ARRL electronics data book

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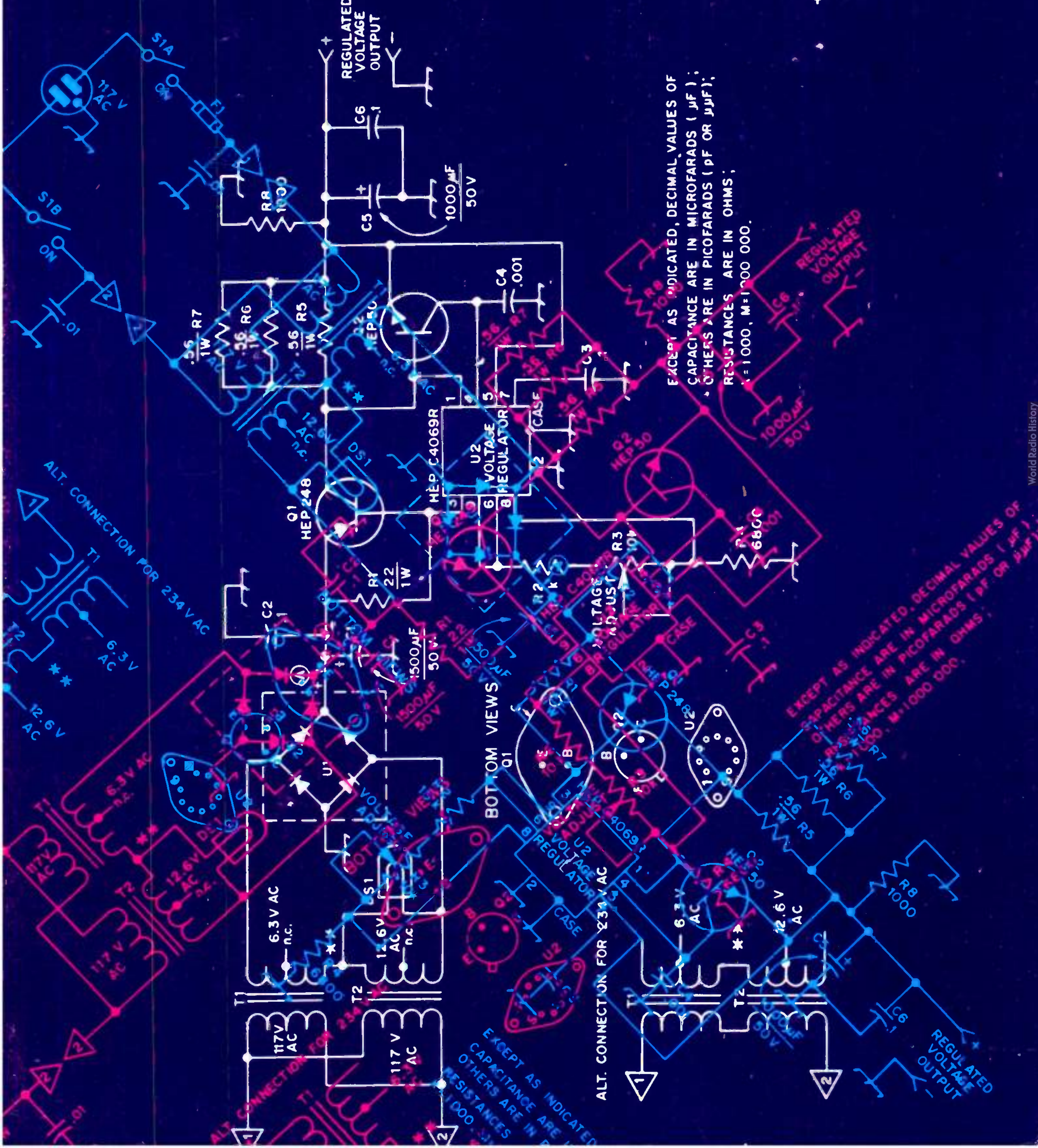

ARRL electronics data book

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Foreword

The state of the electronics art is in large measure a product of the efforts put forth by amateurs since the beginning of radio. More recently, amateurs have contributed significantly to the improvement of communications equipment through innovation, experimental endeavor, and professional design techniques. Regardless of the technical aptitude and formal education levels of amateurs, each is capable of making contributions to the ever-improving state of the art.

It has been a long-established actuality that advances in communications techniques and circuit design have been the products of engineers and experimenters alike. Although a different approach to solving a problem is taken by each type of worker, certain basic precepts must be observed if circuit development is to be as rapid and technically sound as possible. It is helpful, if not essential, to have a basic understanding of mathematical and electrical laws. Ordinarily, it is necessary to maintain a substantial reference library of electrical data if success in the design endeavor is to be realized. Therefore, most experimenters and engineers own data books they have compiled themselves, or refer to similar collections of information which have been prepared and published by others. Whatever the origin of the file, it becomes a portable fountain of useful information which the owner is likely to regard as his or her most valuable booklet.

This book has been written especially for the radio amateur. It contains a potpourri of equations, nomographs, data charts, tables, and explanations of terms and methods. It was prepared to serve as a useful supplement to the other technical books found in a radio amateur's personal library.

Richard L. Baldwin, W1RU
General Manager

Acknowledgment

No book of this kind can be the sole product of one author's knowledge and imagination. In common with many technical publications, this compilation of information is based in part upon lecture and laboratory notations made by the editor over a period of years. During that time, there was no thought toward publication of the data collected, and material was garnered from numerous excellent sources which, with the passage of time, have become obscure or of unknown origin. It is impossible, therefore, to itemize the sources and place credit where it is so justly due.

Some of the material in this file is based on simplification of established design procedures — near rule-of-thumb approaches developed by the author to aid amateurs who are lacking in formal backgrounds in the technical field. Other matter in this book was borrowed from the papers of present and former ARRL Hq. staff members, and from contributions made to *QST* by amateurs the world over.

The author is indebted to RCA and Motorola for their valuable assistance in furnishing data sheets, application notes, and technical books which helped inspire the presentation of considerable data in this volume. Gratitude is expressed also for the help obtained from National Semiconductor and Signetics Corp. through their contributions of data files and technical books on semiconductors. Finally, the writer wishes to acknowledge and express appreciation for the forbearance of W1CKK, the XYL, who spent many lonely nights and weekends while this book was prepared.

The ARRL welcomes any corrections or suggestions for improvement of future editions of this publication.

Doug DeMaw, W1FB

Contents

Chapter 1

Math Aids and Tables

To assure simplification of the data contained in this book, let us say that there are six fundamental components used in electronics equipment — resistors, capacitors, coils, tubes, semiconductors and conductors. The various properties which relate to their use, but not necessarily in the order of prime significance, are *T* (time), *V* (voltage), *I* (current), *X* (reactance), *Z* (impedance), *Q* (quality factor), *P* (power) and phase. There are other terms of major importance, such as *f* (frequency), *Y* (admittance), μ (conductance), *k* (dielectric constant) and θ (phase angle). Most electronics math relates to design work which concerns the six fundamental components mentioned here and utilizes the properties which have been listed as attendant to them in solving equations.

A substantial amount of the data offered in this chapter is meant to serve as peripheral information to the basic math used by engineers and technicians. That is, no attempt shall be made to teach a course in math. Rather, data are given which can be used to shorten the time of calculations, and to enhance accuracy by reducing the number of steps necessary when working a problem.

The choice of material has been based principally on observations made while answering technical inquiries at ARRL hq. and is included to illuminate those blind spots which many amateur builder/designers have.

1) $a = \frac{b}{c}$, $\therefore b = ac$, and $c = \frac{b}{a}$

2) $b = \frac{ad}{c}$, $\therefore c = \frac{ad}{b}$, and $d = \frac{bc}{a}$

3) $a = \frac{1}{d\sqrt{bc}}$, $\therefore a^2 = \frac{1}{d^2\,bc}$,

$b = \frac{1}{d^2 a^2 c}$, $c = \frac{1}{d^2 a^2 b}$, and

$d = \frac{1}{a\sqrt{bc}}$

4) $a = \sqrt{b^2 + c^2}$, $\therefore a^2 = b^2 + c^2$,

$b = \sqrt{a^2 - c^2}$, and $c = \sqrt{a^2 - b^2}$

Fig. 1 — Examples of how equations can be transposed to find various unknown quantities.

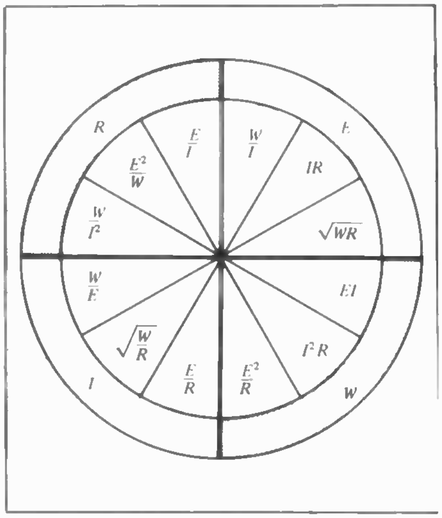


Fig. 2 — Ohm's Law wheel for dc circuits.

Ohm's Law for DC

The terms *I*, *E*, *R* and *W* are significant in Ohm's Law for dc applications. Solutions to various electrical problems are

Table 1

SYMBOL	*MEANING*
X or ·	Multiplied by
÷ or :	Divided by
+	Positive, plus, add
—	Negative, minus, subtract
±	Positive or negative. Plus or minus
∓	Negative or positive. Minus or plus
= or ::	Equals
≡	Identical to
≅ or ≈	Is approximately equal to
≠	Does not equal
>	Is greater than
≫	Is much greater than
<	Is less than
≪	Is much less than
≥	Greater than or equal to
≤	Less than or equal to
∴	Therefore
∠	Angle
∠s	Angles
Δ	Change, increase or decrease
⊥	Perpendicular to
‖	Parallel to
$\|n\|$	Absolute value of *n*
$\sqrt{\ }$	Square root
$\sqrt[3]{\ }$	Cube root

Table 2

Terms and Numerical Equivalents

π	= 3.14159265	$\sqrt{2}$	= 1.4142	$1/\sqrt{8}$	= 0.354
2π	= 6.28318530	$\sqrt{3}$	= 1.7321	$1/\sqrt{9}$	= 0.333
$(2\pi)^2$	= 39.476089	$\sqrt{4}$	= 2.0000	$1/\sqrt{10}$	= 0.316
4π	= 12.5663706	$\sqrt{5}$	= 2.2361	log 1	= 0.000000
π^2	= 9.8690440	$\sqrt{6}$	= 2.4495	log 2	= 0.301030
$\pi/2$	= 1.57079633	$\sqrt{7}$	= 2.6458	log 3	= 0.477121
$\pi/3$	= 1.04719755	$\sqrt{8}$	= 2.8248	log 4	= 0.602060
$\pi/4$	= 0.78539816	$\sqrt{9}$	= 3.0000	log 5	= 0.698970
$1/\pi$	= 0.31830989	$\sqrt{10}$	= 3.1623	log 6	= 0.778151
$1/\pi^2$	= 0.10132118	$1/\sqrt{2}$	= 0.707	log 7	= 0.845098
$\sqrt{\pi}$	= 1.77245385	$1/\sqrt{3}$	= 0.577	log 8	= 0.903090
		$1/\sqrt{4}$	= 0.500	log 9	= 0.954243
		$1/\sqrt{5}$	= 0.447	log 10	= 1.000000
		$1/\sqrt{6}$	= 0.408		
		$1/\sqrt{7}$	= 0.377		

Logs are to the base 10.

Table 3

Standard Resistance Values

Numbers in bold type are ± 10% values. Others are 5% values.

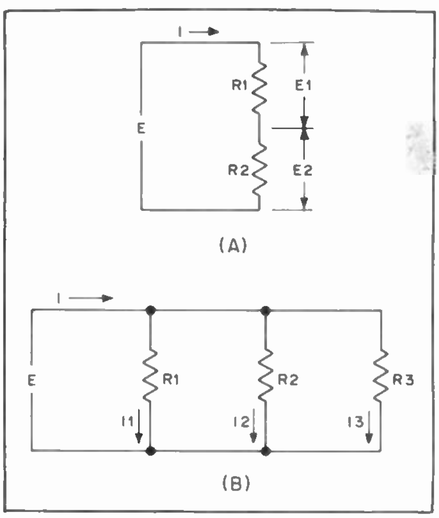
OHMS										*MEGOHMS*				
1.0	3.6	**12**	43	**150**	510	**1800**	6200	**22000**	75000	0.24	0.62	1.6	4.3	11.0
1.1	**3.9**	13	**47**	160	**560**	2000	**6800**	24000	**82000**	**0.27**	**0.68**	**1.8**	**4.7**	**12.0**
1.2	4.3	**15**	51	**180**	620	**2200**	7500	**27000**	91000	0.30	0.75	2.0	5.1	13.0
1.3	**4.7**	16	**56**	200	**680**	2400	**8200**	30000	**100000**	**0.33**	**0.82**	**2.2**	**5.6**	**15.0**
1.5	5.1	**18**	62	**220**	750	**2700**	9100	**33000**	110000	0.36	0.91	2.4	6.2	16.0
1.6	**5.6**	20	**68**	240	**820**	3000	**10000**	36000	**120000**	**0.39**	**1.0**	**2.7**	**6.8**	**18.0**
1.8	6.2	**22**	75	**270**	910	**3300**	11000	**39000**	130000	0.43	1.1	3.0	7.5	20.0
2.0	**6.8**	24	**82**	300	**1000**	3600	**12000**	43000	**150000**	**0.47**	**1.2**	**3.3**	**8.2**	**22.0**
2.2	7.5	**27**	91	**330**	1100	**3900**	13000	**47000**	160000	0.51	1.3	3.6	9.1	
2.4	**8.2**	30	**100**	360	**1200**	4300	**15000**	51000	**180000**	**0.56**	**1.5**	**3.9**	**10.0**	
2.7	9.1	**33**	110	**390**	1300	**4700**	16000	**56000**	200000					
3.0	**10.0**	36	**120**	430	**1500**	5100	**18000**	62000	**220000**					
3.3	11.0	**39**	130	**470**	1600	**5600**	20000	**68000**						

obtained through variations of the basic equation $R = E/I$; or $W = E \times I$. To find other unknowns, transposition of the terms is necessary. Fig. 1 shows how to transpose simple terms. Fig. 2 provides an Ohm's Law wheel which is based on transpositions of $R = E/I$ and $W = E \times I$. Table 1 contains a list of math symbols which will aid the reader to understand the equations found throughout this book. Additional aids are provided in Table 2.

Resistance

Resistors are frequently used in combination (series or parallel configurations), Fig. 3, to take advantage of available values or to increase the overall wattage capability of a resistive circuit element. Also, situations will arise in which a nonstandard resistance value will be required for a particular application. Parallel or series resistor arrangements can be used to obtain the needed value. Table 3 lists the standard 5- and 10-percent tolerance resistor values

Fig. 3 — *R, E* and *I* relationship of series and parallel resistances.



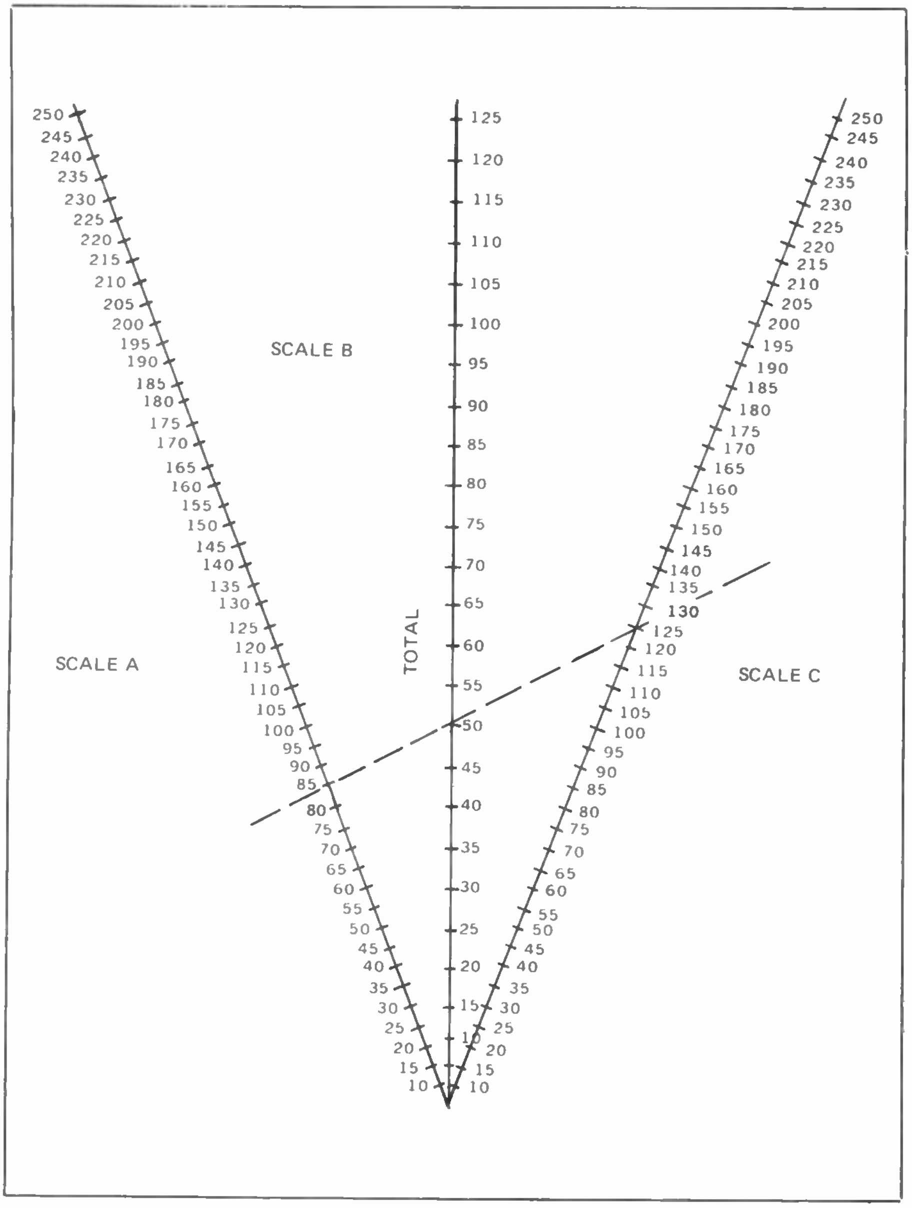


Fig. 4 — Nomograph for determining values of parallel resistances and inductances, and series capacitors. The dashed line shows that a total resistance of 50 ohms is obtained (B scale) when 85 ohms (A scale) is placed in parallel with 125 ohms (C scale). For greater quantities of *R, C* or *L* (250 or greater), add the necessary number of zeros to the numbers of scales A, B and C.

available from manufacturers in ratings up to 2 watts.

The total resistance of series combinations of resistors can be found by

$$R_T = R1 + R2 + R3, \text{etc.} \qquad \text{(Eq. 1)}$$

Therefore, if R1 = 470, R2 = 12,000, and R3 = 100,000 ohms, respectively, the three components in series would amount to 112,470 ohms as illustrated here

$$R_T = 470 + 12{,}000 + 100{,}000$$
$$\therefore R_T = 112{,}470 \text{ ohms} \qquad \text{(Eq. 2)}$$

and the total wattage would be increased over that of any one resistor of the combination. Three identical 1-watt resistors provide a 3-watt rating.

The same is not true of parallel combinations of resistors. The total value of resistance must be determined by

$$R_T = \frac{1}{\frac{1}{R1} + \frac{1}{R2} + \frac{1}{R3}}$$

$$\therefore R_T = \frac{1}{\frac{1}{470} + \frac{1}{12{,}000} + \frac{1}{100{,}000}}$$

$$R_T = 450.25 \text{ ohms} \qquad \text{(Eq. 3)}$$

and the wattage rating of the three units in parallel is dependent mainly on the rating of the lowest ohmic-value resistor, since maximum current flows through the path of least resistance. In this example the total wattage of the three units in combination would be just slightly greater than 1 watt if R1, R2 and R3 were each rated at 1 watt. The total wattage would be 3 only if each resistor was of identical value; e.g., three 470-ohm resistors in parallel (1-watt rating each) would yield 156.66 ohms at 3 watts. When only two resistances are used in parallel, the total value can be obtained by $R_T = R1 \times R2 \div R1 + R2$. Rapid solutions to parallel-resistor problems can be had by using the nomograph of Fig. 4. It is necessary only to place a straightedge across any two *known* values and read the unknown quantity on the remaining scale. The same nomograph can be used to determine unknown values of capacitance in series. The electronics prefixes, symbols, and multipliers given in Fig. 5 can be used to supplement the nomograph of Fig. 4 when computing various denominations of resistance and capacitance versus the scale numbers of the graph.

Prefix	Symbol	Numerical Value	Exponential Value
tera	T	1,000,000,000,000	10^{12}
giga	G	1,000,000,000	10^{9}
mega	M	1,000,000	10^{6}
kilo	K	1,000	10^{3}
hecto	H	100	10^{2}
deka	dk	10	10^{1} or 10
deci	d	0.1	10^{-1}
centi	c	0.01	10^{-2}
milli	m	0.001	10^{-3}
micro	μ	0.000,001	10^{-6}
nano	n	0.000,000,001	10^{-9}
pico	p	0.000,000,000,001	10^{-12}

Fig. 5 – Chart which shows numerical equivalents for metric prefixes.

Power and Energy

Power, the rate of doing work, is equal to voltage multiplied by current. The unit of electrical power is the watt, and is equal to *one volt multiplied by one ampere.* Common fractional and multiple units for power are the *milliwatt* (mW), one one-thousandth of a watt, and the *kilowatt* (kW), one thousand watts. Therefore, the equations for power in milliwatts, watts or kilowatts are $P = E \times I$, $P = E^2 \div R$, and $P = I^2 \times R$. These formulas are pertinent when two terms of each are known . . . the resistance and the current or voltage. *Example:* How much power will be dissipated in a bleeder resistor of 10,000 ohms which is connected across the output of a 275-volt power supply? From the equation

$$P = \frac{E^2}{R} \therefore P = \frac{275^2}{10{,}000},$$

$$P = 7.56 \text{ W} \qquad \text{(Eq. 4)}$$

Fig. 5 can be consulted when converting numerical values of voltage, resistance, current, and power to the standard prefixes (10,000 ohms to 10-kΩ, 0.1 V to 100 mV, 0.4 W to 400 mW and the like).

There are other terms to consider when treating the matter of direct-current power. The kW hour, for example, is defined as work = power in kW × time in hours; and joules, watts and horsepower can be resolved by

1 hp (horsepower) = 33,000 ft.-lb. per minute,
or 550 ft.-lb. per second.
1 ft.-lb. = 1.356 joules.
1 hp = 550 × 1.356 joules.
1 hp = 746 W.
1 W = 1.341×10^{-3} hp.

Still another important consideration when relating to power is *efficiency*. It is the ratio of power output from a circuit to that of its input. Output power will always be less than input power. Efficiency is calculated by Eff. = power output ÷ power input. *Example:* What is the efficiency of a solid-state transmitter which operates at 12-V dc and takes 3 A from the power supply while delivering 10 watts of rf power to the antenna? From the Ohm's Law wheel formula of Fig. 2 the transmitter consumes 36 W: $W = E \times I$, which equals 36. Therefore, the efficiency of the circuit is 10 ÷ 36 = 27.7 percent (0.277 × 100). The foregoing is a measure of the *overall* efficiency of the composite transmitter circuitry. Last-stage efficiency must be determined in a like manner, but with the collector-to-emitter voltage (V_{ce}) and collector current product representing the power input. Thus, if the V_{ce} was 12, and the I_c (collector current) was 1.66 A, power input to the PA stage would be 20 W. Therefore, the efficiency would be 10 ÷ 20 = 0.5, or 50 percent.

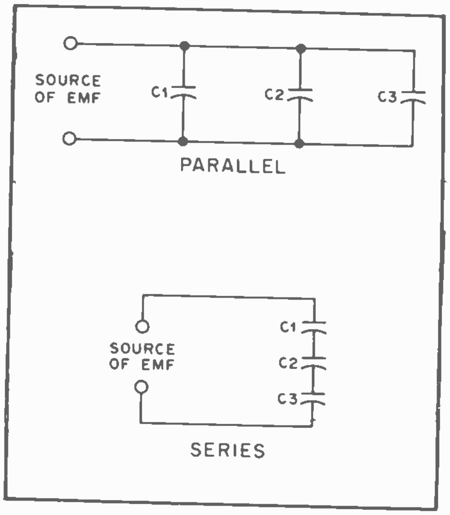


Fig. 6 – Illustrations of capacitors in series and parallel.

Capacitors in Series and Parallel

The terms "parallel" and "series," when used with reference to capacitors, have the same circuit meaning as with resistances. When a number of capacitors are connected in parallel, as in Fig. 6, the total capacitance of the group is equal to the sum of the individual capacitances, so C(total) = C1 + C2 + C3 + C4 + ...

However, if two or more capacitors are connected in series, as in the second drawing, the total capacitance is less than that of the smallest capacitor in the group. The rule for finding the capacitance of a number of series-connected capacitors is the same as that for finding the resistance of a number of *parallel*-connected resistors. That is,

$$C(\text{total}) = \frac{1}{\frac{1}{C1} + \frac{1}{C2} + \frac{1}{C3} + \frac{1}{C4} + \ldots}$$

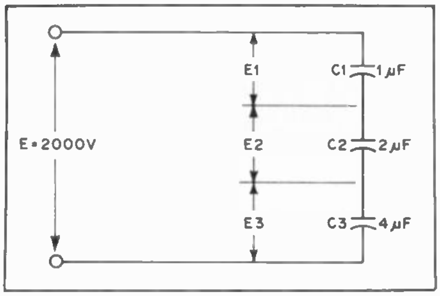


Fig. 7 — Voltage division of capacitors in series.

and, for only two capacitors in series,

$$C(\text{total}) = \frac{C1 \times C2}{C1 + C2} \qquad \text{(Eq. 6)}$$

The same units must be used throughout; that is, all capacitances must be expressed in either µF or pF; both kinds of units cannot be used in the same equation.

Capacitors are connected in parallel to obtain a larger total capacitance than is available in one unit. The largest voltage that can be applied safely to a group of capacitors in parallel is the voltage that can be applied safely to the one having the *lowest* voltage rating.

When capacitors are connected in series, the applied voltage is divided up among them; the situation is much the same as when resistors are in series and there is a voltage drop across each. However, the voltage that appears across each capacitor of a group connected in series is in *inverse* proportion to its capacitance, as compared with the capacitance of the whole group.

Example: Three capacitors having capacitances of 1, 2 and 4 µF, respectively, are connected in series as shown in Fig. 7. The total capacitance is

$$C = \cfrac{1}{\cfrac{1}{C1} + \cfrac{1}{C2} + \cfrac{1}{C3}} = \cfrac{1}{\cfrac{1}{1} + \cfrac{1}{2} + \cfrac{1}{4}} = \frac{1}{1.75}$$

$$= 0.571\ \mu F \qquad \text{(Eq. 7)}$$

Table 4

Standard Capacitance Values

pF	*pF*	*pF*	*pF*	*µF*	*µF*	*WV*	*µF*	*WV DC*	*µF*	*WV DC*
0.3	470	5	470	.001	1000	15	36,000	3	1,650	100
5	500	10	510	.0015	2000	15	27,500	5	2,250	100
6	510	12	560	.0022	3000	15	6,000	10	3,600	100
6.8	560	15	620	.0033	10000	15	8,200	10	5,300	100
7.5	600	18	680	.0047	500	25	10,000	10	8,300	100
8	680	22	750	.0068	1000	25	12,500	10	275	150
10	750	24	820	.01	2000	25	18,000	10	500	150
12	800	27	910	.015	4000	25	1,500	25	700	150
15	820	30	1000	.022	5000	25	2,800	25	1,100	150
18	910	33	1200	.033	500	50	3,800	25	1,550	150
20	1000	39	1500	.047	1500	50	4,500	25	2,500	150
22	1000	47	1800	.068	2000	50	6,000	25	3,600	150
24	1200	51	2000	0.1	300	150	8,500	25	5,600	150
25	1200	56	2200	0.15	150	350	13,500	25	180	200
27	1300	62	2400	0.22	50	350	20,000	25	450	200
30	1500	68	2700	0.33	125	350	31,500	25	550	200
33	1500	82	3000	0.47	10	450	2,100	35	1,000	200
39	1600	100	3300	0.68	20	450	3,100	35	2,450	200
47	1800	120	3900	1.0	40	450	3,600	35	3,800	200
50	2000	150	4300	1.5	80	450	800	50	140	250
51	2200	180	4700	2.0	100	450	1,500	50	275	250
56	2500	200	5100		90	500	2,000	50	375	250
68	2700	220	5600				2,500	50	800	250
75	3000	240	6200				3,300	50	1,900	250
82	3300	270	6800				4,500	50	3,000	250
91	3900	300	7500				7,300	50	325	300
100	4000	330	8200				10,000	50	700	300
120	4300	360	9100				16,500	50	2,600	300
130	4700	390	10000				600	75	100	350
150	4700	430	15000				1,000	75	250	350
180	5000						1,500	75	400	350
200	5000						2,000	75	550	350
220	5600						2,500	75	900	350
240	6800						3,450	75	1,300	350
250	7500						5,500	75	2,000	350
270	8200						8,200	75	150	400
300	10000						12,500	75	325	400
330	10000						400	100	40	450
350	20000						750	100	110	450
360	30000						1,000	100	240	450
390	40000						1,300	100		
400	50000									
470										
A		*B*		*C*	*D*		*E*			

Column A lists disk ceramic. B values are for silver-mica units, and C lists tubular paper and mylar caps. Column D gives values for single-section electrolytics, and E lists high-capacitance electrolytics.

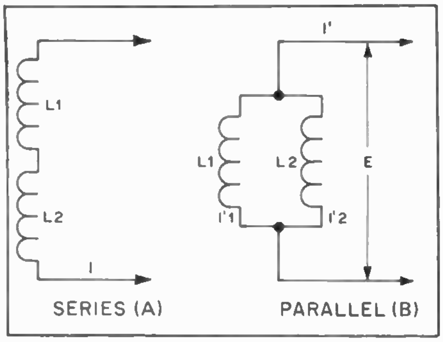


Fig. 8 — Examples of inductances in series and in parallel.

The voltage across each capacitor is proportional to the *total* capacitance divided by the capacitance of the capacitor in question, so the voltage across C1 is

$$E_1 = \frac{0.571}{1} \times 2000 = 1142 \text{ volts} \qquad \text{(Eq. 8)}$$

Similarly, the voltages across C2 and C3 are

$$E_2 = \frac{0.571}{2} \times 2000 = 571 \text{ volts} \qquad \text{(Eq. 9)}$$

$$E_3 = \frac{0.571}{4} \times 2000 = 286 \text{ volts} \qquad \text{(Eq. 10)}$$

totaling approximately 2000, the applied voltage.

Capacitors are frequently connected in series to enable the group to withstand a larger voltage (at the expense of decreased total capacitance) than any individual capacitor is rated to stand. However, as shown by the previous example, the applied voltages do not divide equally among the capacitors (except when all the capacitances are the same) so care must be taken to see that the voltage rating of no capacitor in the group is exceeded.

Table 4 contains a reasonably complete listing of U.S. values of capacitance for disk-ceramic, silver-mica, tubular and electrolytic capacitors. The values given are those which are normally stock values in electronics parts stores. The columns which do not include voltage ratings were so structured to avoid extensive compilations of values versus voltage, which would be necessary in the case of disk-ceramic and silver-mica capacitors. The nomograph of Fig. 4 can be used to determine the value of two capacitors in series. The procedure is the same as for resistors in parallel.

Inductances in Series and Parallel

If the two inductances in Fig. 8A are not magnetically coupled in any way (no mutual inductance), each sets up a number of flux linkages in proportion to its own inductance when the current, *I*, flows. Since the same current flows

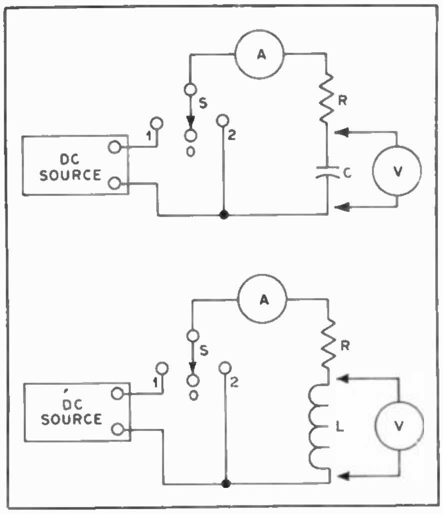


Fig. 9 – Circuit illustrations of *RC* and *RL* time-constant networks.

through both coils, the total number of linkages, and therefore the total inductance, is the sum. The total inductance of any number of inductances (having no mutual coupling) in series is the sum of their individual inductances.

When inductances are connected in parallel, Fig. 8B, the relationships are similar to those for resistances in parallel. (The inductances are assumed to have no mutual coupling.) In this case it is necessary to deal with the *instantaneous* induced voltage, *e*, and *rate of change* of current, here labeled *i'*, instead of the steady voltages and currents common to a parallel-resistance circuit. Using similar reasoning leads to

$$L_T = \frac{1}{\frac{1}{L1} + \frac{1}{L2}} = \frac{1}{\frac{1}{95} + \frac{1}{170}}$$

$$= \frac{1}{.0105 + .00588} = 61\ \mu H \qquad \text{(Eq. 11)}$$

where L1 = 95 μH and L2 = 170 μH.

RC and *RL* Time Constant

It is useful when working with communications equipment to employ timing circuits consisting of resistance and capacitance, or inductance and resistance networks. *RC* timing circuits are the most common ones, and they are used in agc loops, VOX or cw break-in delay hookups, repeater control systems and the like. In some situations a long *charge* (attack) time is needed, wherein resistance and capacitance are connected in series. In other applications the *R* and *C* components are placed in parallel to assure a rapid attack and slow *discharge* (decay) time. The latter is the most applicable to amateur circuits.

When capacitance or inductance are in a circuit which also contains resistance, a finite time is required to build up the field in which energy is stored. Also, a finite time is required for the energy to be withdrawn. This time-dependence is expressed by the *time constant* of a capacitance-resistance or inductance-resistance circuit.

Refer to Fig. 9 where, if the switch *S* is moved from the open position (0) to position 1, current begins to flow into the capacitor *C* through resistance *R*. At the instant of closing *S* the maximum current that can flow is equal to the source voltage divided by *R* and the voltage measured by the voltmeter is zero. The voltage then rises at a rate determined by the rate at which charge can be placed on *C* through *R*, until eventually *V* will read the source voltage. (Theoretically, the capacitor voltage never quite reaches the source voltage; practically, the difference between the two becomes immeasurably small after a time.) If *S* is moved to 0 again, the charge remains on the capacitor indefinitely, depending on the leakage (gradual loss of charge through a path for current flow either in the capacitor itself or externally) The *time in seconds* required for the voltage across *C* to reach 63 percent of the final value during charging is the *time constant* of the *CR* circuit. If *S* is now moved to position 2, the capacitor will discharge through *R* and the same amount of time will be required for the voltage across *C* to decrease to 37 percent of its fully charged value. These figures are based on the fact that the voltage increases and decreases logarithmically.

In the inductance-resistance circuit, Fig. 9, on moving *S* to position 1 from 0 the current begins to rise from zero,

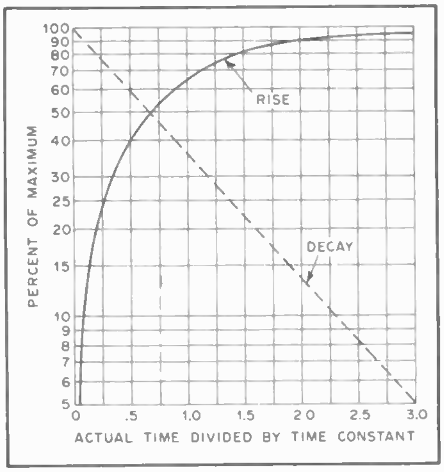


Fig. 10 – Chart showing relationship of time versus rise and decay characteristics (percentages of maximum) for *RC* networks.

being held back at first by the opposing self-induced voltage generated in *L* by the change in current. As time goes on, the rate of change becomes smaller and the current finally is determined solely by *R* and the source voltage. (Again, in theory the current never quite reaches this value, but in practice the difference is too small to be measured after a short time.) The voltage measured across *L* thus starts at the source voltage and decreases to zero. The time constant of such a circuit is defined for current in the *LR* circuit in the same way as for voltage in the *CR* case, and is equal to *L/R*.

If *S* could be moved from 1 to 2 *instantaneously*, the current would at first equal the voltage induced in *L* divided by the circuit resistance *R* (the induced voltage is the result of cutting off the steady current). It would then gradually decrease to zero, reaching 37 percent of its initial value in the time constant of the circuit.

The way in which voltage and current rise or decay in *CR* and *LR* circuits is shown in Fig. 10. As applied to the capacitive circuit in Fig. 9, the "rise" curve shows how, on charging, the voltage *V* increases with time, and the "decay" curve shows how the charging current *A* decreases with time. On *discharge* the "decay" curve shows both current and voltage.

Similarly, in the *LR* circuit the "rise" curve shows how the current builds up, and the "decay" curve gives the voltage change across the inductor when the field is being created.

These curves are plotted against the number of time constants of elapsed time – that is *actual* time divided by the time constant. Thus, if a circuit has a time constant of 0.25 second, the voltage or current would decay to 5 percent of the maximum value in 3 × 0.25 = 0.75 second. Actual rise times are

Table 5

Pk-pk, Pk, RMS and Average AC *E* and *I*

TO OBTAIN / *VOLTAGE DESIGNATOR*	*PK-PK*	*PK*	*RMS*	*AVERAGE*
PK-PK		pk × 2	0.3535 × pk-pk	0.3185 × pk-pk
PK	2 pk	$\frac{\text{pk-pk}}{2}$	0.707 × pk	0.637 × pk
RMS	2.828 × rms	1.41 × rms		0.9 × rms
AVERAGE	3.14 × aver.	1.57 × aver.	1.11 × aver.	

Numerical relations of the various ac voltage terms.

Table 6

Numerical Equivalents Angle Functions

ANGLE	*SIN*	*COS*	*TAN*	*ANGLE*	*SIN*	*COS*	*TAN*
0°	.0000	1.000	.0000	45°	0.7071	0.7071	1.0000
1	.0175	0.9998	.0175	46	0.7193	0.6947	1.0355
2	.0349	0.9994	.0349	47	0.7314	0.6820	1.0724
3	.0523	0.9986	.0524	48	0.7431	0.6691	1.1106
4	.0698	0.9976	.0699	49	0.7547	0.6561	1.1504
5	.0872	0.9962	.0875	50	0.7660	0.6428	1.1918
6	0.1045	0.9945	0.1051	51	0.7771	0.6293	1.2349
7	0.1219	0.9925	0.1228	52	0.7880	0.6157	1.2799
8	0.1392	0.9903	0.1405	53	0.7986	0.6018	1.3270
9	0.1564	0.9877	0.1584	54	0.8090	0.5878	1.3764
10	0.1736	0.9848	0.1763	55	0.8192	0.5736	1.4281
11	0.1908	0.9816	0.1944	56	0.8290	0.5592	1.4826
12	0.2079	0.9781	0.2126	57	0.8387	0.5446	1.5399
13	0.2250	0.9744	0.2309	58	0.8480	0.5299	1.6003
14	0.2419	0.9703	0.2493	59	0.8572	0.5150	1.6643
15	0.2588	0.9659	0.2679	60	0.8660	0.5000	1.7321
16	0.2756	0.9613	0.2867	61	0.8746	0.4848	1.8040
17	0.2924	0.9563	0.3057	62	0.8829	0.4695	1.8807
18	0.3090	0.9511	0.3249	63	0.8910	0.4540	1.9626
19	0.3256	0.9455	0.3443	64	0.8988	0.4384	2.0503
20	0.3420	0.9397	0.3640	65	0.9063	0.4226	2.1445
21	0.3584	0.9336	0.3839	66	0.9135	0.4067	2.2460
22	0.3746	0.9272	0.4040	67	0.9205	0.3907	2.3559
23	0.3907	0.9205	0.4245	68	0.9272	0.3746	2.4751
24	0.4067	0.9135	0.4452	69	0.9336	0.3584	2.6051
25	0.4226	0.9063	0.4663	70	0.9397	0.3420	2.7475
26	0.4384	0.8988	0.4877	71	0.9455	0.3256	2.9042
27	0.4540	0.8910	0.5095	72	0.9511	0.3090	3.0777
28	0.4695	0.8829	0.5317	73	0.9563	0.2924	3.2709
29	0.4848	0.8746	0.5543	74	0.9613	0.2756	3.4874
30	0.5000	0.8660	0.5774	75	0.9659	0.2588	3.7321
31	0.5150	0.8572	0.6009	76	0.9703	0.2419	4.0108
32	0.5299	0.8480	0.6249	77	0.9744	0.2250	4.3315
33	0.5446	0.8387	0.6494	78	0.9781	0.2079	4.7046
34	0.5592	0.8290	0.6745	79	0.9816	0.1908	5.1446
35	0.5736	0.8192	0.7002	80	0.9848	0.1736	5.6713
36	0.5878	0.8090	0.7265	81	0.9877	0.1564	6.3138
37	0.6018	0.7986	0.7536	82	0.9903	0.1392	7.1154
38	0.6157	0.7880	0.7813	83	0.9925	0.1219	8.1443
39	0.6293	0.7771	0.8098	84	0.9945	0.1045	9.5144
40	0.6428	0.7660	0.8391	85	0.9962	.0872	11.43
41	0.6561	0.7547	0.8693	86	0.9976	.0698	14.30
42	0.6691	0.7431	0.9004	87	0.9986	.0523	19.08
43	0.6820	0.7314	0.9325	88	0.9994	.0349	28.64
44	0.6947	0.7193	0.9657	89	0.9998	.0175	57.29

calculated in the same way, using the "rise" curve.

Fig. 11 illustrates the manner in which *C* and *R* can be used to provide a variable time constant (decay time in this example) for a VOX or break-in delay circuit. The maximum decay time depends on the voltage amount supplied initially from CR1 when *R* is set for maximum resistance. That is, *C* must be fully charged to assure the maximum decay time allowed by the values given. The higher the threshold of VR1, the greater the charging voltage needed to make K1 operate and hold during the normal decay period of *C* and *R*.

AC Circuits

The fact that alternating voltage or current is constantly changing in polarity and amplitude makes it necessary to utilize a system of units that will accommodate these variations. The terms *sine wave, phase* and *angular measure* are everyday ones. We deal also with such terms as *peak* (positive and negative), *peak-to-peak* (pk-pk), *root mean square* (rms) and *average* voltages. Table 5 contains data which will enable the amateur to change one term to another, numerically, when working the various equations encountered in routine design applications.

Fig. 12 shows one complete cycle of a sine wave – from zero to peak-positive value, return to zero, then to the peak-negative amount, and back to zero. The number of such cycles per second determines the ac frequency; e.g., 1000 cycles per second = 1000 Hz, or 1 kHz.

The instantaneous voltage or current value of an undistorted sine wave is that amount which exists at a precise point in time somewhere along the al-

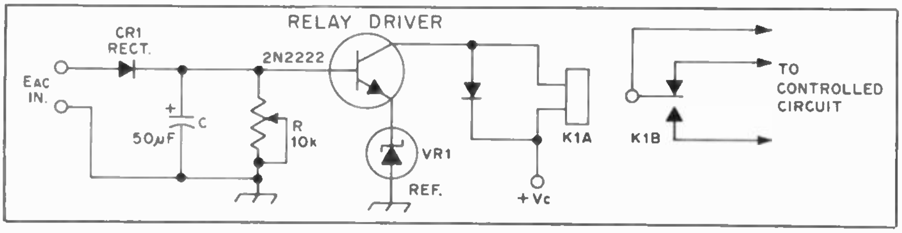


Fig. 11 – Circuit example of a variable time-constant network.

