (lower frequencies) but is not usually observed at frequencies much below 28 Mc., partly because it is masked by other effects. The upper limit of frequency at which useful bending ceases is not known, but transmission by this means is frequent on 56 and 112 Mc.

● 9-7 SPORADIC-E IONIZATION

Description — Under certain conditions small regions or "patches" of unusually dense ionization may appear in the *E* layer of the ionosphere, for reasons not yet clearly understood. This is known as *sporadic E* ionization, and the change in refractive index in such a patch or cloud is frequently great enough to cause waves having frequencies as high as 60 Mc. to be bent back to earth. The dimensions of a sporadic *E* cloud are relatively small,

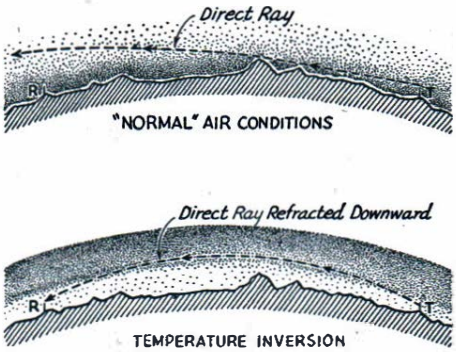


Fig. 908 — Effect of a temperature inversion in extending the range of u.h.f. signals.

hence communication by means of it is restricted to transmitting and receiving localities so situated with respect to the cloud and to each other that a refracted wave path is possible.

The abnormal ionization usually disappears in the course of a few hours. Sporadic *E* ionization is more frequent in the summer than winter, and may occur at any time of the day or night.

Transmission characteristics — Sporadic *E* refraction may take place at all frequencies up to the region of 60 megacycles. At the present time there are no known cases of such refraction on 112 Mc. When sporadic *E* ionization is present, skip distance is greatly reduced (when a wavepath via the cloud is possible to a given receiving location) on the frequencies where transmission is normally by means of the *F*, F_1 and F_2 layers; that is, from about 3.5 to 30 megacycles at night. The skip zone may in fact disappear entirely over most of the high-frequency spectrum, since the critical frequencies may rise to as high as 12 Mc. for sporadic *E*.

At ultra-high frequencies the bending is relatively small compared to lower frequencies, and only wave angles of the order of 5 degrees and less are useful in most cases. The transmitting and receiving points thus must be sufficiently distant from the cloud to enable a wave leaving the transmitter at such angles to strike it, and the cloud should be approximately on, and near the center of, the line joining the transmitter and receiver. Unless the ionization is extremely intense, the minimum distance of transmission on 56 Mc. is of the order of 800 miles and the maximum distance about 1250 miles.

Multi-hop transmission by means of two sporadic *E* clouds properly situated with respect to a transmitter and receiver is possible, but rather rare. Distances up to 2500 miles or so have been attained on 56 Mc. by this means.

Antenna Systems

•10-1 ANTENNA PROPERTIES

Wave propagation and antenna design — For most effective transmission, the propagation characteristics of the frequency under consideration must be given due consideration in selecting the type of antenna to use. These have been discussed in Chapter 9. On some frequencies the angle of radiation and polarization may be of relatively little importance; on others they may be all-important. On a given frequency, the type of antenna best suited for long-distance transmission may not be as good as a different type for shorter-range work.

The important properties of an antenna or antenna system are its polarization, angle of radiation, impedance, and directivity.

Polarization — The polarization of a straight-wire antenna is its position with respect to the earth. That is, a vertical wire transmits vertically polarized waves and a horizontal antenna generates horizontally polarized waves (§ 9-1). The wave from an antenna in a slanting position contains both vertical and horizontal components.

Angle of radiation — The wave angle (§ 9-4) at which an antenna radiates best is determined by its polarization, height above ground, and the nature of the ground. Radiation is not all at one well-defined angle, but rather is dispersed over a more or less large angular region, depending upon the type of antenna. The angle is measured in a vertical plane with respect to a tangent to the earth at the transmitting point.

Impedance — The impedance (§ 2-8) of the antenna at any point is the ratio of voltage to current at that point. It is important in connection with feeding power to the antenna, since it constitutes the load resistance represented by the antenna. At high frequencies it consists chiefly of radiation resistance (§ 2–12). It is understood to be measured at a current loop (§ 2–12) unless otherwise specified.

Directivity — All antennas radiate more power in certain directions than in others. This characteristic, called *directivity*, must be considered in three dimensions, since directivity exists in the vertical plane as well as in the horizontal plane. Thus the directivity of the antenna will affect the wave angle as well as the actual compass directions in which maximum transmission takes place.

Current — The field strength produced by an antenna is proportional to the current flowing in it. Since standing waves are generally present on an antenna, the parts of the wire carrying the higher current therefore have the greatest radiating effect.

Power gain — The ratio of power required to produce a given field strength with a "comparison" antenna, to the power required to produce the same field strength with a specified type of antenna is called the *power gain* of the latter antenna. It is used in connection with antennas intentionally designed to have directivity, and is measured in the optimum direction of the antenna under test. The comparison antenna is almost always a half-wave antenna having the same polarization as the antenna under consideration. Power gain is usually expressed in decibels (§ 3-3).

•10-2 HALF-WAVE ANTENNA

Physical and electrical length — The fundamental form of antenna is a single wire whose length is approximately equal to half the transmitting wave-length. It is the unit from which many more complex forms of antennas are constructed. It is sometimes known as a *Hertz* or *doublet* antenna.

The length of a half wave in space is

$$\textit{length (feet)} = \frac{492}{\textit{Freq. (Mc.)}} \qquad (1)$$

The actual length of a half-wave antenna will not be exactly equal to the half wavelength in space but is usually about 5% less, because of capacitance at the ends of the wire (*end effect*). The reduction factor increases slightly as the frequency is increased. Under average conditions, the following formula will give the length of a half-wave antenna to sufficient accuracy, for frequencies up to 30 Mc.

$$\textit{Length of half-wave antenna (feet)} = \frac{492 \times 0.95}{\textit{Freq. (Mc.)}} \quad \frac{468}{\textit{Freq. (Mc.)}} \qquad (2)$$

At 56 Mc. and higher frequencies the somewhat larger end effects cause a slightly greater reduction in length, so that for these frequencies,

$$\text{Length of half-wave antenna (feet)} = \frac{492 \times 0.94}{\text{Freq. (Mc.)}} = \frac{462}{\text{Freq. (Mc.)}} \quad (3)$$

$$\text{or length (inches)} = \frac{5540}{\text{Freq. (Mc.)}} \quad (4)$$

Current and voltage distribution — When power is fed to such an antenna the current and voltage vary along its length (§ 2-12). The distribution, which is practically a sine curve, is shown in Fig. 1001. The current is maximum at the center and nearly zero at the ends, while the opposite is true of the r.f. voltage.

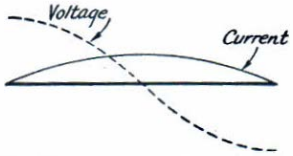


Fig. 1001 — Current and voltage distribution on a half-wave antenna.

The current does not actually reach zero at the current *nodes*, (§ 2-12) because of the end effect; similarly, the voltage is not zero at its node because of the resistance of the antenna, which consists of both the r.f. resistance of the wire (*ohmic resistance*) and the radiation resistance (§ 2-12). Usually the ohmic resistance of a half-wave antenna is small enough, in comparison with the radiation resistance, to be neglected for all practical purposes.

Impedance — The radiation resistance of a half-wave antenna in free space — that is, sufficiently removed from surrounding objects so that they do not affect the antenna's characteristics — is 73 ohms, approximately. The value under practical conditions will vary with the height of the antenna, but is commonly taken to be in the neighborhood of 70 ohms. It is pure resistance, and is measured at the center of the antenna. The impedance is minimum at the center, where it is equal to the radiation resistance, and increases toward the ends (§ 10-1). The end value will depend on a number of factors such as the height, physical construction, and the position with respect to ground.

Conductor size — The impedance of the antenna also depends upon the diameter of the conductor in relation to its length. The figures above are for wires of practicable sizes. If the diameter of the conductor is made large, of the order of 1% or more of the length, the impedance at the center will be raised and the impedance at the ends decreased. This increase in center impedance (of the order of 50% for a diameter/length ratio of 0.025) is accompanied by a decrease in the Q (§ 2-10, 2-12) of the antenna, so that the resonance curve is less sharp. Hence the antenna is capable of working over a wider frequency range. The effect is greater as the diameter/length ratio is increased, and is a property of some importance at ultra-high frequencies where the wavelength is small.

Radiation characteristics — The radiation from a half-wave antenna is not uniform in all directions but varies with the angle with respect to the axis of the wire. It is most intense in directions at right-angles to the wire, and zero along the direction of the wire itself, with intermediate values at intermediate angles. This is shown by the sketch of Fig. 1002, which represents the radiation pattern in free space. The relative intensity of radiation is proportional to the length of a line drawn from the center of the figure to the perimeter. If the antenna is vertical, as shown in the figure, then the field strength (§ 9-1) will be uniform in all horizontal directions; if the antenna is horizontal, the relative field strength will depend upon the direction of the receiving point with respect to the direction of the antenna wire.

●10-3 GROUND EFFECTS

Reflection — When the antenna is near the ground the free-space pattern of Fig. 1002 is modified by reflection of radiated waves from the ground, so that the actual pattern is the resultant of the free-space pattern and ground reflections. This resultant is dependent upon the height of the antenna, its position or orientation with respect to the surface of the ground, and the electrical characteristics of the ground. The reflected waves may be in such phase relationship to the directly-radiated waves that the two completely reinforce each other, or the phase relationship may be such that complete cancellation takes place. All intermediate values also are possible. Thus the effect of a perfectly-reflecting ground is such that the original free-space field strength may be multiplied by a factor which has a maximum value of 2, for complete reinforcement, and having all intermediate values to zero, for complete cancellation. Since waves are always reflected upward from the ground (assuming

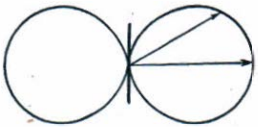


Fig. 1002 — Free-space radiation pattern of half-wave antenna. The antenna is shown in the vertical position. This is a cross-section of the solid pattern described by the figure when rotated on its axis (the antenna). The "doughnut" form of the solid pattern can easily be visualized by imagining the drawing glued to cardboard, with a short length of wire fastened on to represent the antenna. Then twirling the wire will give a visual representation of the solid pattern.

that the surface is fairly level) these reflections only affect the radiation pattern in the vertical plane — that is, in directions upward from the earth's surface — and not in the horizontal plane, or the usual geographical directions.

Fig. 1003 shows how the multiplying factor varies with the vertical angle for several representative heights for horizontal antennas. As the height is increased the angle at which complete reinforcement takes place is lowered until it occurs at a vertical angle of 15 degrees for a height equal to one wavelength. At still greater heights not shown on the chart the first maximum will occur at still smaller angles.

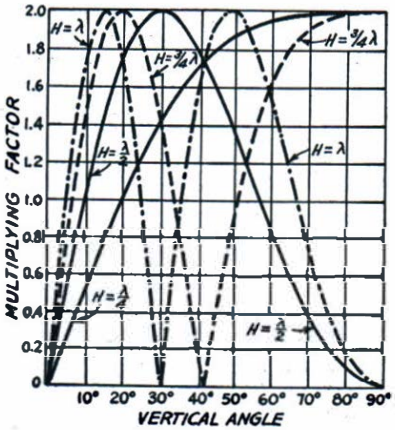


Fig. 1003 — Effect of ground on radiation at vertical angles for four antenna heights. This chart applies only to horizontal antennas, and is based on perfectly conducting ground.

When the half-wave antenna is vertical the maximum and minimum points in the curves of Fig. 1003 exchange positions, so that the nulls become maxima, and vice versa. In this case, the height is taken as the distance from ground to the center of the antenna.

Radiation angle — The vertical angle, or angle of radiation, is of primary importance, especially at the higher frequencies (§ 9-4, 9-5). It is therefore advantageous to erect the antenna at a height which will take advantage of ground reflection in such a way as to reinforce the space radiation at the most desirable angle. Since low radiation angles usually are desirable, this generally means that the antenna should be high; at least ½ wavelength at 14 Mc. and preferably ¾ or 1 wavelength; at least 1 wavelength and preferably higher at 28 Mc. and the ultra-high frequencies. The physical height decreases as the frequency is increased so that good heights are not impracticable; a half wavelength at 14 Mc. is only 35 feet, approximately, and the same height represents a full wavelength at 28 Mc. At 7 Mc. and lower, the higher radiation angles are effective so that again a reasonable antenna height is not difficult of attainment. Heights between 35 and 70 feet are suitable for all bands, the higher figures generally being preferable if circumstances permit their use.

Imperfect ground — Fig. 1003 is based on a ground having perfect conductivity, which is not met with in practice. The principal effect of actual ground is to make the curves inaccurate at the lowest angles; appreciable high-frequency radiation at angles smaller than a few degrees is practically impossible to obtain at heights of less than several wavelengths. Above 15 degrees, however, the curves are accurate enough for all practical purposes, and may be taken as indicative of the sort of result to be expected at angles between 5 and 15 degrees.

The effective ground plane — that is, the plane from which ground reflections can be considered to take place — seldom is the actual surface of the ground but is a few feet below it, depending upon the character of the soil.

Impedance — Waves which are reflected directly upward from the ground induce a current in the antenna in passing and, depending on the antenna height, the phase relationship of this induced current to the original current may be such as either to increase or decrease the total current in the antenna. For the same power input to the antenna, an increase in current is equivalent to a decrease in impedance, and vice versa. Hence the impedance of the antenna varies with height. The theoretical curve of variation of radiation resistance for an antenna above perfectly-reflecting ground is shown in Fig. 1004. The impedance approaches the free-space value as the height becomes large, but at low heights may differ considerably from it.

Choice of polarization — Polarization of the transmitting antenna is generally unimportant on frequencies between 3.5 and 30 Mc.

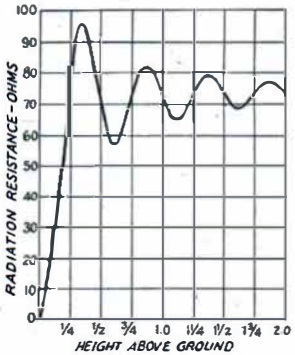


Fig. 1004 — Radiation resistance of a half-wave horizontal antenna as a function of height above perfectly-reflecting ground.

However, the question of whether the antenna should be installed in a horizontal or vertical position deserves consideration on other counts. A vertical half-wave antenna will radiate equally well in all *horizontal* directions, so that it is substantially non-directional in the usual sense of the word. If installed horizontally, however, the antenna will tend to show directional effects, and will radiate best in the direction at right-angles, or broadside, to the wire. The radiation in such a case will be least in the direction toward which the wire points. This can be seen readily by imagining that Fig. 1002 is lying on the ground and that the pattern is looked at from above.

The vertical angle of radiation also will be affected by the position of the antenna. If it were not for ground losses at high frequencies, the vertical half-wave antenna would be preferred because it would concentrate the radiation horizontally. Practically, this theoretical advantage over the horizontal antenna is of little or no consequence.

At 1.75 Mc. vertical polarization will give more low-angle radiation, and hence is better for long-distance transmission; at this frequency the ground wave also is useful and must be vertically polarized. On ultra-high frequencies, direct-ray and lower troposphere transmission require the same type of polarization at both receiver and transmitter, since the waves suffer no appreciable change in polarization in transmission (§ 9-5, 9-6). Either vertical or horizontal polarization may be used, the latter being slightly better for longer distances.

Effective radiation patterns — In determining the radiation pattern it is necessary to consider radiation in both the horizontal and vertical planes. When the half-wave antenna is vertical, the vertical angle of radiation chosen does not affect the *shape* of the horizontal pattern, but only its relative amplitude. When the antenna is horizontal, however, both the shape and amplitude are dependent upon the angle of radiation chosen.

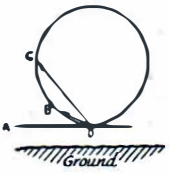


Fig. 1005 — Illustrating the importance of vertical angle of radiation in determining antenna directional effects. Ground reflection is neglected in this drawing.

Fig. 1005 illustrates this point. The "free-space" pattern of the horizontal antenna shown is a section cut vertically through the solid pattern. In the direction *OA*, horizontally along the wire axis, the radiation is zero. At some vertical angle represented by the line *OB*, however, the radiation is appreciable, despite the fact that this line runs *in the same geographical direction* as *OA*. At some higher angle *OC* the radiation, *still in the same geographical direction*, is still more intense. The effective radiation pattern therefore depends upon the angle of radiation most useful and for long-distance transmission is dependent upon the conditions existing in the ionosphere. These conditions may vary not only from day to day and hour to hour, but even from minute to minute. Obviously, then, the effective directivity of the antenna will change along with transmission conditions.

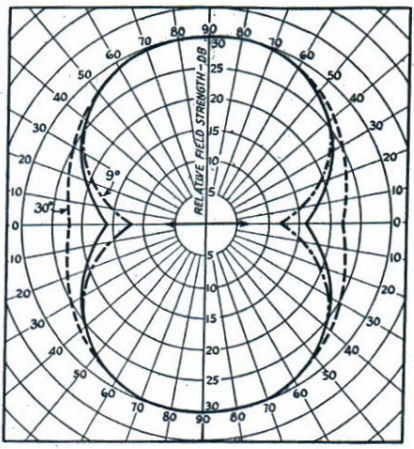


Fig. 1006 — Horizontal pattern of a horizontal half-wave antenna at three vertical radiation angles. Solid line is relative radiation at 15 degrees. Dotted lines show deviation from the 15-degree pattern, for angles of 9 and 30 degrees. The patterns are useful for shape only, since the amplitude will depend upon the height of the antenna above ground and the vertical angle considered. The patterns for all three angles have been proportioned to the same scale, but this does not mean that the maximum amplitudes necessarily are the same. The arrow indicates the direction of the antenna wire.

At ultra-high frequencies, where only extremely low angles are useful for any but sporadic-E transmission (§ 9-7) the effective radiation pattern of the antenna approaches the free-space pattern. A horizontal antenna therefore shows more marked directive effects than it does at lower frequencies, on which high radiation angles are effective.

Theoretical horizontal-directivity patterns for half-wave horizontal antennas at vertical angles of 9, 15, and 30 degrees (representing average useful angles at 28, 14 and 7 Mc. respectively) are given in Fig. 1006. At intermediate angles the values in the affected regions also will be intermediate. Relative field strengths are plotted on a decibel scale (§ 3-3) so that they represent as nearly as possible the actual aural effect at the receiving station.

•10-4 APPLYING POWER TO THE ANTENNA

Direct excitation — When power is transferred directly from the source to the radiating antenna, the antenna is said to be directly excited. While most of the coupling methods (§ 2-11) may be used, the more common ones are shown in Fig. 1007. Power is usually fed to the antenna at either a current or voltage loop (§ 10-2). If at a current loop, the coupling is called *current feed;* if at a voltage loop, it is called *voltage feed.*

Current feed — This is shown in Fig. 1007-A. The antenna is cut at the center and a small coil coupled to the output tank circuit of the transmitter, with adjustable coupling so that the transmitter loading can be controlled. Since the addition of the coil "loads" the antenna, or increases its effective length because of the additional inductance, the series condensers C_1 and C_2 are used to provide electrical means for reducing the length to its original unloaded value; in other words, to cancel the effect of the inductive reactance (§ 2-10).

Voltage feed — In Fig. 1007 at B and C the power is introduced into the antenna at a point of high voltage. In B the end of the antenna is coupled to the output tank circuit through a small condenser; in C a separate tank, connected directly to the antenna, is used. This tank is tuned to the transmitter frequency and should be grounded at one end or at the center of the coil, as shown.

Adjustment of coupling — Methods of tuning and adjustment correspond to those used with transmission lines and are discussed in § 10-6.

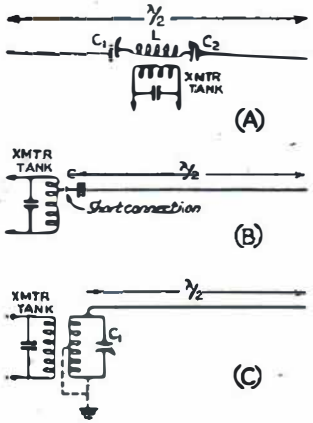


Fig. 1007 — Methods of direct feed to the half-wave antenna. *A*, current feed, series tuning; *B*, voltage feed, capacity coupling; *C*, voltage feed with inductively coupled antenna tank. In *A*, the coupling apparatus is not included in the antenna length.

Disadvantages of direct excitation — Direct excitation is seldom used except on the lowest amateur frequencies because it involves bringing the antenna proper into the operating room and hence into close relationship with the house and electric wiring. This usually means that some of the power is wasted in heating poor conductors in the field of the antenna. Also, it usually means that the shape of the antenna must be distorted so that the expected directional effects are not realized, and likewise means that the height is limited. For these reasons, in high-frequency work practically all amateurs use transmission lines or feeder systems which permit putting the antenna in a desirable location.

•10-5 TRANSMISSION LINES

Requirements — A transmission line is used to transfer power, with a minimum of loss, from its source to the device in which the power is to be usefully expended. At radio frequencies, where every wire carrying r.f. current tends to radiate energy in the form of electromagnetic waves, special design is necessary to minimize radiation and thus cause as much as possible of the power to be delivered to the receiving end of the line.

Radiation can be minimized by using a line in which the current is low, and by using two conductors carrying currents of equal magnitudes but opposite phase so that the fields about the conductors cancel each other. For good cancellation of radiation the two conductors should be parallel and quite close to each other.

Types — The most common form of transmission line consists of two parallel wires, maintained at a fixed spacing of two to six inches by insulating spacers or spreaders at suitable intervals (*open-wire line*). A second type consists of rubber-insulated wires twisted together to form a flexible line without spacers (*twisted-pair line*). A third uses a wire inside and coaxial with a tubing outer conductor, separated from the outer conductor by insulating spacers or "beads" at regular intervals (*coaxial* or *concentric line*). A variation of this type uses solid rubber insulation between the inner and outer conductors, the latter usually being made of metal braid rather than solid tubing so that the line will be flexible. Still another type of line uses a single wire alone, without a second conductor (*single-wire feeder*); in this case radiation is minimized by keeping the line current low.

Spacing of two-wire lines — The spacing between wires of an open-wire line should be small in comparison to the operating wavelength to prevent appreciable radiation. At the same time it is impracticable to make the spacing too small because when the wires

swing with respect to each other in a wind the line constants (§ 2-12) will vary and thus cause a variation in tuning or loading on the transmitter. It is also desirable to use as few insulating spacers as possible to keep the weight of the line to a minimum. In practice a spacing of about six inches is used for 14 Mc. and lower frequencies, with four and two inch spacing being common on the ultra-high frequencies.

Balance to ground — For maximum cancellation of the fields about the two wires it is necessary that the currents be equal in amplitude and opposite in phase. Should the capacity or inductance per unit length in one wire differ from that in the other this condition cannot be fulfilled. Insofar as the line itself is concerned, the two wires will have identical characteristics only when the two have exactly the same physical relationships to ground and to other objects in the vicinity. Thus the line should be symmetrically constructed and the two wires should be at the same height. Line unbalance can be minimized by keeping the line as far above ground and as far from other objects as possible.

To overcome unbalance the line is sometimes transposed, which means that the positions of the wires are interchanged at regular intervals (Fig. 1008). This is more helpful on long than



Fig. 1008 — Method of transposing a two-wire open transmission line to preserve balance to ground and nearby objects.

on short lines, and usually need not be resorted to for lines less than a wavelength or two long.

Characteristic impedance — The square root of the ratio of inductance to capacity per unit length of the line is called the *characteristic* or *surge impedance*. It is the impedance which a long line would present to an electrical impulse induced in the line, and is important in determining the operation of the line in conjunction with the apparatus to which it is connected.

The characteristic impedance of air-insulated transmission lines may be calculated from the following formulas:

Parallel-conductor line

$$Z = 276 \log \frac{b}{a} \qquad (5)$$

where Z is the surge impedance, b the spacing, center to center, and a the radius of the conductor. The quantities b and a must be measured in the same units (inches, etc.). Surge impedance as a function of spacing for lines using conductors of different size is plotted in chart form in Fig. 1009.

Coaxial or concentric line

$$Z = 138 \log \frac{b}{a} \qquad (6)$$

where Z again is the surge impedance. In this case b is the *inside diameter* (not radius) of the outer conductor and a is the *outside diameter* of the inner conductor. The formula is true for air dielectric, and approximately so for a line having ceramic insulators so spaced that the major proportion of the insulation is air.

When a solid insulating material is used between the conductors the impedance decreases, because of the increase in line capacity, by the factor $1/\sqrt{k}$, where k is the dielectric constant of the insulating material.

The impedance of a single-wire transmission line varies with the size of the conductor, its

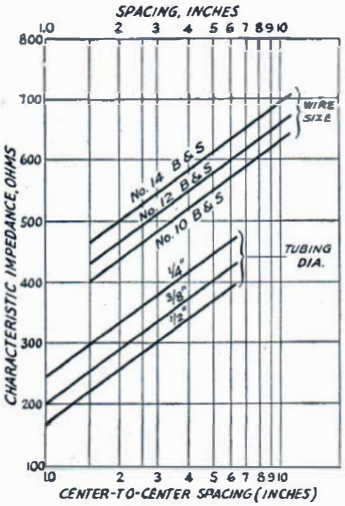


Fig. 1009 — Characteristic impedances of typical spaced-conductor transmission lines. Use outside diameter of tubing.

height above ground, and orientation with respect to ground. An average figure is about 500 ohms.

Electrical length — The electrical length of a line is not exactly the same as its physical length for reasons corresponding to the end effects in antennas. (§ 10-2). Spacers used to separate the conductors have dielectric constants larger than that of air, so that the waves do not travel quite as fast along a line as they would in air. The lengths of electrical quarter waves of various types of lines can be calculated from the formula

$$\text{Length (feet)} = \frac{246 \times V}{\text{Freq. (Mc.)}} \qquad (7)$$

where V depends upon the type of line. For lines of ordinary construction, V is as follows:

Parallel wire line	$V = 0.975$
Parallel tubing line	$V = 0.95$
Concentric line (air-insulated)	$V = 0.85$
Concentric line (rubber-insulated)	$V = 0.56$–0.65
Twisted pair	$V = 0.56$–0.65

Input and output ends — The input end of a line is that connected to the source of power; the output end is that connected to the power-absorbing device. When a line connects a transmitter to an antenna, the input end is at the transmitter and the output end at the antenna; with the same line and antenna connected to a receiver, however, the energy flows from the antenna to the receiver, hence the input end of the line is at the antenna and the output end at the receiver.

Standing-wave ratio — The lengths of transmission lines used at radio frequencies are of the same order as the operating wave lengths and therefore standing waves of current and voltage may appear on the line (§ 2-12). The ratio of current (or voltage) at a loop to the value at a node (*standing-wave ratio*) depends upon the ratio of the resistance of the load connected to the output end of the line, or termination, to the characteristic impedance of the line itself. That is,

$$\text{Standing-wave ratio} = \frac{Z_o}{Z_t} \text{ or } \frac{Z_t}{Z_o} \qquad (8)$$

where Z_o is the characteristic impedance of the line and Z_t is the terminating resistance, Z_t is generally called an impedance, although it must be non-reactive and therefore correspond to a pure resistance for the line to operate as described. This means that the load or termination, when an antenna, must be resonant at the operating frequency.

The formula is given in two ways because it is customary to put the larger number in the numerator so that the ratio will not be fractional. As an example, a 600-ohm line terminated in a resistance of 70 ohms will have a standing wave ratio of 600/70, or 8.57. The ratio on a 70-ohm line terminated in a resistance of 600 ohms would be the same. This means that if the current as measured at a node is 0.1 amp., the current at a loop will be 0.857 amp.

A line terminated in a resistance equal to its characteristic impedance is equivalent to an infinitely long line, consequently there is no reflection and no standing waves appear. The standing wave ratio therefore is 1. The input end of such a line appears as a pure resistance of a value equal to the characteristic impedance of the line.

Reactance, resistance, impedance — The input end of a line may show reactance as well as resistance, and the values of these quantities will depend upon the nature of the load at the output end, the electrical length of the line, and the line characteristic impedance. The reactance and resistance are important in determining the method of coupling to the source of power. Assuming that the load at the output end of the line is purely resistive, which is essentially the case since the load circuit is usually tuned to resonance, a line less than a quarter wavelength long electrically will show inductive reactance at its input terminals when the output termination is *less* than the characteristic impedance, and capacitive reactance when the termination is *higher* than the characteristic impedance. If the line is more than a quarter wave but less than a halfwave long the reverse conditions exist. With still longer lengths the reactance characteristics reverse in each succeeding quarter wavelength. The input impedance is purely resistive if the line is an exact multiple of a quarter wave in length. The reactance at intermediate lengths is higher the greater the standing wave ratio, being zero for a ratio of 1.

Impedance transformation — Regardless of the standing-wave ratio, the input impedance of a line a half-wave long electrically will be equal to the impedance connected at its output end. Such a line (the same thing is true of a line any integral multiple of a half-wave in length) can be considered to be a one-to-one transformer. However, if the line is a quarter-wave (or an odd multiple of a quarter wave) long the input impedance will be equal to

$$Z_i = \frac{Z_o^2}{Z_t}$$

where Z_o is the characteristic impedance of the line and Z_t the impedance connected to the output end. A quarter-wave line therefore can be used as an *impedance transformer*, and by suitable selection of constants a wide range of input impedance values can be obtained. Furthermore, the impedance measured between the two conductors anywhere along the line will vary between the two end values, so that any intermediate impedance value can be selected. This is a particularly useful property since a quarter-wave line may be short-circuited at one end (§ 2-12) and used as a *linear transformer* with adjustable impedance ratio.

Losses — Air-insulated lines operate at quite high efficiency. Parallel-conductor lines average 0.12 to 0.15 db. (§ 3-3) loss per wavelength of line. These figures hold only if the

standing wave ratio is 1. The losses increase with the standing-wave ratio, rather slowly up to a ratio of 15 to 1, but rapidly thereafter. For standing-wave ratios of 10 or 15 to 1 the increase is inconsequential provided the line is well balanced.

Concentric lines with air insulation are excellent when dry, but losses increase if there is moisture in the line. Provision therefore should be made for making such lines airtight, and they should be thoroughly dry when assembled. This type of line has the least radiation loss. The small lines (⅜-inch outer conductor) should not be used at high voltages, hence it is desirable to keep the standing-wave ratio down.

Good quality rubber insulated lines, both twisted pair and coaxial, average about 1 db. loss per wavelength of line. At the higher frequencies, therefore, such lines should be used only in short lengths if losses are important. These lines have the advantages of compactness, ease of installation, and flexibility. Ordinary lampcord has a loss of approximately 1.4 db. per wavelength, when dry, but its losses become excessive when wet. The parallel moulded-rubber type is best from the standpoint of withstanding wet weather. The characteristic impedance of lampcord is between 120 and 140 ohms.

The loss in db. is directly proportional to the length of the line. Thus a line which has a loss of 1 db. per wavelength will have an actual loss of 3 db. if the line is three wavelengths long. In the case of line losses, the length is not expressed in terms of electrical length but in physical length; that is, a wavelength of line, in feet, is equal to 984/Freq. (Mc.) for computing loss. This permits a direct comparison of lines having the same physical length. The electrical lengths, of course, may differ considerably.

Resonant and non-resonant lines — Lines are classified as *resonant* or *non-resonant* depending upon the standing-wave ratio. If the ratio is near 1, the line is said to be non-resonant. Reactive effects will be small and consequently no special tuning provisions need be made for cancelling them (§ 2-10) even when the line length is not an exact multiple of a quarter wavelength. If the standing-wave ratio is fairly large, the input reactance must be cancelled or "tuned out" unless the line is a multiple of a quarter wavelength, and the line is said to be resonant.

• 10-6 COUPLING TO TRANSMISSION LINES

Requirements — The coupling system between a transmitter and the input end of a transmission line must provide means for adjusting the load on the transmitter to the proper value (impedance matching) and for tuning out any reactive component that may be present (2-9, 2-10, 2-11). The resistance and reactance considered are those present *at the input end of the line*, and hence have nothing to do with the antenna itself except insofar as the antenna load may affect the operation of the line (§ 10-5).

Untuned coil — A simple system, shown in Fig. 1010-A, uses a coil of a few turns tightly coupled to the plate tank coil. Since no provision is made for tuning, this system is suitable only for non-resonant lines which show practically no reactance at the input end. Loading on the transmitter may be varied by varying the coupling between the tank inductance and *pickup coil*, as it is frequently called, or by changing the number of turns on the pickup coil. A slight amount of reactance is coupled into the tank circuit by the pickup coil, since the flux leakage (§ 2-11) is high, so that slight retuning of the plate tank condenser may be necessary when the load is connected.

Taps on tank circuit — A method suitable for use with open-wire lines is shown in Fig. 1010-B, where the line is tapped on a balanced tank circuit with taps equidistant from the center or ground point. This symmetry is necessary to maintain line balance to ground (§ 10-5). Loading is increased by moving the taps outward from the center. Any reactance present may be tuned out by readjustment of the plate tank condenser, but this method is not suitable for large values of reactance and therefore direct tapping is best confined to use with non-resonant lines.

Adjustment of untuned systems — Adjustment of either of the above systems is quite simple. Starting with loose coupling, apply power to the transmitter, and adjust the plate tank condenser to minimum plate current. If the current is less than the desired load value, increase the coupling and again resonate the plate condenser. Continue until the desired plate current is obtained, always keeping the plate tank condenser at the setting which gives minimum current.

Pi-section coupling — A coupling system which is electrically equivalent to tapping on the tank circuit, but using a capacity voltage divider in the plate tank circuit for the purpose, is shown in Fig. 1010-C. Since one side of the condenser across which the line is connected is grounded, some unbalance will be introduced into the transmission line. The method is used chiefly with low-power portable sets because it is readily adjustable to meet a fairly wide range of impedance values. A single-ended amplifier, either screen-grid or a grid-neutralized triode (§ 4-7) is required, since the plate tank circuit is not balanced. Coupling is adjusted by varying C_1, re-resonat-

ing the circuit each time by means of C_2, until the desired amplifier plate current is obtained. In general, the coupling will increase as C_1 is made smaller with respect to C_2. Relatively large-capacity condensers are required to give a suitable impedance-matching range while maintaining resonance.

Pi-section filter — The coupling circuit shown in Fig. 1010-D is a low-pass filter capable of coupling between a fairly wide range of impedances. The method of adjustment is as follows: First, with the filter disconnected from the transmitter tank, tune the transmitter tank to resonance, as evidenced by minimum plate current. Then, with trial settings of the clips on L_1 and L_2 (few turns for high frequencies, more for lower) tap the input clips on the final tank coil at points equidistant from the center so that about half the coil is included between them. A balanced tank circuit must be used. Set C_2 at about half scale, apply power, and rapidly rotate C_1 until the plate current drops to minimum. If this minimum is not the desired full-load plate current, try a new setting of C_2 and repeat. If, for all settings of C_2, the plate current is too high or too low, try new settings of the taps on L_1 and L_2, and also on the transmitter tank. Do not touch the tank condenser during these adjustments. When, finally, the desired plate current is obtained,

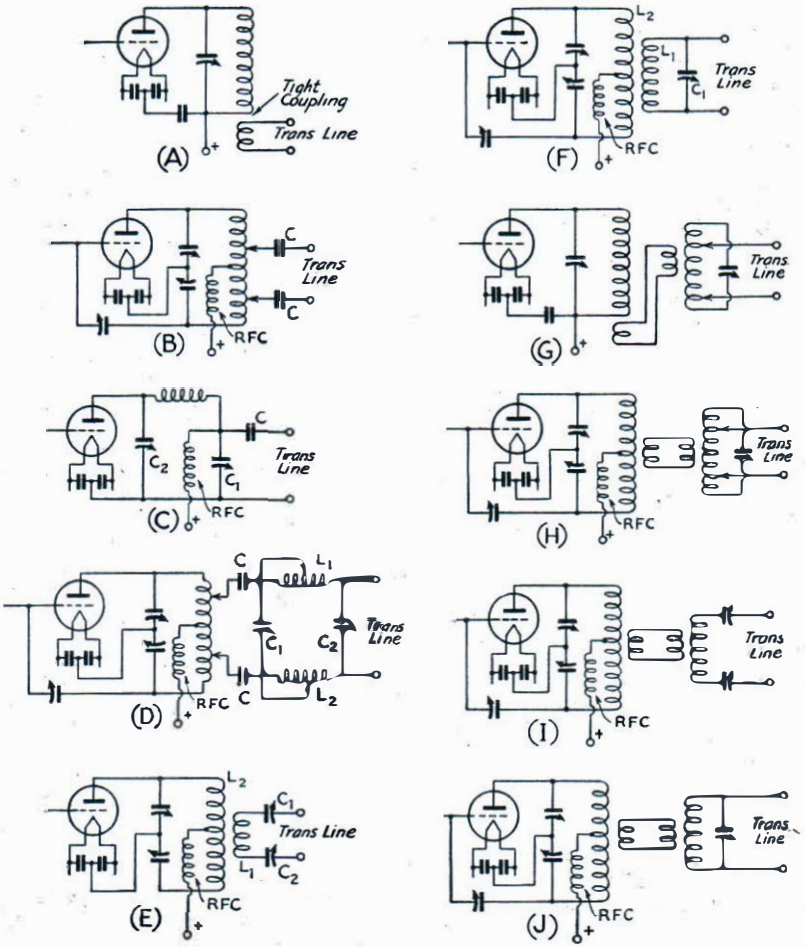


Fig 1010 — Methods of coupling the transmitter to the transmission line. Circuit values and adjustment are discussed in the text. Condensers marked "C" are fixed blocking condensers to isolate the transmitter plate voltage from the antenna; their capacity is not critical, 500 $\mu\mu$fd. to 0.002 μfd. being satisfactory values. Voltage rating should at least equal the plate voltage on the final stage.

set C_1 carefully to the exact minimum plate-current point. *This adjustment is important in minimizing harmonic output.*

With some lengths of resonant lines, particularly those not exact multiples of a quarter wavelength, it may be difficult to get proper loading with the pi-section coupler. Usually, these lengths also will be difficult to feed with other systems of coupling. In such cases, the proper loading often can be obtained by varying the L/C ratio of the filter over a considerably wider range than is used for normal loads.

Series tuning — When the input impedance of the line is low, the coupling method shown in Fig. 1010-E may be used. This system, known as *series tuning*, places the coupling coil, tuning condensers, and load all in series, and is particularly suitable for use with resonant lines when a current loop appears at the input end. Two tuning condensers, as shown, usually are used to keep the line balanced to ground, but one will suffice, the other end of the line being connected directly to the end of L_1.

The tuning procedure with series tuning is as follows: With C_1 and C_2 at minimum capacity, couple the antenna coil L_1 loosely to the transmitter output tank coil and observe the plate current. Then increase C_1 and C_2 simultaneously, until a setting is reached which gives maximum plate current, indicating that the antenna system is in resonance with the transmitting frequency. Readjust the plate tank condenser to minimum plate current. This is necessary because tuning the antenna circuit will have some effect on the tuning of the plate tank. The new minimum plate current will be higher than with the antenna system detuned, but should still be well below the rated value for the tube or tubes. Increase the coupling between L_1 and L_2 by a small amount, readjust C_1 and C_2 for maximum plate current, and again set the plate tank condenser to minimum. Continue this process until the minimum plate current is equal to the rated plate current for the amplifier. Always use the degree of coupling between L_1 and L_2 which will just bring the amplifier plate current to rated value when C_1 and C_2 pass through resonance. The r.f. ammeters should indicate maximum feeder current at the resonance setting; these meters are not strictly necessary, but are useful in indicating the relative power output from the transmitter.

Parallel tuning — When the line has high input impedance, parallel tuning, as shown in Fig. 1010-F, is required. Here the coupling coil, tuning condenser and line are all in parallel, the load represented by the line being directly across the tuned coupling circuit. If the line is non-reactive, the coupling circuit will be tuned independently to the transmitter frequency; line reactance can be compensated for by tuning of C_1 and adjustment, if necessary, of L_1 by means of taps. Parallel tuning is suited to resonant lines when a voltage loop appears at the input end.

The tuning procedure is quite similar to that with series tuning. Find the value of coupling between L_1 and L_2 which will bring the plate current to the desired value as C_1 is tuned through resonance. Again a slight readjustment of the amplifier tank condenser may be necessary to compensate for the effect of coupled reactance.

Link coupling — Where tuning of the circuit connected to the line is necessary or desirable, it is possible to separate physically the line-tuning apparatus and the plate tank circuit by means of link coupling (§ 2-11). This is often convenient from a constructional standpoint, and has the advantage that with proper construction there will be somewhat less harmonic transfer to the antenna since stray capacity coupling is lessened with the smaller link coils.

Figs. 1010-G and H show a method which can be considered to be a variation of Fig. 1010-B. The first (G) is suitable for use with a single-ended plate tank, the second (H) for a balanced tank. The auxiliary tank on which the transmission line is tapped may have adjustable inductance as well as capacity to provide a wide range of reactance variation for compensating for line reactance. The center of the auxiliary tank inductance may be grounded if desired. The link windings should be placed at the grounded parts of the coils to reduce capacity coupling and consequent harmonic transfer. With this inductively-coupled system the loading on the auxiliary tank circuit increases as the taps are moved outward from the center, but since this decreases the Q of the circuit the coupling to the plate tank simultaneously decreases (§ 2-11), hence a compromise adjustment giving proper loading inust be found in practice. Loading also may be varied by changing the coupling between one link winding and its associated tank coil; either tank may be used for this purpose. When the auxiliary tank is properly tuned to compensate for line reactance the plate tank tuning will be practically the same as with no load on the circuit, hence the plate tank condenser need only be readjusted slightly to compensate for the small reactance introduced by the link.

Link coupling also may be used with series and parallel tuning, as shown in Figs. 1010-I and J. The coupling between one link and its associated coil may be made variable to give the same effect as changing the coupling between the plate tank and antenna coils in the ordinary system. The tuning procedure is the same as described above for series and parallel tuning. In the case of single-ended tank cir-

cuits, the input link would be coupled to the grounded end of the tank coil, similarly to the method in Fig. 1010-G.

Circuit values — The values of inductance and capacity to use in the antenna coupling system will depend upon the transmitting frequency, but are not particularly critical. With series tuning (Fig. 1010-E, I) the coil may consist of a few turns of the same construction as is used in the final tank; average values will run from one or two turns at ultra-high frequencies to perhaps 10 or 12 at 1.75 Mc. The number of turns preferably should be adjustable so that the inductance can be changed should it not be possible to reach resonance with the condensers used. The series condensers should have a maximum capacity of 250 or 350 $\mu\mu$fd. at the lower frequencies; the same values will serve even at 28 Mc., although 100 $\mu\mu$fd. will be ample for this and the 14-Mc. band. Still smaller condensers can be used at ultra-high frequencies. Since series tuning is used at a low-voltage point in the feeder system, the plate spacing of the condensers does not have to be large. Ordinary receiving-type condensers are large enough for plate voltages up to 1000, and the smaller transmitting condensers have high-enough voltage ratings for higher-power applications. With high-power 'phone it may be necessary to use condensers having a plate spacing of approximately 0.15 to 0.2 inch.

In parallel-tuned circuits (F, G, H, J) the antenna coil and condenser should be approximately the same as those used in the final tank circuit. The antenna tank circuit must be capable of being tuned independently to the transmitting frequency, and if possible provision should be made for tapping the coil so that the L/C ratio can be varied to the optimum value (§ 2-11) as determined experimentally.

In Fig. 1010-D, C_1 and C_2 may be 100 to 250 $\mu\mu$fd. each, the higher-capacity values being used for lower-frequency operation (3.5 and 1.75 Mc.). Plate spacing should in general be at least half that of the final amplifier tank condenser. For operation from 1.75 to 14 Mc., L_1 and L_2 each should be 15 turns 2½ inches in diameter, spaced to occupy 3 inches length, and tapped every three turns. Approximate settings are 15 turns for 1.75 Mc., 9 turns for 3.5 Mc., 6 turns for 7 Mc., and 3 turns for 14 Mc. The coils may be wound with No. 14 or No. 12 wire. This method of coupling is very seldom used at ultra-high frequencies.

Harmonic reduction — It is important to prevent, insofar as possible, harmonics in the output of the transmitter from being transferred to the antenna system. Untuned (Fig. 1010-A) and directly-coupled (Figs. 1010-B) systems do not discriminate against harmonics, and hence are more likely to cause harmonic radiation than the inductively-coupled tuned systems. Low-pass filter arrangements such as those at C and D, Fig. 1010, do discriminate against harmonics, but the direct coupling frequently is a source of trouble in this respect.

In inductively-coupled systems, care must be taken to prevent capacity coupling between coils. Link coils should always be coupled at a point of ground potential (§ 2-13) on the plate tank coil, and so should series and parallel-tuned coils (E and F) when possible. Capacity coupling can be practically eliminated by the use of a Faraday shield (§ 4-9) between the two coils.

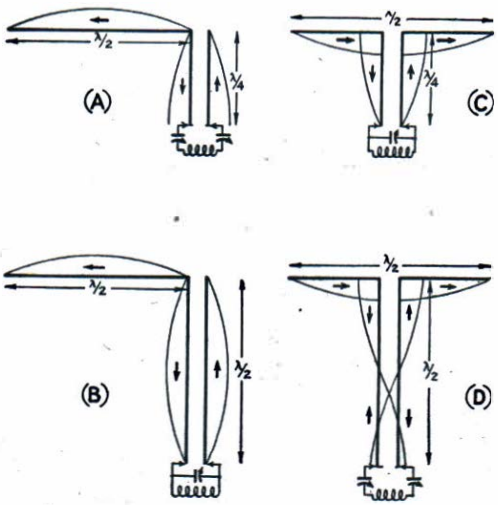


Fig. 1011 — Half-wave antennas fed from resonant lines. *A* and *B*, end feed with quarter- and half-wave lines; *C* and *D*, center feed. The current distribution is shown for all four cases. Arrows indicate instantaneous direction of current flow.

● 10-7 RESONANT LINES

Two-wire lines — Because of its simplicity of adjustment and flexibility with respect to the frequency range over which an antenna system will operate, the resonant line is widely used with simple antenna systems. Constructionally, the spaced or "open" two-wire line is best suited to resonant operation; rubber-insulated lines such as twisted pair will have excessive losses when operated with standing waves.

Connection to antenna — A resonant line is usually — in fact, practically always — connected to the antenna at either a current or voltage loop. This is advantageous, especially when the antenna is to be operated at harmonic frequencies, since it simplifies the problem of determining the coupling system to be used at the input end of the transmission line.

Half-wave antenna with resonant line — It is often helpful to look upon the resonant

line simply as an antenna folded back on itself. Such a line may be any whole-number multiple of a quarter-wave in length; in other words any total wire length which will accommodate a whole number of standing waves. (The "length," however, of a two-wire line is always taken as the length of one of the wires.)

Quarter- and half-wave resonant lines feeding half-wave antennas are shown in Fig. 1011. The current distribution on both antenna and line is indicated. It will be noted that the quarter-wave line has maximum current at one end and minimum current at the other, determined by the point of connection to the antenna. The half-wave line, however, has the same current (and voltage) values at both ends.

If a quarter-wave line is connected to the end of an antenna as shown in Fig. 1011-A, then at the transmitter end of the line the current is high and the voltage low (low impedance) so that series tuning (§ 10–6) can be used. Should the line be a half-wave long, as at 1011-B, current will be minimum and voltage maximum (high impedance) at the transmitter end of the line, just as it is at the end of the antenna. Parallel tuning therefore is required (§ 10-6). The line could be coupled to a balanced final tank through small condensers, as in Fig. 1010-B, but the inductively-coupled circuit is preferable. An end-fed antenna with resonant feeders, as in 1011-A and B, is known as the "Zeppelin," or "Zepp," antenna.

The line also may be inserted at the center of the antenna at the maximum-current point. Quarter- and half-wave lines used in this way are shown at Fig. 1011-C and D. In C, the antenna end of the line is at a high-current, low-voltage point (§ 10-2), hence at the transmitter end the current is low and the voltage high. Parallel tuning therefore is used. The half-wave line at D has high current and low voltage at both ends, so that series tuning is used at the transmitter end.

The four arrangements shown in Fig. 1011 are thoroughly useful antenna systems, and are shown in more practical form in Fig. 1012. In each case the antenna is a half wavelength long, the exact length being calculated from Equations 2, 3 or 4 (§ 10-2) or taken from the charts of Fig. 1015. The line length should be an integral multiple of a quarter wavelength, and may be calculated from Equation 5 (§ 10-5) the result being multiplied by any whole number which gives a total length convenient for reaching from the antenna to the transmitter. If there is an *odd* number of quarter waves on the line in the case of the end-fed antenna,

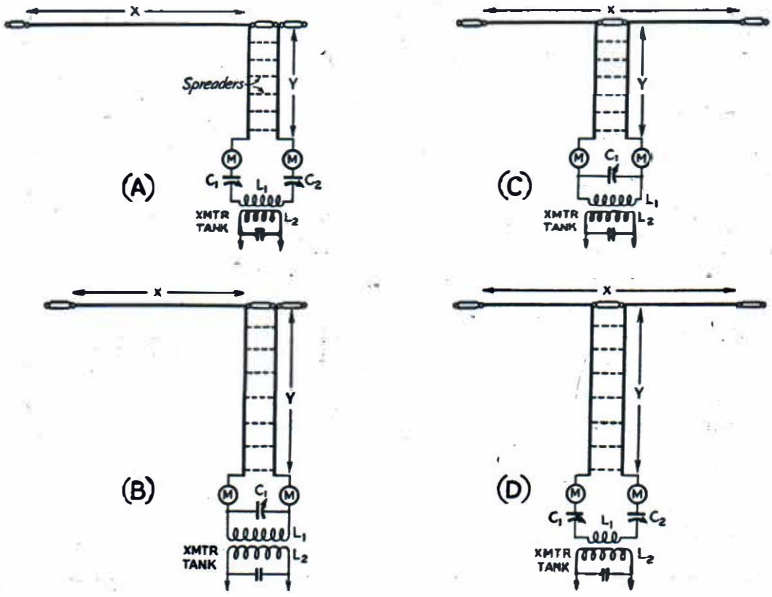


Fig. 1012 — Practical half-wave antenna systems using resonant-line feed. In the center-feed systems, the antenna length "X" does not include the length of the insulator at the center. Line length is measured from the antenna to the tuning apparatus; leads in the latter should be short enough to be neglected. The two meters shown are helpful for balancing feeder currents; however, one is sufficient for tuning for maximum output, and may be transferred from one feeder to the other, if desired. The systems at (A) and (C) are for feeders an odd number of quarter-waves in length; (B) and (D) for feeders a multiple of a half wavelength. The drawings correspond electrically to those of Fig. 1011.

series tuning will be used at the transmitter end; if an *even* number of quarter waves, then parallel tuning is used. With the center-fed antenna the reverse is true.

Practical line lengths — In general, it is best to use line lengths that are integral multiples of a quarter wavelength. Intermediate lengths will give intermediate impedance values and will show reactance as well (§ 10-5). The tuning apparatus is capable of compensating for reactance but it may be difficult to get suitable transmitter loading because simple series and parallel tuning are suitable for only low and high impedances, respectively, and neither will perform well with impedances of the order of a few hundred ohms. Such values of impedance may reduce the *Q* of the coupling circuit to such a point that adequate coupling cannot be obtained (§ 2-11). However, some departure from the ideal length is possible — even as much as 25% of a quarter wave in many cases — without undue difficulty in tuning and coupling. In such cases the type of tuning to use, series or parallel, will depend on whether the feeder length is nearer an odd number of quarter waves or nearer an even number, as well as on the point at which the feeder is connected to the antenna.

Line current — The feeder current as read by the r.f. ammeters is useful for tuning purposes only; the absolute value is of little importance. When series tuning is used the current will be high, but very little current will be indicated in a parallel-tuned system. This is because of the current distribution on the feeders as shown by Fig. 1011. With a given antenna and tuning system, of course, the greatest power will be delivered to the antenna when the readings are highest. However, should the feeder length be changed no useful conclusions can be drawn from comparison between the new and old readings. For this reason any indicator which registers the relative intensity of r.f. current can be used for tuning purposes. Many amateurs, in fact, use flashlight or dial lamps for this purpose instead of meters. They are inexpensive, and when shunted by short lengths of wire so that considerable current can be passed without burnout will serve very well even with high-power transmitters.

Antenna length and line operation — Insofar as the operation of the antenna itself is concerned, departures of a few per cent from the exact length for resonance are of negligible consequence. Such inaccuracies may influence the behavior of the feeder system, however, and as a result may have an adverse effect on the operation of the system as a whole. This is true of the end-fed antennas such as are shown in Fig. 1012-A and -B.

For example, Fig. 1013-A shows the current distribution on the half-wave antenna and quarter-wave feeder when the antenna length is correct. At the junction of the "live" feeder and the antenna the current is minimum so that the currents in the two feeder wires are equal at all corresponding points along their length. When the antenna is too long, as in B, the current minimum occurs at a point on the antenna proper, so that at the top of the live feeder there is already appreciable current flowing, whereas at the top of the "dead" feeder the current must be zero. As a result, the feeder currents are not balanced and some power will be radiated from the line. In C the antenna is too short, bringing the current minimum to a point on the live feeder, so that again the currents are unbalanced. The more serious the unbalance the greater the radiation from the line.

Strictly speaking, a line having an unbalanced connection such as the one-way termination at the end of an antenna cannot be truly

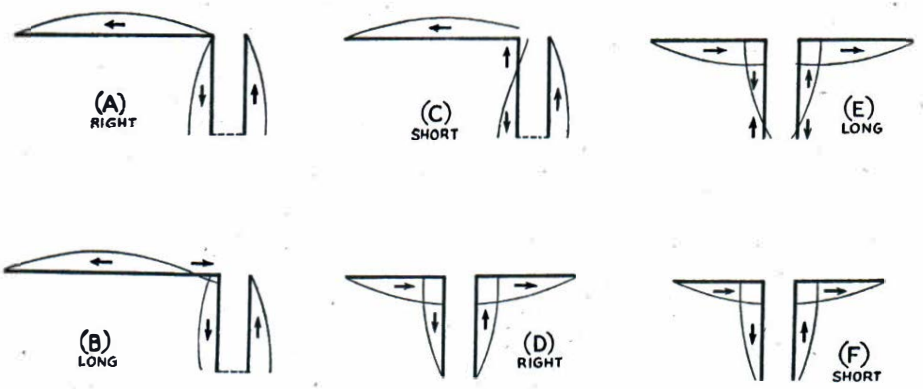


Fig. 1013 — Effect on feeder balance of incorrect antenna length. With center feed, incorrect antenna length does not unbalance the transmission line, as it does with end feed.

balanced even though the antenna length is correct. This is because of the difference in loading on the two sides. The effect is fairly small, however, when the currents are balanced.

If the antenna is fed at the center the undesirable effects of incorrect antenna length balance out so that the line operates properly under all conditions. This is shown in Fig. 1013 at D, E and F. So long as the two halves of the antenna are of equal length, the distribution of current on the feeders will be symmetrical so that no unbalance exists, even for antenna lengths considerably removed from the correct value.

• 10-8 NON-RESONANT LINES

Requirements — The advantages of non-resonant transmission lines — minimum losses, and elimination of the necessity for tuning — make this type of operation attractive. The chief disadvantage of the non-resonant line, aside from the necessity for more care in initial adjustment, is that when "matched" to the ordinary antenna it is matched only for one frequency, or at most for a small band of frequencies on either side of the frequency for which the matching is done. Except for a few special systems, this means that the antenna is unsuitable for work on more than one amateur band.

Adjustment of a non-resonant line is simply that of adjusting the terminating resistance to match the characteristic impedance of the line. To accomplish this, the antenna itself must be resonant at the selected frequency, and the line must then be connected to it in such a way that the antenna impedance as looked at by the line is the right value. The matching may be done by connecting the line at the proper spot along the antenna, by inserting an impedance transforming device between the antenna and line, or by using a line having an impedance equal to the center impedance of the antenna.

In the following discussion of ways in which different types of lines may be matched to the antenna, a half-wave antenna is used as an example. Other types of antennas may be treated by the same methods, making due allowance for the order of impedance that appears at the end of the line with more elaborate systems.

Single-wire feed — In the single-wire-feed system the return circuit is through the ground. There will be no standing waves on the feeder when its characteristic impedance is matched by the impedance of the antenna at the connection point. The principal dimensions are the length of the antenna *L*, Fig. 1014, and the distance *D* from the exact center of the antenna to the point at which the feeder is attached. Approximate dimensions

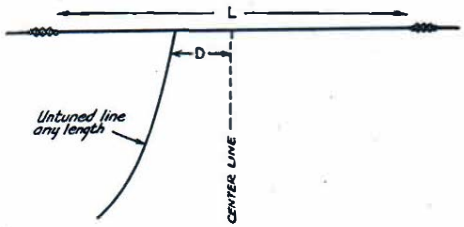


Fig. 1014 — Single-wire-feed system. The length L (one-half wavelength) and D are determined from the chart, Fig. 1015.

can be obtained from Fig. 1015 for an antenna system having a fundamental frequency in the most used amateur bands.

In constructing an antenna system of this type the feeder must run straight away from the antenna (at a right angle) for a distance of at least one-third the length of the antenna. Otherwise the field of the antenna will affect the feeder and cause faulty operation. There should be no sharp bends in the feeder wire at any point.

With the coupling system shown in Fig. 1016-A, adjustment is as follows: Starting at the ground end of the tank coil, the tap is moved towards the plate end until the amplifier draws the rated amount of plate current. The plate tank condenser should be readjusted each time the tap is changed, to bring the plate current to minimum. The amplifier is loaded properly when this "minimum" is the rated current. The condenser in the feeder is for the purpose of insulating the antenna system from the high-voltage plate supply when series plate feed is used. It should have a voltage rating somewhat above that of the plate supply. Almost any capacity greater than 500 $\mu\mu$fd. will be satisfactory. The condenser is unnecessary, of course, if parallel plate feed is used.

Inductive coupling to the output circuit is shown in Fig. 1016-B. The antenna tank circuit should tune to resonance at the operating frequency and the loading is adjusted by varying the coupling between the two tanks, both being kept tuned to resonance.

Regardless of the type of coupling, a good ground connection is essential with this system. Single-wire feed works best over moist ground, and poorly over rock and sand.

Twisted-pair feed — A two-wire line composed of twisted rubber-covered wire can be constructed to have an impedance approximately equal to that at the center of the antenna itself, thus permitting the method of connecting the line to the antenna shown in Fig. 1017. Any discrepancy which may exist between line and antenna impedance can be compensated for by a slight fanning of the line

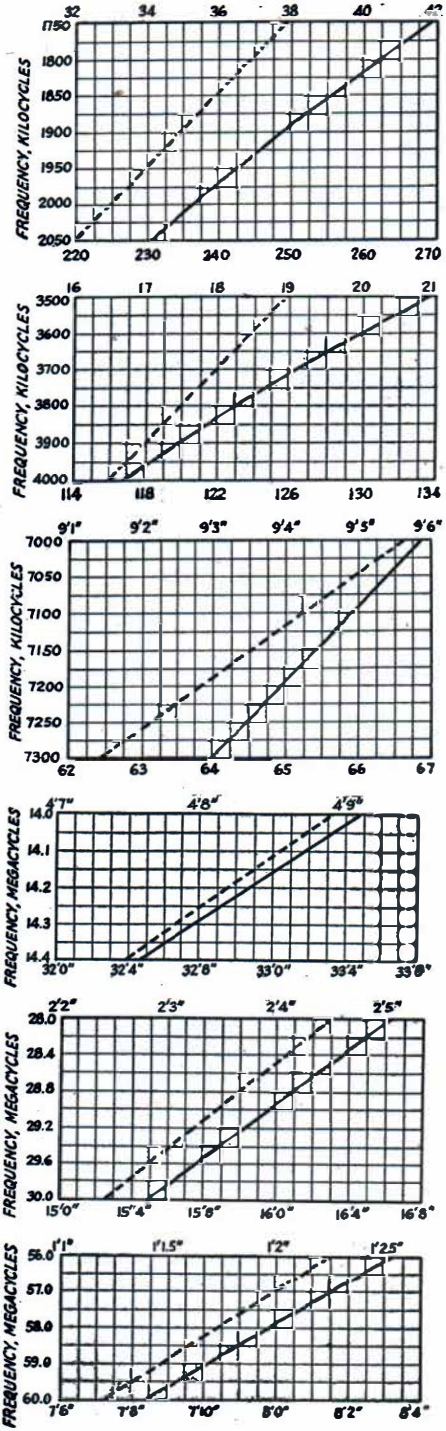


Fig. 1015 — Charts for determining the length of half-wave antennas for use on various amateur bands. Solid lines indicate antenna length (lower scale); dotted lines point of connection for single-wire feeder (upper scale) measured from center of antenna.

where it connects to the two halves of the antenna, as shown at *B* in Fig. 1017.

The twisted line is a convenient type to use, since it is easy to install and the r.f. voltage on it is low because of the low impedance. This makes insulation an easy matter. Special twisted line for transmitting purposes, having lower losses than ordinary rubber-covered wire, is available.

The antenna should be one-half wavelength long for the frequency of operation, as determined by charts of Fig. 1015 or the formulas (§ 10-2). The amount of "fanning" (dimension *B*) will depend upon the kind of cable used; the right value usually will be found between 6 and 18 inches. It may be checked by inserting ammeters in each antenna leg at the junction of the feeder and antenna; the value of *B* which gives the largest current is correct. Alternatively, the system may be operated continuously for a time with fairly high r.f. power input, after which the feeder may be inspected (by touch) for hot spots. These indicate the presence of standing waves, and the fanning should be adjusted until they are eliminated or minimized. Each leg of the feeder forming the triangle at the antenna should be equal in length to dimension *B*.

Coupling between transmitter and transmission line is ordinarily by the untuned coil method shown in Fig. 1010-A (§ 10-6).

Concentric-line feed — A concentric transmission line readily can be constructed to have a surge impedance equal to the 70-ohm impedance at the center of a half-wave antenna. Such a line, therefore, can be connected directly to the center of the antenna, forming the system shown in Fig. 1018.

Solving Equation (6) (§ 10-5) for an air-insulated concentric line shows that, for 70-ohm surge impedance, the inside diameter of the outer conductor should be approximately 3.2 times the outside diameter of the inner conductor. This condition can be fulfilled by using standard $\frac{5}{16}$-inch (outside-diameter) copper tubing for the outer conductor and No. 14 wire for the inner. Ceramic insulating spacers are available commercially for this combination. Rubber-insulated concentric line having the requisite impedance for connection to the center of the antenna also is available.

The operation of such an antenna system is similar to that of the twisted-pair system just described, and the same transmitter-coupling arrangements may be used (§ 10-6).

The outer conductor of the line may be grounded if desired. The feeder system is

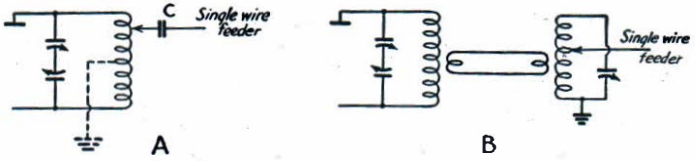


Fig. 1016 — Methods of coupling the single-wire feeder to the transmitter. Circuits are shown for both single-ended and balanced tank circuits.

slightly unbalanced because the inner and outer conductors do not have the same capacity to ground. There should be no radiation, however, from a line having a correct surge impedance.

Delta matching transformer — Because of the extremely close spacing required, it is impracticable to construct an open-wire transmission line which will have a surge impedance low enough to work directly into the center of a half-wave antenna. Such wire lines usually have impedances between 400 and 700 ohms, 600 ohms being a widely-used

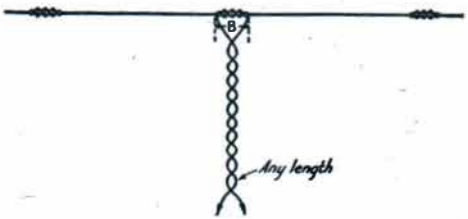


Fig. 1017 — Half-wave antenna center-fed by a twisted pair line.

value. It is therefore necessary to use other means for matching the line to the antenna.

One method of matching is illustrated by the antenna system of Fig 1019. The section *E* is "fanned" to have a gradually increasing impedance so that its impedance at the antenna end will be equal to the impedance of the antenna section *C*, while the impedance at the lower end matches that of a practicable transmission line.

The antenna length *L*, the feeder clearance *E*, the spacing between centers of the feeder wires *D*, and the coupling length *C* are the important dimensions of this system. The system

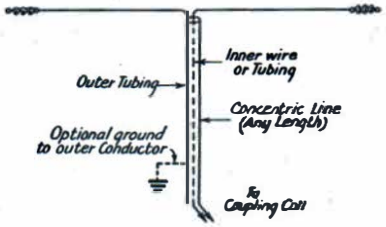


Fig. 1018 — Half-wave antenna with concentric transmission line.

must be designed for exact impedance values as well as frequency values and the dimensions are therefore fairly critical.

The length of the antenna is figured from the formula (§ 10-2) or taken from Fig. 1015.

The length of section *C* is computed by the formula:

$$C\ (feet) = \frac{123}{Freq.\ (Mc.)}$$

The feeder clearance *E* is found from the equation:

$$E\ (feet) = \frac{148}{Freq.\ (Mc.)}$$

The above equations are for feeders having a characteristic impedance of 600 ohms and will not apply to feeders of any other impedance. The proper feeder spacing for a 600-ohm transmission line is computed to a sufficiently close approximation by the following formula:

$$D = 75 \times d$$

where D is the distance between the centers of the feeder wires and *d* is the diameter of the wire. If the wire diameter is in inches the spacing will be in inches and if the wire diameter is in millimeters the spacing will be in millimeters.

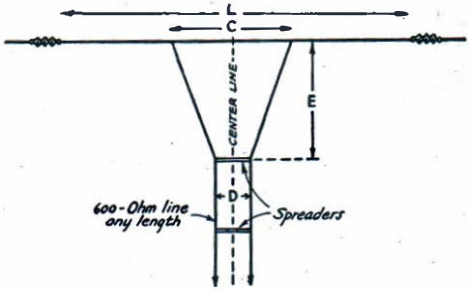


Fig. 1019 — Delta-matched antenna system. The dimensions C, D, and E are given in the text. It is important that the matching section, E, come straight away from the antenna without any bends.

Methods of coupling to the transmitter are discussed in § 10-6, Figs. 1010-C, D, G and H being suitable.

"Q"-section transformer — The impedance of a two-wire line of ordinary construction (400 to 600 ohms) can be matched to the impedance of the center of a half-wave antenna

by utilizing the impedance-transforming properties of a quarter-waveline (§ 10-5). The matching section must have low surge impedance and therefore is commonly constructed of large-diameter conductors such as aluminum or copper tubing, with fairly close spacing. This system is known as the "Q" antenna. It is shown in Fig. 1020. The important dimensions are the length of the antenna, the length of the matching section, *B*, the spacing between the two conductors of the matching section, *C*, and the impedance of the untuned transmission line connected to the lower end of the matching section.

The required surge impedance for the matching section is

$$Z_s = \sqrt{Z_1 Z_2} \qquad (9)$$

where Z_1 is the input impedance and Z_2 the output impedance. A quarter-wave section matching a 600-ohm line to the center of a half-wave antenna (72 ohms) should have a surge impedance of 208 ohms. The spacings

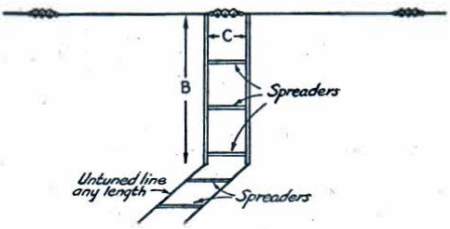


Fig. 1020 — The "Q" antenna with quarter-wave matching section using spaced tubing.

between conductors of various sizes of tubing and wire for different surge impedances are given in graphical form in Fig. 1009. With half-inch tubing, for example, the spacing should be 1.5 inches for an impedance of 208 ohms.