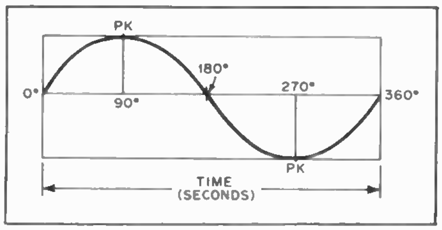


Fig. 13 – Sine-wave relationship of phase angle (one cycle) to time.

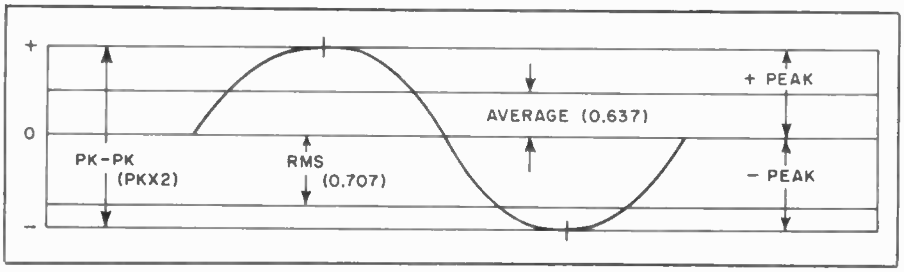


Fig. 12 – Sine-wave (one cycle) example of percentage of peak, average and rms values to pk-pk value.

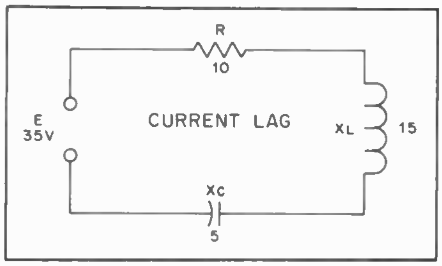


Fig. 14 – Circuit illustration relating to determination of power factor.

Table 7

AC Equivalents of E and I for Pk, RMS and Average

PEAK	RMS	AVERAGE	PEAK	RMS	AVERAGE
1	0.707	0.637	51	36.058	32.482
2	1.414	1.274	52	36.765	33.119
3	2.121	1.911	53	37.472	33.756
4	2.828	2.548	54	38.179	34.393
5	3.535	3.185	55	38.886	35.030
6	4.242	3.822	56	39.593	35.667
7	4.949	4.459	57	40.300	36.304
8	5.656	5.096	58	41.007	36.941
9	6.363	5.733	59	41.714	37.578
10	7.070	6.369	60	42.421	38.214
11	7.777	7.006	61	43.128	38.851
12	8.484	7.643	62	43.835	39.488
13	9.191	9.191	63	44.542	40.125
14	9.898	8.917	64	45.249	40.762
15	10.605	9.554	65	45.956	41.399
16	11.312	10.191	66	46.663	42.036
17	12.019	10.828	67	47.370	42.673
18	12.727	11.465	68	48.077	43.310
19	13.433	12.102	69	48.784	43.947
20	14.140	12.738	70	49.491	44.583
21	14.847	13.375	71	50.198	45.220
22	15.554	14.012	72	50.905	45.857
23	16.261	14.649	73	51.612	46.494
24	16.968	15.286	74	52.319	47.131
25	17.675	15.923	75	53.026	47.768
26	18.382	16.560	76	53.733	48.405
27	19.089	17.197	77	54.440	49.042
28	19.796	17.834	78	55.147	49.679
29	20.503	18.471	79	55.854	50.316
30	21.210	19.107	80	56.561	50.952
31	21.917	19.744	81	57.268	51.589
32	22.625	20.381	82	57.975	52.226
33	23.332	21.018	83	58.682	52.863
34	24.039	21.655	84	59.389	53.500
35	24.746	22.292	85	60.096	54.137
36	25.453	22.929	86	60.803	54.774
37	26.160	23.566	87	61.510	55.411
38	26.867	24.203	88	62.217	56.048
39	27.574	24.840	89	62.924	56.685
40	28.281	25.476	90	63.631	57.321
41	28.988	26.113	91	64.338	57.958
42	29.695	26.750	92	65.045	58.595
43	30.402	27.387	93	65.752	59.232
44	31.109	28.024	94	66.459	59.869
45	31.816	28.661	95	67.166	60.506
46	32.523	29.298	96	67.873	61.143
47	33.230	29.935	97	68.580	61.780
48	33.937	30.572	98	69.287	62.417
49	34.644	31.209	99	69.994	63.054
50	35.351	31.845	100	70.701	63.693

ternating-current wave. Because of this trait it is practical to refer to instantaneous voltage or current values in degrees, zero to 360. This is illustrated in Fig. 13. When doing electronics work, it is often necessary to know the instantaneous voltage or current value at some degree point of the sine-wave excursion. In order to perform such calculations, it is necessary to know the corresponding sine or cosine values of the angles. Those who own electronic calculators with engineering functions will have no difficulty obtaining the values needed. Others may consult the figures in Table 6 for whole-number angle sines and cosines.

Ohm's Law for AC

The procedure in making Ohm's Law calculations for ac design work is the same in principle as that for dc circuits. The primary difference is that the term *impedance* (Z) is used instead of R.

Ac values are generally given in rms amounts. Power is calculated from rms values and is normally referred to as *average power*. When ac is applied across a pure resistance (no reactive component, inductive or capacitive), the voltage and current are identical in phase relationship. Therefore, the standard Ohm's Law equations can be applied: $P = I^2R$, $P = E^2 \div R$, and $P = E \times I$.

A somewhat different situation occurs when the ac circuit includes reactance. Power is dissipated in a purely resistive circuit because the voltage and current are in phase, but when reactance is present the voltage and current are no longer in phase. Hence, when a series ac

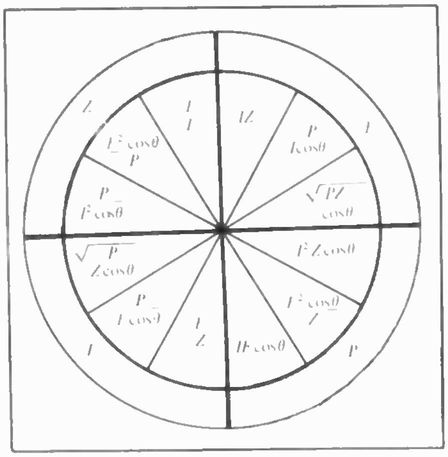


Fig. 15 — Ohm's Law wheel for ac circuits.

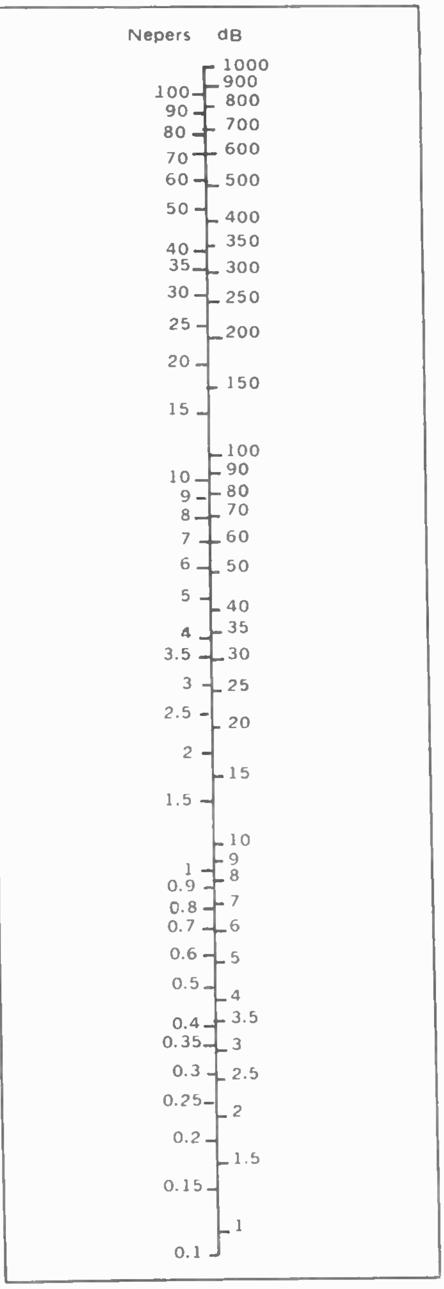


Fig. 16 — Chart for conversion of nepers to dB.

circuit contains reactance and resistance, we no longer obtain *real* or *true* power (P_r) from the calculation. Instead, a result is obtained which is known as *apparent power*, or P_a. The latter is always greater than P_r. Apparent power is generally given as a product of volt-amperes, which are numerically identical to watts. The difference between real power and apparent power is expressed as *power factor*: power factor = $P_r \div P_a$. Power factor will always be between 0 and 1. In the foregoing equation P_r is rendered in watts, and P_a is given in volt-amperes.

Power factor can be determined another way, using the phase angle in degrees, the circuit impedance, and the circuit resistance in ohms: power factor = $R \div Z$ = cosine θ. In this context the phase angle (θ) equals R divided by Z. The methods for determining the power factor of the circuit given in Fig. 14 are illustrated in the following

Table 8

Decibel Equivalents to *E*, *I* and *P* Ratios

− VOLTAGE OR CURRENT RATIO	− POWER RATIO	dB	+ VOLTAGE OR CURRENT RATIO	+ POWER RATIO
1.0000	1.0000	0	1.000	1.000
0.9886	0.9772	0.1	1.012	1.023
0.9772	0.9550	0.2	1.023	1.047
0.9661	0.9333	0.3	1.035	1.072
0.9550	0.9120	0.4	1.047	1.096
0.9441	0.8913	0.5	1.059	1.122
0.9333	0.8710	0.6	1.072	1.148
0.9226	0.8511	0.7	1.084	1.175
0.9120	0.8318	0.8	1.096	1.202
0.9016	0.8128	0.9	1.109	1.230
0.8913	0.7943	1.0	1.122	1.259
0.8810	0.7762	1.1	1.135	1.288
0.8710	0.7586	1.2	1.148	1.318
0.8610	0.7413	1.3	1.161	1.349
0.8511	0.7244	1.4	1.175	1.380
0.8414	0.7079	1.5	1.189	1.413
0.8318	0.6918	1.6	1.202	1.445
0.8222	0.6761	1.7	1.216	1.479
0.8128	0.6607	1.8	1.230	1.514
0.8035	0.6457	1.9	1.245	1.549
0.7943	0.6310	2.0	1.259	1.585
0.7852	0.6166	2.1	1.274	1.622
0.7762	0.6026	2.2	1.288	1.660
0.7674	0.5888	2.3	1.303	1.698
0.7586	0.5754	2.4	1.318	1.738
0.7499	0.5623	2.5	1.334	1.778
0.7413	0.5495	2.6	1.349	1.820
0.7328	0.5370	2.7	1.365	1.862
0.7244	0.5248	2.8	1.380	1.905
0.7161	0.5129	2.9	1.396	1.950
0.7079	0.5012	3.0	1.413	1.995
0.6998	0.4898	3.1	1.429	2.042
0.6918	0.4786	3.2	1.445	2.089
0.6839	0.4677	3.3	1.462	2.138
0.6761	0.4571	3.4	1.479	2.188
0.6683	0.4467	3.5	1.496	2.239
0.6607	0.4365	3.6	1.514	2.291
0.6531	0.4266	3.7	1.531	2.344
0.6457	0.4169	3.8	1.549	2.399
0.6383	0.4074	3.9	1.567	2.455
0.6310	0.3981	4.0	1.585	2.512
0.6237	0.3890	4.1	1.603	2.570
0.6166	0.3802	4.2	1.622	2.630
0.6095	0.3715	4.3	1.641	2.692
0.6026	0.3631	4.4	1.660	2.754
0.5957	0.3548	4.5	1.679	2.818
0.5888	0.3467	4.6	1.698	2.884
0.5821	0.3388	4.7	1.718	2.951
0.5754	0.3311	4.8	1.738	3.020
0.5689	0.3236	4.9	1.758	3.090
0.5623	0.3162	5.0	1.778	3.162
0.5559	0.3090	5.1	1.799	3.236
0.5495	0.3020	5.2	1.820	3.311
0.5433	0.2951	5.3	1.841	3.388
0.5370	0.2884	5.4	1.862	3.467
0.5309	0.2818	5.5	1.884	3.548
0.5248	0.2754	5.6	1.905	3.631
0.5188	0.2692	5.7	1.928	3.715
0.5129	0.2630	5.8	1.950	3.802
0.5070	0.2570	5.9	1.972	3.890
0.5012	0.2512	6.0	1.995	3.931
0.4955	0.2455	6.1	2.018	4.074

− VOLTAGE OR CURRENT RATIO	− POWER RATIO	dB	+ VOLTAGE OR CURRENT RATIO	+ POWER RATIO
0.4898	0.2399	6.2	2.042	4.169
0.4842	0.2344	6.3	2.065	4.266
0.4786	0.2291	6.4	2.089	4.365
0.4732	0.2239	6.5	2.113	4.467
0.4677	0.2188	6.6	2.138	4.571
0.4624	0.2138	6.7	2.163	4.677
0.4571	0.2089	6.8	2.188	4.786
0.4519	0.2042	6.9	2.213	4.898
0.4467	0.1995	7.0	2.239	5.012
0.4416	0.1950	7.1	2.265	5.129
0.4365	0.1905	7.2	2.291	5.248
0.4315	0.1862	7.3	2.317	5.370
0.4266	0.1820	7.4	2.344	5.495
0.4217	0.1778	7.5	2.371	5.623
0.4169	0.1738	7.6	2.399	5.754
0.4121	0.1698	7.7	2.427	5.888
0.4074	0.1660	7.8	2.455	6.026
0.4027	0.1622	7.9	2.483	6.166
0.3981	0.1585	8.0	2.512	6.310
0.3936	0.1549	8.1	2.541	6.457
0.3890	0.1514	8.2	2.570	6.607
0.3846	0.1479	8.3	2.600	6.761
0.3802	0.1445	8.4	2.630	6.918
0.3758	0.1413	8.5	2.661	7.079
0.3715	0.1380	8.6	2.692	7.244
0.3673	0.1349	8.7	2.723	7.413
0.3631	0.1318	8.8	2.754	7.586
0.3589	0.1288	8.9	2.786	7.762
0.3548	0.1259	9.0	2.818	7.943
0.3508	0.1230	9.1	2.851	8.128
0.3467	0.1202	9.2	2.884	8.318
0.3428	0.1175	9.3	2.917	8.511
0.3388	0.1148	9.4	2.951	8.710
0.3350	0.1122	9.5	2.985	8.913
0.3311	0.1096	9.6	3.020	9.120
0.3273	0.1072	9.7	3.055	9.333
0.3236	0.1047	9.8	3.090	9.550
0.3199	0.1023	9.9	3.126	9.772
0.3162	0.1000	10.0	3.162	10.000
0.2985	.08913	10.5	3.350	11.22
0.2818	.07943	11.0	3.548	12.59
0.2661	.07079	11.5	3.758	14.13
0.2512	.06310	12.0	3.981	15.85
0.2371	.05623	12.5	4.217	17.78
0.2239	.05012	13.0	4.467	19.95
0.2113	.04467	13.5	4.732	22.39
0.1995	.03981	14.0	5.012	25.12
0.1884	.03548	14.5	5.309	28.18
0.1778	.03162	15.0	5.623	31.62
0.1585	.02512	16.0	6.310	39.81
0.1413	.01995	17.0	7.079	50.12
0.1259	.01585	18.0	7.943	63.10
0.1122	.01259	19.0	8.913	79.43
0.1000	.01000	20.0	10.000	100.00
.03162	.00100	30.0	31.620	1,000.00
.01	.00010	40.0	100.00	10,000.00
.003162	.00001	50.0	316.20	10^5
.001	10^{-6}	60.0	1,000.00	10^6
.0003162	10^{-7}	70.0	3,162.00	10^7
.0001	10^{-8}	80.0	10,000.00	10^8
.00003162	10^{-9}	90.0	31,620.00	10^9
10^{-5}	10^{-10}	100.0	10^5	10^{10}

equations.

$$Z = \sqrt{R^2 + (X_L - X_C)^2}$$

$$= \sqrt{10^2 + (15-5)^2} = \sqrt{200}$$

$$= 14.1 \text{ ohms} \quad \text{(A)}$$

$$I = \frac{E}{Z} = \frac{35}{14.1} = 2.5 \text{ amperes} \quad \text{(B)}$$

$$P_a = E \times I = 35 \times 2.5 = 87.5 \text{ VA} \quad \text{(C)}$$

$$P_r = I^2 \times R = 2.5^2 \times 10 = 62.5 \text{ W} \quad \text{(D)}$$

$$P_f = \frac{P_r}{P_a} = \frac{62.5}{87.5} = 0.71 \quad \text{(E)}$$

(Eq. 12)

Power factor can be determined by the following procedure also

$$P_f = \frac{R}{Z} = \frac{10}{14.1} = 0.71 \quad \text{(A)}$$

$$\theta = \arctan \frac{X_L - X_C}{R}$$

$$= \arctan \frac{15-5}{10}$$

$$= \arctan 1 = 45^\circ \quad \text{(B)}$$

$$P_f = \cos\theta = \cos 45^\circ \quad \text{(C)}$$

$$= 0.71$$

(Eq. 13)

It can be seen from the foregoing that a power factor of 0.71 is obtained by either method. Table 6 can be consulted to learn the sines, cosines and tangents for angles from 0 to 89.

The method for determining power factor in a parallel circuit is essentially the same as that just treated. The primary difference is that the angle is obtained from the current. The latter is inversely proportional to the impedance (R divided by Z rather than Z divided by R). Fig. 15 contains a handy Ohm's Law wheel for ac-circuit applications.

Another aid for solving problems in this chapter and elsewhere in the book is the list of ac equivalents of E and I given in Table 7. Peak, average, and rms voltages can be compared by using the chart, thereby eliminating the need to use math where absolute precision isn't required.

Working with Decibels

The *decibel* (dB) is a measurement term which has considerable application in electronics work. Logarithmically, it expresses the ratio between two power, voltage, or current amounts. The *bel*, from which dB is derived, is the fundamental unit in a logarithmic scale for expressing the ratio of two power levels. However, such a numerical quantity is rather unwieldy for most electronics calculations, so one-tenth bel (decibel) is used more commonly. Furthermore, the human ear can more easily resolve differences in dB than in bels. The term dBm is also a common one and is used when it is desired to reference power to 1 mW – 0 dBm = 1 mW across a known resistive load, say, 50 ohms.

In some parts of the world the expression *nepers* is used instead of dB. One dB = 0.1151 neper, and one neper = 8.686 dB. The choice of terms is not unlike that of photographers using ASA and DIN ratings for film speed in different parts of the world. The nomograph of Fig. 16 can be used to convert one term to another. Each time *power* is doubled (e.g., 12 W to 24 W), an increase of 3 dB results. Therefore, it is not difficult to make quick calculations mentally. For more precise determinations

$$\text{dB} = 10 \log \frac{P_2}{P_1} = 10 \log \frac{150}{8}$$

$$= 10 \log 18.75 = 10 \times 1.27$$

$$= 12.7 \text{ dB}$$

(Eq. 14)

where P_2 = 150 W and P_1 = 8 W. P_2 is the power output from a circuit, and P_1 is the power input to the circuit. When P_1 is greater in amount than P_2 is, a power loss results, and is given as –dB. If the power levels of P_1 and P_2 were reversed in the foregoing equation, the power *difference* would be –12.7 dB. This condition is referred to as *insertion loss*.

Decibels can be determined from differences in voltage or current. When those terms are used, the multiplier becomes 20 rather than 10, thus

$$\text{dB} = 20 \log \frac{E_2}{E_1} = 20 \log \frac{76 \text{ V}}{15 \text{ V}}$$

$$= 20 \log 5.07 = 20 \times 0.705 = 14 \text{ dB} \quad \text{(A)}$$

or

$$\text{dB} = 20 \log \frac{I_2}{I_1} = 20 \log \frac{112 \text{ mA}}{65 \text{ mA}}$$

$$= 20 \log 1.72 = 20 \times 0.24 \quad \text{(B)}$$

$$= 4.7 \text{ dB}$$

(Eq. 15)

If the reader does not have an electronic calculator or slide rule available, the information in Table 1-8 will be helpful in obtaining dB from various ratios of power, voltage, or current. If precise calculations are required, the equations for determining dB can be used in combination with the common logarithms of Table 9.

Power ratios are not dependent upon source and load impedances. Conversely, *voltage* and *current ratios* in the dB equations must be based on *equal* source and load impedances. In situations where they are not identical, letting R_1 equal *source* impedance, and R_2 equal *load* impedance, the following equations are applicable

$$\text{dB} = 20 \log \frac{I_1\sqrt{R_1}}{I_2\sqrt{R_2}}$$

$$= 20 \log \frac{3\text{A}\sqrt{56 \text{ ohms}}}{0.5\text{A}\sqrt{12 \text{ ohms}}}$$

$$= 20 \log \frac{22.4}{1.7} = 20 \log 12.9$$

$$= 20 \times 1.11 = 22.2 \text{ dB} \quad \text{(A)}$$

or

$$\text{dB} = 20 \log \frac{E_1\sqrt{R_2}}{E_2\sqrt{R_1}}$$

$$= 20 \log \frac{68\text{V}\sqrt{300 \text{ ohms}}}{17\text{V}\sqrt{50 \text{ ohms}}}$$

$$= 20 \log \frac{1178}{120} = 20 \log 9.8$$

$$= 20 \times 0.991 = 19.8 \text{ dB} \quad \text{(B)}$$

(Eq. 16)

It should be remembered that in the examples E_1 and R_2 are always greater in amount than E_2 and R_1, and I_1 and R_1 are always larger than I_2 and R_2.

VU Units

Earlier there was mention of dBm as a unit of measurement. The term uses one milliwatt (mW) as the reference. One variation of this principle is the application of VU, or *volume units*. Zero VU is equal to 1 mW (.001 W) across a 600-ohm load. Audio equipment is frequently complemented by a meter which reads VU directly. Since P_1 (reference amount) is always 1 mW (10^{-3}), the following equation is applicable

$$\text{VU} = 10 \log \frac{P_2}{10^{-3}} = 10 \log \frac{60 \text{ mW}}{.001}$$

$$= 10 \log 60{,}000 = 10 \times 4.78$$

$$= 47.8 \text{ VU} \quad \text{(A)}$$

or simplified

$$\text{VU} = 30 + 10 \log P_2$$

$$= 30 + 10 \log 60 \text{ mW}$$

$$= 30 + 17.8 = 47.8 \text{ VU} \quad \text{(B)}$$

(Eq. 17)

Volume units are not used for most amateur work, but are found in some speech-processing equipment in the more sophisticated amateur stations.

References

The Radio Amateur's Handbook, chapter 2.
Grammer, *A Course in Radio Fundamentals*, ARRL, current edition.
Understanding Amateur Radio, chapter 2.

Table 9

Common Logarithms

N	*0*	*1*	*2*	*3*	*4*	*5*	*6*	*7*	*8*	*9*	*N*
10	0000	0043	0086	0128	0170	0212	0253	0294	0334	0374	10
11	0414	0453	0492	0531	0569	0607	0645	0682	0719	0755	11
12	0792	0828	0864	0899	0934	0969	1004	1038	1072	1106	12
13	1139	1173	1206	1239	1271	1303	1335	1367	1399	1430	13
14	1461	1492	1523	1553	1584	1614	1644	1673	1703	1732	14
15	1761	1790	1818	1847	1875	1903	1931	1959	1987	2014	15
16	2041	2068	2095	2122	2148	2175	2201	2227	2253	2279	16
17	2304	2330	2355	2380	2405	2430	2455	2480	2504	2529	17
18	2553	2577	2601	2625	2648	2672	2695	2718	2742	2765	18
19	2788	2810	2833	2856	2878	2900	2923	2945	2967	2989	19
20	3010	3032	3054	3075	3096	3118	3139	3160	3181	3201	20
21	3222	3243	3263	3284	3304	3324	3345	3365	3385	3404	21
22	3424	3444	3464	3483	3502	3522	3541	3560	3579	3598	22
23	3617	3636	3655	3674	3692	3711	3729	3747	3766	3784	23
24	3802	3820	3838	3856	3874	3892	3909	3927	3945	3962	24
25	3979	3997	4014	4031	4048	4065	4082	4099	4116	4133	25
26	4150	4166	4183	4200	4216	4232	4249	4265	4281	4298	26
27	4314	4330	4346	4362	4378	4393	4409	4425	4440	4456	27
28	4472	4487	4502	4518	4533	4548	4564	4579	4594	4609	28
29	4624	4639	4654	4669	4683	4698	4713	4728	4742	4757	29
30	4771	4786	4800	4814	4829	4843	4857	4871	4886	4900	30
31	4914	4928	4942	4955	4969	4983	4997	5011	5024	5038	31
32	5051	5065	5079	5092	5105	5119	5132	5145	5159	5172	32
33	5185	5198	5211	5224	5237	5250	5263	5276	5289	5302	33
34	5315	5328	5340	5353	5366	5378	5391	5403	5416	5428	34
35	5441	5453	5465	5478	5490	5502	5514	5527	5539	5551	35
36	5563	5575	5587	5599	5611	5623	5635	5647	5658	5670	36
37	5682	5694	5705	5717	5729	5740	5752	5763	5775	5786	37
38	5798	5809	5821	5832	5843	5855	5866	5877	5888	5899	38
39	5911	5922	5933	5944	5955	5966	5977	5988	5999	6010	39
40	6021	6031	6042	6053	6064	6075	6085	6096	6107	6117	40
41	6128	6138	6149	6160	6170	6180	6191	6201	6212	6222	41
42	6232	6243	6253	6263	6274	6284	6294	6304	6314	6325	42
43	6335	6345	6355	6365	6375	6385	6395	6405	6415	6425	43
44	6435	6444	6454	6464	6474	6484	6493	6503	6513	6522	44
45	6532	6542	6551	6561	6571	6580	6590	6599	6609	6618	45
46	6628	6637	6646	6656	6665	6675	6684	6693	6702	6712	46
47	6721	6730	6739	6749	6758	6767	6776	6785	6794	6803	47
48	6812	6821	6830	6839	6848	6857	6866	6875	6884	6893	48
49	6902	6911	6920	6928	6937	6946	6955	6964	6972	6981	49
50	6990	6998	7007	7016	7024	7033	7042	7050	7059	7067	50
51	7076	7084	7093	7101	7110	7118	7126	7135	7143	7152	51
52	7160	7168	7177	7185	7193	7202	7210	7218	7226	7235	52
53	7243	7251	7259	7267	7275	7284	7292	7300	7308	7316	53

Table 10

GREEK NAME	*CAPITAL*	*LOWER CASE*	*DEFINITION*
Alpha	Α	α	Angles, area, coefficients, attenuation constant, absorption factor and the current gain of common-base transistor amplifiers.
Beta	Β	β	Angles, flux density, phase constant and current gain of common-emitter transistor amplifiers.
Gamma	Γ	γ	Angles, conductivity and specific gravity.
Delta	Δ	δ	Variation, density and angles.
Epsilon	Ε	ϵ	Base of natural logarithms and electric intensity.
Zeta	Ζ	ζ	Impedance, coefficients and coordinates.
Eta	Η	η	Hysteresis coefficients, efficiency and surface charge density.
Theta	Θ	θ	Temperature, phase angle, time constant, reluctance and angles.
Iota	Ι	ι	Unit vector.
Kappa	Κ	κ	Dielectric constant, susceptibility.
Lambda	Λ	λ	Wavelength. Attenuation constant.

GREEK SYMBOLS

N	0	1	2	3	4	5	6	7	8	9	N
54	7324	7332	7340	7348	7356	7364	7372	7380	7388	7396	54
55	7404	7412	7419	7427	7435	7443	7451	7459	7466	7474	55
56	7482	7490	7497	7505	7513	7520	7528	7536	7543	7551	56
57	7559	7566	7574	7582	7589	7597	7604	7612	7619	7627	57
58	7634	7642	7649	7657	7664	7672	7679	7686	7694	7701	58
59	7709	7716	7723	7731	7738	7745	7752	7760	7767	7774	59
60	7782	7789	7796	7803	7810	7818	7825	7832	7839	7846	60
61	7853	7860	7868	7875	7882	7889	7896	7903	7910	7917	61
62	7924	7931	7938	7945	7952	7959	7966	7973	7980	7987	62
63	7993	8000	8007	8014	8021	8028	8035	8041	8048	8055	63
64	8062	8069	8075	8082	8089	8096	8102	8109	8116	8122	64
65	8129	8136	8142	8149	8156	8162	8169	8176	8182	8189	65
66	8195	8202	8209	8215	8222	8228	8235	8241	8248	8254	66
67	8261	8267	8274	8280	8287	8293	8299	8306	8312	8319	67
68	8325	8331	8338	8344	8351	8357	8363	8370	8376	8382	68
69	8388	8395	8401	8407	8414	8420	8426	8432	8439	8445	69
70	8451	8457	8463	8470	8476	8482	8488	8494	8500	8506	70
71	8513	8519	8525	8531	8537	8543	8549	8555	8561	8567	71
72	8573	8579	8585	8591	8597	8603	8609	8615	8621	8627	72
73	8633	8639	8645	8651	8657	8663	8669	8675	8681	8686	73
74	8692	8698	8704	8710	8716	8722	8727	8733	8739	8745	74
75	8751	8756	8762	8768	8774	8779	8785	8791	8797	8802	75
76	8808	8814	8820	8825	8831	8837	8842	8848	8854	8859	76
77	8865	8871	8876	8882	8887	8893	8899	8904	8910	8915	77
78	8921	8927	8932	8938	8943	8949	8954	8960	8965	8971	78
79	8976	8982	8987	8993	8998	9004	9009	9015	9020	9025	79
80	9031	9036	9042	9047	9053	9058	9063	9069	9074	9079	80
81	9085	9090	9096	9101	9106	9112	9117	9122	9128	9133	81
82	9138	9143	9149	9154	9159	9165	9170	9175	9180	9186	82
83	9191	9196	9201	9206	9212	9217	9222	9227	9232	9238	83
84	9243	9248	9253	9258	9263	9269	9274	9279	9284	9289	84
85	9294	9299	9304	9309	9315	9320	9325	9330	9335	9340	85
86	9345	9350	9355	9360	9365	9370	9375	9380	9385	9390	86
87	9395	9400	9405	9410	9415	9420	9425	9430	9435	9440	87
88	9445	9450	9455	9460	9465	9469	9474	9479	9484	9489	88
89	9494	9499	9504	9509	9513	9518	9523	9528	9533	9538	89
90	9542	9547	9552	9557	9562	9566	9571	9576	9581	9586	90
91	9590	9595	9600	9605	9609	9614	9619	9624	9628	9633	91
92	9638	9643	9647	9652	9657	9661	9666	9671	9675	9680	92
93	9685	9689	9694	9699	9703	9708	9713	9717	9722	9727	93
94	9731	9736	9741	9745	9750	9754	9759	9763	9768	9773	94
95	9777	9782	9786	9791	9795	9800	9805	9809	9814	9818	95
96	9823	9827	9832	9836	9841	9845	9850	9854	9859	9863	96
97	9868	9872	9877	9881	9886	9890	9894	9899	9903	9908	97
98	9912	9917	9921	9926	9930	9934	9939	9943	9948	9952	98
99	9956	9961	9965	9969	9974	9978	9983	9987	9991	9996	99

GREEK NAME	CAPITAL	LOWER CASE	DEFINITION
Mu	M	μ	Micro, amplification factor and permeability.
Nu	N	ν	Reluctivity. Frequency.
Xi	Ξ	ξ	Coordinates.
Omicron	O	o	—
Pi	Π	π	3.1416 (ratio of circumference to diameter).
Rho	P	ρ	Resistivity. Coordinates.
Sigma	Σ	σ	Summation (u.c.), electrical conductivity, leakage coefficient, surface charge density and complex propagation constant.
Tau	T	τ	Time constant, time phase displacement, density and transmission factor.
Upsilon	Υ	υ	—
Phi	Φ	ϕ	Magnetic flux, angles and scalar potential (u.c.).
Chi	X	χ	Electric susceptibility. Angles.
Psi	Ψ	ψ	Dielectric flux, phase difference, coordinates and angles.
Omega	Ω	ω	Angular velocity, resistance in ohms (u.c.) and solid angles (u.c.).

Chapter 2

Time and Frequency

The wave shown in Fig. 13 of chapter 1 represents one complete cycle of voltage or current over a period of one second.

Table 1

MHz	FEET PER WAVE-LENGTH	INCHES
1300	0	9.0
1215	0	9.7
450	2	2.8
432	2	3.6
420	2	4.0
225	4	4.4
222.5	4	5.0
220	4	5.6
148	6	7.8
147	6	8.4
146	6	8.7
145	6	9.5
144	6	10.0
54	18	2.4
53	18	6.8
52	18	11.0
51	19	3.4
50	19	8.4
29.7	33	1.0
29.0	33	10.8
28.5	34	6.0
28.0	35	1.0
21.45	45	9.0
21.35	46	0
21.25	46	3.6
21.10	46	7.0
21.00	46	9.6
14.35	68	7.0
14.25	69	0
14.10	69	9.0
14.00	70	3.0
7.3	135	0
7.2	136	7.0
7.1	138	7.0
7.0	140	7.0
4.0	246	0
3.9	252	0
3.8	259	0
3.7	266	0
3.6	273	0
3.5	281	0
2.0	492	0
1.9	518	0
1.8	546	7.0

The figures in this table are based on *L* (length in feet) = 984/*f* MHz, the free-space dimension.

It is a *periodic* wave, which means it changes its direction of flow at regular time intervals, reaching equal and alternate positive and negative values. The wavelength or frequency of the ac voltage or current is based on the number of cycles per second (hertz). Thus, one cycle = one hertz (Hz). Similarly, 1000 cycles = one kilohertz (kHz) and 1,000,000 cycles per second = one megahertz (MHz). As the number of Hz increases the wavelength decreases; e.g., 1800 kHz = 166.6 meters (300,000 ÷ 1800 kHz), and 455 kHz = 659 meters. Therefore

$$\lambda\,(\text{meters}) = \frac{300{,}000}{f(\text{kHz})}$$

$$\therefore \lambda = \frac{300{,}000}{455} = 659 \text{ meters} \qquad \text{(A)}$$

or

$$\lambda\,(\text{meters}) = \frac{300}{f(\text{MHz})}$$

$$\therefore \lambda = \frac{300}{0.455} = 659 \text{ meters} \qquad \text{(B)}$$

(Eq. 1)

Radio waves travel at the *same speed as light* – 300,000,000 meters (approximately 186,000 miles) per second. They can be set up by a radio-frequency current flowing in a circuit, because the rapidly changing current creates a magnetic field that changes in a like fashion. The varying magnetic field in turn sets up a varying electric field. Whenever that condition is met, the two fields surge outward at the speed of light.

One can find the number of feet versus wavelength by employing the following simple equation:

$$\lambda\,(\text{feet}) = \frac{984}{f(\text{MHz})}$$

$$\therefore \text{feet} = \frac{984}{1.9} = 518 \qquad \text{(Eq. 2)}$$

and

$$f\,(\text{MHz}) = \frac{984}{\lambda\,(\text{feet})}$$

$$\therefore f = \frac{984}{518} = 1.9 \text{ MHz} \qquad \text{(Eq. 3)}$$

where f = 1.9 MHz.

Wavelength can be converted to frequency from the following:

$$f\,(\text{kHz}) = \frac{300{,}000}{\lambda\,(\text{meters})}$$

$$\therefore f = \frac{300{,}000}{659} = 455 \text{ kHz} \qquad \text{(Eq. 4)}$$

where wavelength = 659 meters.

Time Period

The frequency of an alternating voltage or current has a *time duration* respective to the number of waves. In this context, an audio wave whose frequency is, say, 1500 Hz, has 1500

Table 2

BAND	FREQUENCY (GHz)	WAVELENGTH (cm)
P	0.225 - 0.390	133.3 - 76.9
L	0.390 - 1.550	76.9 - 19.3
S	1.550 - 5.20	19.3 - 5.77
X	5.20 - 10.90	5.77 - 2.75
K	10.90 - 36.00	2.75 - 0.834
Q	36.00 - 46.00	0.834 - 0.652
V	46.00 - 56.00	0.652 - 0.536
W	56.00 - 100.0	0.536 - 0.300

Table of band designators versus frequency in GHz and cm.

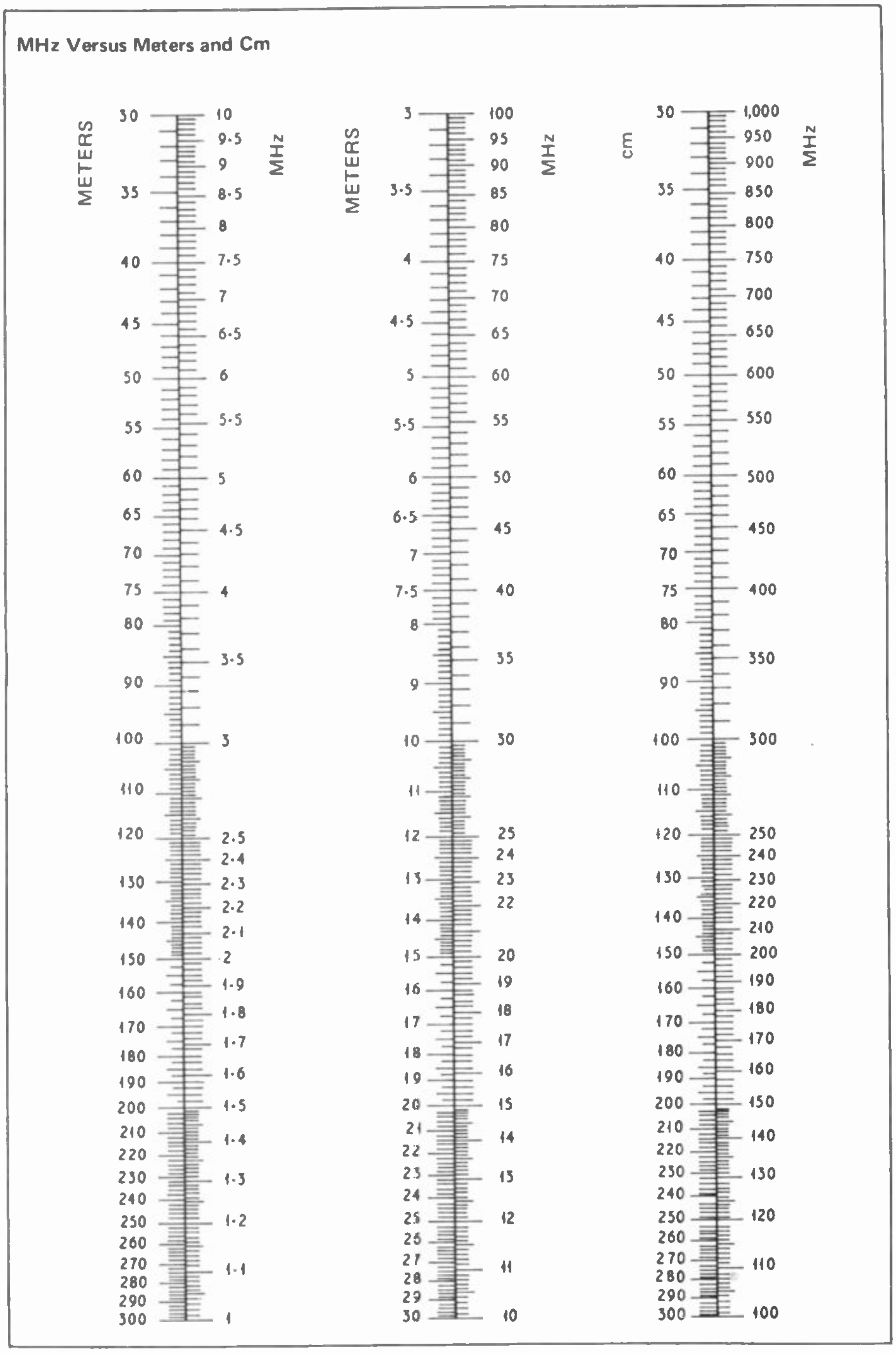


Fig. 1 — Nomograph of frequency in MHz and cm versus meters.

complete cycles per second. The time period for a single cycle would be $T = 1/1500 = .00066$ second, where T is in seconds and f is in Hz, or more simply $T = 1/\text{Hz}$.

Wave Velocity

The *velocity* of a wave can be defined as the distance it covers in a specific amount of time. Because the velocity is identical for all frequencies, the following mathematical relationships hold:

$$V = \lambda f \therefore f = \frac{V}{\lambda}, \text{ and } \lambda = \frac{V}{f} \qquad \text{(Eq. 5)}$$

Alternatively, velocity can be expressed as $V = D/T$, where distance is typically in miles in amateur work, and time is in seconds.

Angular Velocity

In radio work the expression *angular velocity* (Greek symbol omega, ω) is used in a number of equations. A complete ac cycle contains 360 degrees, as illustrated in Fig. 13, chapter 1. There are 2π radians in each 360 degrees. The angular velocity can be determined by

$$\omega = 2\pi \times f$$
$$\therefore \omega = 6.28 \times 1500 \approx 9425 \text{ radians per second} \qquad \text{(Eq. 6)}$$

where $f = 1500$ Hz.

Table 1 gives dimensions in feet and inches of one wavelength for the amateur frequencies between 1.8 and 1300 MHz. One-half wavelength dimensions can be obtained by dividing the feet by 2, and one-quarter wavelength figures can be had when dividing the feet listed by 4.

The nomograph of Fig. 1 can be used to compare frequency in MHz to meters and centimeters. It provides coverage from 1 to 1000 MHz, 3 to 500 meters, and 30 to 300 centimeters.

Table 2 shows the frequency in Gigahertz (GHz) versus *band designators* for the frequency spread from 0.225 to 100 GHz. The wavelengths in centimeters are provided also.

U. S. TV Channel Frequencies

In amateur work it is helpful to know the visual and aural frequencies of the various *TV channels*. The data are useful in identifying spurious-response causes in receivers, and for establishing harmonic relationships between amateur transmitters and TV channels. Channel bandwidths are 6 MHz, and the visual (picture) carrier is 1.25 MHz above the low-frequency limit of each channel. The aural (sound) carrier is 0.25 MHz below the high-frequency edge of each channel and is 4.5 MHz above the visual-carrier frequency. Table 3 con-

Fig. 2 — Classification of the spectrum from 10 kHz to 30,000 MHz.