The length, *B*, of the matching section should be equal to a quarter wavelength, and is given by

$$\frac{\text{Length of } \frac{1}{4} \text{ wave line (feet)}}{} = \frac{234}{\text{Freq. (Mc.)}}$$

The length of the antenna can be calculated from the formula (§ 10-2) or taken from the charts of Fig. 1015.

This system has the advantage of the simplicity of adjustment of the twisted pair feeder system and at the same time the superior insulation of an open-wire system. Figs. 1010-B, D, G (§ 10-6) and H represent suitable methods of coupling to the transmitter.

Linear transformers — Fig. 1021 shows two methods of coupling a non-resonant line to a half-wave antenna through a quarter-

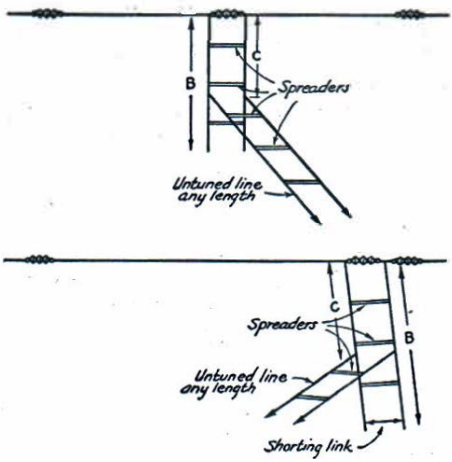


Fig. 1021 — Half-wave antenna systems with quarter-wave open wire matching transformers.

wave linear transformer (§ 10-5) or matching section. In the case of the center-fed antenna the free end of the matching section, *B*, is open (high impedance) since the other end is connected to a low-impedance point on the antenna. With the end-fed antenna the free end of the matching section is closed through a shorting bar or link; this end has low impedance since the other end is connected to a high-impedance point on the antenna (§ 10-7).

In the center-fed system, the antenna and matching section should be cut to lengths found from the equations in § 10-2 and § 10-5. Any necessary on-the-ground adjustment can be made by adding to or clipping off the open ends of the matching section. The matching section in the end-fed system can be adjusted by making the line a little longer than necessary and adjusting the system to resonance by moving the shorting link up and down. Resonance can be obtained by exciting the antenna at the proper frequency from a temporary antenna nearby and measuring the current in the shorting bar by a low-range r.f. ammeter or galvanometer. The position of the bar should be adjusted for maximum current reading. This should be done before the transmission line is attached to the matching section.

The position of the line taps must be determined experimentally, since it will depend upon the impedance of the line as well as the antenna impedance at the point of connection. The procedure is to take a trial point, apply power to the transmitter, and check the transmission line for standing waves. This can be done by measuring the current in the wires, using a device of the type pictured in

Fig. 1022. The hooks (which should be sharp enough to cut through insulation, if any, of the wires) are placed on one of the wires, the spacing between them being adjusted to give a suitable reading on the meter. At any one position along the line the currents in the two wires should be identical. Readings taken at intervals of a quarter wavelength will indicate whether or not standing waves are present.

It will not usually be possible to obtain complete elimination of standing waves when the matching stub is exactly resonant. The line taps should be adjusted for the smallest obtainable standing-wave ratio. Then a further "touching up" of the matching stub tuning will eliminate the remaining standing wave, provided the adjustments are made carefully. The stub must be readjusted because when resonant it exhibits some reactance as well as resistance at all points except at the ends, and the slight lengthening or shortening of the stub is necessary to tune out this reactance. The required readjustment is quite small, however.

When the connection between matching section and antenna is unbalanced, as in the end-fed system, it is important that the antenna be the right length for the operating frequency if a good match is to be obtained (§ 10-7). The balanced center-fed system is less critical in this respect. The shorting-bar method of tuning the center-fed system to resonance may be used if the matching section is extended to a half wavelength, bringing a current loop at the free end.

An impedance mismatch of several per cent is of little consequence so far as power transfer to the antenna is concerned. It is relatively easy to get the standing wave ratio down to 2 or 3 to 1, a perfectly satisfactory condition in practice. Of considerably greater importance is the necessity for getting the currents in the two wires balanced both as to amplitude and phase. If the currents are not the same at corresponding points on adjacent wires, and the loops and nodes do not also occur at corresponding points, there will be considerable radiation loss. This balance can only be brought about by perfect symmetry in the line, particularly with respect to ground. This symmetry should extend to the coupling apparatus at the transmitter. An electrostatic shield between the line and the transmitter coupling coils often will be of value in preventing capacity unbalance, and at the same time will reduce harmonic radiation.

10-9 LONG-WIRE ANTENNAS

Definition — An antenna will be resonant if an integral number of standing waves of current and voltage can exist along its length; in other words, so long as its length is some integral multiple of a half-wavelength. When the antenna is more than a half-wave long, it is usually called a long-wire antenna, or a harmonic antenna.

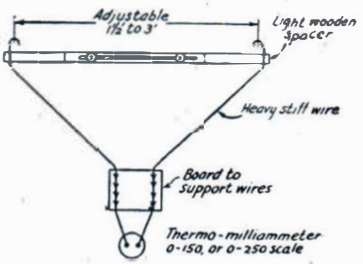


Fig. 1022 — Line-current measuring device for adjustment of untuned transmission lines.

Current and voltage distribution — Fig. 1023 shows the current and voltage distribution along a wire operating at its fundamental frequency (where its length is equal to a half wavelength) and at its second, third and fourth harmonics. For example, if the fundamental frequency of the antenna is 7 Mc., the current and voltage distribution will be as shown at A. The same antenna excited at 14 Mc. would have current and voltage distribution as shown at B. At 21 Mc., the third harmonic of 7 Mc., the current and voltage distribution would be as in C; and at 28 Mc., the fourth harmonic, as in D. The number of the harmonic is the number of half-waves contained in the antenna at the particular operating frequency.

The polarity of current or voltage in each standing wave is opposite to that in the adjacent standing waves. This is shown in the figure by drawing the current and voltage curves successively above and below the antenna (taken as a zero reference line) to indicate that the polarity reverses when the current or voltage goes through zero. Currents flowing in the same direction are *in phase;* in opposite directions, *out of phase.*

It is evident that one antenna may be used for harmonically related frequencies, such as the various amateur bands. The long-wire or harmonic antenna is the basis of multi-band operation with one antenna.

Physical lengths — The length of a long-wire antenna is not an exact multiple of that of a half-wave antenna because the end effects (§ 10-2) operate only on the end sections of the antenna; in other parts of the wire these effects are absent and the wire length is approximately that of an equivalent portion of the wave in space. The formula for the length of a long-wire antenna therefore is:

$$Length\ (feet) = \frac{492\ (N-0.05)}{Freq.\ (Mc.)} \qquad (10)$$

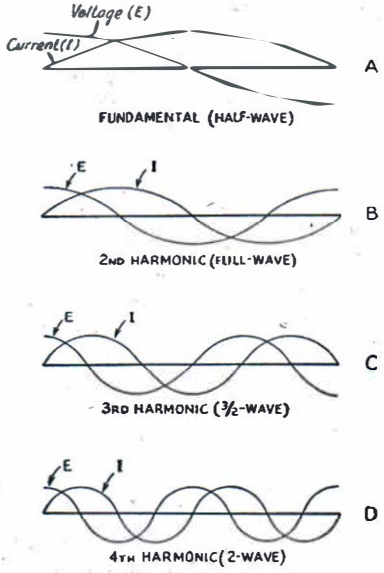


Fig. 1023 — Current and voltage distribution along an antenna operated at various harmonics of its fundamental resonant frequency.

where *N* is the number of *half*-waves on the antenna. From this it is apparent that an antenna cut as a half-wave for a given frequency will be slightly off resonance at exactly twice that frequency (on the second harmonic) because of the different behavior of end effects when there is more than one standing wave on the antenna. For instance, if the antenna is cut to exact fundamental resonance on the second harmonic (full wave) it should be 2.6% longer, and on the fourth harmonic (two-wave), 4% longer. The effect is not very important except for a possible unbalance in the feeder system (§ 10-7) which may result in some radiation from the feeder in end-fed systems.

Impedance and power gain — The radiation resistance as measured at a current loop becomes larger as the antenna length is increased. Also, a long-wire antenna radiates more power in its most favorable direction than does a half-wave antenna in *its* most favorable direction. This power gain is secured at the expense of radiation in other directions. Fig. 1024 shows how the radiation resistance and power in the lobe of maximum radiation vary with the antenna length.

Directional characteristics — As the wire is made longer, in terms of the number of half wavelengths, the directional effects change. Instead of the "doughnut" pattern of the half-wave antenna, the directional characteristic splits up into "lobes" which make various angles with the wire. In general, as the length of the wire is increased the direction of maximum radiation tends to approach the line of the antenna itself.

Directional characteristics for antennas one wavelength, three half-wavelengths, and two wavelengths long are given in Figs. 1025, 1026 and 1027, for three vertical angles of radiation. Note that as the wire length increases the radiation along the line of the antenna becomes more pronounced. Still longer antennas can be considered to be practically "end-on" radiators, even at the lower radiation angles.

Methods of feeding — In a long-wire antenna the currents in adjacent half-wave sections must be out of phase, as shown in Fig. 1028 and Fig. 1023. The feeder system must not upset this phase relationship. This requirement is met by feeding the antenna at either end or at any current *loop*. A two-wire feeder cannot be inserted at a current *node*, however, because this invariably brings the currents in two adjacent half-wave sections in phase; if the phase in one section could be reversed then the currents in the feeders would be in phase and the feeder radiation would not be cancelled out.

Either resonant or non-resonant feeders may be used. With the latter, the systems employing a matching section (§ 10-8) are best. The non-resonant line may be tapped on the matching section as in Fig. 1021 or a "Q" type section, Fig. 1020, may be employed.

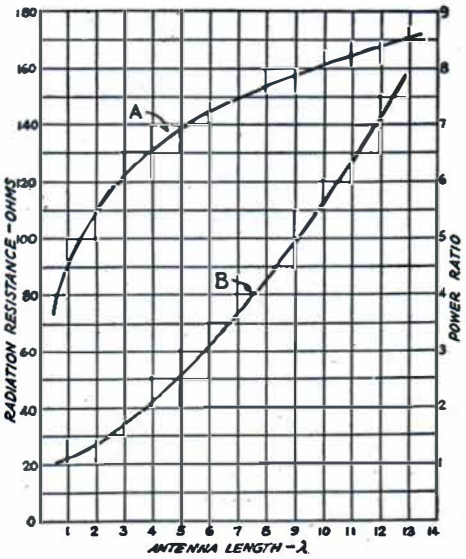


Fig. 1024 — Curve A, variation in radiation resistance with antenna length. Curve B, power in the lobes of maximum radiation for long-wire antennas as a ratio to the maximum of a half-wave antenna.

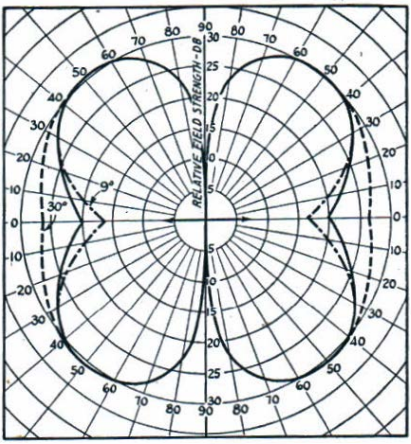


Fig. 1025 — Horizontal patterns of radiation from a full-wave antenna. Solid line, vertical angle 15 degrees; dotted lines, deviation from 15-degree pattern at 9 and 30 degrees.

All three patterns are drawn to the same relative scale; actual amplitudes will depend upon the height of the antenna.

In such case, Fig. 1029 gives the required surge impedance for the matching section. It can also be calculated from Equation 9 (§ 10-8) and the radiation resistance data in Fig. 1024.

Methods of coupling the line to the transmitter are the same as described in § 10-6 for the particular type of line used.

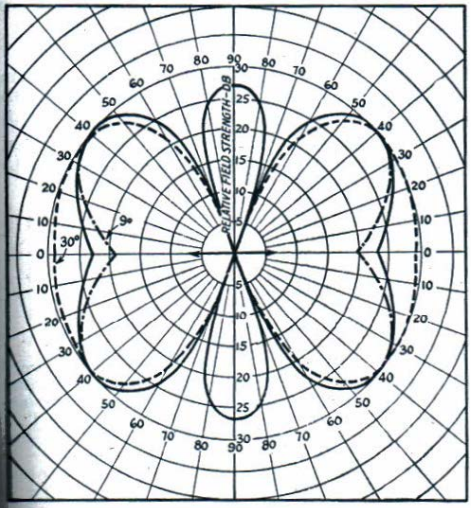


Fig. 1026 — Horizontal patterns of radiation from an antenna three half-wavelengths long. Solid line, vertical angle 15 degrees; dotted lines, deviation from 15-degree pattern at 9 and 30 degrees. The minor lobes coincide for all three angles.

•10-10 MULTI-BAND ANTENNAS

Principles — As suggested in the preceding section, the same antenna may be used for several bands by operating it on harmonics. When this is done, it is necessary to use resonant feeders, since the impedance matching for non-resonant feeder operation can be accomplished only at one frequency unless means are provided for changing the length of a matching section and shifting the point at which the feeder is attached to it. A matching section which is a quarter-wavelength long on one frequency will be a half-wavelength long on twice that frequency, and so on, and changing

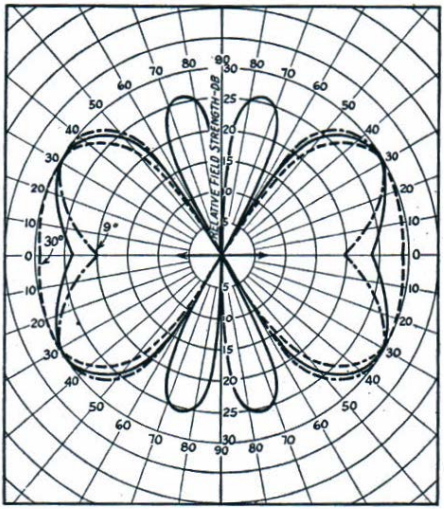


Fig. 1027 — Horizontal patterns of radiation from an antenna two wavelengths long. Solid line, vertical angle 15 degrees; dotted lines, deviation from 15-degree pattern at 9 and 30 degrees. The minor lobes coincide for all three angles.

the length of the wires, even by switching, is inconvenient.

Also, the current loops shift to a new position on the antenna when it is operated on harmonics, further complicating the feed situation. It is for this reason that half-wave antennas center-fed by rubber-insulated lines are practically useless for harmonic operation; on all even harmonics there is a voltage maximum at the feed point and the impedance mismatch is so bad that there is a large standing-wave ratio and consequently high losses in the rubber dielectric.

When the same antenna is used for work in several bands, it must be realized that the directional characteristic will depend on the band in use.

Simple systems — Any of the antenna arrangements shown in § 10-7 may be used for

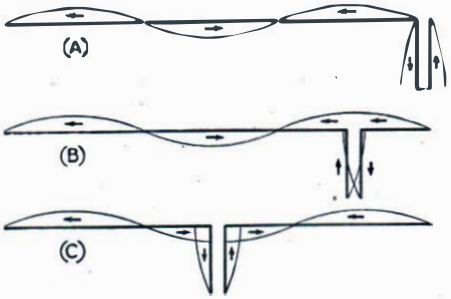


Fig. 1028 — Current distribution and feed points for long-wire antennas. A 3/2-wave antenna is used as an illustration. With two-wire feed, the line may be connected at the end of the antenna or at any current *loop* (not at a current node).

multi-band operation by making the antenna a half wave long at the lowest frequency to be used. The feeders should be a quarter wave, or some multiple of a quarter wave, long at the same frequency. Typical examples, with the type of tuning to be used, are given in Table I. The figures given represent a compromise to give satisfactory operation on all the bands considered, taking into account the change in required length as the order of the harmonic goes up.

A center-fed half-wave antenna will not operate as a long wire on harmonics because of the phase reversal at the feeders previously mentioned (§ 10-9). On the second harmonic, the two antenna sections are each a half wave long, and since the currents are in phase the directional characteristic is different from that of a full-wave antenna even though the overall length is the same. On the fourth harmonic, each section is a full wave long and again because of the direction of current flow the system will not operate as a two-wave antenna. It should not be assumed that these systems are not effective radiators — it is simply that the directional characteristic will not be that of a long-wire having the same overall length. Rather it will resemble the characteristic of one side of the antenna, although this is not exact.

Antennas with a few other types of feed systems may be operated on harmonics for the higher-frequency bands, although their performance is somewhat impaired. The single-wire fed antenna (§ 10-8) may be used in this way; the feeder and antenna will not be matched exactly on harmonics with the result that standing waves will appear on the feeder, but the system as a whole will radiate. The

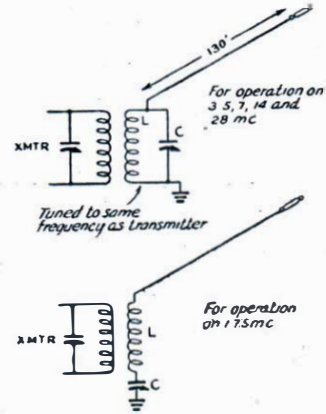


Fig. 1030 — A simple antenna system for five amateur bands. The antenna is voltage fed on 3.5, 7, 14 and 28 Mc., working on the fundamental, second, fourth and eighth harmonics, respectively. For 1.75 Mc. the system is a quarter-wave grounded antenna, in which case series tuning must be used. The antenna wire should be kept well in the clear and should be as high as possible. If the length of the antenna is approximately 260 feet, voltage feed can be used on all five bands.

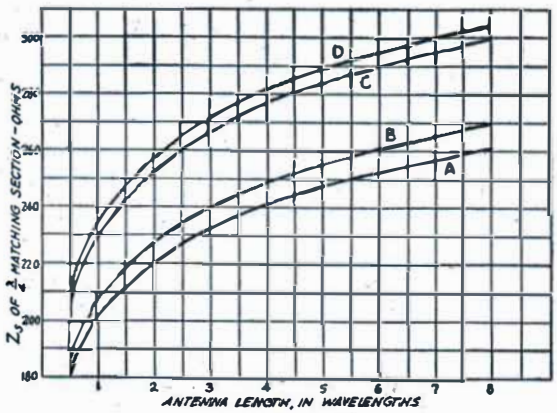


Fig. 1029 — Required surge impedance of quarter-wave matching sections for radiators of various lengths. Curve A is for a transmission line impedance of 440 ohms, Curve B for 470 ohms, Curve C for 580 ohms and Curve D for 600 ohms.

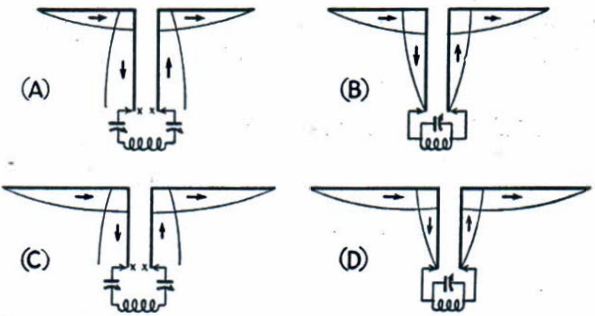


Fig. 1031 — Current distribution on antennas too short for the fundamental frequency. These systems may be used when space for a full half-wave antenna is not available. The current distribution on the second harmonic also is shown to the right of each figure. In A and C, the total length around the system is a half-wavelength at the fundamental frequency. Arrows show instantaneous direction of current flow.

TABLE I

MULTI-BAND RESONANT-LINE FED ANTENNAS

Antenna Length (ft.)	*Feeder Length (ft.)*	*Band*	*Type of Tuning*
With end feed: 243	120	1.75-Mc. 'phone	series
		4-Mc. 'phone	parallel
		14 Mc.	parallel
		28 Mc.	parallel
136	67	3.5-Mc. c.w.	series
		7 Mc.	parallel
		14 Mc.	parallel
		28 Mc.	parallel
134	67	3.5-Mc. c.w.	series
		7 Mc.	parallel
67	33	7 Mc.	series
		14 Mc.	parallel
		28 Mc.	parallel
With center feed: 272	135	1.75 Mc.	parallel
		3.5 Mc.	parallel
		7 Mc.	parallel
		14 Mc.	parallel
		28 Mc.	parallel
137	67	3.5 Mc.	parallel
		7 Mc.	parallel
		14 Mc.	parallel
		28 Mc.	parallel
67.5	34	7 Mc.	parallel
		14 Mc.	parallel
		28 Mc.	parallel

The antenna lengths given represent compromises for harmonic operation because of different end effects on different bands. The 136-foot end-fed antenna is slightly long for 3.5 Mc., but will work well in the region (3500-3600 kc.) which quadruples into the 14-Mc. band. Bands not shown are not recommended for the particular antenna. The center-fed systems are less critical as to length; the 272-foot antenna may, for instance, be used for both c.w. and 'phone on either 1.75 or 4 Mc. without loss of efficiency.

On harmonics, the end-fed and center-fed antennas will not have the same directional characteristics, as explained in the text.

same is true of the delta-matched antenna. The "Q" antenna (§ 10-8) also can be operated on harmonics, but the line cannot operate non-resonant except at the fundamental frequency of the antenna. For harmonic operation the line must be tuned and, therefore, the feeder length is important. The tuning system will depend upon the number of quarter waves on the line, including the "Q" bars. The concentric-line fed antenna (§ 10-8) may be used on harmonics if the concentric line is air-insulated. Its operation on harmonics is similar to that of the "Q." This antenna is not recommended for multi-band operation with a rubber-insulated line, however.

A simple antenna system, without feeders, for operation in five bands is shown in Fig. 1030. On all bands from 3.5 Mc. upward it operates as an end-fed antenna — half-wave on 3.5 Mc., long wire on the other bands. On 1.75 Mc. it is only a quarter-wave in length and must be worked against ground, which in effect replaces the missing half of the antenna. Since on this band it is fed at a high-current point, series tuning (§ 10-6) must be used.

Antennas for restricted space — If the space available for the antenna is not large enough to accommodate the length necessary for a half-wave at the lowest frequency to be used, quite satisfactory operation can be secured by using a shorter antenna and making up the missing length in the feeder system. The antenna itself may be as short as a quarter wavelength and still radiate fairly well, although of course it will not be as effective as one a half-wave long. Nevertheless such a system is useful where operation on the desired band otherwise would be impossible.

Resonant feeders are a practical necessity with such an antenna system, and a center-fed antenna will give best all-around performance. With end feed the feeder currents become badly unbalanced and, since lengths midway between those requiring series or parallel tuning ordinarily must be used to bring the entire system to resonance, coupling to the transmitter often becomes difficult.

With center feed, practically any convenient length of antenna can be used if the feeder length is adjusted to accommodate at least one half-wave around the whole system. Typical cases are shown in Fig. 1031, one for an antenna having a length of one-quarter wave (A) and the other for an antenna somewhat longer (C) but still not a half-wave long. Current distribution is shown for both fundamental and second harmonic. From the points marked *X* resonant feeders any convenient number of quarter waves in length may be extended to the operating room. The sum of the distances on each wire from *X* to the antenna end must equal a half-wave. It is sufficiently accurate to use Equation 2 (§ 10-2) in calculating this length. Note that *X*-*X* is a high-current point on these shortened antennas, corresponding to the center of a half-wave antenna. It is also apparent that the antenna at A is a half-wave antenna on the next higher-frequency band (B).

The practical antenna can be made as in Fig. 1032. Table II gives a few recommended lengths. Remembering the preceding discussion, however, the antenna can be made any convenient length provided the feeder is considered to "begin" at *X*-*X*, and the line length adjusted accordingly.

TABLE II

ANTENNA AND FEEDER LENGTHS FOR SHORT MULTI-BAND ANTENNAS, CENTER-FED

Antenna Length (ft.)	*Feeder Length (ft.)*	*Band*	*Type of Tuning*
137	68	1.75 Mc.	series
		3.5 Mc.	parallel
		7 Mc.	parallel
		14 Mc.	parallel
		28 Mc.	parallel
100	38	3.5 Mc.	parallel
		7 Mc.	series
		14 Mc.	series
		28 Mc.	series or parallel
67.5	34	3.5 Mc.	series
		7 Mc.	parallel
		14 Mc.	parallel
		28 Mc.	parallel
50	43	7 Mc.	parallel
		14 Mc.	parallel
		28 Mc.	parallel
33	51	7 Mc.	parallel
		14 Mc.	parallel
		28 Mc.	parallel
33	31	7 Mc.	parallel
		14 Mc.	series
		28 Mc.	parallel

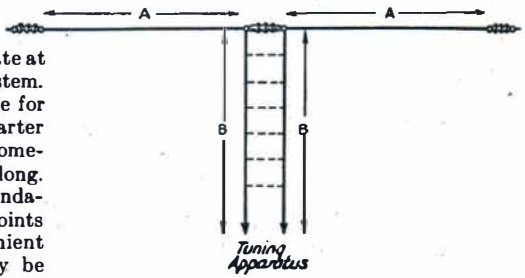


Fig. 1032 — Practical arrangement of a shortened antenna. The total length A + B + B + A, should be a half-wavelength for the lowest-frequency band, usually 3.5 Mc. See Table II for lengths and tuning data.

Bent antennas — Since the field strength at a distance is proportional to the current in the antenna, the high-current part of a half-wave antenna (the center quarter-wave, approximately) does most of the radiating (§ 10-1). Advantage can be taken of this fact when the space available does not permit erecting an antenna a half-wave long. To accomplish it, the ends may be bent, either horizontally or vertically, so that the total length equals a half wave, even though the straightaway horizontal length may be as short as a quarter wave. The operation is illustrated in Fig. 1033. Such an antenna will be a somewhat better radiator than the arrangement of Fig. 1031-A on the lowest frequency, but is not as desirable for multi-band operation because the ends play an increasingly important part as the frequency is raised. The performance of the system in such a case is difficult to predict, especially if the ends are vertical (the most convenient arrangement) because of the combination of horizontal and vertical polarization as well as dissimilar directional characteristics.

● 10-11 LONG-WIRE DIRECTIVE ARRAYS

The "V" antenna — It has been emphasized that as the antenna length is increased the lobe of maximum radiation makes a more acute angle with the wire (§ 10-9). Two such

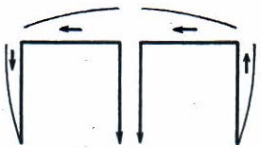


Fig. 1033 — Folded arrangement for shortened antennas. The total length is a half-wave, not including the feeders. The horizontal part is made as long as convenient and the ends dropped down to make up the required length. The ends may be bent back on themselves in feeder fashion to cancel radiation partially. The horizontal section should be at least a quarter-wave long.

wires may be combined in the form of a horizontal "V" so that the main lobes from each wire will reinforce along a line bisecting the angle between the wires. This increases both gain and directivity, since the lobes in directions other than along the bisector cancel to a greater or lesser extent. The horizontal "V" antenna therefore transmits best in either direction (is bi-directional) along the line bisecting the "V" made by the two wires. The power gain depends upon the length of the wires. Provided the necessary space is available, the "V" is a simple antenna to build and operate, and can be used readily on harmonics so that it is suitable for multi-band work. The "V" antenna is shown in Fig. 1034.

Fig. 1035 shows the dimensions that should be followed for an optimum design to obtain maximum power gain for different-sized "V" antennas. The longer systems give good performance in multi-band operation. Angle α is approximately equal to twice the angle of maximum radiation for a single wire equal in length to one side of the "V."

The wave angle referred to in Fig. 1035 is the vertical angle of maximum radiation (§ 10-1). Tilting the whole horizontal plane of the "V" will tend to increase the low-angle radiation off the low end and decrease it off the high end.

The gain increases with the length of the wires, but is not exactly twice the gain for a single long wire as given in Fig. 1024. In the longer lengths, the gain will be somewhat increased because of mutual coupling between the wires. A "V" eight wavelengths on a leg, for instance, will have a gain of about 12 db. over a half-wave antenna, whereas twice the gain of a single 8-wavelength wire would be approximately 9 db.

The two wires of the "V" must be fed out of phase for correct operation. A resonant line may simply be attached to the ends as shown in Fig. 1034. Alternatively, a quarter-wave matching section may be employed and the antenna fed through a non-resonant line (§ 10-8). If the antenna wires are made multiples of a half-wave in length (use Equation 10, § 10-9, for computing the length) the matching section will be closed at the free end.

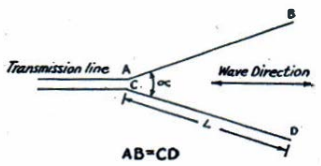


Fig. 1034 — The "V" antenna, made by combining two long wires in such a way that each reinforces the other's radiation. The important quantities are the length of each leg and the angle between legs.

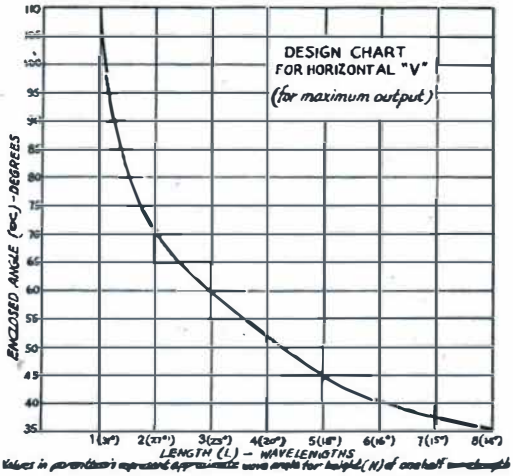


Fig. 1035 — Design chart for horizontal "V" antennas. Enclosed angle between wires versus length of sides.

The rhombic antenna — The horizontal rhombic or "diamond" antenna is shown in Fig. 1036. Like the "V," it requires a good deal of space for erection, but it is capable of giving excellent gain and directivity. It can also be used for multi-band operation. In the terminated form shown in Fig. 1036 it operates, like a non-resonant transmission line, without standing waves, and is uni-directional. It may also be used without the terminating resistor, in which case there are standing waves on the wires and the antenna is bi-directional.

The important quantities influencing the design of the rhombic antenna are shown in Fig. 1036. While several design methods may be used, the one most applicable to the conditions existing in amateur work is the so-called "compromise" method. The chart of Fig. 1037 gives design information based on a given length and wave angle to determine the remaining optimum dimensions for best operation. Curves for values of length of 2, 3 and 4 wavelengths are shown, and intermediate values may be interpolated.

With all other dimensions correct, an increase in length causes an increase in power gain and a slight reduction in wave angle. An increase in height also causes a reduction in wave angle and an increase in power gain but not to the same extent as a proportionate increase in length.

For multi-band work, it is satisfactory to design the rhombic antenna on the basis of 14-Mc. operation, which will permit work on the 7- and 28-Mc. bands as well.

A value of 800 ohms is correct for the terminating resistor for any properly constructed

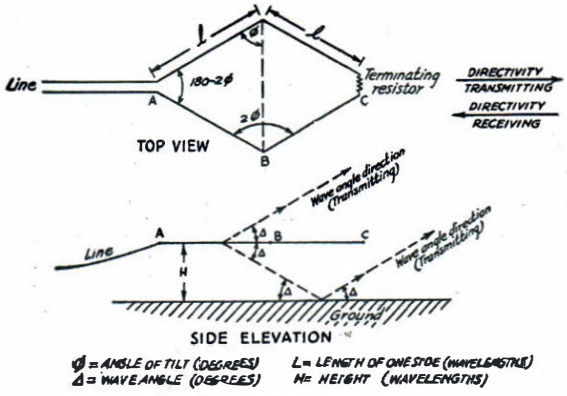


Fig. 1036 — The horizontal rhombic or diamond antenna, terminated.

rhombic, and the system behaves as a pure resistive load under this condition. This terminating resistor must be capable of safely dissipating one-half the power output (to eliminate the rear pattern) and should be non-inductive. Such a resistor may be made up from a carbon or graphite rod or from a long 800-ohm transmission line using resistance wire. If the carbon rod or a similar form of lumped resistance is used the device should be suitably protected from weather effects, i.e., covered with good asphaltic compound and sealed in a small light-weight box or fibre tube. Suitable resistors are available commercially.

For feeding the antenna, the antenna impedance will be matched by an 800-ohm line, which may be constructed from No. 16 wire spaced 20 inches or from No. 18 wire spaced 16 inches. The 800-ohm line is somewhat ungainly to install, however, and may be replaced by an ordinary 600-ohm line with only a negligible mismatch.

Alternatively, a matching section may be installed between the antenna terminals and a low-impedance line. However, when such an arrangement is used it will be necessary to change the matching section constants for each different band of operation.

The same design details apply to the unterminated rhombic as to the terminated type. Resonant feeders are preferable for the unterminated rhombic. A non-resonant line may be used by incorporating a matching section at the antenna, but is not readily adaptable to multi-band work.

Rhombic antennas will give a power gain of 10 db. or more when constructed according to the charts given. In general, the larger the antenna the greater the power gain.

●10-12 DIRECTIVE ARRAYS WITH DRIVEN ELEMENTS

Principles — By combining individual half-wave antennas into an *array* with suitable spacing between antennas (called *elements*) and feeding power to them simultaneously, it is possible to make the radiated fields from the individual elements add in a favored direction, thus increasing the field strength in that

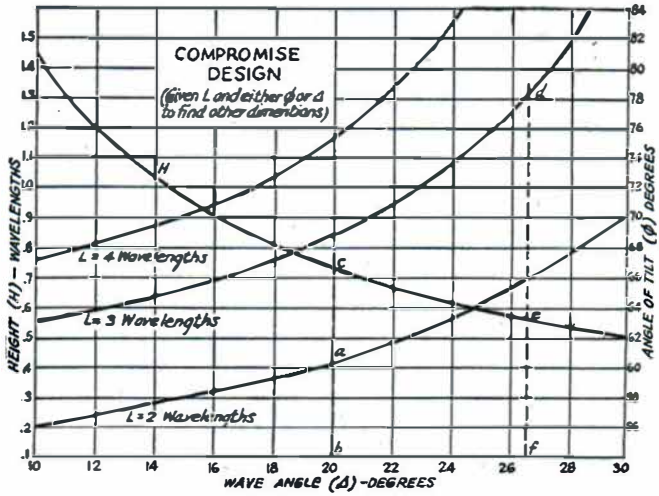


Fig. 1037 — Compromise method design chart for various leg lengths and wave angles. The following examples illustrate the use of the Chart:

(1) Given: Length (L) = 2 wave-lengths.
Desired wave angle (Δ) = 20°.
To Find: H, Φ.
Method:
Draw vertical line through point "a" (L = 2 wavelengths) and point "b" on abscissa (Δ = 20°.) Read angle of tilt (Φ) for point "a" and height (H) from intersection of line ab at point "c" on curve H.
Result:
Φ = 60.5°.
H = 0.73 wavelength.

(2) Given:
Length (L) = 3 wavelengths.
Angle of tilt (Φ) = 78°.
To Find: H, Δ.
Method:
Draw vertical line from point "d" on curve L = 3 wavelengths at Φ = 78°. Read intersection of this line on curve H (point "e" and intersection at point "f" on the abscissa for Δ.
Result:
H = 0.56 wavelength.
Δ = 26.6°.

direction as compared to that produced by one antenna element alone. In other directions the fields will more or less oppose each other, giving a reduction in field strength. Thus the power gain in the desired direction is secured at the expense of a power reduction in other directions.

Besides spacing between elements, the instantaneous direction of current flow (*phase*) in individual elements determines the directivity and power gain. There are several methods of arranging the elements. If they are strung end to end so that all lie on the same straight line, the elements are said to be *collinear*. If they are parallel and all lying in the same plane, the elements are said to be *broadside* when the phase of the current is the same in all, and *end-fire* when the currents are not in phase. Elements which receive power from the transmitter through the transmission line are called *driven elements*.

The power gain of a directive system increases with the number of elements. The proportionality between gain and number of elements is not simple, however, but depends upon the effect of the spacing and phasing upon the radiation resistance of the elements, as well as upon their number.

Collinear arrays — Simple forms of collinear arrays, with the current distribution, are shown in Fig. 1038. The two-element array at *A* is popularly known as "two half-waves in phase." It will be recognized as simply a center-fed antenna operated at its second harmonic. The way in which the number of elements may be extended for increased directivity and gain is shown in Fig. 1038-B. Note that quarter-wave transmission lines are used between each element; these give the reversal in phase necessary to make the currents in individual antenna elements all flow in the same direction at the same instant. Another way of looking at it is to consider that the whole system is a long wire with alternate half-wave sections folded so that they do not radiate. Any phase-reversing section may be used as a quarter-wave matching section for attaching a non-resonant feeder (§ 10-8), or a resonant transmission line may be substituted for any of the quarter-wave sections. Also, the antenna may be end-fed by any of the systems previously described (§ 10-7, 10-8) or any

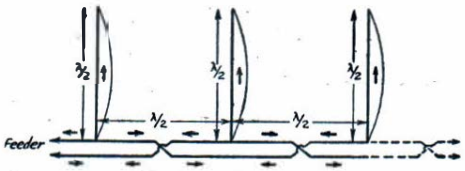


Fig. 1039 — The broadside array using half-wave elements. Arrows indicate direction of current flow. The transposition in feeders is necessary to bring the antenna currents in phase. Any reasonable number of elements may be used. The array is bi-directional perpendicular to the plane of the antenna; i.e., perpendicularly through this page.

element may be center-fed. It is best to feed at the center of the array so that the energy will be distributed as uniformly as possible among the elements.

The gain and directivity depend upon the number of elements and their spacing, center-to-center. This is shown by Table III. Although ¾-wave spacing gives greater gain, it is difficult to construct a suitable phase-reversing system when the ends of the antenna elements are widely separated. For this reason the half-wave spacing is generally used.

TABLE III

THEORETICAL GAIN OF COLLINEAR HALF-WAVE ANTENNAS

Spacing Between Centers of Adjacent Half Waves	*Number of Half Waves in Array vs. Gain in db.*				
	2	3	3	5	6
½ Wave	1.8	3.3	4.5	5.3	6.2
¾ Wave	3.2	4.8	6.0	7.0	7.8

Collinear arrays may be mounted either horizontally or vertically. Horizontal mounting gives horizontal directivity, with vertical directivity the same as for a single element at the same height. Vertical mounting gives the same horizontal pattern as a single element, but concentrates the radiation at low angles. It is seldom possible to use more than two elements vertically at frequencies below 14 Mc. because of the height required.

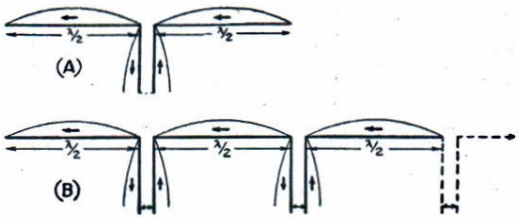


Fig. 1038 — Collinear half-wave antennas in phase. The system at *A* is generally known as "two-half-waves in phase." *B* is an extension of the system; in theory it may be carried on indefinitely, but practical considerations usually limit the number of elements to four.

Broadside arrays — Parallel antenna elements with currents in phase may be combined as shown in Fig. 1039 to form a *broadside* array, so named because the direction of maximum radiation is broadside to the plane containing the antennas. Again the gain and directivity depend upon the number of elements and the spacing, the gain for different spacings being shown in Fig. 1040. Half-wave spacing is generally used, since it simplifies feeding when the array has more than two elements. Table IV gives theoretical gain as a function of the number of elements.

Broadside arrays may be suspended either with the elements all vertical or with them horizontal and one above the other (*stacked*). In the former case the horizontal pattern is quite sharp while the vertical pattern is that of one element alone. If the array is suspended horizontally the horizontal pattern is that of one element while the vertical pattern is sharp, giving low-angle radiation.

Broadside arrays may be fed either by resonant transmission lines (§ 10-7) or through quarter-wave matching sections and nonresonant lines (§ 10-8). In Fig. 1039, note the "crossing over" of the feeder, necessary to bring the elements in proper phase relationship.

Combined broadside and collinear arrays — Broadside and collinear arrays may be combined to give both horizontal and vertical directivity, as well as additional gain. The general plan of constructing such antennas is shown in Fig. 1041. The lower angle of radiation resulting from stacking elements in the vertical plane is desirable at the higher frequencies. In general, doubling the number of elements in an array by stacking will raise the gain 2 to 4 db., depending upon whether vertical or horizontal elements are used — that is, whether the stacked elements are broadside or collinear.

The arrays in Fig. 1041 are shown fed from one end, but this is not especially desirable in the case of large arrays. Better distribution of energy between elements, and hence better all-around performance, will result when the feeders are attached as nearly as possible to the center of the array. Thus in the 8-element array at A the feeders could be introduced at the middle of the transmission line between the second and third set of elements, in which case the connecting line would not be transposed. Or the antenna could be constructed with the transpositions as shown and the feeder connected between the adjacent ends of either the second or third pair of collinear elements.

A four-element array of the general type shown at B is frequently used. It is shown, with the feed point indicated, in Fig. 1042.

End-fire arrays — Fig. 1043 shows a pair of parallel half-wave elements with currents out of phase. This is known as an *end-fire* array

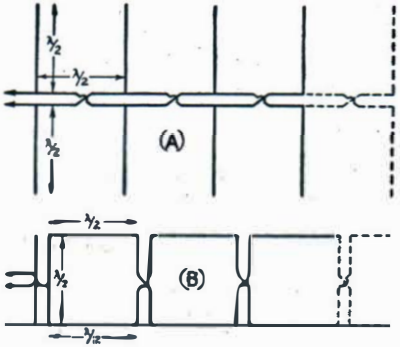


Fig. 1041 — Combination broadside and collinear arrays. A, with vertical elements; B, with horizontal elements. Both arrays give low-angle radiation. Two or more sections may be used.

The gain in db. will be equal, approximately, to the sum of the gain for one set of broadside elements (Table IV) plus the gain of one set of collinear elements (Table III). For example, in A each broadside set has four elements (gain 7 db.) and each collinear set two elements (gain 1.8 db.) giving a total gain of 8.8 db. In B each broadside set has two elements (gain 4 db.) and each collinear set three elements (gain 3.3 db.) making the total gain 7.3 db. The result is not strictly accurate because of mutual coupling between elements, but is good enough for practical purposes.

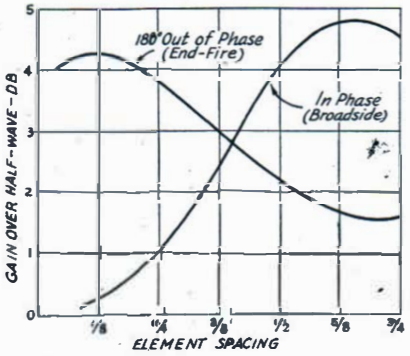


Fig. 1040 — Gain vs. spacing for two parallel half-wave elements.

TABLE IV

THEORETICAL GAIN VS. NUMBER OF BROADSIDE ELEMENTS (HALF-WAVE SPACING)

No. of Elements	*Gain*
2	4 db.
3	5.5 db.
4	7 db.
5	8 db.
6	9 db.

because it radiates best along the line of the antennas, as shown.

The end-fire array may be used vertically or horizontally (elements at the same height) and is well adapted to amateur work because it gives maximum gain with relatively close element spacing. Fig. 1040 shows how the gain varies with spacing. End-fire elements may be combined with additional collinear and broadside elements further to increase the gain and directivity.

Either resonant or non-resonant lines may be used with this type of array, the latter being

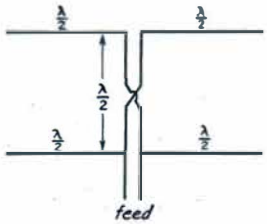


Fig. 1042 — A four-element combination broadside-collinear popularly known as the "lazy H" antenna. A closed quarter-wave stub may be used at the feed point to match into a 600-ohm line, or resonant feeders may be attached at the point shown. The gain over a half-wave antenna is 5 to 6 db.

preferably matched to the antenna through a quarter-wave matching section (§ 10-8).

Checking phasing — Figs. 1041 and 1043 illustrate a point in connection with feeding a phased antenna system which sometimes is confusing. In Fig. 1043 when the transmission line is connected as at A there is no cross-over in the line connecting the two antennas, but when the transmission line is connected to the center of the connecting line the cross-over becomes necessary (B). This is because in B the two halves of the connecting line are simply *branches* of the same line. In other words, even though the connecting line in B is a half-wave in length, it is not actually a half-wave line but *two quarter-wave lines in parallel.* The same thing is true of the untransposed line of Fig. 1041. Note that under these conditions the antenna elements are in phase when the line is not transposed, and out of phase when the transposition is made. The opposite is the case when the half-wave line simply joins two antenna elements and does not have the feed line connected to its center, as in Fig. 1039.

Adjustment of arrays — With arrays of the types just described, using half-wave spacing between elements, it will usually suffice to make the length of each element that given by the equation for a half-wave antenna in § 10-2, while the half-wave phasing

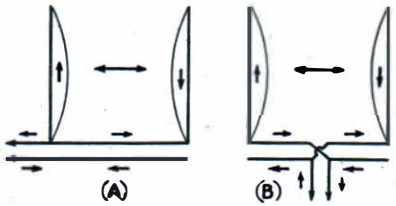


Fig. 1043 — End-fire arrays. They are shown with half-wave spacing to illustrate feeder connections. In practice, closer spacings are desirable, as shown by Fig. 1040. Direction of maximum radiation is shown by the large arrows.

lines between parallel elements can be calculated from the formula

$$\text{Length of half-wave line (feet)} = \frac{492 \times 0.975}{\text{Freq. (Mc.)}} = \frac{480}{\text{Freq. (Mc.)}}$$

The spacing between elements can be made equal to the length of the phasing line. No special adjustments are needed provided the formulas are followed carefully.

With collinear arrays of the type shown in Fig. 1038-B the same formula may be used for the element length, while the quarter-wave phasing section can be calculated from Equation 7 (§ 10-5). If the array is fed at its center it will not be necessary to make any particular adjustments, although if desired the whole system may be resonated by connecting an r.f. ammeter in the shorting link on each phasing section and moving the link back and forth to find the maximum current position. This refinement is hardly necessary in practice so long as all elements are the same length and the system is symmetrical.

Simple arrays — Several simple directive antenna systems using driven elements are in rather wide use among amateurs. They are shown in Fig. 1044. Tuned feeders are assumed in all cases; however, a matching section (§ 10-8) readily can be substituted if a non-resonant transmission line is preferred. Dimensions given are in terms of wavelength; actual lengths readily can be calculated from the equations in § 10-2 for the antenna and Equation 7 (§ 10-5) for the resonant transmission line or matching section. In cases where the transmission line proper connects to the mid-point of a phasing line, only ***half*** the length of the latter is added to the line to find the quarter-wave point.

At A and B are two-element end-fire arrangements using close spacing. They are electrically equivalent; the only difference is in the method of connecting the feeders. B may also be used as a four-element array on the second harmonic, although the spacing is not quite optimum (Fig. 1040) in that case. A close-spaced four-element array is shown at C. It will give

about 2 db. more gain than the two-element array. The antenna at D is designed to take advantage of the greater gain possible with collinear antennas having greater than half-wave center-to-center spacing, but without introducing feed complications. The elements are made longer than a half wave to bring this about. The gain is 3 db. over a single half-wave antenna, and the broadside directivity is quite sharp.

The antennas of A and B may be mounted either horizontally or vertically; horizontal suspension (with the two elements in a plane parallel to the ground) is recommended, since this tends to give low-angle radiation without an unduly sharp horizontal pattern. Thus these systems are useful for coverage over a wide horizontal angle. The system at C, when mounted horizontally, will have a sharper horizontal pattern than the two-element arrays.

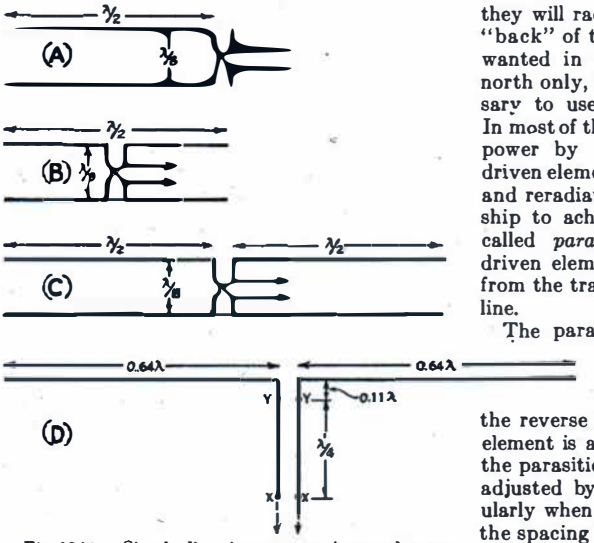


Fig. 1044 — Simple directive systems. A, two-element end-fire array; B, same with center feed, which permits use of the array on the second harmonic, where it becomes a four-element array with quarter-wave spacing C, four-element end-fire array with ⅛-wave spacing. D, extended in-phase antennas ("extended double-Zepp"). The gain of A and B is slightly over 4 db. On the second harmonic, B will give about 5 db. gain. With C, the gain is approximately 6 db., and with D, approximately 3 db.

In the first three, the phasing line contributes about 1/16th wavelength to the transmission line; when B is used on the second harmonic this contribution is ⅛ wavelength. Alternatively, the antenna ends may be bent to meet the transmission line, in which case each feeder is simply connected to one antenna. In D, points Y-Y indicate a quarter-wave point (high current) and X-X a half-wave point (high voltage). The line may be extended in multiples of quarter-waves, if resonant feeders are to be used.

A, B, and C may be suspended on wooden spreaders. The plane containing the wires should be parallel to the ground.

● 10-13 DIRECTIVE ARRAYS WITH PARASITIC ELEMENTS

Parasitic excitation — The antenna arrays described in § 10-12 are bi-directional; that is, they will radiate both to the "front" and the "back" of the antenna system. If radiation is wanted in only one direction (for instance, north only, instead of north-south) it is necessary to use different element arrangements. In most of these the additional elements receive power by induction or radiation from the driven element, generally called the "antenna," and reradiate it in the proper phase relationship to achieve the desired effect. They are called *parasitic* elements, as contrasted to driven elements which receive power directly from the transmitter through the transmission line.

The parasitic element is called a *director* when it reinforces radiation on a line pointing to it from the antenna, and is called a *reflector* when the reverse is the case. Whether the parasitic element is a director or reflector depends upon the parasitic element tuning (which usually is adjusted by changing its length) and, particularly when the element is self-resonant, upon the spacing between it and the antenna.

Gain vs. spacing — The gain of an antenna-

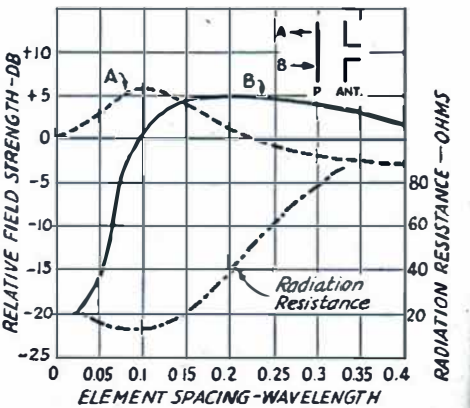


Fig. 1045 — Gain vs. element spacing for an antenna and one parasitic element. Zero db. is the field strength from a half-wave antenna alone. Greatest gain is in the direction A at spacings less than 0.14 wavelength; in direction B at greater spacings. Front-to-back ratio is the difference in db. between curves A and B. Variation in radiation resistance of the driven element also is shown. These curves are for self-resonant parasitic element. At most spacings the gain as a reflector can be increased by slight lengthening of the parasitic element; as a director, by shortening. This likewise improves the front-to-back ratio.

reflector or antenna-director combination varies chiefly with the spacing between elements. The way in which gain varies with spacing is shown in Fig. 1045, for the special case of self-resonant parasitic elements. This chart also shows how the attenuation to the "rear" varies with spacing. The same spacing does not necessarily give both maximum forward gain and maximum backward attenuation. Backward attenuation is desirable when the antenna is used for receiving, since it greatly reduces interference coming from the opposite direction to the desired signal.

Element lengths — The antenna length is given by the formulas in § 10-2. The director and reflector lengths must be determined experimentally for maximum performance. The preferable method is to aim the antenna at a receiver a mile or more distant and have an observer check the signal strength (on the "S" meter) while the reflector or director is adjusted a few inches at a time, until the length which gives maximum signal is found. The attenuation may be similarly checked, the length being adjusted for minimum signal. In general, the length of a director will be about 4% less than that of the antenna, for best front-to-back ratio. The reflector will be about 5% longer than the antenna.

Simple systems — the rotary beam — Practical combinations of antenna, reflector and director are shown in Fig. 1046. Spacings for maximum gain or maximum front-to-back ratio (ratio of power radiated in the desired direction to power radiated in the opposite direction) may be taken from Fig. 1045. In the chart, the front-to-back ratio in db. will be the sum of gain and attenuation at the same spacing.

Systems of this type are popular for rotary beam antennas, where the whole antenna is rotated to permit its gain and directivity to be utilized for any compass direction. They may be mounted either horizontally (plane containing the elements parallel to the earth) or vertically.

Arrays using more than one parasitic element, such as those shown at C and D in Fig. 1046, will give more gain and directivity than is indicated for the single reflector and director by the curves of Fig. 1045. The gain with a properly adjusted three-element array (antenna, director and reflector) will be 5 to 7 db. over a half-wave antenna, while somewhat higher gain still can be secured by adding a second director to make a four-element array. The front-to-back ratio is correspondingly improved as the number of elements is increased.

The elements in close-spaced (less than one-quarter wavelength element spacing) arrays preferably should be made of tubing of half-

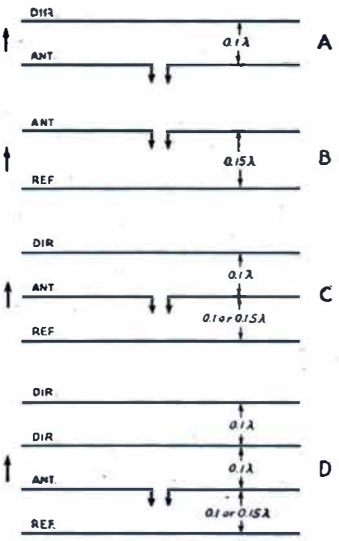


Fig. 1046 — Half-wave antennas with parasitic elements. A, with reflector; B, with director; C, with both director and reflector; D, two directors and one reflector. Gain is approximately as shown by Fig. 1045 in the first two cases and depends upon the spacing and length of the parasitic element. In the three- and four-element arrays a reflector spacing of 0.15 wavelength will give slightly more gain than 0.1-wavelength spacing. Arrows show direction of maximum radiation. The array should be mounted horizontally (these are top views).

to one-inch diameter both to reduce the ohmic resistance (§ 10-2) and to secure mechanical rigidity. If the elements are free to move with respect to each other the array will show detuning effects in a wind.

Feeding close-spaced arrays — While any of the usual methods of feed may be applied to the driven element of a parasitic array, the fact that with close spacing the radiation resistance as measured at the center of the driven element drops to a very low value makes some systems more desirable than others. The preferred methods are shown in Fig. 1047. Resonant feeders are not recommended for lengths greater than a half wavelength.

The quarter- or half-wave matching stubs shown at A and B in Fig. 1047 preferably should be constructed of tubing with rather close spacing, in the manner of the "Q" section. This lowers the impedance of the matching section and makes the position of the line taps somewhat less difficult to determine accurately. This line adjustment should be made only with the parasitic elements in place, and after the correct element lengths have been determined should be checked to compensate for changes likely to occur because of element tuning. The procedure is the same as that described in § 10-8.

The concentric-line matching section at C will work with fair accuracy into a close-spaced parasitic array of 2, 3 or 4 elements without necessity for adjustment. The line is used as an impedance inverting transformer, and if its characteristic impedance is 70 ohms will give an exact match to a 600-ohm line when the resistance at the termination is about 8.5 ohms. Over a range of 5 to 15 ohms the mismatch, and therefore the standing-wave ratio, will be less than 2 to 1. The length of the quarter-wave section should be calculated from Equation 7 (§ 10-5).

The delta matching transformer shown at D is an excellent arrangement for parasitic arrays, and probably is easier to install, mechanically, than any of the others. The positions of the taps (dimension *a*) must be determined experimentally, along with the length *b*, by checking the standing-wave ratio on the line as adjustments are made. Dimension *b* should be about 15% longer than *a*.

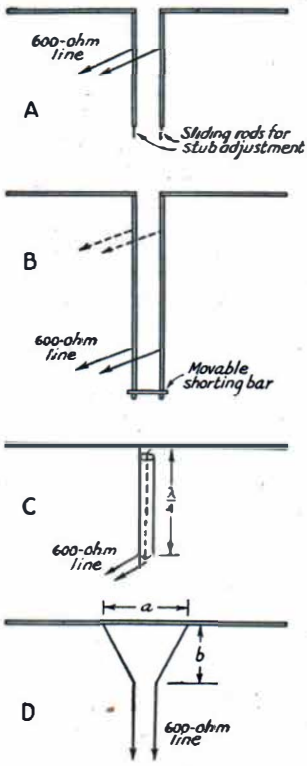


Fig. 1047 — Recommended methods of feeding the driven antenna element in close-spaced parasitic arrays. The parasitic elements are not shown. A, quarter-wave open stub; B, half-wave closed stub; C, concentric-line quarter-wave matching section; D, delta matching transformer.

Sharpness of resonance — Peak performance of a multi-element directive array depends upon proper phasing or tuning of the elements, which in all but the simplest systems can be exact for one frequency only. However, there is some latitude, and most arrays will work well over a relatively narrow band such as 14 Mc. If frequencies in all parts of the band are to be used, the antenna system should be designed for the mid-frequency; on the other hand, if only one frequency in the band will be used the greater portion of the time the antenna might be designed for that frequency and some degree of misadjustment tolerated on the occasionally-used spare frequencies.

When reflectors or directors are used the tolerance is usually less than in the case of driven elements, partly because the parasitic-element lengths are fixed and the operation may change appreciably as the frequency passes from one side of resonance to the other, and partly because the close spacing ordinarily used results in a sharp-tuning system. With parasitic elements operation should be confined to a small region about the frequency for which the antenna is adjusted, if peak performance is to be secured.

Combination arrays — It is possible to combine parasitic elements with driven elements to form arrays composed of collinear driven and parasitic elements and combination broadside-collinear-parasitic elements. Thus two or more collinear elements might be provided with a collinear reflector or director set, one parasitic element to each driven element. Or both directors and reflectors might be used. A broadside-collinear array could be treated in the same fashion.

When combination arrays are built up, a rough approximation of the gain to be expected may be obtained by adding the gains for each type of combination. Thus the gain of two broadside sets of four collinear arrays with a set of reflectors, one behind each element, at quarter-wave spacing for the parasitic elements, would be estimated as follows: From Table III, the gain of four collinear elements is 4.5 db. with half-wave spacing; from Fig. 1040 or Table IV, the gain of two broadside elements at half-wave spacing is 4.0 db.; from Fig. 1045 the gain of a parasitic reflector at quarter-wave spacing is 4.5 db.; the total gain is then the sum, or 13 db. for the sixteen elements. Note that using two *sets* of elements in broadside is equivalent to using two elements, so far as gain is concerned, similarly with sets of reflectors as against one antenna and one reflector. The actual gain of the combination array will depend, in practice, upon the way in which the power is distributed between the various elements, and upon the effect of mutual coupling between elements upon the

radiation resistance of the array, and may be somewhat higher or lower than the estimate.

A great many directive antenna combinations can be worked out by combining elements according to these principles.

●10-14 MISCELLANEOUS ANTENNA SYSTEMS

Grounded antenna — The grounded antenna is used almost exclusively for 1.75-Mc. work, where the length required for a half-wave antenna would be excessive for most locations. An antenna worked "against ground" need be only a quarter-wave long, approximately, because the earth acts as an electrical "mirror" which supplies the missing quarter wave. The current at the ground connection with a quarter-wave antenna is maximum, just as it is at the center of a half-wave antenna.

On 1.75 Mc. the most useful radiation is from the vertical part of the antenna, since vertically-polarized waves are characteristic of ground-wave transmission. It is therefore desirable to make the down-lead as nearly vertical as possible, and also as high as possible. This gives low-angle sky-wave transmission which is most useful for long-distance work at night, in addition to a good ground wave for local work. The horizontal portion contributes to high-angle sky-wave transmission, which is useful for covering short distances on this band at night.

Fig. 1048 shows a grounded antenna with the top folded to make the length equal to a quarter wave. The antenna coupling apparatus consists of the coil L, tuned by the series condenser C, with L inductively coupled to the transmitter tank circuit (§ 10-4, 10-6).

For computation purposes, the *overall* length of a grounded system is given by

$$L\ (feet) = \frac{236}{f\ (Mc.)}$$

This is the *total* length from the far end of the antenna to the ground connection. The length is not critical, since departures of the order of 10% to 20% can be compensated by the tuning apparatus.

The ground should preferably be one with conductors buried deep enough to reach natural moisture. In urban locations, good grounds can be made to water mains where they enter the house; the pipe should be scraped clean and a low-resistance connection made with a tightly-fastened ground clamp. If no water-pipes are available several pipes, six to eight feet long, may be driven into the ground at intervals of six or eight feet, all being connected together. The transmitter should be located so as to make the ground lead as short as possible.

In locations where it is impossible to secure a good ground connection because of sandy soil or other considerations, it is preferable to use a counterpoise or capacity ground instead of an actual ground connection. The counterpoise consists of a system of wires insulated from ground running horizontally above the earth beneath the antenna. The counterpoise should have a sufficient number of wires of sufficient length to cover well the area immediately under the antenna. The wires may be formed into any convenient shape, i.e., they may be spread out fan-shape, in a radial pattern, or three or more parallel wires separated a few feet running beneath the antenna may be used. The counterpoise should be elevated six or seven feet above the ground so it will not interfere with persons walking under it. Connection is made between the usual ground terminal of the transmitter and each of the wires in the counterpoise.

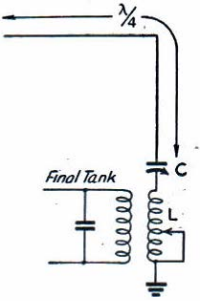


Fig. 1048 — Typical grounded antenna for 1.75 Mc., consisting of a vertical section and horizontal section having a total length (including the ground lead if the latter is more than a few feet long) of one-quarter wavelength. Coil L should have about 20 turns of No. 12 on a three-inch diameter form, tapped every two or three turns for adjustment. C is 250 to 500 $\mu\mu fd.$ variable. The inductive coupling between L and the final tank coil should be variable.

"J" antenna — This antenna, frequently used on ultra-high frequencies when vertical polarization is desired, is simply a half-wave radiator fed through a quarter-wave matching section, (§ 10-8), the whole being mounted vertically as shown in Fig. 1049. Adjustment and tuning are as described in § 10-8. The bottom of the matching section, being practically at zero r.f. potential, can be grounded through a metallic conductor for lightning protection.

Coaxial antenna — With the "J" antenna there is likely to be some radiation from the matching section and transmission line which tends to combine with the radiation from the antenna in such a way as to raise the angle of radiation. As this is undesirable on ultra-high frequencies where the lowest possible radiation angle is essential, the coaxial antenna shown in Fig. 1050 was developed to eliminate feeder radiation. The center conductor of a 70-ohm concentric transmission line is extended one quarter wave beyond the end of the line to act as the upper half of a half-wave antenna, the lower half being supplied by the quarter-wave sleeve, the upper end of which is connected to the outer conductor of the concen-

tric line. The sleeve acts as a shield about the transmission line and very little current is induced on the outside of the line by the antenna field. The line is non-resonant, since its characteristic impedance is the same as the center impedance of the half-wave antenna

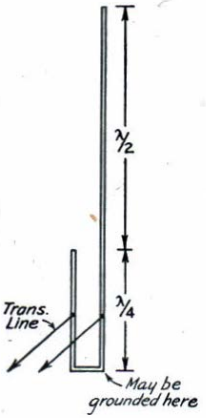


Fig. 1049 — The "J" antenna. It is usually constructed of metal tubing; frequently with the ¾-wave vertical section shown an extension of a grounded metal mast. The stub may be adjusted by a sliding shorting bar.

(§ 10-2). The sleeve may be made of copper or brass tubing of suitable diameter to clear the transmission line. The coaxial antenna is somewhat difficult to construct, but is superior to simpler systems at low radiation angles.

Folded dipole — An arrangement which combines the radiation characteristics of a half-wave antenna with the impedance-transforming properties of a quarter-wave line

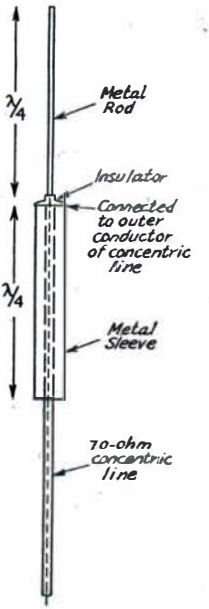


Fig. 1050 — Coaxial antenna. The inner conductor of the 70-ohm line is connected to the quarter-wave metal rod which forms the upper half of the antenna.

(§ 10-5) is shown in Fig. 1051. Essentially, it consists of a center-fed half-wave antenna with another half-wave element connected directly between its ends. The spacing between the two sections should be quite close — not more than a few per cent of the wavelength. As used at ultra-high frequencies, the spacing is of the order of an inch or two with elements constructed of metal tubing.

The impedance at the terminals of the antenna is four times that of a half-wave antenna, or nearly 300 ohms, when the antenna conductors are all the same diameter. A 300-ohm line will therefore be non-resonant when the antenna is connected to its output end (§ 10-5), while the standing-wave ratio with a 600-ohm line will be only 2 to 1.

The total length around the loop formed by the antenna may be calculated by Equation 10 (§ 10-9) for a total length of one wavelength.

Corner reflector antenna — A type of antenna system particularly well-suited to the u.h.f. ranges above 56 Mc., is the "corner" reflector shown in Fig. 1052. It consists of two plane surfaces set at an angle of 90°, with the antenna set on a line bisecting this angle.

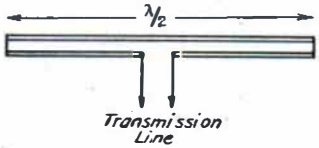


Fig. 1051 — Folded dipole for increasing the value of impedance at the feed point.

The distance of the antenna from the vertex should be 0.5 wavelength, but some compromise designs can be built with closer spacings (see Table V). The plane surfaces do not need to be solid, and can most easily be made of spines spaced about 0.1 wavelength apart. The spines do not have to be connected together electrically.

The resistance of the antenna is raised when a corner reflector is used. The transmission line should be run out at the rear of the reflector to keep the system as symmetrical as possible and thus avoid any unbalance. Two simple antennas which can be used with the corner reflector are shown in Fig. 1053.

The corner reflector can be used with the antenna either horizontal or vertical, and the plane of polarization will be the plane of the antenna. The relative positions of the antenna and reflector must remain the same, however, which means that a support for both horizontal and vertical polarization would require a means for rotating the reflector about its horizontal axis.

Receiving antennas — Because of the high sensitivity of modern receivers a large antenna

is not necessary for picking up signals at good strength. Often it will be found that an indoor wire only 15 or 20 feet long will give quite good results, although a longer wire outdoors is better.

The use of a tuned antenna greatly improves the operation of the receiver because the signal strength is greater in proportion to the stray noises picked up by the antenna than is the case with the antenna of random length. Since the transmitting antenna is usually given the best location, it can be used to great advantage for receiving, especially when a directive antenna is used. A change-over switch or relay connected in the antenna leads can be used to transfer the connection from the receiver to the transmitter while the transmitter is on the air. The directive effects and power gain of directive transmitting antennas are the same for receiving as for transmitting, and should be utilized for best reception.

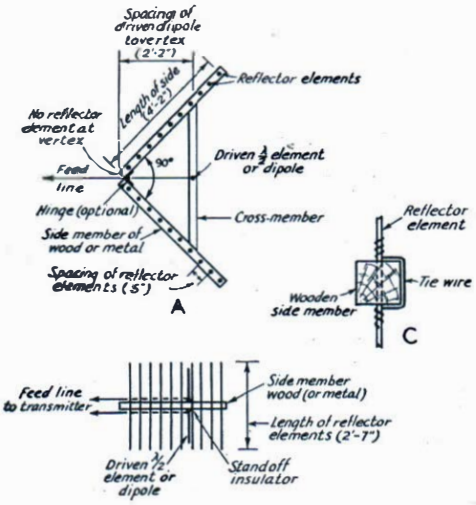


Fig. 1052 — A corner reflector antenna system with grid-type reflector. The reflector elements are stiff wire or tubing. The dimensions are for 224 Mc., and should be doubled for 112 Mc. (See Table V.) The gain of the system is close to 10 db.

TABLE V

Frequency Band	Length of Side	Length of Reflector Elements	Number of Reflector Elements	Spacing of Reflector Elements	Spacing of Driven Dipole to Vertex
224–230 Mc. (1¼ meter)	4′ 2″	4′ 7″	20	5″	2′ 2″
112–116 Mc. (2½ meter)	8′ 4″	5′ 2″	20	10″	4′ 4″
112–116 Mc.* (2½ meter)	6′ 8″	5′ 2″	16	10″	3′ 6″
56–60 Mc. (5 meter)	16′ 8″	10′ 4″	20	1′ 8″	8′ 8″
56–60 Mc.* (5 meter)	13′ 4″	10′ 4″	16	1′ 8″	6′ 11″

Table V. — Dimensions of square-corner reflector for the 224-, 112-, and 56-Mc. bands. Alternative designs are listed for the 112- and 56-Mc. bands. These designs, marked (*), have fewer reflector elements and shorter sides, but the effectiveness is only slightly reduced. There is no reflector element at the vertex in any of the designs.

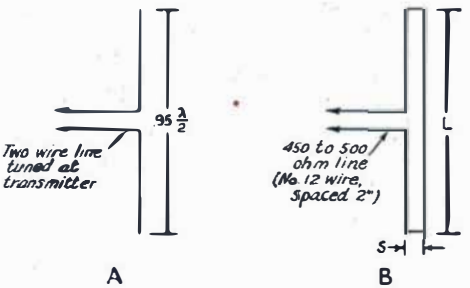


Fig. 1053 — Dipoles suitable for use with the corner reflector antenna system. The length *L* is 25 inches for 224 Mc., *s* = 1 inch for the same band.

Radio Equipment

● A ONE-TUBE REGENERATIVE RECEIVER

THE SIMPLEST receiver capable of giving at all satisfactory results in everyday operation is one consisting of a regenerative detector followed by an audio amplifier. This type of receiver is sufficient for headphone reception, and is quite easy to build and adjust. A dual tube may be used for both stages, thereby reducing cost.

Figs. 1101 to 1105 show such a receiver, using a 6C8G twin-triode tube, one triode section being the regenerative detector and the other the audio amplifier. The circuit diagram is given in Fig. 1103. The grid coil, L_1, is tuned to the frequency of the incoming signal by means of condensers C_1 and C_3, C_1 being the bandsetting or general coverage condenser and C_3 the bandspread condenser. Regeneration is supplied by means of the tickler coil L_2; the variable plate by-pass condenser, C_2, is the regeneration control. The receiver is coupled to the antenna through C_5, a low-capacity trimmer condenser. R_1 and C_4 are the grid leak and grid condenser.

The audio amplifier section of the tube is coupled to the detector by the audio transformer T_1. Bias for the audio stage is supplied by a midget flashlight cell, this type of bias being quite convenient as well as cheaper than other methods. The choke, *RFC*, is necessary to prevent r.f. current from flowing in the primary winding of the audio transformer; without the choke the regeneration control condenser C_2 may be ineffective. A switch, S_1, is provided for turning off the "B" supply when transmitting.

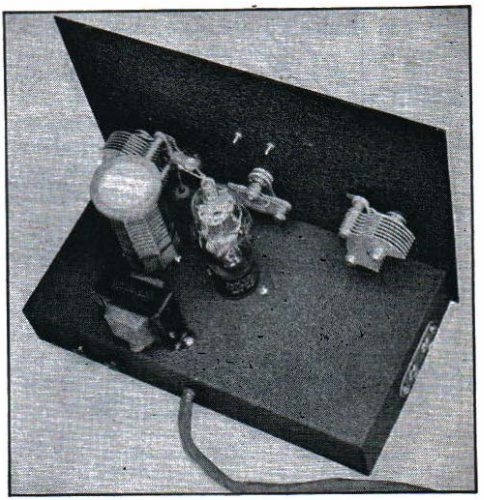


Fig. 1102 — A rear view of the one-tube receiver. The grid condenser and grid leak are supported by their wire leads between the stator plates of the tuning condenser and the grid cap on the tube.

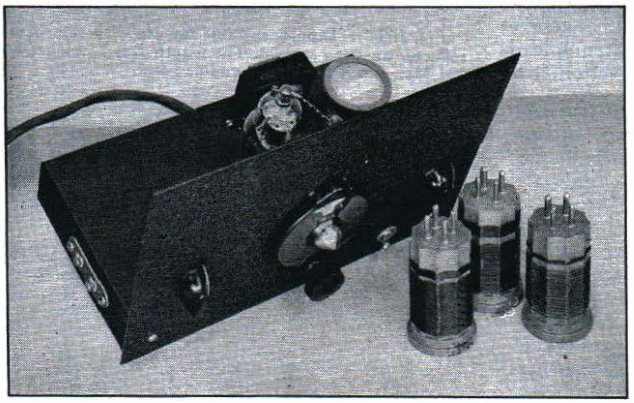


Fig. 1101 — A one-tube regenerative receiver, using a double triode as a regenerative detector and audio amplifier. Plug-in coils for different frequency ranges are shown in front of the receiver.

The construction of the receiver is shown in the photographs. The chassis measures 5½ by 9½ by 1½ inches. The three variable condensers are mounted on the panel three inches from the bottom edge, with C_3 in the center, C_1 at the right and C_2 at the left. All ground connections may be made directly to the chassis, making sure that the paint is scraped away and that good contact is secured.

The headphone connections are made by means of tip jacks mounted on the rear edge of the chassis. Filament and plate power are brought in through a four-wire cable which enters the chassis through the rear edge.

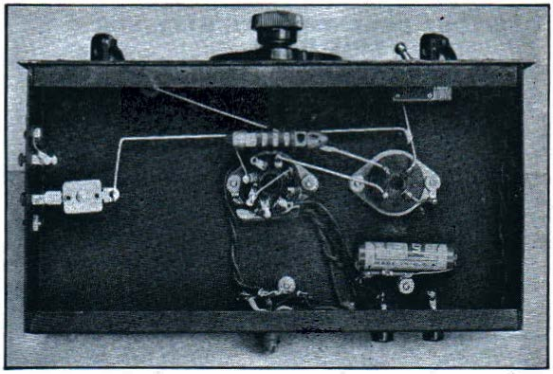


Fig. 1103 — Bottom of chassis view of the one-tube regenerative receiver. The construction and wiring are extremely simple. The power supply cable and the headphone tip jacks are brought out at the rear of the chassis, and the antenna and ground terminals are mounted on one side.

The coils are made as shown in Fig. 1105 and the coil table. Both windings should be in the same direction. L_1 for the B, C and D coils should have its turns evenly spaced to occupy the specified length; the wire may be held in place when the coil is finished by running some Duco cement along the ridges of the coil forms.

The heater supply for the receiver may be either a 6.3-volt filament transformer (the 1-ampere size will be ample) or a 6-volt battery. A 45-volt "B" battery should be used for the plate supply. The "B" current drain is only a few milliamperes, and a medium- or small-size "B" battery will give excellent service.

After the wiring has been checked, the heater and plate supplies, headphones and antenna and ground can be connected to the receiver, and the C coil plugged in. Turn the regeneration condenser, C_2, starting from minimum capacity (plates all out) until the set goes into oscillation. This phenomenon is easily recognizable by a distinct click, thud or hissing sound. The point where oscillation just begins is the most sensitive operating point at that particular dial setting.

The tuning dial may now be slowly turned, the regeneration control knob being varied simultaneously (if necessary) to keep the set just oscillating. A number of stations will probably be heard. A little practice will make tuning easy.

If the set refuses to oscillate, the sensitivity will be poor and no code signals will be heard on the frequencies at which such signals should be expected. It should oscillate easily, however, if the coils are made exactly as shown. It sometimes happens that the antenna takes so much energy from the set that it cannot oscillate, this usually resulting in "holes" in the range where no signals can be picked up (and where the hissing sound cannot be obtained). This can be cured by reducing the capacity of C_5 (unscrewing the adjusting screw) until the detector again oscillates. If it still refuses to oscillate, the coil L_2 must be moved nearer to L_1 or, in extreme cases, a turn or two must be added to L_2. This is best done by rewinding with more turns rather than by trying to add a turn or two to the already-wound coil. For any given band of frequencies, adjust C_5 so that the detector oscillates over the whole range, using as much capacity at C_5 as is possible. This will give the best compromise between dead spots and signal strength. It will be found that less advancing of the regeneration control, C_2, is required at the high-frequency end of a coil range (C_1 at or near minimum capacity) than at the low-frequency end. The best adjustment of the antenna condenser, C_5, and the feedback coil, L_2, is that which requires almost a maximum setting of the regeneration control at the low-frequency end (maximum capacity of C_1) of any coil range.

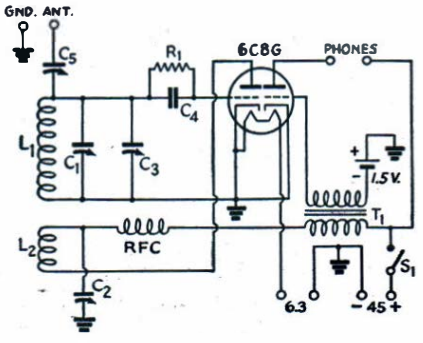


Fig. 1104 — Circuit diagram of the one-tube regenerative receiver.

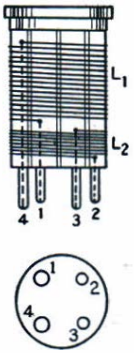
C_1 — 100-$\mu\mu$fd. band-set variable.
C_2 — 100-$\mu\mu$fd. regeneration control variable.
C_3 — 15-$\mu\mu$fd. band-spread variable.
C_4 — 100-$\mu\mu$fd. mica grid condenser.
C_5 — 3-30-$\mu\mu$fd. adjustable mica antenna coupling.
R_1 — 1-megohm, ½-watt grid leak.
L_1, L_2 — Grid coil and tickler coil. See coil table for dimensions.
T_1 — Interstage audio transformer, 3:1 ratio.
S_1 — S.p.s.t. B+ toggle switch.
RFC — 2.5-mh. detector plate r.f. choke.

ONE-TUBE REGENERATIVE RECEIVER COIL DATA

Coil	Grid Winding (L_1)	Tickler (L_2)
A	56 turns No. 22 enamelled	15 turns No. 24 enamelled
B	32 " " " "	8 " " " "
C	18 " " " "	5 " " " "
D	10 " " " "	5 " " " "

All coils wound on 1½-inch diameter forms (Hammarlund SWF-4). Grid windings on coils B, C and D spaced to occupy a length of 1½ inches; grid winding on coil A close-wound. Tickler coils all close-wound, spaced ⅛ inch from bottom of grid winding. See Fig. 1105.

Frequency range Coil: A — 1700 to 3200 kc.
B — 3000 to 5700 kc.
C — 5400 to 10,000 kc.
D — 9500 to 18,000 kc.



BOTTOM OF SOCKET OR COIL FORM

Fig. 1105 — Method of winding coils for the one-tube regenerative receiver. Pin 1 connects to ground, pin 2 to the plate of the detector, pin 3 to RFC and the stator plates of C_2, and pin 4 to the stator plates of C_1 and C_3.

Coil A misses the high-frequency end of the broadcast band, but it is possible to hear police stations and other services. The band is most easily located by listening at night (when there is the most activity), setting C_3 at maximum and slowly tuning with C_1 until some of the police stations are heard. These stations operate on 1712 kc., so that once found they become "markers" for the low-frequency end of the band. Further tuning then should be done with the main tuning dial.

Locating the other ranges is done in exactly the same manner, by searching carefully with C_1 and looking for marker stations of known frequencies. Activity will be greater above 12 Mc. during the day and below that frequency at night.

A suitable antenna for the receiver would be 50 to 75 feet long, and as high and clear of surrounding objects as possible. The ground lead should preferably be short; a ground to a heating radiator or water piping is usually good.

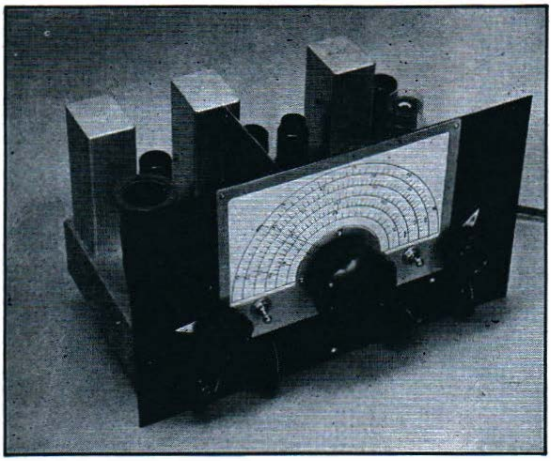


Fig. 1106 — A 7-tube superhet using regeneration to give single-signal reception and an improved image ratio. The dial is of a type that can be directly calibrated for each amateur band. Two plug-in coils are used for each range. The chassis is 11 by 7 by 2 inches and the panel 7 by 12 inches.

• A REGENERATIVE SINGLE-SIGNAL RECEIVER

An inexpensive amateur-band receiver using i.f. regeneration for single-signal reception, is shown in Fig. 1106. Fig. 1108 gives the circuit diagram. Regeneration also is used in the mixer circuit to improve the signal-to-image ratio and to give added gain. This receiver is designed to give the maximum of performance, in the hands of a capable operator, at minimum cost. Selectivity, stability and sensitivity are primary considerations.

The mixer, a 6SA7, is coupled to the antenna and is separately excited by a 6J5 oscillator. There is a single 460-kc. i.f. stage, using a 6SK7 and permeability-tuned transformers. The second detector and first audio amplifier is a 6SQ7 and the audio output tube for loud-speaker operation is a 6F6. The separate beat oscillator circuit uses a 6C5. A VR-105 voltage regulator tube is used to stabilize the plate voltage on the oscillators and the screen voltage on the mixer and i.f. tubes.

To make construction easy and to avoid the necessity for additional trimmer condensers on each coil, the mixer and high-frequency oscillator circuits are separately tuned. Main tuning is by the oscillator band-spread condenser, C_3, which is operated by the calibrated dial. C_2 is the oscillator band-setting condenser. The mixer circuit is tuned by C_1, and regeneration in this circuit is

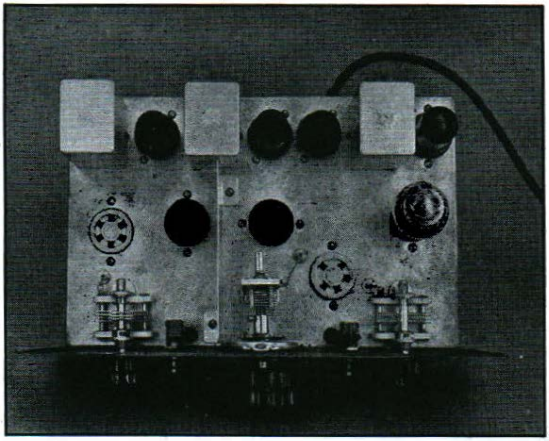


Fig. 1107 — Top view of the 7-tube superhet without coils in place. The band-spread tuning condenser, C_3, is at the front center; at the left is the mixer tuning condenser and at the right the oscillator band-set condenser. The oscillator tube is in the center, with the mixer tube to the left on the other side of a baffle shield. This shield, measuring 4¼ by 4½ inches, is used to prevent coupling between oscillator and mixer. The mixer coil socket is at the left edge of the chassis, and the oscillator coil socket is next to C_3.

The i.f. and audio sections are along the rear edge of the chassis. The transformer in the rear left corner is T_1; next to it is the i.f. tube, then T_2. Next in line is the 6SQ7, followed by the 6C5 beat oscillator, the b.o. transformer, T_3, and finally the 6F6. The VR-105 is just in front of T_3.

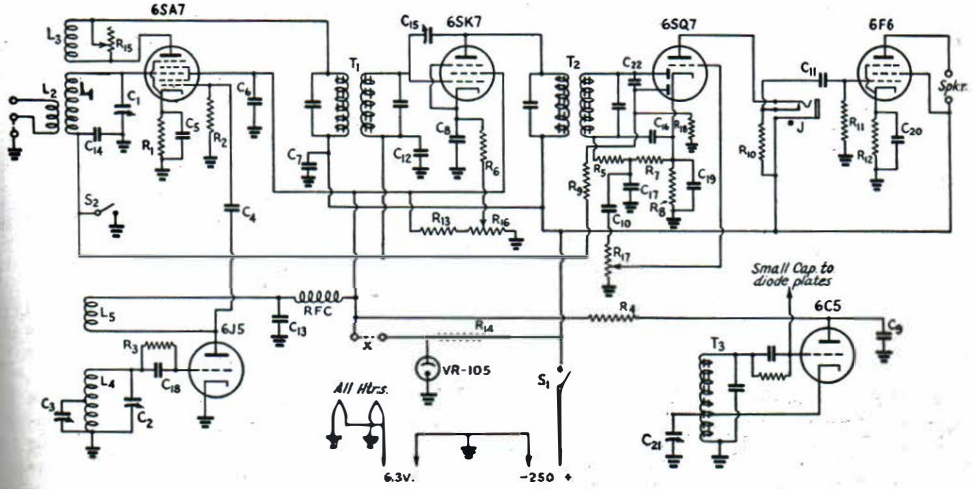


Fig. 1108 — Circuit diagram of the regenerative superhet.

C_1 — 50-$\mu\mu$fd. mixer tuning.
C_2 — 50-$\mu\mu$fd. oscillator band-set.
C_3 — 35-$\mu\mu$fd. oscillator band-spread.
C_4 — 50-$\mu\mu$fd. mica oscillator coupling.
C_5, C_8 — 0.1-μfd. cathode by-pass.
C_6 — 0.1-μfd. mixer screen by-pass.
C_7 — 0.1-μfd. mixer plate by-pass.
C_9 — 0.01-μfd. b.f.o. plate by-pass.
C_{10}, C_{11} — 0.01-μfd. audio coupling.
C_{12} — 0.01-μfd. i.f. amplifier grid by-pass.
C_{13} — 0.005-μfd. oscillator plate by-pass.
C_{14} — 0.005-μfd. mixer grid by-pass.
C_{15} — 3-30-$\mu\mu$fd. adjustable i.f. feedback.
C_{16} — 250-$\mu\mu$fd. diode by-pass.
C_{17} — 100-$\mu\mu$fd. r.f. filter.
C_{18} — 100-$\mu\mu$fd. oscillator grid.
C_{19}, C_{20} — 25-μfd. electrolytic cathode by-pass.
C_{21} — 25-$\mu\mu$fd. b.f.o. tuning.
C_{22} — 100-$\mu\mu$fd. mica diode blocking.
R_1 — 200-ohm, ½-watt cathode bias.
R_2 — 20,000-ohm, ½-watt injection grid leak.
R_3 — 50,000-ohm, ½-watt oscillator grid leak.
R_4 — 50,000-ohm, ½-watt b.f.o. plate dropping.
R_5 — 50,000-ohm, ½-watt diode r.f. filter.
R_6 — 300-ohm, ½-watt cathode bias.
R_7 — 0.2-megohm, ½-watt diode load.
R_8 — 2000-ohm, ½-watt cathode bias.
R_9 — 1.0-megohm, ½-watt a.v.c. filter.
R_{10} — 0.1-megohm, ½-watt plate load.
R_{11} — 0.5-megohm, ½-watt grid.
R_{12} — 450-ohm, 1-watt cathode bias.
R_{13} — 75,000-ohm, 1-watt bleeder.
R_{14} — 5000-ohm, 10-watt voltage dropping.
R_{15} — 10,000-ohm mixer regeneration control.
R_{16} — 25,000-ohm i.f. gain control.
R_{17} — 2-megohm audio volume control.
R_{18} — 2-megohm, ½-watt a.v.c. load.
T_1, T_2 — 465-kc. i.f. transformer, permeability tuned.
T_3 — 465-kc. b.f.o. unit.
RFC — 2.5-mh. r.f. choke.
J — Closed-circuit jack.
S_1 — S.p.s.t. B+ toggle switch.
L_1-L_5, inc. — See coil table.
X indicates jumper inside VR-105 base.

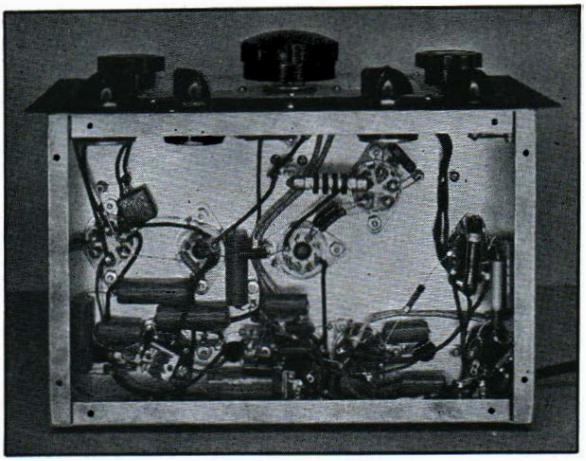


Fig. 1109 — The below-chassis wiring is shown in this view of the 7-tube superhet. The controls along the bottom edge of the panel are, from left to right, the mixer regeneration control, R_{15}, the i.f. gain control, R_{16}, the audio volume control, R_{17}, and the beat-oscillator tuning condenser, C_{21}. The latter has one corner of one rotary plate bent over so that when the condenser plates are fully meshed the condenser is short-circuited, thus stopping oscillation.

controlled by R_{15}, connected across the mixer tickler coil, L_3.

R_{16} is the i.f. amplifier gain control, which also serves as an i.f. regeneration control when this stage is made regenerative. C_{15} is the regeneration condenser; it is adjusted to feed back a small amount of i.f. energy from the plate to the grid of the 6SK7 and thus produce regeneration. If the high selectivity afforded by i.f. regeneration is not wanted, C_{15} may be omitted.

Diode rectification is used in the second detector circuit. One of the two diode plates in the 6SQ7 is used for developing a.v.c. voltage, being coupled through C_{22} to the detector diode. The detector load resistor consists of R_5 and R_7 in series, the tap being used for r.f. filtering of the audio output to the triode section of the tube. R_{18} is the a.v.c. load resistor; R_9, C_{14} and C_{12} constitute the a.v.c. filter circuit. S_2 cuts the a.v.c. out of circuit by grounding the rectifier output. The headphones connect in the plate circuit of the triode section of the 6SQ7. R_{17} is the audio volume control.

The tube heaters are all in parallel, one side of each being grounded right at the tube socket. Only one filament wire need be run from tube to tube.

In wiring the i.f. amplifier, keep the grid and plate leads from the i.f. transformers fairly close to the chassis and well separated. Without C_{15}, the i.f. stage should be perfectly stable and should show no tendency to oscillate at full gain.

The method of winding the coils is shown in Fig. 1110 and complete specifications are given in the coil table.

The i.f. amplifier can be aligned most conveniently with the aid of a modulated test oscillator. First alignment should be made with C_{15} disconnected so that the performance of the amplifier non-regenerative can be checked. A headset or loud speaker can be used as an output indicator. The mixer and oscillator coils should be out of their sockets, and R_{15} should be set at zero resistance.

After the i.f. amplifier is aligned, plug in a set of coils for some range in which there is likely to be a good deal of activity. Set the oscillator padding condenser, C_2, at almost maximum capacity and the mixer regeneration control, R_{15}, for minimum regeneration — no resistance in circuit. Connect an antenna. Switch on the beat oscillator by turning C_{21} out of the maximum position, and adjust the screw on T_3 until the characteristic hiss is heard.

As C_1 is tuned over its range, there should be two points where there will be a definite increase in noise and in the strength of any signals which may be heard. The peak on the low-capacity side corresponds to the image frequency, and the mixer condenser should always be tuned to the peak which occurs at the higher capacity setting.

After the signal peak on C_1 has been identified, tune C_3 over its whole range, following with C_1 to keep the mixer circuit in tune, to see how the band fits the dial. With C_2 properly set, the band edges should fall the same number of main dial divisions from 0 and 100; if the band runs off the low-frequency edge, less capacity is needed at C_2, while the converse is true if the band runs off the high edge. Once the band is properly centered on the dial, the panel may be marked at the appropriate point so that C_2 may be reset readily when changing bands.

To check the operation of the mixer regeneration, tune in a signal on C_3, adjust C_1 for maximum volume, and slowly advance the regeneration control, R_{15}. As the resistance

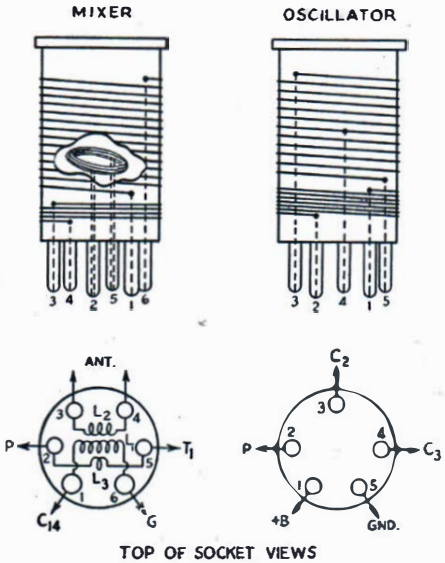


Fig. 1110 — Coil and socket connections for the 7-tube superhet. The small coil inside the mixer coil must be oriented properly for regeneration to take place.

increases retune C_1 to maximum, since the regeneration control will have some effect on the mixer tuning. As regeneration is increased, signals and noise will both become louder and C_1 will tune more sharply. Finally the mixer circuit will break into oscillation when, with C_1 right at resonance, a loud carrier will be heard since the oscillations generated will go through the receiver in exactly the same way as an incoming signal. As stated before, oscillation should occur with R_{15} set at half to three-quarters full scale. In practice, always work with the mixer somewhat below the critical regeneration point and never permit it actually to oscillate. On the lower frequencies, where images are less serious, the tuning is less critical if the mixer is non-regenerative. In this case, always set R_{15} at zero since there will be a range on the resistor where, without definite regeneration, the signal strength will be less than it is with zero resistance.

After the preceding adjustments have been completed the i.f. regeneration may be added. Install C_{15}, taking out the adjusting screw and bending the movable plate to make an angle of about 45 degrees with the fixed plate. Realign the i.f. As the circuits are tuned to resonance the amplifier will oscillate, and each time this happens the gain control, R_{16}, should be backed off until oscillations cease. Adjust the trimmers to give maximum output with the lowest setting of R_{16}. At peak regeneration the signal strength should be about the same, despite reduced gain in the amplifier, as without regeneration at full gain. Too much gain with regeneration will have an adverse effect on selectivity.

COIL DATA FOR 7-TUBE SUPERHET

Band	Coil	Wire Size	Turns	Length	Tap
1.75 Mc.	L_1	24	70	Close-wound	—
	L_2	24	15	" "	—
	L_3	22	15	——	—
	L_4	22	42	Close-wound	Top
	L_5	24	15	" "	—
3.5 Mc.	L_1	22	35	" "	—
	L_2	22	9	" "	—
	L_3	22	12	——	—
	L_4	22	25	1 inch	18
	L_5	22	10	Close-wound	—
7 Mc.	L_1	18	20	1 inch	—
	L_2	22	5	Close-wound	—
	L_3	22	9	——	—
	L_4	18	14	1 inch	6
	L_5	22	6	Close-wound	—
14 Mc.	L_1	18	10	1 inch	—
	L_2	22	5	Close-wound	—
	L_3	22	7	——	—
	L_4	18	7	1 inch	2.4
	L_5	22	4	Close-wound	—
28Mc.	L_1	18	4	1 inch	—
	L_2	22	4	Close-wound	—
	L_3	22	1.5	——	—
	L_4	18	3.6	1 inch	1.4
	L_5	22	2.4	Close-wound	—

All coils except L_3 are 1½ inches in diameter, wound with enamelled wire on Hammarlund SWF Forms. Spacing between L_1 and L_2, and between L_4 and L_5, approximately ⅛ inch. Band-spread taps are measured from bottom (ground) end of L_4.

L_3 for 28 Mc. is interwound with L_1 at the bottom end. L_3 for all other coils is self-supporting, scramble-wound to a diameter of ¾ inch, mounted inside the coil form near the bottom of L_1.

For single-signal c.w. reception, set the beat oscillator so that when R_{16} is advanced to make the i.f. just go into oscillation the resulting tone is the desired beat-note frequency. Then back off on R_{16} to obtain the desired degree of selectivity. Maximum selectivity will be secured with the i.f. just below the oscillating point. The "other side of zero beat" will be very much weaker than the desired side.

Power supply requirements for the receiver are 2.2 amp. at 6.3 volts for the heaters and 80 ma. at 250 volts for the plates. Without the pentode output stage a supply giving 6.3 volts at 1.5 amp. and 250 volts at 40 ma. will be sufficient.

• COMMERCIAL SUPERHETERODYNE RECEIVERS

Although some very advanced types of superheterodyne receivers have been built at home, most of the superheterodynes used by amateurs are of the manufactured type. They range from simple, inexpensive ones with a minimum of tubes to large receivers with many

features. Two of the latter type are shown in Figs. 1111–1116.

The National HRO, shown in Fig. 1111, 1112 and 1113, is a high-gain superheterodyne using plug-in coil gangs for the various frequency ranges. Coils are available for bands of frequencies between 50 kc. and 30 Mc., a range made possible only by the use of plug-in coil gangs. Standard equipment is four sets of coils for 1.7 to 30 Mc. Provision is made on the coil gangs in the high-frequency range for either general coverage or amateur bandspread, depending upon the coil connections used. Two stages of r.f. amplification are used ahead of the mixer, and a separate high-frequency oscillator is used for greater stability than is possible with some of the combination converter types of tubes. The mixer is followed by a crystal filter and two stages of i.f. amplification. The crystal filter is responsible for a high order of selectivity and is used for "single signal" reception of c.w. signals. A double diode, hi-μ pentode is used for the second detector and first audio stage, and is followed by a pentode audio output tube. A beat-frequency oscillator is loosely coupled to the diode second detector.

The coil gang plugs in at the lower center of the panel, as can be seen in Fig. 1111. The controls on the panel, aside from the main tuning dial, are audio and r.f. gain, beat-frequency-oscillator frequency control, and selectivity and phasing controls for the crystal filter. An "on-off" switch is included for turning off the receiver during transmission periods, and another switch turns on the a.v.c. when radiophone signals are being received. A third switch, mounted on the b.f.o. control, turns off the b.f.o. for 'phone reception when the control is set to the end of the scale. A tuning meter mounted on the panel can be switched in during 'phone reception for comparative measurements of carrier strength. The main tuning dial is an ingenious device which must make 10 complete revolutions for 180-degree rotation of the tuning condenser gang. The numbers, visible through small windows on the dial, change every revolution to give consecutive numbering by tens from 0 to 500.

A high order of frequency and alignment stability is obtained by the use of air-dielectric trimmer and padder condensers.

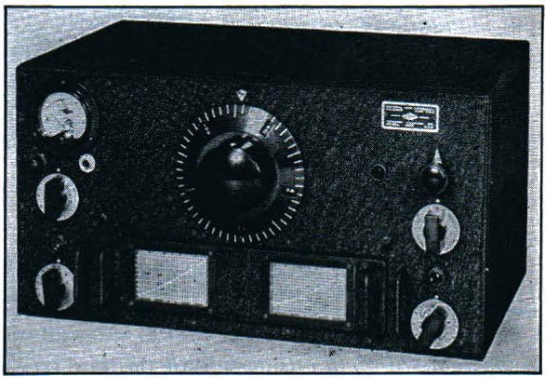


Fig. 1111 — The National HRO 9-tube superheterodyne. The gang of four coils is plugged in under the tuning dial.

The Hammerlund "Super Pro," shown in Figs. 1114, 1115 and 1116, is a 16-tube superheterodyne using coil-switching for band changing. Several models are available — the two common ones cover a range of 0.54 to 20 Mc. and 1.25 to 40 Mc. There are two stages of r.f. amplification ahead of the mixer, and a separate high-frequency oscillator is used. The mixer output feeds into a wide-range crystal filter followed by three stages of i.f. amplifica-

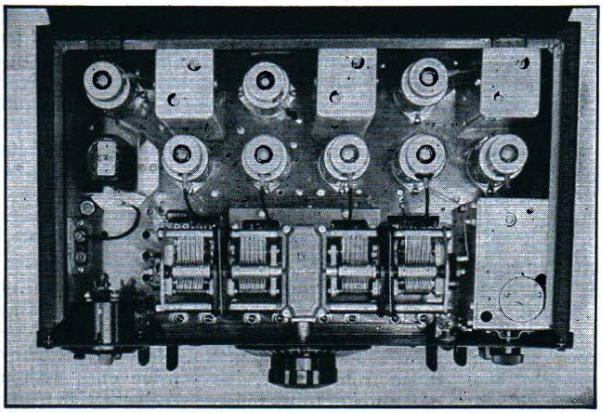


Fig. 1112 — A top view of the HRO receiver. The ganged tuning condensers can be seen either side of the dial mechanism.

tion. The crystal filter is adjustable in steps from a selectivity wide enough to admit 'phone signals to sharp enough for single-signal c.w. reception. To increase further the selectivity range of the i.f. amplifier, the i.f. transformers are given a variable-selectivity characteristic by making the coupling between coils adjustable from the panel. Thus, by turning the i.f. selectivity knob on the panel, the i.f. characteristic can be made broad enough for high-fidelity 'phone reception or sharp enough for something approaching single-signal performance.

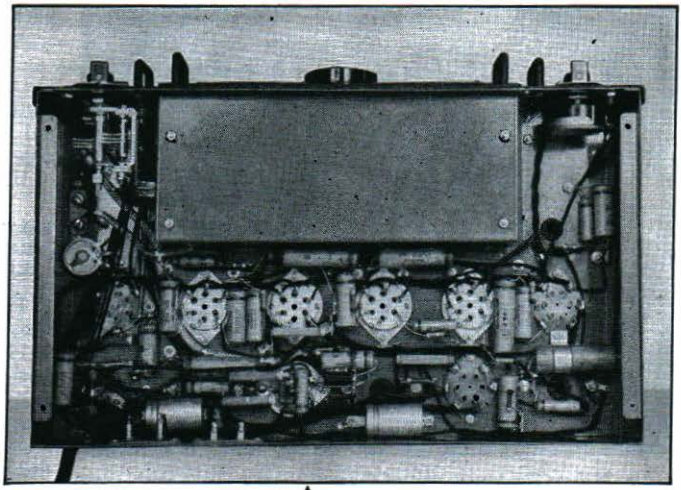


Fig. 1113 — This under-chassis view of the HRO shows the coil compartment and the wiring of the receiver.

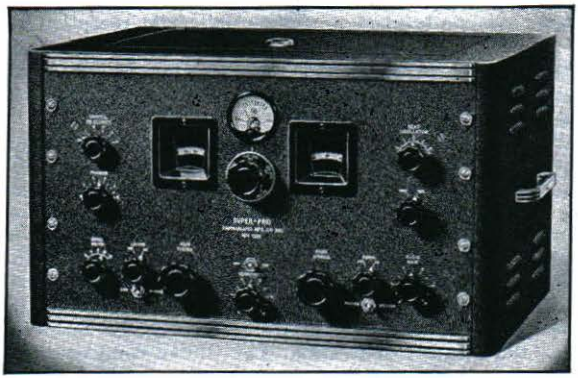
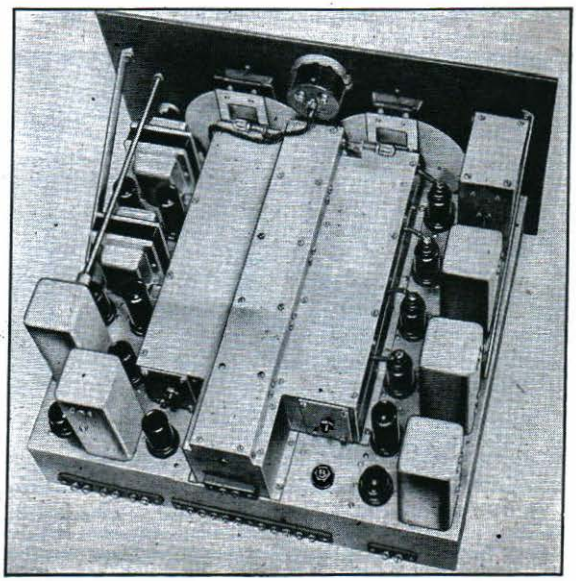


Fig. 1114 — The Hammarlund Super Pro 16-tube superheterodyne. The band-change switch is located between the two dial windows, just below the S meter.

Fig. 1115 — A top view of the Super Pro, showing the shielding of the coils and tuning condensers (center).

The i.f. amplifier is followed by a diode second detector, an audio noise limiter (useful for removing interference of the type caused by automobile ignition and similar sources), and two stages of audio amplification which drive a push-pull pentode output stage. The a.v.c. voltage is obtained from a separate diode rectifier which is driven by an a.v.c. amplifier tapped into the output of the i.f. amplifier. This use of "amplified" a.v.c. results in a wide range of control. A b.f.o. is coupled into the plate circuit of the last i.f. amplifier stage.

The power supply of the Super Pro is a separate unit connected to the receiver through a flexible cable. It is a heavy duty affair with a separate rectifier for the C-bias supply. This rectifier obtains its voltage from a tap on the power transformer. The field



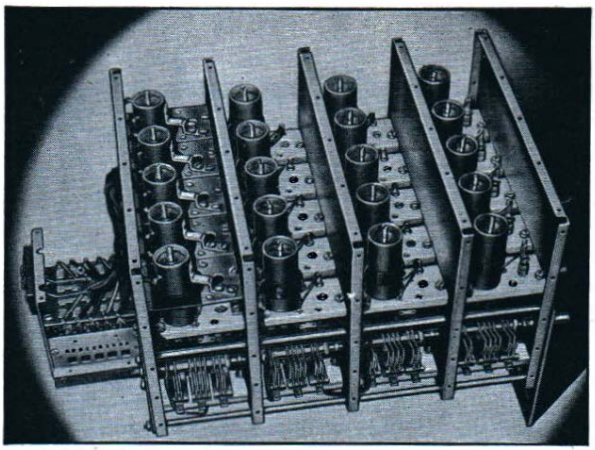


Fig. 1116 — The tuning unit of the Super Pro with the shield cover removed. Each coil is mounted on an Isolantite base along with its associated trimmer condenser.

of the loud speaker serves as one of the two filter chokes in the plate supply filter system.

Two separate gangs of tuning condensers are used for tuning the Super Pro. A "band set" gang is used to set the receiver in the portion of the spectrum one wishes to operate in, and the "band spread" gang is then used to tune slowly over this range. When the band spread dial, which is directly calibrated in frequency, is set to the high-frequency end of an amateur band, the band spread dial will read frequencies in that band. The manufacturer's tolerance for such calibration is less than 0.5% of the highest frequency in each range.

An "S" meter is included for reading comparative carrier strengths, and the sensitivity of the metering system can be adjusted by a control at the rear of the chassis so that the meter will indicate S9 on any signal from 10 to 10,000 microvolts, depending upon the requirements of the operator.

• AN ULTRA-HIGH-FREQUENCY SUPER-REGENERATIVE RECEIVER

The superregenerative receiver shown in Figs. 1117, 1118 and 1119 has excellent sensitivity in both the 112- and 224-Mc. amateur bands, but it is not entirely free from radiation as would be a similar receiver with an r.f. amplifier between the antenna and the detector. However, such a receiver will permit good reception in the u.h.f. range with a minimum of expense, and the radiation is not great enough to bother receivers more than a half mile or so away.

A 9002 u.h.f. triode is used in the detector circuit, followed by two stages of audio amplification. Only one stage of audio is used for headphone reception, and a jack is provided for plugging in the 'phones. The detector circuit is a form of the ultraudion circuit often used for u.h.f. detectors — the grid leak is returned to the positive plate supply rather than to the cathode, which results in slightly smoother superregeneration. The tuning condenser, C_1, is shunted by a mica trimmer, C_2, which is used as a band-set condenser and which allows

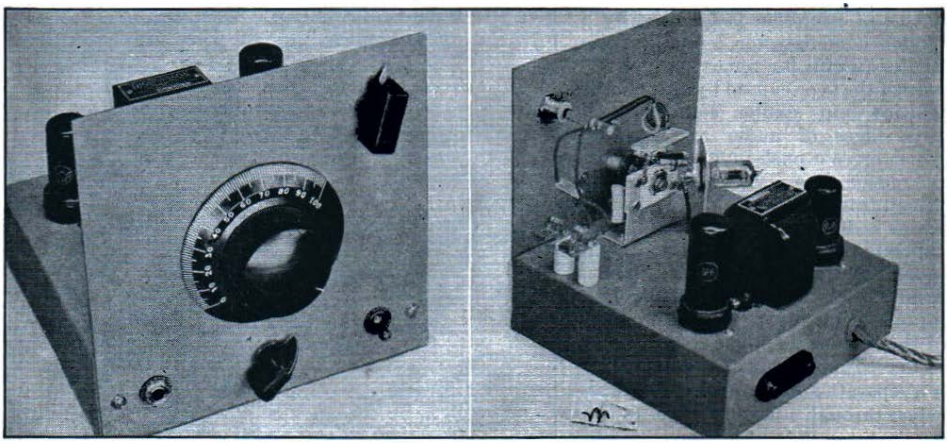


Fig. 1117 — Left — The panel of the two-band superregenerative receiver measures 7 inches square. The knob in the upper right-hand control adjusts the antenna coupling and the knob below the tuning dial controls the regeneration. Right — A view of the back of the two-band superregenerative receiver shows the variable antenna coupling and the placement of parts. Note the 224-Mc. coil in the foreground — the 112-Mc. coil is in the coil socket.

greater bandspread to be obtained with a reasonable value for C_1. Variable antenna coupling is used to aid in matching to different antennas and to make the use of plug-in coils easier. Regeneration can be increased by advancing the regeneration control, R_2, or by loosening the coupling between L_1 and L_2.

The receiver is built on a 7- by 7- by 2-inch chassis. The dial is mounted in the center of the panel and is connected to the tuning condenser by a bakelite flexible coupling. The condenser is mounted on a metal bracket cut out in the shape of a "U" to clear the stator connections of the condenser.

The socket for the plug-in coils is made from the contacts taken from a miniature tube socket. They are obtained by squeezing the socket in a vise until the bakelite cracks, after which they can be easily removed. One of these contacts is soldered to each of the tuning condenser connections and a third is soldered to a lug supported by one of the extra holes of the Isolantite base of the tuning condenser. The only care necessary in mounting the contacts is to see that they are all the same height, so that the plug-in coil will seat well on them. The band-set condenser is mounted by soldering short strips of wire to the ends and then soldering these wires to the tuning condenser terminals.

The polystyrene tube socket for the 9002 is mounted on a metal bracket which is placed close enough to the tuning condenser to allow a very short lead from the tuning condenser to the plate connection and just enough room between the rotor of the condenser and the grid connection of the tube for the grid condenser to fit. The heater and cathode leads are brought down to the underside of the chassis through a rubber grommet.

The variable antenna coupling coil is mounted on a polystyrene rod supported by a shaft bearing. The rod is prevented from moving axially in the bearing by cementing a fiber washer to the shaft and tightening the knob on the other side so that the shaft does not move too freely. The antenna coupling loop should be adjusted so that it will just clear the coils when they are plugged in the socket.

The coils are mounted on small strips of ⅛-inch polystyrene which have three small holes drilled in them corresponding exactly to the tops of the coil sockets. The coil is cemented to the strip with Duco cement at the points where the wire passes through the strip. The No. 18 wire used for the coils will fit snugly in the sockets if the sockets are pinched slightly. A coil socket of this type allows very short leads to be used. The coils are trimmed to the bands by spreading the turns slightly. The

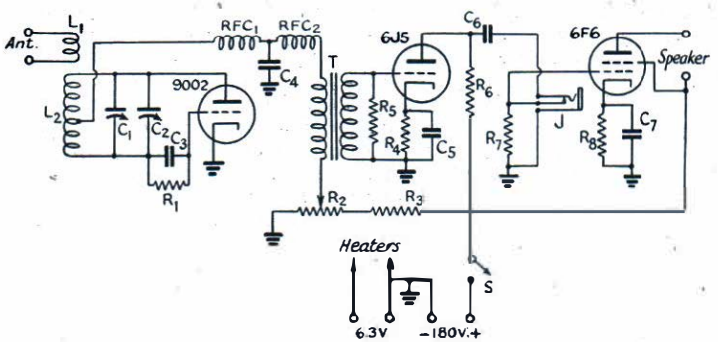


Fig. 1118 — Wiring diagram of the two-band superregenerative u.h.f. receiver.

C_1 — Two-plate band-spread tuning.
C_2 — 3-30-$\mu\mu$fd. adjustable mica band-set.
C_3 — 50-$\mu\mu$fd. grid condenser.
C_4 — 0.003-μfd. interruption-frequency by-pass.
C_5, C_7 — 10-μfd. electrolytic cathode by-pass.
C_6 — 0.01-μfd. audio coupling.
R_1 — 10-megohm, ½-watt grid leak.
R_2 — 50,000-ohm regeneration control.
R_3 — 0.1-megohm, 1-watt voltage dropping.
R_4 — 2500-ohm, ½-watt cathode bias.
R_5 — 0.1-megohm, ½-watt fringe-howl suppressor.
R_6 — 0.1-megohm, ½-watt plate load.
R_7 — 0.1-megohm, ½-watt grid.
R_8 — 500-ohm, 1-watt cathode bias.
J — Closed-circuit 'phone jack.
S — S.p.s.t. B+ toggle switch.
T_1 — Interstage audio transformer, 3:1 ratio.
RFC1 — U.h.f. r.f. choke.
RFC2 — 8-mh. interruption frequency r.f. choke.

L_1 — Antenna coil — 1 turn No. 14 enam. wire, ⅜-inch inside diam.
L_2 — Detector coil — 112 Mc.: 3 turns No. 18 enam., ½-inch diam., spaced to ¼-inch winding length. Tap 1¼ turns from plate end.
224 Mc.: 2 turns No. 18 enam., ¼-inch diam., spaced to ½-inch winding length. Tapped at center.

band-set condenser gives some further range of adjustment and, in the receiver as described, it is screwed down fairly tightly for the 112-Mc. band and loosened about four revolutions for 224 Mc.

Two things will be found to influence the sensitivity of the receiver, the value of C_4 and the degree of antenna coupling. It is recommended that values of C_4 from 0.001 to 0.005 μfd. be tried. The antenna coupling will, of course, vary greatly with the setting of the coil and with the type of antenna that is used, and it is well worth while to tune the antenna circuit and then vary the coupling with the panel control. Tight coupling will usually give better results than loose coupling, and the coupling can be increased almost up to the point where it is impossible to make the detector oscillate with no ill effects except increased radiation and QRM for other receivers in the vicinity.

No audio volume control was included in this receiver because the parts were held down to a minimum, but one could easily be added. In this receiver, the value of R_7 was adjusted until normal loud-speaker output was obtained, and it can be varied to meet anyone's particular requirements.

• A COMMERCIAL U.H.F. RECEIVER

The Hallicrafter S-27 u.h.f. receiver shown in Figs. 1120, 1121 and 1122 affords continuous coverage of 27 to 145 Mc. in three ranges of 27–46, 45–84 and 81–145 Mc. It is a 14-tube receiver of the superheterodyne type, with provision for the reception of either amplitude- or frequency-modulated signals.

Since ordinary receiving tubes are of little value above 60 Mc., special u.h.f. tubes of the "acorn" type are used for r.f. amplifier, mixer and high-frequency oscillator. A front-of-panel antenna trimmer control allows the input circuit to be matched closely to the antenna in use, for maximum gain. The mixer is followed by a two-stage 5.25-Mc. i.f. amplifier which has two degrees of selectivity. A switch on the panel allows the selectivity of the i.f. amplifier to be set at "sharp" for a.m. or narrow-band f.m. reception, or to "broad" for wide-band f.m. reception. Depending upon whether a.m. or f.m. reception is required, the output of the i.f amplifier is fed into a diode second detector or a limiter and discriminator (the latter for f.m. rectification). A b.f.o. is included for c.w. reception, and an audio noise limiter can be switched in for quieter reception of a.m. signals. The two detection methods are followed by audio amplifier stages ending in a push-pull output stage.

A dual-purpose "S" meter serves as a carrier-level meter for a.m. signals and as a tuning meter for f.m. signals. More so than in any other type of receiver, accurate tuning is a requisite of distortionless f.m. reception.

Several degrees of response can be obtained from the audio amplifier through the incor-

Fig. 1119 — Left — A close-up view of the tuning assembly shows how the leads from tuning condenser to tube socket have been kept short and how the coil socket is mounted on the tuning condenser. Hidden by the grid condenser (the 50-μμfd. condenser so prominent in the picture), the plate terminal of the tube socket goes to a lug that has been added to the stator of the tuning condenser. Right — The arrangement of parts under the chassis can be seen in this photograph. The 6J5 socket is on the left and the 6F6 socket is on the right, near the speaker terminals.

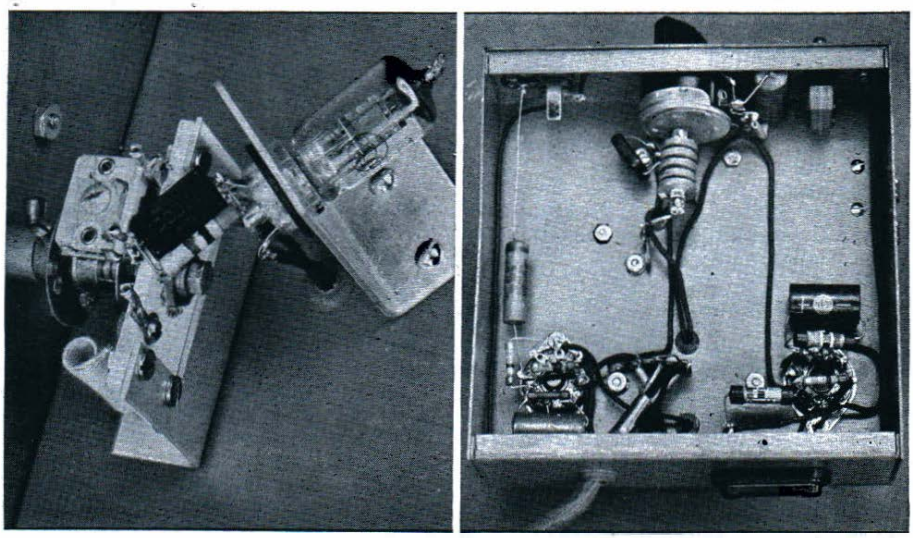




Fig. 1120 — The Hallicrafter S-27 u.h.f. receiver can be used for either a.m. or f.m. reception. The scale at the left indicates the frequency — the center scale is a vernier reading for accurate logging of signals.

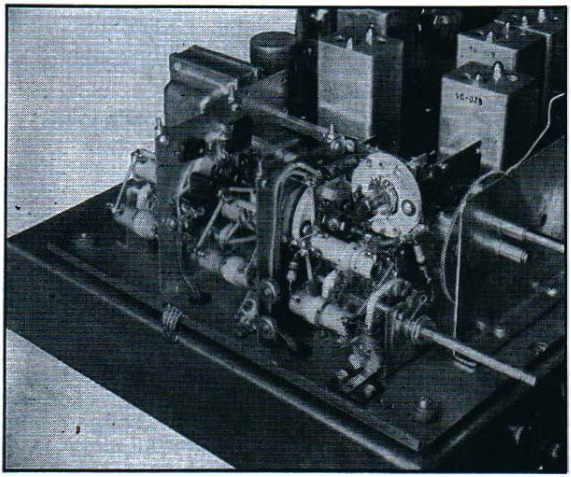


Fig. 1122 — A close-up view of the coil and condenser assembly of the S-27 with the shield cover removed. The acorn tube oscillator can be seen clearly — the mixer and r.f. stage acorn tubes are mounted on the interstage shields.

geared much higher and serves as a vernier indicator of the main dial setting.

Panel controls, other than the tuning, are band switch, a.f. gain, a.m.-f.m.switch, b.f.o. pitch control, tone selector switch, a.v.c. switch, noise limiter on-off switch, b.f.o. switch, send-receive switch and antenna trimmer control.

• RECEIVER POWER SUPPLIES

Power supplies of most manufactured radio receivers are built into

poration of audio filters in the receiver. Setting the tone-control switch at "low" attenuates the higher audio frequencies in the manner of the ordinary tone control. In the "normal" setting, a response similar to the ordinary broadcast receiver is obtained, while in the "high fidelity" position the high-frequency range is extended to include the higher frequencies transmitted by f.m. broadcast stations. A fourth "bass boost" position brings up the lower frequencies below 100 cycles slightly more than they would normally be, to compensate for loud speaker and other acoustical shortcomings.

Only one tuning control is used, in conjunction with two dials. One dial indicates the frequency in megacycles and rotates about 340 degrees for 180-degree rotation of the tuning condenser. The other dial is

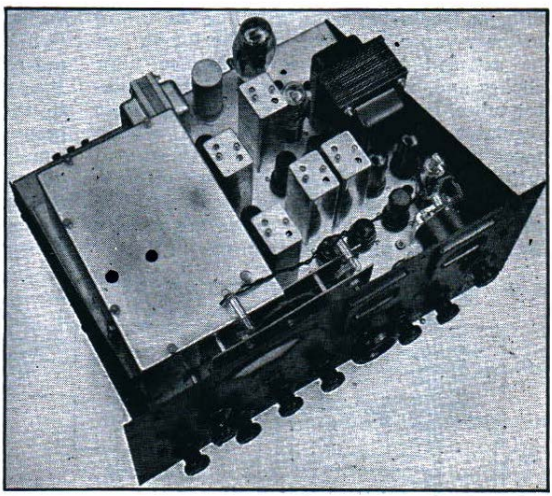


Fig. 1121 — A top view of the S-27, showing the shield covering the coil and condenser assembly and the gears of the dial mechanism.

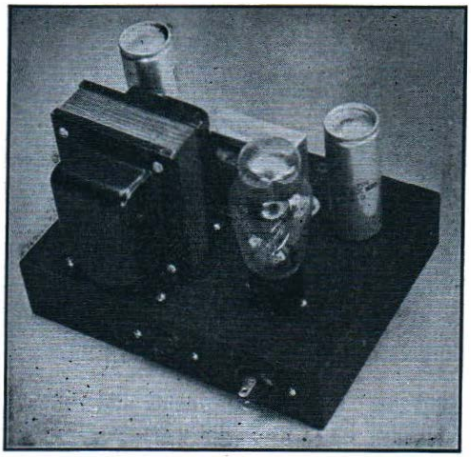


Fig. 1123 — This small receiver power supply will deliver 300 volts at 130 ma. with a choke-input filter and about 450 volts with a condenser-input filter. The chassis measures 7 by 9 by 2 inches. The circuit diagram is shown in Fig. 1124.

the receiver cabinet, thus making the receiver and power supply a single unit. On the other hand, it is often more convenient for the amateur who builds his receivers to make the power supply a separate unit so that it may be used with different receivers that may be built from time to time. In either case, the power supply requirements are substantially the same: correct voltage (within 10% or so), an adequate power capability, low hum and noise level and good regulation. The first two requirements are fulfilled by proper selection of the power transformer and current ratings of the filter chokes, the third depends upon the use of sufficient filter and, in some cases, avoiding the use of mercury-vapor rectifiers, and the fourth can be satisfied by using low-resistance filter chokes and, where necessary, voltage regulator tubes. It is not necessary, however, to use regulator tubes except where the voltage must be held quite close, as in the case of critical biases and the plate voltage of high-frequency and beat-frequency oscillators. For this reason, the regulator tubes are usually used to control only these voltages.

A typical heavy-duty amateur power supply is shown in Figs. 1123 and 1124. It is built upon a small metal chassis, with all of the wiring under the chassis. Heater and high-voltage connections are brought to a four-prong socket, so that the power cable from the receiver can be plugged in, and the connection to the 110-volt line is made through a male plug mounted on the side of the chassis. A switch mounted on one side of the chassis turns on the 110-volt supply to the primary of the transformer. As shown in the wiring diagram, the output voltage will run about 450, too high for most receiver use, but by removing the input condenser, C_1, the output voltage will drop to about 300, correct for most receiver work. Further reduction of voltage can of course be obtained by the use of a dropping resistor (but with an adverse effect on the regulation) or by the use of a lower-voltage transformer.

• A SIMPLE TETRODE OSCILLATOR TRANSMITTER

The unit shown in the photograph of Fig. 1125 represents one of the simplest types of amateur transmitters. The various components are assembled on a plain wooden baseboard.

A simple tetrode crystal-oscillator circuit (§ 4-4) is used and is shown in Fig. 1126. Parallel feed (§ 3-7) is used in both plate and grid circuits so that the only exposed high-voltage points are the plate-circuit r.f. choke and the high-voltage power terminal. Parallel plate feed also permits mounting the plate tank condenser, C_1, directly on the baseboard without insulation. Voltage for the screen is reduced to proper value by means of the dropping resistor, R_2. Bias is obtained entirely from the voltage drop across the cathode resistance, R_1 (§ 3-6). The r.f. chokes are placed so as to be out of the direct field of the plate tank coil. By-pass condensers (§ 2-13) are located close to the points to be by-passed. A common grounding point (§ 2-13) is provided by a wire running the length of the baseboard to which all ground connections shown in the circuit diagram are made.

Since this circuit is not designed for frequency doubling, a separate crystal is re-

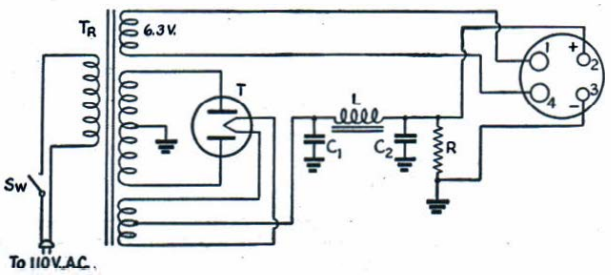


Fig. 1124 — Circuit diagram of a typical receiver power supply.
C1 — 4-μfd. electrolytic input.
C — 8- or 16-μfd. electrolytic output.
L — 10-hy., 150-ma. low-resistance filter choke.
R — 15,000-ohm, 25-watt bleeder.
T — Type 80 rectifier.
Tr — Power transformer: 400 volts each side of center tap at 140 ma., 6.3 volts at 6 amperes and 5 volts at 3 amperes.

quired for each frequency at which it is desired to operate.

Simple direct coupling to the antenna (§ 10-4) is shown in the diagram. Coupling is adjusted by moving the tap up or down on the plate tank coil. As indicated by the dotted

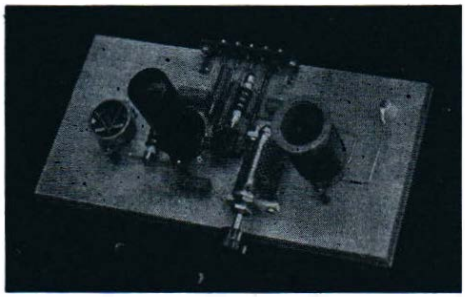


Fig. 1125 — A simple breadboard oscillator transmitter. The crystal is plugged in the tube socket to the left, while the socket to the right holds the plate tank coil, L_1. The 6L6 oscillator tube is near the center. The grid r.f. choke is between the crystal and 6L6, while the plate r.f. choke is to the right of the 6L6. The cathode and screen resistors are to the rear of the 6L6. The blocking condenser, C_2, is between the tube and the tank condenser, C_1, to the left of L_1.

lines, link coupling to a conventional antenna tuner (§ 10-6) may be used, if preferred, by adding a suitable link winding at the bottom of the plate tank-coil form.

Connections to a power unit, such as the one shown in Fig. 1123, may be made by a plug and cable connected to the terminal strip at the rear of the baseboard. Plate voltages up to 450 may be employed when using the type 6L6, while the same arrangement, without change, may be used with the type 6V6 at lower plate voltages.

A meter with a scale of 100 or 200 ma. may be connected in series with the key for checking, for tuning purposes, combined plate and screen currents which flow through the cathode circuit.

With a 6L6 tube and a plate supply delivering 400 volts, the screen voltage will run about 250. The tube will draw about 75 ma. nonoscillating, dipping to about 50 ma. at resonance (§ 4-4) with the antenna disconnected. It should be possible to load up the circuit until the tube draws about 80 ma. at resonance. Under these conditions, the power output on each band should be 15 to 20 watts.

• A TWO-TUBE PLUG-IN COIL EXCITER

In the two-tube exciter or low-power transmitter shown in the photographs of Figs. 1127 and 1128, a 6L6 oscillator is used to drive an 807 as an amplifier-doubler. The circuit diagram is shown in Fig. 1129.

The Tri-tet oscillator circuit (§ 4-5) is chosen because of its ability to supply output at harmonic frequencies of the crystal, as well as at the fundamental. Sufficient output may be obtained at both the second and fourth harmonics (§ 2-7) to drive the 807 amplifier. The amplifier stage is capacitively coupled (§ 4-6) to the oscillator through the coupling condenser, C_4. The 807 is a screened tube, therefore no neutralizing circuit is required (§ 4-7). A link, L_4, is provided at the output for coupling to the grid tank circuit of a following amplifier or an antenna tuner (§ 4-7; § 10-6).

Series plate feed (§ 3-7) is used in both stages, while parallel grid feed is used in the amplifier. The resistances R_4, R_5, R_8 and R_9 form a voltage divider (§ 2-6) to provide suitable voltages for the oscillator screen and plate. Two separate resistors, R_8 and R_9 are used instead of a single resistor in this case because their physical sizes permit suitable space inside the small chassis to be found. A second voltage divider for the amplifier screen is composed of the resistances R_6 and R_7.

Safety bias for the oscillator tube, in case the oscillator circuit ceases to function, is provided by the cathode resistance (§ 3-6), R_3.

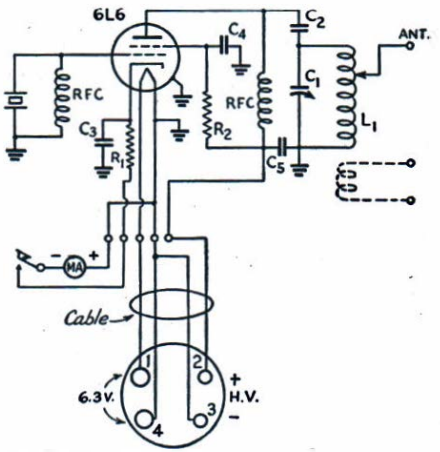


Fig. 1126 — Circuit diagram of the simple tetrode oscillator transmitter.

C_1 — 250-$\mu\mu$fd. plate tank condenser.
C_2 — 0.001-μfd. mica plate blocking condenser.
C_3, C_4, C_5 — 0.01-μfd. paper cathode, screen and plate by-pass condensers.
R_1 — 200-ohm, 2-watt cathode biasing resistor.
R_2 — 15,000-ohm, 2-watt screen voltage-dropping resistor.
RFC — 2.5-mh. r.f. chokes.
L_1 — Plate-tank inductance, 38 μhy., 12 μhy., and 5 μhy., respectively, for 1.75, 3.5 and 7 Mc.
Power connections are made through the plug and cable indicated.

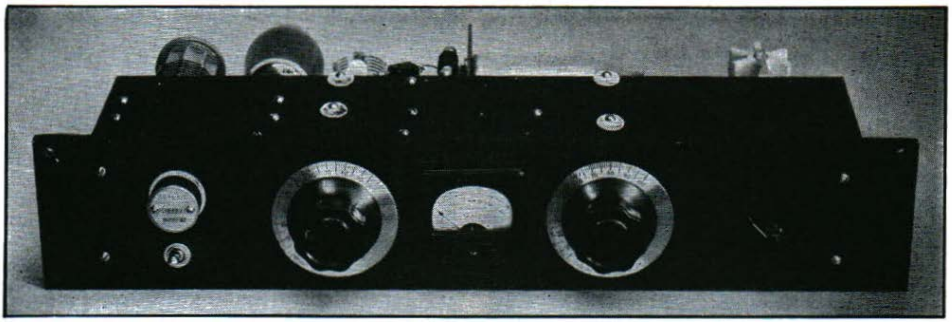


Fig. 1127 — A two-tube plug-in coil exciter, built to conserve space in a relay rack. The crystal socket to the left is submounted in the panel. The dials control the plate tank condensers, C_1 and C_2. The knob to the right is for the meter switch, Sw_2, while the toggle switch, Sw_1, is to the left.

Additional operating bias (§ 4-8) is obtained from the grid-leak resistance (§ 3-6), R_1. In the amplifier circuit, protective bias is furnished from an external 45-volt supply, such as a "B" battery. Additional bias is obtained from the grid leak, R_2, when the circuit is in operation.

A switch is provided so that the milliammeter may be shifted to read either oscillator plate current or amplifier plate current for tuning purposes. With the keying system shown, both stages are keyed simultaneously in the common cathode lead (§ 6-2).

The unit is designed to operate from a single high-voltage power supply delivering 750 volts at 250 ma.

Although the output stage may be operated as a frequency doubler (§ 4-1; § 4-11) when it is necessary to obtain output at the eighth harmonic of the crystal frequency, greater power output may be obtained, without exceeding the dissipation rating of the tube, if it is operated as a straight amplifier.

The unit is constructed in a manner to conserve panel space in a relay rack. The crystal socket is mounted on the panel so that crystals may be conveniently changed. The cathode tank coil, the oscillator tube and the oscillator plate tank coil are grouped closely to permit short connecting wires. L_1 and L_2 are placed so that their axes are at right angles to reduce inductive coupling between the two (§ 2-11). C_1 is mounted inside the chassis so that short leads may be passed through clearance holes in the chassis between its terminals and those of L_2. The 807 tube is mounted horizontally from a small metal panel which, together with the cylindrical can, provides a measure of shielding against external stray capacitive coupling between the amplifier input and output circuits which, if allowed to exist, might cause self oscillation in the amplifier (§ 7-6). The horizontal mounting also permits a short plate lead to the output tank circuit. The output tank coil is placed at the extreme end of the chassis to reduce inductive coupling between the two tank circuits.

In preparing to place the unit in operation, a crystal must be chosen whose fundamental frequency or whose second-, fourth- or eighth-harmonic frequency falls at the output frequency desired. For output at the fundamental

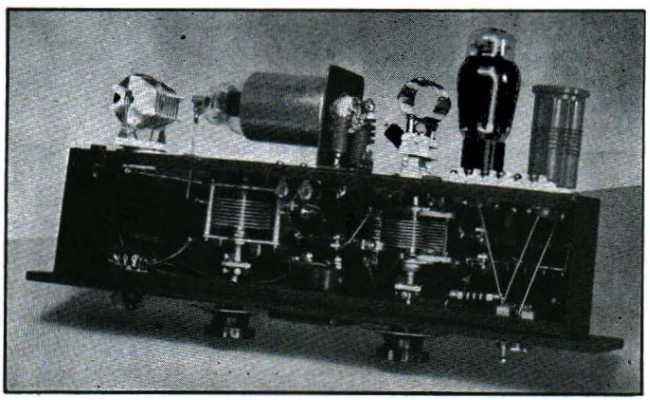


Fig. 1128— Underneath view of the two-tube plug-in coil exciter. The components mounted along the rear edge of the chassis from left to right are: The output tank coil, L_3, the 807 amplifier tube, the oscillator plate tank coil, L_2, the 6L6G oscillator tube and the cathode coil, L_1. Inside the chassis are the two tank condensers, C_1 and C_2, which must be insulated from the chassis, and the various resistors. Insulating couplings are used between the tank-condenser shafts and the two dials.

frequency or the second- or fourth-harmonic frequencies of the crystal, coils must be selected for L_2 and L_3 which will resonate at the desired output frequency. If the output frequency desired is the eighth-harmonic frequency of the crystal, C_1-L_2 should tune to the fourth harmonic, while the C_2-L_3 circuit should be tuned to the desired output frequency. The coil used for L_1 should always correspond to the frequency of the *crystal* in use. When C_1-L_2 is tuned to the fundamental frequency of the crystal, Sw_1 should be closed. The circuit is then that of a simple tetrode crystal oscillator (§ 4-4). The purpose of this is to prevent excessive crystal r.f. currents which may damage the crystal when the oscillator tube is operated at high-power input at the crystal fundamental frequency (§ 4-3).