CLASSIFI-CATION	FREQ. SPREAD	TERI
Very low freq. (Vlf)	10 to 30	kHz
Low freq. (Lf)	30 to 300	kHz
Medium freq. (Mf)	300 to 3000	kHz
High freq. (Hf)	3 to 30	MHz
Very high freq. (Vhf)	30 to 300	MHz
Ultra high freq. (Uhf)	300 to 3000	MHz
Super high freq. (Shf)	3000 to 30,000	MHz

Table 3

	CHANNEL (NO.)	FREQUENCY (MHz)	VISUAL CARRIER (MHz)	AURAL CARRIER (MHz)
	1		(Not assigned)	
VHF	2	54-60	55.25	59.75
VHF	3	60-66	61.25	65.75
VHF	4	66-72	67.25	71.75
	5	76-82	77.25	81.75
	6	82-88	83.25	87.75
	7	174-180	175.25	179.75
	8	180-186	181.25	185.75
	9	186-192	187.25	191.75
	10	192-198	193.25	197.75
	11	198-204	199.25	203.75
	12	204-210	205.25	209.75
	13	210-216	211.25	215.75
UHF	14	470-476	471.25	475.75
UHF	15	476-482	477.25	481.75
UHF	16	482-488	483.25	487.75
	17	488-494	489.25	493.75
	18	494-500	495.25	499.75
	19	500-506	501.25	505.75
	20	506-512	507.25	511.75
	21	512-518	513.25	517.75
	22	518-524	519.25	523.75
	23	524-530	525.25	529.75
	24	530-536	531.25	535.75
	25	536-542	537.25	541.75
	26	542-548	543.25	547.75
	27	548-554	549.25	553.75
	28	554-560	555.25	559.75
	29	560-566	561.25	565.75
	30	566-572	567.25	571.75
	31	572-578	573.25	577.75
	32	578-584	579.25	583.75
	33	584-590	585.25	589.75
	34	590-596	591.25	595.75
	35	596-602	597.25	601.75
	36	602-608	603.25	607.75
	37	608-614	609.25	613.75
	38	614-620	615.25	619.75
	39	620-626	621.25	625.75
	40	626-632	627.25	631.75
	41	632-638	633.25	637.75
	42	638-644	639.25	643.75
	43	644-650	645.25	649.75
	44	650-656	651.25	655.75
	45	656-662	657.25	661.75
	46	662-668	663.25	667.75
	47	668-674	669.25	673.75
	48	674-680	675.25	679.75
	49	680-686	681.25	685.75
	50	686-692	687.25	691.75

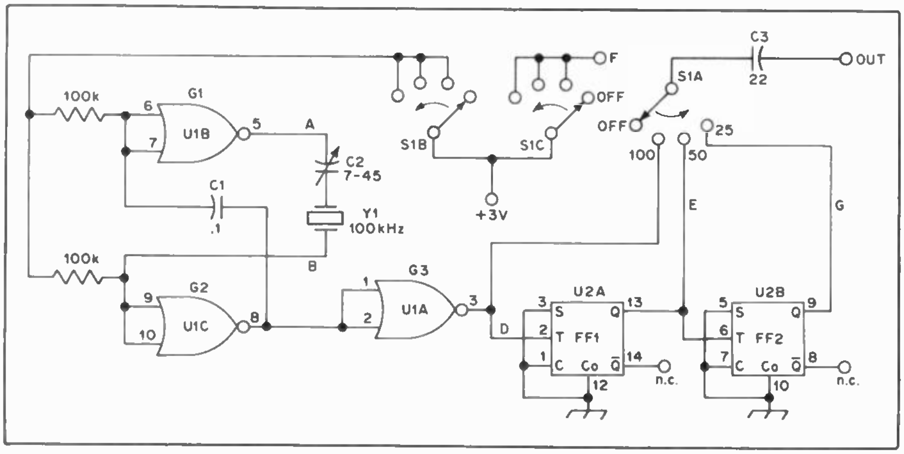


Fig. 3 — Typical circuit of a 100-kHz frequency standard with dividers to provide output at 50- and 25-kHz increments. Constructional details for this kind of equipment can be found in the *Handbook.*

tains a listing for vhf and uhf TV channels.

Spectral Nomenclature

The radio spectrum is broken up into numerous frequency classes. Fig. 2 contains the standard nomenclature applied to the band segments.

Time and Frequency

Accurate time and frequency is important to amateur designers and operators, just as it is to the commercial interests concerned with electronics. In order to maintain universal accuracy of time and frequency, the U.S. and Canada provide regular broadcasts of information which can be used as *standards* of frequency and time. The U.S. stations are WWV at Ft. Collins, Colorado, and WWVH, Kekaha Kauai, Hawaii. Canada maintains its station, CHU, at Ottawa, under the auspices of the National Research Council of Canada. The U.S. stations are operated by the National Bureau of Standards.

Standard Frequency Sources

There are four specific kinds of frequency sources for accurate calibration:

1) Crystal-controlled oscillators with temperature control, and sometimes with electronic divider circuits to provide subdigits of the fundamental frequency; e.g., 100-kHz oscillator with markers at 100, 50, and 25 kHz (and sometimes 10 kHz), Fig. 3.

2) Crystal oscillators locked on an atomic transition frequency.

3) Atomic-beam devices which employ cesium or thallium.

4) Gas-cell devices, most of which utilize rubidium.

Detailed information on the foregoing can be obtained from the ITT book, *Reference Data for Radio Engineers,* recent editions.

There are four recognized standards of time, which are treated in detail in the foregoing reference. They are *Ephemeris, Atomic, Sidereal* (ϕ), and *Universal-Time Scales* (UT).

Fig. 4 gives pertinent data for the WWV and WWVH broadcasts. The operating frequencies are listed in Fig. 4.

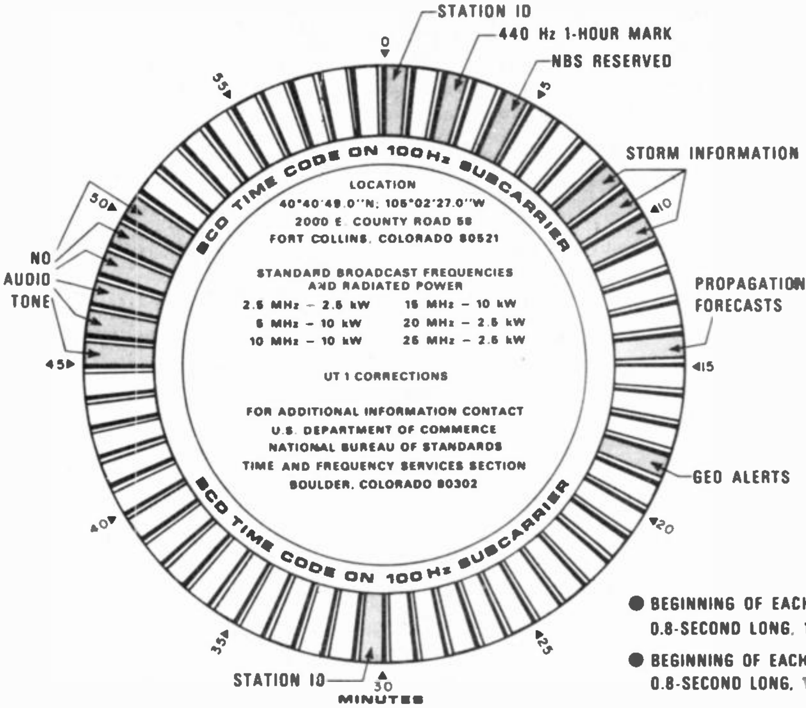
CHU broadcasts can be heard on 3330, 7335 and 14,670 kHz. Detailed information concerning CHU broadcasts can be found in the Time Service Bulletins which are available from the National Research Council of Canada.

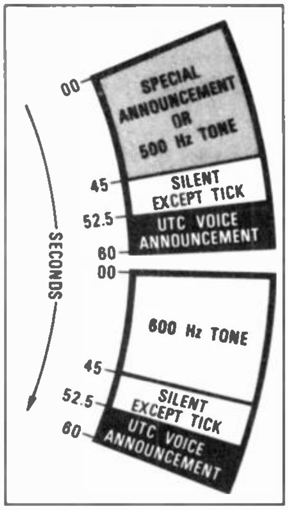
Fig. 5 provides basic information regarding the CHU broadcasts.

Measurement of Frequency

Most amateurs use a 100-kHz secondary frequency standard of the type

WWV BROADCAST FORMAT

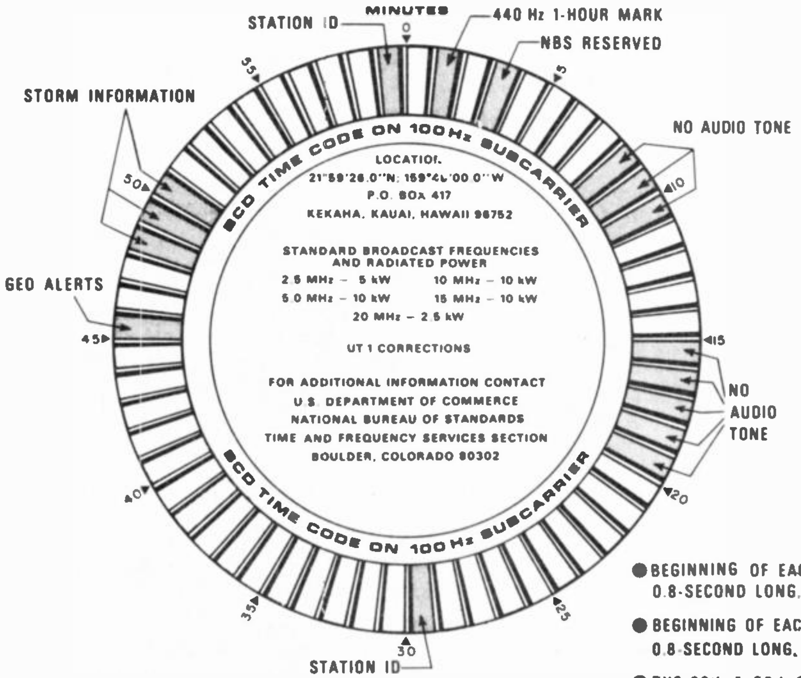


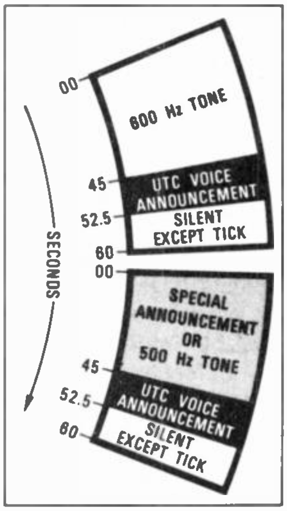


- BEGINNING OF EACH HOUR IS IDENTIFIED BY 0.8-SECOND LONG, 1500-Hz TONE.
- BEGINNING OF EACH MINUTE IS IDENTIFIED BY 0.8-SECOND LONG, 1000-Hz TONE.
- THE 29th & 59th SECOND PULSE OF EACH MINUTE IS OMITTED

VIA TELEPHONE 808–335–4363 NOT A TOLL FREE NUMBER)

WWVH BROADCAST FORMAT





- BEGINNING OF EACH HOUR IS IDENTIFIED BY 0.8-SECOND LONG, 1500-Hz TONE.
- BEGINNING OF EACH MINUTE IS IDENTIFIED BY 0.8-SECOND LONG, 1200-Hz TONE.
- THE 29th & 59th SECOND PULSE OF EACH MINUTE IS OMITTED.

Fig. 4 — The hourly broadcast schedules of WWV and WWVH.

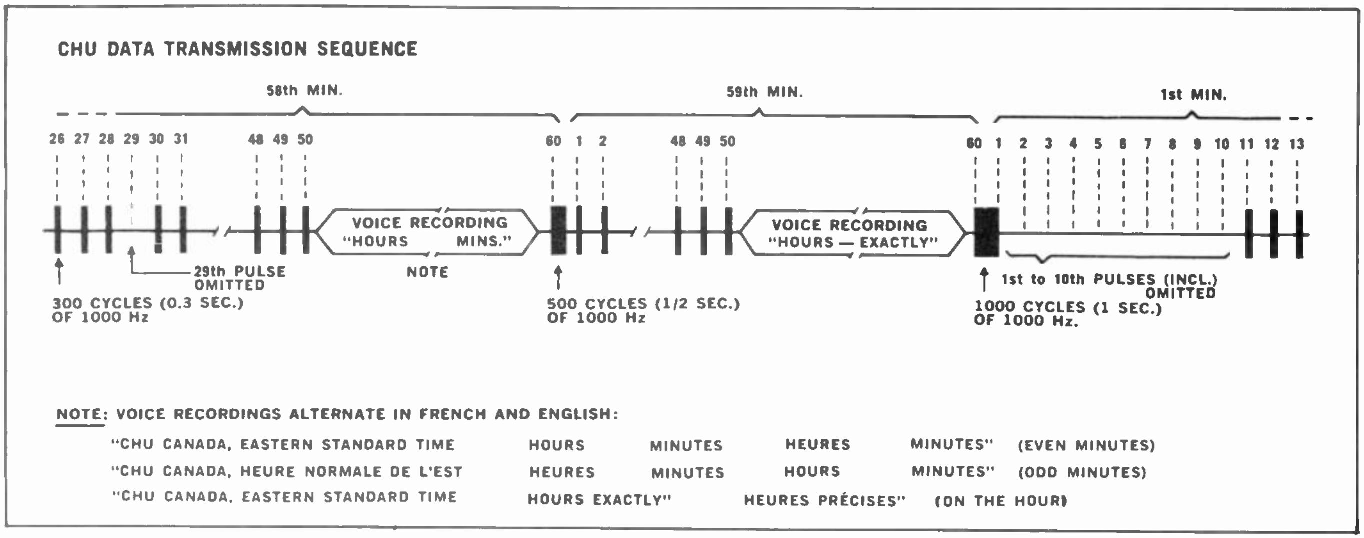


Fig. 5 — Transmission information for Canadian time/frequency station CHU.

illustrated in Fig. 3, or by means of a modern frequency counter with digital readout. Whichever of the two employed, it is wise to check its calibration periodically against WWV or WWVH by zero beating its oscillator against the carrier of the standard station.

Course frequency measurement can be accomplished by way of an absorption wavemeter. Fig. 6 shows a simple circuit which is adequate for amateur use. Constructional information on circuits of this kind can be found in the *Handbook* chapter on measurements.

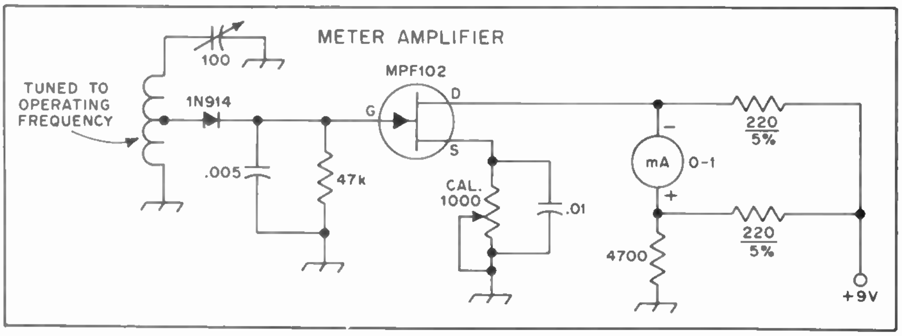


Fig. 6 — Illustrative diagram of an absorption wavemeter for rough frequency measurement in amateur work. Construction information on this kind of circuit can be found in the *Handbook*.

Chapter 3

Radio-Frequency Circuit Data

This chapter treats rf circuits and some of the terms which pertain to them. Greater elaboration of some of the matters discussed here can be found in the *Handbook, Understanding Amateur Radio*, and *A Course in Radio Fundamentals*, all of which are published by the ARRL.

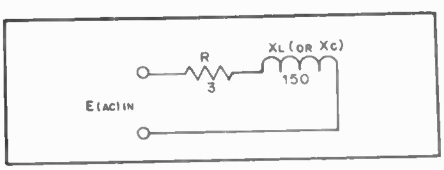
Admittance

Admittance is a common term which expresses mathematically the relative ease with which a circuit permits the flow of alternating current. Although it is used mostly in connection with solid-state circuit design, it applies to any circuit element which is part of a series ac circuit. The equation for determining the value of admittance is

$$Y = \frac{1}{\sqrt{R^2 + X^2}} \quad \therefore \; Y = \frac{1}{\sqrt{3^2 + 150^2}}$$

$$= \frac{1}{\sqrt{9 + 22{,}500}} = \frac{1}{150}$$

$$= .00666 \text{ mho} \qquad \text{(Eq. 1)}$$



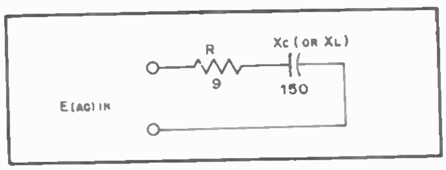
where Y = admittance, R = resistance in ohms, and X = reactance in ohms. Admittance is the reciprocal of impedance, and since impedance is expressed in ohms, admittance is therefore given in mhos. In solid-state circuit work admittance is commonly expressed as a *Y parameter*. Admittance, as the reciprocal of impedance, is expressed as $Y = 1/Z$.

Susceptance

Susceptance is also expressed in mhos. It is the reciprocal of reactance (X). In a series circuit, susceptance (B) can be determined mathematically by

$$B = \frac{-X}{R^2 + X^2} \quad \therefore \; B = \frac{150}{3^2 + 150^2}$$

$$= \frac{150}{9 + 22{,}500} = .00666 \text{ mho} \qquad \text{(Eq. 2)}$$



where B = susceptance, R = ohms, and X = reactance in ohms. If the circuit resistance happens to be zero, one can regard the susceptance as purely the reciprocal of reactance, thus $B = 1/X$. Generally speaking, B will be reactive mhos.

Conductance

Conductance (G) in a dc circuit is the reciprocal of resistance, or $G = 1/R$. In an ac circuit conductance can be regarded as resistive mhos, and is given mathematically as

$$G = \frac{R}{R^2 + X^2} \quad \therefore \; G = \frac{3}{3^2 + 150^2}$$

$$= \frac{3}{22{,}509} = .00013 \text{ resistive mho} \qquad \text{(Eq. 3)}$$

where R = 3 ohms and X = 150 ohms reactive, L or C. The terms conductance and susceptance can be related to resistance by $R = G \div [G^2 + B^2]$. Reactance can be related to conductance and susceptance by $X = B \div [G^2 + B^2]$. In this chapter conductance will be of primary concern in terms of coil wire and its dc resistance versus quality factor (Q) of a tuned circuit.

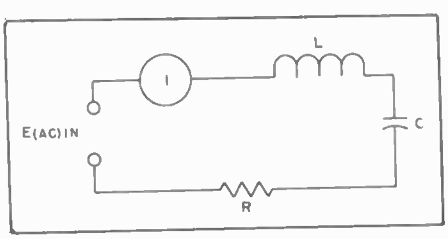


Fig. 1 — Schematic representation of a series-resonant circuit.

Resonance

The circuit of Fig. 1 shows the components of L, R and C connected in series with a voltage E_{ac} which is connected to the circuit. This represents a *series-resonant* condition. When the reactance of C is equal to that of L at the applied frequency, E_{ac}, the circuit is *resonant*. At such time the reactances are 180 degrees opposite one another in phase. Therefore, they cancel each other completely and the current flow is determined entirely by resistance R. At a frequency *lower* than resonance, the X_C will be much larger than the X_L. Therefore, the X_L will be small compared to the resistance of R and X_C.

At some frequency considerably higher than that at resonance, X_C will be very small and X_L will be large. In either situation the current will be low in value, owing to the large amount of net reactance.

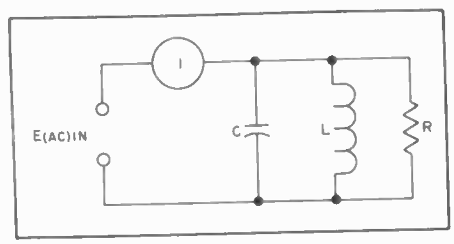


Fig. 2 — Schematic diagram of a parallel-resonant circuit.

Parallel Resonance

Fig. 2 illustrates the configuration for a *parallel-resonant* circuit. The condition of resonance is similar to that of a series-resonant circuit (Fig. 1). In this

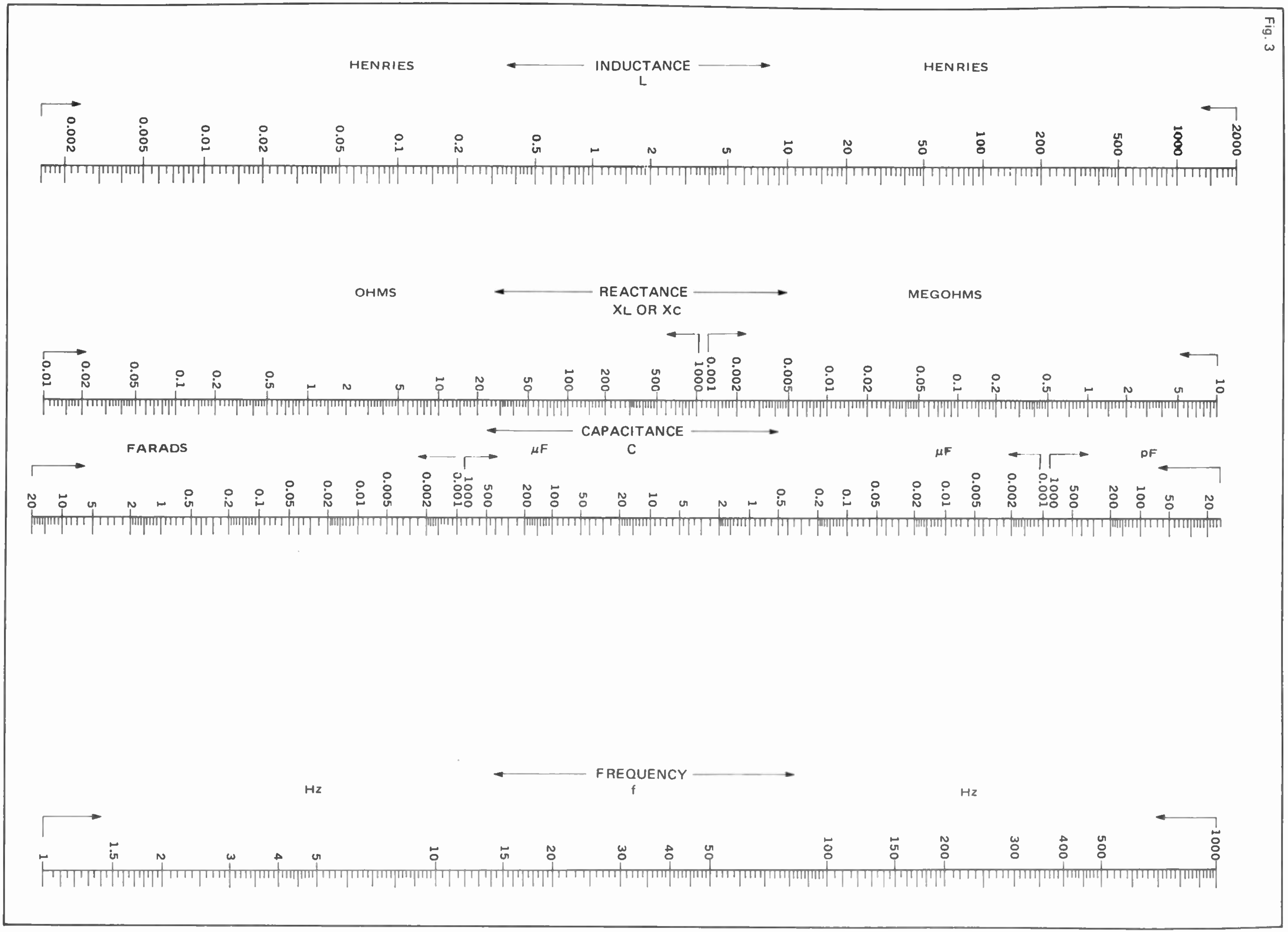

HENRIES
INDUCTANCE
L
HENRIES
OHMS
REACTANCE
XL OR XC
MEGOHMS
CAPACITANCE
FARADS
µF
C
µF
pF
FREQUENCY
Hz
f
Hz

Fig. 3

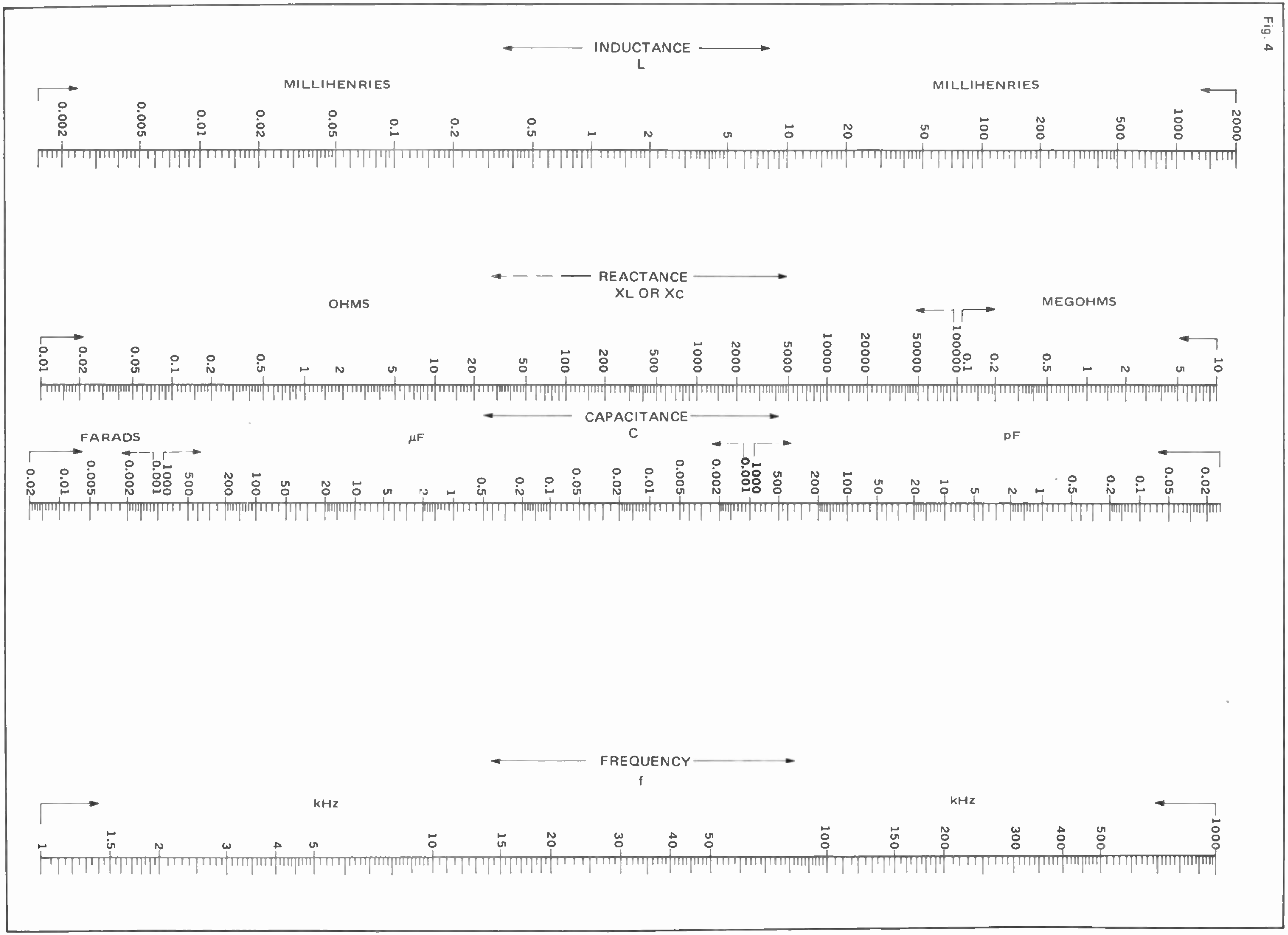

INDUCTANCE
L
MILLIHENRIES
MILLIHENRIES
0.002 0.005 0.01 0.02 0.05 0.1 0.2 0.5 1 2 5 10 20 50 100 200 500 1000 2000
REACTANCE
XL OR Xc
OHMS
MEGOHMS
0.01 0.02 0.05 0.1 0.2 0.5 1 2 5 10 20 50 100 200 500 1000 2000 5000 10000 20000 50000 100000 0.1 0.2 0.5 1 2 5 10
CAPACITANCE
C
FARADS
μF
pF
0.02 0.01 0.005 0.002 0.001 1000 500 200 100 50 20 10 5 2 1 0.5 0.2 0.1 0.05 0.02 0.01 0.005 0.002 0.001 1000 500 200 100 50 20 10 5 2 1 0.5 0.2 0.1 0.05 0.02
FREQUENCY
f
kHz
kHz
1 1.5 2 3 4 5 10 15 20 30 40 50 100 150 200 300 400 500 1000
Fig. 4

Table 1

WIRE GAUGE AWG OR B&S	TURNS PER INCH	WIRE GAUGE AWG OR B&S	TURNS PER INCH
8	7.6	24	46.3
9	8.6	25	51.7
10	9.6	26	58
11	10.7	27	64.9
12	12	28	72.7
13	13.5	29	81.6
14	15	30	90.5
15	16.8	31	101
16	18.9	32	113
17	21.2	33	127
18	23.6	34	143
19	26.4	35	158
20	29.4	36	175
21	33.1	37	198
22	37	38	224
23	41.3	39	248

example, however, the current measured at the point indicated in the drawing is the *least* when X_C and X_L are equal. At resonance the current through L is completely canceled by the out-of-phase (180 degrees) current through C. Thus, the current taken by R is all that flows in the line. Typically, the resistance R will not be an actual resistor, although in some circuit designs it will be one. R is more generally considered to be the resistance of the coil wire which has been transformed into an equivalent parallel resistance. It can also be the antenna or other load resistance *coupled into* the tuned circuit. Therefore, R represents the *total effective resistance* in the circuit.

The equation for resonance is

$$f = \frac{1}{2\pi\sqrt{LC}} \qquad \text{(Eq. 4)}$$

where f = frequency in Hz, L = inductance in henries, C = capacitance in farads and π = 3.14.

The terms of the equation are somewhat unwieldy for most rf calculations required by amateurs. A more suitable formula is

$$f = \frac{10^6}{2\pi\sqrt{LC}} \therefore f = \frac{10^6}{6.28\sqrt{90 \times 125}} = \frac{10^6}{6.28 \times 106} = \frac{10^6}{666} = 1501 \text{ kHz} \qquad \text{(Eq. 5)}$$

where f = kHz, L = μH (90), C = pF (125), and π = 3.14.

The equation can be reduced additionally if we use

$$f = \frac{159.2}{\sqrt{LC}} \quad \therefore f = \frac{159.2}{\sqrt{90 \times .000125}} = \frac{159.2}{\sqrt{.01125}} = \frac{159.2}{0.106} = 1501 \text{ kHz} \qquad \text{(Eq. 6)}$$

where f = kHz, L = μH (90), and C = μF (.000125).

Figs. 3, 4 and 5 contain handy nomographs which can be used with any two known quantities of L, X_C, X_L, C or f. It is necessary only to place a straightedge across the graph, joining the two known quantities.

Fig. 6 will be useful to the amateur who winds his own tuned-circuit inductors. As an example of the nomograph applied, assume a coil diameter of 1 inch, and a winding length of 1-1/2 inches. An inductance of 30 μH is desired, so a line is drawn from the number 1 on D scale to 30 μH on L scale. This line is extended to the axis scale, where it intersects at number 59. This number becomes the reference for the second line, which is drawn from 0.666 (the number obtained by D/length, or 1/1.5) on the K scale to 59 on the axis, then on to the N scale where it intersects at 49. Therefore, 49 close-wound, single-layer coil turns are needed to provide 30 μH of inductance.

If the coil length, diameter and turns count are known, the inductance value in μH for a single-layer, close-wound inductor can be determined from

$$L = \frac{(rN)^2}{9r + 10l} \therefore L = \frac{(0.5 \times 49)^2}{4.5 + 15} = \frac{600.25}{19.5} = 30\,\mu\text{H} \qquad \text{(Eq. 7)}$$

where L = inductance in μH, N = number of turns, r = coil mean radius (inches), and l = length of coil (inches).

For multilayer coils the value of L can be obtained from

$$L = \frac{0.8\,(rN)^2}{6r + 9l + 10b} \qquad \text{(Eq. 8)}$$

where b = depth of coil winding in inches. The remaining terms have the same meaning as when used in the single-layer coil equation.

Table 1 will be handy for selecting a wire size when winding a coil which must occupy a given amount of length. It is based on standard wire gauges, and wire which contains enamel or Formvar insulation.

LC Constants

At times it is convenient to use the numerical value of the LC constant with a group of calculations which involve different L/C ratios for a given frequency. The constant for doing this is shown in the following equation

$$LC = \frac{25{,}330}{f^2} \qquad \text{(Eq. 9)}$$

where L = inductance in μH, C = capacitance in pF, and f = frequency in MHz.

Example: To find the quantity of inductance needed to establish circuit resonance at 7050 kHz (7.05 MHz) with capacitances of 25, 50, 100 and 250 pF, the LC constant is

$$LC = \frac{25{,}330}{7.05^2} = 509.6 \qquad \text{(Eq. 10)}$$

Therefore, with 25 pF – L = 509.6 ÷ C = 509.6 ÷ 25 = 20.4 μH; 50 pF – L = 509.6 ÷ C = 509.6 ÷ 50 = 10.2 μH; 100 pF – L = 509.6 ÷ C = 509.6 ÷ 100 = 5.1 μH; 250 pF – L = 509.6 ÷ C = 509.6 ÷ 250 = 2.0 μH.

Capacitive Reactance

The quantity of electrical charge that can be placed on a capacitor is proportional to the applied emf and capacitance. This amount of charge moves to and fro in the circuit once each cycle. Therefore, the *rate* of charge movement (the current) is proportional to the voltage, capacitance and frequency. If the effects of capacitance and frequency are combined, they comprise a quantity that exhibits a characteristic similar to that of resistance in Ohm's Law. This quantity is called *reactance*, and the unit of designation for it is the ohm. The equation for capacitive reactance is

$$X_C = \frac{1}{2\pi fC} = \frac{1}{6.28 \times 1.9 \times .00015} = \frac{1}{.001789} = 559 \text{ ohms} \qquad \text{(Eq. 11)}$$

where X_C = capacitive reactance in ohms, f = 1.9 MHz, C = .00015 μF (150 pF) and π = 3.14. The reactance

Fig. 5

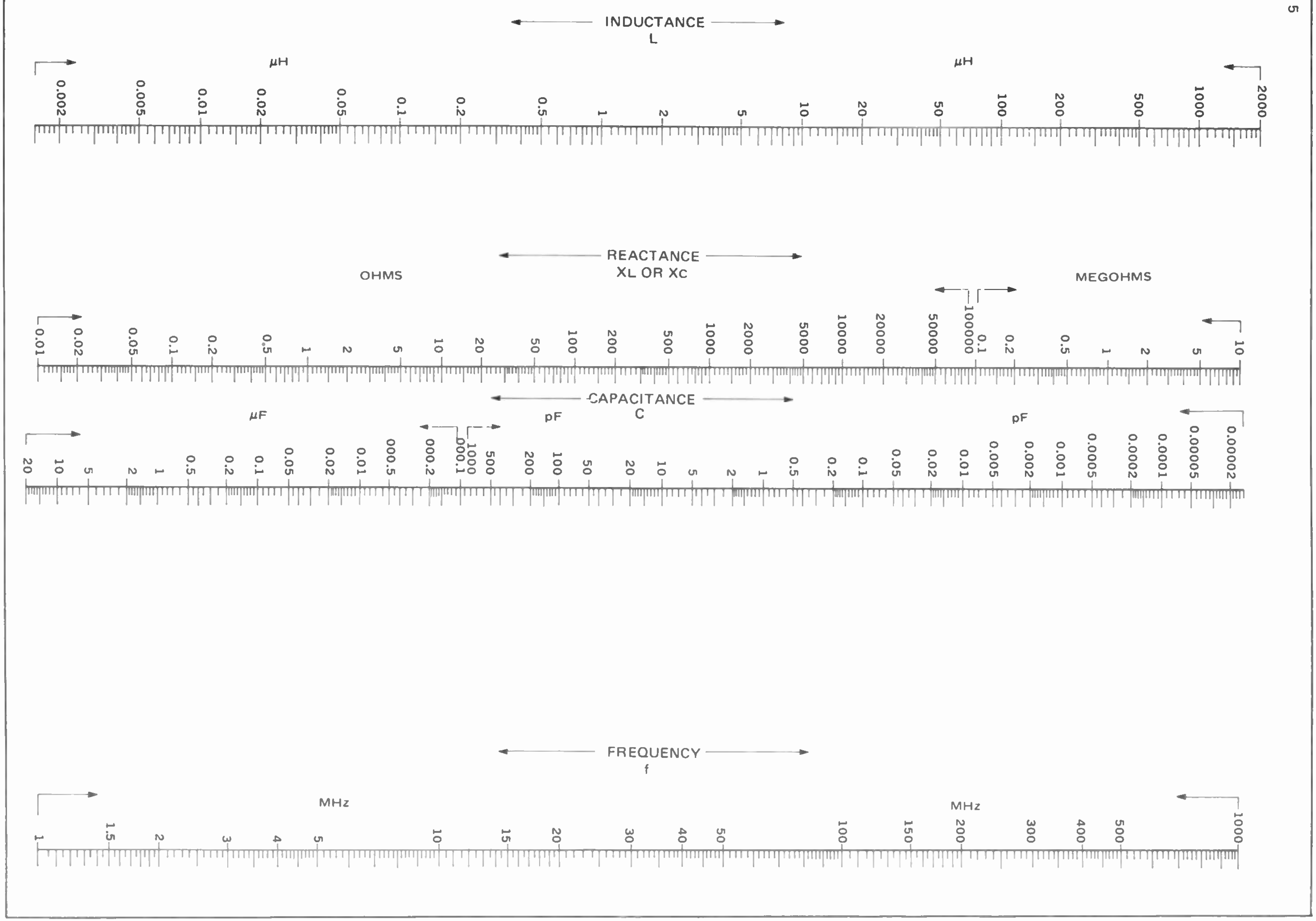
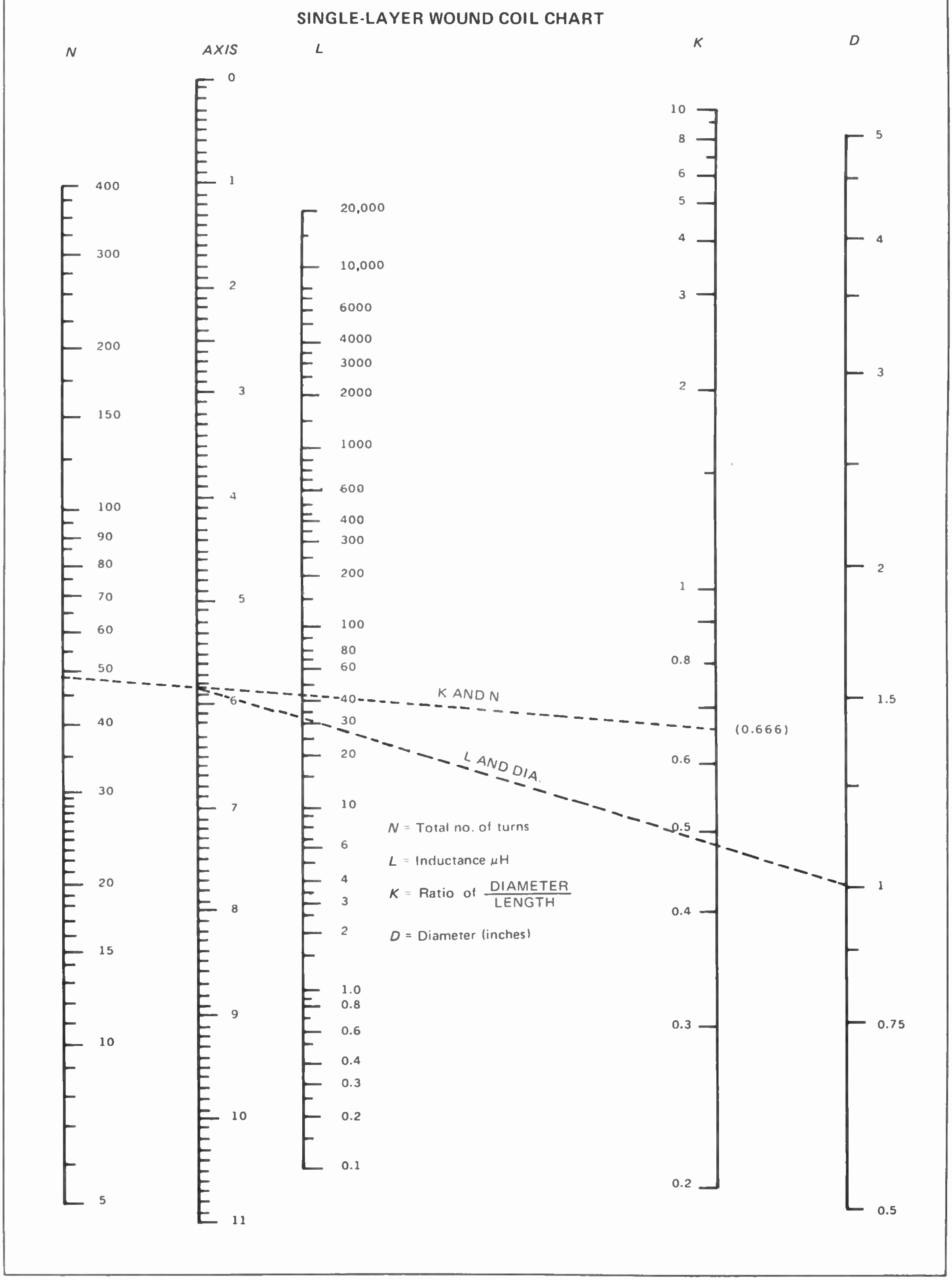

INDUCTANCE
L
μH
0.002 0.005 0.01 0.02 0.05 0.1 0.2 0.5 1 2 5 10 20 50 100 200 500 1000 2000
μH
REACTANCE
XL OR Xc
OHMS
MEGOHMS
0.01 0.02 0.05 0.1 0.2 0.5 1 2 5 10 20 50 100 200 500 1000 2000 5000 10000 20000 50000 100000 0.1 0.2 0.5 1 2 5 10
CAPACITANCE
C
μF
pF
pF
20 10 5 2 1 0.5 0.2 0.1 0.05 0.02 0.01 000.5 000.2 000.1 1000 500 200 100 50 20 10 5 2 1 0.5 0.2 0.1 0.05 0.02 0.01 0.005 0.002 0.001 0.0005 0.0002 0.0001 0.00005 0.00002
FREQUENCY
f
MHz
MHz
1 1.5 2 3 4 5 10 15 20 30 40 50 100 150 200 300 400 500 1000

Fig. 6


SINGLE-LAYER WOUND COIL CHART
N
AXIS
L
K
D
K AND N
L AND DIA.
(0.666)
N = Total no. of turns
L = Inductance µH
K = Ratio of DIAMETER / LENGTH
D = Diameter (inches)

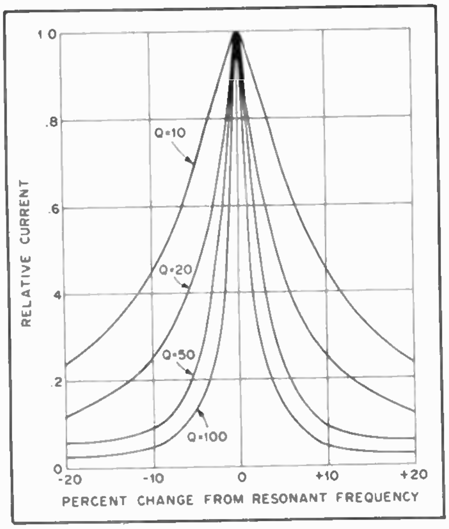


Fig. 7 — Current in series-resonant circuits having different Qs. The current at resonance is assumed to be the same in all cases. The lower the Q, the more slowly the current decreases as the applied frequency is moved away from resonance.

of various capacitors versus operating frequency can be obtained in rough measure from Figs. 3, 4 and 5 of this chapter.

Although the unit of reactance is the ohm, no power is dissipated in reactance. The energy stored in the capacitor during one quarter cycle is returned to the circuit in the next. The X_C equation can be simplified to $X_C = 159.2 \div f\,(\text{kHz}) \times C\,(\mu\text{F})$.

Inductive Reactance

When an ac voltage is applied to an inductance which has no dc resistance (all practical inductors have *some* resistance), the current is 90 degrees out of phase with the applied voltage. In this case — the opposite of that which employs capacitance — the current *lags* 90 degrees behind the voltage. The combined effect of inductance and frequency is called *inductive reactance*. It also is expressed in ohms, and the equation for inductive reactance is

$$X_L = 2\pi f L \therefore X_L = 6.28 \times 1.9 \times 47 = 560 \text{ ohms} \qquad \text{(Eq. 12)}$$

where X_L = inductive reactance in ohms, f = 1.9 MHz, L = 47 μH and π = 3.14.

If the reactance of C or L is known, but the component value for a given frequency is not, the following equations can be used

$$L = \frac{X_L}{2\pi f} \therefore L = \frac{560}{6.28 \times 1.9} = \frac{560}{11.93} = 47\ \mu\text{H} \qquad \text{(A)}$$

$$C = \frac{1}{2\pi f X_C} \therefore C = \frac{1}{6.28 \times 1.9 \times 559} = \frac{1}{6670} = .00015\ \mu\text{F} \qquad \text{(B)} \quad \text{(Eq. 13)}$$

where at A, L = inductance in μH, X_L = inductive reactance in ohms, f = 1.9 MHz and π = 3.14, and where at B, C = capacitance in μF, f = 1.9 MHz, X_C = capacitive reactance in ohms and π = 3.14.

Q and Related Calculations

Q is often referred to as a *quality factor*, but in more definitive terms is the symbol for quantity of electric charge. It can be regarded as a measure of the relationship between stored energy and the rate of dissipation in certain electric elements, structures or materials.

Capacitance Q

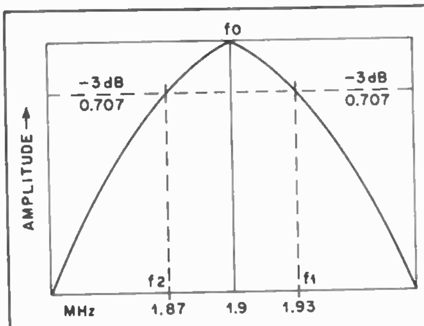
Capacitor quality (Q) is determined by the ratio of capacitive reactance (X_C) to resistance (R). The effective resistance of a capacitor is considered to be in series with the reactance. Q is determined by $Q = X_C \div R$. Thus, the greater the amount of X_C to a specified R quantity, the higher the quality factor, or Q.

The Q of a capacitor has some dependence on the operating frequency, but normally capacitor Q will be relatively high because the effective resistance of most modern capacitors is quite low, the exception being at vhf and above. The simple equation expressed in the foregoing is suitable for most work, but the following is suitable also

$$Q = \frac{\dfrac{1}{(2\pi f C)}}{R} \qquad \text{(Eq. 14)}$$

where f = hertz, C = farads, R = ohms and π = 3.14. Q can also be determined from $Q = 1 \div (2\pi f C) \times R$.

A high-Q capacitor finds its greatest application in tuned circuits that are designed to provide moderate to high selectivity (C and L circuits). High-Q capacitors are necessary in RC active filter circuits for audio peak or rejection applications, and in various kinds of passive LC filter networks.



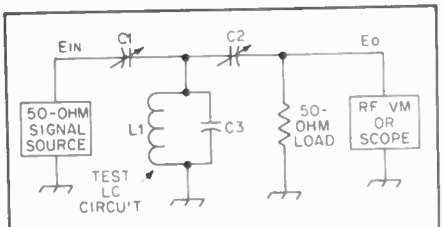
$\Delta f = 30 \text{ kHz} \therefore$
$2\Delta f = 60 \text{ kHz}$

$$Q_u = \frac{f_o}{2\Delta f(-3\text{ dB})} = \frac{1.9}{.06\text{ MHz}} = 31.6$$

$$\text{or } Q_u = \frac{1}{\dfrac{2\Delta f}{f_o}} = \frac{1}{\dfrac{.06\text{ MHz}}{1.9}} = \frac{1}{.0316} = 31.6 \text{ or } Q_u = \frac{f_o}{f_1 - f_2}$$

$$= \frac{1.9}{1.93 - 1.87} = \frac{1.9}{.06} = 31.6$$

Fig. 8 — Curve and equations for determining Q_u of a tuned circuit.



1) $X_{L1} = X_{C3}$ at f_o.
2) Adjust C1, C2 for insertion loss ⩾–30 dB. (C1, C2 typically 5 pF or less for mf and hf tests.)

Fig. 9 — Simple test setup for determining known values of Q_u.

Inductive Q

The Q of an inductor, or coil, is determined by the ratio of inductive reactance (X_L) to resistance. As was true in the discussion of capacitor Q, the X_L and effective R are considered to be in series. Q is determined by $Q = X_L \div R$. Thus, the greater the inductive reactance to a specified R quantity, the higher the Q. In addition to the foregoing equation Q can be found from

$$Q = \frac{2\pi f L}{R} \qquad \text{(Eq. 15)}$$

where f = hertz, L = henries, R = ohms and π = 3.14.

A high-Q inductor finds its greatest application in tuned (LC) circuits which are designed to provide a high degree of selectivity. High-Q coils are also used in wave traps and as narrow-band antenna loading inductors.

Tuned-Circuit Q

Resonant circuits have quality factors (Q) which are dependent in part upon the Q factors of the inductors and capacitors used in them. The quality is expressed as Q_u (unloaded Q) and Q_L (loaded Q). Q_u is the quality factor of a tuned circuit to which no load has been connected (antenna, tube, transistor or whatever). The Q_L condition arises when the tuned circuit is installed in the operating circuit for which it was designed and is, therefore, affected by the impedances reflected into it. The higher the loaded Q the narrower the bandwidth and vice versa.

All coils have *some* ohmic resistance, and so do the coil and capacitor leads. As the operating frequency is raised, the ac resistance of the component leads increases. The total of these resistances is equivalent to a resistor in parallel or series with the tuned circuit, and is designated as R, in ohms.

The effect of Q on sharpness of tuned-circuit resonance can be seen in Fig. 7. Q can be obtained when the reactance of either the inductor or capacitor is known (they are equal but 180 degrees out of phase at resonance), and when R is known. The equation is

$$Q_u = \frac{X}{R} \therefore Q = \frac{590}{3} = 197 \quad \text{(Eq. 16)}$$

where X_C or X_L = 590 and R = 3 ohms.

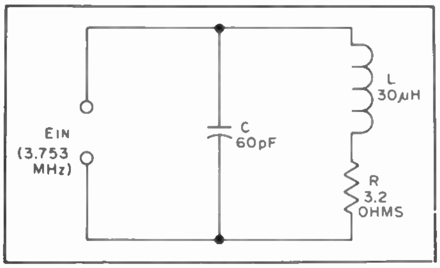


Fig. 10 – Representative circuit of the components contained in the equation for determining the Z of a parallel-resonant circuit.

Resonant-circuit Q is determined by taking points (two) either side of resonance at which the signal amplitude is down to 0.707 (−3 dB) of the peak value, as illustrated in Fig. 8. These points can be found by means of simple test equipment and the method is illustrated in Fig. 9. The network is swept to find the −3 dB points on the curve of Fig. 8. Trimmers C1 and C2 are set for approximately equal amounts of capacitance, consistent with a high order of insertion loss, thereby minimizing the loading on tuned circuit L1C1. Excessive loading would reduce the Q_u and make measurements relatively meaningless.

When the individual Qs of an inductor and a capacitor are known, and when they are to be combined in a tuned circuit, the Q_u can be determined by

$$Q = \frac{Q_{ind} \times Q_{cap}}{Q_{ind} + Q_{cap}} \therefore \frac{35 \times 105}{35 + 105}$$

$$= \frac{3675}{140} = 26.25 \quad \text{(Eq. 17)}$$

where Q_{ind} (coil Q) = 35, and Q_{cap} (capacitor Q) = 105, at tuned-circuit resonance.

The term Q is applicable also to magnetic and dielectric materials. At a specific frequency the Q is 6.28 times the ratio of the maximum stored energy to the energy dissipated in the material per Hz.

The *loaded Q* (Q_L) of a circuit is the value of Q which results when the circuit is connected to a device that dissipates energy (transistor, tube, resistor, antenna, etc.). The loaded Q of a given element is always less than the unloaded Q in linear applications.

Core Materials Versus Q

Slug-tuned inductors, toroidal-wound inductors, and pot-core inductors are used extensively. The magnetic material is either powdered iron or ferrite, and if chosen correctly for the operating frequency it can actually enhance the coil Q by virtue of the magnetic material increasing the effective inductance, thereby reducing the coil resistance. (Fewer turns will be needed for the same inductance.)

Impedance of Parallel-Resonant Circuits

Impedance is expressed in ohms and the symbol for impedance is Z. The fundamental equation for calculating impedance is

$$Z = R + j\left(2\pi fL - \frac{1}{2\pi C}\right)$$

$$\text{or } Z = \sqrt{R^2 + \left(2\pi fL - \frac{1}{2\pi fC}\right)^2} \quad \text{(Eq. 18)}$$

where f = frequency, L = inductance and C = capacitance. Impedance is determined from known values of Q or X and R. Fig. 10 illustrates the equivalent of a parallel-resonant circuit from which Z can be determined. The latter can be found from

$$Z = \frac{{X_L}^2}{R} \text{ and } X_L = 2\pi fL$$

$$= 6.28 \times 3.753 \times 30 = 707 \text{ (ohms)}$$

$$\therefore Z = \frac{707^2}{3.2} = \frac{499941}{3.2}$$

$$= 156{,}239 \text{ (ohms)} \quad \text{(A)}$$

$$\text{or } Z = Q^2 R \text{ and } Q = \frac{X_L}{R} = \frac{707}{3.2} = 221$$

$$\therefore Z = 221^2 \times 3.2 = 156{,}239 \text{ (ohms)} \quad \text{(B)}$$

$$\text{or } Z = \frac{L}{CR} = \frac{30\,\mu\text{H}}{.00006\,\mu\text{F} \times 3.2 \text{ ohms}}$$

$$= 156{,}239 \text{ (ohms)} \quad \text{(C)}$$

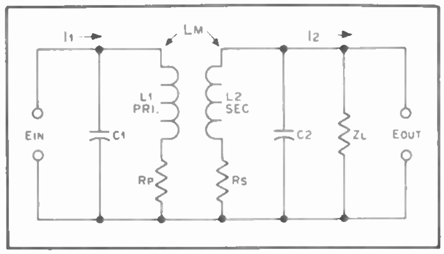


Fig. 11 – Illustration of the components for a mutually coupled circuit. R_p and R_s represent the coil series resistances.

$$\text{or } Z = X_L \times Q = 707 \times 221$$

$$= 156{,}239 \text{ (ohms)} \quad \text{(D)}$$

(Eq. 19)

where L = µH, R = ohms, X_L = reactance, and C = capacitance in µF. It should be noted that in the examples provided for Z the answer is given as 156,239 ohms, which is the result obtained when the various quantities are carried out to at least four decimal places, which they are not in the simplified examples given. The foregoing equations are based on values of Q above approximately 10.

Coupled Circuits

When signal energy is routed from one point in a circuit to another, the energy is *coupled* between those points. The term *coupling coefficient* (k) is used

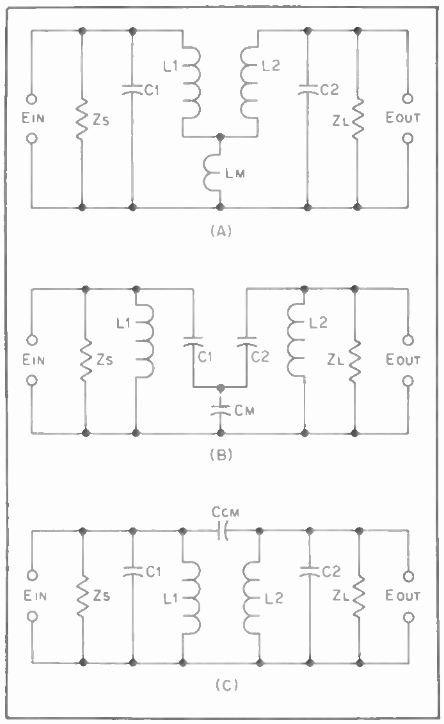


Fig. 12 – Various schemes for mutual coupling. At A, inductive bottom coupling. The circuit of B illustrates capacitive bottom coupling. Top capacitive coupling is shown at C.

in reference to the ratio of the flux which links the second coil from the first coil (L1 and L2 of Fig. 11) to the total flux of the first coil. In other words, the two coils are linked by a mutual or common flux. The equation for determining the mutual inductance (L_M) between the two coils is

$$L_M = k\sqrt{L1 \times L2} = 0.1\sqrt{30\,\mu H \times 30\,\mu H}$$

$$= 0.1 \times \sqrt{900} = 0.1 \times 30 = 3$$

conversely, $k = \dfrac{L_M}{\sqrt{L1 \times L2}} = 0.1$ (Eq. 20)

where L_M = the mutual inductance, L = the coil inductance, and k = the coefficient of coupling (1 or less). When designing transformers with magnetic core material, the objective is to obtain a k factor of unity, or 1. When air-core inductors are used in a transformer, the k factor is quite low – typically 0.01 to 0.1 (1 to 10 percent). It is for this reason that toroidal transformers are ideal for rf applications when ferrite or powdered-iron materials are used.

Other coupling methods are in common use, additional to the mutual-inductance concept. Three of the more popular techniques are shown in Fig. 12. The k factor can be obtained by

$$k = \frac{L_M}{\sqrt{L1 \times L2}} \quad \text{(A)}$$

$$k = \frac{\sqrt{C1 \times C2}}{C_M} \quad \text{(B)}$$

$$k = \frac{C_{CM}}{\sqrt{C1 \times C2}} \quad \text{(C)}$$

(Eq. 21)

A more definitive treatment of this subject can be found in *A Course in Radio Fundamentals*, chapter 14, 5th edition.

Chapter 4

L, C and *R* Networks

This section treats *L, C* and *R* networks for popular amateur radio applications. Networks of *L* and *C* components are necessary for matching a given impedance to another at rf, and are intended to assure maximum transfer of energy from one point in the circuit to another. The *LC* networks treated here are the resonant type, and are for use between the active stages of solid-state and tube-type transmitters and between the transmitter output and the antenna system.

Table 1

Pi-Network Resistive Attenuator (50 ohms)

dB ATTEN.	*R1 (OHMS)*	*R2 (OHMS)*
1	870.0	5.8
2	436.0	11.6
3	292.0	17.6
4	221.0	23.8
5	178.6	30.4
6	150.5	37.3
7	130.7	44.8
8	116.0	52.8
9	105.0	61.6
10	96.2	70.7
11	89.2	81.6
12	83.5	93.2
13	78.8	106.0
14	74.9	120.3
15	71.6	136.1
16	68.8	153.8
17	66.4	173.4
18	64.4	195.4
19	62.6	220.0
20	61.0	247.5
21	59.7	278.2
22	58.6	312.7
23	57.6	348.0
24	56.7	394.6
25	56.0	443.1
30	53.2	789.7
35	51.8	1406.1
40	51.0	2500.0

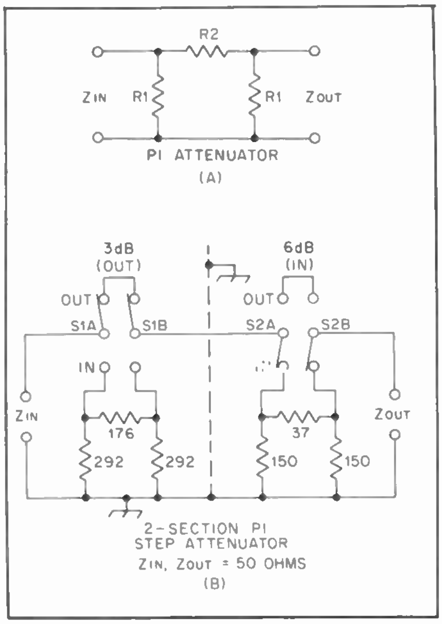


Fig. 1 — Schematic diagram of a pi attenuator (A). Shown at B is a two-section step attenuator using pi sections.

The resistive networks are for use as *rf* and *audio attenuators,* and can be built for whatever power level is necessary by selecting resistors of suitable wattage rating. Alternatively, combinations or series or parallel low-wattage resistors can be used to form the power attenuators. Information on combining resistors is provided in chapter 1 of this book.

Pi-Network Attenuator

Two attenuator networks used frequently by amateurs are the *pi* and *T* types. Each has its subtle advantages and disadvantages, and the builder should choose his network in accordance with the application. Where it is desired to cascade network sections of various attenuation amounts, it can be done in classic step-attenuator form (Fig. 1B). The pi type is a common choice. Another consideration is that the pi type will dissipate power through all three resistances in the event of an accidental open-circuit condition at the output port. This is significant only in cases where a power attenuator is being

Table 2

T Network Resistive Attenuator (50 ohms)

dB ATTEN.	*R1 (OHMS)*	*R2 (OHMS)*
1	2.9	433.3
2	5.7	215.2
3	8.5	132.0
4	11.3	104.8
5	14.0	82.2
6	16.6	66.9
7	19.0	55.8
8	21.5	47.3
9	23.8	40.6
10	26.0	35.0
11	28.0	30.6
12	30.0	26.8
13	31.7	23.5
14	33.3	20.8
15	35.0	18.4
16	36.3	16.2
17	37.6	14.4
18	38.8	12.8
19	40.0	11.4
20	41.0	10.0
21	41.8	9.0
22	42.6	7.8
23	43.4	7.1
24	44.0	6.3
25	44.7	5.6
30	47.0	3.2
35	48.2	1.8
40	49.0	1.0

used. Conversely, the T attenuator will dissipate power through only two of its resistors during an open-circuit period. Therefore, the pi attenuator might survive damage where a T type would not. Furthermore, examination of Tables 1 and 2 will show that each network has decidedly different ratios of resistance at opposite ends of the attenuation scales. The pi attenuator resistances are more practical in terms of ohmic value in some areas of the dB range.

Table 1 is based on a set of calculations which provide attenuation amounts from 1 to 40 dB. The values are for pi attenuators whose characteristic impedances are 50 ohms, bilateral. The resistance values given have been carried out to only the nearest decimal quantity, but should be adequate for all but the most precise work.

Greater precision can be had by using K1PLP's simplified equation to determine the values of resistance required for a network. Also, the equation enables the designer to calculate values for impedances other than 50 ohms. The formula for pi attenuators (with an example) is

$$R1 = Z\,\frac{10^{.05A}+1}{10^{.05A}-1} = 75\left(\frac{10^{.05\times12}+1}{10^{.05\times12}-1}\right)$$

$$= 75\left(\frac{10^{0.6}+1}{10^{0.6}-1}\right) = 75\left(\frac{3.98+1}{3.98-1}\right)$$

$$= 75\left(\frac{4.98}{2.98}\right) = 75 \times 1.671$$

$$= 125.3 \text{ ohms} \qquad \text{(A)}$$

$$R2 = \frac{ZR1\,(10^{.05A}-1)}{Z+R1}$$

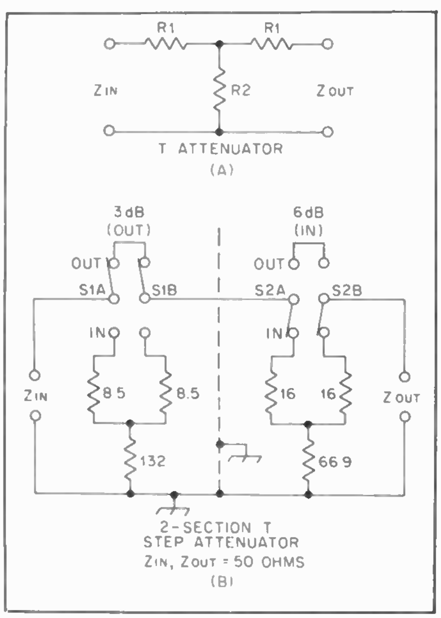


Fig. 2 — Schematic diagram of a T-attenuator network (A). Shown at B is a two-section T attenuator.

$$= \frac{75 \times 125.3\,(10^{.05\times12}-1)}{75+125.3}$$

$$= \frac{9400 \times 2.98}{200.3} = \frac{28012}{200.3}$$

$$= 139.8 \text{ ohms} \qquad \text{(B)}$$

(Eq. 1)

where R is resistance in ohms, A is the desired attenuation amount in dB, and Z is the impedance of the attenuator in ohms. The example shows how to design a 12-dB, 75-ohm attenuator.

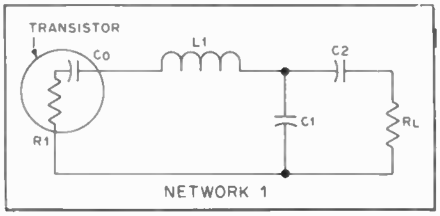


Fig. 3 — Schematic illustration of network 1.

T-Section Attenuator

Fig. 2 shows the configuration of a *T-section attenuator*. It can also be used in a step-attenuator hookup (Fig. 2B). Table 2 lists values for 50-ohm operation between 1 and 40 dB. Resistance values were determined by simplified equations developed by K1PLP. Values listed were carried out to only the nearest decimal quantities, but should be suitable for all but the most precise work.

Greater precision can be had by using the equations given for T attenuators. The equations will permit determination of impedance values other than 50 ohms. The formula for T attenuators (with an example) is

$$R1 = Z\left(\frac{10^{.05A}-1}{10^{.05A}+1}\right)$$

$$= 75\left(\frac{10^{.05\times12}-1}{10^{.05\times12}+1}\right)$$

$$= 75\left(\frac{10^{0.6}-1}{10^{0.6}+1}\right) = 75\left(\frac{3.98-1}{3.98+1}\right)$$

$$= 75\left(\frac{2.98}{4.98}\right) = 75 \times 0.598$$

$$= 44.9 \text{ ohms} \qquad \text{(A)}$$

$$R2 = \frac{R1+Z}{10^{.05A}-1} = \frac{44.9+75}{10^{0.6}-1}$$

$$= \frac{119.9}{3.98-1} = \frac{119.9}{2.98} = 40.2 \text{ ohms} \qquad \text{(B)}$$

(Eq. 2)

where R is resistance in ohms, A is the desired attenuation amount in dB, and Z is the attenuator impedance in ohms. The example shows how to design a 12-dB, 75-ohm attenuator.

For amateur applications, and particularly when power attenuators are employed to reduce exciter drive to a linear amplifier, the resistance values for R1 and R2 of both attenuator types can be selected from the table of standard values in chapter 1. In other words, a 39-ohm resistor can be used in place of a 40.6-ohm type, a 100-ohm value can be used instead of a 104.8-ohm one and so on.

Particular attention should be given to keeping the leads as short as possible to minimize unwanted reactances which can spoil the impedance characteristics of an attenuator. Similarly, unwanted stray capacitances should be kept to a minimum to prevent leakage across the attenuator network, which would degrade the attenuation amount. The latter condition worsens as the operating frequency is increased. In most instances an attenuator should be contained in a well-shielded housing.

LC Impedance-Matching Networks

To ensure maximum power transfer from one circuit point to another, the characteristic impedances of the source and load must be matched as closely as possible. The same is true of vacuum tube or solid-state amplifiers, low or high level. The matter of tuned-circuit Q was treated in chapter 3, and the subject is an important one in connection with *LC* matching networks. The latter are tuned to resonance at the operating frequency of the stages being matched to one another. Network Q is chosen to meet specific operating requirements — harmonic rejection, bandwidth and circuit stability. Respective to the last term, unconditional stability of an amplifier is sometimes hard to realize when using extremely high values of Q in the networks of high-gain unneutralized amplifiers: A tradeoff is often taken between stability and harmonic reduction.

Impedances can be matched by means of the more standard primary/secondary rf transformer circuits, which employ a series- or parallel-resonant primary.

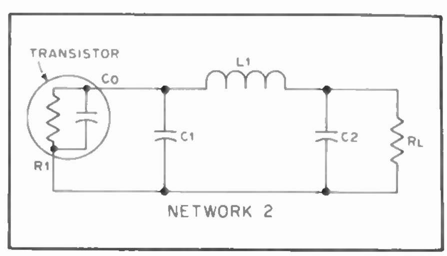


Fig. 4 — Diagram of network 2.

Table 3

Network No. 1 Data

$Q_L = 1$ $R_L = 50$

$R1$	X_{L1}	X_{C1}	X_{C2}
26	26	65	10
27	27	75	14
28	28	85	17.3
29	29	96	20
30	30	108	22.3
32	32	136	26.4
34	34	170	30
36	36	214	33
38	38	272	36
40	40	355	38.7
42	42	479	41
44	44	686	43.5
46	46	1102	45.8
48	48	2351	48

$Q_L = 2$ $R_L = 50$

$R1$	X_{L1}	X_{C1}	X_{C2}
11	22	33	15.8
12	24	39	22.4
13	26	45	27.4
14	28	51	31.6
15	30	58	35.4
16	32	65	38.7
17	34	73	41.8
18	36	81	44.7
19	38	90	47.4
20	40	100	50
21	42	110	52.4
22	44	122	55
23	46	134	57
24	48	147	59
25	50	161	61
26	52	177	63
27	54	194	65
28	56	213	67
29	58	233	69
30	60	256	71
32	64	310	74
34	68	377	77
36	72	464	81
38	76	582	84
40	80	746	87
42	84	995	89
44	88	1409	92
46	92	2241	95
48	96	4739	97

$Q_L = 3$ $R_L = 50$

$R1$	X_{L1}	X_{C1}	X_{C2}
6	18	23	22
7	21	30	31.6
8	24	36	38.7
9	27	43	44.7
10	30	50	50
11	33	58	54.8
12	36	66	59
13	39	75	63
14	42	84	67
15	45	95	71
16	48	105	74
17	51	117	77
18	54	130	81
19	57	143	84
20	60	158	87
21	63	173	89
22	66	190	92
23	69	209	95
24	72	228	97
25	75	250	100
26	78	274	102
27	81	299	105
28	84	327	107
29	87	358	110
30	90	393	112
32	96	473	116
34	102	575	120
36	108	706	124
38	114	882	128
40	120	1129	132
42	126	1502	136
44	132	2124	140
46	138	3372	143
48	144	7119	146

$Q_L = 4$ $R_L = 50$

$R1$	X_{L1}	X_{C1}	X_{C2}
3	12	13	7
4	16	20	30
5	20	27	42
6	24	34	51
7	28	42	59
8	32	51	66
9	36	60	72
10	40	69	77
11	44	80	83
12	48	91	88
13	52	103	92
14	56	115	97
15	60	129	101
16	64	144	105
17	68	159	109
18	72	176	113
19	76	194	117
20	80	214	120
21	84	235	124
22	88	257	127
23	92	282	131
24	96	308	134
25	100	337	137
26	104	368	140
27	108	403	143
28	112	440	146
29	116	482	149
30	120	527	152
32	128	635	157
34	136	770	162
36	144	945	168
38	152	1180	173
40	160	1510	177
42	168	2007	182
44	176	2837	187
46	184	4500	191
48	192	9497	196

Since that method of matching is rather standard, it will not be covered here.

Transistor Matching Networks

Four impedance matching networks will be illustrated in this section, and equations for solving the L and C values are supplied along with tables of reactance values for obtaining the necessary inductance and capacitance amounts for various impedance combinations. The equations and tables are based on $R + j\,0$ conditions. The foregoing is not a typical situation when working with transistors. Complex combinations of reactance (capacitive and inductive) and resistance exist at the input and output terminals of transistor amplifiers, so the information given here represents an approximation of suitable networks. For example, the base terminal (input) of a common emitter rf amplifier contains R_{BB} (base spreading resistance), C_c (case capacitance), R_E (emitter diffusion resistance), and C_{in} (input capacitance). Similar factors affect the output characteristics of transistor amplifiers. Therefore, the network data provided here must be founded on terminal impedances which are *purely resistive*. For most amateur design work, the values will be satisfactory as starting points for final optimization at a given operating frequency. Precise impedance matching can be realized by employing "sloppy networks." That is, the L and C components can be made variable above and below the computed values; then with the circuit operating, L and C can be tweaked for maximum power transfer, or minimum SWR.

Definitive treatments of matching networks and their design methods are given in Motorola *Application Note AN-721*, and in RCA's *Solid-State Power Circuits*, Tech. Series SP-52. The following information is based in part on data given in Motorola *Application Note AN-267*.

Network No. 1

This network is basically an L-section type. It is useful only when the input impedance (R1 of Fig. 3) is less than the output impedance, R_L. To be explicit, it is best suited to matching collector impedances of less than 50 ohms to load impedances of 50 ohms or greater. Network equations for the circuit of Fig. 3 were derived from data given in *Electronics Circuit Analysis*, Vol. 1, by Cutler

$$X_{L1} = Q_L R1 + X_{co} = 4 \times 15 + 300$$

$$= 360 \text{ ohms}$$

where $Q_L = 4$, R1 = 15 ohms, and

$$C_o = 150 \text{ pF}$$

$$(X_{co} = 300 \text{ at } 3.5 \text{ MHz}) \qquad \text{(A)}$$

$$X_{C2} = AR_L = 2.02 \times 50 = 101 \text{ ohms}$$

$$\text{where } A = \sqrt{\left[\frac{R1\,(Q_L{}^2 + 1)}{R_L}\right] - 1}$$

$$= \sqrt{4.1} = 2.02 \qquad \text{(B)}$$

$$X_{C1} = \frac{B}{Q_L + A} = \frac{255}{4 + 2.02} = 42.3 \text{ ohms}$$

where $B = R1\,(Q_L^2 + 1)$

$$= 15 \times 17 = 255 \qquad \text{(C)} \quad \text{(Eq. 3)}$$

X_{co} must be determined for the chosen operating frequency. The value of capacitance can be obtained from the transistor data sheet curves of C versus frequency. The reactance values obtained from the equations can be converted to L and C values as shown in chapter 3.

Table 3 gives reactance values for the components of network 1 (R1 values of 3 to 48 ohms). The highest operating Q_L listed is 4, which is suitable for most amateur circuits. The figures are based on a 50-ohm R_L. For other values of load use the equations just given.

Network No. 2

Fig. 4 shows network 2, which is the pi type that has been used with vacuum-tube amplifiers for many years. This configuration is best suited to impedance-matching situations wherein R1 is larger than R_L. When R1 is less than the output load of the pi network, the values of L become almost impossible to realize from a practical point of view, and C becomes abnormally large in value. Since a pi network is a low-pass filter, it is often preferred in solid-state amplifiers for reduction of harmonic currents at the R_L terminal. More often than not the designer elects to use a double pi-section tank, which is similar in function to a half-wave harmonic filter. The lower the Q, the more practical the component values become at low values of R1. Therefore, a double pi can sometimes be used by making the first section low in Q, then employing a high-Q pi section at the R_L end of the circuit. The following equations are applicable to pi-section matching networks

$$X_{C1} = \frac{R1}{Q_L} = \frac{96}{2} = 48 \text{ ohms}$$

where Q_L = 2 and R1 = 96 ohms (A)

$$X_{C2} = R_L \sqrt{\frac{R1/R_L}{(Q_L^2 + 1) - (R1/R_L)}}$$

$$= 50 \sqrt{\frac{1.92}{5 - 1.92}} = 50\sqrt{0.623}$$

$$= 50 \times 0.789 = 39.5 \text{ ohms}$$

where R_L = 50 ohms (B)

$$X_{L1} = \frac{Q_L R1 + (R1\,R_L/X_{C2})}{Q_L^2 + 1}$$

$$= \frac{192 + 1.265}{5} = \frac{193.2}{5}$$

$$= 38.6 \text{ ohms} \qquad \text{(C)} \quad \text{(Eq. 4)}$$

For the most part X_{co} in this network can be ignored, for it will comprise only a small part of X_{C1} in most applications. Table 4 contains reactance values for pi networks whose Qs are between 1 to 4. R_L is always 50 ohms, and values given for R1 are from 1 to 800 ohms. Other R_L and Q values can be obtained by using the equations.

Network No. 3

Network 3 is a modified L section. It is shown in Fig. 5. It is intended for use in circuits where R1 is less than 50 ohms. This network will provide practical values of L and C where R1 is less in ohmic value than R_L. Equations for solving the network are

$$X_{C1} = Q_L R1 = 3 \times 39 = 117 \text{ ohms}$$

where Q_L = 3 and R1 = 39 ohms (A)

Table 4

Network No. 2 Data

$Q_L = 1$, $R_L = 50$

R1	X_{C1}	X_{C2}	X_L
1	1	5	5
2	2	7	8
3	3	9	10
4	4	10	12
5	5	11	13
10	10	17	20
15	15	21	25
20	20	25	30
25	25	29	34
30	30	33	38
35	35	37	41
40	40	41	44
45	45	45	47
50	50	50	50
55	55	55	52
60	60	61	54
65	65	68	56
70	70	76	58
75	75	87	59
80	80	100	60
85	85	119	60
90	90	150	60

$Q_L = 2$, $R_L = 50$

R1	X_{C1}	X_{C2}	X_L
1	0.5	3	4
2	1	4	5
3	1.5	6	7
4	2	6	8
5	2.5	7	9
10	5	10	14
15	7.5	13	18
20	10	15	22
25	12.5	17	25
30	15	18	28
35	17.5	20	31
40	20	22	34
45	22.5	23	37
50	25	25	40
55	27.5	27	43
60	30	28	45
65	32.5	30	48
70	35	31	50
75	37.5	33	53
80	40	34	55
85	42.5	36	58
90	45	37.5	60
95	47.5	39	62
100	50	41	64
125	62.5	50	75
150	75	61	84
175	87.5	76	93
200	100	100	100
225	112.5	150	105

$Q_L = 3$, $R_L = 50$

R1	X_{C1}	X_{C2}	X_L
1	0.33	2.2	2.5
2	0.67	3.1	3.7
3	1	3.8	4.7
4	1.3	4.4	5.6
5	1.6	5	6.4
10	3.3	7	10
15	5	8.7	13
20	6.6	10.2	15.8
25	8.3	11.4	18.4
30	10	12.6	20.8
35	11.6	13.7	23.2
40	13	14.7	25.5
45	15	15.7	27.8

$Q_L = 3$, $R_L = 50$

R1	X_{C1}	X_{C2}	X_L
50	16.6	16.6	30
55	18	17.5	32
60	20	18.4	34
65	21.6	19.3	36.3
70	23.3	20	38.3
75	25	21	40.3
80	26.6	21.8	42.3
85	28.3	22.6	44
90	30	23.4	46
95	31.6	24	48
100	33.3	25	50
125	41.6	28.8	59
150	50	32.7	67.9
175	58.3	36.6	76.3
200	66.6	40.8	84.4
225	75	45	92.3
250	83.3	50	100

$Q_L = 4$, $R_L = 50$

R1	X_{C1}	X_{C2}	X_L
25	6	8.7	14.3
50	12.5	12.5	23.5
75	18.7	15.5	31.8
100	25	18	39.6
125	31	20.7	47
150	37.5	23	54.3
175	43.7	25.4	61.3
200	50	27.7	68
225	56	30	75
250	62.5	32	81.6
300	75	36.9	94.4
400	100	47	119
500	125	59.7	142
600	150	77.4	163.9
700	175	108	183.7
800	200	200	200

Table 5

Network No. 3 Data

$Q_L = 1$		$R_L = 50$		$Q_L = 3$		$R_L = 50$	
$R1$	X_{C1}	X_{C2}	X_{L2}	$R1$	X_{C1}	X_{C2}	X_{L2}
1	1	7	8	1	3	7	10
2	2	10	11.8	2	6	10	15.8
3	3	12.6	14.8	3	9	12.6	20.8
4	4	14.7	17.5	4	12	14.7	25.5
5	5	16.6	20	5	15	16.6	30
6	6	18.4	22	6	18	18.4	34
7	7	20	24.3	7	21	20	38.3
8	8	21.8	26.3	8	24	21.8	42.3
9	9	23.4	28	9	27	23.4	46
10	10	25	30	10	30	25	50
11	11	26.5	31.8	11	33	26.5	53.7
12	12	28.1	33.3	12	36	28	57.3
13	13	29.6	34.9	13	39	29.6	60.9
14	14	31	36.4	14	42	31	64.4
15	15	32.7	37.9	15	45	32.7	67.9
16	16	34	39.3	16	48	34.3	71.3
17	17	35.8	40.6	17	51	35.8	74.6
18	18	37.5	42	18	54	37.5	78
19	19	39	43	19	57	39	81
20	20	40.8	44.4	20	60	40.8	84.4
21	21	42.5	45.6	21	63	42.5	87.6
22	22	44.3	46.8	22	66	44.3	90.8
23	23	46	47.9	23	69	46	93.9
24	24	48	48.9	24	72	48	96.9
25	25	50	50	25	75	50	100
26	26	52	50.9	26	78	52	102.9
27	27	54	51.9	27	81	54	105.9
28	28	56.4	52.8	28	84	56.4	108.8
29	29	58.7	53.6	29	87	58.7	111.6
30	30	61	54.4	30	90	61	114.4
32	32	66.6	56	32	96	66.6	120
34	34	72.8	57.3	34	102	72.8	125.3
36	36	80	58.4	36	108	80	130.4
38	38	88.9	59.3	38	114	88.9	135.3
40	40	100	60	40	120	100	140
42	42	114.5	60.3	42	126	114.5	144.3
44	44	135.4	60	44	132	135.4	148
46	46	169.5	59.5	46	138	169.5	151.5
48	48	244.9	57.8	48	144	244.9	153.8
$Q_L = 2$		$R_L = 50$		$Q_L = 4$		$R_L = 50$	
1	2	7	9	1	4	7	11
2	4	10	13.8	2	8	10	17.8
3	6	12.6	17.8	3	12	12.6	23.8
4	8	14.7	21.5	4	16	14.7	29.5
5	10	16.6	25	5	20	16.6	35
6	12	18.4	28	6	24	18.4	40
7	14	20	31.3	7	28	20	45.3
8	16	21.8	34.3	8	32	21.8	50.3
9	18	23.4	37	9	36	23.4	55
10	20	25	40	10	40	25	60
11	22	26.5	42.7	11	44	26.5	64.7
12	24	28	45.3	12	48	28	69.3
13	26	29.6	47.9	13	52	29.6	73.9
14	28	31	50.4	14	56	31	78.4
15	30	32.7	52.9	15	60	32.7	82.9
16	32	34.3	55.3	16	64	34.3	87.3
17	34	35.8	57.6	17	68	35.8	91.6
18	36	37.5	60	18	72	37.5	96
19	38	39	62	19	76	39	100
20	40	40.8	64.4	20	80	40.8	104.4
21	42	42.5	66.6	21	84	42.5	108.6
22	44	44.3	68.8	22	88	44.3	112.8
23	46	46	70.9	23	92	46	116.9
24	48	48	72.9	24	96	48	120.9
25	50	50	75	25	100	50	125
26	52	52	76.9	26	104	52	128.9
27	54	54	78.9	27	108	54	132.9
28	56	56.4	80.8	28	112	56.4	136.8
29	58	58.7	82.6	29	116	58.7	140.6
30	60	61	84.4	30	120	61	144.4
32	64	66.6	88	32	128	66.6	152
34	68	72.8	91.3	34	136	72.8	159.3
36	72	80	94.4	36	144	80	166.4
38	76	88.9	97.3	38	152	88.9	173.3
40	80	100	100	40	160	100	180
42	84	114.5	102.3	42	168	114.5	186.3
44	88	135.4	104	44	176	135.4	192.2
46	92	169.5	105.5	46	184	169.5	197.5
48	96	244.9	105.8	48	192	244.9	201.8

$$X_{C2} = R_L\sqrt{\frac{R1}{R_L - R1}} = 50\sqrt{\frac{39}{50-39}}$$

$$= 50\sqrt{3.54} \quad = 94.1 \text{ ohms}$$

where R_L = 50 ohms (B)

$$X_{L1} = X_{C1} + \left(\frac{R1R_L}{X_{C2}}\right) + X_{co}$$

$$= 117 + \left(\frac{1950}{94.1}\right) + 300$$

$$= 117 + 20.7 + 300 = 437.7 \text{ ohms}$$

where X_{co} = 300 ohms at 3.5 MHz (150 pF) (C)

(Eq. 5)

For best accuracy when using this network, the designer should transform the impedance of the device to be matched to series form by combining R1 and X_{co} ($R1 + jX_{co}$). This step is not necessary, of course, when using the network for matching circuits which do not contain transistors or other components which exhibit reactance. Table 5 lists reactance values for the components of network 3. The Q values are 1 through 4, and R1 values range from 1 to 48 ohms.

Network No. 4

This network is the classic T type. It can be used for matching R1 amounts which are above and below the value of R_L. This feature makes it the most flexible network of the four treated here. The T network (Fig. 6) provides high collector efficiencies when it is employed to match an rf power transistor to its load. Fig. 6 shows the circuit. The equations for solving the values of reactance for a given set of R, Q_L and R_L (with example) are

$$X_{L1} = (R1Q_L) + X_{co} = (7.5 \times 4) + 300$$

$$= 30 + 300 = 330 \text{ ohms}$$

where R1 = 7.5 ohms, Q_L = 4, and

X_{co} = 300 at 3.5 MHz (150 pF)

(A)

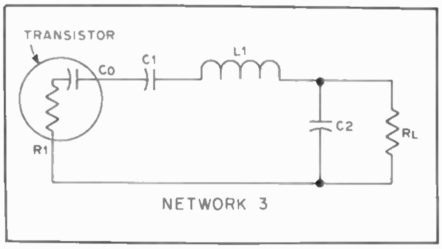


Fig. 5 — Illustration of network 3.

Table 6

Network No. 4 Data

$Q_L = 1$, $R_L = 50$

R1	X_{L1}	X_{L2}	X_{C1}
26	26	10	43.3
27	27	14	42
28	28	17.3	41.5
29	29	20	41.4
30	30	22.3	41.4
32	32	26.4	41.8
34	34	30	42.5
36	36	33	43
38	38	36	44
40	40	38.7	45
42	42	41	46
44	44	43.5	47
46	46	45.8	48
48	48	47.9	49
50	50	50	50
55	55	54.7	52.4
60	60	59	54.9
65	65	63	57.4
70	70	67	69.7
75	75	70.7	62
80	80	74	64.4
85	85	77.4	66.6
90	90	80.6	68.9
95	95	83.6	71
100	100	86.6	73
125	125	100	83.3
150	150	111.8	92.7
175	175	122.4	101.4
200	200	132	109.7
225	225	141.4	117.5
250	250	150	125
275	275	158	132
300	300	165.8	139

$Q_L = 2$, $R_L = 50$

R1	X_{L1}	X_{L2}	X_{C1}
11	22	15.8	23.7
12	24	22.3	24.5
13	26	27.3	25.5
14	28	31.6	26.5
15	30	35.3	27.7
16	32	38.7	28.8
17	34	41.8	29.9
18	36	44.7	31
19	38	47.4	32
20	40	50	33.3
21	42	52.4	34.4
22	44	54.7	35.5
23	46	57	36.6
24	48	59	37.7
25	50	61	38.7
26	52	63	39.8
27	54	65	40.8
28	56	67	41.9
29	58	68.9	42.9
30	60	70.7	43.9
32	64	74	45.9
34	68	77.4	47.9
36	72	80.6	49.8
38	76	83.6	51.7
40	80	86.6	53.5
42	84	89.4	55.4
44	88	92	57
46	92	94.8	59
48	96	97.4	60.7
50	100	100	62.5
55	110	106	66.7
60	120	111.8	70.8
65	130	117	74.8
70	140	122.4	78.6
75	150	127.4	82.4
80	160	132	86
85	170	136.9	89.6
90	180	141.4	93
95	190	145.7	96.6
100	200	150	100
125	250	169.5	115.9
150	300	187	130.6
175	350	203	144.3
200	400	217.9	157
225	450	231.8	169.5
250	500	244.9	181
275	550	257.3	192.3
300	600	269	203

$Q_L = 3$, $R_L = 50$

R1	X_{L1}	X_{L2}	X_{C1}
6	18	22.3	17.4
7	21	31.6	19
8	24	38.7	21
9	27	44.7	23
10	30	50	25
11	33	54.7	26.8
12	36	59	28.6
13	39	63	30.4
14	42	67	32
15	45	70.7	33.9
16	48	74	35.6
17	51	77.4	37.3
18	54	80.6	39
19	57	83.6	40.6
20	60	86.6	42
21	63	89.4	43.8
22	66	92	45.4
23	69	94.8	46.9
24	72	97.4	48.4
25	75	100	50
26	78	102.4	51.4
27	81	104.8	52.9
28	84	107	54.4
29	87	109.5	55.8
30	90	111.8	57
32	96	116	60
34	102	120.4	62.8
36	108	124.5	65.5
38	114	128.4	68
40	120	132	70.8
42	126	136	73.4
44	132	139.6	75.9
46	138	143	78.4
48	144	146.6	80.9
50	150	150	83.3
55	165	158	89
60	180	165.8	94.9
65	195	173	100.5
70	210	180	105.9
75	225	187	111
80	240	193.6	116.4
85	255	200	121.4
90	270	206	126.3
95	285	212	131
100	300	217.9	135.8
125	375	244.9	158
150	450	269	178.8
175	525	291.5	198
200	600	312	216.3
225	675	331.6	233.5
250	750	350	250
275	825	367.4	265.7
300	900	384	280.8

$Q_L = 4$, $R_L = 50$

R1	X_{L1}	X_{L2}	X_{C1}
3	12	7	12.3
4	16	30	14.7
5	20	41.8	17.5
6	24	50.9	20.3
7	28	58.7	23
8	32	65.5	25.6
9	36	71.7	28
10	40	77.4	30.6
11	44	82.7	33
12	48	87.7	35.4
13	52	92.4	37.7
14	56	96.9	40
15	60	101.2	42.3
16	64	105.3	44.5
17	68	109.3	46.7
18	72	113	48.8
19	76	116.8	50.9
20	80	120.4	53
21	84	123.9	55
22	88	127	57
23	92	130.5	59
24	96	133.7	61
25	100	136.9	63
26	104	140	65
27	108	143	66.9
28	112	145.9	68.8
29	116	148.8	70.6
30	120	151.6	72.5
32	128	157	76
34	136	162.4	79.7
36	144	167.6	83
38	152	172.6	86.6
40	160	177.4	90
42	168	182	93.4
44	176	186.8	96.6
46	184	191.3	99.9
48	192	195.7	103
50	200	200	106
55	220	210.3	113.9
60	240	220	121.3
65	260	229.6	128.5
70	280	238.7	135.6
75	300	247.4	142.4
80	320	255.9	148
85	340	264	155.6
90	360	272	162
95	380	279.7	168.3
100	400	287	174.4
125	500	322	203.5
150	600	353.5	230.3
175	700	382.4	255.4
200	800	409	279
225	900	434.4	301.4
250	1000	458	322.8
275	1100	480.8	343.3
300	1200	502.4	362.9

$$X_{L2} = R_L B = 50 \times 1.24 = 62 \text{ ohms}$$

where $B = \sqrt{\left(\frac{A}{R_L}\right) - 1}$

and $A = R1(Q_L^2 + 1)$

$(B = 1.24)$ and $(A = 127.5)$ (B)

$$X_{C1} = \frac{A}{Q_L + B} = \frac{127.5}{4 + 1.24}$$

$= 24.3$ ohms (C)

(Eq. 6)

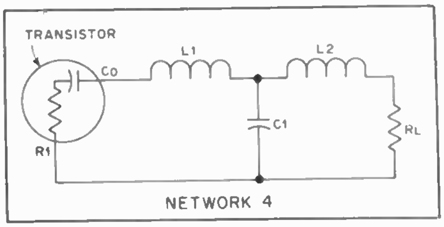


Fig. 6 - The circuit for network 4.