When the unit is operated at 750 volts, the power output obtainable should run between 40 and 55 watts, depending upon the output frequency, if the output stage is operated as a straight amplifier. When the output stage is used as a frequency doubler, the input must be reduced to prevent excessive plate dissipation and, therefore, the power output obtainable in practice will be reduced to between 18 and 25 watts.

• A 450-WATT PUSH-PULL AMPLIFIER

Figs. 1130 and 1131 show two views of a push-pull r.f. amplifier designed for a pair of tubes of the 1500-volt, 150-ma. class, such as the types T40, T55, 812, 8005, RK51, HF100, etc. A similar arrangement, with a plate tank condenser of less plate spacing (§ 4-8), is also suitable for tubes of the 1000-volt, 100-ma. class.

The circuit, which is conventional, is shown in the diagram of Fig. 1132. A tuned tank circuit is provided for the grid circuit, as well as the plate circuit, to permit link coupling to an exciter. A system of plug-in coils is used to shift operating frequency from band to band. To provide sufficient tank-circuit capacity (§ 4-8) without the use of variable condensers of excessively-large physical dimensions, provision has been made for plugging in a fixed air padding condenser in parallel with each of

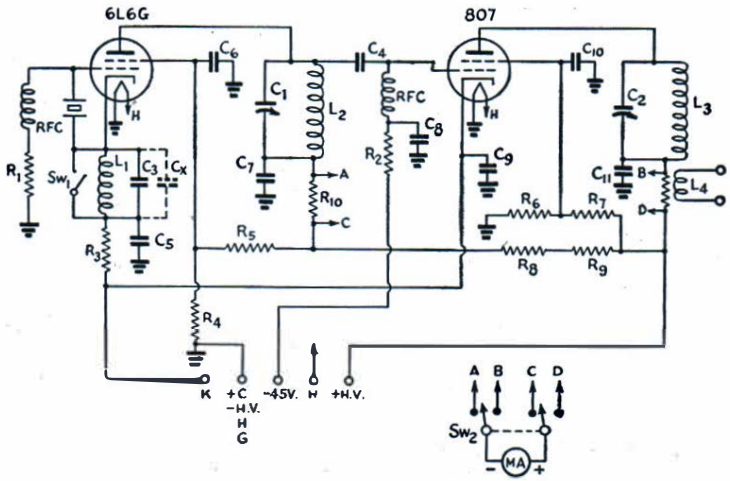


Fig. 1129 — Circuit diagram of the two-tube exciter.

C_1 — 140-$\mu\mu$fd. oscillator plate tank condenser.
C_2 — 150-$\mu\mu$fd amplifier plate tank condenser.
C_3 — 100-$\mu\mu$fd. mica cathode tank condenser.
C_4 — 20-$\mu\mu$fd. mica coupling condenser.
C_5, C_9 — 0.01-μfd., 600-volt, paper cathode by-pass condensers.
C_6, C_{10} — 0.01-μfd., 600-volt, paper screen by-pass condensers.
C_7 — 0.01-μfd., 600-volt, paper oscillator plate by-pass condenser.
C_8 — 0.01-μfd., 600-volt, paper grid-circuit, by-pass condenser.
C_{11} — 0.01-μfd., 1000-volt, paper amplifier plate by-pass condenser.
R_1 — 20,000-ohm, 1-watt oscillator grid leak.
R_2 — 25,000-ohm, 2-watt amplifier grid leak.
R_3 — 200-ohm, 2-watt oscillator cathode biasing resistance.
R_4 — 10,000 ohms, 25-watt (see text).
R_5 — 3500 ohms, 25-watt (see text).
R_6, R_7 — 15,000 ohms, 25-watt (see text).
R_8, R_9 — 1250 ohms, 50-watt (see text).
R_{10}, R_{11} — 10-ohm, 1-watt meter-shunting resistances.
RFC — 2.5-mh. r.f. choke.
Sw_1 — S.p.s.t. toggle switch (see text).
Sw_2 — D.p.d.t. rotary meter switch.
L_1 — Oscillator cathode inductance — 35 μhy., 3.2 μhy., and 1.75 μhy., respectively, for 1.75-, 3.5- and 7-Mc. crystals.
L_2 — Oscillator plate tank inductance — 54 μhy., 15 μhy., 4.2 μhy., 1.25 μhy. and 0.5 μhy., respectively, for 1.75, 3.5, 7, 14 and 28 Mc.
L_3 — Amplifier plate tank inductance — 52 μhy., 16 μhy., 5.7 μhy., 1.5 μhy. and 0.7 μhy., respectively, for 1.75, 3.5, 7, 14 and 28 Mc.

the tank condensers for 1.75-Mc. operation.

The rotor of the split-stator plate tank condenser, C_2, is insulated from ground for d.c., while C_{10} provides a ground path for r.f. currents. This permits a high d.c.-voltage connection to be made to the rotor of C_2, thereby removing the difference of d.c. potential between rotor and stator plates of the tank condenser. This connection permits the use of a tank condenser with less plate spacing, since the peak r.f. voltage is the maximum to appear between the condenser plates (§ 4-4).

The tuned circuits L_3-C_6 and L_4-C_5 are trap circuits tuned to trap out ultra-high-frequency parasitic oscillations which are usually encountered in an amplifier of this type (§ 4-10).

The 100-ma. meter may be switched to read either d.c. grid current or total cathode current. When switched to read cathode current, the meter is shunted by a low resistance, R_2, which is of the correct value to give a meter-scale multiplication of five (see Chapter 12). The shunting resistance, R_1, is of sufficiently high value to have no practical effect upon the reading of the meter when it is switched to read grid current.

The purpose of the disposition of components shown in the photographs is to arrive at an arrangement which will permit both short r.f. connecting leads and good isolation between grid and plate circuits. The plate tank condenser and coil are placed on either side of a vertical partition with clearance holes for the connecting leads. The tubes are placed on the chassis with their plates close to the stator terminals of the plate tank condenser and their sockets submounted so that their grid terminals are close to the stator terminals of the grid tank condenser underneath the chassis. The grid tank coil is mounted on the vertical partition to the left with its axis at right angles to that of the plate tank coil. The chassis

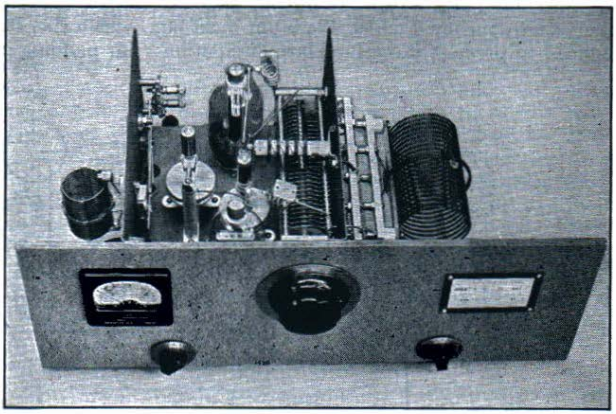


Fig. 1130 — A 450-watt push-pull amplifier. The grid tank inductance, L_1, is to the left and the plate tank inductance, L_2, to the right. The two neutralizing condensers are staggered between the two tubes on the chassis. The plate tank condenser, C_2, is mounted on the right-hand partition. The parasitic trap tank circuits, L_3-C_6 and L_4-C_5, are in the plate leads to the tubes.

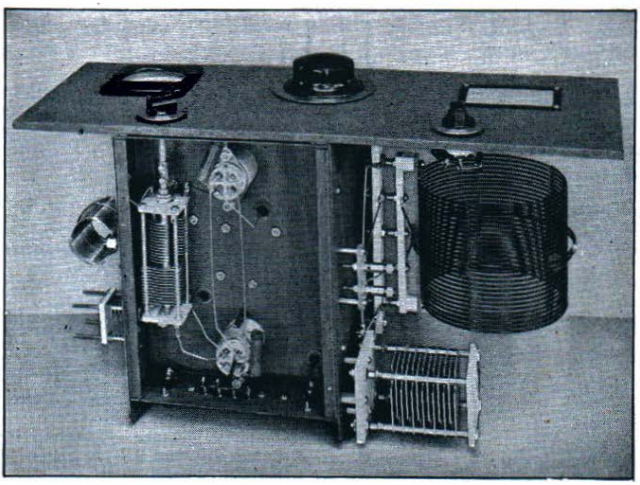


Fig. 1131 — Bottom view of the 450-watt push-pull amplifier, showing the position of the grid tank condenser between the two submounted tube sockets and the two air padding condensers in place for 1.75-Mc. operation.

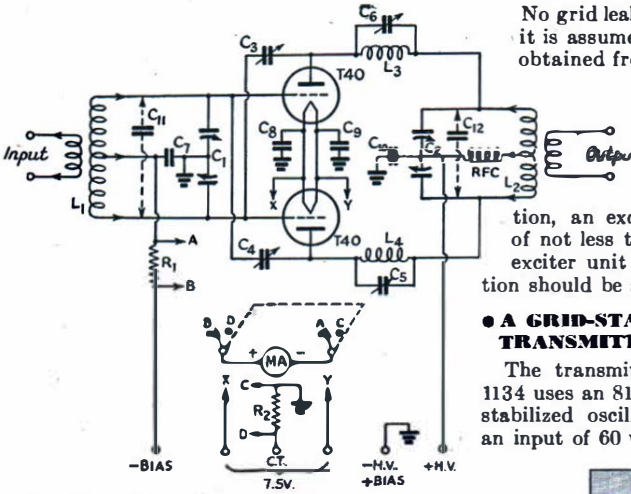


Fig. 1132 — Circuit diagram of the 450-watt push-pull amplifier.

C_1 — 100-$\mu\mu$fd. per-section grid tank condenser, 0.03-inch spacing.
C_2 — 100-$\mu\mu$fd. per-section plate tank condenser, 0.07-inch spacing.
C_3, C_4 — Micrometer-type neutralizing condensers.
C_5, C_6 — 30-$\mu\mu$fd. mica-trimmer parasitic-trap condensers.
C_7 — 0.01-μfd., 600-volt, paper grid-circuit by-pass condenser.
C_8, C_9 — 0.01-μfd., 600-volt, paper filament by-pass condensers.
C_{10} — 0.001-μfd., 7500-volt, mica plate-circuit by-pass condenser.
C_{11} — 50-$\mu\mu$fd., air grid-tank padding condenser for 1.75 Mc., 0.05-inch spacing.
C_{12} — 50-$\mu\mu$fd. air plate-tank padding condenser for 1.75 Mc., 0.125-inch spacing.
R_1 — 25-ohm, 1-watt meter-shunting resistance.
R_2 — Meter-multiplier resistance for 5-times multiplication.
RFC — 1-mh. r.f. choke.
MA — D.c. milliammeter, 100-ma. scale.
L_1 — Grid tank inductance — 70 μhy., 38 μhy., 13 μhy., 4.5 μhy. and 0.8 μhy., respectively, for 1.75, 3.5, 7, 14 and 28 Mc., 3-turn links.
L_2 — Plate tank inductance — 70 μhy., 35 μhy., 14 μhy., 3 μhy. and 1 μhy., respectively, for 1.75, 3.5, 7, 14 and 28 Mc., 2-turn links.
L_3, L_4 — Parasitic trap coils — 4 turns No. 12 wire, ½-inch diameter, ¾-inch long.

and double vertical partitions provide shielding against undesirable couplings (§ 7-6) between input and output circuits, which might otherwise be sufficient to make complete neutralization (§ 4-7) impossible. The neutralizing condensers are placed in a position where short, direct connecting leads are possible. The parasitic-trap components are soldered directly in the leads from the tube plate caps to the stators of C_2.

For operation at maximum input, a plate-voltage supply delivering 1500 volts at 300 ma. is required. Filament supply, depending upon the tubes selected, will also be required.

No grid leak is shown in the diagram, since it is assumed that biasing voltage will be obtained from one of the simple units containing the required grid-leak resistance discussed in the chapter on power supply (§ 8-9).

For maximum plate efficiency with plate modulation, an exciter having an output rating of not less than 25 watts is required. The exciter unit described in the previous section should be suitable.

• A GRID-STABILIZED 815 112-MC. TRANSMITTER

The transmitter shown in Figs. 1133 and 1134 uses an 815 double beam tube in a grid-stabilized oscillator circuit and will run at an input of 60 watts with good efficiency. The

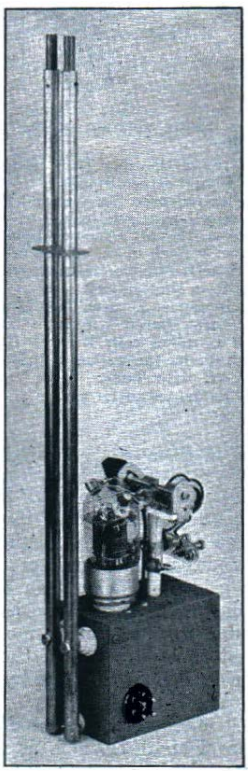


Fig. 1133 — The grid-stabilized 112-Mc. transmitter is mounted on a 3- by 4- by 5-inch metal box, and the box houses the filament transformer and the various fixed condensers, resistors and the r.f. choke. The frequency is changed by adjusting the length of the grid lines by sliding the inner tubes in and out. The power supply cable plugs on the plug mounted on the side of the box. Wires for feedback control run from the plate caps of the 815 close to the grid lines.

circuit is similar to the tuned-grid tuned-plate except that it uses a linear circuit instead of a coil and condenser in the grid circuit. By tapping the grids down on the line the line is loaded lightly and consequently retains its high Q. The 815 does not have a high-enough grid-plate capacity to give all of the necessary feedback, and some additional capacity must be added from plate to grid of both sections of the tube. This is easily done by two short lengths of wire running from the plate terminals to points near the grid lines.

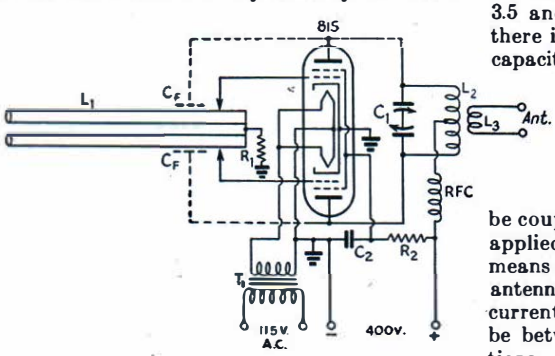


Fig. 1134 — Wiring diagram of the grid-stabilized 2½-meter oscillator.

C_1 — 15-$\mu\mu$fd. per section dual plate tuning.
C_2 — 0.002-μfd. mica screen by-pass.
C_F — Feedback condenser. See text and Fig. 1133.
R_1 — 15,000-ohm, 1-watt grid leak.
R_2 — 25,000-ohm, 10-watt screen dropping.
L_1 — Grid lines: ½-inch diam. copper tubing 23 inches long. Spaced 1 inch on centers; grids tapped 2¾ inches from shorted end.
L_2 — Plate inductance: 2 turns No. 12 enam., 1-inch diam., turns spaced ¾ inch.
L_3 — Antenna coupling coil: 2 turns No. 12 enam., ¾-inch diam., turns spaced ¼ inch.
RFC — U.h.f. r.f. plate choke.
T_1 — 6.3-volt filament transformer.

The grid line is made of half-inch copper tubing and is supported a half inch from the box by three feed-through insulators which also serve as convenient connectors to the grids and to the grid leak. The open ends of the parallel tubings take 3-inch lengths of ⅜-inch diameter tubing which can be moved in and out to adjust the frequency of the oscillator. They are held securely in place by set screws through the half-inch tubing.

The plate condenser is supported by a 3-inch steatite pillar which also acts as a guide for the sliding variable antenna coupling. Two large 866-type plate caps are slid over the pillar and the antenna binding-post assembly is fastened to them by short lengths of No. 12 wire. By sliding this assembly up and down the antenna coupling can be set to any value desired.

There is nothing unusual about the tuning of the transmitter outside of the adjustment of the feedback condensers. This can best be done with a dummy load such as a 25-watt electric lamp connected to the antenna terminals. The lead from the grid leak, R_1, to ground should be opened and a 0-10 millammeter connected in the circuit. Plate voltage can be applied and the plate tuning condenser rotated for maximum output as indicated by the brilliancy of the lamp. The grid current should be between 3.5 and 5 ma. at this point — if it is higher there is too much feedback and the feedback capacity should be reduced by trimming off a short length of the wire or by moving it away from the grid lines. It is not too critical a setting but it should be done before the transmitter is put on the air. After the proper feedback adjustment is found, the antenna can be coupled to the transmitter and modulation applied. The frequency can be checked by means of Lecher wires or a wave-meter. The antenna coupling is tightened until the plate current is 150 ma. and the grid current should be between 3.5 and 5 ma. under these conditions.

The power supply is required to deliver slightly over 165 ma. at 400 volts, and the modulator must give at least 30 watts to modulate fully the oscillator.

• A TRANSMITTER FOR 224 MC.

As one operates on frequencies higher than 116 Mc. he finds considerable difficulty in getting good performance with tubes other than

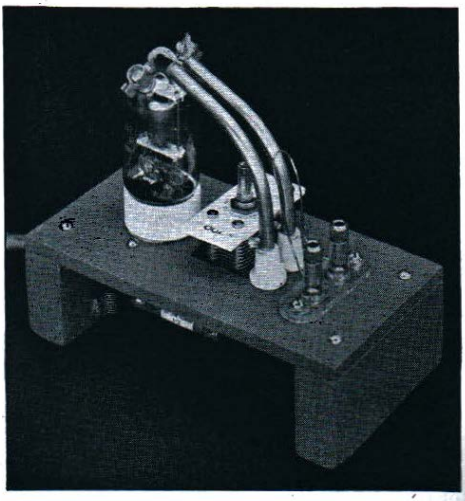


Fig. 1135 — A 224-Mc. transmitter using the HY75. A rectangular hole in the top of the Presdwood chassis allows the tuning condenser to be placed for shortest leads. The tuning condenser is adjusted by an insulated screwdriver.

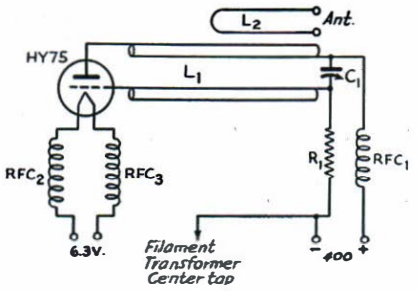


Fig. 1136 — Wiring diagram of the 224-Mc. oscillator.
C_1 — 100-$\mu\mu$fd. midget variable tuning.
R_1 — 5000-ohm, 10-watt wirewound grid leak.
L_1 — Series-tuned tank circuit: ¼-inch copper tubing 3½ inches long, spaced ½ inch on centers.
L_2 — Antenna coupling loop: 2-inch loop No. 16 bare wire.
RFC_1 — U.h.f. r.f. plate choke.
RFC_2, RFC_3 — Filament chokes: 10 turns No. 18 enam. closewound on ½-inch diam., self-supporting.

those designed expressly for u.h.f. operation. However, there are several inexpensive tubes available to amateurs that will perform well on 224 Mc., and the transmitter shown in Figs. 1135 and 1136 shows how the HY75 can be used.

The transmitter is built on a 3½- by 6½-inch strip of ¼-inch Presdwood supported by two strips of 1- by 2-inch wood. A rectangular hole is cut in the center of the Presdwood to accommodate the tuning condenser which is supported by two metal pillars at one end. The tuned circuit consists of two pieces of ¼-inch copper tubing supported at one end by two feed-through insulators. The screws of the feed-through insulators are sweated into the ends of the tubing, and the tuning condenser connects to two lugs right at this point. Connection from the tubing to the grid and plate leads of the tube is made through ½ inch of flexible braid. Filament chokes, the plate r.f. choke and the grid leak are mounted under the chassis.

The antenna coupling consists of a loop of wire parallel to the copper tubing and terminating in the antenna binding posts. The coupling is varied by moving the loop nearer to or farther away from the copper tubing.

The transmitter should first be tested with a dummy load, and a 10-watt electric lamp is excellent for the purpose. The load is connected to the antenna posts and the power supply is then turned on. If everything is connected properly the lamp will light, its brilliancy depending upon the tightness of coupling and the setting of C_1. It will be found that the output is a little better towards the maximum-capacity end of the range of C_1. The frequency coverage of the transmitter should now be checked, by Lecher wires or a wavemeter, to make sure that it will cover the range. The coverage can be adjusted slightly by changing the separation of the copper tubes, but if this is not enough the tubes will have to be made shorter or longer.

The transmitter requires a power supply capable of furnishing 60 ma. at 400 volts, and the modulator should be capable of delivering 12 watts of audio.

Because of its small size, a transmitter of this type can be built right into a rotatable antenna for the 224-Mc. band if desired. It is desirable not to run a feed line for any great distance at this frequency because of the chances for loss in the line.

• A 112-MC. "WALKIE-TALKIE" TRANSMITTER-RECEIVER

Many battery-operated lightweight 112-Mc. stations are of the "transceiver" variety, which usually consists of two tubes, one serving as the transmitter oscillator or receiver detector and the other as modulator or audio amplifier, depending upon the position of the "send-receive" switch. A transceiver of this type has the disadvantage that too many compromises must be made — in the "receive" position the radiation is almost as great as in the "send" position, and the transmitted and received signals must be of the same frequency because only one tuned circuit is used. Further, it is practically impossible to obtain good trans-

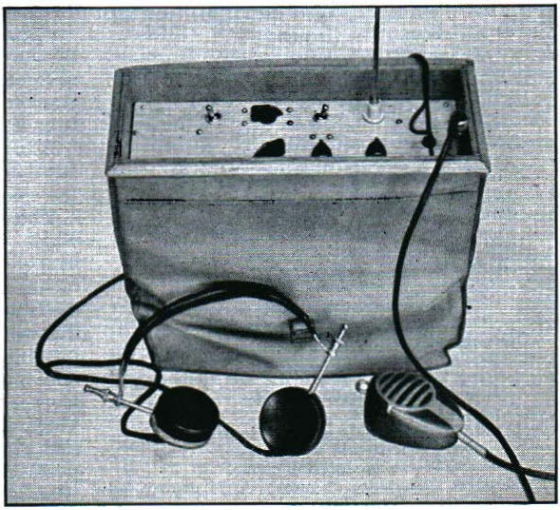


Fig. 1137 — A 112-Mc. pack set ready to go. The station is built into a knapsack which contains two complete sets of batteries.

mitter stability and receiver sensitivity at the same time in such an arrangement.

The transmitter-receiver shown in Figs. 1137, 1138 and 1139 is a big improvement over a transceiver in that it uses only two tubes but has separate tuned circuits for receiving and transmitting, thereby reducing considerably the compromises in performance. In the "receive" position a low-C tank circuit is switched in. This has been adjusted for best sensitivity. In the "send" position a high-C tank circuit which gives a fair order of frequency stability is used. The circuits can be tuned to any desired frequency in the band, and thus no operating limitations are imposed. Optimum antenna coupling to each circuit is assured by individual transmitter and receiver antenna-coupling condensers. The carrying case is large enough to hold two complete sets of batteries; a switch cuts in a new set in the event that batteries begin to fail during operation in the field. However, the batteries are not expensive, and more than 50 hours of operation can be expected from a single set.

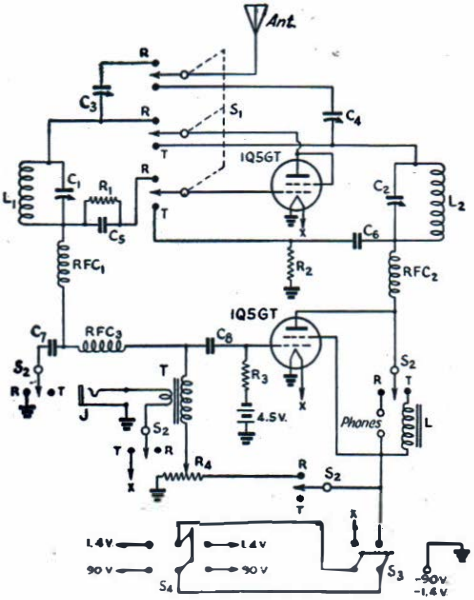


Fig. 1138 — Wiring diagram of the 112-Mc. pack set.

C_1 — 10-$\mu\mu$fd. midget receiver tuning.
C_2 — 35-$\mu\mu$fd. midget transmitter tuning.
C_3 — 3-30-$\mu\mu$fd. adjustable mica receiver antenna coupling.
C_4 — 3-30-$\mu\mu$fd. adjustable mica transmitter antenna coupling.
C_5 — 100-$\mu\mu$fd. receiver grid.
C_6 — 100-$\mu\mu$fd. transmitter grid.
C_7 — 0.004-μfd. interruption-frequency by-pass.
C_8 — 0.01-μfd. audio coupling.
R_1 — 0.5-megohm, ½-watt receiver grid leak.
R_2 — 15,000-ohm, ½-watt transmitter grid leak.
R_3 — 0.25-megohm, ½-watt audio grid.
R_4 — 50,000-ohm regeneration control.
RFC_1 — U.h.f. receiver plate choke.
RFC_2 — U.h.f. transmitter plate choke.
RFC_3 — 80-mh. interruption frequency filter choke.
J — Open-circuit microphone jack.
S_1 — 3-circuit 2-position Isolantite r.f. switch.
S_2 — 4-circuit 2-position bakelite d.c. and audio switch. S_1 and S_2 are ganged.
S_3 — D.p.s.t. on-off toggle.
S_4 — D.p.d.t. power supply selection toggle.
T — S.h. microphone-to-grid transformer.
L — 15-henry, 40-ma. modulation choke.
L_1 — Receiver inductance: 3 turns No. 14 enam., ⅜-inch inside diam., 7/16-inch long.
L_2 — Transmitter inductance: 1 turn No. 12 enam., ½-inch diam.

A 1Q5GT, with plate and screen grid connected together to give a triode characteristic, is used for the r.f. tube. A two-section two-position rotary switch is used for changing from "send" to "receive" and back again. One section of the switch (S_1) has Isolantite insulation and is used to switch the r.f. circuits: the grid and plate of the 1Q5GT and the antenna. The other section of the switch (S_2) is bakelite-insulated and controls the d.c. and audio circuits. In the "receive" position, S_2 connects in the interruption-frequency feedback condenser, C_7 (removed during transmission because it by-passes too much of the audio from the microphone transformer), opens the microphone circuit to reduce battery drain, closes the receiver regeneration-control supply circuit and switches the output of the audio tube (also a 1Q5GT) from the modulation choke, L, to the headphones.

In the "receive" position, the superregenerative detector is impedance-coupled to the audio amplifier tube by the secondary of the microphone transformer, T (acting as a coupling impedance), coupling condenser C_8 and grid resistor R_3. Regeneration is controlled by the setting of R_4. Bias for the audio tube is obtained from a small 4.5-volt battery. When transmitting, the plate circuit of the audio tube is coupled to the oscillator across L.

A toggle switch, S_3, turns on the station, and another toggle switch, S_4, is used to switch to the spare supply of batteries.

The station is built on a plywood panel which is fastened to a plywood box fitted snugly in the knapsack. The box extends 2 inches above the panel, thus protecting the controls from accidental movement. The dimensions of the box will, of course, be determined by the size of the knapsack. The box does not extend to the bottom of the knapsack. Corner posts, of ½-inch square wood, are provided for fastening the panel in place. Glue and brads are used to hold the box and corner posts together. The box is prevented from slipping into the knapsack by tacking strips of ¼-inch quarter-round trim around the top edges of the box.

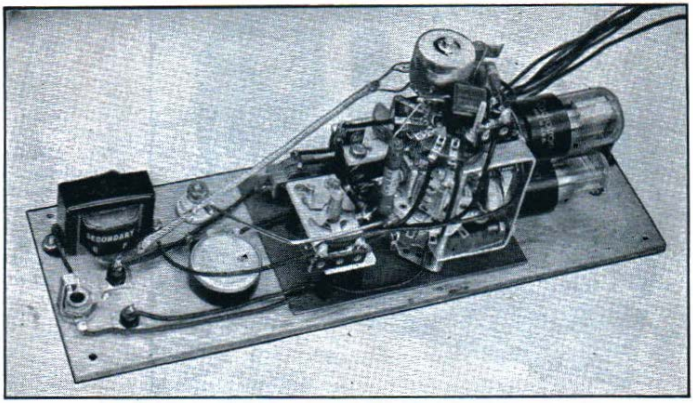


Fig. 1139 — A view of the back of the 112-Mc. pack-set panel. The receiver tuning condenser can be seen to the left of the "send-receive" switch. The two tubes are mounted on brackets on the switch. The microphone transformer, microphone jack, headphone tip jacks and regeneration control can be seen at the left-hand edge. The modulation transformer, L_2 is directly behind the two tubes.

The arrangement of the parts in the r.f. portion of the station is centered around the "send-receive" switch because of the necessity for short r.f. leads. The two tube sockets are supported by metal brackets mounted on the switch, thus placing the r.f. tube socket directly over the Isolantite switch section and the audio tube over the bakelite switch section. The brackets can be made of ½-inch strips of 1/16-inch thick aluminum or copper. The plywood panel is backed up by a 3½- by 5-inch plate of aluminum to reduce hand-capacity effects. The tuning condensers, C_1 and C_2, are mounted on 1-inch pillars as close to the Isolantite switch as possible.

Considerable care must be exercised in wiring the r.f. portions of the station because additional inductance in the leads will prevent the obtaining of proper *L-C* ratios in the tank circuits. Short leads of No. 12 or 14 wire should be used. The d.c. wiring is relatively unimportant and can be run wherever convenient. The lead from RFC_3 to the audio transformer had best be shielded to prevent r.f. pickup along its length.

The antenna is a 38-inch length of automobile antenna. It is supported by a porcelain feedthrough insulator. The antenna is tapped to take a 6-32 screw, and a long 6-32 screw is passed through the insulator. The antenna can be unscrewed from the assembly when not in use.

To test the station, connect the batteries, antenna, headphones and microphone and throw S_1-S_2 to the "receive" position. Set the antenna condensers, C_3 and C_4, to the minimum-capacity position. When S_3 is closed and the tubes heat up (which takes only an instant), the usual superregenerative hiss should be heard as R_4 is advanced. If no hiss is heard, it may be necessary to try different values of C_7 or R_1, providing, of course, that the wiring has been thoroughly checked and found to be correct. The receiver inductance, L_1, can now be adjusted by squeezing the turns together or pulling them apart until the tuning range covered by C_1 includes the 112-Mc. amateur band. Then close C_3 until it is necessary to advance R_4 well towards the end of its range to obtain regeneration.

Upon switching to "send," the inductance L_2 should be trimmed until the transmitter tunes to 112 Mc. with C_2 almost completely meshed. This will give the best stability obtainable with this rig. It is probable that C_4 can be closed up to very near full capacity without any tendency for the transmitter to go out of oscillation. If full capacity does prevent oscillation, C_4 can be opened a bit.

Since the range of the "walkie-talkie" depends to a great extent upon the location of the station, it is highly desirable to operate the unit from as high a point as possible. A change in height of only a few feet may increase the signal strength enough to make communication possible where it would not be sufficient from a lower position. In a crowded city, the range may be only a few blocks, but in open terrain and between elevated points, the range will run up to one-half mile or more.

• A 40-WATT SPEECH AMPLIFIER OR MODULATOR

Fig. 1140 is the photograph of high-gain audio amplifier which may be used to plate modulate a low-power final amplifier or to drive a high-power Class-B modulator. The circuit diagram of this unit will be found in Fig. 1141.

The microphone output is fed to the grid circuit of a high-gain voltage-amplifier stage using a 6J7 pentode. Resistance coupling (§ 5-9) is used between the output of this stage and the input of a second stage with a 6J5 triode. The 6J5 is transformer coupled to the input of a push-pull amplifier which fur-

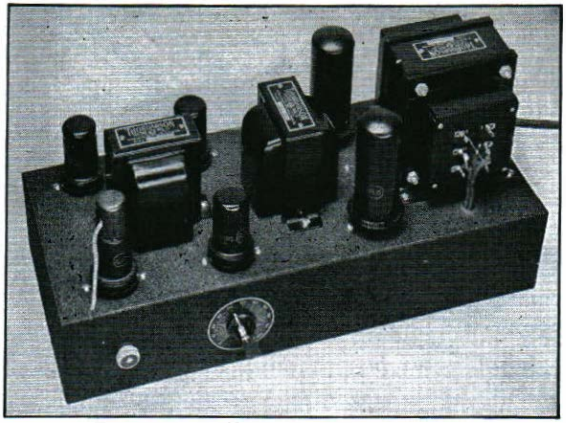


Fig. 1140 — A 40-watt speech amplifier or modulator of inexpensive construction. The 6J7 and first 6J5 are at the front, near the microphone connector and volume control, respectively. T_1 is just behind them, and the push-pull 6J5s are at the rear of the chassis behind T_1. T_2, the 6L6s, and T_3 follow in order to the right.

nishes sufficient power to drive a pair of 6L6s operating as Class AB_2 power amplifiers. Gain is controlled by the potentiometer, R_6, which adjusts the signal voltage delivered to the grid of the 6J5. Operating bias is obtained from the voltage drop across cathode resistors (§ 3-6) in all stages except the output stage. Bias for this stage must be obtained from a steady-voltage, low-resistance source, such as a battery, to prevent instantaneous changes in biasing voltage when the grids are driven positive and grid current flows (§ 5-6).

No by-pass condenser is required across the cathode resistor of the push-pull stage, R_8, because, with a circuit of this type, the sum of the instantaneous cathode currents of both tubes does not change and, therefore, degeneration cannot take place.

T_1 is of the type commonly known as an interstage audio transformer with a turns ratio which provides a step up of signal voltage from the plate of the single 6J5 to the grids of the push-pull stage. T_2 is a step-down transformer designed especially to provide the required driving voltage for the 6L6s with good regulation over the excitation cycle (§ 5-6). The output transformer, T_3, is chosen with a turns ratio to provide the correct operating load resistance for the 6L6s, depending upon the value of the load connected across the secondary of the transformer.

R_{10} and R_{11} comprise a voltage divider

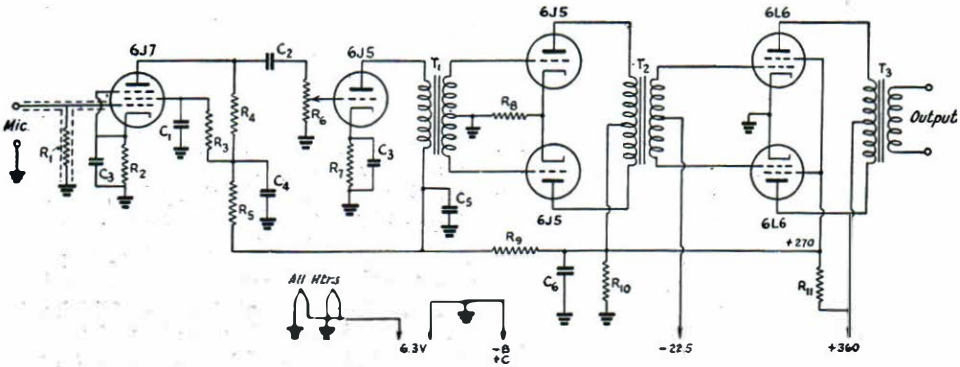


Fig. 1141 — Circuit diagram of the 40-watt speech amplifier modulator.

C_1 — 0.1-μfd. screen by-pass condenser.
C_2 — 0.01-μfd. coupling condenser.
C_3 — 20-μfd., 50-volt electrolytic cathode by-pass condenser.
C_4, C_5, C_6 — 8-μfd., 450-volt electrolytic audio-filter condensers.
R_1 — 5-megohm, ½-watt grid resistance.
R_2 — 1300-ohm, ½-watt cathode biasing resistance.
R_3 — 1.5-megohm, ½-watt screen voltage-dropping resistance.
R_4 — 0.25-megohm, ½-watt plate load resistance.
R_5 — 50,000-ohm, ½-watt plate decoupling resistance.
R_6 — 1-megohm volume control.
R_7 — 1500-ohm, 1-watt cathode biasing resistance.
R_8 — 750-ohm, 1-watt cathode biasing resistance.
R_9 — 12,000-ohm, 1-watt decoupling resistance.
R_{10} — 20,000 ohms, 25-watt (see text).
R_{11} — 1500 ohms, 10-watt (see text).
T_1 — Interstage audio transformer, single plate to p.p. grids, 3:1 ratio.
T_2 — Driver transformer, p.p. 6J5s to 6L6s, Class AB_2.
T_3 — Output transformer, multi-tap.

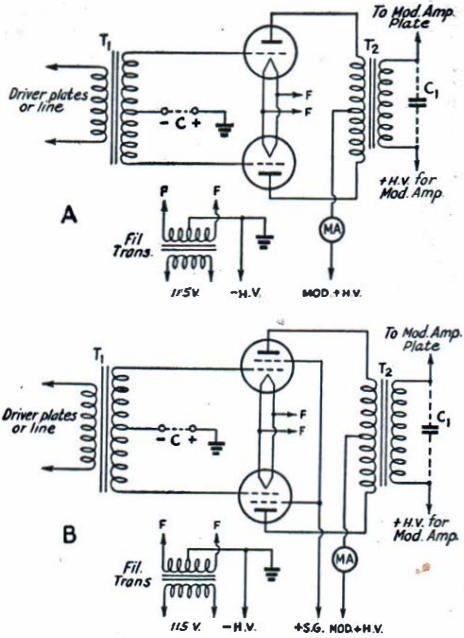


Fig. 1142 — Class-B modulator circuit diagrams. The circuit for triodes is shown at A, the circuit for tetrodes at B.

(§ 2-6) for obtaining lower plate voltage for the tubes of the first three stages. R_5 and R_9 and C_4, C_5 and C_6 form a filtering system for prevention of coupling between stages via power circuits (§ 5-9).

The transformers and tubes are arranged on the chassis so as to permit reasonably short connecting leads. The input circuit of the 6J7 is shielded as completely as possible to prevent the picking up of hum (§ 5-9). The transformers are arranged with their cores at right angles to reduce the possibility of feed-back coupling. Resistors and condensers are mounted underneath the chassis.

The voltage gain provided by the unit is sufficient to operate the output stage at a rated power output of 40 watts from the input signal from a crystal microphone.

Besides heater voltage for the six tubes, the power unit supplying this amplifier should deliver 360 volts with good regulation over the current range of approximately 140 ma. to 265 ma.

The power output obtainable is sufficient to plate modulate an input of 80 watts to a Class-C r.f. amplifier. It may also be used to drive a high-power Class-B modulator.

● CLASS-B MODULATORS

Class-B modulator circuits are practically identical, no matter what the power output of the modulator. The diagrams of Fig. 1142, therefore, will serve for almost any modulator of this type. The circuit for triodes is shown at A, while that for tetrodes is shown at B. When small tubes with indirectly-heated cathodes are used, the cathodes should be connected to ground.

An output transformer should be chosen which will permit matching (§ 5-3) the rated modulator load impedance to the modulating impedance of the r.f. amplifier and, similarly, a driver transformer selected which will couple the driver stage properly to the Class-B grids.

The input transformer, T_1, may couple directly between the driver tube and the modulator grids or may be designed to work from a low-impedance (200- or 500-ohm) line. In the latter case, a tube-to-line output transformer must be used at the driver stage. This type of coupling is recommended only when the driver must be at a considerable distance from the modulator, because the second transformer not only introduces additional losses but also further impairs the voltage regulation.

Condenser C_1 in these diagrams will give a "tone-control" (§ 7-5) effect and filter off high-frequency side-bands (splatter) caused by distortion in the modulator or preceding speech-amplifier stages. Values in the neighborhood of 0.002 to 0.005 μfd. are suitable. The voltage rating should be adequate for the peak voltage across the transformer secondary. The plate by-pass condenser in the modulated amplifier will serve the same purpose.

The plate power supply for the modulator should have good voltage regulation and

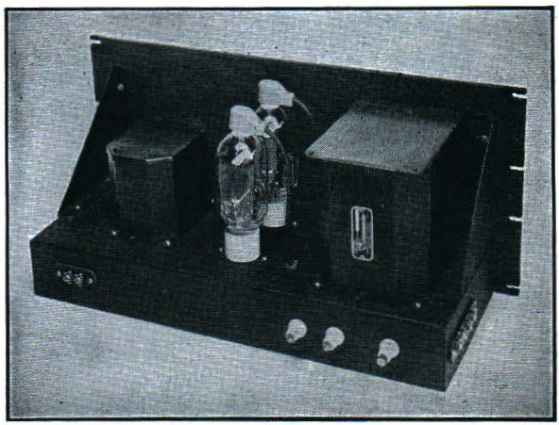


Fig. 1143 — A conventional chassis arrangement for low and medium power Class-B modulators. The layout follows the circuit diagram.

should be well filtered. It is particularly important, in the case of a tetrode Class-B stage, that the screen supply have excellent regulation to prevent distortion. The screen voltage should be set as exactly as possible to the recommended value.

When "C" bias is required for the modulator, the bias source must have very low resistance. Batteries are the most suitable source of bias. In cases where the voltage values are right, regulator tubes such as the VR-75, VR-105, etc., may be connected across a tap on an a.c. bias supply (§ 8-9) and will hold the bias voltage steady under grid-current conditions. Generally, however, zero-bias modulator tubes are preferable not only because no bias supply is required but also because the loading on the driver stage is less variable and driver distortion is consequently reduced.

An example of modulator construction is shown in Fig. 1143.

•POWER SUPPLY UNITS

An example of medium-power plate-supply (see § 8-1 to 8-7) construction is shown in Fig. 1145. The unit pictured varies slightly from the types most frequently encountered in that it is a duplex supply from which two independent voltages may be obtained. Either section, however, by itself is typical of the form in use in the great majority of installations.

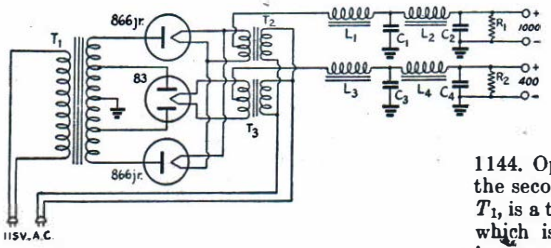


Fig. 1144 — Circuit diagram of the combination 1000- and 400-volt power supply.

C_1, C_2 — 2 μfd., 1000-volt.
C_3 — 4 μfd. electrolytic, 600-volt working.
C_4 — 8 μfd. electrolytic, 600-volt working.
L_1, L_3 — 5/20 hy. swinging choke, 150-ma.
L_2, L_4 — 12 hy. smoothing choke, 150-ma.
R_1 — 20,000 ohms, 75-watt.
R_2 — 20,000 ohms, 25-watt.
T_1 — High-voltage transformer, 1075 and 500 volts r.m.s. each side of center, 125- and 150-ma. simultaneous current rating.
T_2 — Filament transformer, 2.5 volts, 5-amp.
T_3 — Filament transformer, 5 volts, 4-amp.

The circuit diagram is shown in Fig. 1144. Operating from the low-voltage taps of the secondary of the high-voltage transformer, T_1, is a type-83 full-wave rectifier, the output of which is fed into the double-section, choke-input smoothing filter, consisting of L_3-C_3 and L_4-C_4. The filament voltage for the rectifier tube is supplied by the filament transformer, T_3. A pair of type-866 jr. half-wave rectifiers, whose filaments are supplied from T_2, operates from the high-voltage taps of T_1 and the output is fed through a second similar filter system consisting of L_1-C_1 and L_2-C_2. The resistances R_1 and R_2 are the bleeder resistances designed to assist in voltage regulation of the output and to discharge the filter condensers when the power supply is turned off.

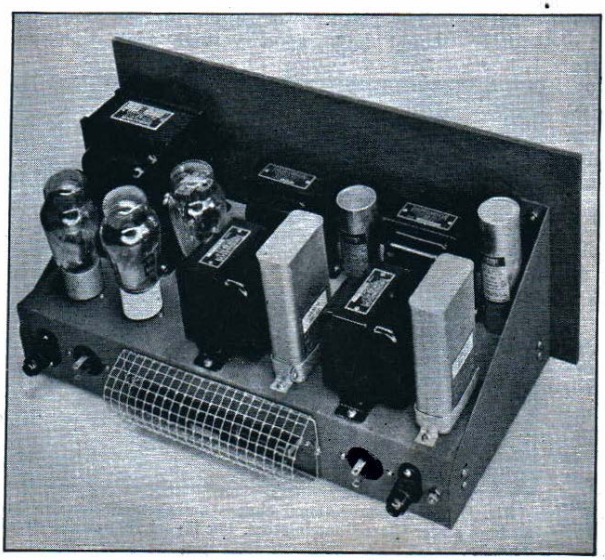


Fig. 1145 — This power supply makes use of a combination transformer and dual filter system delivering 1000 volts at 125 ma. and 400 volts at 150 ma. simultaneously. The circuit diagram is shown in Fig. 1144. The 1000-volt bleeder resistance is mounted on the rear edge of the chassis with a protective guard made of screening to provide ventilation. Safety terminals are used for the two high-voltage terminals. Ceramic sockets should be used for the 866 jrs.

The arrangement of power-supply components is seldom critical. The important points of construction in the unit shown are the use of ceramic sockets for the high-voltage rectifiers (where the insulation of bakelite or fibre sockets would be insufficient to insure against voltage break-down to the chassis), the placing of the high-voltage bleeder resistance, R_1, where it may be adequately ventilated and the use of specially-insulated positive high-voltage terminals and a protective screen around R_1 to reduce the chances of accidental contact by the operator.

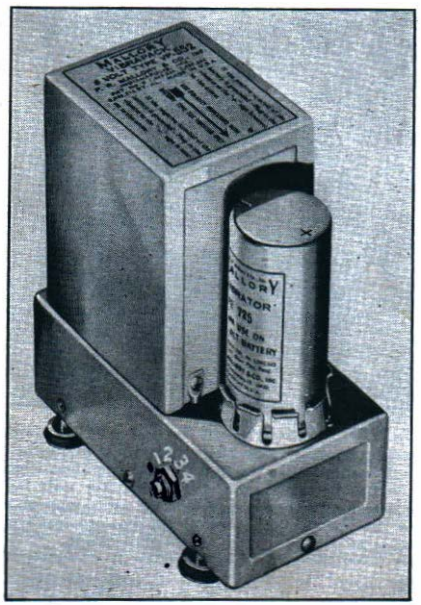


Fig. 1146 — The Mallory VP-552 vibrator supply operates from a 6-volt d.c. source and delivers 300 volts at 100 ma. maximum. Lower values of voltage and current can be obtained by proper setting of the switch on the side of the chassis. This particular unit includes no output filter but does contain all of the necessary hash filters.

The plugs showing in the rear edge of the chassis are for the 115-volt power connections for the plate transformer and the two rectifier filament transformers which are mounted underneath the chassis.

• EMERGENCY POWER SUPPLIES

For emergency and field operation, gasoline-engine-driven generators are almost universally used when the power demand is above 100 watts. However, for low-powered operation above 5 watts (dry batteries are generally used when the power demand is less than 5 watts) the most universally acceptable self-contained power source is the storage battery. It has high initial capacity and can be recharged, so that its effective life is practically infinite. It can be used to provide filament or heater power directly, and plate power through associated devices such as vibrator-transformers, dynamotors and genemotors, and a.c. converters. For emergency work, a storage battery is a particularly successful power source, since practically no matter what the circumstances, such batteries are available. In a serious emergency it would be possible to obtain 6-volt storage batteries as long as there were automobiles to borrow them from. For this reason, the 6-volt storage battery makes an excellent unit around which to design the low-powered portable or emergency station.

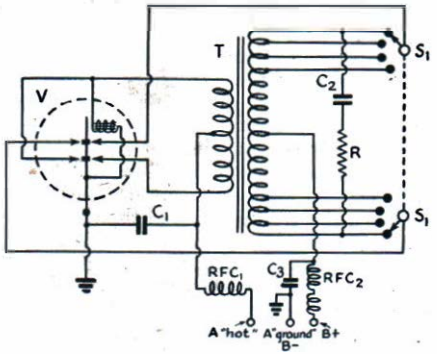


Fig. 1147 — Wiring diagram of the VP-552 vibrator supply.
C_1 — 0.5-μfd., 50-volt vibrator hash suppression.
C_2 — 0.007-μfd., 1600-volt buffer.
C_3 — 0.02-μfd., 1000-volt r.f. filter.
R_1 — 5,000-ohm, 1-watt buffer.
RFC_1 — Low-resistance "A" hash filter.
RFC_2 — "B" hash filter choke.
S_1 — Two-pole 4-position voltage selector.
T — Transformer.
V — Vibrator.

For maximum efficiency and usefulness, the power drain on the storage battery should be limited to 15 or 20 amperes from the ordinary 100- or 120-ampere-hour, 6-volt battery. This should provide a carrier power when transmitting of 20 to 30 watts, which is usually adequate. In connecting the battery, heavy leads of the automotive-cable type should be used to minimize the voltage drop; ordinary car-receiver leads are definitely not satisfactory. Similarly, heavy-duty low-resistance switches are required.

Vibrator Power Supplies — The vibrator power supply consists of a specially-designed transformer combined with a vibrating interrupter. When the unit is connected to a storage battery, the circuit is made and reversed rap-

idly by the vibrator contacts and the square-wave d.c. which flows in the primary of the transformer causes an alternating voltage to be developed in the secondary. This high-voltage a.c. is in turn rectified, either by a vacuum-tube rectifier or by an additional synchronized pair of vibrator contacts, and filtered, providing outputs as high as 400 volts at 200 ma. Tube rectifiers are ordinarily used only when the negative side of the circuit cannot be grounded, a requirement with the self-rectifying type. The high-voltage filter circuit is usually identical with that of an equivalent power source operating from the a.c. line. Noise-suppression equipment, serving to minimize r.f. disturbances, is incorporated in the manufactured units.

Some of the commercial units include a hum filter and some do not, but the design of this filter is, for the most part, conventional. A typical commercial unit of the self-rectifying type is shown in Figs. 1146 and 1147. The vibrator supplies used with automobile receivers are satisfactory for receiver application but usually are not desirable for use with a transmitter except where the power requirements are slight. The efficiency of vibrator packs runs from 60% to 75%. Vibrator supplies are not intended to withstand much overloading, but fusing of the battery cable will eliminate any danger of failure through overloading.



Fig. 1149 — Rear view of the 150-watt transmitter. Immediately below the antenna-tuner unit at the top is a unit containing a single-tube final amplifier, which is driven by the multi-tube all-band exciter unit below. The speech-amplifier and modulator units may be seen below the exciter, with one of the power-supply units showing at the lower edge of the photograph.

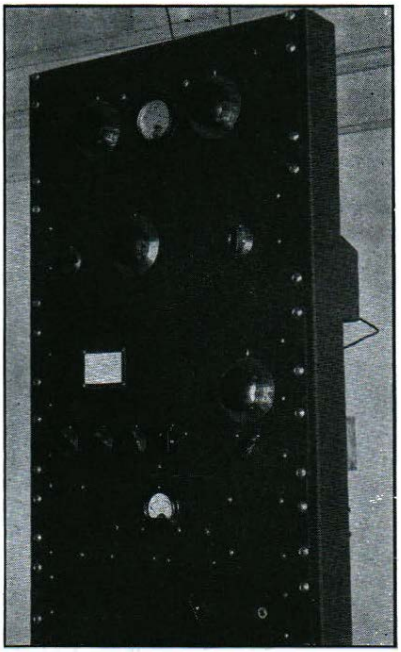


Fig. 1148 — A complete 150-watt rack transmitter for 'phone and c.w.

Dynamotors and Genemotors — A dynamotor is a double-armature high-voltage generator, the additional winding operating as a driving motor. It is usually operated from a 6-, 12- or 32-volt battery, and may deliver voltages from 300 to 1000 or more. Dynamotors have been widely used in military work and many of those in amateur use derive from such origins.

The genemotor is a refinement of the dynamotor designed especially for automobile receiver, sound truck and similar applications. It has found wide acceptance among amateurs as a source of transmitting power, having good regulation and efficiency combined with economy of operation. It is also used in connection with portable receiver installations, although a rather high inherent noise level limits this application in sensitive amateur high-frequency receivers.

Genemotors are made to fill almost every need. Their cost, at amateur net prices, runs from about eight to twenty-four dollars. Stand-

ard models range from 135 volts at 30 ma. to 300 volts at 200 ma. or 500 volts at 200 ma. Parallel and series operation of identical units to provide higher capacity is entirely practical. The normal efficiency averages around 50%, increasing to better than 60% in the higher-power units. The regulation is comparable to well-designed a.c. supplies; it is largely dependent upon external *IR* drops.

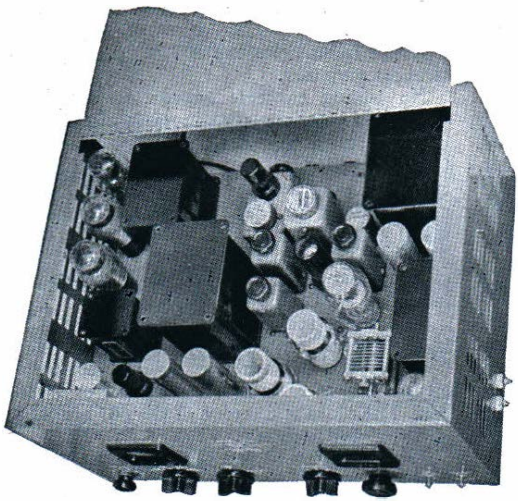


Fig. 1150 — The Hallicrafter HT6 transmitter.

Successful operation of dynamotors and genemotors implies heavy, direct leads, mechanical isolation to reduce vibration, and thorough r.f. and ripple filtration (the purchase of manufactured filter units is recommended). The shafts and bearings should be thoroughly "run in" before regular operation is attempted, and the tension of the bearings should be checked occasionally.

• COMPLETE TRANSMITTERS

A complete radio transmitter may range in size and power from a simple receiving-tube oscillator of almost pocket size coupled directly to the antenna and operating from low-voltage battery supply to large multistage installations feeding elaborate antenna systems and operating with a power input of several hundred kilowatts, although an input of 1000 watts to the output amplifier is the maximum allowed amateur stations by law. The output of the transmitter may be keyed for telegraphic communication or modulated, by one of several different systems, for voice or, in the case of broadcasting stations, for voice and musical transmissions.

One of the simplest forms of transmitter for covering more than very short distances is the simple oscillator transmitter shown in Fig. 1125. Since the operation of crystals ground to frequencies lower than 10 Mc. is more reliable than that of higher-frequency crystals, it is common practice to use the lower-frequency crystals in conjunction with frequency-multiplying stages in the transmitter when output at the higher frequencies is desired. This practice is also often followed in a transmitter employing a variable-frequency self-excited oscillator, because better frequency stability may be obtained with the oscillator operating at a lower frequency, and in amateur transmitters, which operate in harmonically-related bands, to obtain output in any one of several bands from a single crystal. Such frequency multiplication is often done at low-power levels with a series of relatively-small tubes in a unit commonly referred to as an exciter. An example of such an exciter is shown in Fig. 1127.

When greater output than that obtainable from the output stage of the exciter is desired, an amplifier, such as the one shown in Fig. 1130, operating at the output frequency of the

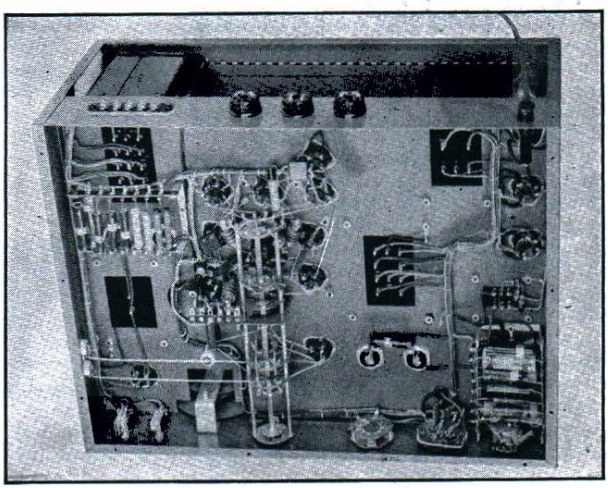


Fig. 1151 — Wiring underneath the HT6.

exciter may be added to step up the power delivered to the antenna.

When modulation of the output wave is desired, audio-frequency units, such as those shown in Figs. 1140 and 1143, are added, which, together with the power-supply units, complete the transmitter.

Transmitters may be found built in many different forms, depending upon the service for which they are designed, but all types will be found to follow the same general plan here outlined. One commonly-used arrangement is the rack-and-panel system pictured in the photographs of Figs. 1148 and 1149. In this system, the transmitter is made up of several units of standard dimensions which fit into a pre-drilled frame or rack also of standard dimensions. The arrangement permits the removal or change of certain units without disturbing other units of the transmitter.

A commercially-built transmitter, of which a wide variety is normally available for various services is shown in the photographs of Figs. 1150 and 1151. It is completely self-contained with modulator and power supply and has an output rating of 25 watts. Provision is made for obtaining output in any amateur band from 1.75 Mc. to 56 Mc. with suitable crystals and coils.

The r.f. circuit employed in this unit is somewhat similar to that of Fig. 1129. A type 6L6 tube is used in the oscillator to drive an 807 output amplifier. Grid and cathode connections are brought out from the oscillator to allow a variety of circuits. On the four lower amateur-frequency bands, the 6L6 is operated as a simple crystal oscillator with output at the crystal frequency. If desired, the crystal may be replaced by a special plug-in unit which converts the circuit to that of an electron-coupled, variable-frequency oscillator.

For 28-Mc. output, the oscillator circuit is converted to the Tri-tet type by means of a special plug-in unit, so that oscillator output at the second harmonic of a 14-Mc. crystal may be obtained.

For 56-Mc. output, the 6L6 is used as a frequency doubler, while a special unit consisting of a separate crystal oscillator is plugged in.

At all frequencies the 807 is operated as a straight amplifier.

No tuning controls appear on the front panel. The various plug-in tank circuits are broadly tuned to permit satisfactory output to be obtained at any frequency within the specified band without the necessity for accurate tuning for that frequency. Provision is made so that pre-tuned units for any three bands may be plugged in simultaneously and output in any one of these bands may be selected by a switch on the front panel.

The audio section consists of a two-stage triode speech amplifier, with sufficient gain for a crystal microphone, and a modulator using a pair of 6L6s as Class AB power amplifiers. The plate and screen of the 807 are modulated simultaneously.

The cabinet contains two power units, one supplying the r.f. section, while the other supplies the audio section. A milliammeter is provided which may be switched to read the plate current of the oscillator, r.f. amplifier or modulator, or grid current of the r.f. amplifier.

The output-coupling system is suitable for use with resonant antenna systems or systems employing "flat" transmission lines. Apparatus must be supplied externally when the unit is to be used with antenna systems requiring tuning.

Measurements and Measuring Equipment

•12-1 MEASUREMENT TECHNIQUE

THE NATURE of radio and electronic equipment is such that visual inspection is of little value in determining satisfactory performance. A wide variety of test and measuring apparatus, based on the same principles as are used in the equipment itself, has been devised to enable accurate evaluation of the performance of experimental equipment as well as to facilitate production testing and adjustment and, ultimately, efficient maintenance in the field.

Because of the variety of such measuring and test equipment and its widespread application in all branches of the art, it is essential for the radio technician to be familiar with its principles and use.

Fundamentally, the process of measurement is that of comparing a quantity with a reference standard. Measuring equipment divides into two types: (1) fixed *standards* giving a reference point of known accuracy, with associated equipment for making comparisons, and (2) direct-reading instruments or *meters* calibrated in terms of the quantity being measured.

The basic quantities to be measured are: (1) frequency; (2) the primary electrical quantities: current, voltage and power; (3) circuit constants: resistance, impedance, capacity, inductance and Q; and (4) waveform. This chapter will describe commonly-used methods of measuring these quantities, and the application of these methods to the testing of specific types of equipment: receivers, transmitters, antennas, tubes.

•12-2 FREQUENCY MEASUREMENT

Frequency (§ 2-7) is measured by counting the number of cycles or oscillations per second. Since this cannot be done directly, except at very low frequencies, in practice the measurement is made (a) by noting the response of a selective resonant device, such as a tuned circuit (absorption frequency meter, Wien bridge, etc.) or mechanical resonator (tuning fork, vibrating reed, etc.) previously calibrated in terms of frequency, or (b) comparing the unknown with a known frequency from a separate source, either matching it directly by varying a calibrated source (heterodyne frequency meter), or measuring the difference between it and a fixed source (frequency standard), the frequency of which is known with high precision, by interpolation.

Calibrated Receiver — In the absence of more elaborate frequency-measuring equipment, a calibrated receiver may be used to indicate the approximate frequency of an oscillator. If the receiver is well-made and has good inherent stability, a band-spread dial calibration can be relied on to within perhaps 0.2 per cent. Some manufactured models having factory calibration may be used to even closer limits. For most accurate measurement the oscillator should be unmodulated and maximum response in the receiver indicated by a carrier-operated tuning indicator (§ 7-13), the receiver beat-oscillator being turned off.

In checking transmitting frequency the receiving antenna should be disconnected. If the signal is too strong and blocks the receiver, the transmitter frequency may be checked by listening to the oscillator, with power amplifier turned off.

Absorption frequency meters — The simplest type of frequency meter consists of a coil and condenser, tunable over the frequency range desired. A frequency meter of this type, when tuned to the frequency of the transmitter and loosely coupled to the tank coil, will extract a small amount of energy from the tank. The energy thus extracted can be used to light a small flashlight lamp. Maximum current will flow in the lamp when the frequency meter is tuned exactly to the transmitter frequency, hence the brightness of the lamp indicates resonance. A more accurate indication may be obtained by substitution of a thermo-galvanometer or vacuum-tube volt-meter as the indicator. A crystal detector can also be used.

Although this type of frequency meter is not well adapted to precise measurement of frequency, it is useful for checking (1) the fundamental frequency of an oscillating circuit, (2) presence and order of amplitude of harmonics, (3) frequency of parasitic oscillations, (4) neu-

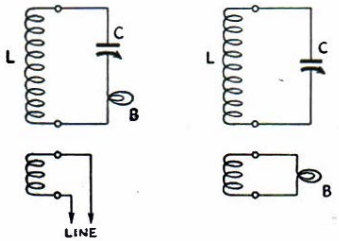


Fig. 1201 — A simple absorption frequency meter circuit is shown at left. It is used in transmitter checking with link line coupling to the circuit being checked. Circuit at right has bulb indicator loosely coupled to tuned circuit, giving a sharper resonance point.

B — 1.4-volt 50-ma. dial light.
C — 150-$\mu\mu$fd. variable.
L — Coils covering high-frequency spectrum with overlapping ranges, wound on 1½-inch dia. forms.

Freq. Range	*Wire Size*	*No. of Turns*	*Length of Winding*	*Link* [1]
1.1–3.5 Mc.	No. 28 e.	81¾	1⅞"	17 t.
2.5–8.0 Mc.	No. 24 t.	37¾	1⅝"	11 t.
4.5–14 Mc.	No. 20 t.	17¾	1½"	6 t.
7.5–25 Mc.	No. 16 t.	8¾	1¼"	4 t.
22–70 Mc.	No. 16 t.	2¾	1"	2 t.
40–120 Mc.	No. 16 t.	¾	—	¾ t.

[1] Closewound, No. 30 d.s.c., ¼-inch from bottom end of primary winding.



Fig. 1202 — A sensitive absorption frequency meter with a crystal detector-rectifier and d.c. milliammeter indicating circuit. Individual calibration charts mounted on each coil form make the meter direct-reading. A toggle switch connects a 10-ma. shunt across the 0–1 ma. meter; the 10-ma. range is used for preliminary readings, to avoid burning out meter or crystal. The meter gives indications several feet from a low-power oscillator.

tralization of an amplifier, (5) field strength on a qualitative basis, (6) presence of r.f. in undesired places such as power wiring, or any other application where detection of a small amount of r.f. and measurement of its frequency provides useful information.

Calibration of the absorption frequency meter is most easily accomplished with a receiver of the regenerative type to which the coil in the meter can be coupled. With the detector oscillating weakly, the frequency meter should be brought near the detector coil and tuned over its range until a setting is found which causes the detector to stop oscillating. The coupling between meter and receiver should then be loosened until the stoppage of oscillations occurs at only one spot on the meter tuning dial. The meter is then tuned to the frequency at which the receiver is set. If the receiver is set on several stations of known frequency, a number of points for a calibration curve can be obtained for each coil.

The same method may be used with a superheterodyne receiver, but it is necessary to remember that the oscillator frequency differs from the signal frequency by the intermediate frequency. For instance, if the receiver dial reads 6500 kc. and the receiver i.f. is 456, the oscillator frequency will be 6956 kc., which is the frequency to be marked on the meter calibration scale. It is necessary to know if the oscillator is on the high or low side of the incoming signal; in most receivers the high side is used throughout, but some receivers shift to the low side on the high-frequency ranges.

If the oscillator coils in the receiver are not accessible, the frequency meter may be capacity coupled through a few turns of insulated wire wrapped around the frequency-meter coil with one end of the wire placed near the stator plates of the oscillator condenser.

For transmitter frequency checking, a flashlight lamp or other indicator is not entirely necessary, since resonance will be indicated by a change in the plate current of the stage being checked as the meter is tuned through resonance. However, for locating parasitic oscillations, determining the relative amplitude of harmonics, checking neutralization, locating

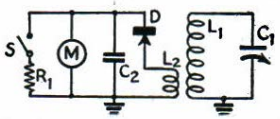


Fig. 1203 — Absorption frequency meter with crystal-detector indicator.

C_1 — 140-$\mu\mu$fd. variable.
C_2 — 0.001-μfd. mica.
D — Fixed crystal detector (Philmore).
L_1, L_2 — Same as in Fig. 1201.*
M — 0–1 ma. d.c. milliammeter.
R_1 — 3-ohm shunt; see general data on meter shunts.
S — S.p.s.t. toggle switch.

Crystal polarity must be determined by experiment; if meter reads backwards, reverse crystal connections.

* Experiment with number of turns on L_2 for maximum current indication is necessary to compensate for variations in impedance of individual crystal detectors.

stray r.f. fields, etc., a sensitive indicator is indispensable.

The inherent errors in the absorption-type frequency meter ordinarily limit its useful accuracy to about 1%.

Lecher wires — At ultrahigh frequencies it is possible to determine frequency by actually measuring the length of the waves generated. The measurement is made by observing standing waves on a two-wire transmission line or Lecher-wire system. Such a line shows pronounced resonance effects, and it is possible to determine quite accurately the current loops (points of maximum current) as shown in Fig. 1204. The distance between two consecutive current loops is equal to one-half wavelength. Thus the wavelength can be read off directly in meters (inches x 39.37 if a yardstick is used) or centimeters for the very short wavelengths.