Table 6 provides values of reactance for the *C* and *L* components of the T network. The table is based on *Q*s from 1 to 4, and R1 values from 12 to 300 ohms.

Pi and Pi-L Networks for Tube Circuits

When working with vacuum-tube rf amplifiers, it is common practice to use pi and pi-L networks (Fig. 7) for matching the plate of an amplifier to the load. In the interest of harmonic reduction the designer can use Q_L values which are relatively high – 10 to 16 being typical. The pi-network equations given earlier for transistor amplifiers are suitable for designing tube pi networks as well. The equations will yield values of reactance for *L* and *C* which will closely approximate the reactances of the values listed in Table 7. The *L* and *C* values provided in Table 6 were supplied by W6FFC.

Table 7

Pi and Pi-L Network Values

Tube Load Impedance (Operating *Q*)

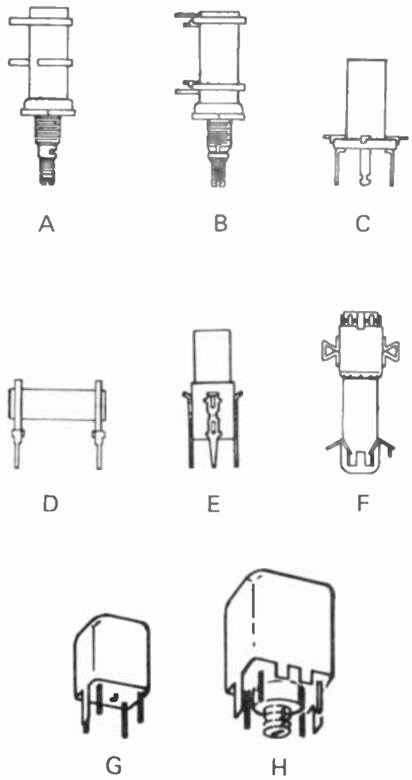
	MHz	*1500(12)*	*2000(12)*	*2500(12)*	*3000(12)*	*3500(12)*	*4000(12)*	*5000(13)*	*6000(14)*	*8000(16)*
C1	3.5	420	315	252	210	180	157	126	114	99
	7	190	143	114	95	82	71	57	52	45
	14	93	70	56	47	40	35	28	25	22
	21	62	47	37	31	27	23	19	17	15
	28	43	32	26	21	18	16	13	12	10
C2	3.5	2117	1776	1536	1352	1203	1079	875	862	862
	7	942	783	670	583	512	451	348	341	341
	14	460	382	326	283	247	217	165	162	162
	21	305	253	216	187	164	144	109	107	107
	28	210	174	148	128	111	97	72	70	70
L1	3.5	5.73	7.46	9.17	10.86	12.53	14.19	17.48	19.18	21.98
	7	3.14	4.09	5.03	5.95	6.86	7.77	9.55	10.48	12.02
	14	1.60	2.08	2.56	3.03	3.49	3.95	4.85	5.33	6.11
	21	1.07	1.39	1.71	2.02	2.34	2.64	3.25	3.56	4.09
	28	0.77	1.01	1.24	1.46	1.69	1.91	2.34	2.57	2.95

Tube Load Impedance (Operating *Q*)

	MHz	*1500(12)*	*2000(12)*	*2500(12)*	*3000(12)*	*3500(12)*	*4000(12)*	*5000(12)*	*6000(12)*	*8000(12)*
C3	3.5	406	305	244	203	174	152	122	102	76
	7	188	141	113	94	81	71	56	47	35
	14	92	69	55	46	40	35	28	23	17
	21	62	46	37	31	26	23	18	15	12
	28	43	32	26	21	18	16	13	11	8
C4	3.5	998	859	764	693	638	593	523	472	397
	7	430	370	329	298	274	255	225	203	171
	14	208	179	159	144	133	123	109	98	83
	21	139	119	106	96	89	82	73	65	55
	28	95	81	72	66	60	56	50	45	38
L2	3.5	7.06	9.05	10.99	12.90	14.79	16.67	20.37	24.03	31.25
	7	3.89	4.97	6.03	7.07	8.10	9.12	11.13	13.11	17.02
	14	1.99	2.54	3.08	3.61	4.13	4.65	5.68	6.69	8.68
	21	1.33	1.69	2.05	2.41	2.76	3.10	3.78	4.46	5.78
	28	0.96	1.22	1.48	1.74	1.99	2.24	2.73	3.22	4.17
L3	3.5	4.45	4.45	4.45	4.45	4.45	4.45	4.45	4.45	4.45
	7	2.44	2.44	2.44	2.44	2.44	2.44	2.44	2.44	2.44
	14	1.24	1.24	1.24	1.24	1.24	1.24	1.24	1.24	1.24
	21	0.83	0.83	0.83	0.83	0.83	0.83	0.83	0.83	0.83
	28	0.60	0.60	0.60	0.60	0.60	0.60	0.60	0.60	0.60

Table 8

STYLE	*FORM MATERIAL*	*FORM DIMENSIONS (IN.)*	*MOUNTING*	*TERMINALS*
A	Ceramic	0.205 × 0.593	Bushing	2 Wire rings
B	Ceramic	0.205 × 0.593	Bushing	2 Collars with 1 Terminal on each
B	Ceramic	0.205 × 0.593	Bushing	2 Collars with 2 Terminals on each
A	Ceramic	0.260 × 0.858	Bushing	2 Wire rings
B	Ceramic	0.260 × 0.858	Bushing	2 Collars with 1 Terminal on each
B	Ceramic	0.260 × 0.858	Bushing	2 Collars with 2 Terminals on each
A	Ceramic	0.375 × 1.060	Bushing	2 Wire rings
B	Ceramic	0.375 × 1.060	Bushing	2 Collars with 1 Terminal on each
B	Ceramic	0.375 × 1.060	Bushing	2 Collars with 2 Terminals on each
A	Phenolic (PBG)	0.250 × 0.860	Bushing	2 Wire rings
A	Phenolic (PBG)	0.375 × 1.125	Bushing	2 Wire rings
C	Alpha Cellulose	0.162 × 0.400	Printed Circuit	3 Terminals
C	Alpha Cellulose	0.162 × 0.400	Printed Circuit	4 Terminals
D	Alpha Cellulose	0.162 × 0.520	Printed Circuit	2 Collars with 2 Terminals on each
C	Alpha Cellulose	0.211 × 0.625	Printed Circuit	4 Terminals
D	Alpha Cellulose	0.211 × 0.680	Printed Circuit	2 Collars with 2 Terminals on each
C	Ceramic	0.260 × 0.625	Printed Circuit	3 Terminals
C	Ceramic	0.260 × 0.625	Printed Circuit	4 Terminals
D	Ceramic	0.260 × 0.625	Printed Circuit	2 Collars with 2 Terminals on each
E	Alpha Cellulose	0.281 × 0.750	Printed Circuit	3 Terminals
E	Alpha Cellulose	0.281 × 0.750	Printed Circuit	4 Terminals
F	Nylon	0.285 × 1.250	Clip	2 Terminals



Courtesy of J. W. Miller Co., Compton, CA 90224

Table 9

1/4" Dia. Slug-Tuned, Style-A Ceramic-Form Coils

MINIMUM CORE POSITION INDUCTANCE MINIMUM	Q MIN.	TEST FREQUENCY	MAXIMUM CORE POSITION INDUCTANCE MAXIMUM	Q MIN.	TEST FREQUENCY	R OHMS MAX.	MAX. CURRENT mA	MIN. Fo MHz
0.440 µH	80	25 MHz	.760 µH	52	25 MHz	.03	1600	142
1.10 µH	72	25 MHz	1.50 µH	40	7.9 MHz	.06	1000	96
1.70 µH	51	7.9 MHz	2.70 µH	36	7.9 MHz	0.11	636	80
3.10 µH	56	7.9 MHz	4.80 µH	33	7.9 MHz	0.23	400	58
5.50 µH	60	7.9 MHz	8.60 µH	33	7.9 MHz	0.49	256	45
9.90 µH	52	7.9 MHz	15 µH	41	2.5 MHz	1.5	100	32
17 µH	47	2.5 MHz	23 µH	53	2.5 MHz	2.3	100	19
26 µH	48	2.5 MHz	33 µH	51	2.5 MHz	2.9	100	16
38 µH	50	2.5 MHz	57 µH	48	2.5 MHz	3.4	100	12
66 µH	44	2.5 MHz	114 µH	40	.79 MHz	4.1	100	5.2
120 µH	46	0.79 MHz	190 µH	40	.79 MHz	5.7	100	4.1
209 µH	45	0.79 MHz	314 µH	32	.79 MHz	7.7	100	3.2
350 µH	53	0.79 MHz	475 µH	41	.79 MHz	10	100	2.9
528 µH	44	0.79 MHz	760 µH	40	.79 MHz	14	100	2.5

3/8" Dia. Slug-Tuned, Style-A Ceramic-Form Coils

MINIMUM CORE POSITION INDUCTANCE MINIMUM	Q MIN.	TEST FREQUENCY	MAXIMUM CORE POSITION INDUCTANCE MAXIMUM	Q MIN.	TEST FREQUENCY	R OHMS MAX.	MAX. CURRENT mA	MIN. Fo MHz
0.990 µH	18	25 MHz	1.50 µH	57	7.9 MHz	.04	1600	75
1.60 µH	67	7.9 MHz	3.10 µH	52	7.9 MHz	.08	1000	51
3.30 µH	70	7.9 MHz	6.50 µH	44	7.9 MHz	0.17	636	41
7.30 µH	50	7.9 MHz	14 µH	35	2.5 MHz	0.57	256	26
16 µH	52	2.5 MHz	29 µH	56	2.5 MHz	2.2	100	7.6
33 µH	51	2.5 MHz	66 µH	37	2.5 MHz	3.1	100	4.9
74 µH	42	2.5 MHz	124 µH	44	0.79 MHz	4.5	100	3.7
138 µH	47	0.79 MHz	238 µH	44	0.79 MHz	6.6	100	3.0
270 µH	57	0.79 MHz	451 µH	43	0.79 MHz	8.9	100	2.2
495 µH	56	0.79 MHz	760 µH	41	0.79 MHz	12	100	2
825 µH	41	0.79 MHz	1.30 mH	27	0.25 MHz	20	100	1.7
1.40 µH	40	0.25 MHz	2.00 mH	33	0.25 MHz	25	100	1.6

Courtesy of J. W. Miller Co., Compton, CA 90224

Table 10
Miniature Style-E PC-Mount Rf Inductors

MINIMUM CORE POSITION INDUCTANCE MINIMUM	Q MIN.	TEST FREQUENCY	MAXIMUM CORE POSITION INDUCTANCE MAXIMUM	Q MIN.	TEST FREQUENCY	R OHMS MAX.	MAX. CURRENT mA	MIN. Fo MHz
0.095 µH	77	25.0 MHz	0.125 µH	94	25.0 MHz	.02	4100	350
0.130 µH	68	25.0 MHz	0.170 µH	92	25.0 MHz	.02	1600	300
0.185 µH	88	25.0 MHz	0.265 µH	100	25.0 MHz	.02	1600	230
0.285 µH	88	25.0 MHz	0.410 µH	93	25.0 MHz	.03	1000	198
0.420 µH	100	25.0 MHz	0.580 µH	80	25.0 MHz	.03	2500	150
0.540 µH	101	25.0 MHz	0.850 µH	89	25.0 MHz	.03	1600	136
0.640 µH	101	25.0 MHz	1.00 µH	78	25.0 MHz	.03	1600	118
0.760 µH	98	25.0 MHz	1.25 µH	70	7.9 MHz	.04	1600	114
1.20 µH	65	7.9 MHz	1.87 µH	70	7.9 MHz	.06	1000	89
1.65 µH	61	7.9 MHz	2.75 µH	65	7.9 MHz	0.14	400	77
2.40 µH	64	7.9 MHz	4.10 µH	60	7.9 MHz	0.17	400	62
3.40 µH	68	7.9 MHz	5.80 µH	60	7.9 MHz	0.24	400	53
4.60 µH	64	7.9 MHz	8.50 µH	56	7.9 MHz	0.39	250	45
5.60 µH	64	7.9 MHz	10.0 µH	57	2.5 MHz	0.64	160	40
7.10 µH	68	7.9 MHz	12.5 µH	55	2.5 MHz	0.77	160	38
10.0 µH	58	2.5 MHz	18.7 µH	95	2.5 MHz	1.68	100	11.7
14.8 µH	61	2.5 MHz	27.5 µH	90	2.5 MHz	1.91	100	8.4
22.0 µH	60	2.5 MHz	41.0 µH	75	2.5 MHz	2.34	100	6.7
31.0 µH	58	2.5 MHz	58.0 µH	68	2.5 MHz	2.72	100	5.6
43.5 µH	56	2.5 MHz	85.0 µH	55	2.5 MHz	3.30	100	4.6
61.0 µH	48	2.5 MHz	100.0 µH	88	790.0 kHz	3.89	100	4.3
76.0 µH	52	2.5 MHz	125.0 µH	90	790.0 kHz	4.39	100	3.8
105.0 µH	57	790.0 kHz	187.0 µH	92	790.0 kHz	5.46	100	3.3
160.0 µH	63	790.0 kHz	275.0 µH	90	790.0 kHz	6.70	100	2.9
240.0 µH	66	790.0 kHz	410.0 µH	90	790.0 kHz	8.30	100	2.5
360.0 µH	68	790.0 kHz	580.0 µH	81	790.0 kHz	10.50	100	2.1
530.0 µH	66	790.0 kHz	850.0 µH	75	790.0 kHz	12.90	100	1.75
700.0 µH	64	790.0 kHz	1.00 mH	80	250.0 kHz	14.90	100	1.70
910.0 µH	66	790.0 kHz	1.25 mH	85	250.0 kHz	17.10	100	1.61
990.0 µH	35	250.0 kHz	1.87 mH	60	250.0 kHz	28.20	65	.73
1.60 mH	39	250.0 kHz	2.75 mH	62	250.0 kHz	34.80	65	.62
2.40 mH	41	250.0 kHz	4.10 mH	60	250.0 kHz	42.90	65	.60
3.40 mH	42	250.0 kHz	5.80 mH	57	250.0 kHz	51.60	65	.53
5.15 mH	42	250.0 kHz	8.50 mH	50	250.0 kHz	63.60	65	.50
7.40 mH	42	250.0 kHz	10.00 mH	50	79.0 kHz	75.60	65	.40
9.80 mH	40	250.0 kHz	12.50 mH	52	79.0 kHz	87.30	65	.38
12.00 mH	39	79.0 kHz	18.70 mH	55	79.0 kHz	111.0	65	.32
12.10 mH	20	79.0 kHz	27.50 mH	51	79.0 kHz	197.0	33	.26
18.20 mH	24	79.0 kHz	41.00 mH	54	79.0 kHz	244.0	33	.21
27.50 mH	28	79.0 kHz	58.00 mH	56	79.0 kHz	302.0	33	.20
40.00 mH	32	79.0 kHz	85.00 mH	56	79.0 kHz	378.0	33	.16
50.00 mH	34	79.0 kHz	100.00 mH	58	79.0 kHz	423.0	33	.15
62.00 mH	35	79.0 kHz	125.00 mH	56	79.0 kHz	468.0	33	.14

Courtesy of J. W. Miller Co., Compton, CA 90224

Table 11

Single-Pi Ferrite-Core Rf Chokes

L ± 5%	*MIN. Q AT FREQ. kHz*	*MIN. Fo MHz*	*MAX. R OHMS*	*APPROX. COIL DIA.*	*MAX. mA*
100 μH	140 @ 790	7.7	2.0	17/64	250
120 μH	140 @ 790	7.0	2.2	17/64	250
150 μH	160 @ 790	6.3	2.5	17/64	250
180 μH	160 @ 790	6.0	2.9	9/32	250
220 μH	160 @ 790	5.6	3.2	19/64	250
270 μH	160 @ 790	5.1	3.6	19/64	250
330 μH	160 @ 790	4.8	3.8	5/16	250
390 μH	160 @ 790	4.4	4.2	21/64	225
470 μH	160 @ 790	3.9	4.8	21/64	200
560 μH	160 @ 790	3.6	5.3	11/32	200
680 μH	160 @ 790	3.4	6.0	23/64	200
820 μH	150 @ 790	3.1	6.8	3/8	200
1.0 mH	140 @ 790	2.6	7.5	13/32	150
1.2 mH	140 @ 250	2.2	7.8	31/64	150
1.5 mH	140 @ 250	2.0	8.8	33/64	150
1.8 mH	140 @ 250	1.8	11.0	35/64	150
2.2 mH	100 @ 250	1.5	12.0	3/8	150
2.7 mH	100 @ 250	1.4	13.5	13/32	125
3.3 mH	100 @ 250	1.3	15.1	7/16	125
3.9 mH	95 @ 250	1.2	18.0	29/64	125
4.7 mH	95 @ 250	1.0	21.5	15/32	100
5.6 mH	95 @ 250	1.0	25.0	1/2	80
6.8 mH	95 @ 250	.90	29.0	17/32	80
8.2 mH	70 @ 250	.91	30.0	31/64	80
10.0 mH	65 @ 250	.85	34.0	33/64	80

Single-Layer Rf Chokes

L	*MIN. Q @ F*	*MIN. Fo. MHz*	*R OHMS MAX.*	*1 mA MAX.*	*CORE MATERIAL*
1.0 μH	45 @ 7.9 MHz	238	0.20	1000	Phenolic
1.2 μH	45 @ 7.9 MHz	218	0.22	950	Phenolic
1.5 μH	45 @ 7.9 MHz	200	0.25	900	Phenolic
1.8 μH	45 @ 7.9 MHz	180	0.28	850	Phenolic
2.2 μH	45 @ 7.9 MHz	165	0.30	800	Phenolic
2.7 μH	45 @ 7.9 MHz	149	0.50	700	Phenolic
3.3 μH	45 @ 7.9 MHz	136	0.70	600	Phenolic
3.9 μH	45 @ 7.9 MHz	127	0.80	500	Phenolic
4.7 μH	50 @ 7.9 MHz	114	1.00	400	Phenolic
5.6 μH	50 @ 7.9 MHz	104	1.80	350	Phenolic
6.8 μH	50 @ 7.9 MHz	94.8	1.85	300	Phenolic
8.2 μH	50 @ 7.9 MHz	86.1	1.90	275	Phenolic
10 μH	50 @ 7.9 MHz	78.5	3.00	250	Phenolic
12 μH	30 @ 2.5 MHz	71.6	3.60	200	Phenolic
15 μH	30 @ 2.5 MHz	63.9	6.00	150	Phenolic
18 μH	30 @ 2.5 MHz	57.5	7.50	100	Phenolic
22 μH	85 @ 2.5 MHz	36.1	2.00	500	Iron
27 μH	80 @ 2.5 MHz	33.2	1.85	450	Iron
33 μH	80 @ 2.5 MHz	30.6	2.00	460	Iron
39 μH	90 @ 2.5 MHz	26.6	2.60	400	Iron
47 μH	90 @ 2.5 MHz	24.9	3.50	350	Iron
56 μH	90 @ 2.5 MHz	23.6	3.75	300	Iron
68 μH	90 @ 2.5 MHz	21.6	4.00	250	Iron
82 μH	100 @ 2.5 MHz	19.0	5.10	200	Iron
100 μH	100 @ 2.5 MHz	17.8	6.00	150	Iron

Courtesy of J. W. Miller Co., Compton, CA 90224

Inductors for Networks

The inductors chosen for a given network application must satisfy the requirements of power-handling capability, circuit Q, adjustability, and physical size. For the "sloppy networks" discussed earlier in this chapter the builder may elect to use slug-tuned coils for the inductances. In networks where it isn't necessary to make L variable, air-wound or toroidal-wound inductors can be employed. The self-shielding and high-Q properties of the latter may be of special interest in some situations.

Table 8 lists some of the more common coil forms with powdered iron or ferrite slugs. A catalog of numerous capacitors, blank coil forms, and pre-wound inductors can be obtained from the J. W. Miller Co., Compton, CA 90224. Table 8 contains a partial listing of the forms available.

Table 9 lists standard values of variable inductance for 1/4-inch diameter ceramic-form coils. Given also are values of variable inductance available on 3/8-inch diameter ceramic forms.

For use in printed-circuit board projects, the builder may desire miniature pc-mount variable inductors. Table 10 lists a variety of inductance values which should be useful in amateur construction work.

Rf chokes are frequently used as part of an LC network. They are used to provide a high-impedance feed circuit when applying operating voltage to the collector or base of a transistor, or when supplying operating voltage to the grid or plate of a tube. Table 11 gives data on the standard values of inductance for miniature rf chokes. In some networks the choke values will be suitable for use as the inductors.

Some rf chokes used in vacuum-tube circuits must be capable of passing considerable dc current. Table 12 lists standard choke values and provides information on the dc-current capabilities of each value given.

Table 12

Freq. Selective Rf Chokes (600 mA)

INDUCTANCE ±10%	*FREQ. MHz*
266 μH	1.8-5
208 μH	3-13
84 μH	7-20
38.5 μH	15-30
24 μH	25-40
8.2 μH	30-90
1.72 μH	75-180
0.82 μH	160-340
0.22 μH	350-500

Ceramic Form Rf Chokes

INDUCTANCE ±5% @ 1000 Hz	*RESISTANCE MAX.*	*RATED CURRENT*
0.25 mH	9.6	150 mA
0.50 mH	14.4	150 mA
1.0 mH	20.4	150 mA
2.5 mH	33.6	150 mA
5.0 mH	54.0	150 mA
0.25 mH	4.20	300 mA
0.50 mH	6.60	300 mA
1.0 mH	11.5	300 mA
2.5 mH	20.4	300 mA
5.0 mH	22.8	300 mA

Single-Layer-Wound Chokes

μH	*OHMS MAX.*	*mA*
1	.036	300
2.5	0.108	300
5	0.300	300
10	1.14	300

Progressive-Wound Chokes

μH	*OHMS MAX.*	*mA*
25	1.92	300
50	2.52	300
100	3.60	300

Multiple Pi-Wound Chokes

mH	*OHMS MAX.*	*mA*
0.25	9.6	200
0.5	14.4	200
1	20.4	200
1.5	25.2	200
2.5	33.6	200
5	50.4	125
7.5	98.4	125
10	114	125
25	192	125

Courtesy of J. W. Miller Co., Compton, CA 90224

Table 13

Powdered-Iron Toroid Cores

Inductance and Turns Formula

The number of turns (N) necessary to obtain a specific inductance (L) can be calculated by

$$N = 100\sqrt{\frac{L}{L_{100}}}$$

L = desired inductance (μH)

L_{100} = core inductance (μH per 100 turns)

Inductance Per 100 Turns ±5%

NO.	*MIX-41*	*MIX-3*	*MIX-15*	*MIX-1*	*MIX-2*	*MIX-6*	*MIX-10*	*MIX-12*
T-200	755 μH	360 μH			120 μH	105 μH		
T-184	1640 μH	720 μH			240 μH			
T-157	970 μH	420 μH			140 μH	115 μH		
T-130	785 μH	330 μH	250 μH	200 μH	110 μH	96 μH		
T-106	900 μH	405 μH	330 μH	280 μH	135 μH	116 μH		
T-94	590 μH	248 μH		160 μH	84 μH	70 μH	58 μH	32 μH
T-80	450 μH	180 μH	170 μH	115 μH	55 μH	45 μH	34 μH	22 μH
T-68	420 μH	195 μH	180 μH	115 μH	57 μH	47 μH	32 μH	21 μH
T-50	320 μH	175 μH	135 μH	100 μH	50 μH	40 μH	31 μH	18 μH
T-37	240 μH	110 μH	90 μH	80 μH	42 μH	30 μH	25 μH	15 μH
T-25	200 μH	100 μH	85 μH	70 μH	34 μH	27 μH	19 μH	13 μH
T-12	90 μH	60 μH		48 μH	24 μH	19 μH	12 μH	8.5 μH

Magnetic Properties

Iron Powder Cores

MATERIAL	*COLOR CODE*	*PERMEABILITY*	*TEMPERATURE STABILITY*	*TYPICAL FREQUENCY RANGE*	*OPTIMUM FREQUENCY RANGE*
41' HA'	Green	μ = 75	975 ppm/c	10 kHz-100 kHz	20 kHz- 50 kHz
3' HP'	Gray	μ = 35	370 ppm/c	20 kHz- 2 MHz	50 kHz-500 kHz
15' GS6'	Rd. & Wh.	μ = 25	190 ppm/c	20 kHz- 5 MHz	500 kHz- 1 MHz
1' C'	Blue	μ = 20	280 ppm/c	40 kHz- 5 MHz	1 MHz- 2 MHz
2 E'	Red	μ = 10	95 ppm/c	200 kHz- 30 MHz	2 MHz- 10 MHz
6' SF'	Yellow	μ = 8	35 ppm/c	2 MHz- 50 MHz	10 MHz- 20 MHz
10' W'	Black	μ = 6	150 ppm/c	4 MHz-100 MHz	20 MHz- 40 MHz
12' Irn-8'	Gr. & Wh.	μ = 3	170 ppm/c	10 MHz-200 MHz	40 MHz- 90 MHz
0' Ph'	Tan	μ = 1	—	50 MHz-300 MHz	90 MHz-150 MHz

Courtesy of Amidon Assoc., N. Hollywood, CA 91607

Other Magnetic-Core Inductances

Ferrite and powdered-iron toroid cores are preferred for use in most low-power modern *LC* networks. The core selected for a given application must exhibit certain properties, consistent with the performance needed. The considerations are size, permeability (μ factor), and saturation characteristics. Ferrite cores are the most common for use in broadband transformers (discussed later in this book), but are useful in tuned circuits as well. Powdered-iron toroid cores are used most frequently in rf networks which are resonant.

Table 13 contains data on powdered-iron cores. Information is provided on color codes, material designators, optimum frequency of use, and sample inductance amounts for 100 turns of wire wound on various cores.

Additional design data are contained in Table 14. Here the builder can learn what sizes are available for most amateur circuit work. A catalog of cores and prices can be obtained from Amidon Associates, N. Hollywood, CA 91607.

Knowledge of the properties of ferrite toroids is of value to the amateur designer. Table 15 shows most of the parameters which bear significance. Of particular interest is the matter of permeability and frequency rating.

Table 16 carries a partial listing of the core sizes available in ferrite material. Dimensions are supplied in metric and English units. The cores are available "tumbled" (smooth edges) as are the powdered-iron ones listed earlier.

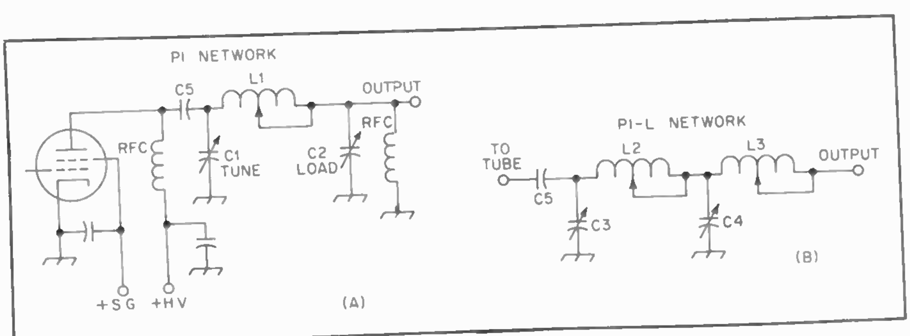


Fig. 7 — Diagram of tube-type pi network at A. The circuit at B is that of a pi-L network.

Air Inductors

Standard values of inductance are available in air-wound single-layer form. Various wire sizes and diameters can be obtained for use in *LC* networks of the variety treated in this chapter. Ordinarily, the Q_u of this type of coil is very high. Air-core inductors are used extensively in vhf solid-state circuits, and in hf and vhf tube circuits. Table 17 lists the commercial air-core inductors which are available.

Table 14

Powdered-Iron Toroid Cores

Red E Cores-500 kHz to 30 MHz (μ = 10)

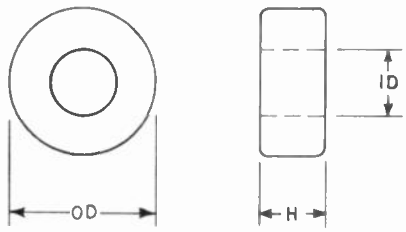
NO.	*OD (IN.)*	*ID(IN.)*	*H (IN.)*
T-200-2	2.00	1.25	0.55
T-94-2	0.94	0.56	0.31
T-80-2	0.80	0.50	0.25
T-68-2	0.68	0.37	0.19
T-50-2	0.50	0.30	0.19
T-37-2	0.37	0.21	0.12
T-25-2	0.25	0.12	.09
T-12-2	0.125	.06	.05

Black W Cores-30 MHz to 200 MHz (μ = 7)

NO.	*OD (IN.)*	*ID (IN.)*	*H (IN.)*
T-50-10	0.50	0.30	0.19
T-37-10	0.37	0.21	0.12
T-25-10	0.25	0.12	.09
T-12-10	0.125	.06	.05

Yellow SF Cores-10 MHz to 90 MHz (μ = 8)

NO.	*OD (IN.)*	*ID (IN.)*	*H (IN.)*
T-94-6	0.94	0.56	0.31
T-80-6	0.80	0.50	0.25
T-68-6	0.68	0.37	0.19
T-50-6	0.50	0.30	0.19
T-26-6	0.25	0.12	.09
T-12-6	0.125	.06	.05



Number of Turns vs. Wire Size and Core Size

Approximate maximum number of turns — single layer wound — enameled wire.

WIRE SIZE	*T-200*	*T-130*	*T-106*	*T-94*	*T-80*	*T-68*	*T-50*	*T-37*	*T-25*	*T-12*
10	33	20	12	12	10	6	4	1		
12	43	25	16	16	14	9	6	3		
14	54	32	21	21	18	13	8	5	1	
16	69	41	28	28	24	17	13	7	2	
18	88	53	37	37	32	23	18	10	4	1
20	111	67	47	47	41	29	23	14	6	1
22	140	86	60	60	53	38	30	19	9	2
24	177	109	77	77	67	49	39	25	13	4
26	223	137	97	97	85	63	50	33	17	7
28	281	173	123	123	108	80	64	42	23	9
30	355	217	154	154	136	101	81	54	29	13
32	439	272	194	194	171	127	103	68	38	17
34	557	346	247	247	218	162	132	88	49	23
36	683	424	304	304	268	199	162	108	62	30
38	875	544	389	389	344	256	209	140	80	39
40	1103	687	492	492	434	324	264	178	102	51

Actual number of turns may differ from above figures according to winding techniques, especially when using the larger size wires. Chart prepared by Michael J. Gordon, Jr. – WB9FHC

Courtesy of Amidon Assoc., N. Hollywood, CA 91607

Table 15

Magnetic Properties of Toroid Cores (Ferrite)

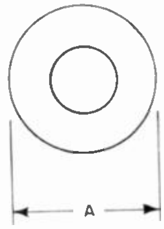
PROPERTY	*UNIT*	*SYMBOL*	*63*	*61*	*43*	*72*	*75*
Initial Permeability		μ_i	40	125	950	2000	5000
Maximum Permeability		μ_M	125	450	3000	3500	8000
Saturation Flux Density @ 13 oer.	Gauss	B_s	1850	2350	2750	3500	3900
Residual Flux Density	Gauss	B_r	750	1200	1200	1500	1250
Curie Temperature	°C	T_c	500	300	130	150	160
Volume Resistivity	OHM-CM	ρ	1×10^8	1×10^8	1×10^5	1×10^2	5×10^2
Optimum Frequency Range	MHz		15-25	.2-10	.01-1	.001-1	.001-1
Specific Gravity			4.7	4.7	4.5	4.8	4.8
Loss Factor	$\frac{1}{\mu_i Q}$		9.0×10^{-5} @ 25 MHz	2.2×10^{-5} @ 2.5 MHz	2.5×10^{-5} @ .2 MHz	9×10^{-6} @ .1 MHz	5×10^{-6} @ .1 MHz
Coercive Force	Oersteds	H_c	2.40	1.60	0.30	0.18	0.18
Temp. Coefs. of Initial Permeability	%/°C (20°C-70°C)		0.10	0.10	0.20	0.60	

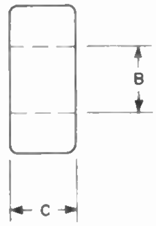
Courtesy of Amidon Assoc., N. Hollywood, CA 91607

Table 16

Ferrite Toroid Cores and Properties

TYPE NO.		PHYSICAL DIMENSIONS			MAGNETIC DIMENSIONS		
		A DIM.	*B DIM.*	*C DIM.*	A_e (in.)^2 (mm)^2	l_e (in.)^2 (mm)	V_e (in.)^2 (mm)^3
FT-23	in.	0.225/.235	0.115/.125	.055/.065	.00330	0.529	.001747
	mm	5.71/5.96	2.92/3.17	1.39/1.65	2.13	13.44	28.63
FT-37	in.	0.367/.383	0.182/.192	0.120/.130	.01175	0.846	.00994
	mm	9.32/9.72	4.62/4.87	3.04/3.30	7.58	21.49	162.9
FT-50	in.	0.490/.510	0.274/.289	0.183/.193	.0206	1.190	.0245
	mm	12.45/12.95	6.95/7.34	4.65/4.90	13.29	30.23	401.6
FT-82	in.	0.810/.840	0.508/.532	0.243/.257	.0381	.207	.0789
	mm	20.57/21.33	12.90/13.51	6.17/6.52	24.58	52.58	1293
FT-114	in.	1.112/1.172	0.728/.768	0.284/.306	.0581	2.92	0.1695
	mm	28.24/29.76	18.49/19.50	7.21/7.77	37.49	74.17	2778





Available in the Following Materials

43
61
63
72
75

Symbols & Their Definitions

A_e = Effective magnetic cross-sectional area.
l_e = Effective magnetic path length.
V_e = Effective magnetic volume.
A_L = Inductance in 10^{-9} Henries for 1 turn.
A_s = Surface area of core exposed for cooling.
A_w = Total window area of core.

Courtesy of Amidon Assoc., N. Hollywood, CA 91607

Table 17

Air-Core Inductors

MODEL NUMBER	*COIL DIAMETER (IN.)*	*TURNS PER INCH*	*LENGTH (IN.)*	*WIRE SIZE*	*INDUCTANCE (APPROX.) (μH)*
3001	1/2	4	2	18	0.40
3002	1/2	8	2	18	0.96
3003	1/2	16	2	20	3.2
3004	1/2	32	2	24	13.7
3005	5/8	4	2	16	0.56
3006	5/8	8	2	18	1.4
3007	5/8	16	2	20	4.9
3008	5/8	32	2	24	19.2
3009	3/4	4	3	16	0.94
3010	3/4	8	3	18	2.9
3011	3/4	16	3	20	10.9
3012	3/4	32	3	24	42.5
3013	1	4	3	16	1.9
3014	1	8	3	18	4.8
3015	1	16	3	20	19.9
3016	1	32	3	24	73.0
3017	1-1/4	4	4	14	2.56
3018	1-1/4	8	4	16	9.4
3019	1-1/4	16	4	18	37.5
3020	1-1/4	32	4	24	145
3021	1-3/4	4	4	14	4.5
3022	1-3/4	8	4	14	17.2
3023	1-3/4	16	4	18	72.0
3024	1-3/4	32	4	24	280
3025	2	6	10	12	33
3026	2	8	10	14	60
3027	2	10	10	16	92
3029	2-1/2	6	10	12	52
3030	2-1/2	8	10	14	92
3031	2-1/2	10	10	16	142
3033	3	6	10	12	74
3034	3	8	10	14	135
3035	3	10	10	16	200

Courtesy Barker and Williamson Co.

Chapter 5

Transformers

Transformers take many forms – audio, narrow-band rf, wide-band rf and power types. A significant characteristic of each class is the primary and secondary voltages of the transformer. The voltage ratio can be translated into turns ratio or impedance ratio. When the primary voltage is known, the secondary voltage can be obtained by $E_s = E_p \times n$ where E_s = secondary voltage, E_p = primary voltage, and n is the turns ratio of the windings. The turns ratio is obtained by dividing E_p by E_s.

A 60-Hz power transformer has 150 primary turns and 475 secondary ones. If 117 volts ac is applied to the primary winding, the secondary voltage will be found by

$$n = \frac{475}{150} = 3.166$$

$$\therefore E_s = E_p \times n = 117 \times 3.166$$

$$= 370.4 \text{ V} \qquad \text{(Eq. 1)}$$

If an audio transformer with the same turns ratio was used, the impedance ratio of the windings would be determined by

$$Z_{ratio} = n^2 = 3.166^2 = 10.02 \qquad \text{(Eq. 2)}$$

Therefore, if the primary impedance was 500 ohms, the secondary impedance would be $Z_p \times Z_{ratio}$. The secondary impedance would be 5010 ohms.

Transformers can be air wound (no magnetic core) for rf applications. Powdered-iron or ferrite core material is also used at rf, and so is brass (at vhf and above). Iron, Hypersil, powdered-iron and ferrite materials are used at audio and below.

Transformer Efficiency

A lossless transformer should have the same power in the secondary that is found in the primary. As was stated earlier in this book, efficiency (percentage) = $P_1 \div P_2$. Therefore, transformer efficiency = $P_s \div P_p$ where P_s is the secondary-winding power and P_p is the power in the primary. A lossless transformer, which can seldom be realized completely, would be represented by $E_p I_p = E_s I_s$.

A step-down transformer has a turns ratio of 6 to 1. The primary voltage is 117, and the primary current is 0.5 A. How can one determine the secondary current? The secondary voltage is $117 \div 6$, or 19.5 V. Therefore

$$E_p I_p = E_s I_s \;\therefore\; I_s = \frac{E_p I_p}{E_s}$$

$$= \frac{117 \times 0.5}{19.5} = \frac{58.5}{19.5} = 3\text{A} \qquad \text{(Eq. 3)}$$

This example assumes no transformer losses. It can be seen that the secondary current increased by the same ratio that the secondary voltage decreased.

If the transformer impedances are known, but the voltage or turns ratios are not, the latter can be found from

$$n = \sqrt{\frac{Z1}{Z2}} = \sqrt{\frac{500}{10{,}000}} = 0.223$$

where Z1 = 500 ohms, and Z2 = 10,000 ohms (Eq. 4)

The turns ratio must be 1 to 0.223, step down, total primary to total secondary.

Table 1 gives pertinent data on standard wire sizes for winding transformers. The nearest British SWG equivalents are listed in the table.

Table 2 contains a listing of standard values of voltage and current for U.S. filament transformers. The designer can use the table for determining the physical sizes of transformers for a given project.

Data for miniature transistor-circuit audio transformers are provided in Table 3. The information supplied concerning primary and secondary impedances should be useful in designing amplifier stages to work with the impedances available. Some power transformers which are available in the USA are itemized in Table 4. Voltage, current, weight and size data are given.

Core Characteristics

Toroid cores are made from powdered-iron or ferrite materials. Their greatest application in amateur work is in rf circuitry. Manufacturers of toroid cores can provide data sheets which contain pertinent information about their cores – permeability (μ), physical dimensions, cross-sectional area (A_e), gauss (b), inductance index (A_L), and other important facts. Table 5 lists the terms and their meanings.

There are specific limits to the amount of power a given core can handle before being saturated or

Table 1

Copper-Wire Information

WIRE SIZE A.W.G. (B&S)	DIA. IN MILS[1]	CIRC. MIL AREA	TURNS PER LINEAR INCH[2] ENAM.	S.C.E.	D.C.C.	CONT.-DUTY CURRENT[3] SINGLE WIRE IN OPEN AIR	CONT.-DUTY CURRENT[3] WIRES OR CABLES IN CONDUITS OR BUNDLES	FEET PER POUND, BARE	OHMS PER 1000 FT. 25°C	CURRENT CARRYING CAPACITY[4] AT 700 C.M. PER AMP.	DIA. IN MM.	NEAREST BRITISH S.W.G. NO.
1	289.3	83690	–	–	–	–	–	3.947	0.1264	119.6	7.348	1
2	257.6	66370	–	–	–	–	–	4.977	0.1593	94.8	6.544	3
3	229.4	52640	–	–	–	–	–	6.276	0.2009	75.2	5.827	4
4	204.3	41740	–	–	–	–	–	7.914	0.2533	59.6	5.189	5
5	181.9	33100	–	–	–	–	–	9.980	0.3195	47.3	4.621	7
6	162.0	26250	–	–	–	–	–	12.58	0.4028	37.5	4.115	8
7	144.3	20820	–	–	–	–	–	15.87	0.5080	29.7	3.665	9
8	128.5	16510	7.6	–	7.1	73	46	20.01	0.6405	23.6	3.264	10
9	114.4	13090	8.6	–	7.8	–	–	25.23	0.8077	18.7	2.906	11
10	101.9	10380	9.6	9.1	8.9	55	33	31.82	1.018	14.8	2.588	12
11	90.7	8234	10.7	–	9.8	–	–	40.12	1.284	11.8	2.305	13
12	80.8	6530	12.0	11.3	10.9	41	23	50.59	1.619	9.33	2.053	14
13	72.0	5178	13.5	–	12.8	–	–	63.80	2.042	7.40	1.828	15
14	64.1	4107	15.0	14.0	13.8	32	17	80.44	2.575	5.87	1.628	16
15	57.1	3257	16.8	–	14.7	–	–	101.4	3.247	4.65	1.450	17
16	50.8	2583	18.9	17.3	16.4	22	13	127.9	4.094	3.69	1.291	18
17	45.3	2048	21.2	–	18.1	–	–	161.3	5.163	2.93	1.150	18
18	40.3	1624	23.6	21.2	19.8	16	10	203.4	6.510	2.32	1.024	19
19	35.9	1288	26.4	–	21.8	–	–	256.5	8.210	1.84	0.912	20
20	32.0	1022	29.4	25.8	23.8	11	7.5	323.4	10.35	1.46	0.812	21
21	28.5	810	33.1	–	26.0	–	–	407.8	13.05	1.16	0.723	22
22	25.3	642	37.0	31.3	30.0	–	5	514.2	16.46	0.918	0.644	23
23	22.6	510	41.3	–	37.6	–	–	648.4	20.76	0.728	0.573	24
24	20.1	404	46.3	37.6	35.6	–	–	817.7	26.17	0.577	0.511	25
25	17.9	320	51.7	–	38.6	–	–	1031	33.00	0.458	0.455	26
26	15.9	254	58.0	46.1	41.8	–	–	1300	41.62	0.363	0.405	27
27	14.2	202	64.9	–	45.0	–	–	1639	52.48	0.288	0.361	29
28	12.6	160	72.7	54.6	48.5	–	–	2067	66.17	0.228	0.321	30
29	11.3	127	81.6	–	51.8	–	–	2607	83.44	0.181	0.286	31
30	10.0	101	90.5	64.1	55.5	–	–	3287	105.2	0.144	0.255	33
31	8.9	80	101	–	59.2	–	–	4145	132.7	0.114	0.227	34
32	8.0	63	113	74.1	62.6	–	–	5227	167.3	.090	0.202	36
33	7.1	50	127	–	66.3	–	–	6591	211.0	.072	0.180	37
34	6.3	40	143	86.2	70.0	–	–	8310	266.0	.057	0.160	38
35	5.6	32	158	–	73.5	–	–	10480	335	.045	0.143	38-39
36	5.0	25	175	103.1	77.0	–	–	13210	423	.036	0.127	39-40
37	4.5	20	198	–	80.3	–	–	16660	533	.028	0.113	41
38	4.0	16	224	116.3	83.6	–	–	21010	673	.022	0.101	42
39	3.5	12	248	–	86.6	–	–	26500	848	.018	.090	43
40	3.1	10	282	131.6	89.7	–	–	33410	1070	.014	.080	44

[1] A mil is .001 inch. [2] Figures given are approximate only; insulation thickness varies with manufacturer. [3] Max. wire temp. of 212° F and max. ambient temp. of 135° F. [4] 700 circular mils per ampere is a satisfactory design figure for small transformers, but values from 500 to 1000 cm are commonly used.

damaged. For a particular core the maximum ac excitation can be determined by

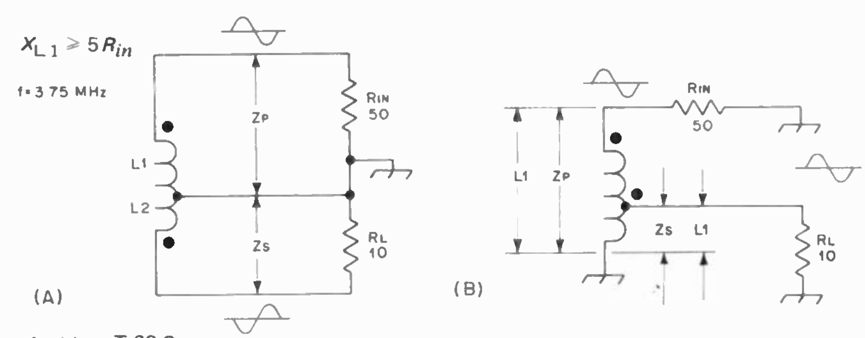
$$B_{max(ac)} = \frac{E_{rms} \times 10^8}{4.44 f N_p A_e} \text{ (gauss)} \quad \text{(Eq. 5)}$$

where A_e = equivalent area of magnetic path in cm^2, E_{rms} = applied voltage, N_p = number of core turns, f = frequency in Hz, and B_{max} = maximum flux density (gauss).

The foregoing is applicable to inductors which do not have dc flowing in the winding along with ac. Where both ac and dc currents flow

$$B_{max(total)} = \frac{E_{rms} \times 10^8}{4.44 f N_p A_e} + \frac{N_p I_{dc} A_L}{10 A_e} \quad \text{(Eq. 6)}$$

where I_{dc} = dc current through winding, and A_L = the manufacturer's inductance index for the core being used.



Amidon T-68-2

A_L = 57 µH

1) $Z_{ratio} = \frac{Z_p}{Z_s} = \frac{50}{10} = 5:1$

2) $n = \sqrt{Z_{ratio}} = \sqrt{5} = 2.23:1$

3) $L1 = \frac{5R_{in}}{2\pi f} = \frac{5 \times 50}{6.28 \times 3.75} = \frac{250}{23.55} = 10.6\ \mu H$ (min.)

4) $N_{L1} \doteq 100 \sqrt{\frac{L1}{A_L}} = 100 \sqrt{\frac{10.6}{57}} = 100 \times 0.431 = 43$ turns

5) $N_{L2} = \frac{N_{L1}}{n} = \frac{43}{2.23} = 19.28$ turns

Fig. 1

Table 2

Filament Transformers

TYPE	SECONDARY VOLTS	AMPS	SIZE	WT. LBS.
P-4026	2.5	1.5	2-7/8 X 1-1/2 X 1-5/8"	3/4
P-6133	2.5 CT	5.0	2-11/16 X 3-3/16 X 2-1/4"	1-1/2
P-3024	2.5 CT	10.0	3-3/16 X 2-5/8 X 2-5/8"	2-1/2
P-3060	2.5 CT	10.0	3-1/2 X 2-7/8 X 2-1/2"	2-1/2
P-6454	2.5 CT	10.0	3-5/8 X 2-1/2 X 3-1/8"	2-1/2
P-6467	5.0 CT	3.0	3-3/16 X 2 X 2"	1-1/2
P-3026	5.0 CT	3.0	2-5/8 X 2-5/8 X 2-5/8"	2-11/16
P-6455	5.0 CT	6.0	3-3/16 X 2-1/4 X 2-5/8"	2
P-6135	5.0 CT	10	3-1/8 X 2-1/2 X 2-7/8"	3
P-6468	5.0 CT	30.0	3-9/16 X 3-7/8 X 4-3/16"	4-1/3
P-6492	5.0 CT	30.0	4-3/4 X 3-3/4 X 3-7/8"	7-1/2
P-6465	6.3 CT	0.6	1-3/8 X 2-3/8 X 1-3/8"	0.4
P-8190	6.3	1.2	2 X 3-1/4 X 1-3/4"	1
P-8191	6.3	1.2	2 X 3-1/4 X 1-3/4"	1
P-6134	6.3 CT	1.2	1-5/8 X 2-3/4 X 1-3/4"	1
P-6462	6.3	3.0	3-5/8 X 2-5/8 X 3"	2
P-5014	6.3 CT	3.0	3-1/8 X 2-1/2 X 2-1/2"	2
P-6466	6.3 CT	3.0	2 X 3-5/16 X 2"	1-1/2
P-4019	6.3 CT	4.0	2-5/8 X 3-3/16 X 2-5/8"	2-3/4
P-3064	6.3 CT	6.0	3-1/8 X 2-1/2 X 2-7/8"	2-1/2
P-6456	6.3 CT	6.0	2-1/4 X 3-5/8 X 2-1/2"	2
P-6308	6.3 CT	10.0	3-1/2 X 2-7/8 X 2-7/8"	3-1/2
P-6309	6.3 CT	20.0	4-5/8 X 3-3/4 X 3"	7
P-5015	7.5 CT	4.0	3-1/8 X 2-1/2 X 2-7/8"	3
P-6138	7.5 CT	8.0	3-1/8 X 2-7/8 X 3-7/8"	4.7
P-8380	10.0 CT	3.0	2-1/4 X 3-3/4 X 2-3/8"	1.6
P-5016	10.0 CT	4.0	3-1/2 X 2-7/8 X 2-5/8"	3-1/2
P-6458	10.0 CT	5.0	3-1/16 X 2-1/2 X 2-7/8"	3
P-6139	10.0 CT	8.0	3-7/8 X 3-1/8 X 3-1/8"	5
P-6461	10.0 CT	10.0	3-7/8 X 3-1/4 X 3-5/8"	5
P-8130	12.6 CT	2.0	2 X 3-1/4 X 2"	1-1/2
P-8180	25.2 CT	1.0	2-1/8 X 3-1/4 X 2"	1.4
P-8358	12.6 CT	3.0	3-5/8 X 2-1/8 X 2-1/4"	2
P-6469	25.2	1.0	2 X 3-1/4 X 2"	1-1/2
P-8357	25.2 CT	2.0	4 X 2 X 2-5/8"	2-1/4
P-8388	25.2 CT	2.8	2-1/4 X 4 X 2-5/8"	2.2
P-8389	6.3	1.0	1-1/2 X 2-7/8 X 2-3/8"	0.6
P-8609	26.8 CT	1.0	2-1/8 X 3-1/4 X 2-13/16"	1.3
P-8606	26 V CT	–	1-3/8 X 2-1/8 X 1-3/4"	0.25
P-8607	26 V CT	0.25	1-3/4 X 2-7/8 X 2-3/8"	0.70
P-8605	48 V CT	1.1	2-3/8 X 4 X 3-9/16"	2.3

Chicago-Stancor Co.

It should be noted that the various core materials are rated for an optimum range of operating frequencies. The Q of the toroidal inductor is determined in part by the kind of core substance used. The tables in chapter 4 contain this kind of data.

Autotransformers

The most basic kind of transformer is the *autotransformer*. Fig. 1 gives two versions of the hookup. At A, a reversal in phase occurs between input and output. At B, the phase is the same at both transformer ports. A design example is given in Fig. 1 to show how to calculate the number of turns for each winding in a hypothetical example, where 50 ohms must be matched to 10 ohms. Other impedance levels can be matched similarly. In all broadband transformers, it is necessary to assure ample reactance in the winding of the highest impedance to prevent losses and an excessive SWR. As a general rule the reactance of high-impedance winding should be four or five times the intended impedance of the winding. This rule holds for the transformers (broadband) described elsewhere in this chapter. The arrangement at A of Fig. 1 may be preferred if a phase reversal presents no problems. The currents in the two windings are of opposite phase, thereby reducing the current which flows in the common lead. In the circuit at B the current of *both* windings flows through the lower, or common, leg.

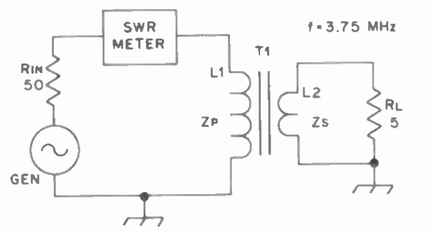
Conventional Transformers

Transformers with separate windings on a common core (not bifilar, or multifilar) are called *conventional transformers* in the present vernacular. Fig. 2 provides circuit and design data for such a transformer. Design data in the example are based on the use of an Amidon T-68-2 powdered-iron core, as was the case with the circuit in Fig. 1. It should be noted that although the term A_L is used in the equation, Amidon uses L_{100} in place of A_L. An SWR meter can be inserted in series with the primary, L1, to permit experimental adjustment of the turns ratio once the loads are connected. This method can be used to compensate for unknown reactances which may be present in the circuit with which the transformer will be used (input to transistor amplifier, etc.).

The inductance amount can be determined for a given number of turns of wire if the A_L is not known. The basic equation for determining maximum inductance (L_m) is

$$L_m = 0.4\pi N_p^2 \mu_e \frac{A_e\,(\text{cm}^2)}{l_e\,(\text{cm})} \times 10^{-8} \text{ henries} \qquad \text{(Eq. 7)}$$

Fig. 2



1) $Z_{ratio}\ (\text{T1}) = \dfrac{Z_p}{Z_s} = \dfrac{50}{5} = 10$

$\therefore n = \sqrt{Z_{ratio}} = \sqrt{10} = 3.16{:}1$

2) $\text{L1} = \dfrac{5R_{in}}{2\pi f} = \dfrac{5 \times 50}{6.28 \times 3.75}$

$= \dfrac{250}{23.55} \approx 10.6\ \mu\text{H (min.)}$

3) $N_{\text{L1}} = 100\sqrt{\dfrac{\text{L1}}{A_L}} = 100\sqrt{\dfrac{10.6}{57}}$

$= 100 \times 0.431 = 43$ turns

4) $N_{\text{L2}} = \dfrac{N_{\text{L1}}}{n} = \dfrac{43}{3.16} = 13.6$ turns

A_L = mfg. constant
N = no. of turns
n = turns ratio
$X_{\text{L1}} \geqslant 5R_{in}$
core = Amidon T-68-2 toroid (A_L = 57), from Table 13.

where N_p = number of primary turns, μ_e = effective core permeability, A_e = effective cross-sectional area of core, and l_e = length of flux path in core.

The equation can be shortened considerably by using the manufacturer's A_L expression. Inductance in μH can be found by

$$L_m = N_p{}^2 A_L \times 10^{-4}\ \mu\text{H}$$

$$= 31^2 \times 57 \times .0001 = 5.47\ \mu\text{H} \qquad \text{(Eq. 8)}$$

where an Amidon T-68-2 core (A_L = 57) is used with a 31-turn winding. The fundamental equation can be used in situations where A_L is not listed on the data sheet. Table 6 provides data on inductance values for three popular sizes of Amidon powdered-iron cores for use at mf and hf. Two core mixes are specified, types 2 and 6.

Another kind of conventional transformer is shown in Fig. 3. The technique illustrated is one used by TRW Electronic Components, Lawndale, CA 90260. A definitive treatment of the design is contained in *TRW Application Note* CT-113-71. The principle is similar to that shown in Fig. 2, except that a U-shaped brass tube serves as the low-impedance center-tapped winding. Two stacks of high-μ small-diameter toroids are slipped over the tubes, side by side as shown at B, then the tubes are joined at one end by means of a pc-board plate. The primary winding, L1, is looped through both brass tubes. L2 in Fig. 3 is equivalent to a 1-turn, center-tapped winding. The turns ratio is effected by using the necessary number of turns at L1. A permeability of 950 (ferrite) is suitable for this design. The high μ is needed to assure ample reactance in each winding for the lowest operating frequency. The transformer of Fig. 3 is suitable for use from 1.8 to 30 MHz.

Narrow-Band Transformers

Magnetic core material can be used in narrow-band (tuned) rf transformers. The core material, powdered-iron or ferrite, is chosen for the frequency of use, and must be consistent with the circuit Q needed. Narrow-band transformers are preferred in circuits which must discriminate as much as possible against unwanted harmonic currents (bandpass and low-pass networks in particular).

Fig. 4 shows a narrow-band transformer. The design parameters are practically the same as for the networks discussed in chapter 4. An advantage in using toroidal narrow-band transformers is that the toroidal transformer exhibits self-shielding properties, thereby reducing the need for external shielding devices.

Table 3

Transistor Interstage Transformers

MFR'S. TYPE	*IMP. IN OHMS PRI.*	*SEC.*	*MAX. PRI. mA DC*	*POWER WATTS*	*MFG. CTRS.*
TA-2	100 CT	10 CT	100	0.25	1-13/16
TA-3	100	1000 CT	100	0.25	2
TA-4	500 CT	5000 CT	12	.03	2
TA-19	100 CT	10 CT	–	0.15	25/32
TA-24	500 CT	50,000	–	0.15	25/32
TA-28	1500	500 CT	–	0.15	25/32
TA-30	5000 CT	7500 CT	–	0.15	25/32
TA-31	5000 CT	10,000 CT	–	0.15	25/32
TA-32	5000 CT	80,000 CT	–	0.15	25/32
TA-34	10,000	200 CT	–	0.15	25/32
TA-35	10,000	2000 CT	–	0.15	25/32
TA-36	10,000	3000 CT	–	0.15	25/32
TA-38	500 CT	150 CT	–	0.15	25/32
TA-46	100,000	1500 CT	–	0.30	1-3/8

Transistor Driver Transformers

MFR'S. TYPE	*IMP. IN OHMS PRI.*	*SEC.*	*MAX. PRI. mA DC*	*POWER WATTS*	*MFG. CTRS.*
TA-7	100	100 CT	100	0.5	2-3/8
TA-16	20	36 CT	400	1.0	1-3/4
TA-59	500 CT	200 CT	50	0.5	1-3/4
TA-61	–	–	–	–	2

Transistor Output Transformers

MFR'S. TYPE	*IMP. IN OHMS PRI.*	*SEC.*	*MAX. PRI. mA DC*	*POWER WATTS*	*MFG. CTRS.*
TA-10	2000 CT	16/8/4	–	0.2	1-3/8
TA-12	20 CT	8	500	10	2
TA-21	500 CT	4/8/16	–	0.15	25/32
TA-26	1250	4/12	–	0.15	25/32
TA-33	10,000 CT	4/8/16	–	0.15	25/32
TA-39	100 CT	4/8/16	–	0.30	1-3/8
TA-42	500 CT	4/8/16	–	0.30	1-3/8
TA-43	700 CT	4/8/16	–	0.30	1-3/8
TA-62	25	4	400	4	2-3/8

Chicago-Stancor Co.

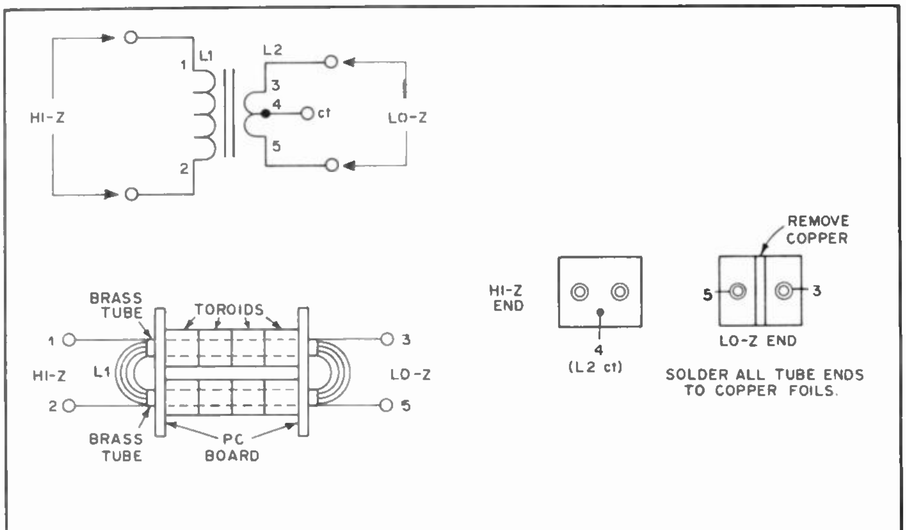
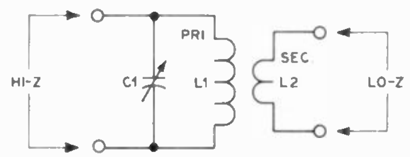


Fig. 3
Toroid cores have μ 950 for 1.8-30 MHz.
L1 – Wind turns through brass tubing.
L2 – Brass tubing and Hi-Z end pc board.

Unbal. to bal. broadband step-down tran. $X_{L1} \geqslant 5 \times R$ of source at lowest f_o (for 50 ohms, $X_{L1} \geqslant 250$ ohms). $X_{L2} \geqslant 5 \times$ LO-Z load at lowest f_o.

Fig. 4



NARROW-BAND RF TRANSFORMER

$X_{L1} = X_{C1}$ at f_o.

$$n = \sqrt{\frac{Z_p}{Z_s}}$$

n = turns ratio

Capacitive-Divider Transformer

A narrow-band transformer can be built with a single winding by using a capacitive divider to match a low-impedance source to a high-impedance load, or vice-versa. Fig. 5 illustrates the circuit and provides equations for solving a design problem in which this configuration is desired. This kind of circuit is used frequently for matching one transistor stage to another. It has a band pass characteristic and is, therefore, useful in minimizing harmonic energy.

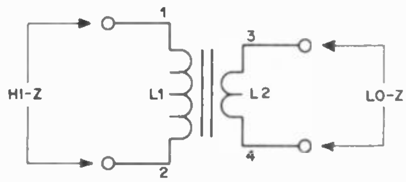
A conventional broadband transformer is illustrated in Fig. 6 to show how the primary and secondary windings are placed on a toroid core. A layer of insulating material is sometimes used between the core and the first winding, and again between the primary and secondary windings to prevent arcing or short circuits.

Broadband Transmission-Line Transformers

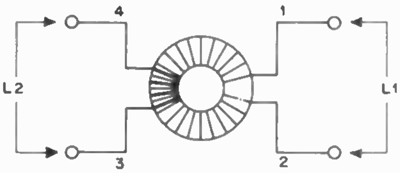
Transmission-line transformers are used more often in solid-state work than are the *conventional* types. The former are somewhat more efficient, and thus their popularity. Detailed data on the theory of transmission-line transformers can be found in the references at the end of this chapter.

By way of brief explanation, the windings are bifilar, multifilar, or made from sections of coaxial cable. The windings are sometimes formed by twisting the wires a given number of turns per inch to effect a specific line impedance. At other times the wires of a winding are simply laid side by side on the core, spaced apart to effect a particular impedance characteristic. When the impedances of the transmission lines are not precisely correct for a given transformer application, it may become necessary to add compensating capacitors to the circuit for good per-

Fig. 6



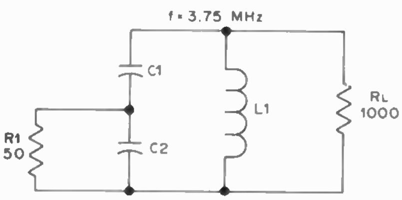
CONVENTIONAL BROADBAND TRANSFORMER
(A)



L1 – TO OCCUPY ENTIRE CORE
L2 – WOUND OVER L1 WINDING
(B)

$X_{L1}, X_{L2} \geqslant 5 \times$ respective design R at lowest f_o.

Fig. 5



$$R1 < R_L, \text{ and } Q_L > \sqrt{\frac{R_L}{R1} - 1}$$

1) Select a bandwidth (e.g., 0.5 MHz)

2) $Q_L = \frac{f}{f_2 - f_1} = \frac{3.75}{4.0 - 3.5} = \frac{3.75}{0.5} = 7.5$

3) $C_T = \frac{Q_L}{2\pi f\left[\frac{R_L}{2}\right]} \times 10^6 = \frac{7.5}{6.28 \times 3.75\left[\frac{1000}{2}\right]} \times 10^6$

$\frac{7.5}{11,775} \times 10^6 = .000636 \times 10^6 = 636$ pF

4) $\frac{C2}{C1} \cong \sqrt{\frac{R_L}{R1} - 1} = \sqrt{\frac{1000}{50} - 1} = 3.472$

$\therefore$ C1 = x and C2 = 3.472 x and CT = $\frac{C1 \times C2}{C1 + C2}$

$\therefore 636 = \frac{x\,(3.472\,x)}{1\,x + 3.472\,x} = \frac{3.472}{4.472}\left[\frac{x^2}{x}\right] = 0.776\,(x)$

$\therefore 636 = 0.776\,x$ and $x = \frac{636}{0.776} = 820$

$\therefore$ C1 = x = 820 pF

5) C2 = C1 × 3.472 = 820 × 3.472 = 2847 pF

6) $X_{L1} = X_{CT} \qquad X_{CT} = \frac{1}{2\pi f\,(C\,10^{-6})}$

$= \frac{1}{6.28 \times 3.75 \times (636 \times 10^{-6})} = \frac{1}{23.55 \times .000636}$

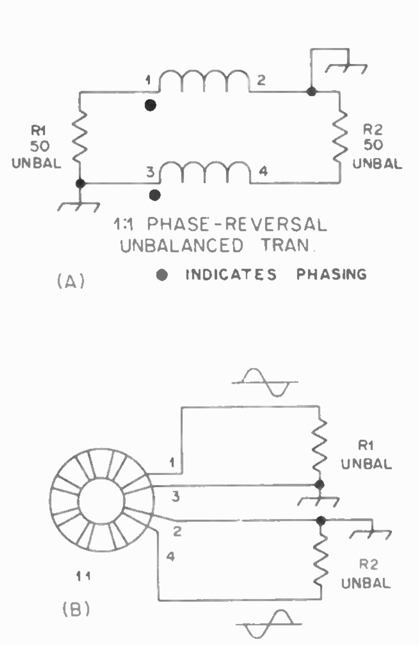
$= \frac{1}{.01497} = 66.76$ ohms $\therefore X_{L1} = 66.76$ ohms

$\therefore L1 = \frac{X_{L1}}{2\pi f} = \frac{66.76}{6.28 \times 3.75} = 2.83\ \mu H$

Proof: $f = \frac{10^6}{2\pi\sqrt{LC}} = \frac{10^6}{6.28\sqrt{2.83 \times 636}}$

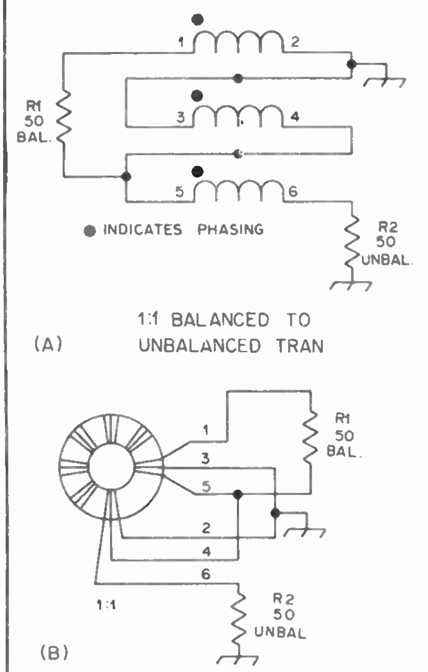
$= \frac{10^6}{266.4} = 3753.3$ kHz

Fig. 7



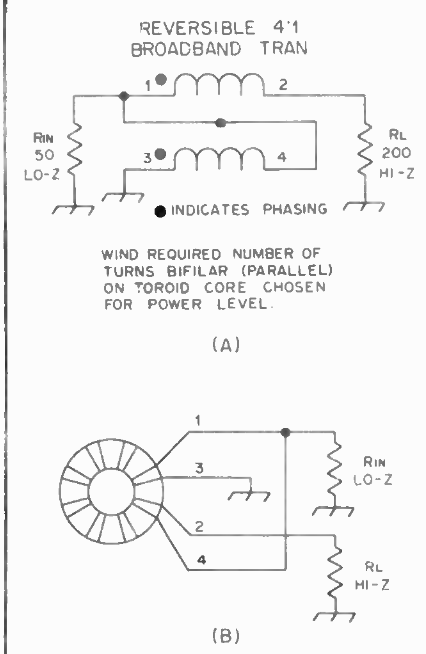
X_L (windings) $\geqslant$ 5 × R1 at lowest f_o (X_L = 250 this example). Bifilar winding.

Fig. 8



X_L (windings) $\geqslant$ 5 × R1 at lowest f_o ($X_L \geqslant$ 250 this example). Trifilar winding.

Fig. 9



X_L (windings) $\geqslant$ 5 × Hi-Z term. At lowest f_o ($X_L \geqslant$ 1000 ohms in this example).

formance at the high end of the frequency range for which the transformer was built.

Phase-Reversal Transformer

In circuits where the phase must be reversed on a 1:1 relationship, and where each side of the circuit is single-ended, the transformer of Fig. 7 is suitable. The reactance of the windings would be four or five times greater than the value of R1.

Balanced-to-Unbalanced 1:1 Transformer

When it is necessary to transform from an unbalanced to balanced condition at one impedance value, the circuit of Fig. 8 can be employed. This kind of transformer is useful for feeding a split dipole antenna from a coaxial transmission line.

Unbalanced 4:1 Transformer

An impedance transformation ratio of 4:1 can be secured by using the circuit of Fig. 9. Since the terminals are of the unbalanced variety, this transformer can be used effectively between single-ended stages of solid-state transmitters where 4:1 transformations are applicable. A transformer of this type is often used in matching coaxial feed lines to low-impedance vertical antennas.

Balanced-to-Unbalanced 4:1 Transformer

One of the better-known baluns is the balanced-to-unbalanced 4:1 transformer. Its popularity started in the day of the Windom antenna, for feeding a 75-ohm transmitter from a 300-ohm balanced antenna feed. The circuit is shown in Fig. 10.

Unbalanced-to-Balanced 4:1 Transformer

The inverse of the transformer illustrated in Fig. 10 is the unbalanced-to-balanced 4:1 transformer. It is shown in Fig. 11. This one finds frequent application at very low impedance levels, such as between a 50-ohm source (unbalanced) and a balanced lower-impedance load (bases of push-pull transistors).

Table 4

Power Transformers

MFR'S. TYPE	*PLATE VOLTS*	*CT mA*	*FIL. 1 V.*	*FIL. 1 A.*	*FIL. 2 V. CT*	*A*	*MFG. CTRS.*	*WT. LBS.*
PC-8403	500	70	5.0	2.0	6.3 CT	2.5	2 X 2-1/16	3-1/4
PC-8404	520	90	5.0	2.0	6.3 CT	3.0	2-1/4 X 2-1/4	4
PC-8405	540	120	5.0	3.0	6.3 CT	3.5	2-1/2 X 2-3/16	5
PC-8410	720	120	5.0	3.0	6.3 CT	3.5	2-1/2 X 2-7/16	5-1/2
PC-8411	750	150	5.0	3.0	6.3 CT	4.5	2-3/4 X 2-13/16	5-3/4
PC-8412	800	200	5.0	3.0	6.3 CT	5.0	3 X 2-13/16	8-1/4
PC-8413	800	250	5.0	4.0	6.3 CT	5.0	3 X 3-5/16	10
PC-8414	1200	200	5.0	3.0	6.3 CT	3.0	3 X 3-1/16	8-1/4
PC-8422	650	150	5.0	3.0	6.3 CT	5.0	2-1/2 X 2-9/16	5.8
PS-8415	125*	15	–	–	6.3	0.6	2	3/4
PS-8416	250	25	–	–	6.3	1.0	2-3/8	1
PA-8421	125*	50	–	–	6.3	2.0	3-1/8	1-1/2
PC-8417	440	50	–	–	25.2	0.5	2 X 1-9/16	2-1/4
PC-8418	460	50	–	–	6.3	2.5	2 X 1-9/16	2-1/4
PM-8418	460	50	–	–	6.3	2.5	2 X 2-1/2	2-1/4
PC-8419	480	70	–	–	6.3	3.0	2 X 1-13/16	2-1/2
PM-8419	480	70	–	–	6.3	3.0	2 X 2-1/2	2-1/2
PC-8420	520	90	–	–	6.3	4.0	2 X 2-1/4	3-1/2
PM-8420	520	90	–	–	6.3	4.0	2-1/4 X 2-13/16	3-1/2

*Half-wave type
Chicago-Stancor Co.

Table 5

Nomenclature and Symbology

SYMBOL	UNITS	TERM	DESCRIPTION
A_c	in², cm²	Available winding area of core	Cross-sectional area (perpendicular to direction of wire), available for winding turns on a particular core.
A_{CB}	in², cm²	Available winding area of bobbin	Cross-sectional area (perpendicular to direction of wire), available for winding turns on a particular bobbin.
A_e	cm²	Effective area of core	The cross-sectional area that an equivalent gapless core (of uniform magnetic and geometric properties) would have.
A_g	cm²	Effective gap area	The equivalent area through which the (assumed uniform) flux in a magnetic core gap passes. Corrects for fringing effects.
A_L	mH/1,000 turns	Inductance index	Relates inductance to turns for a particular core and gap.
A_m	cm²	Effective area of magnetic path	In a gapped structure, the equivalent cross-sectional area of the magnetic part of the path, assumed uniform.
A_p	in², cm²	Available winding space for primary	That portion of the total available winding area allotted to the transformer primary winding.
A_x	in², cm²	Cross-sectional area of wire	The cross-sectional area of the conductor part of the wire.
B	gauss	Magnetic flux Density	The flux density (lines/cm²) in a magnetic circuit, measured at a given point.
B_{max}	gauss	Maximum flux density	The value of flux density corresponding to the peak of the applied excitation.
B_O	gauss	Magnetic flux density	Maximum values of flux density for $\mu = \mu e$ in transformer core.
B_t	gauss	Pulse-excited flux density	The value of flux density corresponding to the instant of termination of an applied rectangular pulse.
BW	hertz	Bandwidth	Frequency range over which response of a tuned transformer is considered to be uniform ($BW = 2\Delta f = \frac{1}{Q}$).
C_t	farads	Equivalent shunt capacitance	The equivalent shunt capacitance represented by the interwinding capacitance of a transformer.
E_m	Volts, rms	Magnetizing voltage	Effective value of alternating voltage applied across primary magnetizing inductance (see L_m).
E_p	Volts, rms	Transformer primary terminal voltage	Effective value of alternating voltage applied across primary terminals of a transformer.
E_{pmax}	Volts, peak	Peak primary	Peak value of E_p (for sine waves, $E_{pmax} = E_p\sqrt{2}$).
E_s	Volts, rms	Transformer secondary terminal voltage	Effective value of alternating voltage applied across secondary (output) terminals of a transformer.
f	hertz	Frequency	Alternations per second of voltage or current
f_1	hertz	Low-frequency cut-off	Lowest frequency at which the sinusoidal amplitude vs frequency characteristic of a transformer is within 3 dB of the mid-frequency level.
f_2	hertz	High-frequency cut-off	Highest frequency at which the sinusoidal amplitude vs frequency characteristic of a transformer is within 3 dB of the mid-frequency level.
H	oersteds	Magnetization	The magnetizing force that produces the magnetic flux in a transformer or inductor core. Ampere-turns per cm.
H_{max}	oersteds	Peak magnetization	The peak value of the magnetizing force.
H_o	oersteds	Dc Magnetization	The dc magnetizing force applied to a core (unidirectional current flowing in its windings).
I_c	Amperes rms	Core loss current	Component of magnetizing current accounting for power lost in core material.
I_{DC}	Amperes	Direct current	The magnitude of the unidirectional component of a current.
I_m	Amperes, rms	Magnetizing current	Component of primary alternating current devoted to magnetization of transformer core.
I_p	Amperes, peak	Transformer primary current	Total alternating current flowing in transformer primary winding. Vector sum of magnetizing (I_m) and load ($I_s \div n$) currents
I_{pmax}	Amperes, peak	Peak input current	The peak value of the transformer input current.
I_s	Amperes, rms	Transformer secondary current	Total alternating current flowing in transformer secondary winding.
K	Numeric	Dielectric constant of a substance	The ratio of the capacitance of a capacitor using that substance as a dielectric to the capacitance of the same structure with a vacuum as the dielectric.
k_1	(gauss)²	μ vs. B factor	Factor relating μ and B for a given power inductor design requirement.

SYMBOL	UNITS	TERM
k_2	Volt-amperes per Hz X 2π	Core-selection factor
L_m	henries	Self-inductance
L_O	henries	Air inductance
L_t	henries	leakage inductance
l_e	in., cm.	Length of flux path
l_g	in., cm.	Length of air gap
l_m	in., cm.	Length of magnetic path
M	ratio	Load/magnetizing current ratio
N_p	numeric	Primary turns
N_s	numeric	Secondary turns
n	numeric	Turns ratio
P_d	percent	Droop
P_{load}	watts	Load power
P_o	watts	Power dissipation factor
P_t	watts	Total power dissipation
Q	numeric	Quality factor
R_C	Ohms	Equivalent core loss resistance
R_g	Ohms	Generator resistance
R_L	Ohms	Load resistance
R_p	Ohms	Resistance of primary winding
R_s	Ohms	Resistance of secondary winding
T_{max}	°C	Maximum operating temperature
T_{rise}	°C	Temperature rise
t_p	seconds	Pulse duration
t_r	seconds	Pulse rise time
V_e	in.³, cm.³	Effective core volume
X_t	Ohms	Leakage reactance
α	numeric	Attenuation constant
αt	numeric	Total attenuation
Δf	hertz	Half-bandwidth
μ	numeric	Permeability
μ_{av}	numeric	Average permeability
μ_e	numeric	Effective permeability

Courtesy of Ferroxcube Corp.

DESCRIPTION

Factor relating the Volt-Ampere rating of a core to the frequency.

Self-inductance of a transformer primary winding ($L_m = N \frac{2}{P} A_L \times 10$ '' henries)

L for $\mu = 1$ (air; Thus, $L = \mu_o L_o$.

Self-inductance that appears in series with the windings due to leakage flux that does not contribute to the transfer of energy in a transformer.

Effective length of a uniform flux path.

Effective length of a uniform air gap in a magnetic circuit.

Effective length of a uniform magnetic path.

The ratio of the load component ($I_s \div n$) to the magnetizing component (I_m) in the transformer primary current (I_p).

Total number of complete turns in transformer primary winding.

Total number of complete turns in transformer secondary winding.

Ratio of transformer primary/secondary turns. Equivalent to primary/secondary voltage ratio (neglecting resistance of windings).

Percent Droop in the nominally flat portion of the waveform of a pulse delivered by a transformer, compared with the input pulse waveform.

The power dissipated in the external load connected to the transformer secondary terminals.

The power dissipation level that will cause a temperature rise of 50°C above ambient in a given core.

The total power dissipated in an inductor (Winding loss + core loss). Neglects dielectric losses.

Ratio of reactance to equivalent series resistance in an impedance.

The value of resistance across which the magnetizing excitation voltage would dissipate a power equal to the core loss of a transformer or inductor.

Effective internal series resistance of the excitation source.

Equivalent resistance of external load connected between transformer secondary terminals.

Dc resistance of transformer primary.

Dc resistance of transformer secondary.

The sum of the maximum expected ambient temperature, and the winding temperature rise in the core material.

Rise in temperature of an inductor or transformer core due to internal power dissipation.

Duration of the longest nominally flat portion of a pulse waveform.

Longest allowable time for a pulse to rise from 10% to 90% of its final amplitude.

The effective volume of a magnetic core ($V_e = l_e \times A_e$) defined as the volume of a uniform magnetic path l_e long, having a cross-sectional area A_e.

The reactance of transformer leakage inductance L_l ($X_l = 2\pi f L_l$).

Mid-frequency reduction in signal amplitude from primary terminal voltage to voltage established across the load.

Mid-frequency reduction in signal amplitude from generator internal EMF to transformer output terminal voltage.

Difference between center frequency and low frequency cut-off point (f_1) on resonance curve of tuned transformer.

Ratio of magnetic flux density to the excitation producing it ($\mu = B/H$).

The slope of the straight line between 0, 0 and $B = B_{max}$ for a given core material.

($\mu_{av} = \frac{B_{max}}{H_{max}}$)

The effective permeability of the core, at low excitation (in the linear region) for a uniform magnetic path.

Unbalanced-to-Unbalanced 9:1 Transformer

Fig. 12 shows how to build a 9:1 unbalanced broadband transformer. This kind is most often used in feeding a single transistor stage from a low-impedance source. It should be noted that the transmission line pair of L1/L2 is twice the length of the L3/L4 winding.

Balanced-to-Balanced 9:1 Transformer

This transformer is useful when feeding the push-pull bases of a transistor amplifier pair from push-pull driver stage collectors. Fig. 13 illustrates how the 9:1 transformer is configured.

Balanced-to-Unbalanced 9:1 Transformer

This often-used transformation scheme is found in many solid-state transmitters. It is employed between a 50-ohm unbalanced source and push-pull bases of a transistor amplifier. Fig. 14 shows the details for such a transformer.

16:1 Impedance Transformer

Separate 4:1 transformers can be used in cascade to provide a 16:1 ratio. It is possible also to wind a 16:1 transformer on a single core (see example in Fig. 16). The hookup is given in Fig. 15. This configuration is useful in feeding a low-impedance antenna or transistor amplifier from a 50-ohm unbalanced source.

Variable-Ratio Transformer

The transformer highlighted in Fig. 16 is one developed by W2FMI for experimental work. It is useful for general applications of testing and impedance matching. It was first described in W2FMI's paper at IEEE Intercon-75, Session L.

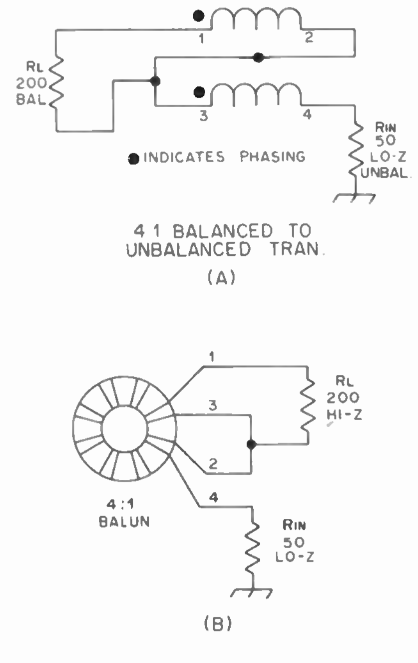
Single-Ended Hybrid Combiner

When it is desired to isolate one signal source from another while feeding the combination to a single load, the circuit of Fig. 17 can be used. When the input signals are on different frequencies, the power is split evenly between R3 and R4. When the input voltage is on the same frequency, and with equal amplitudes and phase, all of the power is dissipated in R4.

Broadband Mixer Transformer

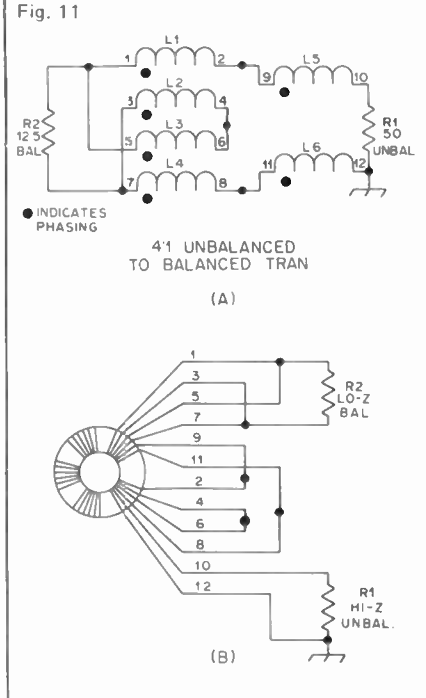
Trifilar broadband transformers are often used in diode-quad doubly balanced mixers or product detectors. Such a circuit is shown in Fig. 18. T1 and T2 are identical transformers. The windings should be placed on core material (ferrite in most applications) with a sufficiently high μ to assure a reactance four or five times the charac-

Fig. 10



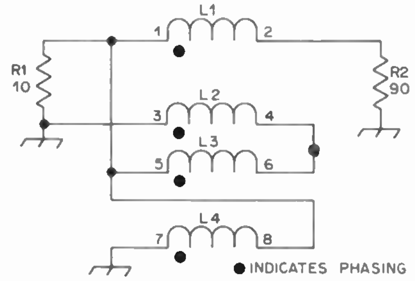
X_L (windings) $\geqslant$ 5 X Hi-Z term. At lowest f_o ($X_L \geqslant$ 1000 ohms this example). Bifilar wound turns.

Fig. 11

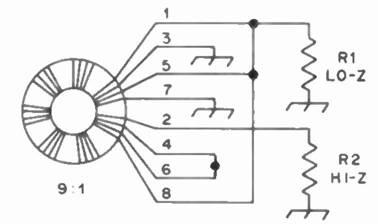


L1/L2, L3/L4, L5/L6 each are 50-ohm bifilar pairs or 50-ohm coax.

L1/L2 pair are twice the length of L3/L4.



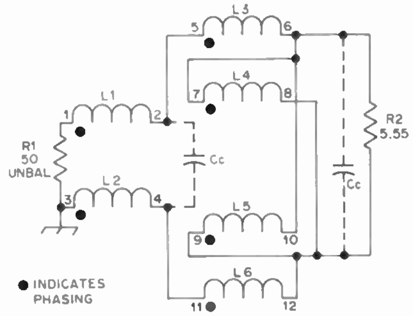
9:1 UNBALANCED TO UNBALANCED TRAN.
(A)



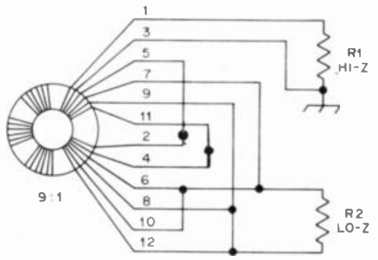
(B)

L1/L2 and L3/L4 lines must be 3 X R1 (30 ohms in this example).