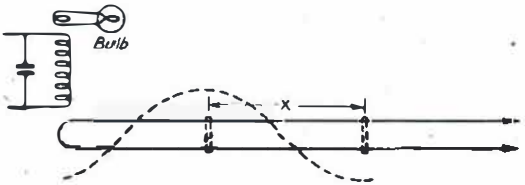


Fig. 1204 — Lecher wire system for measuring wavelength at u.h.f. Typical standing-wave distribution is shown, with positions of the shorting bar at current loops indicated. The distance "X" equals one-half wavelength.

The line should be at least a wavelength long at the lowest frequency to be measured and entirely air-insulated except where it is supported at the ends. The wires may be stretched tightly between any two convenient supports, using a spacing of 1 to 1½ inches. The positions of the current loops are found by means of a "shorting bar" which is slid along the line to vary its effective length (Fig. 1205).

Resonance indications may be obtained in several ways. A convenient and fairly sensitive indicator can be made by soldering the ends of a one-turn loop of wire to a flashlight bulb, then coupling the loop to the tank coil until the bulb glows moderately brightly. A similar coupling loop should be connected to the ends of the Lecher wires and brought near the tank coil. Then the shorting bar should be slid along the wires outward from the transmitter until the lamp shows a sharp loss in brightness. This point should be marked and the shorting bar moved out until a second dip is obtained. The distance between the two points will be equal to half the wavelength. If the measurement is made in inches, the frequency will be

$$F_{Mc.} = \frac{5906}{\text{length (inches)}}$$

If the length is measured in meters

$$F_{Mc.} = \frac{150}{\text{length (meters)}}$$

In checking a superregenerative receiver, the Lecher wires may be similarly coupled to the receiver coil. As the bar is slid along the wires a spot will be found where the receiver goes out of oscillation. The distance between two such spots is equal to a half wavelength.

Lecher wire measurements may easily be made to an accuracy of 1% or better. If sufficient care is used, measurements accurate to 0.1% at 112 Mc. are possible, representing a linear distance of about 1 millimeter. This is accomplished by loosening the coupling for the final adjustment until indications are just discernible. It is helpful to use a highly sensitive indicator. The crystal-detector absorption frequency meter previously described will enable closer measurements when used as a resonance indicator than will the flashlight bulb indicator, for example.

Heterodyne frequency meters—For more accurate measurement of transmitter frequency, a heterodyne frequency meter should be used. This is a small oscillator with a precise frequency calibration covering the lowest frequeney band used, completely shielded. It

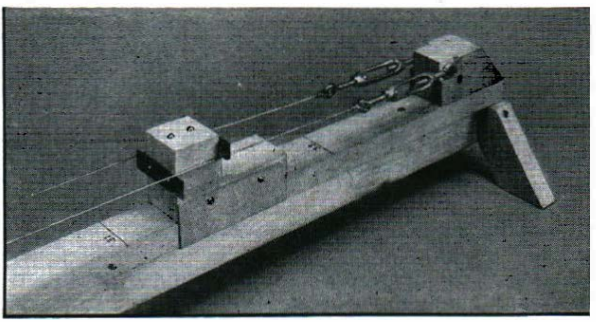


Fig. 1205 — One end of a typical Lecher wire system, showing turnbuckles for maiutaining tension. Wires are No. 16 bare copper. The shorting bar is of brass with a sharp edge for better contact and precise indication; the slider and side-guides keep the bar at right angles to the wire. A horizontal strip of bakelite at the back keeps the wires right against the bar. A transparent celluloid centimeter rule is cemented to the front of the slider. The beam is marked off in decimeter (10 cm.) units. The sum of the reading of the slider and the lowest adjacent numeral on the beam gives the wavelength.

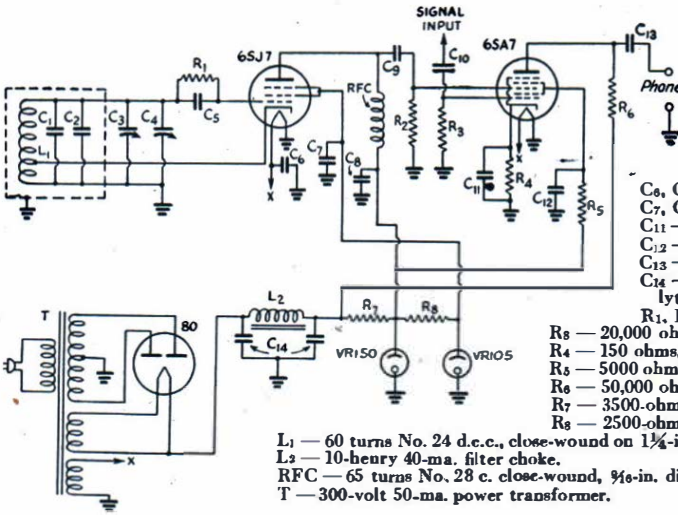


Fig. 1206 — Electron-coupled heterodyne frequency meter circuit diagram.

C_1 — 350-$\mu\mu$fd. zero temp.
C_2 — 40-$\mu\mu$fd. negative temp.
C_3 — 100-$\mu\mu$fd. midget.
C_4 — 50-$\mu\mu$fd. trimmer.
C_5 — 100-$\mu\mu$fd. silver-mica.
C_6, C_8 — 0.005-μfd. mica.
C_7, C_9, C_{10} — 0.002-μfd. mica.
C_{11} — 25-μfd., 25-volt electrolytic.
C_{12} — 0.01-μfd., 400-volt paper.
C_{13} — 0.5-μfd., 400-volt paper.
C_{14} — Dual 8-μfd., 450-volt electrolytic.
R_1, R_2 — 1 megohm, ½-watt.
R_3 — 20,000 ohms, ½-watt.
R_4 — 150 ohms, ½-watt.
R_5 — 5000 ohms, 1-watt.
R_6 — 50,000 ohms, 2-watt.
R_7 — 3500-ohm, 10-watt wire-wound.
R_8 — 2500-ohm, 5-watt wire-wound.
L_1 — 60 turns No. 24 d.c.c., close-wound on 1¼-in. dia. form tapped at 12 turns.
L_2 — 10-henry 40-ma. filter choke.
RFC — 65 turns No. 28 c. close-wound, 9/16-in. dia.
T — 300-volt 50-ma. power transformer.

must be so designed and constructed that it can be accurately calibrated and will retain its calibration over long periods of time.

The signal from this oscillator (or a harmonic thereof) is fed into a receiver or simple detector together with the signal to be measured, and the two frequencies are heterodyned. When the frequency meter oscillator is tuned to zero beat with the signal, its frequency or the harmonic multiple is the same as the unknown.

The oscillator used in the frequency meter must be very stable. Mechanical considerations are most important in its construction. No matter how good the instrument may be electrically, its accuracy cannot be depended upon if it is flimsily built. Inherent frequency stability can be improved by avoiding the use of phenolic compounds and plastics (bakelite, polystyrene, etc.) in the oscillator circuit, employing only high-grade ceramics for insulation. Plug-in coils or switches are not ordinarily used; instead, a solidly-built and firmly-mounted tuned circuit is permanently installed and the oscillator panel and chassis reinforced for rigidity.

To obtain high accuracy the frequency meter must have a dial that can be read precisely to at least one part in 500; ordinary dials like those used on transmitters and inexpensive receivers are not capable of such precision without the addition of vernier scales. The dial should have fine lines for division marks, and an indicator set close to the dial scale so that the readings will not appear different because of parallax when the dial is viewed from different angles.

A stable oscillator suitable for use in the frequency meter is one using the electron-coupled circuit (§ 4-2). The oscillation frequency is practically independent of moderate variations in supply voltages, provided the plate and screen voltages are properly proportioned. It is possible to take output from the plate with but negligible effect on the frequency of the oscillator. A third feature is that strong harmonics are generated in the plate circuit so that the meter is useful over a number of frequency bands. A typical electron-coupled frequency meter circuit is shown in Fig. 1206.

When the frequency meter is first turned on some little time is required for the tube to reach its final operating temperature; during this period the frequency of oscillation will drift slightly. It is desirable, therefore, to allow the meter to "warm up" for about a half hour before calibrating, or before making measurements in which the utmost accuracy is desired. Better still, the frequency meter can be left on permanently. The power consumption is negligible, and the long-time stability will be vastly improved.

Although some frequency drift is unavoidable, it can be minimized by the use of voltage regulator tubes in the power supply and low-drift silvered-mica or zero temperature-coefficient fixed condensers in the tuned circuit. A small negative temperature-coefficient capacity can be included to compensate for residual drift.

Calibration of the frequency meter is readily accomplished if a low-frequency standard (discussed later in this chapter) is available, the required calibration points being supplied by harmonics from the standard. The frequency meter is tuned to zero-beat (§ 7-4) with these harmonics, using either a built-in detector or the station receiver to combine the two signals to provide an audible beat. When a sufficient

number of points have been established, they may be marked on graph paper and a calibration curve drawn. For maximum convenience, a direct-reading dial scale can be constructed.

If no frequency standard is available, calibration points may be obtained from other sources of known frequency, such as the transmitter crystal oscillator, harmonics of local broadcasting stations, or even checks by other amateurs on the air. As many such points as possible should be secured, so that inaccuracies will average out.

In use, the signal from the frequency meter can be fed into the receiver by connecting a wire from the plate of the oscillator through a very small capacity to the input of the receiver. The signal to be measured is then tuned in in the usual way and the frequency meter adjusted to zero-beat.

For convenience in checking the frequency of the transmitter or other local oscillators which generate sufficiently strong signals, it is desirable to incorporate a detector in the frequency meter which will combine the signals and deliver the audio beat-note output to headphones or a visual zero-beat indicator. A frequency converter tube such as the 6L7 or 6SA7 is especially suited for this purpose.

With a stable oscillator, a precision dial and frequent and careful calibration, an overall accuracy of 0.05 to 0.1% may be expected of the heterodyne frequency meter. The principal limiting factors are the precision with which the calibrated dial can be read and the "re-set" stability of the tuned circuit.

Frequency standards — To make more precise frequency measurements, particularly of amateur-band limits, a secondary frequency standard is required. This is a highly stable low-frequency oscillator, usually operated at 50 or 100 kc., the harmonics of which provide reference points every 50 or 100 kc. throughout the spectrum. A 1000-kc. frequency is often added to facilitate preliminary identification of frequency ranges, especially on u.h.f.

An electron-coupled oscillator built according to the principles previously outlined for frequency meters, equipped with a tuned circuit for 50 or 100 kc., will serve as a simple and inexpensive standard. A standard of this type is inherently more accurate than a heterodyne frequency meter because (a) the low-frequency oscillator has better inherent stability and (b) the frequency setting once made is not thereafter changed, eliminating the re-set and calibration errors.

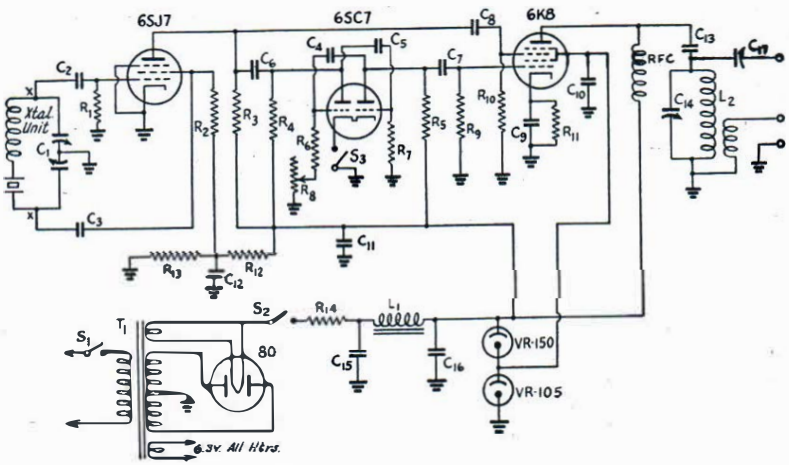


Fig. 1207 — Circuit diagram of a precision crystal-controlled 100-kc. frequency standard.

C_1 — Dual 365-$\mu\mu$fd. variable, compact broadcast type.
C_2, C_3 — 0.01-μfd., 400-volt paper.
C_4, C_5 — 0.001-μfd. midget mica.
C_6, C_7 — 10-$\mu\mu$fd. midget mica.
C_8 — 50-$\mu\mu$fd. midget mica.
C_9, C_{10}, C_{11}, C_{12} — 0.1-μfd. 400-volt paper.
C_{13} — 0.002-μfd. midget mica.
C_{14} — 140-$\mu\mu$fd. variable.
C_{15}, C_{16} — 8-μfd., 450-volt elec.
C_{17} — 3-30-$\mu\mu$fd. trimmer.
R_1 — 1 megohm, ½-watt.
R_2, R_3 — 0.5 megohm, 1-watt.
R_4, R_5 — 50,000 ohms, 1-watt.
R_6, R_7 — 20,000 ohms, ½-watt.
R_8 — 15,000-ohm potentiometer.
R_9 — 0.3 megohm, ½-watt.
R_{10} — 0.1 megohm, ½-watt.
R_{11} — 800 ohms, ½-watt.
R_{12} — 25,000 ohms, 1-watt.
R_{13} — 50,000 ohms, 1-watt.
R_{14} — 1500 ohms, 10-watt.
RFC — 2.5 mh. r.f. choke.
S_1, S_2, S_3 — S.p.s.t. toggle.
T_1 — Power transformer, 250 v., 40 ma.
L_1 — 7-henry, 40-ma. filter choke.
L_2 — Plug-in coil tuned to frequency in use. Output link is adjusted to give desired signal strength in receiver.

The crystal is a Bliley SOC-100 (oscillator coil in same mounting). For checking 1000-kc. points, a 150 microhenry coil (No. 30 d.c.c. on 1½-inch form) may be substituted for the crystal unit, connected between points X-X in the diagram. With C_1 near maximum, circuit will tune to 1000 kc.

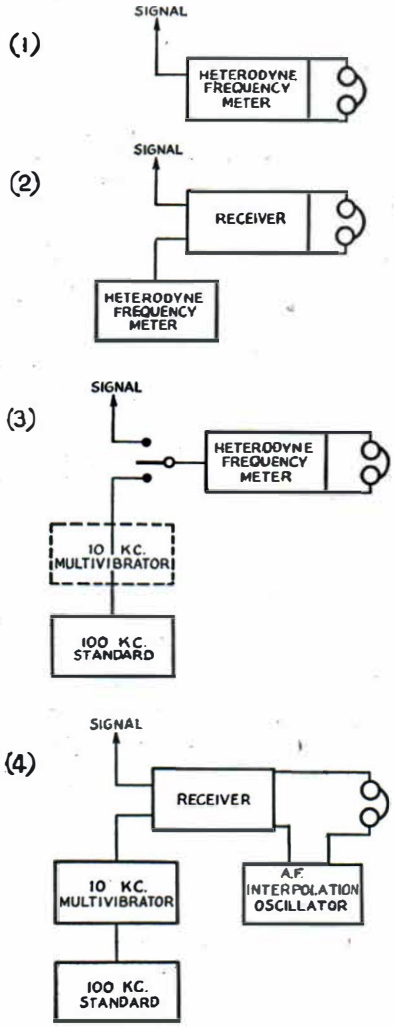


Fig. 1208 — **Frequency measurement methods.**

A frequency meter (with built-in detector) used alone is the simplest arrangement for checking the frequency of local oscillators (1). With a receiver (2) incoming received signals can be measured as well. A heterodyne frequency meter can also be used as a linear interpolation oscillator in conjunction with a 100-kc. standard (3), with or without a 10-kc. multivibrator. The standard provides accurate check points on the frequency meter scale. Alternatively, a receiver (if adequately calibrated) may be substituted for the frequency meter. For greatest precision, method (4) is used with an interpolation audio oscillator having a linear scale.

With careful design and construction, high precision can be attained with methods (3) and (4). Using (3), the accuracy can be 0.01% (100 parts in a million). Method (4) is accurate to 10 parts in a million with ordinary equipment; precision laboratory apparatus is reliable to better than 1 part in a million.

For highest accuracy, the most suitable instrument is a crystal-controlled standard, provided with a 10-kc. multivibrator (§ 3-7) for frequency division. Such a standard will mark 10-kc. intervals throughout the communications spectrum. The frequency of a signal can then be checked by noting its location with respect to two adjacent 10-kc. points on the dial of a calibrated receiver or heterodyne frequency meter and estimating the exact frequency by interpolation.

In the adjustment of the frequency standard at least a 15-minute warm-up period should be allowed. For initial adjustment, its output may be coupled into a receiver operating on the broadcast band and the oscillator tuned to zero beat with a broadcasting station on a frequency that is a multiple of 100 kc. (800 kc., 900 kc., 1000 kc., etc.) If the oscillator is self-excited, a second station 100 kc. away should be checked, to make certain that the oscillator is working on 50 or 100 kc. rather than another frequency which gives an odd harmonic. Since broadcasting stations are required to stay within 20 cycles of assigned frequency, the maximum error of such a source will be less than 30 parts in one million.

For highest precision, the standard may be calibrated on the National Bureau of Standards WWV standard frequency transmissions, which are accurate to better than 1 part in 10 million. These transmissions may be tuned in on a receiver operating on 5 Mc. (receiver beat oscillator off) and the standard adjusted until its harmonic is exactly at zero-beat with WWV. The calibration should be rechecked whenever precise measurements are to be made.

In adjusting the multivibrator, two adjacent 100-kc. points are first noted on the dial of a calibrated receiver. The multivibrator is then turned on, and its frequency control (R_8 in Fig. 1207) set at half scale. The number of separate audio beats *between* the two marked 100-kc. points is then counted. If it is a number other than nine (indicating 10-kc. intervals), readjust R_8 until nine beats are observed. Mark this point. Note also the points on the R_8 scale where 8 and 10 beats occur, indicating approximately 11- and 9-kc. separation. The odd frequencies are occasionally useful in checking frequencies very close to the 10-kc. harmonics where the low beat-frequency makes it difficult to secure zero-beat, particularly when an interpolation oscillator is used. Mathematical calculation is required to determine the exact frequency.

In practice the 100-kc. points can usually be identified as being louder than the 10-kc. multivibrator harmonics. This identification process can be facilitated by applying audio modulation to the 100-kc. signal only, causing

the modulated points to stand out because of the distinctive tone.

Interpolation — When measuring exact frequencies with the aid of a frequency standard and multivibrator providing equi-spaced harmonic points, it is necessary to determine the exact location of the unknown frequency by interpolation between adjacent standard harmonics. This can be done (a) by use of a calibrated receiver or heterodyne frequency meter with a scale that is linear with frequency, or (b) by comparison of the audio beat frequency with a calibrated audio oscillator.

In method (a), the points at which the unknown frequency and the nearest lower and higher harmonics appear on the dial of the receiver or frequency meter are noted, as shown in Fig. 1209. Knowing the exact frequencies of the harmonic points f_1 and f_2, the unknown frequency, f_x, can be determined as follows

$$f_x = f_1 + \frac{S_x - S_1}{S_2 - S_1}(f_2 - f_1)$$

where S_1 is the dial setting for f_1, S_2 for f_2 and S_x for f_x.

Method (b) consists of beating the standard and unknown frequencies in a detector and measuring the resulting audio frequency by zero-beating with a calibrated audio oscillator having a linear frequency range covering half the difference between adjacent harmonics (0–5000 cycles with a 10-kc. multivibrator), as shown in Fig. 1209. The measured frequency is then equal to the reading of the audio oscillator added to or subtracted from the nearest standard harmonic, as the case may be. To determine whether to add or subtract this audio difference it is necessary that the frequency be known to better than 5 kc. from the receiver (or auxiliary heterodyne frequency meter) calibration.

In addition to the beat note resulting from the nearest adjacent harmonic, f_1, there will also be another higher beat from f_2. However, by tuning the receiver midway between f_1 and f_x, its adjacent-channel selectivity will discriminate against f_2 and reduce the higher beat note to a negligible level.

The interpolation audio oscillator should have a scale that reads linearly with frequency (as opposed to the logarithmic scale commonly found in laboratory oscillators). A beat-frequency oscillator (Fig. 1228) with a straight-line capacity tuning condenser in series with the correct value of fixed capacity will have a nearly linear scale. Some forms of resistance-capacity oscillators can also be made to have such a scale.

A suitable detector is a pentagrid converter (§ 7-9) with some form of zero-beat indicator in the plate circuit. The interpolation audio oscillator is connected to the oscillator grid, the audio beat note from the receiver being applied to the signal grid.

Zero-Beat Indicators — Use of the heterodyne method of frequency comparison requires a means for determining when the known and unknown frequencies are synchronized; i.e., when they are at zero beat. The point at which zero-beat occurs can be determined approxi-

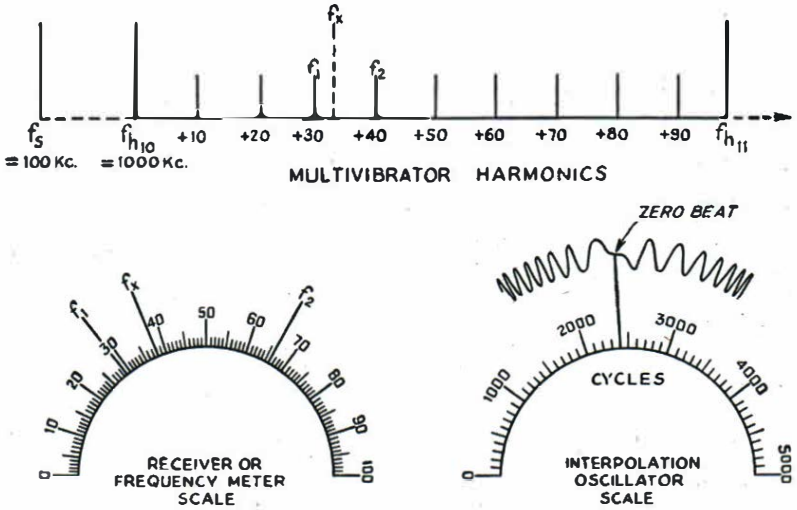


Fig. 1209— Use of interpolation methods in measuring frequencies between standard harmonics. At top is shown the relative location of the frequency-standard fundamental and harmonics in the spectrum, together with the multivibrator harmonics, as related to the unknown frequency under measurement (f_x). At the left is shown a small segment of this spectrum as it appears on the dial of a calibrated receiver or heterodyne frequency meter, and at right the appearance of the audio oscillator dial when using the comparison audio beat-note method.

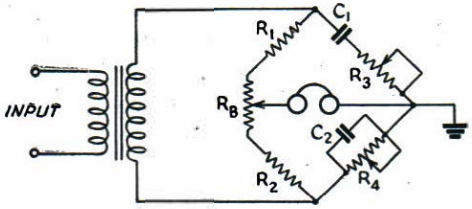


Fig. 1210 — Wien bridge for audio frequency measurements.

R_1 — 2000 ohms, ½-watt.
R_2 — 1000 ohms, ½-watt.
R_3, R_4 — 10,000-ohm wire-wound variable.
R_B — 25-ohm potentiometer.
C_1, C_2 — 0.01-μfd. for 2,000–20,000 cycles
0.1-μfd. for 200–2,000 cycles
1.0-μfd. for 20-200 cycles
The input transformer should be of the balanced-to-ground type with an electrostatic shield between primary and secondary.

mately by listening to the output of the receiver or detector with headphones or loud speaker. For greatest accuracy some form of auxiliary visual zero-beat indicator is desirable, however. This may be a rectifier-type a.f. voltmeter with copper-oxide or diode rectifier (§ 2-3), a neon tube "flasher" or an electron-ray tube (§ 7-13) with its triode grid connected to the receiver output. The headphones will still be required for preliminary adjustments since the visual indicator usually responds only to frequencies under about 25 cycles.

Audio-Frequency-Measurement — The measurement of unknown audio frequencies can also be accomplished by either direct or comparison methods. Direct measurements are best made with an a.c. bridge (§ 2-11). The most suitable form is the Wien bridge as shown in Fig. 1210. The a.c. voltage to be measured is applied across the input terminals and the variable resistors R_3 and R_4 varied simultaneously until balance is obtained as shown by zero response in the indicator. Then

$$f = \frac{1}{2\pi R C}$$

where R is the resistance effectively in the circuit at R_3 and C is equal to the value of C_1. R_1 and R_2 may be ganged on one shaft and provided with a scale calibrated in terms of frequency for each value of C (the ranges being selected by a 2-gang switch). R_B is a small balancing potentiometer used to compensate for minor variations in the two arms. The indicator may be a pair of headphones as shown, although a sensitive visual indicator will give greater range and higher accuracy.

Where a calibrated audio oscillator is available, measurements may be made by comparison as previously described in this chapter. If no electrical frequency standard is available, the audio frequency can be converted into sound through a power amplifier and loudspeaker and measured by aural comparison with a properly-tuned piano, remembering that middle C is 256 cycles and each octave above or below doubles or halves the frequency. Intermediate points can be obtained by multiplying each successive half-note above C in any octave by 1.05946 (e.g., if C is 1, C# equals 1.05946, D equals 1.1225, etc.).

The cathode-ray oscilloscope (§ 3-9) is extremely useful in measuring frequencies by the comparison method when a reliable standard

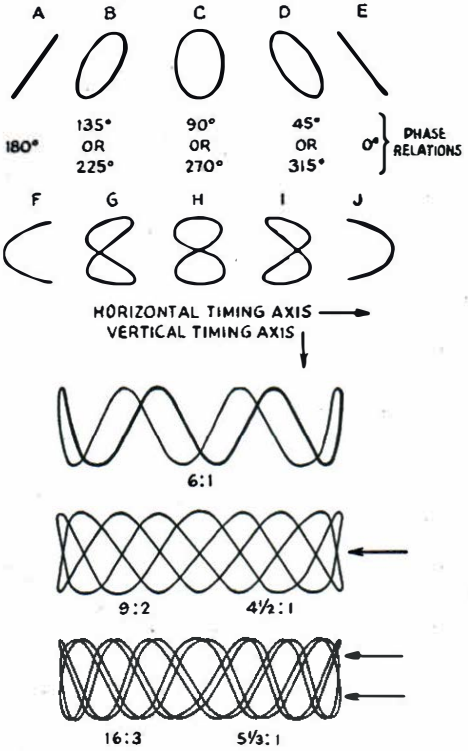


Fig. 1211 — Lissajou's figures as used in measuring audio frequencies by comparison with a known source on a cathode-ray oscilloscope. Figures A through E illustrate the pattern produced by different phase relationships when the two voltages have a 1:1 frequency ratio. Figures F through J show the same phase relationships with a 2:1 frequency ratio, the higher frequency being applied to the vertical plates. The next figure shows a ratio of 6:1, determined by counting the peaks of the waves in the horizontal plane (in this case the higher frequency is applied to the horizontal plates). Complex ratios are identified by one or more cross-overs, as indicated by the arrows opposite the 9:2 and 16:3 figures. In principle, frequency ratios are determined by counting both horizontal and vertical peaks (number of cross-overs plus 1). Care must be taken not to confuse the back lines (return trace shown by light line in 6:1 figure) in counting cross-overs. This can be done by counting only the peaks travelling in the same direction across the screen when the frequency is adjusted so that the pattern rotates slowly.

source is available. Applying voltages from the unknown and the standard to the opposite pairs of cathode-ray tube deflecting plates results in patterns of varying form termed Lissajou's figures. By proper interpretation of these figures, as shown in Fig. 1211, frequency ratios up to 10 to 1 can be obtained conveniently. Thus with a 1000-cycle oscillator calibration points between 100 and 10,000 cycles are possible. The 60-cycle a.c. supply can be used as a calibration source up to 600 cycles or so.

Frequency Monitoring — Contrasted with the problem of frequency measurement as a single operation is that of continuously monitoring the frequency of an oscillator, particularly in broadcasting and other single-frequency transmitters. This requires an automatic device continuously reading frequency deviation. Common practice is to provide a highly-stable temperature-controlled standard having a frequency slightly different (usually 1 kc.) from the assigned carrier frequency. Both carrier and standard frequencies are fed into a detector and the resulting beat measured on a direct-indicating audio-frequency meter. Variations in the audio beat then show deviations of the carrier frequency in either direction. Alternatively, selective circuits and vacuum-tube relays controlling panel lamps may be used instead of the meter, the lamps flashing to indicate deviations beyond pre-set upper and lower limits.

●12-3 MEASUREMENT OF CURRENT, VOLTAGE AND POWER

D.c. instruments — Instruments for measuring direct current (§ 2-6) are based on the d'Arsonval moving-coil principle, comprising an indicating pointer moving across a calibrated scale, actuated by the flow of current through a coil located in a constant magnetic field.

Ammeters and voltmeters are basically identical instruments, the difference being in the method of connection. An ammeter is connected in series with the circuit and measures the current flow. A voltmeter is a milliammeter (ammeter reading one-thousandth of an ampere) which measures the current through a high resistance connected across the source to be measured; its calibration is in terms of voltage drop in the resistance, or *multiplier*.

The ranges of both voltmeters and ammeters can be extended by the use of external resistors, connected in series with the instrument in the case of a voltmeter or in shunt in the case of an ammeter. Fig. 1212 shows at (A) the manner in which a shunt is connected to extend the range of an ammeter and at (B) the connection of a voltmeter multiplier.

To calculate the value of a shunt or multiplier, it is necessary to know the resistance of the meter. If it is desired to extend the range of a voltmeter, the value of resistance which must be added in series is given by the formula:

$$R = R_m (n - 1)$$

where R is the multiplier resistance, R_m the resistance of the voltmeter, and n the scale

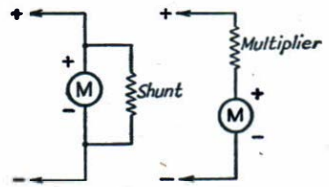


Fig. 1212 — How voltmeter multipliers and milliammeter shunts are connected.

multiplication factor. For example, if the range of a 10-volt meter is to be extended to 1000 volts, n is equal to 1000/10 or 100.

If a milliammeter is to be used as a voltmeter, the value of series resistance can be found by Ohm's law (§ 2-6)

$$R = \frac{1000\ E}{I}$$

where E is the desired full-scale voltage and I the full-scale current reading of the instrument in milliamperes.

To increase the current range of a milliammeter, the resistance of the shunt can be found from the formula:

$$R' = \frac{R_m}{n - 1}$$

where R_m is the meter resistance as before.

A portable combination milliammeter-voltmeter having several ranges is extremely useful for experimental purposes and for trouble-shooting in receivers and transmitters. As a voltmeter such an instrument should have high resistance so that very little current will be drawn in making voltage measurements. A low-resistance voltmeter will give inaccurate readings when connected across a high-resistance circuit. A resistance of 1000 ohms per volt is satisfactory for most uses; a 0-1 milliammeter or 0-500 microammeter (0-0.5 ma.) is the basis of most multi-range meters of this type. Microammeters having a sensitivity of 0-50 μa., giving a voltmeter resistance of 20,000 ohms per volt, are found in units available at reasonable cost. Multipliers for the various ranges are selected by switches.

The various current ranges on a multi-range instrument are also obtained by using a number of shunts individually switched in parallel with the meter. Particular care must be taken to minimize contact resistance.

When d.c. voltage and current are accurately

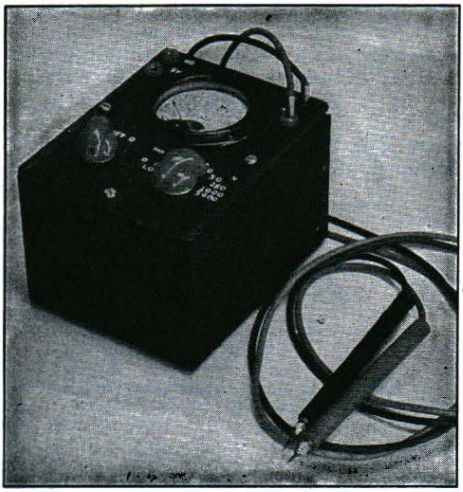


Fig. 1213 — A typical inexpensive multi-range combination volt-ohm-milliammeter, showing meter with 1-ma. basic range and knobs for range switch and zero re-setting resistor. Flexible test leads with insulated-handle test probes are used to make connection to the circuit under test.

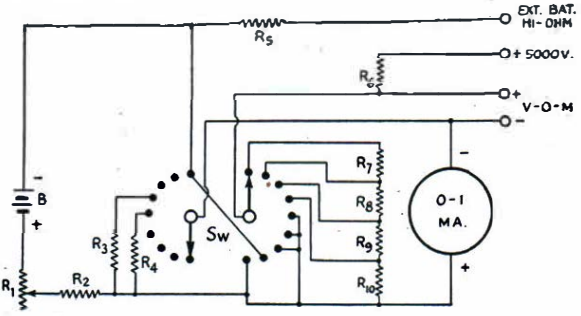


Fig. 1214 — Circuit of the low-cost V-O-M.

R_1 — 2000-ohm wire-wound variable.
R_2 — 3000 ohms, ½-watt.
R_3 — 10-ma. shunt, 3.6 ohms.
R_4 — 100-ma. shunt, 0.33 ohms.
R_5 — 40,000 ohms, ½-watt.
R_6 — 4 megohms, 4-watt (four 1-meg., 1-watt resistors in series).
R_7 — 0.75 megohm, 1-watt (0.5 megohm and 0.25 megohm, ½-watt in series).
R_8 — 0.2 megohm, ½-watt.
R_9 — 40,000 ohms, ½-watt.
R_{10} — 10,000 ohms, ½-watt.
SW — 9-point 2-pole switch (Mallory-Yaxley 3109).
B — 4.5 volts (Burgess 5360).

known, the power can be stated by simple application of Ohm's law: $P = EI$ (§ 2-6).

A.c. instruments — D.c. meters will not respond on alternating current, and it is therefore necessary either to rectify the a.c. and measure the resulting d.c. or to use special instruments that will indicate on a.c. (§ 2-8).

A.c. ammeters and voltmeters utilize the moving iron-vane principle. Since the maximum sensitivity is 15 to 25 ma., making the ohms-per-volt 40 to 67, iron-vane voltmeters consume substantial power. They are suitable for measuring filament and line voltages, but cannot be used in circuits which are unable to sustain a measuring load. Iron-vane meters are not accurate above a few hundred cycles.

For measurements where iron-vane meters are not suitable, special devices enabling the use of d.c. movements are employed. The most common of these for the power and audio frequency range is the full-wave copper-oxide rectifier, which converts a low-resistance 0-1 d.c. milliammeter into a high-resistance 0-0.909 a.c. milliammeter, making possible the construction of a.c. voltmeters having a sensitivity of 1000 ohms per volt and an accuracy of about 5%. The design of multipliers for such a voltmeter must allow for the fact that the rectifier resistance varies with current. Two scales are usually provided, one for use above 50 volts and one below. The frequency error is approximately 0.5% per 1000 cycles.

A.c. power measurements are more complex than for d.c., the simple multiplication of current and voltage being in error unless the load is purely resistive. If the current and a.c. impedance are known, the power is I^2Z. For ordinary amateur power calculations, such as the input to a power transformer, the product of a.c. voltage and current can be considered sufficiently accurate. Commercial power measurements are made on a wattmeter, which is a complex instrument relating average current and voltage in terms of a single reading.

R.f. instruments — The measurement of very high-frequency a.c. or r.f. quantities involves special problems. Practical instru-

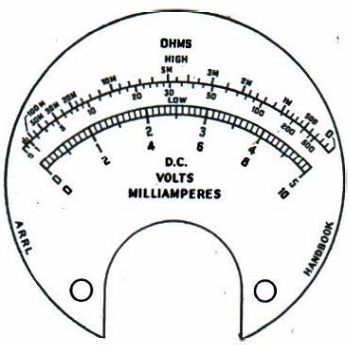


Fig. 1215 — Scale of the typical multi-purpose meter. It is read in accordance with the range in use. In this meter the ranges are "high" (0–250,000) and "low" (0–500) ohms, 100, 10 and 1 ma., and 10, 50, 250, 1000 and 5000 volts.

ments read in terms of d.c. from a conversion device.

R.f. current is usually measured by means of a thermoammeter. This is a sensitive d.c. microammeter connected to a thermocouple associated with a heater made of a short piece of resistance wire. Thermoammeters have been made with a sensitivity of 1 ma., but the ranges used by amateurs for measuring antenna current, etc., are from 0-0.5 amperes up.

The most suitable r.f. voltmeter is a peak-reading vacuum-tube voltmeter (Fig. 1216). When properly designed, its accuracy is limited at r.f. only by the variation of the input resistance with frequency. The peak diode voltmeter has little error even at 60 Mc. The same is true of the self-biased and slide-back types if tubes having low input capacity are used. The oscilloscope can also be used as an r.f. voltmeter for potentials of several volts or more.

R.f. power measurements can be made by measuring the current through a resistor or reactance of known value (§ 4-9). Approximate power measurements may be made by using ordinary 115-volt light bulbs as a substitution or "dummy" load, connected either singly or in series-parallel to provide the required resistance and power rating. The approximate resistance of the bulb can be computed from its wattage rating at 60 cycles. Special non-inductive resistance units, enclosed in vacuum bulbs mounted on standard tube bases, with resistances of 73 and 600 ohms at power ratings up to 100 watts, are available for this purpose. For higher power the units may be connected in series-parallel.

Where the substitution load method is impractical, r.f. power can be measured by multiplying the current through a thermoammeter in the circuit by the r.f. voltage across the circuit as indicated by an r.m.s. meter (or 70.7% of the reading on a peak voltmeter).

Another method of measuring r.f. power is the photometric method. A calibrated light-sensitive cell (a photographer's exposure meter is suitable) is used to measure the relative brilliance of an electric light bulb as a substitution load and its normal brilliance on 115-volt 60-cycle supply.

Vacuum-tube voltmeters — The most useful instrument for the measurement of both d.c. and a.c. voltages is the vacuum-tube voltmeter. Its chief advantages are (a) negligible power taken from the circuit under measurement and (b) accuracy over a wide frequency range including r.f.

The v.t.v.m. operates by virtue of the change in plate current caused by a change in grid voltage (§ 3-2). In the measurement of a d.c. voltage, the voltage — termed the "signal" — represents simply a change in grid bias. In

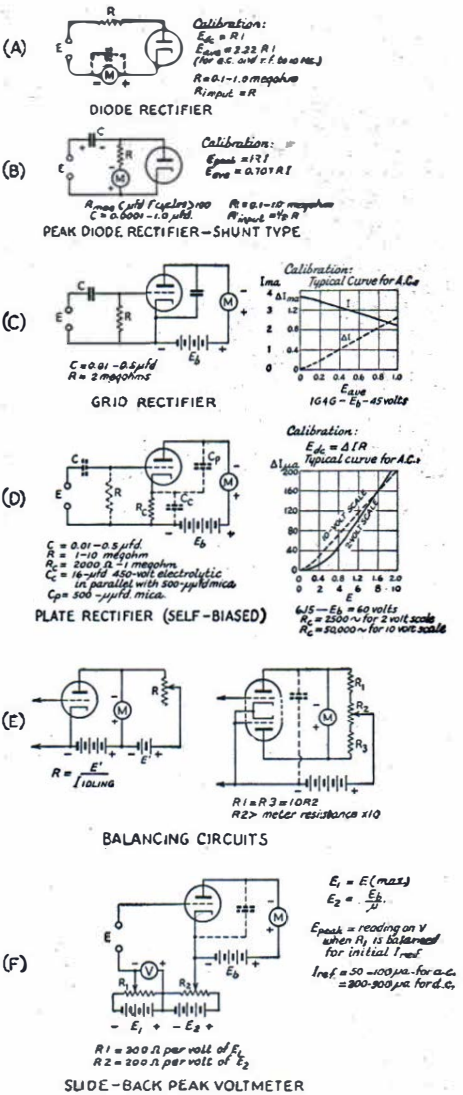


Fig. 1216 — **Vacuum-tube voltmeter circuits.**

the case of a.c., the tube acts as a rectifier and the measurement is in terms of rectified d.c.

Representative vacuum-tube voltmeter circuits are shown in Fig. 1216. The simple diode rectifier (A) can be almost any vacuum tube; in a triode or multi-grid type, all electrodes except the control grid are connected to cathode (or negative filament). A Type 30 or 1G4G tube with a flashlight cell for filament supply makes a convenient portable unit. A tube with low input capacity (1N5G, 6T7G, 954) should

be used for high frequencies. The frequency range is limited by the tube input capacity shunting the load resistance. Calibration is linear above 2 or 3 volts provided the load resistance exceeds 0.1 megohm. The meter *M* should be a sensitive microammeter (0-100 or 200 μa.); a 0-1 ma. meter can be used with reduced sensitivity.

The peak diode voltmeter at (B), shunt-connected to eliminate the need for a d.c. return in the measured circuit, reads peak a.c. voltage. The input resistance is comparable to that of the simple diode for equivalent sensitivity, but the high-frequency error is less. The time constant of the *RC* circuit should be at least 100 for the lowest frequency to be measured (*RCF>100*). Typical values are 0.5 megohm and 0.5 μfd. for audio, and 0.1 megohm and 0.05 μfd. (mica) for r.f. and i.f.

The grid rectification circuit shown at (C) can be considered equivalent to the diode rectifier of (B) followed by a zero-bias triode amplifier. The sensitivity is greatly increased over the ordinary diode. The input resistance is low with small inputs (0.1 to 1.0 megohm) because of grid current. The plate current is at maximum when idling and decreases with signal. This circuit is useful chiefly because it can be used with inexpensive meters. The instrument can be calibrated from a known 60-cycle source; the scale is square-law for small signals, becoming linear with increasing input. The value of *R* is non-critical. *C* should have a reactance small compared with *R* at the operating frequency, i.e., 0.01 μfd. mica from 1 kc. up, 0.1 μfd. paper for low a.f. For d.c. *C* is, of course, omitted. A high-μ tube is preferable, to reduce the idling or no-signal plate current.

The self-biased plate-rectification or reflex voltmeter at (D) has a very high input resistance and fair sensitivity. It is normally connected directly across the circuit to be measured; if this does not provide a d.c. return a coupling circuit must be added as shown by dotted lines (C – 0.01 μfd., R – 10 megohms or more). A low-μ tube is preferable, to minimize contact potential and grid current. The cathode resistance R_c controls sensitivity; the higher it is the more nearly linear and stable will be the calibration. A range switch can be provided, connecting in various values of cathode resistance from 2000 ohms to 0.5 megohm to give full-scale ranges from about 2.5 to 250 volts. The plate and cathode bypasses may be 0.001-μfd. mica condensers, the cathode being shunted by a 10-μfd. electrolytic for 60-cycle calibration and low a.f. measurements.

The no-signal plate current present in the circuits of (C) and (D) can be balanced out by bridge or bucking circuits, typical forms of which are shown at (E). An auxiliary battery (or section of the voltage divider, in an a.c. power supply) is connected back to the meter through a variable resistor, providing a controllable opposite current flow which can be made to equal exactly the residual plate current of the tube. When used with (C), this balance circuit allows the meter terminals to be reversed, thereby making it read forward instead of backward with signal. The resistor *R* should be not less than ten times the internal resistance of the meter.

At the right in (E) an automatic balancing circuit is shown wherein a duplicate triode (usually the second section of a twin triode) takes the place of the adjustable resistor *R*. Current flow through R_1 and R_2 being equal and opposite with no signal, there is no potential across the meter and consequently no current flow. When a voltage is placed on the grid of the voltmeter triode this balance is disturbed, however, and the meter registers current flow. A small zero-setting resistor, R_3, is provided to correct for any discrepancies in tubes or resistors. The values of R_1 and R_2 will depend on the plate supply voltage available; the higher they are, the better the sensitivity and stability. The minimum value is several times the meter resistance.

The "slide-back" voltmeter at (F) is a comparison instrument in which the peak value of an a.c. or r.f. voltage is read in terms of a d.c. substitution voltage; the voltmeter tube and milliammeter, *M*, merely indicate when the two are equal. With the input terminals shorted and R_1 set so that *V* reads zero, the tube is biased nearly to cut-off by adjustment of R_2. The residual plate current is the reference current ($I_{ref.}$) or "false zero." If an a.c. voltage, *E*, is now applied across the input terminals, plate rectification of the positive peaks will cause the plate current to rise. By adjusting R_1, additional bias voltage may be introduced to balance out the a.c. voltage. The additional bias required to bring the plate current back to the reference value ($I_{ref.}$) is equal to the peak value of the signal being measured. In operation R_1 should be set (after setting $I_{ref.}$) so that all of E_1 is in the circuit, to avoid burning out the milliammeter when the signal is applied. After the unknown voltage has been connected the bias is reduced by R_1 until the reference current is reached. The slide-back voltmeter is capable of high accuracy and has the advantage of requiring no a.c. calibration.

Oscilloscopes — Perhaps the most useful of all measuring devices is the cathode-ray oscilloscope. (§ 3-9) Although relatively expensive, its applications are so numerous that it can replace a number of other less satisfactory types of measuring equipment. It can be used on d.c., a.c. and r.f., and is particularly suited

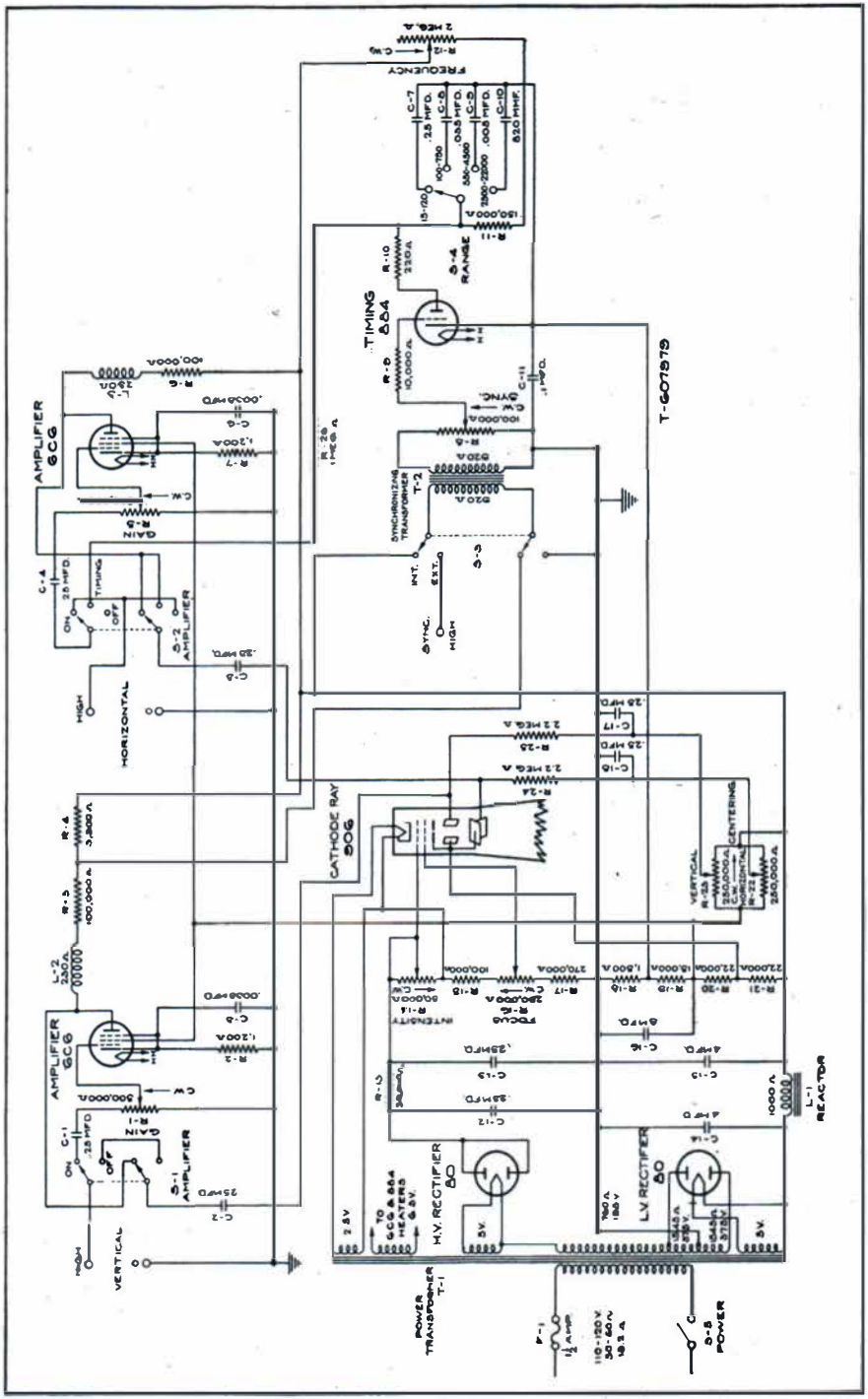


Fig. 1217 — Circuit diagram of a commercially-built cathode-ray oscilloscope with sweep circuit and voltage amplifiers. (Courtesy R.C.A.)



Fig. 1218 — A representative oscilloscope of the type widely used in amateur, service and commercial applications, using a 906 3-in. cathode-ray tube. The circuit is shown in Fig. 1217. All necessary controls are located on the front panel, including intensity, focus, timing circuit synchronization, frequency and range, and centering, gain and input selection controls for both vertical and horizontal axes. (Courtesy R.C.A.)

to a.f. and r.f. measurements because of the high input resistance and small frequency error.

The oscilloscope is in effect a complex voltmeter capable of measuring any two voltages simultaneously by the deflection of a weightless electron-beam pointer. Moreover, because this pointer projects its indication on a retentive luminous screen, the measurements include the additional factor of time. It is possible therefore to see the actual form of one or more repetitive cycles of an a.c. voltage by means of the oscilloscope, and to measure thereby not only its amplitude but also its frequency and waveform.

When used as a simple voltmeter the signal is applied to the vertical plates and its amplitude measured in terms of the height of the resulting trace. Approximate measurements can be made by calibrating the sensitivity of the cathode-ray in volts per inch. This varies with the anode voltage and type of tube; typical figures for small tubes are 25 to 75 volts per inch, peak-to-peak. The initial calibration can be made with a variable d.c. voltage and comparison voltmeter.

● 12-4 R, Z, C, L AND Q MEASUREMENTS

It is frequently necessary to measure the components used in the construction of amateur equipment — resistors, condensers, coils, etc. — both as a means of identification and in checking accuracy. The advanced amateur will also be interested in measuring impedances and the characteristics of devices of his own construction or under other than rated conditions.

Resistance — The volt-ammeter, ohmmeter and Wheatstone bridge methods are commonly used in measuring resistance. In the volt-ammeter method, the resistance is determined from Ohm's Law by measuring the current through the resistor when a known d.c. voltage is applied. The resistance can be determined with a voltmeter alone if its internal resistance is known

$$R = \frac{eR_m}{E} - R_m$$

where R is the resistance under measurement,
E is the voltage read on the meter,
e is the series voltage applied, and
R_m is the internal resistance of the meter (full-scale reading in volts x ohms-per-volt).

The ohmmeter is a practical application of this method, with a low-current d.c. voltmeter and a source of voltage (usually dry cells), connected in series with the unknown resistance. If the meter reads full-scale with the connecting leads shorted, insertion of the resistance under measurement will cause the reading to decrease in proportion to the amount of resistance inserted. The scale can therefore be calibrated in ohms.

In Fig. 1219-A, the series resistance is adjusted until the milliammeter reads full-scale when the test leads are shorted. When the meter reading changes as the battery ages, this resistance is reduced, compensating for the change. In (B) the series resistance is kept constant but the sensitivity of the meter is varied to compensate for the changing voltage. The circuit of (C) is useful for measuring resistances below a few hundred ohms. The unknown resistance is connected as a shunt across the meter, reducing the current reading. Values of a fraction of an ohm can be read in this way.

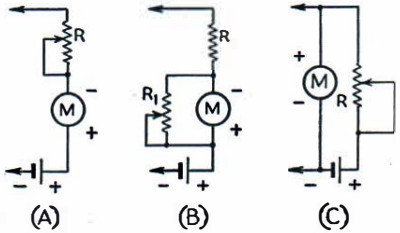


Fig. 1219 — Ohmmeter circuits. (A) Series ohmmeter with series compensation. (B) Series ohmmeter with shunt compensation. (C) Shunt ohmmeter for measuring low resistances.

The ratio of resistances which can be measured on a single ohmmeter range averages about 100 to 1, or from one-tenth to ten times the center-scale value.

Only approximate measurements can be made with an ohmmeter. For greater accuracy the unknown resistor may be compared with a standard resistance of known accuracy by means of a Wheatstone bridge (§ 2-11). If resistance measurements only are to be made, the bridge may be powered by a battery and a milliammeter used for the balance indicator. If reactances are also to be measured, an a.c. source is required (Fig. 1221).

Capacity and inductance — The capacity of condensers and the inductance of coils can be measured (a) in terms of their reactances, (b) by comparison with a standard, and (c) by substitution methods.

The reactance method is simplest but least accurate. The method is similar to the d.c. ohmmeter, except that impedance is measured instead of resistance. In Fig. 1220, at (A) the unknown reactance is placed in series with an a.c. rectifier-type voltmeter across the 115-volt a.c. line. With a 1000-ohms-per-volt meter, capacities can be identified from approximately 0.001-μfd. to 0.1-μfd. At (B) the reactance is connected in series with a 1000-ohm resistance; the proportionate voltage drop across this resistance indicates the reactance of condensers from 0.1 μfd. to 10 μfd. and inductances from 0.5 henry to 50 henries, when Q is greater than 10. Because the lower end of the scale of a rectifier-type meter is somewhat crowded, a better reading can be had by using the connection at (C) for large reactances. Approximate calibrations for each connection may be made by checking typical condensers and coils of known values and drawing calibration curves for the voltmeter in use.

The reactance method at best gives only approximate indications of inductance and capacity. For accurate measurements, an a.c. bridge must be used.

A simple bridge for the measurement of R, C and L is shown in Fig. 1221. Its accuracy will depend on the precision of the standards, the sensitivity of the detector or balance indicator, the voltage and frequency of the a.c. source, and the ratio of the unknown value to the standard. The signal source can be a 1000-cycle audio oscillator with low harmonic content and the detector a pair of headphones or an electron-ray tube.

For maximum accuracy the ratio of the unknown to the standard should be kept small, so that R is read near the center of its scale. The ratio can be as high as 10 to 1 in either direction with good accuracy, and an indication can be had even at 100 to 1. Additional standards can be included for other ranges if desired.

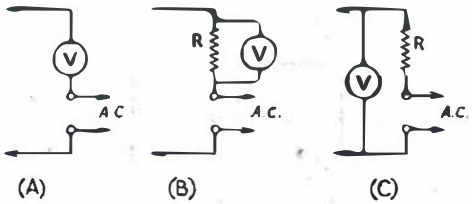


Fig. 1220 — Reactance-measurement circuits for checking capacity and inductance values.

The potentiometer R must be calibrated as accurately as possible in terms of the *ratio* of resistance on either side of its mid-point, which may be arbitrarily marked 10. If the potentiometer is next set at 500 ohms, the ratio of resistances is 1 to 10 and the scale may be marked 1. The corresponding point on the other end of the scale is marked 100. Intermediate points are similarly marked according to the resistance ratios. These ratios will then correspond with the ratio of the unknown resistance, inductance or capacity to the standard in use, when the bridge has been balanced for a null indication on the detector.

Since direct current flowing through a coil changes its inductance, allowance must be made for this in measuring choke coils and transformers carrying d.c.

Condensers should be checked for leakage as well as capacity. This check must be made with the rated d.c. voltage applied, a microammeter being connected in series with the high voltage source. The resistance of good paper condensers should be above 50 megohms per microfarad, that of mica above 100.

The condition of electrolytic condensers can be checked roughly with an ohmmeter. With the positive terminal of the condenser connected to the positive of the ohmmeter bat-

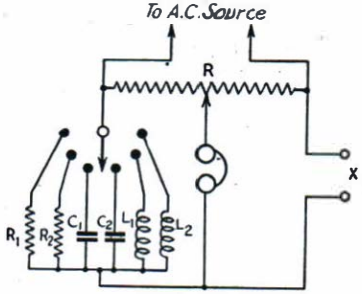


Fig. 1221 — Simple a.c. bridge for measuring resistance, inductance and capacity.
C_1 — 0.01-μfd. mica.
C_2 — 1.0-μfd. paper.
R — 10,000-ohm linear wire-wound potentiometer.
R_1 — 100 ohms, wire-wound (1% accuracy).
R_2 — 10,000 ohms, wire-wound (1%).
L_1 — 125-millihenry iron-core r.f. choke.
L_2 — 12-henry iron-core choke (Thordarson T-49C91).

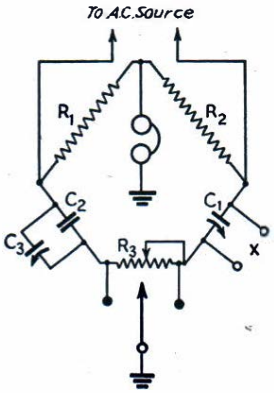


Fig. 1222 — Substitution-type capacity bridge.
C_1 — 1000-$\mu\mu$fd. straight-line-capacity condenser (may be two-gang 500-$\mu\mu$fd. with sections in parallel).
C_2 — 900-$\mu\mu$fd. silver mica.
C_3 — 100-$\mu\mu$fd. variable trimmer.
R_1, R_2 — 5000-ohm wire-wound (1%).
R_3 — 1000-ohm wire-wound potentiometer.

tery, high-voltage electrolytics should show a resistance of 0.5-megohm or so; low-voltage cathode by-pass condensers should be over 0.1 megohm. Electrolytics can also be checked by measuring the leakage current when the rated d.c. polarizing voltage is applied. It should read about 0.1 ma. per μfd. The maximum for a useful unit is about 0.5 ma. per μfd. Low leakage current also indicates a faulty unit. Electrolytic condensers which have lain idle on the shelf will show leakage currents as high as 2 ma. per μfd. per 100 volts. After "aging" for a few minutes at rated d.c. voltage they should return to normal, however.

The measurement of small capacities under 0.001 μfd. is not possible with a bridge of the type previously described because stray reactances cause errors. A more accurate bridge for measurement of small capacities is shown in Fig. 1222. It is of the substitution type with a calibrated air condenser, C_1, for the variable arm. C_2 is a fixed reference capacity. C_3 is used to balance out stray capacity including that of the leads to C_x. The bridge is first balanced by adjusting C_3, with C_1 at maximum capacity and the leads to C_x in place. C_x is then connected and the bridge again balanced by adjusting C_1. The difference in capacity (ΔC) of C_1 between its new setting and its maximum capacity represents the capacity of C_x.

It is impossible to get a zero null indication from the detector unless the resistances as well as the capacities of the two condensers being compared are equal. R_3 is therefore included to aid in achieving a resistive balance. Generally speaking, R_3 will be in the C_2 leg when measuring a mica condenser and in the C_1 leg for an air condenser. The bridge is brought into balance by alternately varying the standard capacity C_1 and equalizing the power factors by R_3 until zero indication is obtained.

The bridge can be made direct-reading in $\mu\mu$fd. by using a dial with 100 divisions and a 10-division vernier installed so that 0 on the dial corresponds to maximum capacity on C_1. Then as the capacity of C_1 is decreased to compensate for the addition of C_x, ΔC is numerically equal to the dial reading times 10. The true capacity of C_1 will depart from linearity with the dial setting as it nears zero, but the percentage error remains small up to at least 90 on the dial ($C_x < 900$ $\mu\mu$fd). The overall accuracy can be made better than 1%.

L, C and Q measurements at r.f. — The low-frequency a.c. bridge method of measuring inductance is of value only with high-inductance coils for use at power and audio frequencies. I.f. and r.f. coils must be measured at frequencies corresponding to those at which they are used.

The method commonly employed is to determine the frequency at which the coil resonates when connected across a capacity of known value. This may be done (a) by connecting the coil-condenser combination in a negative-resistance oscillator (§ 3-7) and observing the resulting oscillation frequency on a calibrated receiver, or (b) by connecting the coil to a calibrated condenser, supplying the circuit with r.f. power from a suitable oscillator, and tuning the condenser until resonance is indicated by maximum indication on a vacuum-tube voltmeter (Fig. 1223). With the capacity known in $\mu\mu$fd. and the resonant frequency in kc., the apparent inductance of the coil in microhenries can be computed

$$L = \left(\frac{159,160}{f_2}\right)^2 \frac{1}{C}$$

The apparent inductance thus computed is in error, however, in that it also includes the distributed capacity of the coil. This will be discovered if a similar measurement is made at another frequency (for example, the harmonic of f_1), for it will be found that a different value of inductance results. However, by combining the two measurements the true inductance can be found

$$L = \frac{1}{8422 f_2} \times \frac{1}{C_1 - C_2}$$

when f_2 is the second harmonic of f_1, C_1 is the capacity required to tune to f_1 and C_2 to f_2.

A convenient source of r.f. power for the two-frequency method of inductance measurement is the transmitter exciter unit, provided it has good second harmonic output. The oscillator output and link circuit (shown inside dashed lines in Fig. 1223) should either be shielded or

sufficiently remote from the measuring circuit so that the vacuum-tube voltmeter shows no indication when there is no coil in the circuit. The calibrated condenser must, of course, have sufficient capacity to tune over a 2-to-1 frequency range. It can be calibrated by a bridge such as the substitution-type capacity bridge of Fig. 1222.

The resonance method can also be used for accurate measurement of capacity. A standard coil of suitable inductance must be provided; the exact value is not important. The standard condenser C_1 is first tuned to resonance with the oscillator frequency. The unknown capacity, C_x, is then added in parallel and the capacity of C_1 reduced until the circuit again resonates at the oscillator frequency. The difference between the two settings (ΔC) represents the capacity of C_x.

The arrangement of Fig. 1223 is additionally useful in that it can be used as a Q meter, measuring r.f. resistance and impedance (§ 2-10).

As is shown by Fig. 1224, resistance in a tuned circuit broadens the resonance curve. By measuring the frequency difference between the two points at which the output voltage equals 70.7% of the peak voltage (where the resistance in the circuit equals its reactance), the Q of coils and condensers can be computed.

There are two methods of determining these points. One involves the use of a calibrated variable frequency oscillator to determine the bandwidth in terms of frequency change and the other a calibrated variable condenser to measure the capacity change.

When the calibrated variable oscillator and v.t.v.m. are used, the frequency and r.f. voltage at resonance are first noted. The oscillator frequency is then varied on either side of resonance until the v.t.v.m. reads 70.7% of its initial value. Then Q is equal to the frequency divided by the bandwidth, or

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

where Δf is the difference between the frequencies f_1 and f_2.

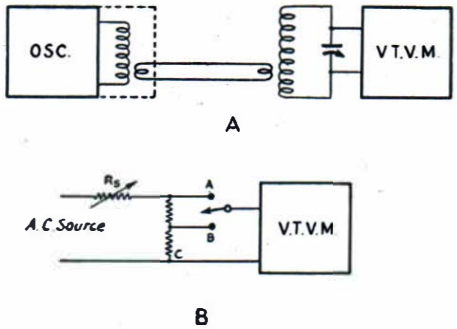


Fig. 1223 — (A) Circuit used in measuring inductance, capacity and Q at r.f. The calibrated variable-frequency oscillator should have a tuning range in excess of 2-to-1. (B) Circuit for calibrating the v.t.v.m. for Q measurements from 60-cycle a.c. R_{AC} is 70.7% of R_{AC}. With the switch in position A, R_S is adjusted to give a voltmeter deflection near the upper part of its scale; this is the peak-deflection reference point. The switch is then turned to position B, and the new reading noted. By making a number of measurements with different initial input levels, a graph can be plotted showing peak and 70.7% readings for a wide range of inputs.

When the frequency of the oscillator is fixed and a calibrated variable condenser is used, the capacity at resonance (C_r) is noted, as well as that on either side at which the voltmeter reads 70.7% of maximum.
Then

$$Q = \frac{2\,C_r}{C_2 - C_1}$$

The foregoing applies when measuring the Q of coils. Actually, the figure of Q thus derived is not that for the coil alone but for the tuned circuit as a whole, including the condenser. The Q of the standard condenser must therefore be kept high. An efficient air condenser with steatite or mycalyx insulation is required; it should be operated near maximum capacity. Short, heavy leads must be used and the stray capacities kept as low as possible.

The Q of other air condensers and of mica condensers can be determined by first measuring the Q of the circuit with a standard coil in place, then connecting C_x in parallel with C and

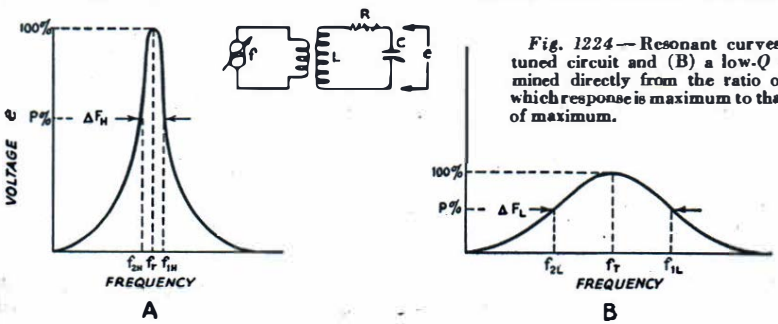


Fig. 1224 — Resonant curves of (A) a high-Q tuned circuit and (B) a low-Q circuit. Q is determined directly from the ratio of the frequency at which response is maximum to that at which it is 70% of maximum.

again measuring the Q. The Q of the unknown condenser is

$$Q_x = \frac{(C_1 - C_2)\, Q_1\, Q_2}{C_1\,(Q_1 - Q_2)}$$

Low-Q mica and paper condensers ($Q <$ 1000) can be measured by inserting the unknown in series with L and C. Q_1 is measured with a shorting bar across the unknown; the bar is then removed and Q_2 determined. Then

$$Q_x = \frac{(C_2 - C_1)\, Q_1\, Q_2}{C_1\, Q_1 - C_2\, Q_2}$$

If C_2 is larger than C_1, the reactance is inductive rather than capacitive; i.e., the "condenser" is actually an inductance at the measurement frequency.

The r.f. resistance, reactance and impedance of other components can be measured by the same methods. If an external r.f. impedance (such as an antenna or transmission line or an r.f. choke) is inserted in a coil-condenser circuit, it will both detune the circuit and broaden its resonance curve. By observing the capacity required to bring the circuit back to resonance and measuring the additional resistance introduced by re-measuring its Q, the reactive and resistive components of the external impedance can be computed.

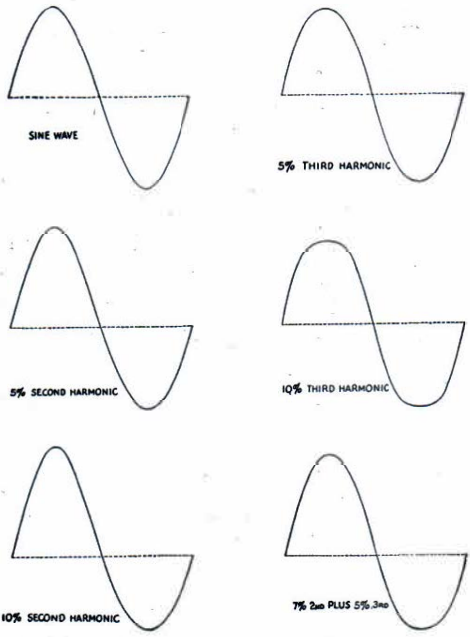


Fig. 1225 — Waveforms as viewed on the cathode-ray oscilloscope, showing the distinguishing characteristics of various types of harmonic distortion. The wave shape is symmetrical with odd-harmonic distortion, asymmetrical with even.

Using a standard coil and condenser suitable for the operating frequency, connect the unknown quantity across C_1 (for high resistances) or in series with L and C (for low resistances), and proceed as previously outlined. If C_1 must be increased to restore resonance, the reactance of the unknown is inductive; if it must be decreased, the reactance is capacitive.

•12-5 WAVEFORM

Wave Analyzers — In working with alternating currents it is often necessary to know not only the peak or average values of voltage and current, but also something of the waveform (§ 2-7). By the use of a wave analyzer it is possible to measure the amplitude and frequency of the fundamental and each of the important harmonic components of a complex electrical waveform.