Fig. 12



9:1 BALANCED TO UNBALANCED TRAN
(A)

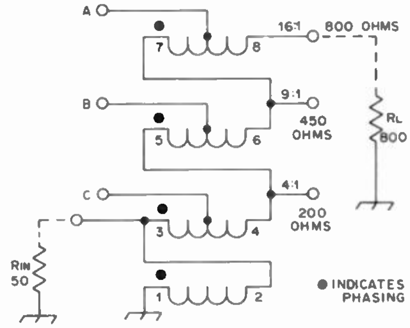


(B)

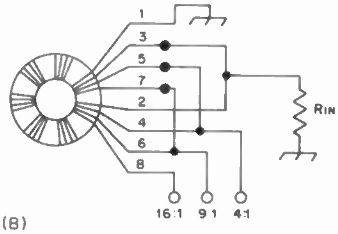
L1/L2 have 50-ohm bifilar winding. L3/L4/L5/L6 are quadfilar and should be 16.6 ohms for R2 given. If not, C_C can be added for reactance compensation.

Fig. 14

Taps A, B and C located to provide various ratios from 1.5:1 to 16:1.



REVERSIBLE VARIABLE-Z BROADBAND TRAN
(A)

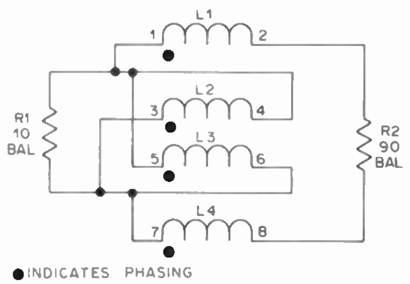


(B)

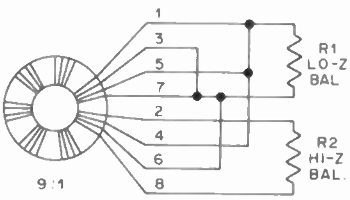
X_L (windings) $\geqslant$ 5 X Hi-Z term at lowest f_o ($X_L \geqslant$ 4000 ohms in this example). Wind required number of turns quadfilar (parallel) on toroid core chosen for power level.

Fig. 16

Fig. 13



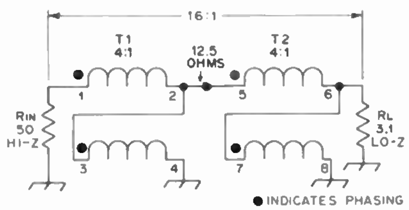
9:1 BALANCED TO BALANCED TRAN
(A)



(B)

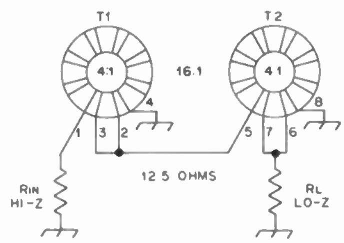
L1/L2 and L3/L4 pairs must have impedance of 3 X R1 (30 ohms in this example).

Fig. 15



CASCADED 4:1 BROADBAND TRANFORMERS GIVE 16:1 Z RATIO

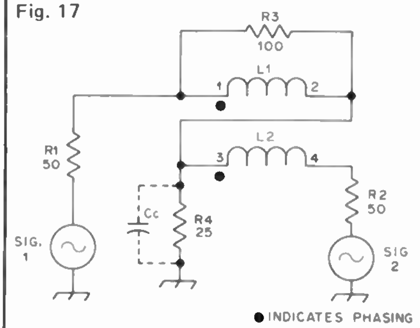
(A)



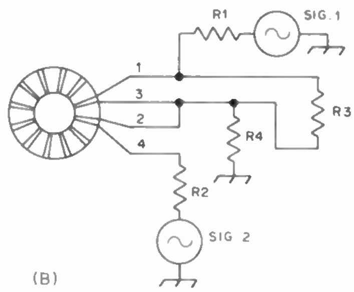
(B)

X_L (windings) $\geqslant$ 5 X Hi-Z term at lowest f_o ($X_L \geqslant$ 250 in this example).
Wind T1, T2 bifilar.

Fig. 17



SINGLE-ENDED HYBRID COMBINER
(A)



(B)

Inductance of windings:

$$L = \left[\frac{R1}{2\pi f_{(min)}}\right]\sqrt{S-1}$$

where S = power ratio.
L1/L2 bifilar pair.
Impedance of pair = R1, R2.
R1 – R2 isolation ≅ 40 dB.

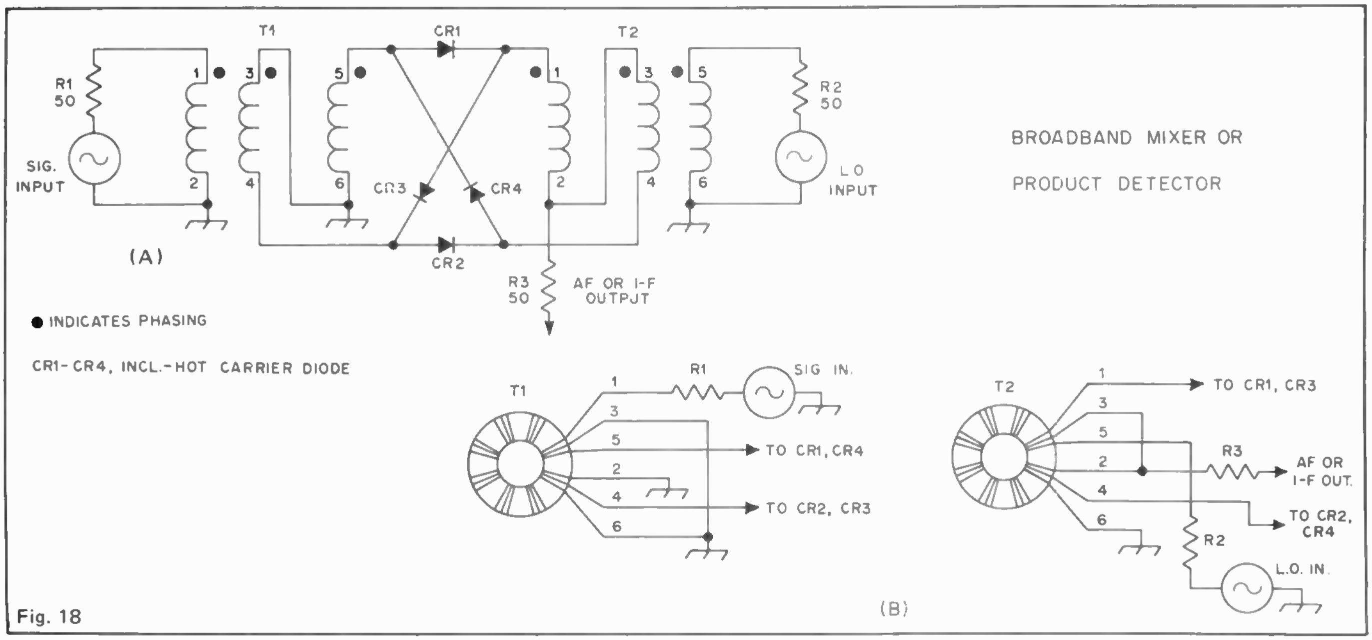


Fig. 18

Table 6

CORE TURNS NO.	µH T-50-2 (A_L = 50)	µH T-68-2 (A_L = 57)	µH T-80-2 (A_L = 55)	µH T-200-2 (A_L = 120)	µH T-50-6 (A_L = 40)	µH T-68-6 (A_L = 47)
5	0.125	0.142	0.137	0.3	0.1	0.117
10	0.5	0.57	0.55	1.2	0.4	0.47
15	1.125	1.28	1.23	2.7	0.9	1.05
20	2.0	2.28	2.20	4.8	1.6	1.88
25	3.12	3.56	3.43	7.5	2.5	2.93
30	4.5	5.13	4.95	10.8	3.6	4.23
35	6.1	6.98	6.73	14.7	4.9	5.75
40	8.0	9.12	8.8	19.2	6.4	7.52
45	10.1	11.54	11.1	24.3	8.1	9.51
50	12.5	14.25	13.75	30.0	10.0	11.75
55	15.1	17.24	16.6	36.3	12.1	14.2
60	18.0	20.52	19.8	43.2	14.4	16.9
70	24.5	27.93	26.9	58.8	19.6	23.0
80	32.0	36.48	35.2	76.8	25.6	30.0
90	40.5	46.17	44.5	97.2	32.4	38.0
100	50.0	57.0	55.0	120.0	40.0	47.0

teristic 50-ohm impedance of the circuit. An X_L of 250 is suitable for the lowest operating frequency at R1 or R2. In the example, R1, R2 and R3 are not physical resistances. Rather, they represent the characteristic impedance of the three mixer ports with signal applied. At signal and local-oscillator frequencies in the hf region, a suitable core material is ferrite with a permeability of 125. At mf and lf, ferrites with a μ of 950 or greater are often used.

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Chapter 6

Filter Design

Filters are used extensively by amateur and commercial designers. They are used to convert double-sideband energy to ssb energy in transmitters, to provide i-f selectivity in receivers, and to attenuate undesired frequencies in a host of applications. Additional uses are in eliminating RFI and TVI caused by amateur stations.

This chapter treats an assortment of filter types and illustrates various methods for applying filters to amateur and commercial equipment. An attempt has been made, where applicable, to simplify the design methods by restricting the parameters around which the design is carried out. In all instances the performance of the filters should be entirely satisfactory for amateur work. Those desiring to follow the classic design techniques used in the engineering world will find the references given in this chapter of considerable value.

Simplified Bandpass-Filter Design

The material in this section of the chapter is built in part upon some notes contributed by Hayward, W7ZOI. The procedure shows how to design by means of simple steps some band-pass filters which are applicable to amateur circuits.

In amateur work the more common approach to filter design has been to use the classic image-parameter filter by synthesizing a low-pass filter for a particular terminating impedance, then transforming the result to a band-pass structure. The method is entirely suitable for wide-bandwidth design work, but when a narrow-bandwidth filter is needed the component values resulting from that design approach often become impractical, and the losses in the filter elements can distort the desired response.

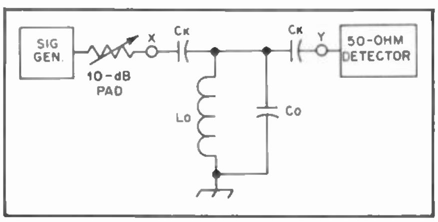


Fig. 1 — Test setup for determining unloaded Q of a resonator.

Modern band-pass filters are designed by coupling a number of resonant circuits (resonators) together with minimal restrictions as far as impedance levels at the ends of the filter are concerned. Such filters are predistorted — the loading and coupling amounts are adjusted to take into account the unloaded Q (Q_u) of the resonators. A complete design of this type of filter is extremely comprehensive in scope, but by accepting some restrictions it is possible to design two- and three-pole Butterworth filters that will provide good performance. In chapter 4 it was shown how to measure the Q_u of an arbitrary quantity of inductance across which the necessary amount of C has been placed to establish resonance ($2\pi f = 1 \div \sqrt{LC}$).

Procedure

In using the test setup for determining Q_u (Fig. 1), select a value of C_K which is vastly smaller than that of C_o (e.g., C_o = 200 pF and C_K = 1 pF). The power to the detector will be 30 or 40 dB lower than that available at point X of Fig. 1. The generator is swept to find the center frequency, then to f_1 and f_2 — the points at which the detected voltage or current is down 0.707 from f_o. A better approach would be to use a 3-dB pad at circuit point Y. Then, Q_u will equal $f_o \div f_2 - f_1 = f_o \div BW_u$ (unloaded bandwidth). The principle applies when measuring any resonator at any frequency, and results can be as accurate as those obtained by means of a laboratory type of Q meter.

Normalized Q

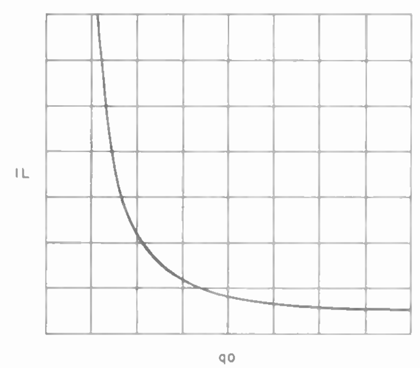
As an example, we will design a filter which contains several resonators. The f_o is 10 MHz and the Q_u is 100. It shall be assumed that the filter bandwidth is 500 kHz (0.5 MHz). Therefore, the loaded Q of the filter is

$$Q_F = \frac{f_o}{BW} = \frac{10}{0.5} = 20 \qquad \text{(Eq. 1)}$$

The normalized Q (q_o) is defined as

$$q_o = \frac{Q_u}{Q_F} = \frac{100}{20} = 5 \qquad \text{(Eq. 2)}$$

With the foregoing information the designer can proceed to the next step.



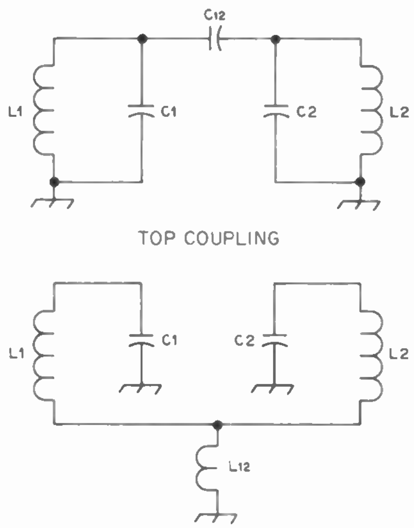
q_o	I_L(dB)
14.1	0.915
7.07	1.94
4.71	3.09
3.54	4.40
2.83	6.02
2.36	7.96
2.02	10.46

Fig. 2 — Nomograph of insertion loss versus q_o.

The Double-Tuned Circuit

The starting point is with two identical tuned circuits, the elements of which are L_o, C_o, and $R_{p(u)}$, where $R_{p(u)} = Q_u (2\pi f_o \times L_o)$ and $2\pi f_o = 1 \div \sqrt{L_o C_o}$. Calculate the unloaded bandwidth (BW_u) as $BW_u = f_o \div Q_u$.

Next, design the filter for a loaded



$$C_{12} = C_o \times \frac{BW_L}{f_o} \times K_{12}$$

For bottom-coupled circuits using inductive coupling:

$$L_{12} = L_o \times \frac{BW_L}{f_o} \times K_{12}$$

Fig. 3 — Examples of top and bottom coupling for a two-resonator circuit.

bandwidth (BW_L) which is greater than BW_u. Once a filter bandwidth (BW_F) is chosen, it can be known that $q_o = Q_u \div Q_F = BW_L \div BW_u$.

Insertion Loss

Based on the foregoing, the designer can define the insertion loss (IL) of the filter. Fig. 2 illustrates the relationship between IL and q_o. It can be seen that a trade-off occurs: the closer Q_L is to Q_u, the higher the IL.

With knowledge of the value of C_o the designer can calculate the coupling capacitor value, C_{12} (derived from a dependency on C1 and C2). The term K_{12} represents the normalized coefficient of coupling for this circuit and is 0.707. Fig. 3 shows the circuit and contains the equation for determining the value of C_{12}. It should be noted that the terms C_{12} and K_{12} are not to be thought of as C twelve and K twelve. Rather, they are C one-two and K one-two, respectively.

In a double-tuned circuit like that of Fig. 3, *each* end of the filter must be terminated such that the single-loaded Q of the resonators is $Q_1 = Q_2 = \sqrt{2} \times f_o \div BW_L$. This differs from Q_F by $\sqrt{2}$.

As an example, if the first resonator is to be terminated in an R_o of 50 ohms, and L1 has a specified number of turns (N_1), a link can be added to L1 which will transform to provide a Q_L which equals Q_1. It may be shown that the number of unity-coupling turns on the link will be

$$N_L = \sqrt{\frac{R_o N_1^2}{Q_1\ 2\pi f L_o}\left(\frac{Q_u}{Q_u - Q_1}\right)} \quad \text{(Eq. 3)}$$

By following this procedure virtually any terminal impedance can be accommodated. A series capacitor can be used for the same purpose as the link just discussed. Fig. 4 shows the circuit configuration. Looking left from the dashed line, the Z is seen as $Z = R_o - j \div 2\pi f C_L$.

This same impedance will result from a very large parallel resistance and a shunt capacitance. The two are related by the usual series-to-parallel equivalent conversions

$$Y = \frac{1}{Z} = \frac{Z^*}{ZZ^*} = G + jB$$

$$G = \frac{1}{R_p} \text{ and } B = \frac{1}{X_p} \quad \text{(Eq. 4)}$$

where Y = admittance, Z = impedance, * = complex conjugate, G = conductance, B = susceptance, R_p = parallel resistance, and X_p = parallel reactance.

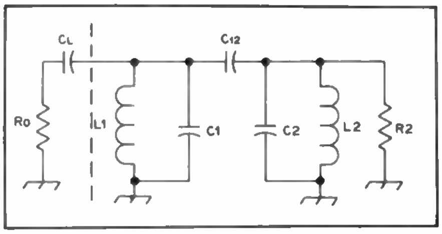


Fig. 4 — Two-resonator filter which uses capacitive coupling to match the filter to a load (R_L and C_L).

R_p can be given as

$$R_{p(ext.)} = Q_1\ 2\pi f L \left(\frac{Q_u}{Q_u - Q_1}\right) \quad \text{(Eq. 5)}$$

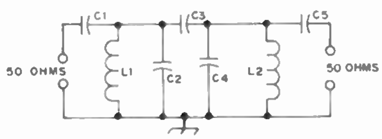
where $Q_u \div Q_u - Q_1$ corrects for the inherent resonator losses. Knowing $R_{p(ext.)}$ and R_o, it follows that

$$X_{CL} = \sqrt{R_{p(ext.)}\ R_o - R_o^2} \quad \text{(Eq. 6)}$$

In this example the capacitance of C_L will detune the resonator. In this relation $C1 = C_o - C_{12} - C_L$.

The loading of the other end of the filter is handled in exactly the same way. However, combinations of

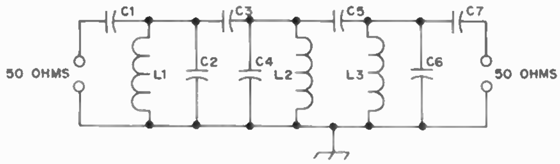
Table 1



$C2 = C_o - C1 - C3$
$C4 = C_o - C3 - C5$

BW_L (3 dB)	L1, L2 (µH)	C_o (pF)	C1 (pF)	C3 (pF)	C5 (pF)	R (L1-C2 and L2-C4)
1.8-1.9 MHz	5.12	1446	291	55.3	291	1804 ohms
1.8-1.85 MHz	5.12	1485	192	28.8	192	4120 ohms
1.8-1.9 MHz	8.74	847	221	32.4	221	3006 ohms
1.8-1.85 MHz	8.74	870	148	16.8	148	7020 ohms
3.5-3.6 MHz	5.12	393	75.1	7.8	75.1	7190 ohms
3.5-3.6 MHz	8.74	230	56.6	4.58	56.6	12.6 k ohms
3.5-3.7 MHz	5.12	352	110	15	110	3280 ohms
3.5-3.7 MHz	8.74	224	83.4	8.8	83.4	5670 ohms
3.8-4.0 MHz	5.12	325	93.2	11.8	93.2	3890 ohms
3.8-4.0 MHz	8.74	191	70.7	6.9	70.7	6700 ohms
5.0-5.2 MHz	8.74	111	40.1	3.09	40.1	12.1 k ohms
7.0-7.1 MHz	2.05	248	26.6	2.5	26.6	14.4 k ohms
7.0-7.2 MHz	2.05	245	42.4	4.88	42.4	5600 ohms
7.0-7.3 MHz	2.05	242	53.2	7.17	53.2	3550 ohms
7.0-7.2 MHz	8.74	57.5	19.1	1.14	19.1	27.4 k ohms
10.7-11.1 MHz	2.07	102.7	25.9	2.67	25.9	6400 ohms
10.8-11.0 MHz	2.07	102.7	16.7	1.33	16.7	15.3 k ohms
14.0-14.2 MHz	0.8	158.7	14.4	1.59	14.4	12.3 k ohms
14.0-14.2 MHz	2.08	61	9.2	0.61	9.2	30 k ohms
14.0-14.4 MHz	2.08	61	14.7	1.2	14.7	11.6 k ohms
16.0-16.5 MHz	0.81	118	20.4	2.57	20.4	4700 ohms
19.0-19.5 MHz	0.49	139	17.6	2.56	17.6	4490 ohms
19.0-20.0 MHz	0.49	136	26.6	4.92	26.6	1930 ohms
19.0-21.0 MHz	0.49	129	37.8	9.15	37.8	938 ohms
21.0-21.3 MHz	0.49	115	9.2	1.15	9.2	13.3 k ohms
21.0-21.5 MHz	0.49	114	14.1	1.9	14.1	5690 ohms
21.0-21.5 MHz	0.82	68.4	11.4	1.14	11.4	8700 ohms
28.0-28.5 MHz	0.48	65.3	7.2	0.82	7.2	12.4 k ohms
28.0-29.0 MHz	0.48	64	11.8	1.6	11.8	4500 ohms
41.0-42.0 MHz	0.49	29.9	4.6	.051	4.6	13.7 k ohms
41.0-43.0 MHz	0.49	29.3	7.6	0.1	7.6	4980 ohms

Table 2



BW_I (3 dB)	L1, L2, L3 (μH)	C2 (pF)	C4 (pF)	C6 (pF)	C1 (pF)	C7 (pF)	C3 (pF)	C5 (pF)	R (L1-C2 AND L3-C6)
1.8-1.825 MHz	14	431	541	477	115	68.7	4.8	5.4	11.7 k ohms 32.7 k ohms
1.8-1.850 MHz	14	360	522	416	172	117	10.8	10.0	5100 ohms 11.1 k ohms
1.8-1.850 MHz	5.12	1186	1430	1289	273	286	26.9	28.6	2000 ohms 5400 ohms
1.8-1.9 MHz	5.12	982	1336	1117	406	277	57.9	52.5	946 ohms 1981 ohms
1.8-1.850 MHz	8.74	646	838	724	209	130	15.9	16.7	3500 ohms 9000 ohms
1.8-1.9 MHz	8.74	505	783	605	308	212	33.9	30.7	1600 ohms 3400 ohms
1.8-2.0 MHz	8.74	314	684	438	426	311	65.2	55.9	827 ohms 1500 ohms
3.5-3.6 MHz	8.74	146	221	174	79.3	51.2	4.5	4.5	6400 ohms 15.4 k ohms
3.5-3.7 MHz	8.74	99.6	206	135	115	80.5	9.3	8.3	3000 ohms 6000 ohms
3.8-4.0 MHz	8.74	85.9	177	116	97.5	67.8	7.3	6.5	3600 ohms 7300 ohms
3.5-3.6 MHz	5.12	280	377	316	105	68.8	7.8	7.6	3700 ohms 8600 ohms
3.5-3.7 MHz	5.12	214	352	261	152	107	16	14.1	1800 ohms 3500 ohms
3.8-4.0 MHz	5.12	184	302	224	129	90	12.5	11.2	2000 ohms 4200 ohms
3.9-4.0 MHz	5.12	228	306	258	84	54	5.5	5.6	4700 ohms 11.2 k ohms
5.0-5.2 MHz	8.74	52.7	105	71.1	55.6	37.4	3.1	3.0	6400 ohms 14 k ohms
7.0-7.1 MHz	2.05	208	244	220	38.4	25.8	2.1	2.6	7000 ohms 15 k ohms
7.0-7.2 MHz	2.05	181	236	201	59.1	39.1	4.9	4.7	2900 ohms 6600 ohms
7.0-7.3 MHz	2.05	161	227	184	73.7	68	7.5	7.0	1870 ohms 3860 ohms
7.0-7.2 MHz	8.74	99.6	206	135	115	80.5	9.3	9.3	13.7 k ohms 36.4 k ohms
10.7-11.1 MHz	2.076	63.9	97.4	75.4	36	24.8	2.7	2.5	3300 ohms 7000 ohms
10.8-11.0 MHz	2.076	77.8	100	86.9	23.7	14.5	1.2	1.3	7600 ohms 20.3 k ohms
14.0-14.2 MHz	2.08	47.4	60	51.7	13.2	8.9	0.5	0.64	14.5 k ohms 32 k ohms
14.0-14.2 MHz	0.8027	137	156	143	20.9	13.7	1.2	1.7	5800 ohms 13.6 k ohms
14.0-14.4 MHz	2.08	38.6	58	45.6	20.5	13.6	1.2	1.16	6000 ohms 13.7 k ohms
16.0-16.5 MHz	0.811	87.3	113.2	97	28.4	18.8	2.6	2.5	2400 ohms 5500 ohms
19.0-19.5 MHz	0.491	112	134	121	25	15.3	2.4	2.6	2200 ohms 5900 ohms
19.0-20.0 MHz	0.491	93	126	106	37.2	25.3	5.1	4.7	1000 ohms 2100 ohms
19.0-21.0 MHz	0.491	66.7	111	83.1	52.7	37.7	9.9	8.5	507 ohms 943 ohms
21.0-21.4 MHz	0.491	96.2	112	101.6	17.3	11.6	1.25	1.59	3800 ohms 8400 ohms
21.0-21.5 MHz	0.820	51.3	66.2	57.2	16.0	10.1	1.0	1.1	4400 ohms 11.1 k ohms
21.0-21.6 MHz	0.491	89	109	97.4	22.6	14.1	2.1	2.2	2200 ohms 5700 ohms
28.0-28.5 MHz	0.485	54	63.9	59	10.5	5.4	0.61	0.87	5800 ohms 2200 ohms
28.0-29.0 MHz	0.485	46	61.1	52	16.6	10.7	1.6	1.6	2300 ohms 5500 ohms
41.0-42.0 MHz	0.491	22.8	29	26	6.7	3.5	0.38	0.54	6400 ohms 2500 ohms
41.0-43.0 MHz	0.491	17.6	27.3	21.4	10.7	6.9	9.6	9.7	2500 ohms 6100 ohms
50.0-52.0 MHz	0.491	12.3	18.8	15.1	7.0	4.2	0.5	0.55	4000 ohms 11 k ohms

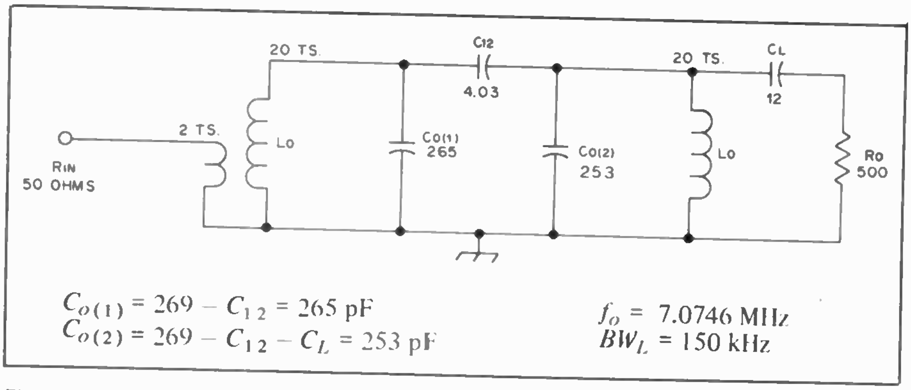


Fig. 5 – Circuit example of a two-resonator filter.

methods may be used, and the input termination need not equal the output load.

A Design Example

To follow a rather casual design approach, we shall assume that some Amidon T-68-6 cores ($A_L = 47$, from Table 6) are available, and that 20 turns of wire are wound on each. Each inductor has a measured Q_u of 275 at 7 MHz, and a double-tuned filter is desired. The 3-dB points (f_1 and f_2) shall fall at 7.0 and 7.15 MHz, respectively. Therefore

$$f_o = \sqrt{f_1 f_2} = \sqrt{7.0 \times 7.15}$$

$$= 7.0746 \text{ MHz} \qquad \text{(Eq. 7)}$$

If 20 turns are wound on the cores, then $L_o = A_L \times 10^{-4} \times N^2 = 47 \times .0001 \times 400 = 1.88\ \mu$H.

Next we will find the reactance of L_o: $X_{Lo} = 2\pi f_o L_o = 6.28 \times 7.0746 \times 1.88 = 83.5$ ohms.

Proceeding

$$C_o = \frac{1}{2\pi f_o X_{Lo}}$$

$$= \frac{1}{6.28 \times 7.0746 \times 83.5} = \frac{1}{3709.7}$$

$$= .000269\ \mu\text{F} \times 10^6 = 269 \text{ pF} \qquad \text{(Eq. 8)}$$

Now it is necessary to learn what C_{12} is

$$C_{12} = C_o \times \frac{BW_L}{f_o} \times K_{12}$$

$$= 269 \times \frac{0.15}{7.0746} \times 0.707$$

$$= 4.03 \text{ pF} \qquad \text{(Eq. 9)}$$

where K_{12} is the coupling coefficient for this filter.

Resonator Q is found from

$$Q_1 = Q_2 = \sqrt{2} \times \frac{f_o}{BW_L} = 1.414$$

$$\times \frac{7.0746}{0.15} = 66.7 \qquad \text{(Eq. 10)}$$

Now, if it is desired to link couple one end of the filter to a 50-ohm load (R_{in})

$$N_{link} = \sqrt{\frac{R_{in} N^2}{Q_{Lo} X_{Lo} \left(\frac{Q_u}{Q_u - Q_{Lo}}\right)}}$$

$$= \sqrt{\frac{50 \times 20^2}{66.7 \times 83.5 \left(\frac{275}{275 - 66.7}\right)}}$$

$$= \sqrt{\frac{20{,}000}{7351.6}} = 1.65 \text{ turns} \qquad \text{(Eq. 11)}$$

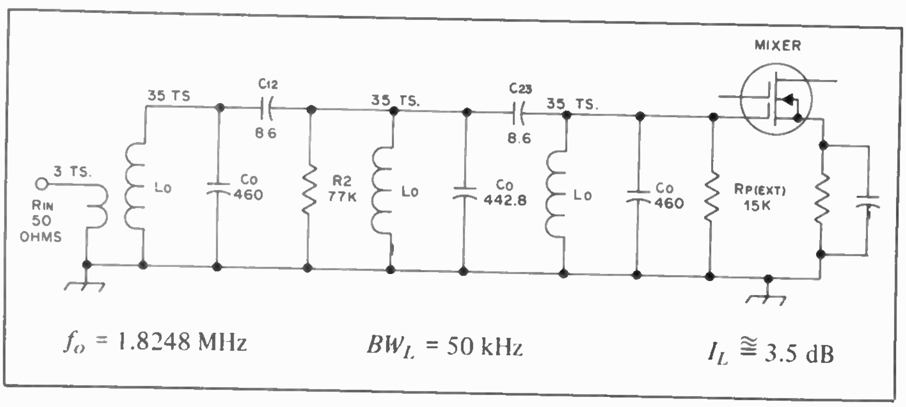


Fig. 6 – Example of a three-pole filter as applied to a receiver front-end circuit.

Two turns will be close enough in number to meet the design requirement.

Should the designer wish to terminate the filter in 50 ohms at each end, a second two-turn link can be added, one on each resonator.

However, if one end is terminated in 50 ohms, and the other end is to be connected to, say, a 500-ohm load (R_o), a series coupling capacitor can be used instead of a link, as shown by the equation given earlier

$$R_{p(ext.)} = Q_2\, 2\pi f_o L_o \left(\frac{Q_u}{Q_u - Q_1}\right)$$

$$= 66.7 \times 6.28 \times 7.0746$$

$$\times 1.88 \left(\frac{275}{275 - 66.7}\right)$$

$$= 5571 \times 1.32 = 7355 \text{ ohms}$$

$$\therefore X_{CL} = \sqrt{R_{p(ext.)}\ R_o - R_o^2}$$

$$= \sqrt{7355 \times 500 - 500^2}$$

$$= \sqrt{3{,}427{,}500}$$

$$= 1851 \text{ ohms}$$

$$\therefore C_L = \frac{1}{2\pi f_o X_{CL}}$$

$$= \frac{1}{6.28 \times 7.0746 \times 1851}$$

$$= \frac{1}{82{,}237} = .00001215\ \mu\text{F} \times 10^6$$

$$= 12.15 \text{ pF} \qquad \text{(Eq. 12)}$$

Table 3

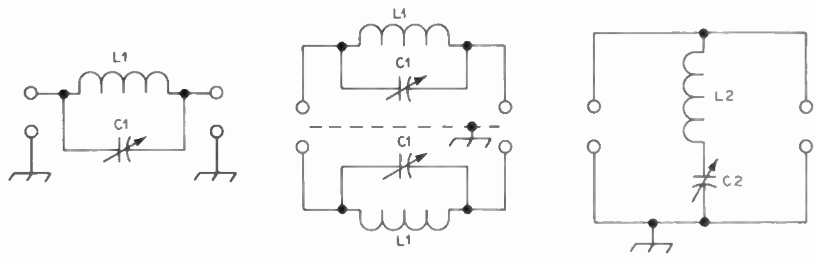
BW_x	Ω	*ATTEN.*
20 kHz	0.4	.02 dB
35 kHz	0.7	0.48 dB
50 kHz	1	3.01 dB
70 kHz	1.4	9.31 dB
100 kHz	2.0	18.13 dB
125 kHz	2.5	23.9 dB
150 kHz	3.0	28.6 dB
200 kHz	4.0	36.1 dB
300 kHz	6.0	46.7 dB
400 kHz	8.0	54.2 dB

A capacitor value of 12 pF is adequate. The completed circuit is shown in Fig. 5.

The Three-Resonator Filter

The design of filters which contain three or more pole pairs is similar to that of the two-pole example. The major exception is that in the two-pole design constants were used in the equations relative to C_{12}, Q_1 and Q_2. They were, respectively, $\sqrt{2} \div 2$, and $\sqrt{2}$. However, when the number of resonators (N_R) is equal to or greater than 3, the "constants" are complicated functions of the

Table 4



MHz	L1 (μH)	C1 (pF) MAX.	L2 (μH)	C2 (pF) MAX.
50-54	0.79	20	3.18	8
28-30	1.42	30	5.68	10
21.0-21.4	1.89	45	7.5	15
14-14.4	2.84	45	11.37	15
7-7.3	5.68	100	22.74	30
3.5-4.0	11.37	200	45.4	50
1.8-2.0	22.11	400	88.4	100
1.0-1.6	39.8	700	159.2	200
0.55-1.0	72.3	1500	289.5	300

normalized resonator Q and q_o. The constants are taken from Zverev, *Handbook of Filter Synthesis*, John Wiley and Sons, New York, NY.

For simplified amateur design work we will restrict the q_o to 5. Therefore, BW_L must be equal to or greater than five times BW_u. In order to satisfy this criterion, we must place a resistor across the center resonator of the circuit in Fig. 6.

Given values of C_o and L_o for a center frequency of f_o

$$C_{12} = C_{23} = C_o \times \frac{BW_L}{f_o} \times 0.68,$$

$$Q_1 = \frac{f_o}{BW_L} \times 0.822, \text{ and } Q_3$$

$$= \frac{f_o}{BW_L} \times 1.71 \qquad \text{(Eq. 13)}$$

The loading of resonators 1 and 2 is handled as it was when designing the two-pole filter. The resistor, R2, across the center resonator is used to degrade the Q_u to five times Q_L

$$R2 = 5Q_F f_o L_o \left(\frac{Q_u}{Q_u - 5Q_F}\right) \qquad \text{(Eq. 14)}$$

where Q_u = basis unloaded Q of the resonator, and $Q_F = f_o \div BW_L$.

A Design Example

The designer can assume in this example that some Amidon T-106-2 toroid cores are available, and that they will provide a Q_u of 330 at 1.8 MHz. It will be assumed also that the –3-dB points of the filter will occur at an f_1 of 1.8 MHz and an f_2 of 1.85 MHz. Therefore BW_L will be 50 kHz (.05 MHz).

$$f_o = \sqrt{f_1 f_2} = \sqrt{1.8 \times 1.85}$$

$$= 1.8248 \text{ MHz} \qquad \text{(Eq. 15)}$$

We will wind 35 turns on the cores. The inductance of L_o will be based on A_L, which from Table 6 is 135

$$L_o = A_L \times 10^{-4} \times N^2$$

$$= 135 \times .0001 \times 1225 = 16.537\,\mu\text{H}$$

$$\therefore X_{Lo} = 2\pi f_o L_o = 189.5 \text{ ohms} \qquad \text{(Eq. 16)}$$

Next we will find the value of C_o:

$$C_o = \frac{1}{(2\pi f_o)^2 X_{Lo}} = \frac{1}{2171.6}$$

$$= .00046\,\mu\text{F} \times 10^6 = 460 \text{ pF} \qquad \text{(Eq. 17)}$$

The values of C_{12} and C_{23} can be determined by

$$C_{12} = C_{23} = C_o \times \frac{BW_L}{f_o} \times 0.68 \text{ (coupling coeff.)}$$

$$= 460 \times \frac{.05}{1.8248} \times 0.68 = 8.6 \text{ pF} \qquad \text{(Eq. 18)}$$

Proceeding with the design order

$$Q_1 = \frac{f_o}{BW_L} \times 0.822$$

$$= \frac{1.8248}{.05} \times 0.822 = 30$$

$$\text{and } Q_3 = \frac{f_o}{BW_L} \times 1.71$$

$$= \frac{1.8248}{.05} \times 1.71 = 62.4 \qquad \text{(Eq. 19)}$$

As an example, say the filter is to be used between a 50-ohm antenna and a MOSFET mixer. The link for coupling at R_{in} will contain the number of turns found by

$$N_{link} = \sqrt{\frac{R_{in} N^2}{Q_{Lo} X_{Lo}\left(\frac{Q_u}{Q_u - Q_{Lo}}\right)}}$$

$$= \sqrt{\frac{50 \times 35^2}{30 \times 189.5 \left(\frac{330}{330 - 30}\right)}}$$

$$= \sqrt{9.794} = 3.12 \text{ turns} \qquad \text{(Eq. 20)}$$

The opposite end of the filter will have to be terminated in 14,532 ohms, as determined by

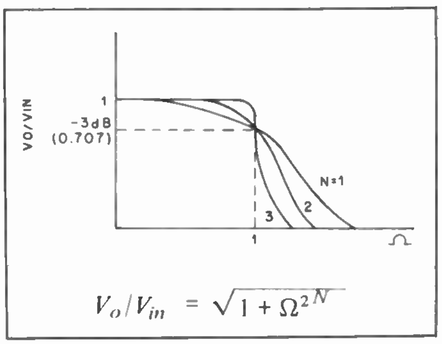


Fig. 7 – Curves illustrating selectivity of 1, 2 and 3 resonators in a filter.

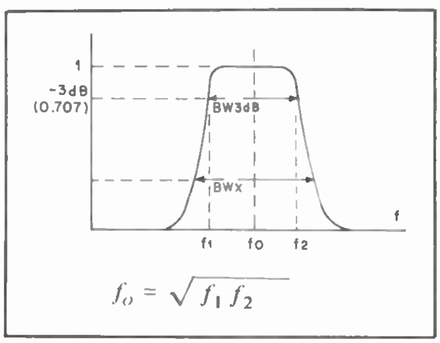


Fig. 8 – Characteristic curve for BW_{3dB} and BW_x.

Table 5

$Q_L = 1$
R_{in}, R_o = 50 ohms
X_L = 50 ohms
$X_{C1, C3}$ = 50 ohms
X_{C2} = 25 ohms

BAND (METERS)	*L1, L2 (μH)*	*C1, C3 (pF)*	*C2 (pF)*
160	3.98	1592	3184
80 (cw)	2.15	860	1721
75 (phone)	1.99	796	1592
40	1.09	436	872
20	0.55	221	443
15	0.372	149	298
10	0.268	107	214
6	0.157	63	126

$$R_{p(ext.)} = Q_3 2\pi f_o L_o \left(\frac{Q_u}{Q_u - Q_3}\right)$$

$$= 62.4 \times 6.28 \times 1.8248 \times 16.53$$

$$\left(\frac{330}{330 - 62.4}\right) = 11{,}815.2 \times 1.23$$

$$= 14{,}532 \text{ ohms} \qquad \text{(Eq. 21)}$$

A 15,000-ohm standard-value resistor will be suitable.

The final design step requires calculating the value of R2

$$R2 = 5Q_F 2\pi f_o L_o \left(\frac{Q_u}{Q_u - 5Q_F}\right)$$

$$= 5 \times 36.5 \times 6.28 \times 1.8248 \times 16.53$$

$$\left(\frac{330}{330 - 182.5}\right) = 34{,}570.8 \times 2.23$$

$$= 77{,}093 \text{ ohms} \qquad \text{(Eq. 22)}$$

where

$$Q_F = \frac{f_o}{BW_L} = \frac{1.8248}{.05} = 36.5$$

$$\text{and } Q_u = 330 \qquad \text{(Eq. 23)}$$

A value of 77,000 ohms will suffice.

Fig. 6 illustrates the three-pole filter as it might be used in a practical application. It should be noted that if the classic and somewhat complex filter equations had been used there would be a slight difference in the computed values of C_{12} and C_{23}. However, in amateur work they can be made the same in value without any performance degradation observed. It is recommended that if the nearest standard capacitor value is desired at C_{12} and C_{23}, the designer should select the nearest *lower* value. In this case one could place in series a 12- and 27-pF silver mica capacitor to obtain a value of 8.3 pF. Examination of Fig. 6 will show that C_o of the center resonator is not 460 pF on the diagram. The actual value is 460 pF, but it is necessary to subtract C_{12} and C_{23} from the C_o value of 460 pF (460 minus 8.6 minus 8.6 = 442.8) to prevent detuning of the resonator. A practical filter would contain high-Q trimmer capacitors at all three C_o points to permit final tweaking of the resonators. In the example given here, where q_o has been degraded to 5, mica compression trimmers would be suitable. As we stated earlier in this book, the Q of a resonant circuit is dependent in part upon the Q of the capacitor used in the circuit.

Table 1 lists values of L, C, BW_L and R for two-pole Butterworth band-pass filters. Through knowledge of the R values, it is possible to match the filters to nearly any external impedance or combination of two external impedances, as stated earlier in this chapter.

Table 2 contains similar data for three-pole Butterworth band-pass filters. Both tables list ranges of frequencies within the various hf amateur bands. Additional frequencies are given for use in VFO and heterodyne-mixer output circuits. The data in the tables were obtained by means of a digital computer program, and are based on classic design rather than the simplified approach just treated (Q has not been purposely degraded.)

Shape Factor

The filter-response shape is also significant to the designer. The response for the voltage-transfer ratio of a Butterworth characteristic is given by

$$\frac{V_o}{V_{in}} = \frac{1}{\sqrt{1 + \Omega^{2N}}} \qquad \text{(Eq. 24)}$$

where Ω is a variable parameter, and N is the number of resonators in the filter. Of special importance is the point where the value of Ω is equal to 1, since $V_o \div V_{in}$ (voltage output and voltage input) will be equal to $1 \div \sqrt{2}$ or 0.707. This is called the half-power (−3 dB) *frequency* in the case of a low-pass filter, or the −3-dB *bandwidth* of a band-pass filter. Fig. 7 shows that the response curves for all values of N pass through this point.

In low-pass filter design work Ω can be found by $\Omega = f_X \div f_{3dB}$, where f_{3dB} is the frequency at which $V_o \div V_{in}$ is down by 3 dB, and where f_X is the new variable.

In the case of a band-pass filter, Ω can be obtained from $\Omega = BW_X \div BW_{3dB}$, where BW_X is some *arbitrary* bandwidth, and BW_{3dB} is the 3-dB bandwidth of the filter. This condition can be seen in Fig. 8.

The formula for a Butterworth filter can be expressed as Attenuation (dB) = $10 \log_{10} (1 + BW_X \div BW_{3dB})$.