The most widely-used instrument for waveform analysis is the cathode-ray oscilloscope equipped with a linear sweep circuit. As shown in § 3-9, the pattern traced on the screen represents the actual electrical shape of the wave under study. By study of this shape the character of the wave can be determined. Fig. 1225 shows typical patterns shapes resulting from different kinds of harmonic content.

The oscilloscope does not readily enable analysis of separate small harmonic components and in laboratory work special wave analyzers are used for this purpose. In the heterodyne type of wave analyzer the voltage under measurement is heterodyned with the signal from a calibrated variable oscillator and the resultant applied to a square-law detector or balanced modulator which delivers its output to a highly-selective amplifier using piezoelectric crystal filters or sharply-tuned circuits. The frequency of the component under measurement is indicated by the variable oscillator setting, while the amplitude is read on a vacuum-tube voltmeter. Since it first magnifies the relative separation between the harmonic components by heterodyning, and then uses a highly-selective amplifier with a pass-band of but a few cycles, the individual components of a complex voltage can be measured even when the fundamental lies in the lowest audio range.

Simpler wave analyzers of the resonant type are satisfactory on higher frequencies. R.f. harmonics may be measured qualitatively with an ordinary tuned circuit equipped with a sensitive resonance indicator. An absorption frequency meter or field-intensity meter may be used for the purpose (§ 12-2, § 12-8). By tuning the instrument to the fundamental and each harmonic in turn, keeping the pick-up constant, a useful indication of relative amplitudes can be obtained.

Modulation Percentage — Determination of the percentage of modulation in a modulated

wave (§ 5-2) requires analysis of the wave envelope. Use of the oscilloscope for this purpose is described in § 5-10. Another form of modulation percentage indicator is shown in Fig. 1230-B. Essentially, it constitutes a peak diode voltmeter which reads the r.f. carrier steady-state level when the plug is in the left-hand jack, while in the right-hand position the meter reads the average rectified component of the audio modulation. The relative amplitudes of the reading indicate the modulation percentage.

• 12-6 RECEIVER CHARACTERISTICS

Measurements in connection with receiving equipment come under two heads: (1) overall performance, and (2) servicing and alignment.

Performance Measurement — Laboratory measurements of overall performance require (1) a standard-signal generator, which is a test oscillator covering the frequency range in use, capable of being modulated 30% at 400 cycles and equipped with an output attenuator accurately calibrated in microvolts; (2) a standard dummy antenna approximating an ordinary receiving antenna, consisting of a 20 μh. inductance, shunted by 400 $\mu\mu$fd. and 400 ohms, in series with a 200 $\mu\mu$fd. condenser, used to couple the signal generator to the receiver antenna input; (3) an output-power measuring device consisting of an a.f. voltmeter connected across a resistance representing the output load of the receiver; and (4) a beat-frequency audio oscillator continuously variable from 30 to 15,000 cycles. A wave analyzer is also desirable for measuring audio distortion.

With this equipment the receiver characteristics listed in § 7-2 can be measured and charted. Sensitivity is stated in terms of the input in microvolts of a 30%-modulated signal required to give an audio output of 0.5 watt (0.05 watt for battery receivers, etc.) at various frequencies throughout the range. Selectivity curves (Fig. 701) are taken by varying the signal-generator frequency either side of resonance in increments of 1 to 10 kc., increasing the signal input each time to keep the output constant. Fidelity curves are made by modulating the signal generator at 30% with a continuously-variable beat-frequency oscillator and recording the audio output of the receiver relative to the normal response at 400 cycles. The signal-to-noise ratio is determined by measuring the noise output at maximum gain with no signal, and then applying a signal sufficient to double the output. Frequency stability is measured by methods described in § 12-2, zero-beating the receiver oscillator against a stable standard and observing the drift (a) during the warm-up period of the receiver, (b) with variation in power supply voltage, and (c) with various input signal levels.

Testing and Alignment — The measurement of receiver performance requires precision laboratory equipment beyond the scope of the average amateur. Sufficient apparatus for servicing receivers should be available in every amateur station, however. This may be as little as a multi-range volt-ohm-milliammeter, a test-signal source of some description, and a few spare tubes.

For the alignment of tuned circuits a simple test oscillator is required, preferably one that can be modulated by a 400-cycle audio oscillator. A rectifier-type voltmeter can be used for the output meter.

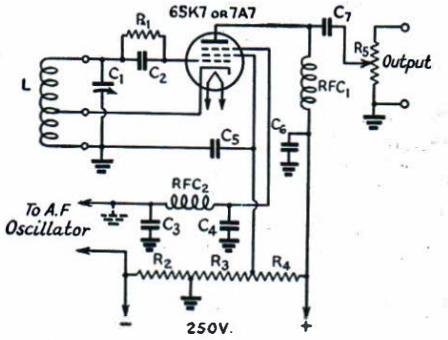


Fig. 1226 — I.f. test signal generator circuit.

C_1 — 100-$\mu\mu$fd. variable with 200-$\mu\mu$fd. fixed in parallel.
C_2 — 100-$\mu\mu$fd. midget mica.
C_3, C_4 — 250-$\mu\mu$fd. midget mica.
C_5 — 0.005-μfd. mica.
C_6 — 0.1-μfd., 400-volt paper.
C_7 — 500-$\mu\mu$fd. midget mica.
R_1 — 50,000 ohms, ½-watt.
R_2 — 2000 ohms, ½-watt.
R_3 — 20,000 ohms, 1-watt.
R_4 — 20,000 ohms, 2-watt.
R_5 — 500-ohm carbon potentiometer.
L — 440–510 kc.: 140 turns No. 30 enamel close-wound on 1½-inch plug-in form. Cathode tap 35 turns from ground end.
1400–1550 kc.: 42 t. No. 20 d.s.c., tap at 10 t.
4500–5500 kc.: 11 t. No. 18 spaced, tap at 3 t.
RFC_1 — 2.5-mh. r.f. choke.
RFC_2 — 25-mh. r.f. choke.

The frequency meter is a suitable signal source for r.f. alignment provided the harmonic amplitude on the higher-frequency bands is great enough. A harmonic amplifier and output attenuator are useful in this application.

The i.f. test oscillator circuit shown in Fig. 1226 consists of a simple e.c.o. with plug-in coils. The output level is controlled by a potentiometer so connected as to present a constant input resistance to the receiver. The oscillator should be shielded so that direct pick-up is minimized. It is helpful to make all ground returns to a heavy copper strap which is connected to the cabinet at only one point — the output ground terminal. The plug-in-coil should be separately shielded.

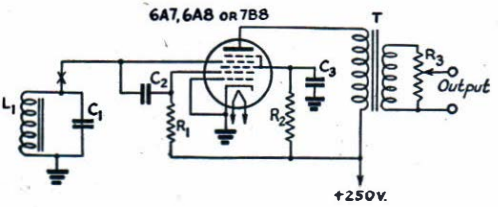


Fig. 1227 — Negative-resistance audio-frequency oscillator.
C_1 — 0.15-μfd., 400-volt paper.
C_2, C_3 — 0.1-μfd., 400-volt paper.
C_3 — 0.25-μfd., 200-volt paper.
R_1, R_2 — 50,000 ohms, 1-watt.
R_3 — 50,000-ohm volume control.
L_1 — 1.2-henry choke (Thordarson T-14C61 with iron core removed).
T — Output transformer (interstage audio, 1:3 ratio).

The test oscillator may be suppressor-grid modulated by applying approximately 10 volts of audio (for 50% modulation), as shown in the diagram. The suppressor-grid is biased 10-volts negative for modulated use; if an unmodulated signal is desired, the upper terminal can be grounded as indicated. This will increase the output from the oscillator. Conversely, if the output potentiometer does not attenuate the signal sufficiently, additional d.c. negative bias can be applied between the modulation terminals.

Ordinarily there is no requirement for precise calibration of the test oscillator. In i.f. alignment most communications receivers are equipped with a crystal filter and the oscillator frequency is set to correspond with the crystal response. If the receiver contains no crystal filter, the oscillator should be set at the design i.f. as closely as its calibration will permit.

With an unmodulated test signal the output indicator can be the "S"-meter in the receiver, a microammeter in the detector or a.v.c. circuit, or a vacuum-tube voltmeter. It is not advisable to use the receiver beat oscillator to generate an audible note for output indications. When a modulated test signal is used, the output indicator can be a copper-oxide rectifier-type voltmeter which reads the a.f. voltage across the rated output load resistance. Power output can be computed as previously described.

The a.f. modulating source for the test oscillator can be any audio oscillator capable of delivering 10 to 20 volts at the standard receiver checking frequency of 400 cycles.

Audio Oscillators — A simple oscillator of the negative-resistance or two-terminal variety using a pentagrid tube is shown in Fig. 1227. A frequency of approximately 400 cycles is generated with the tuned circuit shown.

Fig. 1228 shows a beat-frequency type of audio oscillator with a continuously-variable range from 0 to 15,000 cycles, of the kind used in making fidelity measurements and for other purposes requiring a wide-range variable a.f. source. The electron-coupled oscillators are initially tuned to an identical frequency in the

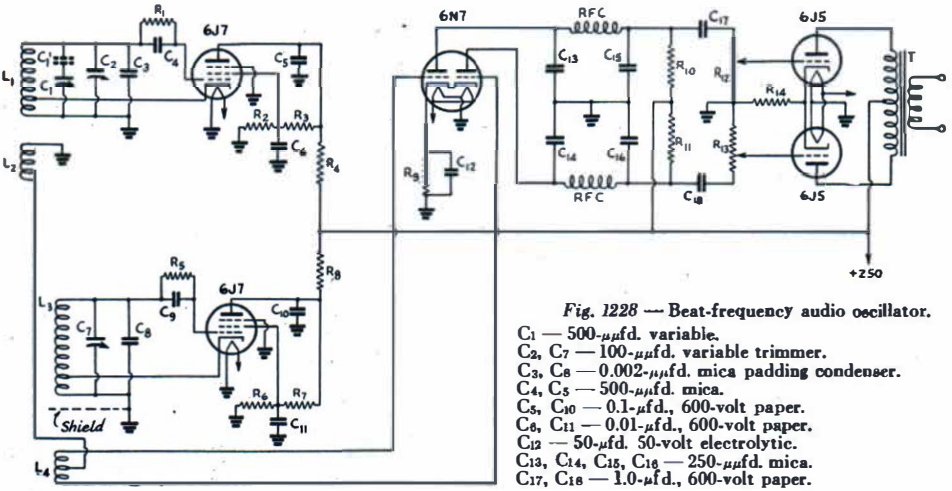


Fig. 1228 — Beat-frequency audio oscillator.
C_1 — 500-μμfd. variable.
C_2, C_7 — 100-μμfd. variable trimmer.
C_3, C_8 — 0.002-μμfd. mica padding condenser.
C_4, C_5 — 500-μμfd. mica.
C_5, C_{10} — 0.1-μfd., 600-volt paper.
C_6, C_{11} — 0.01-μfd., 600-volt paper.
C_{12} — 50-μfd. 50-volt electrolytic.
C_{13}, C_{14}, C_{15}, C_{16} — 250-μμfd. mica.
C_{17}, C_{18} — 1.0-μfd., 600-volt paper.

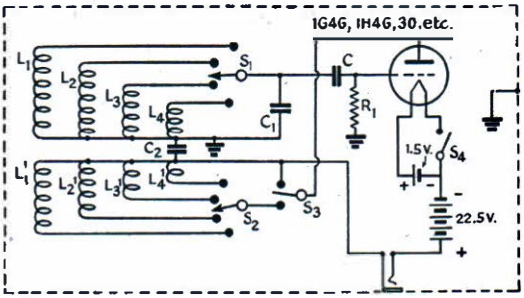
R_1, R_5, R_9 — 50,000 ohms, ½-watt.
R_2, R_3 — 20,000 ohms, 1-watt.
R_4 — 2000 ohms, ½-watt.
R_6, R_7, R_8 — 40,000 ohms, ½-watt.
R_{10}, R_{11} — 0.1 megohm, ½-watt.
R_{12}, R_{13} — 25,000-ohm potentiometer.
R_{14} — 500 ohms, ½-watt.
RFC — 125-mh. iron-core r.f. choke.
T — High-fidelity push-pull interstage or tube-to-line transformer, according to application.
L_1–L_3 — 1-mh. shielded coil with cathode tap (may be 456-kc. receiver beat-oscillator coil, with self-contained trimmer used as C_2).
L_2 — 100 turns No. 32 wire close-wound over L_1.
L_4 — 20 turns wound over L_3 (separated by electrostatic shield).

Fig. 1229 — C.w. and 'phone monitor.

C_1 — 50-$\mu\mu$fd. midget variable.
C_2 — 0.002-μfd. midget mica.
C_3 — 100-$\mu\mu$fd. midget mica.
R_1 — 1 megohm, ½-watt.
S_1–S_2 — 2-section 4-position rotary band switch.
S_3 — S.p.d.t. low-capacity switch.
S_4 — Toggle switch.

Band	*L*	*L'*
1750	90 t. No. 30 e.	24 t.
3500	50 t. No. 30 d.c.c.	16 t.
7000	30 t. No. 22 d.c.c.	10 t.
14000	10 t. No. 22 d.c.c.	6 t.

All coils close wound on 1-inch diameter forms, grouped around S_1–S_2 with adjacent coils at right angles. L and L' approximately ¼-in. apart.



vicinity of 100 kc., the variable tuning condenser C_1 being set at zero. The oscillator outputs are coupled to the balanced modulator which passes the beat frequency along to the push-pull audio amplifier. In construction the two oscillator circuits should be made as nearly identical as possible, so that any tendency toward drift will be equal in both. The tuned circuits should be isolated from tube heat and made mechanically stable. An electrostatic shield (§ 4-9) on the fixed-oscillator output aids in reducing pulling at very low frequencies. A voltage-regulated power supply is desirable.

●12-7 TRANSMITTER CHARACTERISTICS

The transmitter characteristics requiring measurement are carrier frequency, d.c., a.c. and r.f. voltage, current and power, keying and modulation quality, modulation percentage distortion, carrier noise and spurious radiations. Instruments for the measurement of most of these quantities have been discussed under the appropriate headings.

Keying and modulation checks may be made by several methods; the two commonly used by amateurs are aural checks with monitors, and visual checks with the oscilloscope. For data on the use of the oscilloscope in checking modulation using "wave envelope" and "trapezoid" methods, see § 5-10.

Monitors — A monitor is a miniature receiver, usually having only a single tube, enclosed with its batteries in some sort of metal box which serves as a shield. The requirements for a satisfactory monitor for checking c.w. signals are not difficult to satisfy. It should oscillate steadily over the bands on which the station is to be active; the tuning should not be excessively critical, although the degree of band-spread ordinarily considered desirable for receivers is not essential; the shielding should be complete enough to permit the monitor to be placed near the transmitter and still give a good beat note when tuned to the fundamental frequency of the transmitter (this is often impossible with the receiver because the pick-up is too great); and it should be constructed solidly so that it can be moved around the station without necessity for retuning when listening to a signal.

The circuit of a simple monitor with band-switching covering four amateur bands is shown in Fig. 1229. Any 1.5- or 2-volt filament triode can be used, as well as any batteries of a size that will fit into the container selected. A plate-tickler switch (S_3) is provided to make the monitor non-oscillating when checking 'phone signals. If desired, a regeneration control could be incorporated (§ 7-4).

Any type of simple detector with a means for picking up a small amount of r.f. from the transmitter can be used as a 'phone monitor. A satisfactory monitor can be constructed using a diode rectifier and untuned pick-up coil, as shown in Fig. 1230-A. Headphones are used for listening checks. The monitor can also be employed as an over-modulation indicator by use of the 0–1 milliammeter. The pick-up coil is loosely coupled to the tank circuit of the final r.f. amplifier until the milliammeter reads approximately 0.9 ma. The speech amplifier is supplied with a 400-cycle sine-wave tone from an audio oscillator such as that shown in Fig. 1227, and the gain control turned up until the monitor meter starts to rise, indicating overmodulation.

The circuit at 1230-B indicates the percentage of modulation directly. The a.c. milliammeter is first plugged into the left-band jack and the pick-up coupling adjusted to give a full-scale meter reading on the unmodulated carrier. Then the meter is plugged into the right-hand jack and the transmitter modulated by a tone or speech signal. The modulation percentage will be 140 times the reading of the meter; e.g., for 100% modulation the meter will read approximately 0.7 ma. In measuring the percentage of modulation with speech the inertia of meter will cause it to undershoot on peaks; the maximum swing should therefore be limited to something less than 0.7 ma.

Power — The power input to a radio transmitter is rated as the total d.c. plate input ($E \times I$) to the final stage. The power output is rated as the actual a.c. power measured at the

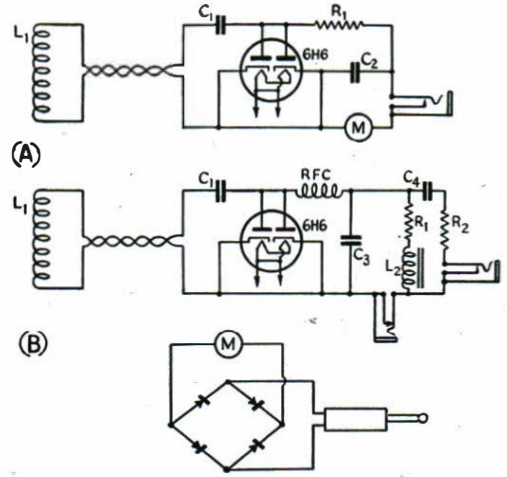


Fig. 1230 — (A) Simple diode 'phone monitor and overmodulation indicator.
C_1 — 0.005-μfd. midget mica.
C_2 — 0.01-μfd. paper.
R_1 — 0.15 megohm, ½-watt.
L_1 — Pick-up coil (enough turns of hook-up wire to give 1-ma. deflection on meter when loosely coupled to final tank, connected to unit through twisted-pair line).
M — 0–1 ma. d.c. milliammeter.

(B) Modulation percentage indicator.
C_1, R_1 and L_1 same as above.
C_3 — 0.005-μfd. midget mica.
C_4 — 1.0-μfd. paper.
R_2 — 0.25 megohm, ½-watt. (Should equal R_1 plus impedance of L_2 at modulation test frequency.)
L_2 — 30–50 henry iron-core choke.
M — 0-1 ma. a.c. milliammeter (d.c. meter with copper-oxide rectifier).

output terminals of the transmitter unit proper when connected to its normal load circuit or equivalent. A c.w. transmitter is rated in terms of its power capability under keying conditions, defined as intermittent operation with the key down 50% of the time. An amplitude-modulated transmitter is rated in terms of the average power under modulation when being modulated to its specified capability (the modulation capability and distortion at the test frequency being stated).

Instruments and methods of measuring both d.c. and r.f. power have been described in § 12-3. Additional methods used to measure transmitter power are the calorimeter and anode-dissipation methods. In the calorimeter method a non-inductive resistor carrying the r.f. power is cooled by water or other liquid surrounding it. The power dissipated is then calculated from the temperature rise of the water. The anode-dissipation method is used in connection with large water-cooled tubes. The total power delivered to the filament grid and plate circuits is first measured. The power dissipated in the cooling fluid is then measured in terms of its temperature rise and the difference between the total input power and that dissipated in the water is considered to be the useful power output delivered to the load.

• 12-8 ANTENNA MEASUREMENTS

Antenna measurements are made for the purposes (a) of securing maximum transfer of power to the antenna from the transmitter, and (b) of adjusting directional antennas to conform with design conditions. Measurements are therefore made of the current (power) in the antenna, voltage and current relationships, resistance, and radiated field intensity. Related to measurement of the antenna proper is the measurement of transmission line characteristics, chiefly involving impedance and resistance.

Elaborate field-strength measuring sets calibrated in microvolts-per-meter or similar units have been devised for use in commercial work, incorporating standard-signal generators and standard-gain amplifiers for measuring field intensity by the comparison method. Ordinary antenna checks and adjustments can be made by a simple field-intensity meter, however.

The instruments described for r.f. measurement (thermocouple ammeter, vacuum-tube voltmeter, L, C and Q meter) are all applicable to antenna measurement.

Antenna resistance (§ 10-1) can be determined by inserting a non-inductive resistor of known value and a thermoammeter in series in the antenna, preferably at a current loop. Then the antenna resistance in ohms is

$$R_a = \frac{I_k}{I - I_k} R_k$$

where I_k = antenna current with the resistance R_k in the circuit and I = the current with the resistance short-circuited.

With the resistance and current known, the antenna power can be calculated:

$$P_a = I^2 R_a$$

The effective antenna reactance (as well as the resistance) can be measured by the substitution method outlined under § 12-4. The measurement must be made at the operating frequency or band of frequencies.

Antenna impedance can also be measured by feeding power to the antenna from a source of known impedance through a uniform low-loss quarter-wave line (§ 10-5). The current or voltage is measured at both ends, whereupon

$$Z = \frac{Z_i I_i}{I_o} = \frac{Z_i E_o}{E_i}$$

where the subscripts i and o indicate the input (oscillator) and output (antenna) ends of the line.

Field-intensity meters — In adjusting antenna systems for maximum radiation and in exploring radiation patterns use is made of a field intensity meter. Basically a field-intensity meter is a vacuum-tube voltmeter provided with a tuned input circuit. It is used to indicate the relative intensity of the radiation field under actual radiating conditions. It is particularly useful on the ultra-high frequencies and in adjusting directional antennas. Checks with a field-intensity meter should be made at points not less than several wavelengths distant from the antenna and at heights corresponding with the desired angle of radiation. Standard comparative tests are made at a distance of one mile with one watt input to the antenna.

The absorption frequency meter shown in Figs. 1202-03 may be used as field strength meter if it is provided with a pick-up antenna. This can be a short length of brass rod or an automotive-type antenna mounted on a stand-off insulator and connected to the stator of the tuning condenser through a small trimmer. The crystal detector is not linear, so that a given increase in current does not indicate a directly proportional increase in field strength.

A more sensitive field-intensity meter of use in examining the field-strength patterns of lower-frequency antenna systems is shown in Fig. 1231. It consists of a diode rectifier and d.c. amplifier in the same envelope. The initial plate current reading is in the neighborhood of 1.4 ma.; with signal input, the current dips downward. The scale reading is linear with signal voltage, a characteristic that is advantageous in making comparative measurements. Radiated power variations will be as the square of the field-voltage indication.

Power gain in antenna systems is usually expressed in terms of decibels. A field intensity meter that reads directly in db. is shown in Fig. 1232. It consists of a self-biased linear triode voltmeter followed by a variable-μ d.c.

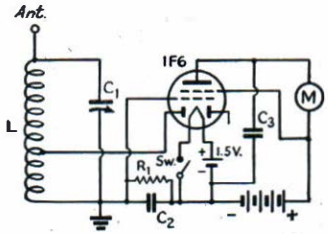


Fig. 1231 — Sensitive field-intensity meter.
C_1 — 50-$\mu\mu$fd. midget variable.
C_2 — 250-$\mu\mu$fd. midget mica.
C_3 — 0.002-μfd. midget mica.
R_1 — 1 megohm, ½-watt.
L — Coil to tune to frequency in use in conjunction with C_1. Diode tap in center of coil.
M — 0–1.5 milliamperes.

amplifier tube. Because of the logarithmic grid voltage-plate current characteristic of this tube, a 0-1 ma. meter in its plate circuit can be calibrated arbitrarily with an approximate linear db. scale as shown.

The scale covers approximately 25 db. and is linear over a range of about 20 db. At very small signals it departs from linearity, and 0 db. is therefore placed at 90% of the scale. A variable meter shunt compensates for variations in tubes and battery voltages. In use this resistor is initially adjusted to give a full-scale reading of 1 ma. The signal pick-up is then made such as to cause the meter to indicate 0 db. Alternatively, the initial reading can be set arbitrarily at 10 db.; adjustments will then be indicated as losses or gains in relation to that figure.

●12-9 TUBE CHARACTERISTICS

Accurate measurement of the operating characteristics of vacuum tubes (§ 3-2) under dynamic conditions requires the use of a vacuum-tube bridge, in which the tube coefficients constitute the "unknown" arm. Such a bridge is a complex and costly affair.

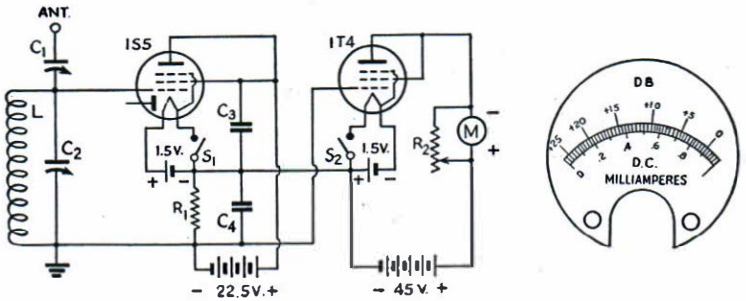


Fig. 1232 — Logarithmic field-intensity meter with db. calibration using miniature dry-cell type tubes.
C_1 — 3-30-$\mu\mu$fd. mica trimmer.
C_2 — 50-$\mu\mu$fd. midget variable.
C_3, C_4 — 500-$\mu\mu$fd. midget mica.
R_1 — 10 megohms, ½-watt.
R_2 — 1000-ohm wire-wound variable.
L — Coil to tune to frequency in use.
S_1, S_2 — Toggle switches or d.p.s.t. switch.
M — 0-1 ma. d.c. milliammeter.

Fortunately, however, the need for measuring vacuum-tube characteristics as such is rarely encountered in ordinary radio work, since complete data concerning standard types is published by the manufacturers.

The principal need for checking a tube lies in determining when, through aging or other defect, it departs from the original characteristics. The best method of checking a receiving or transmitting tube is by direct comparison in its own socket with a new tube of known quality under actual operating conditions. Any other test falls short of an actual performance test.

A number of commercial tube checkers of the type used by servicemen are on the market. In purchasing one the following qualifications should be sought: (1) complete facilities for checking shorts between any pair of electrodes; (2) a transconductance rather than an "emission" test (the emission of a tube may vary widely with no effect on its performance, while genuinely faulty tubes may show rated emission); (3) provision for checking plate and screen currents under typical conditions (at rated voltages); (4) gas and noise tests.

Commercial tube-checkers are elaborate assemblies bearing all standard socket types and switching arrangements to supply correct element potentials for all types of tubes. In the absence of such a unit, for an occasional need the amateur can assemble a circuit using an existing power source in accordance with Fig. 1233 to make a reasonably accurate standard transconductance test. A pentode tube is shown; for other types omit or add electrode connections as required. The voltages applied should correspond with those listed in the published tube tables within 5% (expecially grid voltage, plate voltage for triodes and screen voltage for pentodes). With the

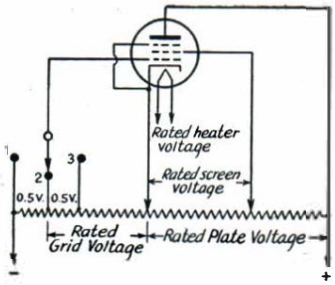


Fig. 1233 — Circuit for measuring vacuum-tube transconductance.

grid switch in No. 2 position, the plate and screen currents should read near the rated values; wide variations from normal indicate a defective tube.

To make the transconductance test, note the plate current with the grid switch alternately on positions 3 and 1, changing the bias from exactly 0.5 volt less than rated bias to exactly 0.5 volt more. The resulting plate current change multiplied by 1000 equals the transconductance in micromhos. This value can be checked against the tables. Tubes will usually operate satisfactorily until the transconductance falls to 70% of the rated value.

Pentagrid and heptode frequency converters may be checked by this method if the rated d.c. electrode voltages are applied. The oscillator section can be checked separately by noting the oscillator-anode current change.

Diodes can be checked by applying 50 volts of 60-cycle a.c. between plate and cathode, in series with a 0.25-megohm load shunted by a 2-μfd. condenser, and reading the rectified current on a 0-1 ma. d.c. meter. A reading of 0.2 and 0.25 ma. indicates a satisfactory tube.

Workshop Practice

• TOOLS

WHILE THE GREATER the variety of tools available, the easier and, perhaps, the better the job may be done, with a little thought and care it is possible to turn out a fine piece of equipment with comparatively few common hand tools. A list of tools which will be found indispensable in the construction of equipment will be found on this page. With these tools it should be possible to perform any of the required operations in preparing panels and metal chassis for assembly and wiring. A few additional tools will make certain operations easier.

The following list will be found helpful in making a selection:

Bench vise, 4-in. jaws
Tin shears, 10-in., for cutting thin sheet metal
Taper reamer, ½-in., for enlarging small holes
Taper reamer, 1-in., for enlarging holes
Countersink for brace
Carpenter's plane, 8- to 12-in., for woodworking
Carpenter's saw, cross-cut
Motor-driven emery wheel for grinding
Long-shank screwdriver with screw-holding clip for tight places
Set of socket wrenches for hex nuts
Wood chisel, ½-in.
Cold chisel, ½-in.
Wing dividers, 8-in. for scribing circles
Set machine-screw taps and dies
Folding rule, 6-ft.
Dusting brush

Several of the pieces of light woodworking machinery, often sold in hardware stores and mail-order retail stores, are ideal for amateur radio work, especially the drill press, grinding head, band and circular saws and joiner.

• CARE OF TOOLS

A few minutes with the oil stone or emery wheel now and then will maintain the fine cutting edges of knives, drills, chisels, etc.

Drills should be sharpened at frequent intervals so that grinding is kept at a minimum each time. This makes it easier to maintain the rather critical surface angles for best cutting with least wear.

INDISPENSABLE TOOLS

Long-nose pliers, 6-in.
Diagonal cutting pliers, 6-in.
Screwdriver, 6- to 7-in., ¼-in. blade
Screwdriver, 4- to 5-in., ⅛-in. blade
Scratch awl or ice pick for marking lines
Combination square, 12-in. for laying out work
Hand drill, ¼-in. chuck or larger, 2-speed type preferable
Electric soldering iron, 100 watts
Hacksaw, 12-in. blades
Center punch for marking hole centers
Hammer, ball peen, 1-lb. head
Heavy knife
Yardstick or other straight edge
Carpenter's brace with adjustable hole cutter or socket-hole punches (see text)
Pair small "C" clamps for holding work
Large coarse flat file
Large round or rat-tail file, ½-in. diameter
Three or four small and medium files, flat, round, half-round, triangular
Drills, particularly ¼-in., and Nos. 18, 28, 33, 42 and 50
Combination oil stone for sharpening tools
Solder and soldering paste (non-corroding)
Medium-weight machine oil

The soldering iron may be kept in good condition by keeping the tip well tinned with solder and not allowing it to run at full voltage for long periods when it is not being used. After each period of use, the tip should be removed and cleaned of any scale, which may have accumulated. An oxidized tip may be cleaned by dipping in sal ammoniac while hot and wiping clean with a rag. Should the tip become pitted, it should be filed until smooth and then tinned by dipping it in solder.

All tools should be wiped occasionally with an oily cloth to prevent rust.

• USEFUL MATERIALS

Small stocks of various miscellaneous materials will be required from time to time. Most of them may be purchased from hardware or radio-supply stores. A representative list follows:

½-in. by 1/16-in. brass strip for brackets, etc. (half-hard for bending)
¼-in. square brass rod or ½-in. by ½-in. by 1/16-in. angle brass for corner joints
¼-in.-diam. round brass rod for shaft extensions

Machine screws: Round-head, flat-head with nuts to fit. Most useful sizes, 4–36, 6–32, and 8–32 in lengths from ¼-in. to 1½-in. (Nickeled iron will be found satisfactory except in strong r.f. fields where brass should be used.)

Plain washers and lock washers for screws

Bakelite and hard rubber scraps

Soldering lugs, panel bearings, rubber grommets, lug terminal strips, cambric tubing

Machine screws, nuts, washers, soldering lugs, etc., are most reasonably purchased in quantities of a gross.

• CHASSIS CONSTRUCTION

With a few essential tools and proper procedure, it will be found that building radio gear on a metal chassis is no more of a chore than building with wood and a more satisfactory job results.

Much trouble and energy can be saved by spending plenty of time on the planning end of the job. When all details have been worked out, the actual construction will be greatly simplified.

Cover the top of the chassis with a piece of wrapping paper, or preferably cross-section paper, folding the edges down over the sides of the chassis and fastening with adhesive tape. Next, assemble parts to be mounted on top of the chassis and move them about until a satisfactory arrangement has been found, keeping in mind any parts which are to be mounted underneath so that interferences in mountings will be avoided. Place condensers and other parts with shafts extending to the panel first and arrange so that the controls will form the desired pattern on the panel. Be sure to line up the shafts square with the chassis front. Locate any partition shields and panel brackets next and then sockets with their shields, if used, and other parts, marking accurately the mounting-hole centers of each on the paper. Watch out for condensers whose shafts do not line up with the mounting holes. Do not forget to mark the centers of socket holes and holes for leads under i.f. transformers, etc., as well as holes for wiring leads.

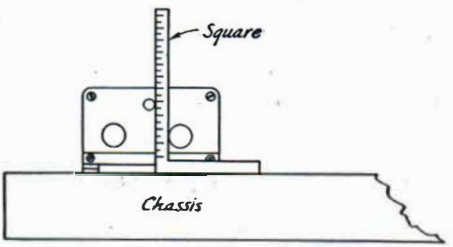


Fig. 1301 — Method of measuring heights of shafts. If the square is adjustable, the end of the scale should be set flush with the face of the bead.

By means of the square, lines indicating accurately the centers of shafts should be extended to the front of the chassis and marked on the panel at the chassis line by fastening the panel temporarily. The hole centers may now be punched in the chassis with the center-punch. After drilling, the parts which require mounting underneath may be located and the mounting holes drilled, making sure by trial that no interferences exist with parts mounted on top. Mounting holes along the front edge of the chassis should be transferred to the panel by once again fastening the panel to the chassis and marking from the rear.

Next mount on the chassis the condensers and any other parts with shafts extending to the panel, and measure accurately the height of the center of each shaft above the chassis as illustrated in Fig. 1301. The horizontal displacement of shafts having already been marked on the chassis line on the panel, the vertical displacement may now be measured from this line and the shaft centers marked on the back of the panel and the holes drilled. Holes for any other panel equipment coming above the chassis line may now be marked and drilled and the remainder of the apparatus mounted.

• DRILLING AND CUTTING HOLES

In drilling holes in metal with the hand drill, it is important that the centers be well located with the center punch so that the drill point will not "walk" away from the center when starting the hole. The drill should be held at right angles to the surface being drilled. Care should be used to prevent too much pressure with small drills which bend or break easily. When the drill starts to break through, special care should be used and it is often an advantage to shift a two-speed drill to low gear at this point. Holes near ¼-in. in diameter may be started with a smaller drill and reamed out with a larger drill.

The chuck of the usual type of hand drill is limited to ¼-in. drills. Although it is rather tedious, the ¼-in. hole may be filed out to larger diameters with round files. Another possible method with limited tools is to drill a series of small holes with the hand drill along the inside of the diameter of the large hole, placing the holes as close together as possible. The center may then be knocked out with a cold chisel and the edges smoothed up with a file. Taper reamers which fit in the carpenter's brace make the job much easier. A large rat-tail file may be clamped in the brace. This makes a very good reamer for holes up to the diameter of the file if the file is revolved counter-clockwise.

NUMBERED DRILL SIZES

Number	Diameter (mils)	Will Clear Screw	Drilled for Tapping Iron, Steel or Brass*
1	228.0	—	—
2	221.0	12–24	—
3	213.0	—	14–24
4	209.0	12–20	—
5	205.0	—	—
6	204.0	—	—
7	201.0	—	—
8	199.0	—	—
9	196.0	—	—
10	193.5	10–32	—
11	191.0	10–24	—
12	189.0	—	—
13	185.0	—	—
14	182.0	—	—
15	180.0	—	—
16	177.0	—	12–24
17	173.0	—	—
18	**169.5**	**8–32**	—
19	166.0	—	12–20
20	161.0	—	—
21	159.0	—	10–32
22	157.0	—	—
23	154.0	—	—
24	152.0	—	—
25	149.5	—	10–24
26	147.0	—	—
27	144.0	—	—
28	**140.5**	**6–32**	—
29	**136.0**	—	**8–32**
30	128.5	—	—
31	120.0	—	—
32	116.0	—	—
33	**113.0**	**4–36 4–40**	—
34	111.0	—	—
35	**110.0**	—	**6–32**
36	106.5	—	—
37	104.0	—	—
38	101.5	—	—
39	099.5	3–48	—
40	098.0	—	—
41	096.0	—	—
42	**093.5**	—	**4–36 4–40**
43	089.0	2–56	—
44	086.0	—	—
45	082.0	—	3–48
46	081.0	—	—
47	078.5	—	—
48	076.0	—	—
49	073.0	—	2–56
50	**070.0**	—	—
51	067.0	—	—
52	063.5	—	—
53	059.5	—	—
54	055.0	—	—

* Use one size larger drill for tapping bakelite and hard rubber.

For socket holes and other large round holes, an adjustable cutter designed for the purpose may be used in the brace. When the cutter is well sharpened, it makes the job easy. Occasional application of machine oil in the cutting groove usually helps. The cutter should first be tried out on a block of wood to make sure that it is set for the correct diameter. The best device of all for cutting socket holes is the socket-hole punch. The preferred type works by pressure applied by turning a screw with a wrench.

Square or rectangular holes may be cut out by using the series of small holes previously described, but more easily by drilling a ½-in. hole inside each corner, as illustrated in Fig. 1302, and using these holes for starting and turning the hacksaw. The socket-hole punches may also be of considerable assistance in cutting out large rectangular openings.

The burrs or rough edges, which usually remain after drilling or cutting holes, may be removed with a file, or sometimes more conveniently with a sharp knife or chisel. It is a good idea to keep an old wood chisel sharpened up for this purpose.

• CUTTING THREADS

Brass rod may be threaded or the damaged threads of a screw repaired by the use of *dies*. Holes of suitable size (see drill chart) may be threaded for screws by means of *taps*. Either are obtainable in any standard machine-screw

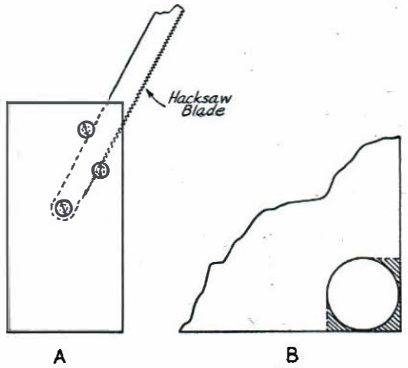


Fig. 1302 — Cutting rectangular holes in a chassis. If the corner holes are filed out as shown in the shaded portion of B, it will be possible to start the hacksaw blade along the cutting line. A shows a single-ended handle for a hacksaw blade.

size. A set usually consists of taps and dies for 4–36, 6–32, 8–32, 10–32 and 14–20 sizes with a suitable holder. The die may be started easily by filing a taper or bevel on the end of the rod. In tapping a hole, extreme care should be used to prevent breaking the tap. The tap should be kept at right angles to the surface of the material and rotation should be reversed a revolution or two whenever the tap starts to turn hard. With care, holes may be tapped rapidly by clamping the tap in the chuck of the hand drill and using slow speed. Machine oil applied to the tap usually makes cutting easier and sticking less troublesome.

• CUTTING AND BENDING SHEET METAL

If a sheet of metal is too large to be conveniently cut with a hacksaw, it may be marked with scratches as deep as possible along the line of the cut on both sides of the

sheet and then clamped in a vise and worked back and forth until the sheet breaks at the line. Do not carry the bending too far until the break begins to weaken, otherwise, the edge of the sheet may become bent. A pair of iron bars or pieces of heavy angle stock, as long or longer than the width of the sheet, used in the vise will make the job easier. "C" clamps may be used to keep the bars from spreading at the ends. The rough edges may be smoothed up with a file or by placing a large piece of emery cloth or sandpaper on a flat surface and running the edge of the metal back and forth over the sheet.

Bends are made similarly. The sheet should be scratched on both sides, but not too deeply.

• CLEANING AND FINISHING METAL

Parts made of aluminum may be cleaned up and given a satin finish, after all holes have been drilled, by placing them in a solution of lye for half to three-quarters of an hour. Three or four tablespoonfuls of lye should be used to each gallon of water. If more than one piece is treated in the same bath, each piece should be separated from the others so as to expose all surfaces to the solution. Overlapping of pieces may result in spots or stains.

• CRACKLE FINISH

Wood or metal parts may be given a crackle finish by applying one coat of clear Duco or Tri-Seal and allowing it to dry over night. A coat of Kem Art Metal Finish is then sprayed or put on thickly with a brush, taking care that the brush marks do not show. This should be allowed to dry for two or three hours and the part should then be baked in a household oven at 225 degrees for one and one-half hours. This will produce a regular commercial job. This finish comes in several different colors and should be obtainable through any dealers handling Sherwin-Williams paint products.

• HOOK-UP WIRE

A popular type of wire for receivers and low-power transmitters is that known as "push-back" wire which comes in sizes of No. 18 or 20 which is sufficiently large for all power circuits except filament. The insulating covering, which is sufficient for circuits where voltages do not exceed 300 or 400, may be pushed back a few inches at the end making cutting of the insulation unnecessary when making a connection. Filament wires should be of sufficiently large conductor to carry the current without appreciable voltage drop. Rubber-covered house-wire, sizes No. 14 to No. 10, is suitable for heavy-current transmitting tubes, while No. 18 to No. 14 flexible wire is satisfactory for receivers and low-drain transmitting tubes where the total length of wire is not excessive.

Stiff bare wire, sometimes called *bus-wire*, is most favored for the high r.f.-potential wiring of transmitters and, where practicable, in receivers. It comes in sizes No. 14 and No. 12 and is usually tin-dipped. Soft-drawn antenna wire may also be used. Kinks or bends may be removed by stretching 10 or 15 feet of the wire and then cutting into small usable lengths.

The insulation covering power wiring which will carry high transmitter voltages should be appropriate for the voltage involved. Wire with rubber and varnished cambric covering, similar to ignition cable, is usually available at radio dealers. Smaller sizes have sufficient insulation to be safe at 1000 to 1500 volts, while the more heavily insulated types should be used for voltages above 1500.

• WIRING TRANSMITTERS AND RECEIVERS

It is usually advisable to do the power-supply wiring first. The leads should be bunched together in cable form as much as possible and kept down close to the surface of the chassis. Chassis holes for wires should be lined with *rubber grommets* to fit the hole to prevent chafing of the insulation. In cases where power-supply leads have several branches, it is often convenient to use *fibre terminal strips* as anchorages. These strips also form handy mountings for wire-terminal resistors, etc. When any particular unit is provided with a nut or thumb-screw terminal, solder-lug wire terminals to fit are useful.

High-potential r.f. wiring should be well spaced from the chassis or other grounded metal surfaces and should run as directly as possible between the points to be connected without fancy bends. When wiring balanced or push-pull circuits, care should be taken to make the r.f. wiring on each side of the circuit as symmetrical as possible. When it is necessary to pass r.f. wiring through the chassis, a *feed-through* insulator of low-loss material should be used, or the hole in the chassis should be of sufficient size to provide plenty of air space around the wire. Large-diameter rubber grommets may be used to prevent accidental short-circuit to the chassis.

By-pass condensers should be connected directly to the point to be by-passed and grounded immediately at the nearest available mounting screw, making certain that the screw makes good electrical contact with the chassis. In using tubular paper by-pass condensers, care should be taken to connect the side marked "ground" or "outside foil" to ground.

Blocking and coupling condensers should be mounted well spaced from the chassis.

High-voltage wiring should be done in such a manner that exposed points are kept at a minimum and so that those which cannot be avoided are rendered as inaccessible as possible to accidental contact.

• SOLDERING

The secret of good soldering is in making certain that all members of the joint and the tip of the iron are clean and in allowing time for the *joint*, as well as the solder, to attain sufficient temperature. Sufficient heat should be applied so that the solder will melt when it comes in contact with the wire forming the joint without the necessity for touching the solder to the iron. Soldering paste, if the non-corroding type, is extremely useful when used correctly. In general, it should not be used for radio work except when it is necessary to make the soldered joint with one hand. In this case, the joint should first be warmed slightly and the soldering paste applied with a piece of wire. Only the soldering paste which melts from the warmth of the joint should be used. If the soldering iron is clean, it will be possible to pick up a drop of solder on the tip of the iron which can be applied to the joint with one hand, while the other is used to hold the connecting wires together. The use of excessive soldering paste causes the paste to spread over the surface of adjacent insulation causing leakage or breakdown of the insulation. Except where absolutely necessary, solder should never be depended upon for the mechanical strength of the joint; the wire should be wrapped around the terminals or clamped with soldering terminals.

• CONSTRUCTION NOTES

Lockwashers should be used under nuts to prevent loosening with use, particularly when mounting tube sockets or plug-in coil receptacles subject to frequent strain.

If a control shaft must be extended or insulated, a flexible shaft coupling with adequate insulation must be used. Satisfactory support for the shaft extension may be provided by means of a *metal* panel bearing made for the purpose. Never use panel bearings of the non-metal type unless the condenser shaft is grounded. *The metal bearing should be connected to the chassis with a wire or grounding strip.* This prevents any possible danger.

• COIL WINDING

Dimensions for coils for various inductances or frequencies may be determined from the data given in Chapters Four and Fifteen.

The number of turns specified should be spaced out to fill the specified length on the form. The length specified should be marked on the form and holes drilled opposite the pines to which the ends of the winding are to connect. Scrape one end of the wire and pass it through the lower hole in the form to the pin to which the bottom end of the winding is to connect and solder this end fast. Unroll an amount of wire approximately sufficient to make the winding and clamp the spool in a vise so it will not turn. The wire should be pulled out straight and the winding started by turning the form in the hands and walking up toward the vise. A fair tension should be kept on the wire at all times. The spacing can be judged by eye. If, as the winding progresses, it becomes evident that the spacing is going to be incorrect to fill the required length, the winding may be started over again with a different spacing. If the spacing is only slightly off, the winding may be finished, the top end fastened and the spacing corrected by pushing each turn. When complete, the turns should be fastened permanently in place with coil cement. After a little practice, the job of determining the correct spacing will not be difficult.

Sometimes it becomes necessary to adjust the number of turns on a coil experimentally to fit a particular job. The easiest way to do this is to bring a wire out from one of the pins, extending through the hole in the form for a half-inch or so. The end of the winding may then be soldered to this extension, rather than to the pin itself, and the nuisance of repeatedly fishing the wire through to the pin avoided until the correct size of the winding has been determined.

Learning the Code

You are about to learn a new language — the language of code. In learning you should consider it a language of sound only, for the sole form in which it is written is its English equivalent.

By far the best way to learn this new language is to secure the help of a competent operator skilled in code. If that is possible, you should never have to look at an alphabetic table of code equivalents; you should be taught by sound. Since everyone cannot get tutoring, however, it is necessary for us to teach this sound language as best we can by means of the printed page.

The Continental (International Morse) code is the type used in all radio work. It consists of various combinations of two different-length sounds properly interspersed with spaces, forming letters of the alphabet, numbers, etc. In nearly all radio work, this sound takes the form of a high-pitched audio beat or oscillation, usually ranging from 300 to 1200 cycles, depending upon the receiving operator's individual preference. Lacking an audio oscillator itself, the best way to simulate the high-pitched sound of code in headphones is to whistle. Since a whistle cannot be very well shown in print, terms have been coined which closely duplicate the actual letter-sounds in code. We speak of the short sound as "dit," of the longer one (three times as long, actually) as "dah."

For purposes of explanation to you and to whomever might assist you, let us say that the sound "dit" is pronounced as "it" with a "d" before it. The sound "dah" is pronounced with "ah" as in "father." The sound "dah" is always stressed or accented — not in a different tone of voice, but slightly drawn out and the least bit louder. The sound "dit" is pronounced as rapidly and sharply as possible; for purposes of easy combination, as a prefix, it is often shortened to "di." When combinations of the sounds appear as one letter say them smoothly but rapidly, remembering to make the sound "di" staccato, and allowing equal stress to fall on every dah. There should never be a space or hesitation between dits and dahs of the same letter.

These are simply convenient vocal terms to use in duplicating the sound of code. You should never think of a letter as being composed of sounds, but rather as a sound itself. You don't think of the spoken letter "u," for example, as being composed of two separate and distinct sounds — yet actually it is made up of the pure sounds "ee" and "oo," spoken in rapid succession. You learned the spoken letter "u" as a sound unit itself. Similarly, you should learn code letters as sounds themselves, and not as combinations of other sounds. A skilled operator does not think of dits and dahs when copying, but actually hears the headphones speak words to him. He has mastered this new language. The sound "dah didididit dit" is just as familiar to him as the sound of the spoken word "the," for which it stands.

• LEARN BY SOUND

Do not use the "didah" tables below for memorization, for if you do you will have a visual concept of code, instead of a sound concept; when you hear a sound you will have to convert it first into "didah" language, and then into letters. If you learn the sounds directly with their letters, however, without using visual reference, you will eliminate this extra step while learning.

So, while we must print a table of code equivalents, memorize them in sound. If possible, get someone who knows the code to start you off. A member of your family or a friend (preferably not one wishing to learn code) will suffice if he will study these paragraphs and practice the voice exercises a short time. Then ask him to pronounce the sounds to you, identifying them with the letters for which they stand. Take a few letters at a time, such as in the groups suggested below. As your "instructor" sings out the sounds to you, you should be able to call out the letters, or vice versa. This is excellent practice. It will come harder at first, possibly, than other ways, but you are set for a successful code career if you learn by sound. In other words, after you read this chapter do not go back and memorize the code tables. Don't even take another look. Get someone to call out the sounds to you, either by "didah" language or whistling, and preferably a combination of both.

Before taking up letter-sounds, let us observe some of the element sounds. Practice saying to yourself the sound "dididididi . . ." (dits in rapid succession). It should sound like a blast from a machine-gun: staccato, evenly-

spaced, precise. To make certain you get correct timing, start tapping the top of a table continuously in smooth, even sequence, like clockwork; if a metronome is available, it will serve admirably. The tapping (or metronome ticks) should be at about 100 per minute, or a bit less than two per second. Synchronize your tonguing of dits at four per metronome beat. Be careful not to say "didididi . . . didididi . . . ," thereby leaving a space after every fourth dit; they should be as even and regular as the metronome beat, but four times as fast. If it will help, slightly accent the dit which coincides with the metronome beat. When you can do this easily, begin repeating "dahdah-dahdahdah . . ." so that the beginning of each second dah synchronizes with a metronome tick (or table rap). Let the dahs run smoothly and make them of equal length. Do not let them become choppy; your voice should be almost continuous, broken only for that short instant your tongue cuts off the tone to make the "d" sound.

Now alternate, repeating dits for a moment and then switching to dahs, without stopping to take a breath. This will show you the proper proportion of dits to dahs at this speed, or any speed to which the metronome is set; that is, four dits or two dahs per metronome tick. You should now practice alternating this voice simulation with whistling. Again, make your whistled dits very short and staccato, your whistled dahs smooth and full.

Another excellent practice exercise is the alternating of single dits and dahs. Practice repeating (and then whistling) "didahdidah-didahdidah . . . ," and then "dahdidahdidah-didahdi . . . ." Here again the tone should be almost continuous, the dit as short as possible. An hour, in several shorter periods, is not too much time for these exercises. If you can master them, you will have no difficulty in forming the various letter and number sounds.

• LETTER SOUNDS

When you have mastered the timing and rhythm exercises explained above, you can begin to combine these groups and form letter sounds. In doing so remember that this is a new type of spoken language you are learning, not a collection of short and long code symbols taken from a printed page. In fact, it might be better to forget that you are learning "code" and think of it instead as the "didah" language.

Let us take a few letters and numbers to start with, as shown in the following table. Remember that you should look at this table only briefly while reading this text; when you come to study, have someone hold this book and coach you in memorizing the code-sound of each letter.

5	dididididit
0	dahdahdahdahdah
e	dit
t	dah
a	didah
r	didahdit

The "5" should merely be five staccato dits in the same sequence and speed you practiced above; the zero should similarly be five dahs. "A" is our first character using both sounds. Make that di very short, the dah the usual length, properly stressed; it should have the same metric swing as does the word "to-day," rapidly spoken and strongly accenting "day." Similarly, the "didahdit" of "r" should have the same metric swing as "repeated," the second syllable strongly stressed, the final one not accented and as short as you can make it.

Spend at least a half-hour on this group; and preferably more, though your practice should then be split up into shorter periods. Take your time and learn the sounds of these letters thoroughly. Repeat them in "didah" language, and then whistle the characters.

You should immediately begin practicing copying down the characters your "instructor" calls out to you in sound. Copy them simultaneously speaking the letter, if you like. For the first copying practice it is well to print; in the transmission of non-English text, which you often will be getting for letter-practice, longhand letters are sometimes confused with each other. For example, "l" is mistaken for "e," an uncrossed "t" for "l," etc. The U. S. Signal Corps requires printing ability by its field operators, since much of the text handled is in code groups and errors would cause much difficulty in reading the actual message. Print until you are able to receive about ten words per minute, above which speed you will want to resort to longhand to keep up with the text.

Learn to print the characters rapidly and without conscious effort; your mental effort must be on reading the code sounds and not on your finger movements. Never look back over your copy when receiving, nor try to guess what word is coming; copy what is sent. If you do not immediately recognize a character, skip it and devote your full attention to the next one; if you try to remember it, you doubtless will miss several letters in a row. By checking your written copy with the transmitted text, you can determine what letters

are giving you trouble and give special attention to them.

Here is another group of characters. Before beginning their study, have your "instructor" review you on the first group. Intersperse learned sounds with new ones when studying this next group:

s	didadit
l	didahdidit
u	dididah
q	dahdahdidah
j	didahdahdah

The first letter obviously is simply three staccato dits. "L" is more difficult; its metric swing is like that of "fraternity"; again, make dits rapidly, particularly the final ones, and stress the dah. "U" is similar to "s" with the final dit changed to a dah; practice the two letters interchangeably to get the rhythm. "Q" and "j" should be smooth; stress each dah equally, make the dit short.

Study these in the same manner as the previous group. You have nearly twice as many letters to remember now, so your progress will be a bit slower. Don't rush; learn each sound thoroughly before proceeding to the next. Speed will come later, and it will come rapidly if you learn by sound. There are quite a number of words which can be made from the letters so far studied, and examples are given for each successive lesson on a later page. Practice saying the sounds to yourself, particularly between study sessions. Occasionally you should have an operator who knows code check you on your progress.

Here is a third group, to be taken up only after you have learned the previous eleven characters well:

h	dididadit
o	dahdahdah
n	dahdit
c	dahdidahdit
v	dididadah

The first two letters should give no trouble. Be careful not to have a space in the sound for "n"; nor to make the dit anything but short. "C" can best be simulated by remembering our earlier exercise of alternating single dits and dahs; practice that again momentarily, and you should have no trouble.

When learning the sound of "v," practice "h" and then make the final dit a dah; practice the two letters intermittently as you did "u" and "s," repeating in your mind what letter is concerned with each sound.

Here is a fourth group:

i	didit
b	dahdididit
y	dahdidahdah
p	didahdahdit

By now you should have developed sufficient timing sense and code consciousness that you can pronounce new sounds without difficulty. Simply remember to keep characters smooth, without spaces.

In between practice periods, when convenient, notice street signs and posters containing these letters; when you see one you know, call out its code-sound. Keep this practice principle always in mind and use it a great deal; you will find a goodly amount of idle time you can put to good use — riding back and forth to work, or walking to the corner drugstore. Or even if you are not in the vicinity of signs, there are a number of short words you can practice saying to yourself in sound language. Make your character formation snappy, leave a recognizable space between letters, plenty of space between words. Use both whistle and voice technique. We want to get you thinking subconsciously of code whenever you see letters and words.

Now, another group:

w	didahdah
k	dahdidah
z	dahdahdidit
m	dahdah

When these have been thoroughly learned, you may proceed to the final group:

d	dahdidit
x	dahdididah
f	dididahdit
g	dahdahdit

You already have learned the two simplest numerals: five and zero. The others are:

1	didahdahdahdah
2	dididahdahdah
3	didididahdah
4	dididididah
6	dahdidididit
7	dahdahdididit
8	dahdahdahdidit
9	dahdahdahdahdit

From here, your progress will be principally a matter of practice. Listen in on the short-wave bands for commercial automatic-tape stations sending slowly. While the schedule varies, you will usually find suitable stations in the daytime between 7,000 and 20,000 kilocycles, and at night between 4,000 and 10,000 kilocycles. Do not be surprised if you seem to copy "solid" but your text does not make sense; much commercial transmission is in coded groups or foreign language. Tape sending is precision itself; the more you listen the more will become fixed in your mind the sound of code perfectly sent and the more easily will you be able to simulate it.

• LEARNING TO SEND

It is important that you should learn the correct sound of code letters thoroughly before ever touching a telegraph key. If you do not know how a code letter should sound, no amount of playing with a key will teach you. When you are at the point where you can unhesitatingly call out each letter as your "instructor" pronounces or whistles the sound, you are ready to learn how to handle a key. First, however, you need some device to produce a tone. A buzzer set will serve the purpose, but a much better tone source is the vacuum-tube oscillator since it duplicates the audio beat note a radio operator copies. Construction data on both types are given near the end of this chapter.

A telegraph key is simply an on-off switch in convenient form for rapid manipulation. Pressing the key knob closes the electrical contact and produces the "mark" or sound; releasing it allows the contact to break, producing the space or no-signal period. Correct key adjustment is that adjustment which fits your particular touch, and it is important that you arrive at the correct adjustment before attempting to use the key. A reliable rule is that there should be a vertical movement of about one-sixteenth inch at the key knob. This is measured from the top surface of the knob. It is set by the rear screw adjustment. When the knob's top surface goes down about one-sixteenth inch upon pressing the key, you have the approximate "average" spacing between the key contacts. In making any key adjustment, be sure to loosen the lock nuts first, so that you do not strip the threads. Tighten all lock nuts when you complete the adjustment. The contact points should be perfectly aligned by means of the side screws. There should be a very slight degree of side play, between the two side screws. These screws should be tightened, then loosened just a bit so that the key moves freely and does not bind. Recheck the contacts after this adjustment to be sure they are true, making any slight readjustment necessary.

Although the "one-sixteenth inch" rule is a good one to follow for first adjustment, the amount of vertical movement can later be changed to suit your particular fist. The spring tension, likewise, must be set for the individual operator. Some prefer a heavier spring than others. The primary consideration is to send *good code;* how you have your key adjusted to do this depends on what you find best for you. However, it should be remembered that too heavy a spring tends to make your sending "choppy," causing you to "clip" your dabs and dits as well as being tiring for long periods of sending. Similarly, too light a spring tends to cause you to run characters together, there being insufficient control of the key. Remember, *you* are making the characters, the key isn't! Generally speaking, a somewhat heavy spring allows better control, particularly of dots. With a spring of "feather-weight" tension, the dits are likely to run away from you and you will find yourself slurring them. No treatise on key adjustment ever can solve the individual's problems in this line. Only by personal use can you find the correct adjustment for yourself.

There is a definite sending posture which should be observed. Sit upright in your chair, square with the operating table, with your arm on a line with the key. The key knob should be about eighteen inches from the front edge of the operating table, allowing room for the elbow to rest on the table. The muscle of the forearm should support the weight, and the wrist should be off the table. A table about thirty inches in height is best. The key may be fastened by means of wood screws directly to the table if one is available for permanent use. Otherwise, it may be fastened to a rectangular piece of thin board such as three-ply veneer, about six inches wide and two feet long.

The manner of grasping a key knob is also the choice of the individual operator. When learning, place the thumb against the left edge of the key knob, the first finger on top of the knob at the rear and lapping over the rear edge just a bit; and the second finger against the right edge of the knob, about in the center or slightly to the rear of center. In no event should the grip be tense. The first and second

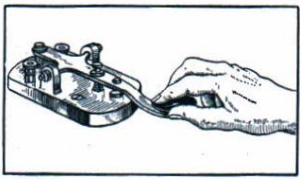


Fig. 1401 — Illustrating the correct position of the hand and fingers in using a telegraph key.

fingers should be slightly arched, not held out straight. The third and fourth fingers should be permitted to curl naturally toward the palm of the hand, but they should not be tightly clenched. Keep the fingers, hand and wrist relaxed at all times.

Now that you are all set, you can begin sending practice. Before taking up letter-sounds, it is best to achieve smoothness and facility with element sounds, as we did in learning by voice. Begin by making a series of dits — ten or more in a row. Make them evenly spaced, precise, a bit slower than the speed at which you learned the voice-sounds. Work on both dits and dahs, and then alternate, just as you did in voice work. Here again a metronome, or a substitute such as someone rapping on the table for you, will be a valuable timing guide. Remember to synchronize your sending at four dits, or two dahs, per beat. Make the beats about 60 per minute — one per second — as compared to the 100 in receiving practice. Synchronize your sending with the spoken sounds, if you wish. Keep your wrist flexible; allow it to bob up and down with your sending. If you find that your fingers or whole arm are doing the work, stop, and start over again: Be particularly careful to make dahs smooth and full; the tone should be almost continuous, broken only for that tiny instant the key contacts open.

When you have achieved smoothness in these exercises, begin on letter-sounds in the same sequence as you learned the voice, starting out with the five and the zero. Aim at perfect rhythm in duplicating just what you have learned by voice. In particular, be careful not to leave spaces between elements of letters. Practice sending material from a book, and also non-English text such as the cipher groups printed in this chapter. Keep your speed at the level at which you make a minimum of errors. Accuracy and perfection come first — speed will come with practice. Here again, it is well to have someone skilled in code check your progress occasionally.

• ADVANCED PRACTICE

To become expert in transmitting good code, after you have thoroughly learned each letter and numeral and can both send and copy letters without hesitation, your best practice is to listen to commercial automatic tape stations. Perfectly-sent code can be accomplished only by a machine, and you want to get fixed in your mind, indelibly, the correct formation of each and every code character and in particular the associated spaces. Notice the formation of each letter, the spaces left between letters and words, and the proportion in length of dits to dahs.

One of the best methods for deriving this association is to find a commercial or other tape station sending at about your maximum receiving speed. Listen to the transmissions as you would at a musical concert, concentrating on assimilating every detail.

The spaces between words may seem exaggerated, simply because you have probably been running yours together. A score of other details where the automatic transmission is different than yours will very likely show up in the same text. From all this you will learn where your own faults lie and be able to correct them.

If you can locate a tape station sending double (repeating each word) you can get excellent rhythm practice. Set up an audio oscillator alongside your short-wave receiver so you can hear both simultaneously. As each word comes through the receiver fix it in your mind; then, as the tape repeats it, send the same word on your oscillator simultaneously with the tape, as closely to perfect synchronism as possible. Perhaps you will find yourself leaving too much or too little space between characters, or making certain dahs too long — these are the most common errors. Remember that all inaccuracies are yours, and profit accordingly. By such constant practice you will learn the proper rhythm and precision of perfect code. It's bound to work itself, subconsciously, into your sending.

Probably the most important single factor in sending ability is this sense of proper spacing. While it can be much more easily obtained by listening to tape than by visual study, the student should understand the mechanical relation of the various marks and spaces, as shown in the accompanying chart.

In transmitting text there of course is a need for code symbols for punctuation marks, and some special procedure signals to facilitate rapid transmission. The important ones are:

period	didahdidahdidah
comma	dahdahdididahdah
question mark	dididahdahdidit
double dash	dahdidididah
end of message	didahdidahdit
end of work	didididahdidah
wait	didahdididit
invitation to transmit	dahdidah

Others which you might encounter in future code work, and which you therefore should learn after you become reasonably proficient with the previous characters, are:

hyphen	dahdididididah
parentheses	dahdidahdahdidah
colon	dahdahdahdididit
semicolon	dahdidahdidahdit
quotes	didahdididahdit
error	dididididididit
apostrophe	didahdahdahdahdit
fraction bar	dahdididahdit

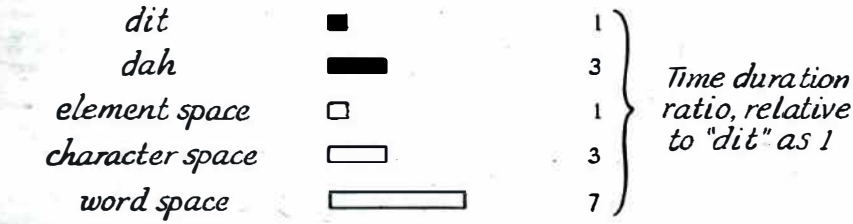
• HIGH-SPEED OPERATION

Perfect copy at high speeds should be the eventual objective of every operator. Ability to read rapid code "in your head" means little; what counts is what you can transcribe to paper correctly. Since the limit of writing ability is about 30 words per minute, one must resort to a typewriter for copy at higher speeds. The first essential is touch-typing ability; no "two-fingered" typist ever became a really good operator. If you have not been schooled in touch typing, therefore, try to find some class giving the necessary instruction; often you can find an evening public high school offering such courses. Remember there should be a margin of about twenty words per minute between your straight typing speed and the code speed which you can expect to obtain on the "mill."

Assuming you have touch typing ability, then, you simply substitute the typewriter for your pencil. Use standard letter-size paper, and write double-space. Do not try to capitalize any letters at first; all lower case type will suffice until you become proficient. Using some source of copy at about your normal receiving speed, such as a code machine or a commercial short-wave station if you can locate one, practice copying smoothly and evenly, preferably a letter or two behind the transmitted text. Do not listen and then type ferociously for a second . . . and listen . . . and type hurriedly again. Your typing must be dissociated, consciously, from your code reception. After you are able to handle this first speed, pick stations sending a bit faster (or step up the code machine), so that you get about 90 per cent of the text; when you copy solid, again step up the speed.

But all the ability is not in typing; you must be able to read the code as well. In high-speed copying a new principle is involved. It is one you will reach automatically if you progress sufficiently far in your practice, but it is worth explanation here.

When one first learns the code by sound, he learns letters first. (In some cases, students studying visually learn *parts* of letters first, but we have tried to obviate that error by teaching code sounds in letter-units.) With a





good deal of practice, one may slowly increase his speed of copying until he reaches a certain point — differing with various individuals — which is the maximum speed at which he can copy individual letters without having them seem to run together or blur in his mental thought. The average is around 28 words per minute.

Progress beyond that point must be on a new principle of copying, then. It is simply the process of copying by *word*-sounds instead of letter-sounds. An operator capable of receiving, say, 25 words per minute, can listen to 35 or 40 w.p.m. text and easily pick out the shorter, more common words such as "the," "and," "but," and so on. He can do so only because he is copying *word*-sounds and not letter-sounds. A skilled operator does not hear letters, but actual syllables and words. The code sound "didit dahdit dahdahdit" (the familiar suffix "ing"), for example, is mentally heard by the skilled operator as a complete sound and not as three different letters; again, it is as if someone had pronounced the syllable to him.

There is no secret to the attainment of this ability, except continued practice. You can help yourself, though, by having someone send to you rapidly (about 10 w.p.m. above your normal receiving speed) the common words and syllables such as and, of, the, to, a, in, is, it, for, ing, ion, of, on, end, that, and so on. Nothing can equal the practice obtainable by copying commercial tape transmissions, however.

It should be pointed out that one important prerequisite to high-speed copying is ability to spell. Since you do not hear letters, but entire word-sounds, spelling of a word on the typewriter is up to you. True, it was sent only one way, but since you did not hear individual letters you do not know exactly which were sent; you know only what the word sounded like to you. You must know the spelling of words before you can recognize them from their code sounds and be able to transcribe them correctly to the typewritten page.

The complement of high-speed receiving is of course high-speed sending. You should never let your sending speed outstrip your receiving speed, however; if that does happen, it simply means that your conception of the code is mechanical rather than in terms of sound. A semi-automatic key, or "bug," can be used for speeds higher than obtainable on a "straight" key, but in no event should a student attempt to handle a "bug" until he has mastered the regular key. The semi-automatic key is a mechanical device which produces a series of dits (when the proper lever is tripped) by means of a vibrating contact. Dahs are made in the usual manner, although the operating knobs work in a horizontal instead of a vertical motion. Relieved of the fatigue of tense motions involved in making rapid dits, the fingers are able to tap out code of much higher speed.

The position of the "bug" should be similar to that of the straight key. The arm again should be relaxed, and the right side of the hand should rest on the table immediately in front of the key. To the left of the control paddle is the thumb, which when moved to the right trips the dit vibrator; to the right of the paddle are the index and second fingers (some operators prefer to use only one), which when moved to the left operate the dah contact, similar to a straight key set on edge.

Operating motions consist of an easy roll of the wrist and hand from left to right, and return. With the key connected to a tone source, practice sending a series of dits, then of dahs, then of alternating single sounds — just as we did in learning voice sounds. Take letter-sounds in that sequence, as well. Before attempting to use a "bug" in actual operating, you should have become proficient with it through practice with an audio oscillator and headphones.

•TONE SOURCES FOR PRACTICE

A buzzer set, connected as shown in the diagram of Figure 1402, will serve the purpose of a signal source to be keyed for code practice.

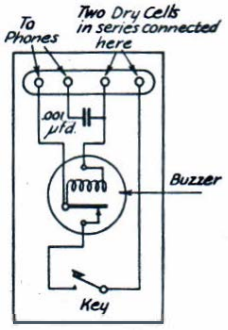


Fig. 1402 — Circuit of a buzzer code practice set. The headphones are connected across the coils of the buzzer with a condenser in series. The size of this condenser determines the strength of the signal in the 'phones. If the value shown gives an excessively loud signal, it may be reduced to 500 μμfd. or even 250 μμfd.

A much better tone, however, since it is identical to the audio beat note a radio operator reads, is that produced by a vacuum-tube audio oscillator. Beside a pair of headphones and key, the parts required are an old audio transformer, grid resistor and condenser, tube and socket. For battery operation, the tube may be any 1½- or 2-volt filament type, such as a 1G4G, 1LE3, '30, etc., power being sup-

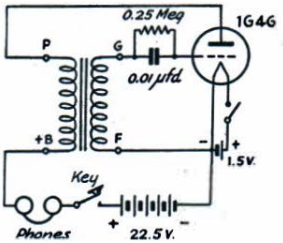


Fig. 1403 — Wiring diagram of a simple audio oscillator for use as a code practice set.

plied by a No. 6 dry cell and a 22½-volt B battery. The parts may be mounted on a baseboard, as shown in the layout drawing; or they may be enclosed in a portable carrying-case in which event flashlight cells and ten or more volts of B battery will suffice for power. If nothing is heard in the 'phones when the key is depressed, reverse the leads going to either transformer winding; reversing both sets of leads will have no effect. If desired, an oscillator powered from the 110-volt line may be used; the hookup of a simple one is shown in Figure 1405.

• A CODE-INSTRUCTION TABLE

The preceding material has been written primarily for the individual who must learn his code with little or no outside help. If a number of persons wish to learn the code as a group, or to develop speed after learning, and if some competent instructor can be located, the best method of instruction is by means of a code table.

Any such table should be so wired as to permit the instructor to send to the whole class, but by a little special wiring many other things are possible. In the one shown here, for instance, each student can practice sending, independently and to himself; yet the instructor, by means of the selector switch, can listen in on each student in turn, can break in on him and correct his errors. Moreover, simply by throwing a few switches, facing pairs of students may join in two-way conversations, and each such pair may be monitored in turn by

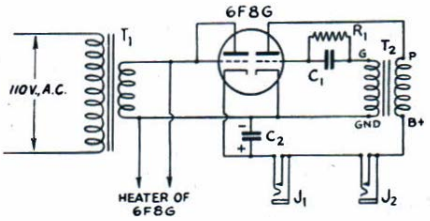


Fig. 1405 — Wiring diagram of the a.c. oscillator.
C_1 — 250-μμf'd. mica.
C_2 — 25-μfd. electrolytic, 25 volts.
J_1, J_2 — Small closed-circuit jacks.
R_1 — 0.15 megohms, ½-watt.
T_1 — 6.3-volt, 1-ampere transformer (Thordarson T-19F80).
T_2 — Small audio transformer (Thordarson T-13A34).

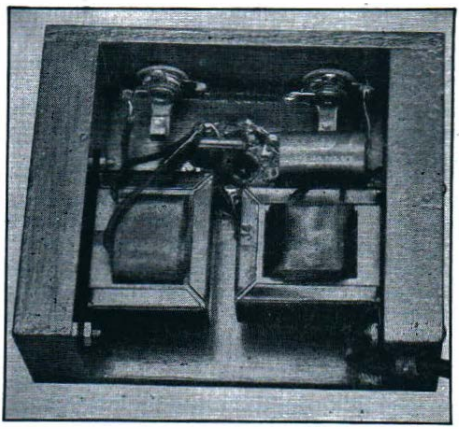


Fig. 1406 — The a.c. oscillator may be built on a home-made chassis, of which this is a bottom view, consisting of 1 x 2 white pine supporting strips and a baseboard of ¼-inch plywood. The 6F8G is the only component mounted on top the chassis.

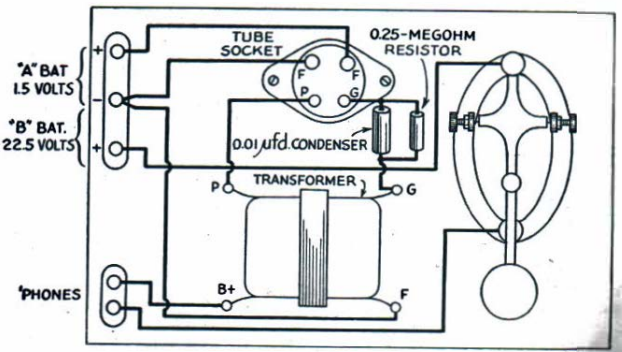


Fig. 1404 — Layout of the audio-oscillator code practice set. All parts can be screwed onto a wooden baseboard approximately 5 x 7 inches.

Example practice words, paragraphed in groups corresponding to the sequence of letters learned:

ate rat tear 50 era rate art 05 tare

sell request jar lure ruse stare suet squeal jet queer slate jute quart lesser quell just tesla sales

coverlet alone cancel vocal hover the collect never that shone lathe sheet these conquest there neutral severe tenet runner enclose reluctant jocular love helen lunar conclave recluse quiver

bottle corncob yesteryear honey this poppy jitters honey battery bay sissy reconciliate council bony phone join supper pave ship pay capon nylon coin pyrites copy boil pebble cavity vicious isotope

work buzzer zero alow hark warmer kind jerk suzanne bump quirk make wink simmer map skim milk tomorrow wholesome jam qualms lowly wharves pompom war mark womanish causeway know ersatz mink waltz

dock kind finger dolores dexterous kidder fixtures goodly golf jigger foggy jinx stuffing dog flight draught fling fox faddist god guffaw dagger mexican doddering textile exchange paradox xylograph

the instructor — all on the one table and from the common oscillator. Bigger "nets" can be switched at will, while other students at the same table continue practice either singly or in pairs, all under control and supervision.

The instructor should sit at one end of the table, facing down its length, the students arranged along each side. A center partition and cross-partitions make a sort of "private office" of each operating position, each with its 'phones, key and switch. The oscillator is at, or handy to, the instructor's end of the table, his controls on a low panel in front of him, running across the table.

At each student's position, a key and headset in series are bridged across the line carrying the continuously-running audio tone. The headsets (and keys) may be paralleled by closing the inter-connecting switches, each switch being associated with the position of the same number. A tap from each position is taken off to the multi-point switch (*S*) at the instructor's position so that the latter can also place his headset and key in parallel with those at any of the ten positions.

When the instructor sends to the entire class all the small switches are closed, connecting all positions in parallel. With all switches open each student can practice sending by himself, and the instructor can listen to any student by setting the selector switch *S* to the appropriate position. This also puts the instructor in parallel with the student at that position so that the instructor can "work" the student.

Suppose now that students are to practice together in groups of two. Closing Sw_2 connects Positions 1 and 2; closing Sw_4 connects 3 and 4; Sw_6 connects 5 and 6, and so on. The odd-numbered switches would be open in this case. The instructor can listen in on any group by setting the selector switch to either position in the group, and again can break in on the work. Larger groups can easily be formed; for instance, closing switches Nos. 2, 3 and 4 will connect positions 1, 2, 3 and 4 together, closing Sw_6 and Sw_7 will connect positions 5, 6 and 7 together, and so on.

The number of positions can be extended indefinitely by following the same wiring system. Separate tables, grouping perhaps ten men to a table, can be used if the room is too small to accommodate a single table for a large class. In such a case the instructor could have a central position with a separate selector switch for each table.

The code-practice oscillator described earlier in this chapter is suitable only for two or three sets of 'phones; it is, therefore, necessary to

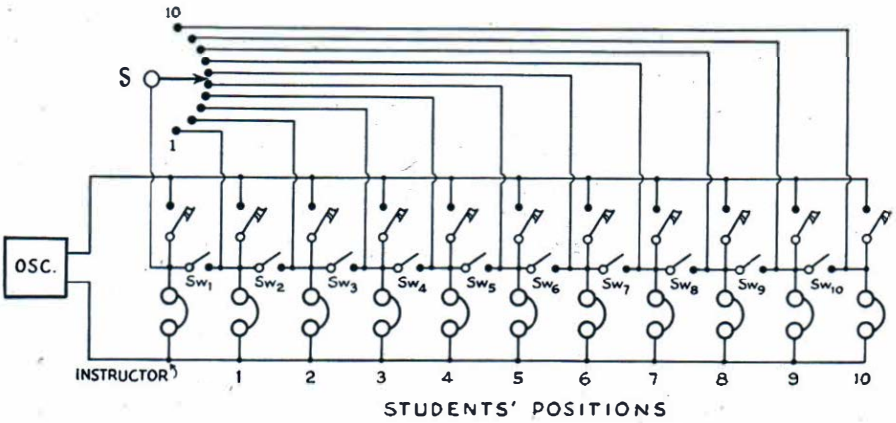


Fig. 1407 — Wiring diagram of the code-instruction table.

PRACTICE CIPHER GROUPS

Cipher groups make better practice material than plain English because you can't foresee the next letter. The groups below, taken from the operator's manual of the Signal Corps, are representative both of the practice material used in service schools and of ciphered messages actually used in the services. Their breakdown into 5-letter units makes it easy for you to determine your approximate speed, figured on the basis of 5 characters to a word.

With someone to send to you, this material of course provides the best possible copying practice. Check back for your errors, concentrate on the letters that are proving difficult for you. If you find yourself memorizing some of the combinations or their order, use each group backwards or start from the bottom of the page. There are plenty of practice possibilities in the following list, even for the experienced operator who is seeking higher speed on the typewriter.

OKICQ	CEMID	9 7 2 Ø 1	ALCRJ	MAIDP	JRNOL
2 4 6 8 Ø	JXEFY	EFXNK	DLZIK	7 6 3 2 1	EWSKM
CKTOG	EFMEY	LCKQH	PHVTX	7 Ø 3 6 5	HZGNA
AXBTR	4.9 2 8 5	OCYAA	HIMAD	OHWIM	SFMCG
Ø 3 7 5 9	JNABD	ZIJBI	CPNZI	VTNEE	FMEMI
CWXCK	JBCYD	8 4 7 Ø 7	JBOAH	HELKF	ZDLYS
DKAKX	PGZUN	KDKNG	GYEVZ	4 7 3 8 2	KTLES
WQYFZ	CTILL	FHFUX	ARJZU	MRFXE	WBPCM
OMIWG	THQVI	NTVMG	DXAQN	4 Ø 3 6 7	SPEJN
HSPCQ	HMOUS	WOHZR	1 1 9 Ø 2	OBRDK	VIODO
XGLDT	TVCPT	IPAZQ	Ø 6 1 4 3	PVFKQ	CUBEZ
8 Ø 7 Ø 1	CPTKO	FNZQJ	CPTKO	RBEFP	AXTRX
RKMOZ	2 Ø 1 8 4	PGMAE	IRAGO	KDIEY	KCNYQ
EKJCD	VNFBH	1 Ø 2 7 3	PZOKA	MDIDH	EIURY
DJHNG	DIEUY	DXNZE	OQUIW	Ø 9 1 6 5	MCNBS
HAHGJ	JCUWH	9 8 2 3 4	MCNUY	QZASU	IHDJI
LDOSJ	8 3 7 4 6	JEHDY	BCNBX	CVFZA	SCQZW
1 2 3 1 Ø	LKDOP	MCNCB	ZAOIZ	ZLKAM	ZJHFG
BAYQT	WEQFQ	PAOKU	3 Ø 2 1 Ø	JDMNF	DKUER
QNBAV	1 7 8 6 3	MDHNH	TQRYW	EIOUW	MXNBG
2 5 8 9 4	MVHVD	UEYTA	ZMNZB	ZXVGX	OIEUQ
JXHGQ	APLKM	QYEUR	NDBGY	UYRHI	LKCCN
9 8 Ø 1 5	NDHBH	QWZXN	KDIOE	DYAFH	LPWCA
MBUDF	JTIOE	MNOPA	CVQAR	1 2 7 5 4	OEVAN
AEIOU	MNLFG	KQZAX	NASTF	PONDL	FAWZX
1 3 5 7 9	YOUMK	LARDO	CFGIL	NPQRT	QWERT
CJRNO	VOIUM	MQECV	CADXA	SNKAD	CUTYR
YOUMK	1 2 7 Ø 9	BEXZB	TTLOR	KOBDZ	RGLBH

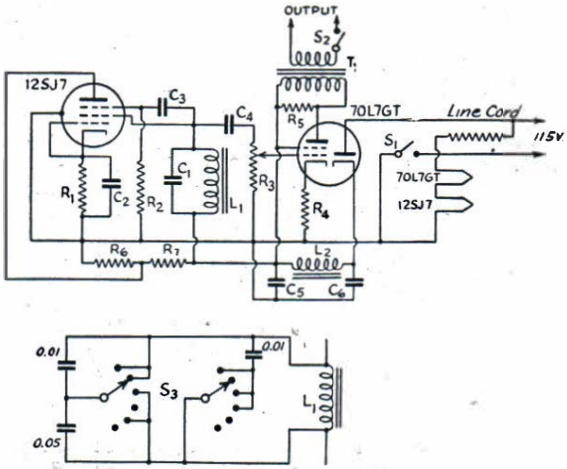


Fig. 1408 — Circuit of code-practice oscillator suitable for large groups.

C_1 — 0.05-μfd. paper (for tone of approximately 650 cycles with L_1 = 1 henry); substitute lower circuit for variable tone.
C_2 — 25-μfd. electrolytic, 25-volt.
C_3, C_4 — 0.01-μfd. paper.
C_5, C_6 — 30-μfd. electrolytic, 150-volt.
R_1 — 2000 ohms, 1-watt.
R_2 — 0.25 megohm, ½-watt.
R_3 — 0.25-megohm volume control.
R_4 — 150 ohms, 1-watt.
R_5 — 3000 ohms, 1-watt.
R_6, R_7 — 10,000 ohms, 1-watt.
L_1 — 1 henry, approx. (see text).
L_2 — 8-henry 55-ma. filter choke.
T_1 — Universal output transformer, tube to voice coil (set for matching 8-ohm v.c. to 2000-ohm plate load).
S_1, S_2 — S.p.s.t. toggle.
S_3 — 2-pole 6-position wafer switch.
Line Cord — 220-ohm.

build a special type of oscillator for use with the code table. A circuit diagram suitable for handling up to thirty or forty headsets is shown in the diagram herewith. Operating directly from the 115-volt line, it consists of a 12SJ7 oscillator followed by a 70L7GT power amplifier and rectifier. The pitch of the tone is determined by the constants of the tuned circuit L_1C_1. The inductance should be of the order of 1 or 2 henrys for use with readily-available paper condensers. In the unit shown in the photographs this inductance is an ordinary small filter choke (Thordarson T-14C61) with the straight section of the core removed and a piece of wood of the same size and shape substituted so that the mounting clamp can be replaced. This gives a choke of approximately 1 henry inductance.

A choice of tones should be available to avoid monotony. Variable tone is secured by

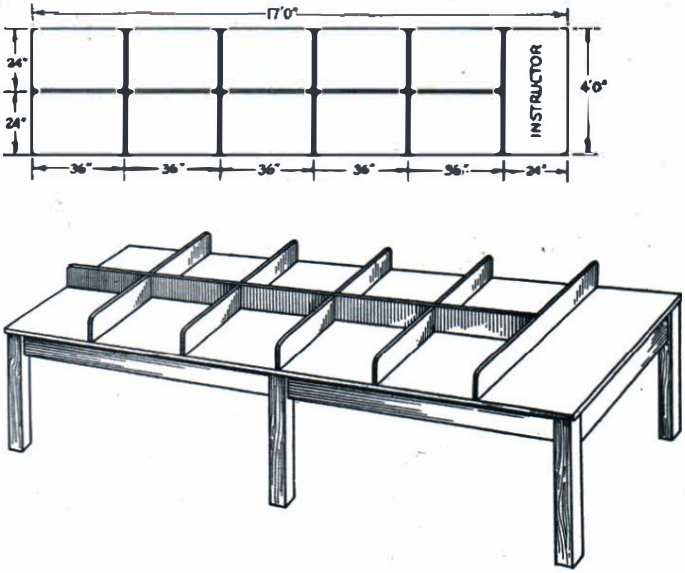


Fig. 1409 — A suggested form of construction for the code-instruction table. Dimensions may be varied to suit individual needs. Legs may be made of 4 by 4 stock; top and side rails of 1-inch boards. Appropriate bracing should be provided underneath. A coat or two of shellac after the table is finished will prevent shrinkage.

means of the switching arrangement shown below the main wiring diagram. With three condensers of the values indicated, an assortment of six tones ranging from about 600 to 1600 cycles can be obtained. If only one tone is needed, however, simply use the constants shown in the main diagram.

The output switch S_2 enables the instructor to cut off the tone from the entire class, and thus serves as a simple means of attracting attention when group work or individual sending practice is being carried on.

No specific construction is indicated for the table, since bracing and other details will depend upon the size, and the builder's preferences. The important thing is to be sure each position provides enough room for the student to work comfortably. In particular there should be sufficient depth so that the key can be placed the proper distance — about 18 inches — from the edge of the table.

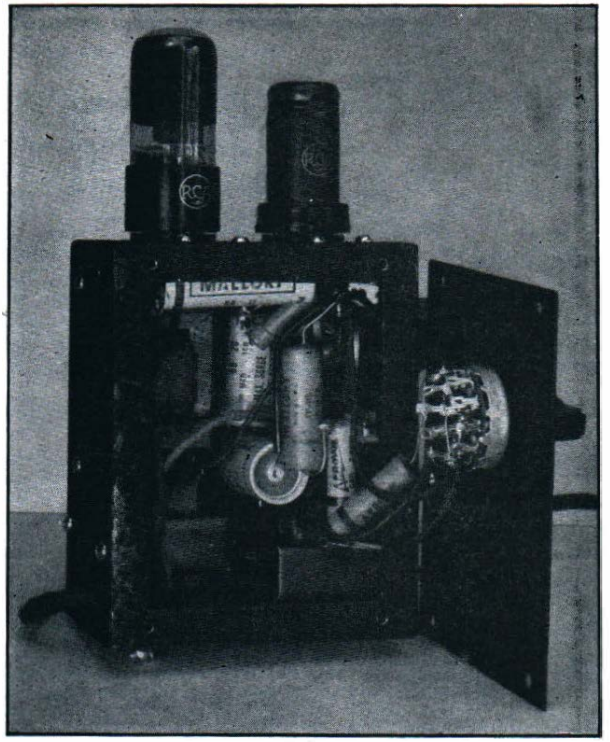


Fig. 1410 — The code-practice oscillator circuit of Fig. 1408 built into a 3 by 4 by 5 box. All wiring is insulated from the case.

The interconnecting switches, Sw_1, Sw_2, etc., can be ten-eent store s.p.s.t. knife switches and the instructor's selector switch a wafer-type unit. The latter can be obtained in the single-pole type with as many as 23 contacts. Since no special precautions need be. taken with respect to insulation or voltage drop, probably the most economical wiring job can be done by using ordinary bell wire. In cases where the students bring their own headsets it would be advisable to mount a tip-jack assembly at each position to facilitate connection.

Scales Used in Expressing Signal Strength and Readability

(See QRK and QSA in the Q Code)

Strength		*Readability*	
QSA1	Barely perceptible.	QRK1	Unreadable.
QSA2	Weak.	QRK2	Readable now and then.
QSA3	Fairly good.	QRK3	Readable with difficulty.
QSA4	Good.	QRK4	Readable.
QSA5	Very good.	QRK5	Perfectly readable.

● "Q" CODE

IN THE REGULATIONS accompanying the existing International Radiotelegraph Convention there is a very useful internationally-agreed code designed to meet major needs in international radio communication. This code follows. The abbreviations themselves have the meanings shown in the "answer" column. When an abbreviation is followed by an interrogation mark (?) it assumes the meaning shown in the "question" column.

Abbreviation	*Question*	*Answer*
QRA	What is the name of your station?	The name of my station is .....
QRB	How far approximately are you from my station?	The approximate distance between our stations is ..... nautical miles (or ..... kilometers).
QRC	What company (or Government Administration) settles the accounts for your station?	The accounts for my station are settled by the ..... company (or by the Government Administration of .....).
QRD	Where are you bound and where are you from?	I am bound for ..... from .....
QRG	Will you tell me my exact frequency (wave-length) in kc/s (or m)?	Your exact frequency (wave-length) is ..... kc/s (or ..... m).
QRH	Does my frequency (wave-length) vary?	Your frequency (wave-length) varies.
QRI	Is my note good?	Your note varies.
QRJ	Do you receive me badly? Are my signals weak?	I cannot receive you. Your signals are too weak.
QRK	What is the legibility of my signals (1 to 5)?	The legibility of your signals is .... (1 to 5).
QRL	Are you busy?	I am busy (or I am busy with .....). Please do not interfere.
QRM	Are you being interfered with?	I am being interfered with.
QRN	Are you troubled by atmospherics?	I am troubled by atmospherics.
QRO	Shall I increase power?	Increase power.
QRP	Shall I decrease power?	Decrease power.
QRQ	Shall I send faster?	Send faster (..... words per minute).
QRS	Shall I send more slowly?	Send more slowly (..... words per minute).
QRT	Shall I stop sending?	Stop sending.
QRU	Have you anything for me?	I have nothing for you.
QRV	Are you ready?	I am ready.
QRW	Shall I tell ..... that you are calling him on ..... kc/s (or ..... m)?	Please tell ..... that I am calling him on ..... kc/s (or ..... m).
QRX	Shall I wait? When will you call me again?	Wait (or wait until I have finished communicating with .....) I will call you at ..... o'clock (or immediately).
QRY	What is my turn?	Your turn is No. ..... (or according to any other method of arranging it).
QRZ	Who is calling me?	You are being called by .....
QSA	What is the strength of my signals (1 to 5)?	The strength of your signals is ..... (1 to 5).
QSB	Does the strength of my signals vary?	The strength of your signals varies.
QSD	Is my keying correct; are my signals distinct?	Your keying is incorrect; your signals are bad.
QSG	Shall I send ..... telegrams (or one telegram) at a time?	Send ..... telegrams (or one telegram) at a time.
QSJ	What is the charge per word for ..... including your internal telegraph charge?	The charge per word for ..... is ..... francs, including my internal telegraph charge.
QSK	Shall I continue with the transmission of all my traffic, I can hear you through my signals?	Continue with the transmission of all your traffic, I will interrupt you if necessary.
QSL	Can you give me acknowledgment of receipt?	I give you acknowledgment of receipt.
QSM	Shall I repeat the last telegram I sent you?	Repeat the last telegram you have sent me.
QSO	Can you communicate with ..... direct (or through the medium of .....)?	I can communicate with ..... direct (or through the medium of .....).
QSP	Will you retransmit to ..... free of charge?	I will retransmit to ..... free of charge.
QSR	Has the distress call received from ..... been cleared?	The distress call received from ..... has been cleared by .....
QSU	Shall I send (or reply) on ..... kc/s (or m) and/or on waves of Type A1, A2, A3, or B?	Send (or reply) on ..... kc/s (or ..... m) and/or on waves of Type A1, A2, A3, or B.
QSV	Shall I send a series of VVV .....?	Send a series of VVV .....

Abbreviation	Question	Answer
QSW	Will you send on ..... kc/s (or ..... m) and/or on waves of Type A1, A2, A3, or B?	I am going to send (or I will send) on ..... kc/s (or ..... m) and/or on waves of Type A1, A2, A3, or B.
QSX	Will you listen for ..... (call sign) or ..... kc/s (or ..... m)?	I am listening for ..... (call sign) on ..... kc/s (or ..... m).
QSY	Shall I change to transmission on ..... kc/s (or ..... m) without changing the type of wave? or Shall I change to transmission on another wave?	Change to transmission on ..... kc/s (or ..... m) without changing the type of wave or Change to transmission on another wave.
QSZ	Shall I send each word or group twice?	Send each word or group twice.
QTA	Shall I cancel telegram No. ..... as if it had not been sent?	Cancel telegram No. ..... as if it had not been sent.
QTB	Do you agree with my number of words?	I do not agree with your number of words; I will repeat the first letter of each word and the first figure of each number.
QTC	How many telegrams have you to send?	I have ..... telegrams for you (or for .....).
QTE	What is my true bearing in relation to you? or What is my true bearing in relation ..... (call sign)?	Your true bearing in relation to me is .... degrees or Your true bearing in relation to ..... (call sign) is ..... degrees at ..... (time) or
	What is the true bearing of ..... (call sign) in relation to ..... (call sign)?	The true bearing of ..... (call sign) in relation to ..... (call sign) is ..... degrees at ..... (time).
QTF	Will you give me the position of my station according to the bearings taken by the direction-finding stations which you control?	The position of your station according to the bearings taken by the direction-finding stations which I control is ..... latitude ..... longitude.
QTG	Will you send your call sign for fifty seconds followed by a dash of ten seconds on ..... kc/s (or ..... m) in order that I may take your bearing?	I will send my call sign for fifty seconds followed by a dash of ten seconds on ..... kc/s (or ..... m) in order that you may take my bearing.
QTH	What is your position in latitude and longitude (or by any other way of showing it)?	My position is ..... latitude ..... longitude (or by any other way of showing it).
QTI	What is your true course?	My true course is ..... degrees.
QTJ	What is your speed?	My speed is ..... knots (or ..... kilometers) per hour.
QTM	Send radioelectric signals and submarine sound signals to enable me to fix my bearing and my distance.	I will send radioelectric signals and submarine sound signals to enable you to fix your bearing and your distance.
QTO	Have you left dock (or port)?	I have just left dock (or port).
QTP	Are you going to enter dock (or port)?	I am going to enter dock (or port).
QTQ	Can you communicate with my station by means of the International Code of Signals?	I am going to communicate with your station by means of the International Code of Signals.
QTR	What is the exact time?	The exact time is .....
QTU	What are the hours during which your station is open?	My station is open from ..... to .....
QUA	Have you news of ..... (call sign of the mobile station)?	Here is news of ..... (call sign of the mobile station).
QUB	Can you give me in this order, information concerning: visibility, height of clouds, ground wind for ..... (place of observation)?	Here is the information requested .....
QUC	What is the last message received by you from ..... (callsign of the mobile station)?	The last message received by me from ..... (call sign of the mobile station) is .....
QUD	Have you received the urgency signal sent by ..... (call sign of the mobile station)?	I have received the urgency signal sent by ..... (call sign of the mobile station) at ..... (time).
QUF	Have you received the distress signal sent by ..... (call sign of the mobile station)?	I have received the distress signal sent by ..... (call sign of the mobile station) at ..... (time).
QUG	Are you being forced to alight in the sea (or to land)?	I am forced to alight (or land) at ..... (place).
QUH	Will you indicate the present barometric pressure at sea level?	The present barometric pressure at sea level is ..... (units).
QUJ	Will you indicate the true course for me to follow, with no wind, to make for you?	The true course for you to follow, with no wind, to make for me is ..... degrees at ..... (time).
QUK	Can you tell me the condition of the sea observed at ..... (place or coördinates)?	The sea at ..... (place or coördinates) is .....
QUL	Can you tell me the swell observed at ..... (place or coördinates)?	The swell at ..... (place or coördinates) is .....
QUM	Is the distress traffic ended?	The distress traffic is ended.

Special abbreviations adopted by the American Radio Relay League.

QST General call preceding a message addressed to all amateurs and ARRL Members. This is in effect "CQ ARRL."

QRR Official ARRL "land SOS." A distress call for use by stations in emergency zones only.

Miscellaneous Data

THIS CHAPTER represents a compilation of miscellaneous data and reference information intended to illustrate and supplement the basic material throughout the remainder of this *Handbook*.

Inductance (L)

The formula for computing the inductance of air-core radio coils is:

$$L = \frac{0.2\ A^2N^2}{3A + 9B + 10C}$$

where: L is the inductance in microhenrys
A is the mean diameter of the coil in inches
B is the length of winding in inches
C is the radial depth of winding in inches
N is the number of turns.

The quantity C may be neglected if the coil is a single-layer solenoid, as is nearly always the case with coils for high frequencies.

For example, assume a coil having 35 turns of No. 30 d.s.c. wire on a receiving coil form having a diameter of 1.5 inches. Consulting the wire table, we find that 35 turns of No. 30 d.s.c. will occupy a length of one-half inch. Therefore,

$$A = 1.5$$
$$B = .5$$
$$N = 35$$

and

$$L = \frac{0.2 \times (1.5)^2 \times (35)^2}{(3 \times 1.5) + (9 \times .5)}$$

or 61.25 microhenrys.

To calculate the number of turns of a single-layer coil for a required value of inductance:

$$N = \sqrt{\frac{3A + 9B}{0.2A^2} \times L}.$$

More rapid and convenient calculations in designing coils can be made with the ARRL *Lightning Radio Calculator* (Type A).

Condenser Capacity (C)

The formula for the capacitance of a condenser is:

$$C = 0.0088\ \frac{kA}{d}\ (n-1)\ \mu\text{fd}.$$

where: A = area of one side of plate (sq. cm.
n = total number of plates
d = separation of plates (cm.)
k = dielectric constant of dielectric.

When A is the area of one side of one plate *in square inches* and d is the separation of the plate *in inches*,

$$C = 0.02235\ \frac{kA}{d}\ (n-1)\ \mu\mu\text{fd}.$$

The dielectric constant determines the quantity of charge which a given separation and area of plates will accumulate for a given applied voltage. "k" is the ratio of the capacitance of a condenser with a given dielectric to its capacitance with air dielectric.

Table of Dielectric Constants

Dielectric	"k"	*Power Factor* [1]	*Puncture Voltage* [2]
Air (normal pressure)	1.0	—	19.8–22.8
Amber	2.0	0.2–0.5	—
Asphalts	2.7–3.1	2.3 [3]	25–30
Bakelite — See Phenol			
Beeswax	2.9–3.2	—	—
Casein plastics [4]	6.1–6.4	5.2–6	165
Caster oil	4.3–4.7	7	380
Celluloid	4–16	5–10	—
Cellulose Acetate	6–8	3–6	600
Cellulose Nitrate	4–7	2.8–5	300
Ceresin wax	2.5–2.6	0.12–0.21	
Enamel (wire)	—	—	500–750
Fibre	5–7.5	4.5–5	150–180
Glass:			
Cobalt	7.3	0.7	—
Common window	7.6–8	1.4	200–250
Crown	6.2–7	1 [3]	500
Electrical	4–5	0.5	2000
Flint	7–10	0.4	—
Nonex	4.2	0.25	—
Photographic	7.5	0.8–1	—
Plate	6.8–7.6	0.6–0.8	—
Pyrex	4.5	0.7	335
Gutta Percha	2.5–4.9	—	200–500
Lucite	2.5–3	—	—
Mica	2.5–8	0.01–0.06	—
Mica (clear India)	6.4–7.3	0.01–0.02	600–1500
Mycalex	6–8	0.2–0.3	250
Paper	2.0–2.6	—	—
Paraffin wax (solid)	1.9–2.6	0.1–0.3	300
Phenol: [5]			
Pure	5	1	—
Asbestos base	7.5	15	90–150
Black molded	5–5.5	3.5	400–500
Fabric base	5–6.5	3.5–10	150–500
Mica-filled	5–6	0.8–1	475–600
Paper base	3.8–5.5	2.5–4	650–750
Yellow	5.3–5.4	0.36–0.7	500
Polystyrene [6]	2.4–2.9	0.02	500
Porcelain (dry process)	6.2–7.5	0.7–15	40–100

Porcelain (wet process)	6.5–7	0.6	150
Pressboard (untreated)	2.9–4.5	—	125–300
Pressboard (oiled)	5	—	750
Quartz (fused)	4.2–5.1	0.03	200
Rubber (hard) [7]	2–3.5	0.5–1	450
Shellac	2.5–4	0.09	900
Steatite [8]	6.1	0.06–0.2	150–315
Titanium Dioxide [9]	90–170	0.1	—
Urea Formaldehyde resins [10]	5–7	2.4	300–400
Varnished cloth (black)	2	2	550
Varnished cloth (yellow) [11]	2.5	3	440
Vinyl resins	4	1.4–1.7	400–500
Vitrolex	6.4	0.3	—
Wood (dry oak)	2.5–6.8	3.85	—
Wood (paraffined maple)	4.1	—	115

[1] At 1 Mc.

[2] In kilovolts per inch. Most data applies to relatively thin sections and cannot be multiplied directly to give breakdown for thicker sections without added safety factor.

[3] At 1 kc.

[4] Includes such products as Aladdinite, Galalith, Erinoid, Lactoid, etc.

[5] Phenolaldehyde products include Acrolite, Bakelite, Celeron, Dielecto, Durez, Durite, Formica, Micarta, Synthane, Textolite, etc. Yellow bakelite is so-called "low-loss" bakelite.

[6] Includes Amphenol 912A, QuartzQ, Styron, Trolitul, Victron, etc.

[7] Also known as Ebonite.

[8] Soapstone — Alberene, Alsimag, Isolantite, Lava, etc.

[9] Rutile. Used in low-temperature-coefficient fixed condensers.

[10] Includes Aldur, Beetle, Plaskon, Pollopas, Prystal, etc.

[11] Includes Empire cloth.

RMA RADIO COLOR CODES

Standard color codes have been adopted by the Radio Manufacturers Association for the identification of the values and connections of standard components.

Resistors and Condensers:

For identification of resistance and capacitance values of small carbon-type resistors and midget mica condensers, numbers are represented by the following colors:

0 — Black	5 — Green
1 — Brown	6 — Blue
2 — Red	7 — Violet
3 — Orange	8 — Gray
4 — Yellow	9 — White

Three colors are used on each resistor to identify its value. The body color represents the first figure of the resistance value; one end or tip is colored to represent the second figure; a colored band or dot near the center of the resistor gives the number of zeros following the first two figures. A 25,000-ohm resistor, for example, would be marked as follows: body, red (2); tip, green (5); band, orange (3 zeros).

Small mica condensers usually are marked with three colored dots, with an arrow or other symbol indicating the sequence of colors. Readings are in micromicrofarads ($\mu\mu$fd.), with the color code as above. For example, a 0.00025-μfd. (250-$\mu\mu$fd.) condenser would be marked as follows: red (2), green (5), brown (1 zero).

An auxiliary color code has been established covering the *tolerances* of resistors and condensers and the *voltage ratings* of condensers. Tolerances are indicated by the following colors, which appear as a fourth dot or band:

Brown	1%	Violet	7%
Red	2%	Gray	8%
Orange	3%	White	9%
Yellow	4%	Gold	5%*
Green	5%	Silver	10%*
Blue	6%	None	20%*

* Commonly used on resistors.

Condenser voltage ratings are indicated by a supplementary dot or band as follows:

Brown	100 volts	Violet	700 volts
Red	200 volts	Grey	800 volts
Orange	300 volts	White	900 volts
Yellow	400 volts	Gold	1000 volts
Green	500 volts	Silver	2000 volts
Blue	600 volts	None	500 volts

In some instances condensers are identified by a second (lower) row of three dots or a series of three narrow color bands to the right of the wide capacitance bands. In these cases the first color of the second row or series indicates the number of cipbers following the capacitance digits, the second color indicates tolerance and the third voltage rating.

I.F. Transformers:

Blue — plate lead.
Red — B + lead.
Green — grid (or diode) lead.
Black — grid (or diode) return.

NOTE: If the secondary of the i.f.t. is center-tapped, the second diode plate lead is green-and-black striped, and black is used for the center-tap lead.

A.F. Transformers:

Blue — plate (finish) lead of primary.
Red — B + lead (this applies whether the primary is plain or center-tapped).
Brown — plate (start) lead on center-tapped primaries. (Blue may be used for this lead if polarity is not important.)
Green — grid (finish) lead to secondary.
Black — grid return (this applies whether the secondary is plain or center-tapped).
Yellow — grid (start) lead on center-tapped secondaries. (Green may be used for this lead if polarity is not important.)

NOTE: These markings apply also to line-to-grid, and tube-to-line transformers.

Loudspeaker Voice Coils:

Green — finish.
Black — start.

Field Coils:

Black and Red — start.
Yellow and Red — finish.
Slate and Red — tap (if any).

Power Transformers:

1. Primary Leads ..... *Black*
   If tapped:
   Common ..... *Black*
   Tap ..... *Black and Yellow Striped*
   Finish ..... *Black and Red Striped*
2. High-Voltage Plate Winding ..... *Red*
   Center-Tap ..... *Red and Yellow Striped*
3. Rectifier Fil. Winding ..... *Yellow*
   Center-Tap ..... *Yellow and Blue Striped*
4. Fil. Winding No. 1 ..... *Green*
   Center-Tap ..... *Green and Yellow Striped*
5. Fil. Winding No. 2 ..... *Brown*
   Center-Tap ..... *Brown and Yellow Striped*
6. Fil. Winding No. 3 ..... *Slate*
   Center-Tap ..... *Slate and Yellow Striped*

LC Constants

The product of the inductance and capacity required to resonate at a given frequency is known as the LC constant. If any value

of inductance (or capacity) is divided into that constant the quotient is the capacity (or inductance) required for resonance at that frequency. The following table gives LC constants for amateur and intermediate frequencies in terms of microhenries and micromicrofarads:

Frequency Kc.	*LC Constant μh. × μμfd.*	*Frequency Mc.*	*LC Constant uh. × μμfd.*
100	2,533,030.	7.0	516.944
455	122,355.	7.3	475.399
1000	25,330.3	14.0	129.236
1600	9,894.64	14.4	122.185
1750	8,271.12	28.0	32.3090
1900	7,016.68	30.0	28.1448
2000	6,332.57	56.0	8.07726
3500	2,067.78	60.0	7.03620
4000	1,583.14	112.0	2.01931
5000	1,013.21	116.0	1.88245

For other frequencies:

$$LC = \frac{25330.3}{(\text{Freq.}_{Mc.})^2}$$

(The above data are contributed by Henry R. Hesse, W2ERY.)

Symbols for Electrical Quantities

Admittance	Y, y
Angular velocity ($2\pi f$)	ω
Capacitance	C
Conductance	G, g
Conductivity	v
Current	I, i
Difference of potential	E, e
Dielectric constant	K or ϵ
Energy	W
Frequency	f
Impedance	Z, z
Inductance	L
Magnetic intensity	H
Magnetic flux	Φ
Magnetic flux density	B
Magnetomotive force	F
Mutual inductance	M
Number of conductors or turns	N
Permeability	μ
Phase displacement	θ or Φ
Power	P, p
Quantity of electricity	Q, q
Reactance	X, x
Reactance, Capacitive	X_c
Reactance, Inductive	X_L
Resistance	R, r
Resistivity	P
Susceptance	b
Speed of rotation	n
Voltage	E, e
Work	W

Letter Symbols for Vacuum Tube Notation

Grid potential	E_g, e_g
Grid current	I_g, i_g
Grid conductance	g_g
Grid resistance	r_g
Grid bias voltage	E_c
Plate potential	E_p, e_p
Plate current	I_b, I_p, i_p
Plate conductance	g_p
Plate resistance	r_p
Plate supply voltage	E_b
Emission current	I_e
Mutual conductance	g_m
Amplification factor	μ
Filament terminal voltage	E_f
Filament current	I_f
Grid-plate capacity	C_{gp}
Grid-cathode capacity	C_{gk}
Plate-cathode capacity	C_{pk}
Grid capacity (input)	C_g
Plate capacity (output)	C_p

NOTE — Small letters refer to instantaneous values.

Units of Length

English	*Metric*
1 mil = 0.001 inch = 0.0254 millimeter	1 millimeter = 39.37 mils
1 inch = 2.54 centimeters	1 centimeter = 0.3937 inch = 0.0328 foot
1 foot = 30.48 centimeters	1 meter = 3.28 feet
1 yard = 0.9144 meter	= 1.094 yards
1 mile = 1.6093 kilometers	1 kilometer = 0.6214 mile
1 micron = 10^{-6} meter = 0.0001 centimeter = 10,000 Angstrom units (A°)	
1 Angstrom = 10^{-10} meter = 10^{-8} centimeter = 0.0001 micron	

Relative Electrical Conductivity of Metals at Ordinary Temperatures

(Based on Copper as 100)

Aluminum (28; pure)	59	Iron (cast)	2–12
Aluminum (alloys):		Iron (wrought)	11.4
Soft-annealed	45–50	Lead	7
Heat-treated	30–45	Manganin	3.7
Brass	28	Mercury	1.66
Cadmium	19	Molybdenum	33.2
Chromium	55	Monel	4
Climax	1.83	Nichrome	1.45
Cobalt	16.3	Nickel	12–16
Constantin	3.24	Phosphor Bronze	36
Copper (hard drawn)	89.5	Platinum	15
Copper (annealed)	100	Silver	106
Everdur	6	Steel	3–15
German Silver (18%)	5.3	Tin	13
Gold	65	Tungsten	28.9
Iron (pure)	17.7	Zinc	28.2

Approximate relations:
An increase of 1 in A.W.G. or B. & S. wire size increases resistance 25%.
An increase of 2 increases resistance 60%.
An increase of 3 increases resistance 100%.
An increase of 10 increases resistance 10 times.

Current Capacity of Power Wiring

The National Board of Fire Underwriters has established the following as maximum current densities for commonly-used sizes of copper wire in electrical power circuits:

Gauge No. B. & S.	Circular Mil Area	Amperes	
		Rubber Insulation	Other Insulation
1	83690	100	150
2	66370	90	125
4	41740	70	90
6	26250	50	70
8	16510	35	50
10	10380	25	30
12	6530	20	25
14	4107	15	20
16	2583	6	10
18	1624	3	6

Greek Alphabet

Since Greek letters are used to stand for many electrical and radio quantities, the names and symbols of the Greek alphabet with the equivalent English characters are given.

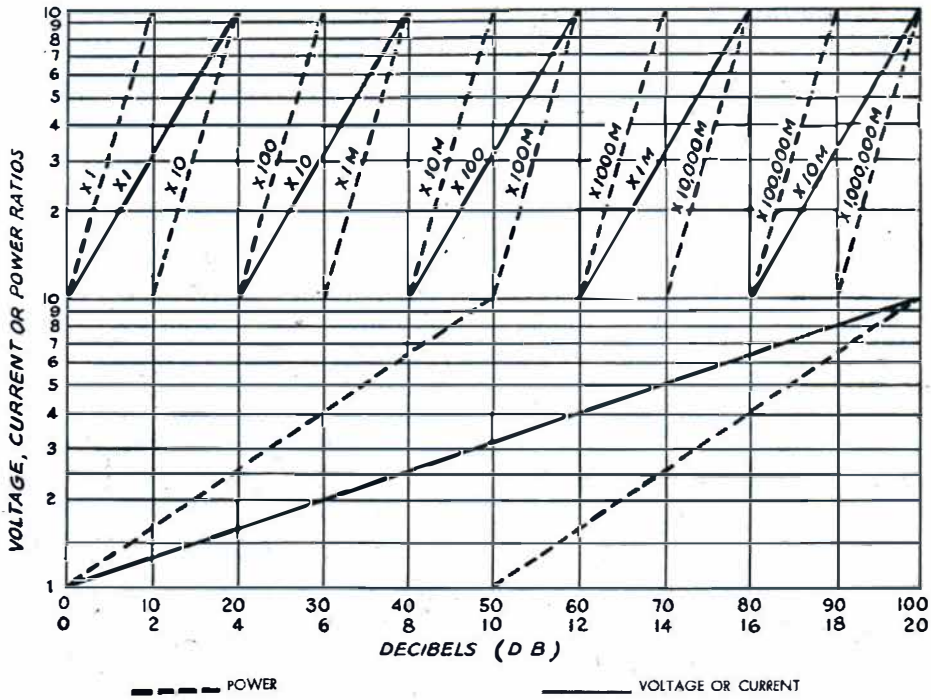
Greek Letter	Greek Name	English Equivalent
A α	Alpha	a
B β	Beta	b
Γ γ	Gamma	g
Δ δ	Delta	d
E ϵ	Epsilon	e
Z ζ	Zeta	z
H η	Eta	é
Θ θ	Theta	th
I ι	Iota	i
K κ	Kappa	k
Λ λ	Lambda	l
M μ	Mu	m
N ν	Nu	n
Ξ ξ	Xi	x
O o	Omicron	ŏ
Π π	Pi	p
P ρ	Rho	r
Σ σ	Sigma	s
T τ	Tau	t
T υ	Upsilon	u
Φ ϕ	Phi	ph
X χ	Chi	ch
Ψ ψ	Psi	ps
Ω ω	Omega	ō

Multiples and Sub-Multiples

Ampere	= 1,000,000 microamperes
Ampere	= 1,000 milliamperes
Cycle	= 0.000,001 megacycle
Cycle	= 0.001 kilocycle
Farad	= 1,000,000,000,000 micro-microfarads
Farad	= 1,000,000 microfarads
Farad	= 1,000 millifarads
Henry	= 1,000,000 microhenrys
Henry	= 1,000 millihenrys
Kilocycle	= 1,000 cycles
Kilovolt	= 1,000 volts
Kilowatt	= 1,000 watts
Megacycle	= 1,000,000 cycles
Megohm	= 1,000,000 ohms
Mho	= 1,000,000 micromhos
Mho	= 1,000 millimhos
Microampere	= 0.000,001 ampere
Microfarad	= 0.000,001 farad
Microhenry	= 0.000,001 henry
Micromho	= 0.000,001 mho
Micro-ohm	= 0.000,001 ohm
Microvolt	= 0.000,001 volt
Microwatt	= 0.000,001 watt
Micromicrofarad	= 0.000,000,000,001 farad
Micromicro-ohm	= 0.000,000,000,001 ohm
Milliampere	= 0.001 ampere
Millihenry	= 0.001 henry
Millimho	= 0.001 mho
Milliohm	= 0.001 ohm
Millivolt	= 0.001 volt
Milliwatt	= 0.001 watt
Volt	= 1,000,000 microvolts
Volt	= 1,000 millivolts
Watt	= 1,000,000 microwatts
Watt	= 1,000 milliwatts
Watt	= 0.001 kilowatt

Metric Prefixes

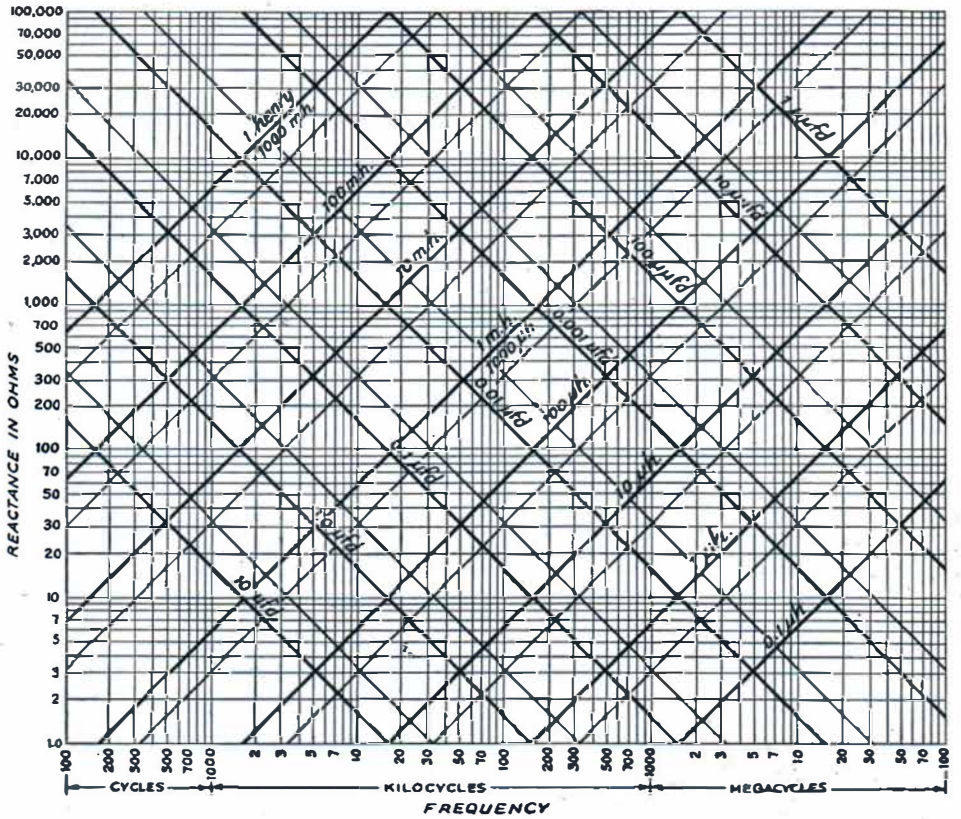
μ	$\frac{1}{1,000,000}$	One-millionth	micro-
m	$\frac{1}{1,000}$	One-thousandth	milli-
c	$\frac{1}{100}$	One-hundredth	centi-
d	$\frac{1}{10}$	One-tenth	deci-
	1	One	uni-
dk	10	Ten	deka-
h	100	One hundred	hekto-
k	1,000	One thousand	kilo-
	10,000	Ten thousand	myria-
M	1,000,000	One million	mega-



The chart above is direct-reading in terms of decibels for all power, voltage or current ratios. The top scale goes from 0 to 100 db. and is useful for very large ratios; the lower scale permits closer reading between 0 and 20 db., or one cycle of the extended scale. Solid lines show voltage or current ratios; dotted lines, power ratios. To find db. gain, divide output power by corresponding input power and read db. value for this ratio, using the appropriate curve (i.e., " × 1" for ratios from 1 to 10, " × 10" for ratios from 10 to 100, " × 100" for ratios from 100 to 1000, and so on). To find db. loss, as where output is less than input, divide input value by output value. Current and voltage ratios in db. can be found similarly, provided the input and output impedances are the same. Power, voltage and current values must be in the same units (watts, millivolts, microamperes, etc.).

ABBREVIATIONS FOR ELECTRICAL AND RADIO TERMS

Alternating current	a.c.	Megohm	MΩ
Ampere (amperes)	a.	Meter	m.
Antenna	ant.	Microfarad	μfd.
Audio frequency	a.f.	Microhenry	μh.
Centimeter	cm.	Micromicrofarad	μμfd.
Continuous waves	c.w.	Microvolt	μv.
Cycles per second	c.p.s.	Microvolt per meter	μv/m.
Decibel	db.	Microwatt	μw.
Direct current	d.c.	Milliampere	ma.
Electromotive force	e.m.f.	Millivolt	mv.
Frequency	f.	Milliwatt	mw.
Ground	gnd.	Modulated continuous waves	m.c.w.
Henry	h.	Ohm	Ω
High frequency	h.f.	Power	P.
Intermediate frequency	i.f.	Power factor	p.f.
Interrupted continuous waves	i.c.w.	Radio frequency	r.f.
Kilocycles (per second)	kc.	Ultra-high frequency	u.h.f.
Kilowatt	kw.	Volt (volts)	v.
Megacycle (per second)	Mc.	Watt (watts)	w.



REACTANCE CHART

By use of this chart, the approximate reactance of any capacity from 1.0 μμfd. to 10 μfd. at any frequency from 100 cycles to 100 megacycles can be read directly. The reactance of inductances from 0.1 microhenry to 10 henries can also be read. Intermediate values can be estimated by interpolation. In making interpolations remember that the rate of change between lines is logarithmic. Use the frequency or reactance scales as a guide in estimating intermediate values on the capacity or inductance scales.

This chart can also be used to find the approximate resonance frequencies of LC combinations or the frequency to which a given coil and condenser will tune. First locate the respective slanting lines for the capacity and inductance. The point where they intersect, i.e., where the reactances are equal, is the resonant frequency (projected downward and read on the frequency scale).

● **Numerical Values**

π =	3.1416	$(2\pi)^2$ =	39.4784	$\frac{1}{\log_{10} \epsilon}$ = 2.3026	$\sqrt{2}$ = 1.4142
$1/\pi$ =	0.3183	$\log_{10} \pi$ =	0.4971	$1/2\pi$ = 0.1592	$\sqrt{3}$ = 1.7321
π^2 =	9.8696	$\log_{10} (\pi/2)$ =	0.1961	$(1/2\pi)^2$ = 0.0253	$1/\sqrt{2}$ = 0.7071
$1/\pi^2$ =	0.1013	$\log_{10} \pi^2$ =	0.9943		$1/\sqrt{3}$ = 0.5773
π^3 =	31.0063	$\log_{10} \sqrt{\pi}$ =	0.2486	Area of Circle = $(\pi/4) \times D^2 = 0.7854\ D^2$	
$1/\pi^3$ =	0.0323	ϵ =	2.7183	Volume of Sphere = $(\pi/6) \times D^3 = 0.5236\ D^3$	
$\sqrt{\pi}$ =	1.7725	$1/\epsilon$ =	0.3679	1 Radian = 57° 17′ 44″ .8	
$1/\sqrt{\pi}$ =	0.5642	ϵ^2 =	7.3890	= 57° 17′ .7468	
$\sqrt{\pi/2}$ =	1.2533	$\sqrt{\epsilon}$ =	1.6487	= 57° .29578	
2π =	6.2832	$\log_{10} \epsilon$ =	0.4343	1° = 0.01745 Radian	

Standard Metal Gauges

Gauge No.	American or B. & S.[1]	U. S. Standard[2]	Birmingham or Stubs[3]
1	.2893	.28125	.300
2	.2576	.265625	.284
3	.2294	.25	.259
4	.2043	.234375	.238
5	.1819	.21875	.220
6	.1620	.203125	.203
7	.1443	.1875	.180
8	.1285	.171875	.165
9	.1144	.15625	.148
10	.1019	.140625	.134
11	.09074	.125	.120
12	.08081	.109375	.109
13	.07196	.09375	.095
14	.06408	.078125	.083
15	.05707	.0703125	.072
16	.05082	.0625	.065
17	.04526	.05625	.058
18	.04030	.05	.049
19	.03589	.04375	.042
20	.03196	.0375	.035
21	.02846	.034375	.032
22	.02535	.03125	.028
23	.02257	.028125	.025
24	.02010	.025	.022
25	.01790	.021875	.020
26	.01594	.01875	.018
27	.01420	.0171875	.016
28	.01264	.015625	.014
29	.01126	.0140625	.013
30	.01003	.0125	.012
31	.008928	.0109375	.010
32	.007950	.01015625	.009
33	.007080	.009375	.008
34	.006350	.00859375	.007
35	.005615	.0078125	.005
36	.005000	.00703125	.004
37	.004453	.006640625	....
38	.003965	.00625	....
39	.003531	.....	....
40	.003145	.....	....

[1] Used for aluminum, copper, brass and non-ferrous alloy sheets, wire and rods.

[2] Used for iron, steel, nickel and ferrous alloy sheets, wire and rods.

[3] Used for seamless tubes; also by some manufacturers for copper and brass.

Decimal Equivalents of Fractions

1/32	.03125	17/32	.53125
1/16	.0625	9/16	.5625
3/32	.09375	19/32	.59375
1/8	.125	5/8	.625
5/32	.15625	21/32	.65625
3/16	.1875	11/16	.6875
7/32	.21875	23/32	.71875
1/4	.25	3/4	.75
9/32	.28125	25/32	.78125
5/16	.3125	13/16	.8125
11/32	.34375	27/32	.84375
3/8	.375	7/8	.875
13/32	.40625	29/32	.90625
7/16	.4375	15/16	.9375
15/32	.46875	31/32	.96875
1/2	.5	1	1.0

Effect of Coil Shields on Inductance

It is well known that enclosing a coil in a shield decreases the inductance of the coil. An easily-applied graphical method of determining the extent of the decrease has been worked out by the Radiotron Division of RCA Manufacturing Company and published as a tube application note.[1]

Considering the shield as a single turn having low resistance compared to its reactance, the following formula for inductance of the coil within the shield can be worked out:

$$L = L_a(1-K^2)$$

where L is the desired inductance, L_a is the inductance of the coil outside the shield, and K^2 is a factor depending upon the geometric dimensions of the coil and shield. Values of K^2 have been plotted as a family of curves in the chart reproduced on the opposite page. The notations are as follows:

b — length of winding of coil
a — radius of coil
A — radius of shield

The curves are sufficiently accurate for all practical purposes throughout the range shown *when the length of the shield is greater than that of the coil by at least the radius of the coil.* If the shield can is square instead of circular, A may be taken as 0.6 times the width of one side. The reduction factor, K^2, is plotted against $b/2a$ (ratio of length to diameter of coil), for a series of values of a/A, the ratio of coil radius to shield radius (or coil diameter to shield diameter).

The following example will illustrate the use of the chart. Assume an r.f. coil 1½ inches long and ¾ inch in diameter to be used in a shield 1¼ inches in diameter. The inductance-reducing effect of the shield is to be calculated. The values are:

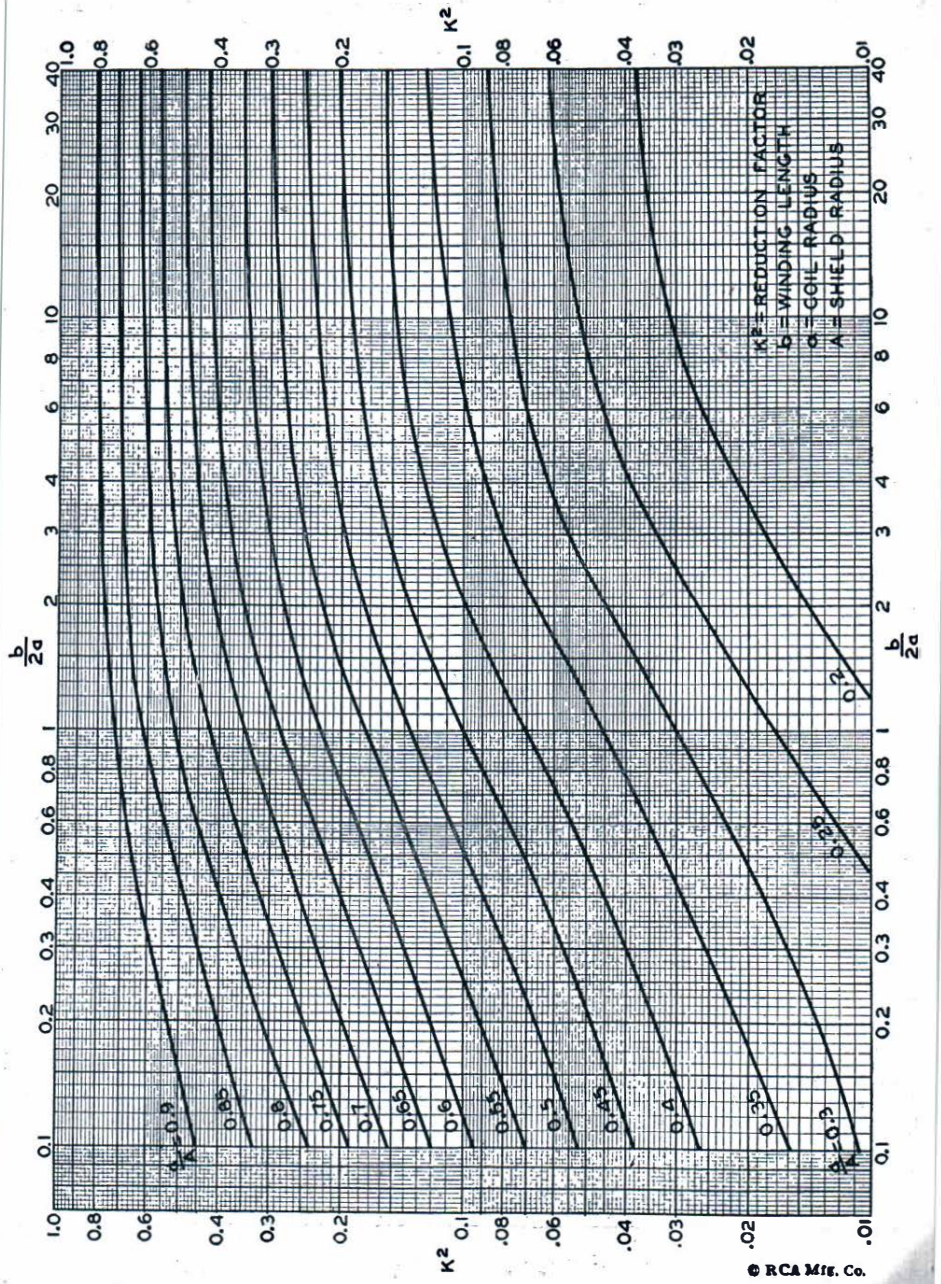
$$b = 1.5$$
$$a = 0.375$$
$$A = 0.625$$
$$b/2a = 1.5/0.75 = 2$$
$$a/A = 0.375/0.625 = 0.6$$

From the curves, K^2 is 0.28; the inductance of the coil is therefore reduced 28% by the shield, or conversely, the inductance of the shield coil is 72% of its unshielded value.

The reduction in inductance does not become serious with coils of $b/2a$ ratios of 2 or less, until the shield diameter becomes less than twice the coil diameter. With an a/A ratio of 0.5, the reduction in inductance will be of the order of 15%.

[1] Application Note No. 48, Copyright, 1935, RCA Manufacturing Co., Inc.

CURVES FOR DETERMINATION OF DECREASE IN INDUCTANCE PRODUCED BY A COIL SHIELD



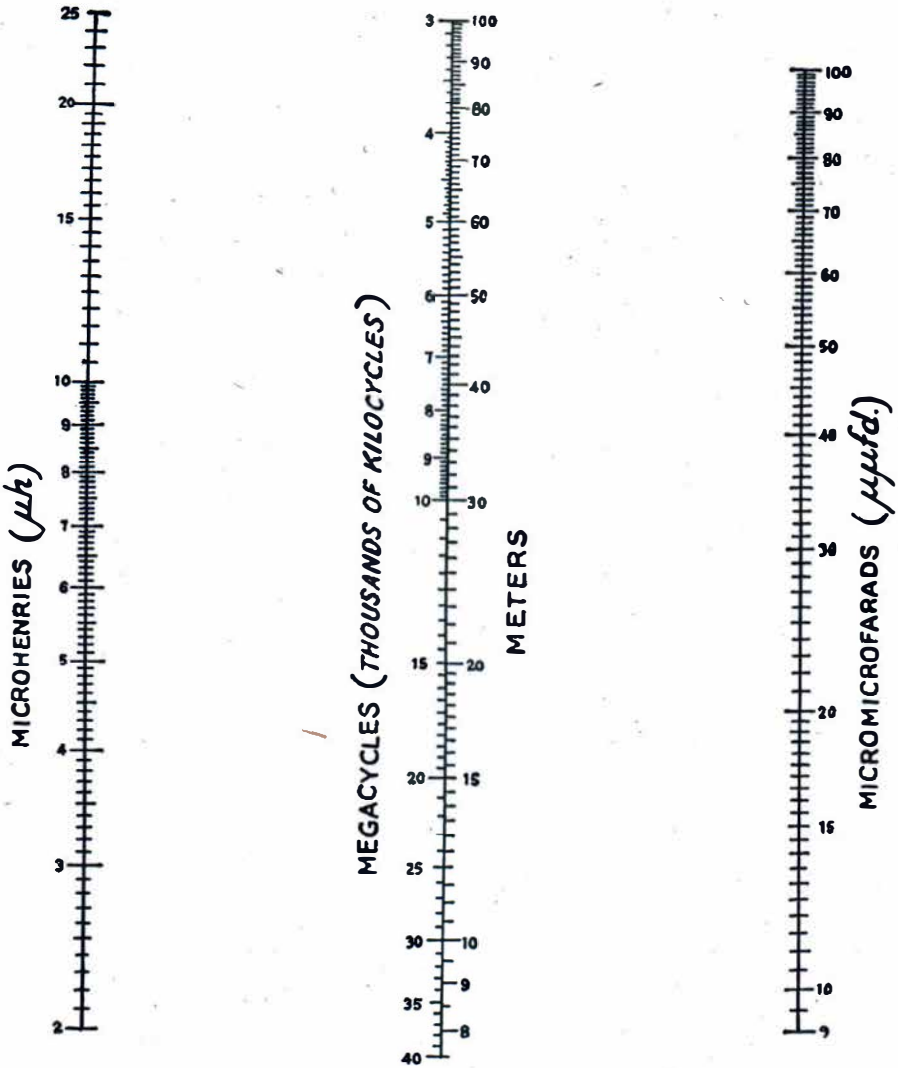
COPPER WIRE TABLE

Gauge No. B. & S.	Diam. in Mils [1]	Circular Mil Area	Turns per Linear Inch [2]				Turns per Square Inch [2]			Feet per Lb.		Ohms per 1000 ft. 25° C.	Current Carrying Capacity at 1500 C.M. per Amp. [3]	Diam. in mm.	Nearest British S.W.G. No.
			Enamel	S.C.C.	D.S.C. or S.C.C.	D.C.C.	S.C.C.	Enamel S.C.C.	D.C.C.	Bare	D.C.C.				
1	289.3	83690	—	—	—	—	—	—	—	3.947	—	.1264	55.7	7.348	1
2	257.6	66370	—	—	—	—	—	—	—	4.977	—	.1593	44.1	6.544	8
3	229.4	52640	—	—	—	—	—	—	—	6.276	—	.2009	35.0	5.827	4
4	204.3	41740	—	—	—	—	—	—	—	7.914	—	.2533	27.7	5.189	5
5	181.9	33100	—	—	—	—	—	—	—	9.980	—	.3195	22.0	4.621	7
6	162.0	26250	—	—	—	—	—	—	—	12.58	—	.4028	17.5	4.115	8
7	144.3	20820	—	—	—	—	—	—	—	15.87	—	.5080	13.8	3.665	9
8	128.5	16510	7.6	—	7.4	7.1	—	—	—	20.01	19.6	.6405	11.0	3.264	10
9	114.4	13090	8.6	—	8.2	7.8	—	—	—	25.23	24.6	.8077	8.7	2.906	11
10	101.9	10380	9.6	—	9.3	8.9	87.5	84.8	80.0	31.82	30.9	1.018	6.9	2.588	12
11	90.74	8234	10.7	—	10.3	9.8	110	105	97.5	40.12	38.8	1.284	5.5	2.305	13
12	80.81	6530	12.0	—	11.5	10.9	136	131	121	50.59	48.9	1.619	4.4	2.053	14
13	71.96	5178	13.5	—	12.8	12.0	170	162	150	63.80	61.5	2.042	3.5	1.828	15
14	64.08	4107	15.0	—	14.2	13.8	211	198	183	80.44	77.3	2.575	2.7	1.628	16
15	57.07	3257	16.8	—	15.8	14.7	262	250	223	101.4	97.3	3.247	2.2	1.450	17
16	50.82	2583	18.9	18.9	17.9	16.4	321	306	271	127.9	119	4.094	1.7	1.291	18
17	45.26	2048	21.2	21.2	19.9	18.1	397	372	329	161.3	150	5.163	1.3	1.150	18
18	40.30	1624	23.6	23.6	22.0	19.8	493	454	399	203.4	188	6.510	1.1	1.024	19
19	35.89	1288	26.4	26.4	24.4	21.8	592	553	479	256.5	237	8.210	.86	.9116	20
20	31.96	1022	29.4	29.4	27.0	23.8	775	725	625	323.4	298	10.35	.68	.8118	21
21	28.46	810.1	33.1	32.7	29.8	26.0	940	895	754	407.8	370	13.05	.54	.7230	22
22	25.35	642.4	37.0	36.5	34.1	30.0	1150	1070	910	514.2	461	16.46	.43	.6438	23
23	22.57	509.5	41.3	40.6	37.6	31.6	1400	1300	1080	648.4	584	20.76	.34	.5733	24
24	20.10	404.0	46.3	35.3	41.5	35.6	1700	1570	1260	817.7	745	26.17	.27	.5106	25
25	17.90	320.4	51.7	50.4	45.6	38.6	2060	1910	1510	1031	903	33.00	.21	.4547	26
26	15.94	254.1	58.0	55.6	50.2	41.8	2500	2300	1750	1300	1118	41.62	.17	.4049	27
27	14.20	201.5	64.9	61.5	55.0	45.0	3030	2780	2020	1639	1422	52.48	.13	.3606	29
28	12.64	159.8	72.7	68.6	60.2	48.5	3670	3350	2310	2067	1759	66.17	.11	.3211	30
29	11.26	126.7	81.6	74.8	65.4	51.8	4300	3900	2700	2607	2207	83.44	.084	.2859	31
30	10.03	100.5	90.5	83.3	71.5	55.5	5040	4660	3020	3287	2534	105.2	.067	.2546	33
31	8.928	79.70	101	92.0	77.5	59.2	5920	5280	—	4145	2768	132.7	.053	.2268	34
32	7.950	63.21	113	101	83.6	62.6	7060	6250	—	5227	3137	167.3	.042	.2019	36
33	7.080	50.13	127	110	90.3	66.3	8120	7360	—	6591	4697	211.0	.033	.1798	37
34	6.305	39.75	143	120	97.0	70.0	9600	8310	—	8310	6168	266.0	.026	.1601	38
35	5.615	31.52	158	132	104	73.5	10900	8700	—	10480	6737	335.0	.021	.1426	38–39
36	5.000	25.00	175	143	111	77.0	12200	10700	—	13210	7877	423.0	.017	.1270	39–40
37	4.453	19.83	198	154	118	80.3	—	—	—	16660	9309	533.4	.013	.1131	41
38	3.965	15.72	224	166	126	83.6	—	—	—	21010	10666	672.6	.010	.1007	42
39	3.531	12.47	248	181	133	86.6	—	—	—	26500	11907	848.1	.008	.0897	43
40	3.145	9.88	282	194	140	89.7	—	—	—	33410	14222	1069	.006	.0799	44

[1] A mil is 1/1000 (one thousandth) of an inch.

[2] The figures given are approximate only, since the thickness of the insulation varies with different manufacturers.

[3] The current-carrying capacity at 1000 C.M. per ampere is equal to the circular-mil area (Column 3) divided by 1000.



RELATION BETWEEN INDUCTANCE, CAPACITY AND FREQUENCY

With this chart and a straight-edge any of the above quantities can be determined if the other two are known. For example, if a condenser has a minimum capacity of 15 μμfd. and a maximum capacity of 50 μμfd., and it is to be used with a coil of 10 μh. inductance, what frequency range will be covered? The straight-edge is connected between 10 on the left-hand scale and 15 on the right, giving 13 mc. as the high-frequency limit. Keeping the straight-edge at 10 on the left-hand scale, the other end is swung to 50 on the right-hand scale, giving a low-frequency limit of 7.1 mc. The tuning range would, therefore, be from 7.1 mc. to 13 mc., or 7100 kc. to 13,000 kc. The center scale also serves to convert frequency to wavelength.

The range of the chart can be extended by multiplying each of the scales by 0.1 or 10. In the example above, if the capacities are 150 and 500 μμfd. and the inductance 100 μh., the range becomes approximately 231 to 422 meters or 0.7 to 1.3 Mc. Alternatively, 1.5 to 5 μμfd. and 1 μh. will give 71 to 130 Mc.

● USE OF LOG TABLES

The accompanying four-place table will give results which are accurate to three significant figures without interpolation and to four figures with interpolation. The use of log tables has been discussed in § 1-6. Only the mantissas are given in the tables; the decimal point (omitted in the tables for convenience) should be placed to the left of the mantissas in every case. The characteristic of any log can be found readily if the number is factored by a power of ten (§ 1-5) with the significant factor assigned a value between 1 and 10. For example:

$$256{,}000{,}000 = 2.56 \times 10^{8} \qquad \log = 8.4082$$

$$0.000000256 = 2.56 \times 10^{-7} \qquad \log = \bar{7}.4082$$

If the number has only two digits it will be found in the "N" column and the corresponding logarithm will be beside it in the "0" column. For example, the mantissa of any number having only two significant figures, such as 5400, is 0.7324; in this example the characteristic is 3 so the com-

FOUR-PLACE

N	0	1	2	3	4	5	6	7	8	9	Proportional Parts 1	2	3	4	5	6	7	8	9
10	0000	0043	0086	0128	0170	0212	0253	0294	0334	0374	*4	8	12	17	21	25	29	33	37
11	0414	0453	0492	0531	0569	0607	0645	0682	0719	0755	4	8	11	15	19	23	26	30	34
12	0792	0828	0864	0899	0934	0969	1004	1038	1072	1106	3	7	10	14	17	21	24	28	31
13	1139	1173	1206	1239	1271	1303	1335	1367	1399	1430	3	6	10	13	16	19	23	26	29
14	1461	1492	1523	1553	1584	1614	1644	1673	1703	1732	3	6	9	12	15	18	21	24	27
15	1761	1790	1818	1847	1875	1903	1931	1959	1987	2014	*3	6	8	11	14	17	20	22	25
16	2041	2068	2095	2122	2148	2175	2201	2227	2253	2279	3	5	8	11	13	16	18	21	24
17	2304	2330	2355	2380	2405	2430	2455	2480	2504	2529	2	5	7	10	12	15	17	20	22
18	2553	2577	2601	2625	2648	2672	2695	2718	2742	2765	2	5	7	9	12	14	16	19	21
19	2788	2810	2833	2856	2878	2900	2923	2945	2967	2989	2	4	7	9	11	13	16	18	20
20	3010	3032	3054	3075	3096	3118	3139	3160	3181	3201	2	4	6	8	11	13	15	17	19
21	3222	3243	3263	3284	3304	3324	3345	3365	3385	3404	2	4	6	8	10	12	14	16	18
22	3424	3444	3464	3483	3502	3522	3541	3560	3579	3598	2	4	6	8	10	12	14	15	17
23	3617	3636	3655	3674	3692	3711	3729	3747	3766	3784	2	4	6	7	9	11	13	15	17
24	3802	3820	3838	3856	3874	3892	3909	3927	3945	3962	2	4	5	7	9	11	12	14	16
25	3979	3997	4014	4031	4048	4065	4082	4099	4116	4133	2	3	5	7	9	10	12	14	15
26	4150	4166	4183	4200	4216	4232	4249	4265	4281	4298	2	3	5	7	8	10	11	13	15
27	4314	4330	4346	4362	4378	4393	4409	4425	4440	4456	2	3	5	6	8	9	11	13	14
28	4472	4487	4502	4518	4533	4548	4564	4579	4594	4609	2	3	5	6	8	9	11	12	14
29	4624	4639	4654	4669	4683	4698	4713	4728	4742	4757	1	3	4	6	7	9	10	12	13
30	4771	4786	4800	4814	4829	4843	4857	4871	4886	4900	1	3	4	6	7	9	10	11	13
31	4914	4928	4942	4955	4969	4983	4997	5011	5024	5038	1	3	4	6	7	8	10	11	12
32	5051	5065	5079	5092	5105	5119	5132	5145	5159	5172	1	3	4	5	7	8	9	11	12
33	5185	5198	5211	5224	5237	5250	5263	5276	5289	5302	1	3	4	5	6	8	9	10	12
34	5315	5328	5340	5353	5366	5378	5391	5403	5416	5428	1	3	4	5	6	8	9	10	11
35	5441	5453	5465	5478	5490	5502	5514	5527	5539	5551	1	2	4	5	6	7	9	10	11
36	5563	5575	5587	5599	5611	5623	5635	5647	5658	5670	1	2	4	5	6	7	8	10	11
37	5682	5694	5705	5717	5729	5740	5752	5763	5775	5786	1	2	3	5	6	7	8	9	10
38	5798	5809	5821	5832	5843	5855	5866	5877	5888	5899	1	2	3	5	6	7	8	9	10
39	5911	5922	5933	5944	5955	5966	5977	5988	5999	6010	1	2	3	4	5	7	8	9	10
40	6021	6031	6042	6053	6064	6075	6085	6096	6107	6117	1	2	3	4	5	6	8	9	10
41	6128	6138	6149	6160	6170	6180	6191	6201	6212	6222	1	2	3	4	5	6	7	8	9
42	6232	6243	6253	6263	6274	6284	6294	6304	6314	6325	1	2	3	4	5	6	7	8	9
43	6335	6345	6355	6365	6375	6385	6395	6405	6415	6425	1	2	3	4	5	6	7	8	9
44	6435	6444	6454	6464	6474	6484	6493	6503	6513	6522	1	2	3	4	5	6	7	8	9
45	6532	6542	6551	6561	6571	6580	6590	6599	6609	6618	1	2	3	4	5	6	7	8	9
46	6628	6637	6646	6656	6665	6675	6684	6693	6702	6712	1	2	3	4	5	6	7	7	8
47	6721	6730	6739	6749	6758	6767	6776	6785	6794	6803	1	2	3	4	5	5	6	7	8
48	6812	6821	6830	6839	6848	6857	6866	6875	6884	6893	1	2	3	4	4	5	6	7	8
49	6902	6911	6920	6928	6937	6946	6955	6964	6972	6981	1	2	3	4	4	5	6	7	8
50	6990	6998	7007	7016	7024	7033	7042	7050	7059	7067	1	2	3	3	4	5	6	7	8
51	7076	7084	7093	7101	7110	7118	7126	7135	7143	7152	1	2	3	3	4	5	6	7	8
52	7160	7168	7177	7185	7193	7202	7210	7218	7226	7235	1	2	2	3	4	5	6	7	7
53	7243	7251	7259	7267	7275	7284	7292	7300	7308	7316	1	2	2	3	4	5	6	6	7
54	7324	7332	7340	7348	7356	7364	7372	7380	7388	7396	1	2	2	3	4	5	6	6	7
N	0	1	2	3	4	5	6	7	8	9	1	2	3	4	5	6	7	8	9

* Interpolation in this section of the table is inaccurate.

plete log is 3.7324. For numbers having three significant figures, locate the first two in the "N" column and move to the right, reading the mantissa under the appropriate digit; e.g., log 5430 = 3.7348, the characteristic being found as before. For numbers having four significant figures, proceed as before and add to the mantissa the figure given on the same line in the "Proportional Parts" column under the appropriate digit. For example, log 5434 = 3.7351.

In determining numbers whose logarithms are given (such numbers are called *antilogarithms* and are written $\log^{-1}$) the procedure given above is reversed. For example, given 1.9304 as the logarithm of a number to be found, inspection of the table shows that the mantissa 0.9304 corresponds to the number 852; the characteristic is 1 so there are two integral places, hence the number is 85.2. If the given log had been 1.9306, there is no exactly corresponding mantissa in the table, but the next lower one is 0.9304. The difference is 2 (0.0002, actually), and on referring to the columns of proportional parts on the same line it is found that a difference of two may indicate either 3 or 4 as the fourth digit in the antilogarithm. Hence the antilogarithm of 1.9306 is either 85.23 or 85.24. This is the maximum uncertainty which will be encountered with four-place tables; in most cases the fourth figure will have no uncertainty.

LOGARITHMS

N	0	1	2	3	4	5	6	7	8	9	Proportional Parts 1	2	3	4	5	6	7	8	9
55	7404	7412	7419	7427	7435	7443	7451	7459	7466	7474	1	2	2	3	4	5	5	6	7
56	7482	7490	7497	7505	7513	7520	7528	7536	7543	7551	1	2	2	3	4	5	5	6	7
57	7559	7566	7574	7582	7589	7597	7604	7612	7619	7627	1	2	2	3	4	5	5	6	7
58	7634	7642	7649	7657	7664	7672	7679	7686	7694	7701	1	1	2	3	4	4	5	6	7
59	7709	7716	7723	7731	7738	7745	7752	7760	7767	7774	1	1	2	3	4	4	5	6	7
60	7782	7789	7796	7803	7810	7818	7825	7832	7839	7846	1	1	2	3	4	4	5	6	6
61	7853	7860	7868	7875	7882	7889	7896	7903	7910	7917	1	1	2	3	4	4	5	6	6
62	7924	7931	7938	7945	7952	7959	7966	7973	7980	7987	1	1	2	3	3	4	5	6	6
63	7993	8000	8007	8014	8021	8028	8035	8041	8048	8055	1	1	2	3	3	4	5	5	6
64	8062	8069	8075	8082	8089	8096	8102	8109	8116	8122	1	1	2	3	3	4	5	5	6
65	8129	8136	8142	8149	8156	8162	8169	8176	8182	8189	1	1	2	3	3	4	5	5	6
66	8195	8202	8209	8215	8222	8228	8235	8241	8248	8254	1	1	2	3	3	4	5	5	6
67	8261	8267	8274	8280	8287	8293	8299	8306	8312	8319	1	1	2	3	3	4	5	5	6
68	8325	8331	8338	8344	8351	8357	8363	8370	8376	8382	1	1	2	3	3	4	4	5	6
69	8388	8395	8401	8407	8414	8420	8426	8432	8439	8445	1	1	2	2	3	4	4	5	6
70	8451	8457	8463	8470	8476	8482	8488	8494	8500	8506	1	1	2	2	3	4	4	5	6
71	8513	8519	8525	8531	8537	8543	8549	8555	8561	8567	1	1	2	2	3	4	4	5	5
72	8573	8579	8585	8591	8597	8603	8609	8615	8621	8627	1	1	2	2	3	4	4	5	5
73	8633	8639	8645	8651	8657	8663	8669	8675	8681	8686	1	1	2	2	3	4	4	5	5
74	8692	8698	8704	8710	8716	8722	8727	8733	8739	8745	1	1	2	2	3	4	4	5	5
75	8751	8756	8762	8768	8774	8779	8785	8791	8797	8802	1	1	2	2	3	3	4	5	5
76	8808	8814	8820	8825	8831	8837	8842	8848	8854	8859	1	1	2	2	3	3	4	5	5
77	8865	8871	8876	8882	8887	8893	8899	8904	8910	8915	1	1	2	2	3	3	4	4	5
78	8921	8927	8932	8938	8943	8949	8954	8960	8965	8971	1	1	2	2	3	3	4	4	5
79	8976	8982	8987	8993	8998	9004	9009	9015	9020	9025	1	1	2	2	3	3	4	4	5
80	9031	9036	9042	9047	9053	9058	9063	9069	9074	9079	1	1	2	2	3	3	4	4	5
81	9085	9090	9096	9101	9106	9112	9117	9122	9128	9133	1	1	2	2	3	3	4	4	5
82	9138	9143	9149	9154	9159	9165	9170	9175	9180	9186	1	1	2	2	3	3	4	4	5
83	9191	9196	9201	9206	9212	9217	9222	9227	9232	9238	1	1	2	2	3	3	4	4	5
84	9243	9248	9253	9258	9263	9269	9274	9279	9284	9289	1	1	2	2	3	3	4	4	5
85	9294	9299	9304	9309	9315	9320	9325	9330	9335	9340	1	1	2	2	3	3	4	4	5
86	9345	9350	9355	9360	9365	9370	9375	9380	9385	9390	1	1	2	2	3	3	4	4	5
87	9395	9400	9405	9410	9415	9420	9425	9430	9435	9440	0	1	1	2	2	3	3	4	4
88	9445	9450	9455	9460	9465	9469	9474	9479	9484	9489	0	1	1	2	2	3	3	4	4
89	9494	9499	9504	9509	9513	9518	9523	9528	9533	9538	0	1	1	2	2	3	3	4	4
90	9542	9547	9552	9557	9562	9566	9571	9576	9581	9586	0	1	1	2	2	3	3	4	4
91	9590	9595	9600	9605	9609	9614	9619	9624	9628	9633	0	1	1	2	2	3	3	4	4
92	9638	9643	9647	9652	9657	9661	9666	9671	9675	9680	0	1	1	2	2	3	3	4	4
93	9685	9689	9694	9699	9703	9708	9713	9717	9722	9727	0	1	1	2	2	3	3	4	4
94	9731	9736	9741	9745	9750	9754	9759	9763	9768	9773	0	1	1	2	2	3	3	4	4
95	9777	9782	9786	9791	9795	9800	9805	9809	9814	9818	0	1	1	2	2	3	3	4	4
96	9823	9827	9832	9836	9841	9845	9850	9854	9859	9863	0	1	1	2	2	3	3	4	4
97	9868	9872	9877	9881	9886	9890	9894	9899	9903	9908	0	1	1	2	2	3	3	4	4
98	9912	9917	9921	9926	9930	9934	9939	9943	9948	9952	0	1	1	2	2	3	3	4	4
99	9956	9961	9965	9969	9974	9978	9983	9987	9991	9996	0	1	1	2	2	3	3	3	4
N	0	1	2	3	4	5	6	7	8	9	1	2	3	4	5	6	7	8	9

GOOD BOOKS

Every amateur should maintain a carefully selected bookshelf; a few good books, consistently read and consulted, will add immeasurably to the interest and knowledge of the owner. We suggest a selection among the following works, all of which have been carefully chosen and are recommended in their various fields.

Fundamentals:

Audel's New Radioman's Guide, by E. P. Anderson. For one who wants to get a working knowledge of radio. Not an engineering text, but filled with useful information for the practicing radioman. 765 pages, 519 illustrations. 1940. Audel, $4.00.

Fundamentals of Radio, by F. E. Terman. An elementary version of the author's "Radio Engineering" with simplified treatment intended for readers of limited mathematical ability. 458 pages, 278 illustrations. 1938. McGraw-Hill, $3.75.

Getting Acquainted With Radio, by Alfred Morgan. Gives the neophyte in radio the basic principles of the science in popular language. 285 pages, 130 illustrations. 1940. Century, $2.50.

Principles of Radio, by Keith Henney. This popular book covers the range from the fundamentals of electricity to modern concepts of modulation and detection. 495 pages, 311 illustrations. 3rd edition. 1938. McGraw-Hill, $3.50.

Theory and Engineering:

Electrical Communication, by Arthur L. Albert. General treatment of the whole field of electrical communications, both wire (telegraphy and telephony) and wireless (radio). 534 pages, 397 illustrations. 2nd edition, 1940. Wiley, $5.00.

Electrical Engineers Handbook: Communication and Electronics, by Pender and McIlwain. Engineering reference book covering all phases of communication and electronics. 1022 pages, 981 illustrations. 1936. Wiley, $5.00.

Radio Engineering, by F. E. Terman. A comprehensive treatment covering all phases of radio. The recognized authority in its field. 813 pages, 475 illustrations. 2nd edition, 1937. McGraw-Hill, $5.50.

Radio Engineering Handbook, Keith Henney, Editor. An authoritative handbook for radio engineers, with technical data on all aspects of radio. 945 pages, 837 illustrations. 1941. McGraw-Hill, $5.00.

Principles of Electron Tubes, by H. J. Reich. The theory, characteristics and applications of electron tubes and their circuits. Includes data not elsewhere available. 397 pages, illustrated. 1941. McGraw-Hill, $3.50.

Experiments and Measurements:

Measurements in Radio Engineering, by F. E. Terman. A comprehensive discussion of measurement problems encountered in engineering practice, with emphasis on basic principles. 400 pages, 208 illustrations. 1935. McGraw-Hill, $4.00.

Radio Frequency Electrical Measurements, by H. A. Brown. A laboratory course in r.f. measurements for communications students. Contains practical information on methods. 384 pages, 177 illustrations. 2nd edition, 1938. McGraw-Hill, $4.00.

The Cathode-Ray Tube at Work, by John F. Rider. Cathode-ray tube theory, sweep circuits, a.c. wave patterns and oscilloscopes, including actual photographs of screen patterns. 322 pages, 444 illustrations. 1935. Rider, $2.50.

Commercial Operating and Equipment:

Radio Operating Questions and Answers, by Nilson and Hornung. Gives answers to questions in the FCC study guide covering all six elements of the commercial examinations. 415 pages, 87 illustrations. 7th edition, 1940. McGraw-Hill, $2.50.

The Radio Manual, by G. E. Sterling. A practical handbook, especially valuable to the commercial and broadcast operator, covering principles, methods and apparatus. Illustrated. 1120 pages, 1938. D. Van Nostrand, $6.00.

Books Dealing with Specialized Radio Topics:

Aeronautic Radio, by Myron F. Eddy. Supplies the information needed by students, pilots, mechanics, operators and executives. Prepares for aviation radio license exams. 502 pages, 198 illustrations. 1939. Ronald, $4.50.

Frequency Modulation, by John F. Rider. Practical exposition of the subject from the serviceman's standpoint, but including worthwhile theoretical background. 136 pages, illustrated. 1940. Rider, $1.00.

Mathematics for Electricians and Radiomen, by Nelson M. Cooke. Furnishes the student with a sound mathematical foundation and shows how to apply this knowledge to practical problems. 604 pages, illustrated. 1942. McGraw-Hill, $4.00.

Principles of Television Engineering, by D. G. Fink. Covers the television system from the camera to the viewing screen in the receiver, with descriptions of modern equipment. 541 pages, 313 illustrations. 1940. McGraw-Hill, $5.00.

Radio as a Career, by J. L. Hornung. A realistic discussion of the opportunities to be found in the various radio fields and the relationship of the radio amateur to these fields. 212 pages. 1940. Funk & Wagnalls, $1.50.

Servicing Superheterodynes, by John F. Rider. Theory and practice of superheterodynes, with adjustment and trouble-shooting data. 278 pages. Rider, $1.00.

Publishers addresses:

D. Appleton-Century Co., 35 West 32nd St., New York.
D. Van Nostrand Co., 250 Fourth Ave., New York.
Funk & Wagnalls Co., 354 Fourth Ave., New York.
John F. Rider, 404 Fourth Ave., New York.
McGraw-Hill Book Co., 330 West 42nd St., New York.
Ronald Press, 15 East 26th St., New York.
Theo. F. Audel & Co., 49 West 23rd St., New York.
John F. Wiley & Sons, 440 Fourth Ave., New York.

Index

PAGE

A Battery ... 42
A.C. ... 28–30
A.C.-D.C. Power Supplies ... 152
AT-Cut Crystals ... 61
Abbreviations, Radio and Electrical ... 272
Abscissae ... 18
Absorption ... 157
Absorption Frequency Meters ... 225–227
Acorn Tubes ... 121
Adjacent-Channel Selectivity ... 112
Air-Gap Crystal Holder ... 61–62
Alignment, Receiver ... 137–138, 141–142, 243–244
Alternating Current ... 28–30
Alternating Current, Ohm's Law for ... 30–31
Alternating Current Spectrum ... 30
Ammeters ... 23, 32, 233, 235
Ampere ... 23
Ampere Turns ... 25
Amplication Factor ... 43
Amplification, Vacuum Tube ... 43, 44–46
Amplifier Classifications ... 46–48
Amplifiers (*see* "Receivers," "Transmitters" and "Radiotelephony")
Amplitude Modulation ... 84
Analyzers, Wave ... 242–243
Angle, Amplifier Operating ... 71, 79–80
Angle of Radiation ... 158, 162, 164
Angular Velocity ... 29
Anode ... 23
Anode-Dissipation Power Measurement ... 246
Antenna Systems:
  Angle of Radiation ... 158, 162, 164
  Beam Antennas ... 184–194
  Bent Antennas ... 184
  Broadside Arrays ... 187, 188
  Coaxial Vertical Radiator ... 193–194
  Collinear Arrays ... 187–188
  Compact Antennas ... 183–184
  Conductor Size ... 163
  "Corner" Reflector Antennas ... 194
  Coupling to Receiver ... 116
  Coupling to Transmitter ... 169–172
  Current Distribution ... 163, 179
  Directive Antennas ... 184–194
  Directivity ... 162
  Director ... 190
  Doublet ... 162
  Driven Elements ... 187
  Dummy Antennas ... 76
  End Effect ... 162
  End-Fire Arrays ... 187, 188–189
  Feed Systems:
    Balance to Ground ... 167
    Characteristic Impedance ... 167
    Concentric Lines ... 166, 176–177
    Current ... 166
    Delta Matching Transformer ... 177
    Harmonic Resonance ... 39
    Height ... 163–164
    Impedance ... 40, 167, 168
    Length ... 167–168, 189
    Losses ... 168–169
    Linear Transformers ... 168, 178–179
    Non-Resonant Lines ... 169, 175–179
    Open-Wire Line ... 166, 172
    "Q" Antenna ... 177–178
    Reactance ... 168
    Resistance ... 168
    Resonant Lines ... 169, 172–175
    Single-Wire Feed ... 166, 175
    Standing-Wave Ratio ... 168
    Stub Matching ... 178–179
    Tuned Transmission Lines ... 172–175
    Twisted-Pair Line ... 166, 175–176
    Untuned Transmission Lines ... 171–179
    Voltage ... 166
  Folded Dipole ... 194
  Ground Effect ... 164–165
  Grounded Antennas ... 193
  Hertz Antenna ... 162
  Impedance ... 162, 163, 164, 181
  J Antenna ... 162
  Length ... 162–163, 174–175, 179, 191, 193
  Long-Wire Antennas ... 179–182, 184–186
  Marconi (Grounded) Antennas ... 193
  Measurements ... 246–248
  Multi-Band Antennas ... 182–183
  Parasitic Elements ... 190–193
  Patterns, Radiation ... 165
  Phased Arrays ... 186–190
  Polarization ... 156, 162, 164–165
  Power Gain ... 162, 181
  Radiation Characteristics ... 163
  Radiation Resistance ... 40
  Receiving ... 194–195
  Reflection, Ground ... 163–164
  Reflector ... 190
  Rhombic or Diamond Antennas ... 185–186
  Rotary-Beam Antennas ... 191
  Standing Waves ... 39
  Transmission Lines ... 166–169
  V Antennas ... 184–185
  Voltage Distribution ... 163, 179
Anti-Nodes ... 39
Arc-Back ... 144
Arrays, Antenna ... 186–195
Atoms ... 21
Atmospheric Bending ... 160, 161
Attraction ... 22
Audio Frequencies ... 30
Audio Frequency Measurement ... 232–233
Audio Image ... 129
Audio Limiting ... 97, 134
Audio Oscillators ... 245, 260–261, 264–265
Autodyne Reception ... 111
Automatic Volume Control ... 131–133
Auto-Transformer ... 33
Average Current Value ... 29
Axes, Cathode-Ray Tube ... 55–56

B Battery ... 42
Back-Wave ... 106
Baffle Shield ... 37
Balanced Transmission Line ... 167
Balancing Circuits, Meter ... 236
Band Changing ... 121
Band-Pass Filters ... 38
Band Spread ... 121–122
Band Width ... 112
Bands, Antenna Lengths for ... 176, 183, 184, 195
Base (of Logarithm) ... 15
Batteries ... 24
Beam Antennas (*see* "Antenna Systems")
Beam Tubes ... 49
Beat Frequencies ... 40–41
Beat Frequency Audio Oscillator ... 243, 245
Beat Note ... 111
Beat Oscillator ... 124, 131
Bent Antennas ... 184
Bias ... 43, 44, 52, 92
Bias Calculation ... 50, 71–72
Bias Modulation ... 88–89
Bias Supplies ... 150–152
Birdies ... 124, 138–139
Bleeder ... 145
Blocked-Grid Keying ... 108
Blocking ... 118
Blocking Condensers ... 41
Body Capacity ... 67, 116
Books ... 280
Brackets ... 11–12
Break-In ... 107
Bridge Circuits ... 38–39, 232, 239–240
Bridge Rectifiers ... 145
Broadside Arrays ... 188
Buffer Amplifier ... 58
Buffer Condenser ... 154
Bus Wire ... 252
Button, Microphone ... 93
Buzzer Code Practice Set ... 260
By-Pass Condensers ... 41, 119–120

C-L Ratios ... 35, 53, 72–73
C. W. Reception ... 111, 116–117, 136–137
Calorimeter Method of Power Measurement ... 246
Cancelling Out Terms ... 12
Capacitive Coupling ... 36, 66
Capacitive Reactance ... 31

PAGE

Capacity 22–23
Capacity, Distributed 31–32
Capacity, Inductance and Frequency Chart 277
Capacity, Interelectrode 43
Capacity Measurement 239–240, 241
Capacity of Condenser, Computing 268
Capacity-Resistance Time Constant 27–28
Carbon Microphones 93
Carrier 58, 84
Cascade Amplifiers 44
Cathode 23, 24, 42, 49–50
Cathode Bias 50
Cathode By-Pass 50
Cathode Keying 108–109
Cathode Modulation 86, 89–90
Cathode-Ray Oscilloscopes 55–57, 98–101, 236–238
Cathode-Ray Tubes 54–55
Cell 24
Center-Tap, Filament 50
Center-Tap Full-Wave Rectifier 144–145
Center-Tap Keying 108–109
Center-Tap Modulation 89–91
Channel 85–86
Characteristic Curves 42
Characteristic Impedance 167
Characteristic (of Logarithm) 16
Charges, Electrical 21–23
Charts and Tables:
- Abbreviations for Electrical and Radio Terms 272
- Antenna and Feeder Lengths 176, 183, 184, 195
- Antenna Gain 187, 188, 190
- Band-Width, Typical I. F. 129
- Cathode Modulation Performance Curves 90
- Characteristic Impedance 167, 180
- Coil Shields, Effect on Inductance 274–275
- Coil-Winding Data 74
- Color Code for Radio Parts 269
- Conductivity of Metals 270
- Continental Code 255–257
- Conversion Factors 271
- Current Capacity of Power Wiring 271
- Decibel Chart 272
- Decimal Equivalents 274
- Dielectric Constants 268–269
- Drill Sizes 251
- Electrical Symbols 270
- Gain of Directive Antennas 187, 188, 190
- Gauges, Metal 274
- Greek Alphabet 271
- Inductance, Capacity and Frequency Chart 277
- LC Constants 270
- Length, Units of 270
- Line-of-Sight U. H. F. Range 160
- Log Tables 278–279
- Metric Prefixes 271
- Multiples and Sub-Multiples 271
- Numerical Values 273
- Q Code 266–267
- Radiation Angles 164, 165
- Radiation Patterns 182
- Radiation Resistance 164, 181
- Reactance Chart 273
- Resistance-Coupled Amplifier Data 97
- Rhombic Antenna Design 186
- Schematic Symbols ..... *Frontispiece*
- Surge Impedance 167, 180
- Tank Circuit Capacity 73
- Tap Sizes 249
- Tolerances (Color Code) 269
- Tools 249
- V-Antenna Design Chart 185
- Vacuum Tube Symbols 270
- Voltage Ratings (Color Code) 269
- Wire Table 269

Chassis Layout 250
Chirp, Keying 107, 110
Choke Coil 41
Choke-Coupled Modulator 87–88
Choke, Filter 147
Choke-Input Filter 145, 146–148
Cipher Groups 263
Circuit Diagram Symbols ..... *Frontispiece*
Circuit, Electric 21
Circuits, Receiving (*see* "Receivers")
Circuits, Transmitting (*see* "Transmitters")
Circuits, Tuned 33–35
Class A, AB, B, and C Amplifiers 46–48
Class B Modulators 91–92, 219–220
Clicks, Keying 106–107, 110
Closed Circuit 21
Coaxial Line 166, 176–177
Coaxial Vertical Radiators 193–194

PAGE

Code Learning:
- Audio Oscillators 260–261, 264–265
- Buzzer Practice Set 260
- Cipher Groups 263
- Code Table 261–265
- High-Speed Operation 259–260
- Practice Cipher Groups 263
- Practice Words 262
- Receiving 254–257
- Sending 257–259
- Sounds 254–257
- Tone Sources 260–265

Code Table 261–265
Code, Q 266–267
Code Practice Sets 260–265
Coefficient 11
Coefficient of Coupling 36
Coefficient, Temperature 61
Coil Shields, Effect on Inductance 274
Coils (*see* "Inductance")
Coils, Winding 253
Collecting Terms 12
Collinear Arrays 187–188
Color Codes, RMA 269
Colpitts Circuit 51
Combined A.C. and D.C. 40
Complex Waves 40
Concentric Line Circuits 40, 81–82
Concentric Transmission Line 166, 176–177
Condenser 22–23
Condenser Capacitance (Computing) 268
Condenser Color Code 269
Condenser-Input Filter 145–146
Condenser Microphones 94
Condenser Series and Parallel Connections 27
Condenser Reactance 31
Condensers, Electrolytic 145
Condensers, Testing 239–240
Condensers, Voltage Rating 74, 145
Conductance 43
Conduction 23
Conduction, Electrolytic 24
Conduction, Gaseous 23
Conduction, Thermionic 24
Conductivity of Metals 270
Conductors 21
Constant, Time 27–28
Continental Code 255–257
Control Grid 42
Controlled-Rectifier Keying 108
Conversion Efficiency 125
Conversion Factors 271
Converters, Frequency 124–126
Coördinates 18
Copper-Oxide Rectifier Meter 234
Core 25
Corner Reflector Antenna 194
Coupled Circuits 35–39
Coupling 35–37
Coupling, Antenna to Receiver 116
Coupling, Antenna to Transmitter 169–172
Coupling Condenser 36
Coupling, Interstage 166
Crackle Finish 252
Critical Angle 159
Critical Coupling 36–37
Critical Frequency 157
Critical Inductance 147
Cross-Modulation 47
Crystal Detector 225
Crystal, Failure to Oscillate 63
Crystal Filters 129–131, 136–137
Crystal Microphones 93–94
Crystal Oscillators 52, 62–65
Crystal-Controlled Transmitter Construction (*see* "Transmitters" *and* "Ultra-High Frequencies")
Crystals, Piezo-Electric 35, 61–62
Current, Alternating 28–30
Current Capacity of Wiring (Table) 271
Current, Direct 23–24
Current Feed for Antennas 166
Current Flow 21, 23–24
Current in A.C. Circuits 28
Current Lag and Lead 30
Current Measurement 28, 32, 233–235
Cut-Off Frequency 38
Cut-Off, Plate Current 43
Cycle 28

D'Arsonval Movement 233
Damping 34

PAGE

Dead Spots ... 116
Decibel ... 18, 46
Decimals ... 7–9
Decoupling ... 96
Decrement ... 34
Deflection Plates ... 54
Deflection Sensitivity ... 54
Degeneration ... 45–46
Delayed A. V. C. ... 132
Delta Matching Transformer ... 177
Demodulation ... 111
Detection ... 111, 112–118, 131
Deviation Ratio ... 103
Deviation-Reading Frequency Meter ... 233
Diagrams, Schematic Symbols for ... *Frontispiece*
Diamond Antenna ... 185–186
Dielectric ... 22, 23
Dielectric Constant ... 22
Dielectric Constants (Table) ... 268–269
Dies ... 251
Diffraction ... 156, 160–161
Diode Detectors ... 113
Diode Voltmeters ... 235–236
Diodes ... 42
Dipole, Folded ... 194
Direct Coupling ... 36
Direct Current ... 23
Direct Feed for Antennas ... 166
Direct-Ray Transmission ... 159–160
Direction of Current Flow ... 24
Directional Antennas (*see* "Antenna Systems")
Director, Antenna ... 190
Disassociation ... 23
Discriminator ... 140–141
Dissipation, Plate ... 44
Distortion ... 44
Distributed Inductance and Capacitance ... 31–32
Dividers, Voltage ... 27
Doubler, Frequency ... 58, 79–80
Double Resonance ... 77
Double Superheterodyne ... 124
Doublet Antenna ... 162
Downward Modulation ... 101–102
Drift, Frequency ... 53, 61, 228
Drill Sizes (Table) ... 251
Drilling Holes ... 250–251
Driven Elements ... 187
Driver ... 66, 91
Driving Power ... 45, 72, 91–92
Drop, Voltage ... 27, 143
Dry Cell ... 24
Dummy Antennas ... 76, 243
Duplex-Diode Triodes and Pentodes ... 49
Duplex Power Supplies ... 153
Dynamic Characteristics ... 43
Dynamic Instability ... 52–53
Dynamic Microphones ... 94
Dynamotors ... 222

E ... 26
E-Layer ... 158, 161
Eddy Current ... 37
Effective Current Value ... 29
Efficiency, Amplifier ... 71
Efficiency, Transformer ... 32
Electric Charges ... 21–23
Electric Circuit ... 21
Electrical Length ... 167–168
Electrical Units and Symbols ... 270, 272
Electricity ... 21
Electricity, Static ... 21–23
Electrolytic Condensers ... 45
Electrolytic Condensers, Testing ... 239–240
Electrolytic Conduction ... 23–24
Electromagnetic Deflection ... 54
Electromagnetic Radiation ... 40
Electromagnetism ... 24–26
Electromotive Force (E.M.F.) ... 22
Electrons ... 21
Electron-Coupled Oscillator ... 59–60, 61
Electron Flow ... 23
Electron Gun ... 54
Electron-Ray Tubes ... 132
Electronic Conduction ... 24
Electrostatic Coupling ... 36
Electrostatic Deflection ... 54
Electrostatic Field ... 22
Electrostatic Shield ... 76
Element, Antenna ... 186–187
Emission, Electron ... 24
End Effect ... 162
End-Fire Arrays ... 187, 188–189

PAGE

Envelope, Modulation ... 84
Excitation ... 44, 45, 52, 74–75
Exciter ... 58
Exciter Units (*see* "Transmitters")
Exponential Functions ... 17–18
Exponents ... 9, 14–15

F-Layers ... 158
Factor ... 9
Fadeouts ... 158
Fading ... 159
Farad ... 22
Faraday Shield ... 76
Feed, Series and Parallel ... 52
Feed-Through Insulators ... 252
Feedback ... 45–46, 51–52
Feeders and Feed Systems (*see* "Antenna Systems")
Fidelity ... 112
Fidelity, Measurement of ... 243
Field Intensity Meters ... 247–248
Field, Electrostatic ... 22
Field, Magnetic ... 24–26
Field, Strength ... 156
Field-Strength Measuring Sets ... 246
Filament ... 24, 49–50
Filament Supply ... 143
Filter, Crystal ... 129–131
Filters ... 38, 145–148
Filters, R.F. ... 109
Final Amplifier ... 58
First Detector ... 124–126
Five Meters (*see* "Ultra High Frequencies")
Flat Response ... 93
Flexible Couplings ... 253
Flow, Current ... 24
Fluorescent Screen ... 54
Flux Density, Magnetic ... 26
Fly-Back ... 56
Flywheel Effect ... 34
Focusing Electrode ... 54
Folded Dipole ... 194
Force, Electromotive ... 22
Force, Lines of ... 22, 24–25
Formulas:
- A.C. Average, Effective and Peak Values ... 29
- Amplification Factor ... 43
- Antenna Impedance Measurement ... 247
- Antenna Length ... 162, 163, 179, 193
- Antenna Power Measurement ... 247
- Antenna Resistance Measurement ... 247
- Bias ... 50
- Bias Supply Bleeder ... 150
- Bridge Balance ... 38, 39
- Capacitive Reactance ... 31
- Capacity of Condenser ... 268
- Cathode Bias ... 50
- Characteristic Impedance ... 167
- Coefficient of Coupling ... 36
- Combined A.C. and D.C. ... 40
- Complex Wave ... 40
- Coupling Transformer Turns Ratio ... 91, 92
- Critical Inductance ... 147
- Crystal Frequency-Thickness Ratio ... 61
- Decibel ... 18, 46
- Delta Matching Transformer Design ... 177
- Filter Design ... 147–148
- Frequency ... 33
- Grid Impedance ... 72
- Grid Leak Bias ... 50
- Impedance ... 31
- Impedance Matching ... 33, 86, 168, 178
- Impedance Ratios ... 32
- Impedance Transformation ... 168
- Inductance, Apparent ... 240
- Inductance Calculation ... 268
- Inductance, True ... 241
- Inductive Reactance ... 31
- Input Filter Choke ... 147
- Interpolation, Frequency ... 231
- Lecher Wires ... 227
- Linear Matching Sections ... 168
- Modulation Impedance ... 86
- Modulation Percentage ... 84, 99
- Modulator Transformers, Turns Ratio ... 91, 92
- Multipliers (Meter) ... 233
- Mutual Conductance ... 43
- Ohmmeter (from voltmeter) ... 238
- Ohm's Law (A.C.) ... 31
- Ohm's Law (D.C.) ... 26
- Parallel Impedance ... 35
- Piano Frequencies ... 232
- Plate Resistance ... 43

Index

PAGE

Power ... 26
Power Factor ... 31
Power Supply Output Voltage ... 147
Power Supply Tranaformer Voltage ... 148
Q ... 35, 241, 242
Q Antenna ... 178
R/C Time Constant ... 27
Reactance ... 31
Regulation ... 143
Resonance ... 33
Resonant Impedance ... 35
Rhombic Antenna ... 186
Ripple ... 146, 147
Series, Parallel and Series-Parallel Capacities ... 27
Series, Parallel and Series-Parallel Inductance ... 27
Series, Parallel and Series-Parallel Resistances ... 27
Shunts, Meter ... 233
Standing-Wave Ratio ... 168
Surge Impedance ... 167
Time Constant ... 27–28
Transformer Volt-Ampere Rating ... 148
Transmission Line Length ... 168, 178, 189
Transmission Line Spacing ... 167
Turns Ratio ... 32
U.H.F. Range ... 160
Voltage Dividers ... 27, 152
Voltage Regulation ... 143
Voltage Regulator Limiting Resistor ... 149
Wavelength ... 33
Wavelength-Frequency Conversion ... 39
Wien Bridge ... 232
Formulas, Explanation ... 10–14
Frequency ... 28–29, 39
Frequency Converters ... 124–126
Frequency Deviation ... 103
Frequency Distortion ... 44
Frequency Divider ... 53
Frequency Drift ... 53, 61, 228
Frequency, Inductance and Capacity Chart ... 277
Frequency Measurement:
Absorption Frequency Meters ... 225–227
Audio Frequencies ... 232–233
Frequency Standards ... 229–231
Heterodyne Frequency Meters ... 227–229
Interpolation ... 231
Lecher Wires ... 227
Monitoring ... 233
Zero-Beat Indicators ... 231–232
Frequency Modulation ... 84
Deviation Ratio ... 103
Discriminator ... 140–141
Limiter ... 139–140
Methods ... 104–105
Principles ... 102–104
Reactance Modulator ... 104
Reception ... 139–142
Frequency Multipliers ... 58, 79–80
Frequency Response ... 44, 93
Frequency Spectrum ... 30
Frequency Stability ... 52–53, 107, 228
Frequency Standards ... 229–231
Frequency-Wavelength Conversion ... 39
Full-Wave Rectifiers ... 144–145
Functions (Mathematical) ... 16–18
Fundamental Frequency ... 29, 39
Fundamental Radio System ... 21

Gain Control ... 96–97, 120
Galvanometers ... 28
Ganged Tuning ... 122
Gaseous Conduction ... 23
Gaseous Regulator Tubes ... 148–149
Genemotors ... 222
General Coverage Tuning ... 122
Graphs ... 18–20
Grid ... 42
Grid Bias ... 43, 44, 50, 52, 71–72
Grid-Bias Modulation ... 86, 88–89
Grid-Control Rectifiers ... 49, 108
Grid Current ... 43
Grid Driving Power ... 45
Grid Emission ... 77
Grid Excitation ... 44, 45
Grid Impedance ... 72
Grid Keying ... 108
Grid Leak ... 50
Grid Leak Detectors ... 113–114
Grid Neutralization ... 67, 68
Grid-Plate Oscillator ... 64
Grid-Plate Transconductance ... 43
Grid Swing ... 46
Ground Effect ... 163–165

PAGE

Ground Potential ... 41
Ground Wave ... 156–157
Grounded Antennas ... 193
Grounds ... 41
Gun, Electron ... 54

Hairpin Coupling Loop ... 82
Half-Wave ... 39
Half-Wave Antenna ... 162–163
Half-Wave Rectifiers ... 144
Hand Capacity ... 67
Harmonic Distortion ... 44
Harmonic Frequencies ... 29, 39
Harmonic Generation ... 79–80
Harmonic Operation of Antennas ... 39, 179
Harmonic Resonance ... 39
Harmonic Suppression ... 76–77, 172
Hartley Circuit ... 51, 59
Hash ... 154
Headphone Impedance ... 118
Heater ... 50
Heating Effect ... 26
Henry ... 25
Hertz Antenna ... 162
Heterodyne Frequency Meters ... 227–229
Heterodyne Reception ... 111
High-Frequency Oscillator (Receiver) ... 123–124, 126–127
High-Pass Filters ... 38
High Speed Code Operation ... 259–260
High-Vacuum Rectifiers ... 143
Hiss Noise ... 120, 134
Holes, Drilling and Cutting ... 250–251
Hook-Up Wire ... 252
Hum ... 98, 102, 116

I ... 26
Image, A.F. ... 129
Image, R.F. ... 124
Image Ratio ... 124, 133
Impedance ... 31, 34–35
Impedance, Antenna ... 162, 163, 164, 181
Impedance, Grid ... 72
Impedance Matching ... 33, 168
Impedance Ratio, Transformer ... 32–33
Impedance, Tank ... 60, 71, 72–73
Impulse Noise ... 134
Index (of a Root) ... 9
Indicators, Tuning ... 132–133
Inductance ... 25
Inductance Calculation ... 268
Inductance, capacity and Frequency Chart ... 277
Inductance, Distributed ... 31–32
Inductance Measurement ... 239–241
Inductance Winding Charts ... 74
Induction ... 25
Inductive Coupling ... 36
Inductive Neutralization ... 68–69
Inductive Reactance ... 31
Infinite Impedance Detector ... 114–115
Input Capacity, Tube ... 43, 45
Input Choke ... 147
Input Resistance, Tube ... 121
Instability ... 52–53
Instantaneous Plate Current ... 144
Instruments:
A.C. ... 32, 234–235
Ammeters ... 28, 233–235
Audio Oscillators ... 245
Capacity Bridges ... 39, 239–240
D.C. ... 28, 233–234
Field-Intensity Meters ... 247–248
Frequency Meters ... 225–229
Frequency Standards ... 229–231
I.F. Test Oscillator ... 244–245
Inductance Bridge ... 239
L. C and Q Meter ... 240–242
Multi-Range V-O-M ... 234
Ohmmeters ... 238–239
Oscilloscopes ... 55–57, 98–101, 236–238
Reactance Meter ... 239
Resistance Bridges ... 239
Vacuum-Tube Voltmeters ... 235–236
Voltmeters ... 28, 233–234
Wave Analysers ... 242–243
Wheatstone Bridge ... 239
Wien Bridge ... 232
Insulators ... 21
Integer ... 8
Interelectrode Capacitances ... 43
Intermediate Frequency ... 123, 127–128
Intermediate Frequency Amplifiers ... 111, 124, 127–131
Intermediate Frequency Test Oscillator ... 244–245

PAGE

Interpolation (Frequency Measurement) 231
Interpolation (Graphs) 20
Interruption Frequency 117–118
Interstage Coupling 66
Interstage Transformers, I.F. 128–120
Inverse Peak Voltage 144
Inversion, Temperature 161
Inverted Amplifier 71
Inverter, Phase 97
Ionization 23
Ionosphere 157–158
Ions 23
Iron-Vane Meters 234

J Antenna 193

Key Chirps 107, 110
Key Clicks 106–107, 110
Key, How to Use 257–258
Keying:
  Back Wave 106
  Break-In 107
  Key-Click Reduction 109
  Methods:
    Cathode or Center-Tap 108
    Blocked Grid 108
    Grid-Controlled Rectifier 108
    Plate 107
    Primary 108
    Screen Grid 107
    Waveform 106
    Suppressor Grid 108
  Oscillator Keying 110
  Parasitic Clicks 110
Kilocycle 29
Kilohm 26
Kilovolt 26
Kilowatt 26
Kilowatt-Hour 26

LC Constants (Table) 270
L-C Ratios 35, 53, 72–73
LD-Cut Crystals 61
L-Section Filters 38
Lag Circuits 109
Lag, Current 30
Lag, Keying 106
Layer Height 157–158
Lead, Current 30
Leakage, Checking Condenser 240
Leakage Reactance 32
Lecher Wires 227
Limiter 139–140
Limiting, Audio 97, 134
Limiting (Noise) Circuits 134–136
Line-of-Sight Transmission 160, 161
Line Voltage Adjustment 154–155
Linear Amplifiers 47
Linear Antenna Transformers 168, 178–179
Linear Circuits 39–40, 80–82
Linear Time Base 56, 363
Linearity 85, 86, 90, 100, 103
Lines of Force 22, 24–25
Lines, Resonant 40
Link Coupling 36, 66, 171–172
Lissajou's Figures 232–233
Load 21, 33, 43, 44, 143
Load Impedance 44, 71
Local Oscillator (Receiver) 123–124, 126–127
Locking 53
Log Paper 19
Log Tables 278–279
Logarithmic Functions 17–18
Logarithms 15–16
Long-Wire Antennas 179–182
Loops 39
Loose Coupling 36
Low-Level Modulators 92
Low-Pass Filters 38
Lumped-Constant Circuits 82–83

Magic-Eye (Electron-Ray) Tubes 132
Magnetic Field 24–25
Magnetic Storms 158
Magnetism 24–26
Magnetomotive Force 25, 26
Mantissa 16
Matching Impedances 33
Materials 249–250
Mathematics:
  Abscissae 18
  Base 15

PAGE

Brackets 11–12
Cancelling Out 12
Characteristic 12
Coefficient 16
Collecting Terms 11
Coördinates 12
Decibels 18
Decimals 18
Exponential Functions 7–9
Exponents 17–18
Formulas 9, 14–15
Functions 10–14
Graphs 16–18
Index 18–20
Interpolation 9
Log Paper 20
Logarithmic Functions 19
Logarithms 17–18
Mantissa 15–16
Notation 16
Ordinates 7, 10
Plotting Graphs 18
Polar Coördinates 18
Power Function 20
Powers 17
Radical Sign 9, 11, 15
Roots 9
Scales 9–10, 11, 15
Signs 18–20
Square Root 10–11
Transposition 9–10
Variables 12–13
Maximum Current Value 16–17
Measurements: 29
  Antenna 246–248
  Capacity 239–241
  Current 28, 32, 233–235
  Frequency 225–233
  Inductance 239–241
  Modulation 98, 102, 243, 245–246
  Power 28, 234, 235, 246
  Q 241–242
  Receivers 243–245
  Resistance 238–239, 242, 247
  Technique 225
  Transmitters 245–246
  Tubes 248
  Voltage 28, 233–238
Measuring Instruments 28, 32, 225–243
Megacycle 29
Megohm 26
Memorizing the Code 254
Mercury Vapor Rectifiers 49, 144
Metals, Relative Conductivity of 270
Meters (*see* "Instruments")
Metric Prefixes 271
Mho 43
Microammeters 28, 233
Microampere 26
Microfarad and Micromicrofarad 22
Microhenry 25
Micromhos 43
Microphones 84, 92–94
Microvolt 26
Microwatt 26
Milliammeters 28
Milliampere 26
Millihenry 25
Millivolt 26
Milliwatt 26
Mixers 124–126
Modulation: 98–102
  Adjustments and Testing 84–88
  Amplitude Modulation 85
  Capability 89–91
  Cathode Modulation 85
  Characteristic 84
  Envelope 102–105
  Frequency Modulation 88, 89
  Grid-Bias Modulation 86
  Impedance 98, 102, 245–246
  Measurements 84
  Methods 84
  Percentage 86–88
  Plate Modulation 86
  Power 91–92, 217–220
Modulators 26
Molecular Friction 21
Molecules 233
Monitoring Frequencies 100
Monitors 96, 139
Motorboating

Index

PAGE

Mounting Crystals ... 61–62
Moving Coil Meters ... 234
Moving Iron Vane Meters ... 234
Mu ... 43
Multi-Band Antennas ... 182–184
Multi-Element Tubes ... 48–49
Multi-Hop Transmission ... 150
Multi-Purpose Tubes ... 49
Multipliers, Frequency ... 58, 79–81
Multipliers, Voltmeter ... 28, 233
Multivibrator ... 52, 239
Mush ... 136
Mutual Conductance ... 43
Mutual Inductance ... 36

Negative Charges ... 22
Negative Resistance ... 53
Negative Resistance Oscillators ... 53
Neutralization ... 67–70
Neutralizing Bridge Circuits ... 39
Neutralizing Condensers ... 69
Nodes ... 30
Noise ... 97–98, 102, 120–121, 134
Noise Ratio ... 112, 120–121, 133
Noise Reduction ... 134–135
Non-Resonant Lines ... 175–179
Non-Synchronous Vibrators ... 154
Notation ... 7, 10
Nucleus ... 21
Numerical Values ... 273

Ohm ... 26
Ohm's Law ... 26
Ohm's Law for A.C. ... 31
Ohmmeters ... 238–239
Open Circuit ... 21
Open-Wire Line ... 166
Operating Angle, Amplifier ... 71, 79–80
Operating Point ... 44
Ordinates ... 18
Oscillating Detector ... 115–117
Oscillation ... 50–51
Oscillator Keying ... 110
Oscillators ... 50–53
Oscilloscopes ... 57–57, 98–101, 236–238
Output Capacity, Tube ... 43
Output Limiting ... 97
Output Transformer ... 44
Overmodulation ... 85
Overmodulation Indicators ... 243, 245–246

P ... 26
Padding Condenser ... 122
Parallel Amplifiers ... 46
Parallel Antenna Tuning ... 171
Parallel Capacitors ... 27
Parallel Circuits ... 33–34
Parallel Feed ... 52
Parallel Impedance ... 33, 34–35
Parallel Inductances ... 27
Parallel Resistances ... 26–27
Parallel Resonance ... 33–34
Parasitic Elements ... 190
Parasitic Oscillators ... 77–79
Patterns, Oscilloscope ... 55–56
Patterns, Radiation ... 165
Peak Current Value ... 29
Peak Diode Voltmeter ... 236
Peak Plate Current ... 144
Peak Voltage Rating ... 73–74
Pentagrid Converters ... 125
Pentode Amplifiers ... 66–67
Pentode Crystal Oscillators ... 62–65
Pentodes ... 49
Period ... 28
Permeability ... 25–26
Persistence, Screen ... 54
Phase ... 29–30
Phase Angle ... 30
Phase Difference ... 30
Phase Inversion ... 97
Phase Modulation ... 84
Phase Relations, Amplifiers ... 45
Phased Antennas ... 186–190
Phasing Control ... 130
'Phone (*see* "Radiotelephony")
Photometric Power Measurements ... 235
Piano Frequencies ... 225
Pickup Coil ... 169
Pi-Section Antenna Couplers ... 169–171
Pi-Section Filters ... 38
Pierce Oscillator ... 63

PAGE

Piezo-Electricity ... 35
Piezo-Electric Crystals ... 61–62
Piezo-Electric Microphone ... 93–94
Pitch ... 24, 42
Plate ... 42
Plate Current ... 101–102
Plate Current Shift ... 114
Plate Detectors ... 44
Plate Dissipation ... 45, 52
Plate Efficiency ... 107
Plate Keying ... 43
Plate Load Resistance ... 86–88
Plate Modulation ... 67–68
Plate Neutralization ... 43
Plate Resistance ... 143
Plate Supply ... 18
Plotting Graphs ... 7
Point, Decimal ... 20
Polar Coördinates ... 22
Polarity ... 156, 162, 164–165
Polarization ... 22
Pole ... 22
Positive Charges ... 82–83
Pot Oscillator ... 52
Potential Difference ... 27
Potentiometers ... 26
Power ... 45
Power Amplification ... 110
Power Amplifiers, Receiver ... 31
Power Factor ... 17
Power Functions ... 102
Power Gain, Antenna ... 45
Power Input ... 234, 235, 246
Power Measurement ... 52, 75–76, 246
Power Output ... 45
Power Sensitivity ... 107
Power Supply Keying ... 107
Power Supply ... 220–221
Constructional ... 24
Dry Batteries ... 222
Dynamotors ... 221
Emergency ... 222
Genemotors ... 143–155
Principles ... 208
Receiver Power Supplies ... 24
Storage Batteries ... 154
Vibrators ... 220–222
Vibrator Power Supplies ... 9, 11, 15
Powers (of a Number) ... 263
Practice Cipher Groups ... 262
Practice Code Words ... 271
Prefixes, Metric ... 133–134
Preselection ... 32, 35–36
Primary ... 108
Primary Keying ... 156–161
Propagation ... 125
Pulling ... 40
Pulsating Current ... 143
Pure D.C. ... 252
Push-Back Wire ... 46
Push-Pull Amplifiers ... 69
Push-Pull Neutralization ... 79
Push-Push Doubler ... 31, 35, 37, 60, 72, 80
Q ... 177–178
Q Antenna ... 266–267
Q Code ... 241–242
Q Measurement ... 39
Quarter Wave ... 39
Quasta Crystals (*see* "Crystals")
Quench Frequency ... 117–118

R ... 20
R.M.S. Current Value ... 29
Radian ... 29
Radiation ... 40
Radiation Angle ... 158, 162
Radiation Resistance ... 40
Radiator (*see* "Antennas")
Radical Sign ... 9
Radio Frequency Amplifiers (Receivers) ... 119–121
Radio Frequency Choke Coil ... 41
Radio-Frequency Resistance ... 34
Radiotelephony:
Adjustments and Testing ... 98–105
Class-B Modulators ... 91–92, 219–220
Constructional:
40-watt Speech Amplifier-Modulator ... 217–219
112 Mc. "Walkie-Talkie" ... 215–217
Class-B modulators ... 219–220
Microphones ... 92–94
Modulation ... 84–91
Monitors ... 245–246
Principles ... 84–105

PAGE

Reception 137
Speech Amplifiers 94–98
Range, U.H.F. 160
Ratio Arms 39
Ray, Direct and Reflected 159–160
Reactance 30–31
Reactance, Capacitive 31
Reactance Chart 272
Reactance, Inductive 31
Reactance Meter 33
Reactance Modulator 239
Receivers:
  Antennas for 194–195
  Construction al:
    1-Tube 2-Stage Regenerative 196–198
    7-Tube Regenerative Single-Signal 198–202
    9-Tube National HRO 202–203
    16-Tube Hammarlund Super-Pro 203–204
    (*See also* "Ultra-High Frequencies — Receivers")
  Frequency Measurement with 225
  Measurements 243–245
  Power Supplies 208
  Principles and Design 111–142
  Testing and Alignment 243–245
Receiving Code 254–257
Reflex Voltmeter 236
Regenerative I.F. 129
Rectification 42
Rectifiers 42, 49, 143–144
Rectifier Circuits 144–145
Reed, Vibrating 225
Reflected Impedance 33
Reflected Waves 39
Reflection of Radio Waves 156, 157–160, 163
Reflector 160
Refraction of Radio Waves 157
Regeneration 45
Regeneration Control 115–116
Regenerative Detectors 115
Regenerative Receivers 111
Regulation, Voltage 143
Regulator, Voltage 148–149
Relaxation Oscillator 53
Relay, Keying 107
Reluctance 26
Repulsion 22
Resistance 21, 26
Resistance Capacity Filter 38
Resistance-Capacity Time Constant 27–28
Resistance-Coupled Amplifier Data (Chart) 97
Resistance Coupled Audio Amplifiers 95–96
Resistance Coupling 36, 95–96
Resistance Measurement 238–239, 242, 247
Resistance, Negative 53
Resistance, Radiation 40
Resistance, R.F. 34
Resistances in Series and Parallel 26–27
Resistor 26
Resistor Color Code 269
Resonance 33
Resonance Curve 112
Resonant Circuits 33–35
Resonant Frequency 33
Resonant Impedance 35
Resonant Line Circuits 40
Resonant Transmission Lines 172–175
Response, Frequency 44
Reversal, Phase 45
Rhombic Antenna 185–186
Ribbon Microphones 94
Ripple 145
Rochelle-Salt Crystals 93
Root-Mean-Square Value 29
Roots 9–10, 11, 15
Rotary Antennas 191

S Meters 132–133
Saturation Point 42
Sawtooth Wave 55–56
Scales 18–20
Schematic Symbols ..... *Frontispiece*
Screen Grid 48
Screen-Grid Keying 107
Second Detector 124, 131, 134–135
Secondary 32, 35–36
Secondary Emission 48
Selectivity 34, 112, 129
Selectivity Control 131
Selectivity, Measurement of 243
Self-Bias 50
Self-Controlled Oscillators 59–61
Self-Induction 25

PAGE

Self-Oscillation 50–51
Self-Quenching 118
Sending Code 257–259
Sensitivity, Measurement of 243
Sensitivity, Microphone 92–93
Sensitivity, Receiver 112
Series Antenna Tuning 171
Series Capacities 27
Series Circuits 33
Series Feed 52
Series Inductances 27
Series Resistances 26–27
Series Resonance 33
Servicing Receivers 137–139
Sharpness of Resonance 34
Sheet Metal Working 251–252
Shielding 37–38, 76–77, 102, 139
Shielding, Effect on Inductance 274
Shields, Electrostatic 76–77
Short Wave Receivers (*see* "Receivers")
Shot Noise 134
Shunts, Meter 28, 233
Sidebands 85–86, 104–104, 106
Signal 44
Signal Generators 243–246
Signal-Handling Capability 113
Signal-Noise Ratio 112, 120–121, 133
Signal-Noise Ratio, Measurement 243
Signal-Strength Indicators 132–133
Signs 10–11
Silencers, Noise 134–136
Sine Curve 28
Sine Wave 28
Single-Signal Reception 129–131
Single-Wire Feed 166, 175
Skin Effect 34
Skip Distance 158–159
Sky Wave 156, 158–159
Slide-Back Voltmeter 236
Smoothing Choke 147
Solar Cycle 158
Soldering 253
Sounds, Code 254–257
Space Charge 42
Spectrum, A.C. 30
Speech Amplifiers 94–98, 217–219
Sporadic-E Ionization 161
Square Root 9–10
Squegging 118, 139
Stability, Frequency 52–53
Stability, Receiver 112, 139
Stabilization, Voltage 148–150
Stage Gain 95
Standard Signal Generator 243
Standards, Frequency 220–231
Standing Waves 39
Standing Wave Ratio 169
Starting Voltage 148
Static Characteristics 43
Static Charge 21–22
Static Electricity 21–23
Storage Batteries 24
Stub Antenna Matching 178–179
Substitution Measurements 240
Sunspot Cycle 158
Super-Control Tubes 49
Superheterodyne 111, 123–124
Superimposed Currents 40–41
Superregeneration 117–118
Superregenerative Detectors 111–112, 117–118
Suppressor Grid 49
Suppressor-Grid Keying 108
Suppressor-Grid Modulation 89
Suppressor, Noise 134–136
Surge Impedance 167
Sweep 55
Sweep Circuits 56
Swing, Grid 46
Swinging Choke 147
Symbols, Electrical 270
Symbols, Schematic ..... *Frontispiece*
Synchronization 56–57
Synchronous Vibrators 154

Tables and Charts (*see* "Charts and Tables")
Talking Level, Microphone 93
Tank Circuits 52, 60, 72–73
Taps 249
Tapped Circuits 33
Temperature Coefficient of Crystals 61
Temperature Inversion 161
Termination, Line 168

Index

PAGE

Test Equipment 233–248
Test Oscillator 244–245
Tetrode Amplifiers 66–67
Tetrode Crystal Oscillators 62–63
Tetrodes 48
Textbooks 280
Thermal Agitation 120
Thermionic Conduction 24
Thermionic Emission 24
Thermocouple Ammeters 32, 235
Thickness of Crystals 61
Threads, Cutting 251
Threshold Control 136
Thump Filters 109
Tickler 51
Tight Coupling 36–37
Time Constant 27–28
Tolerances, Color Code 269
Tone Control 119
Tone Sources 260–265
Tools 249
Trace, Cathode-Ray 54, 56
Tracking 127
Transceivers 215
Transconductance, Grid-Plate 43
Transformer Color Codes 269
Transformer Coupling 30, 86, 96
Transformer Design 148
Transformerless Power Supplies 152
Transformers 32–38
Transformers, I.F. 128–129
Transmission Lines 166–179
Transmitters:
  Antenna Coupling 169–172
  Constructional:
    Complete Transmitter Assemblies 223–224
      Hallicrafter HT-6 25-watt 224
    Exciters or Low-Power Transmitters:
      15-Watt Output 6L6 Bread-Board 209
      50-Watt Output 6L6-807 Plug-In Coil 209–212
    Power Amplifiers:
      450-Watt Input Push-Pull 212–213
  (*See also* "Ultra-High Frequencies — Transmitters")
  Measurements 245–246
  Principles and Design 58–63
Transposition 12–13
Trapezoidal Pattern 98–100
Triangular Spectrum 103
Triggering Voltage 56
Trimmer Condenser 122
Triodes 42–44
Triode Crystal Oscillators 62
Triple-Tuned Transformers 128–129
Tri-Tet Oscillator 63–65
Troposphere Refraction 161
Trouble Shooting (Receivers) 137–139, 243, 244
Trough Line Oscillator 82
Tube Checking 248
Tube Keyer 109, 265
Tube Noise 120–121
Tubes (*see* "Vacuum Tubes")
Tuned Circuits 33–35
Tuned-Grid Tuned-Plate Circuit 52
Tuned R.F. Receiver 111
Tuning Fork 225
Tuning Indicators 132–133
Tuning Receivers 116–117, 136–137, 142
Tuning Transmitters 60–61, 63, 64–65, 74–77
Turns Ratio 32
Twisted Pair Feed 166, 175–176
Two-Conductor Line 80–81

Ultraudion Circuit 52

PAGE

Ultra-High Frequencies:
  Constructional:
    Receivers:
      224 & 112 Mc. Superregenerative 204–206
      Hallicrafter S-27 FM-AM Superhet 207–208
    Transmitter-Receivers:
      112-Mc. Battery "Walkie-Talkie" 215–217
    Transmitters:
      112-Mc. 815 Linear Oscillator 213–214
      224-Mc. HY75 Oscillator 214–215
Upward Modulation 101–102

V Antennas 184–185
V-Cut Crystals 61
Vacuum Tube Voltmeters 235–236
Vacuum Tubes 24, 42–57
Vacuum Tube Testing 248
Vacuum Tube Symbols 270
Variable-μ Tubes 49
Variable Selectivity 129, 130–131
Velocity, Angular 29
Velocity Microphone 94
Velocity of Radio Waves 39
Vibrating Reed 225
Vibrator Power Supplies 221–222
Vibrators 154
Virtual Height 157–158
Volt 22
Volt-Amperes 31, 148
Voltage Amplification 45
Voltage Dividers 27, 152
Voltage Drop 27
Voltage-Doubling Circuits 152–153
Voltage Feed for Antennas 166
Voltage Measurement 28, 233–238
Voltage Ratings (Color Code) 269
Voltage Ratio, Transformer 32
Voltage Regulation 143
Voltage Rise 34
Voltage-Stabilized Power Supplies 148–150
Voltmeters 28, 233–234, 235–238
Volume Control 96–97, 131–133

WWV Standard Frequency Transmissions 230
Watt 26
Watt-Hour 29
Wattmeter 28, 234
Wave Analyzers 242–243
Wave Angle 158
Wave Envelope Pattern 98–100
Wave Propagation 156–161
Waveform 29, 44, 100, 106
Wavelength 39
Wavelength-Frequency Conversion 39
Wavemeter (*see* "Frequency Meters")
Wheatstone Bridge 38–39, 239
Wien Bridge 232
Winding Coils 253
Wire Table 276
Wiring Apparatus 252–253
Wiring, Current Capacity 271
Wiring Diagrams ..... *Frontispiece*
Working Voltage 145
Workshop Practice 249–253

X 31
X-Cut Crystals 61

Y-Cut Crystals 61

Z 31
Zero Beat 117
Zero-Beat Indicators 231–232
Zeppelin Feed for Antennas 173