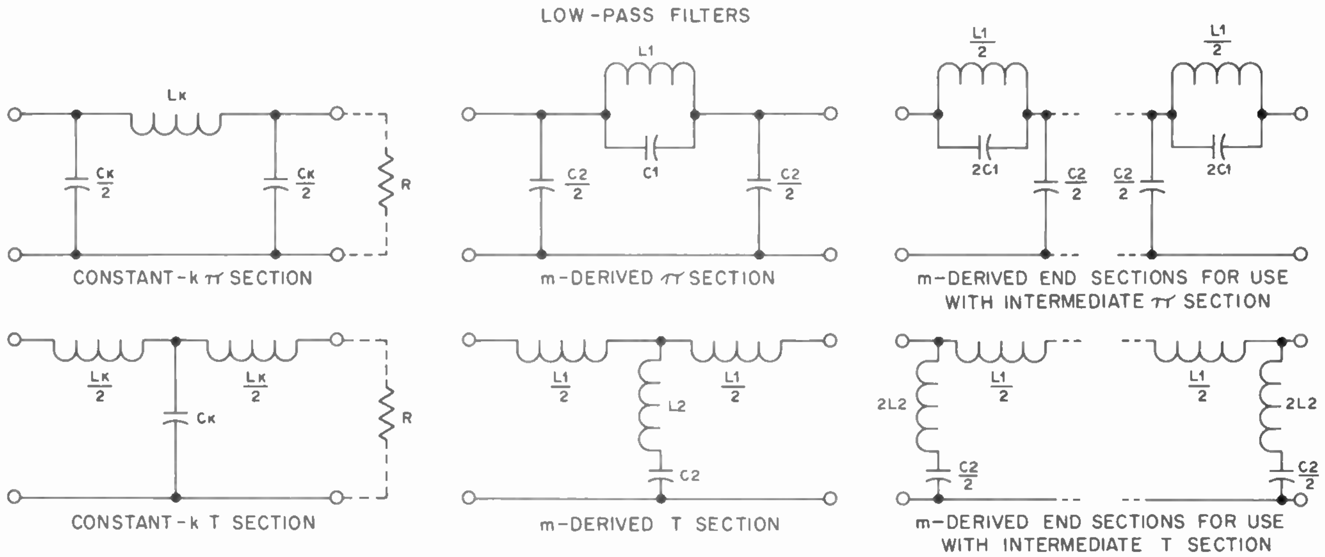
To illustrate, consider a 3-pole filter with a BW_{3dB} of 50 kHz. Assume that the f_o is 1800 kHz, and it is desired to know how much rejection the filter will provide in eliminating the unwanted effects of a strong Loran signal at 1900 kHz. First, the value of BW_X must be learned. BW_X is given by f_2 minus f_1, where $f_o = \sqrt{f_1 \times f_2}$. In this situation f_2 is equal to 1900 kHz – the frequency where the attenuation is desired. Then, f_1 will be given by $f_1 = f_o^2 \div f_2 = 1800^2 \div 1900 = 1705.26$ kHz. BW_x is then equal to 1900 minus 1705.26 = 194.7 kHz. Thus, Ω is equal to 194.7 ÷ 50 = 3.89. Table 3 shows that the closest value of Ω corresponding to the number is 4, and the rejection at this value is 36.1 dB. Therefore, the filter will reject the LORAN signal at 1900 kHz by some 36 dB.

Image-Parameter Filters

Many amateurs are familiar with the image-parameter filters shown in Fig. 9. They are based on transmission-line concepts, and when compared to modern filters of the variety described earlier in this chapter, they are best classified as "approximate" designs. Modern filters are totally predictable with respect to bandpass characteristics when they are terminated by proper resistances. The same is not true of an image-parameter design. Nonetheless, the latter type is useful in amateur work and is simple to design. Fig. 9 contains a catalog of configurations and equations for image-parameter filters. A design example is offered to illustrate the manner in which a filter can be calculated.

A High-Pass Filter

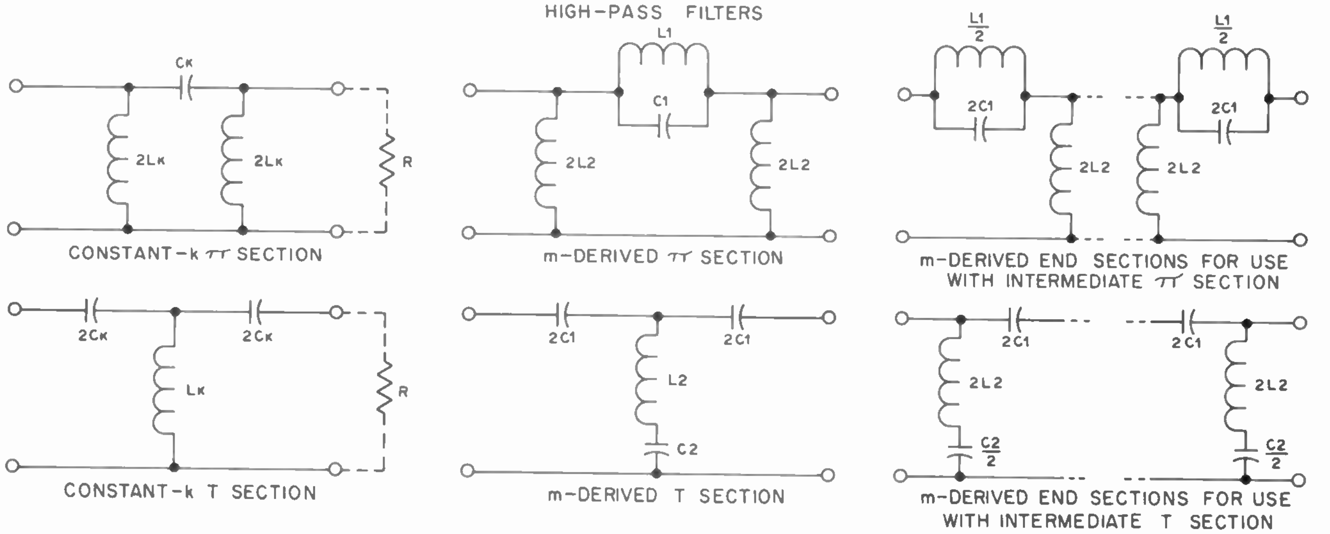
The design goal in this example is to build a 50-ohm filter which will be used at the input of a 160-meter receiver to attenuate the broadcast-band signals which would otherwise overload the receiver front end and cause IMD and desensitization. From Fig. 9 we select a filter shown under the heading "High-Pass Filters." Our cutoff frequency shall be 1650 kHz, thereby allowing the 160-meter signals to pass, but greatly attenuating the bc-band energy present



$$L_K = \frac{0.318\,R}{f_c} \qquad C_K = \frac{3.18\,(10^5)}{f_cR}$$

$$L_1 = mL_K \qquad C_1 = \frac{1-m^2}{4m}\,C_K \qquad L_2 = \frac{1-m^2}{4m}\,L_K \qquad C_2 = mC_K$$

$$L_1 = mL_K \qquad C_1 = \frac{1-m^2}{4m}\,C_K \qquad L_2 = \frac{1-m^2}{4m}\,L_K \qquad C_2 = mC_K$$



$$L_K = \frac{0.0796\,R}{f_c} \qquad C_K = \frac{7.96\,(10^4)}{f_cR}$$

$$L_1 = \frac{4m}{1-m^2}\,L_K \qquad C_1 = \frac{C_K}{m} \qquad L_2 = \frac{L_K}{m} \qquad C_2 = \frac{4m}{1-m^2}\,C_K$$

$$L_1 = \frac{4m}{1-m^2}\,L_K \qquad C_1 = \frac{C_K}{m} \qquad L_2 = \frac{L_K}{m} \qquad C_2 = \frac{4m}{1-m^2}\,C_K$$

Fig. 9 — Basic filter sections and formulas for image-parameter designs.

on the antenna. The progression (Fig. 10) is

$$L_K = \frac{.0796 \times R}{f_C} = \frac{3.98}{1.65} = 2.4\ \mu\text{H}$$

$$C_K = \frac{7.96\,(10^4)}{f_cR} = \frac{79{,}600}{1.65 \times 50} = 965\ \text{pF}$$

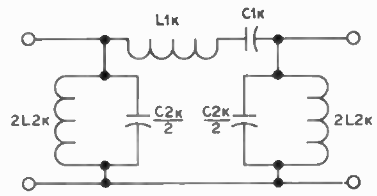
$$\therefore C1, C2 = 2C_K = 1930\ \text{pF} \qquad \text{(Eq. 25)}$$

where f_c is the desired cutoff frequency (low side) in MHz, L_K is the inductance in μH, and R is the resistance in ohms.

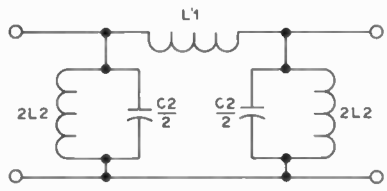
Tunable Filters

In some amateur applications a fixed-tuned filter may not be desired. The builder may wish to enhance the circuit selectivity by using a narrow-band tunable filter. In such cases a single tuned filter rarely contains more than three resonators. Top coupling can be used (DeMaw, "More Receiver Design Notes," *QST* for June, 1974, page 23), or the designer may elect to use bottom coupling (Sabin, "The Solid-State Receiver," *QST* for July, 1970, page 35). Fig. 11 shows a two-resonator top-

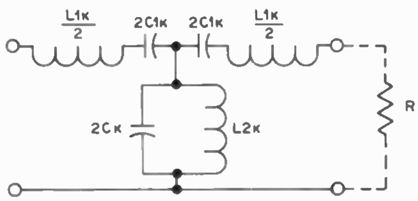
BAND-PASS FILTERS



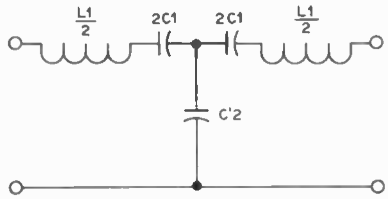
CONSTANT-k π SECTION



THREE-ELEMENT π SECTION



CONSTANT-k T SECTION



THREE-ELEMENT T SECTION

$$L_{1K} = \frac{0.318\,R}{(f_2 - f_1)}$$

$$L_1 = \frac{0.318\,R}{f_1 + f_2}$$

$$C_{1K} = \frac{7.96\,(f_2 - f_1)\,10^4}{f_1\,f_2 R}$$

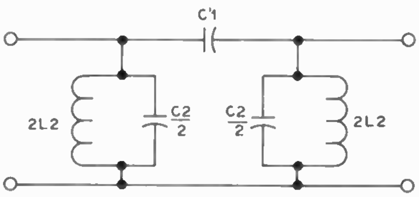
$$C_1 = \frac{7.96\,(f_2 - f_1)\,10^4}{f_1\,R}$$

$$L_{2K} = \frac{0.0796\,(f_2 - f_1)\,R}{f_1\,f_2}$$

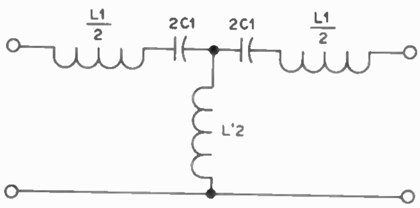
$$L_1 = L_{1K} \qquad L_2 = \frac{0.0796\,(f_2 - f_1)\,R}{{f_1}^2}$$

$$C_{2K} = \frac{3.18\,(10^5)}{(f_2 - f_1)\,R}$$

$$C_2 = C_{2K} \qquad C'_2 = \frac{3.18\,(10^5)}{(f_1 + f_2)\,R}$$



THREE-ELEMENT π SECTION



THREE-ELEMENT T SECTION

$$L_1 = \frac{0.318\,f_1\,R}{f_2\,(f_2 - f_1)}$$

$$C_1 = C_{1K}$$

$$L'_2 = \frac{0.0796\,(f_1 + f_2)\,R}{f_1\,f_2}$$

$$L_2 = L_{2K}$$

$$C'_1 = \frac{7.96\,(f_1 + f_2)\,10^4}{f_1\,f_2\,R}$$

$$C_2 = \frac{3.18\,f_1\,(10^5)}{f_2\,(f_2 - f_1)\,R}$$

coupled tunable filter, and one which contains three resonators and is bottom coupled. Either coupling system will provide good performance, and the coupling amount should be set to assure a single-hump f_o' response. The lighter the coupling between resonators, the less chance for pass-band ripple. The three-resonator filter shown in Fig. 11 was adjusted experimentally for a desired insertion loss of 5 dB by adjusting the input and output loading amounts.

It will be noted that a three-section variable capacitor can be used to tune the filter containing three resonators. This is possible by paralleling inductances L2 and L3 to obtain an inductance value equal to that of L1 and L4. These filters are frequently referred to as "tunable Cohn filters." For more info on Cohn filters see, Cohn, "Dissipation Loss in Coupled Resonator Filters," *Proc. IRE*, August, 1959, page 1342.

Tunable filters are narrow-band devices. They must be optimized for bandwidths on the order of 200 kHz or less. If made wider, the selectivity and insertion loss characteristics will degrade markedly from the design center.

Wave-Trap Filters

Wave traps are effective in eliminating many kinds of amateur-related interference. Table 4 shows a single parallel-tuned trap which can be used in coaxial or single-wire lines to reject an undesired frequency. Two such traps can be used in a balanced line (TV ribbon). A series-tuned trap is shown also. It can be employed in place of a parallel-tuned trap, or in combination with one, for greater rejection of a particular frequency. The desired frequencies will not be affected by the traps.

Filters of this variety are useful in trapping amateur signals at the input of TV front ends, fm tuners, and a-m radios. They are often found in heterodyne exciters where they are used to trap unwanted products at the mixer output or elsewhere in the circuit. Traps can be installed in the feeders of amateur antennas to reject commercial bc-station energy which might otherwise overload amateur receivers.

These filters are principally single-frequency devices when tuned to the interference frequency. The greater the tuned-circuit Q, the narrower the trap bandwidth, and the more effective the rejection capability. High-Q coils and capacitors are recommended for best results. Traps used in antenna lines should be contained in shield enclosures, and the enclosure should be attached to an earth ground.

The trimmer values listed in Table 4 will permit resonating of the traps across the MHz ranges listed for each. In some stubborn cases of amateur TVI, a trap tuned to the interfering amateur frequency may be more effective than a high-pass filter in preventing TV front-end overloading.

AC-Line Filter

It is often necessary to utilize an ac-line filter to prevent interference from being carried through the mains. Such a device helps to prevent rf from transmitters from entering the ac line and being conveyed to house radios, TV sets, and hi-fi equipment, directly or by means of powerline radiation of rf. Additionally, a line filter will reduce interference to amateur receivers caused

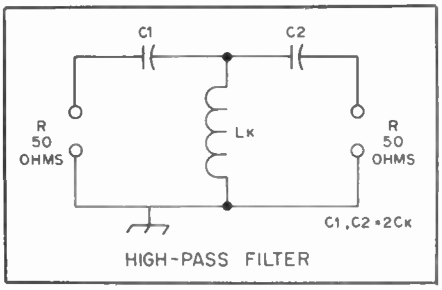


Fig. 10 — Circuit for a simple high-pass filter.

by line noise and bc-band energy which might be present.

The filter illustrated in Fig. 12 is a low-pass type, designed for a Q_L of 1, and a roll-off frequency of 500 kHz. Because it is a low-pass filter, the 60-Hz line frequency will not be attenuated. Impedance matching in an ac line is not a significant matter, so an arbitrary impedance of 50 ohms is used in the filter design. This, plus a Q of 1, gives X_L and X_C values of 50.

The f_c is the cutoff frequency in MHz, C is in µF, and X is in ohms. The wire size used in the coils must be large enough in cross section to handle the current taken by the equipment with which the filter is used (see wire table elsewhere in this book).

C1 and C2 should be 1000 V disk ceramic types. The filter should be enclosed in a metal case, and an earth ground should be affixed to the box. For best results the filter should be installed as close to the equipment as practicable. In some instances a line filter can be built inside the equipment chassis, just where the ac line enters. This filter will attenuate all rf energy above 500 kHz. For stringent filtering jobs, the builder can cascade two or more of these filter sections, although such a severe measure is seldom necessary.

Helical Resonators

A helical resonator is in physical terms a high-Q single-layer (solenoid) coil which is contained in a cylindrical or rectangular shield compartment. Electrically, these resonators are one-quarter wavelength devices, and function in the same manner as a quarter-wavelength coaxial-line resonator or cavity resonator. Helical resonators offer the distinct advantage of being much smaller in size than the more common coaxial-line resonator. Therefore, they lend themselves nicely to use in vhf and uhf receiver front ends, and in transmitter output circuits. Values of Q can be achieved from several hundred to, say, 1000. This makes practical the application of two or more helical resonators in a filter configuration. Coupling between filter sections can be effected by means of an aperture in the shield wall between resonators, by tapping the coils near the low-impedance ends, using links, or by probe coupling at the high-impedance ends of the resonators. The aperture-coupling method is preferred by many, for it permits varying the coupling amount by adjusting the size of the aperture. Furthermore, it eliminates the possibility of Q degradation by attachment of leads to the resonator coils.

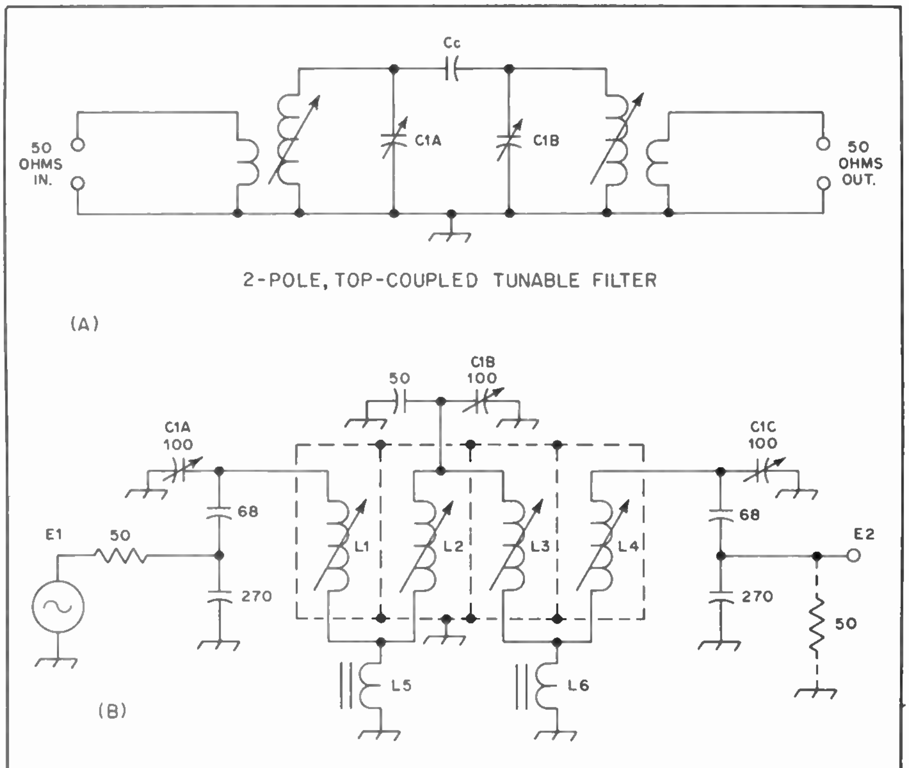
The material offered in this section of the chapter is based on information found in a paper by MacAlpine and Schildknecht, "Coaxial Resonators with Helical Inner Conductor," *Proc. IRE, Vol. 47, No. 12*, p. 2100, December, 1959. Some of the graphics used in this discussion are based on those used in the *IRE* paper.

Resonance is dependent for the most part on distributed inductance and capacitance, which sets the helical resonator apart from the more conventional lumped-L/C resonant circuit. High-Q trimmer capacitors are often used at the high-impedance ends of the resonators to provide precise tuning. Ideally, an air-dielectric capacitor should be used, and the value of capacitance required for resonance should be as small as possible in order to preserve the circuit Q. A theoretically perfect helical resonator would have no tuning capacitor associated with it and would be resonant at the desired frequency through its natural distributed capacitance and inductance.

One end of the coil should be connected to the shield enclosure in the best electrical manner possible. For amateur work it is suggested that the resonator be designed in accordance with the equations given here, but that the f_o be chosen slightly below desired resonance. This will permit use of a small trimmer at the high-impedance end of the resonator for frequency alignment.

Fig. 13A shows a helical resonator and defines the dimensions pertinent to the equations used in design procedure. At B in Fig. 13 is a two-resonator filter. The shield enclosures should be seamless to assure highest Q. Although the bottom end of the container need not be closed, it is recommended that the builder do so.

The unloaded Q of a helical resonator



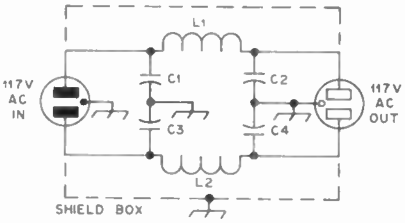
L1, L4 – 60 µH (NOM)
L2, L3 – 120 µH (NOM.)
L5, L6 – 145 µH (NOM.)

fo = 1.9 MHz

$dB = 20 \log \frac{E1}{E2}$

kHz FREQ.	-dB
2050	-45
2000	-35
1950	-16
1900	0
1850	-17
1800	-37
1750	-50

Fig. 11 — At A, an example of a tunable top-coupled, two-pole filter. The circuit at B shows a three-pole tunable Cohn filter which employs bottom coupling (L5 and L6).



X_L = 50 ohms
X_C = 50 ohms

$$L1, L2 = \frac{X_L}{2\pi f_c} = \frac{50}{6.28 \times 0.5} = 15 \mu H$$

$$C1, C2, C3, C4 = \frac{1}{2\pi f_C X_C} = \frac{1}{6.28 \times 0.5 \times 50} = .006 \mu F$$

Fig. 12 — Example of a "brute-force" low-pass ac-line filter. One half of this filter is suitable for dc-lead filtering.

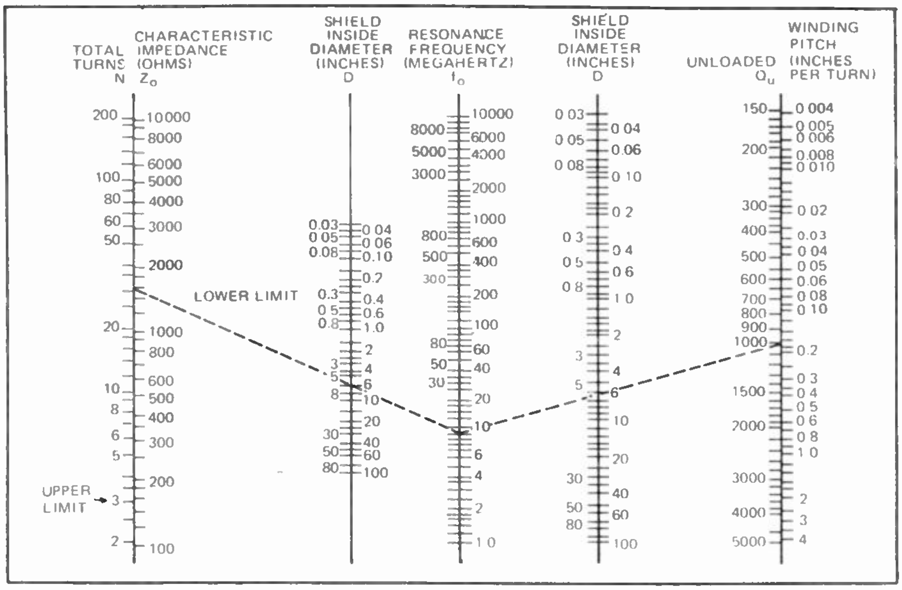


Fig. 14 — Nomograph for designing helical-resonator filters.

which is contained in a cylindrical copper or silver-plated brass shield can be obtained from $Q_u = 50D\sqrt{f_o}$ where D is the inner diameter of the shield and f_o is the resonant frequency in MHz. Thus, if a D of 3 inches were used, and f_o were 100 MHz, the Q_u would be 1500.

For practical helical resonators the ratio of d/D should be greater than 0.45 and less than 0.6. Also, b/d should be greater than 1, where b is the axial length of the coil in inches. The coil wire thickness (d_o) divided by the center-to-center spacing between turns (τ) should be greater than 0.4 and less than 0.7 when b/d = 1.5. When b/d = 4, d_o/τ should be greater than 0.5 and less than 0.7.

To determine the total number of turns on the coil, $N = 1900 \div f_oD$ for a d/D equal to 0.55, and a b/d greater than 1. Thus, if the quantity of D was 3 inches, and f_o was 100 MHz, N would equal 6.33 turns.

To determine winding pitch: $\tau = 1/n$, where n is the turns per inch. The impedance (Z_o) can be found from: $Z_o = 98{,}000/f_oD$ ohms. Therefore, if D = 3 inches, and f_o = 100 MHz, Z_o must equal 326 ohms, when d/D = 0.55, and b/d = 1.5.

Fig. 14 contains a nomograph for designing helical resonators. It can be seen that practical resonators can be constructed for frequencies down to 10 MHz or slightly lower.

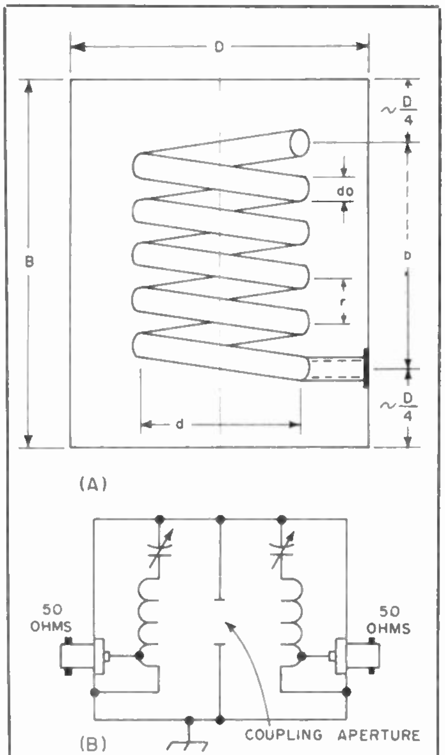


Fig. 13 — Details of helical-resonator design.

Band-Reject Filter

Band-reject filters are useful in preventing unwanted rf energy from entering a circuit. That is, an interfering band of frequencies may be present in such strength as to impair receiver performance if allowed to reach the receiver front end. An example would be if an amateur lived near a short-wave commercial broadcast station which operated between 40 and 20 meters. A band-reject filter could be placed in the receiver antenna line to remove most of the unwanted signal energy, but frequencies above and below those the filter would reject would pass without attenuation. Such a filter is in principle the inverse of a band-pass filter.

Fig. 15 shows the circuit for a band-reject filter, and gives equations for calculating the C and L values. The design is based on the image-parameter filter concept.

As an example of design, suppose the builder was having receiver overload problems from some nearby a-m broadcast stations. He would build a filter which would reject the frequencies from 550 to 1600 kHz. His receiver has a 50-ohm antenna-line characteristic, so he would design the filter for a 50-ohm bilateral impedance. In order to find L_{1K} he would use

$$L_{1K} = \frac{0.318 \times 50}{1.6 - 0.55} = \frac{15.9}{1.05} = 15 \mu H$$

$$X_{L1K} = 89, \text{ and } \frac{L_{1K}}{2} = 7.5 \mu H \quad \text{(Eq. 26)}$$

where R is in ohms, f is in MHz, and L is in μH.

C_{1K} would be found by

$$C_{1K} = \frac{7.96 \times 1.05 \times 10^4}{0.55 \times 1.6} = \frac{83{,}580}{44}$$

$$= 1899 \text{ pF}$$

$$X_{C1K} = 89, \text{ and } 2C_{1K} = 3798 \text{ pF} \quad \text{(Eq. 27)}$$

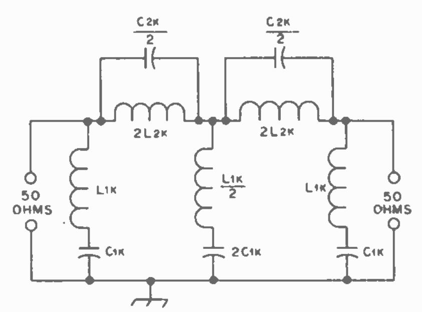
where R is in ohms, f is in MHz and C is in pF.

L_{2K} is obtained in a similar manner:

$$L_{2K} = \frac{.0796 \times 1.05 \times 50}{0.88} = \frac{4.179}{0.88}$$

$$= 4.7 \mu H$$

$$X_{L2K} = 56, \text{ and } 2L_{2K} = 9.4 \mu H \quad \text{(Eq. 28)}$$



$$L_{1K} = \frac{0.318\,R}{(f_2 - f_1)} \text{ and}$$

$$C_{1K} = \frac{7.96\,(f_2 - f_1)\,10^4}{f_1 f_2 R}$$

$$L_{2K} = \frac{.0796\,(f_2 - f_1)\,R}{f_1 f_2}$$

$$\text{and } C_{2K} = \frac{3.18\,(10^5)}{(f_2 - f_1)\,R}$$

f_1, f_2 = -3dB points of filter curve

Fig. 15 — Circuit details of a band-reject filter.

and C_{2K} is calculated by

$$C_{2K} = \frac{3.18\,(10^5)}{1.05 \times 50} = \frac{318{,}000}{52.5}$$

$$= 6057 \text{ pF}$$

$$\frac{X_{C2K}}{2} = 56, \text{ and } \frac{C_{2K}}{2} = 3028.5 \text{ pF} \qquad \text{(Eq. 29)}$$

The filter should be housed in an rf-tight metal box and connected to an earth ground. Although toroidal or pot-core inductors are ideal for the circuit, solenoid coils can be used with good results. The pot-core or toroid inductors provide their own shielding characteristics, but if solenoid coils are used, it is wise to shield the individual filter sections from one another.

Strip-Line Bandpass Filter

Vhf and uhf high-Q band-pass filters can be made by following strip-line design methods. It is less difficult to build a practical filter, in terms of power-handling capability, when using the strip-line concept as opposed to the helical resonator kind of filter treated earlier in the chapter. For use at 50 and 144 MHz, the quarter-wavelength strip-line filter of Fig. 16 is suitable. However, at 220 MHz and higher, it is difficult to realize a practical filter unless a half-wavelength format is adopted, as shown in Fig. 17. One problem with the quarter-wavelength style at 220 and above is that the box enclosure tends to become a resonant cavity by itself, and the unwanted resonance may fall in the desired filter pass band, thereby spoiling the characteristics of the filter. Furthermore, by using half-wavelength dimensions the effects of unwanted stray capacitance within the filter are less pronounced.

Although strip-line filters can be built for a variety of transmission-line impedances (they are transmission-line sections of some characteristic impedance), this discussion will center on an arbitrary 70-ohm characteristic. Detailed data on this subject can be found in *Reference Data for Radio Engineers*, p. 22-27, Fifth Edition, Howard W. Sams & Co., Inc.

A Design Example

A 2-meter band-pass filter is desired for use between the transmitter output and a 50-ohm transmission line. A metal box is available, and the dimensions are 1 inch high, 14 inches long, and 2 inches wide. Since h/H = 0.25 for a 70-ohm strip line the thickness of L3 (Fig. 16) will be 1/4 inch. The width of L3 must be W/H = 0.4. Since H is 1 inch, W will equal 0.4 inch (13/32 in. or 10.3 mm). L3 should be centered in the box, and the low-impedance end of L3 must be soldered securely to one end of the box. The housing should be made of aluminum, copper, or silver-plated brass. L3 should be made of the same material as the case.

The length of L3 can be found from

$$L3_{(inches)} = \frac{2808}{f_{MHz}} \times 0.65$$

$$= \frac{2808}{145} \times 0.65$$

$$= 12.58 \text{ inches} \qquad \text{(Eq. 30)}$$

where f is in MHz.

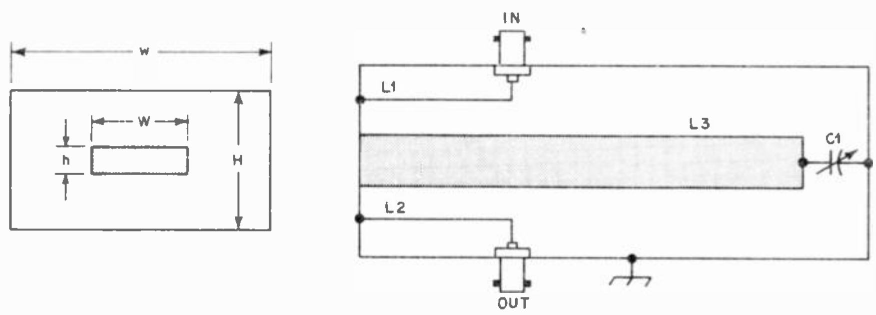
The length of coupling lines L1 and L2 will be 0.25 X L3, or 3.1 inches. This will permit ample coupling to L3. Insertion loss will depend on the coupling amount, as will the bandpass characteristics. L1 and L2 can be moved closer to or farther from L3 to obtain the desired coupling characteristic. For most applications an insertion loss of 0.1 to 0.5 dB can be used.

Capacitor C1 of Fig. 16 should have wide spacing for power levels above 25 watts. Air-dielectric capacitors are preferable in the interest of maintaining high Q and minimum filter losses. In the example just given, C1 should have a maximum capacitance of 10 pF, and could be made from two rectangular plates of copper, each being approximately 3/4 X 1-1/2 inches in size. Copper disks could be used if a box with greater H was used. A small double-spaced 10- or 15-pF air variable could be used at C1 for low-power applications. With the dimensions given in this example, a *maximum* power of 75 watts is suggested.

Fig. 17 shows a half-wave strip-line filter and contains equations for a 70-ohm design. Details for C1 and the adjustment of L1 and L2 are the same as for the quarter-wavelength filter.

Fig. 18 illustrates how a 50-MHz strip-line filter can be compressed to make it a more practical size. The design rules are the same as for the filter in Fig. 16.

Strip-line bandpass filters can be tuned properly by inserting an SWR indicator between the filter input and the transmitter. The filter is tuned for minimum SWR.



$$\frac{h}{H} = 0.25$$

$$\frac{W}{H} = 0.4$$

$$w \geqslant 2H$$

$$L3 = \frac{2808}{f_{MHz}} \times 0.65 \text{ (inches)}$$

$$L1, L2 = 0.25 \times L3 \text{ (inches)}$$

Fig. 16 — Design details for a quarter-wavelength transmission-line type of filter (strip-line filter).

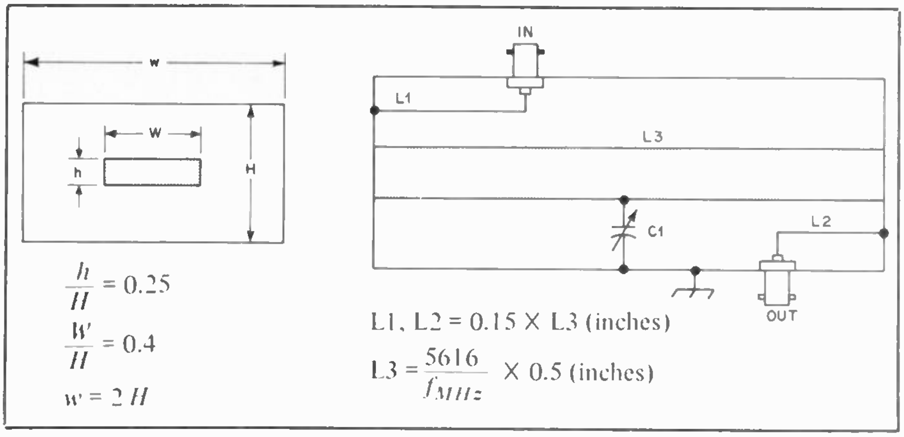


Fig. 17 — Configuration and design data for a half-wavelength transmission-line filter.

Half-Wave Harmonic Filter

An effective means by which to attenuate harmonic energy at the output of a transmitter, mixer, amplifier stage, or similar, is to place a half-wave filter in the output line. For the sake of simplicity in defining such a filter, it can be thought of as two pi networks in series, or two low-pass filters in series. Unlike the more common low-pass TVI filter, half-wave filters must be designed for the band of operation. The classic TVI low-pass filter is suitable for all amateur bands below its cutoff frequency. The latter is typically 40 MHz.

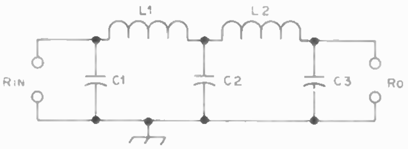
A half-wave filter is normally designed for a loaded Q of 1, thereby assuring a broad frequency response. With a Q of 1, X_L and X_c will always be equal to the impedance for which the filter is designed. That is, if a filter is needed for a 50-ohm transmission line, and the loaded Q is 1, X will be 50. The center capacitor of a half-wave filter will have an X_c of one half the end capacitors because two like impedances are joined at that point. Thus, for a 50-ohm filter the X_c for the center capacitor will be 25.

Table 5 provides L, C and f data for half-wave filters from 160 through 6 meters for use in a 50-ohm line. In each case f is chosen for the high-frequency limit of the band; e.g., the 160-meter filter uses 2.0 MHz as f.

Table 6 contains similar data, but is based on a 75-ohm line impedance. It will be seen that nonstandard capacitor values are obtained from the design when one converts X into μH and pF. For the most part, the nearest standard capacitor value will suffice. If precise design is wanted, trimmer capacitors can be used at each C point, or standard-value fixed capacitors can be used in parallel with low-value trimmers for on-the-nose adjustment.

Table 6

$Q_L = 1$
$R_{in}, R_o = 75$ ohms
$X_L = 75$ ohms
$X_{C1,C3} = 75$ ohms
$X_{C2} = 37.5$ ohms



BAND (METERS)	L1, L2 (μH)	C1, C3 (pF)	C2 (pF)
160	5.97	1061	2123
80 (cw)	3.22	573	1147
75 (phone)	2.98	530	1061
40	1.63	290	581
20	0.82	147	294
15	0.559	99	198
10	0.402	71.4	142
6	0.236	42.1	84.2

Low-Pass TVI Filter

As a precautionary measure, at least, many amateurs install a low-pass TVI filter in the transmitter output line. Such a filter must pass without significant attenuation all of the amateur frequencies below the filter cutoff frequency. The latter is generally near 40 MHz. Energy above the cutoff frequency should be attenuated significantly.

The subject of low-pass TVI filter design and application was treated in considerable depth by Grammer, W1DF, in his three-part series in *QST* for February, March and April of 1950 ("Eliminating TVI with Low-Pass Filters"). The papers represent excellent reference material for the designer.

A Practical Filter

The circuit of Fig. 19 is based on the "insertion-loss design" concept rather than on an image-parameter one. It is described more completely by the designer, Welsh (WB6HRM), in his January, 1966, *QST* article, "An Effective Low-Pass Filter."

The frequencies of maximum attenuation are 40.5, 47.5 and 78.1 MHz. The characteristic impedance of the unit is 50 ohms. Overall, the attenuation level of energy above 40 MHz is 50 dB or greater.

The coils are made from no. 14 enameled copper wire, and are formed on a 1/2-inch dia. mandrel. L1 has 8 turns while L2 and L3 each have 6 turns. After the coils are formed, the capacitors are soldered across them, and the parallel branches are initially tuned to resonance by adjusting the turns spacing until a grid-dip meter indicates resonance at the frequencies shown in Fig. 19. The coil/capacitor assemblies are then mounted in the chassis (individually), and the resonant frequency is checked again. Finally, the shunt capacitors are mounted and soldered in.

The filter is housed in a 5 X 3 X 2-inch aluminum Minibox. Aluminum shields are used to provide isolation between filter sections. Each shield is secured to the Minibox at eight places to assure isolation and prevent "hot spots." The paint is removed from along the edges of the cover to insure good metallic contact between the overlapping flanges, when the unit is assembled.

If silver-mica, 600-volt capacitors are used in the filter, power capability will be 50 W at 28 MHz, 150 W at 21 MHz, and 300 W at 14 MHz and below. By replacing the silver-mica capacitors with APC style trimmers of appropriate value, the power capability can be extended considerably.

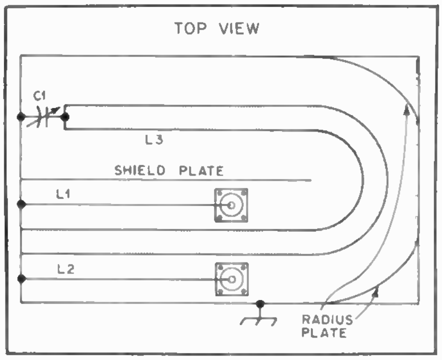


Fig. 18 — Structural details for compacting a 50-MHz strip-line filter.

Table 7

PART & TYPE NUMBERS	MIN. 3 dB BW @ 25 C (kHz)	MIN. 4 dB BW OTR (kHz)	MAX. 60 dB BW @ 25 C (kHz)	MAX. 60 dB BW OTR (kHz)	MAX. RV @ 25 C (dB)	MAX. RV OTR (dB)	MAX. IL @ 25 C (dB)	MAX. IL OTR (dB)	MIN. 60 dB SBR (kHz)	S & L ±5% (Ω)	RES. CAP. ±5% (pF)
526-9689-010 F455FD-04	0.375	0.375	3.5	4.0	3.0	4.0	10.0	12.0	445-F60L F60H-465	2,000	350 350
526-9690-010 F455FD-12	1.2	1.2	8.7	9.5	3.0	4.0	10.0	12.0	445-F60L F60H-465	2,000	350 350
F526-9691-010 F455FD-19	1.9	1.9	5.4	5.9	3.0	4.0	10.0	12.0	445-F60L F60H-465	2,000	330 330
526-9692-010 F455FD-25	2.5	2.5	6.5	7.0	3.0	4.0	10.0	12.0	445-F60L F60H-465	2,000	510 510
526-9693-010 F455FD-29	2.9	2.9	7.0	8.0	3.0	4.0	10.0	12.0	445-F60L F60H-465	2,000	510 510
526-9694-010 F455FD-38	3.8	3.8	9.0	10.0	3.0	4.0	10.0	12.0	445-F60L F60H-465	2,000	1000 1000
526-9695-010 F455FD-58	5.8	5.8	14.0	15.0	3.0	4.0	10.0	12.0	445-F60L F60H-465	2,000	1100 1100

OTR = Operating Temperature Range, RV = Ripple Voltage, IL = Insertion Loss, SBR = Stop Band Range, S & L = Source and Load.

Courtesy of Collins Radio Co.

Table 8

Standard 10.7-MHz Models

13 kHz BANDWIDTH	15 kHz BANDWIDTH	30 kHz BANDWIDTH	NO. POLES (RESONATORS)	DIMENSIONS W X L X H (INCHES)
2194F	2195F	2196F	2	0.185 X 0.432 X 0.450
1463	1464	1465	4	0.47 X 0.59 X 0.59
1466	1467	1468	4	0.61 X 0.89 X 0.22
1469	1470	1471	6	0.47 X 0.59 X 0.59
1472	1473	1474	6	0.61 X 1.31 X 0.22
1475	1476	1477	8	0.47 X 0.73 X 0.59
1478	1479	1480	8	0.61 X 1.75 X 0.22
1433	1434	1435	8	1.02 X 2.39 X 0.52

Models 1433-35 interchange with commonly used conventional crystal filters. PTI also offers standard models at many other popular frequencies including 5.26, 6.46, 8.0, 16.9, 20.0, 21.4, 30.0, 32.0, 75 and 156 MHz.

Courtesy of Piezo Technology, Inc.

Table 9

APPLICATION	SSB TRAN.	SSB REC.	CW OR DIGITAL DATA	A-M	A-M	CW	FM
FILTER TYPE	XF-9A	XF-9B	XF-9NB	XF-9C	XF-9D	XF-9M	XF-9E
NO. OF CRYSTALS	5	8	8	8	8	4	8
6-dB BANDWIDTH	2.5 kHz	2.4 kHz	0.5 kHz	3.75 kHz	5.0 kHz	0.5 kHz	12 kHz
PASSBAND RIPPLE	< 1 dB	< 2 dB	< 0.5 dB	< 2 dB	< 2 dB	< 1 dB	< 2 dB
INSERTION LOSS	< 3 dB	< 3.5 dB	< 6.5 dB	< 3.5 dB	< 3.5 dB	< 5 dB	< 3 dB
TERM. IMPEDANCE	500 Ω	500 Ω	500 Ω	500 Ω	500 Ω	500 Ω	1200 Ω
RIPPLE CAPACITORS	30 pF	30 pF	30 pF	30 pF	30 pF	30 pF	30 pF
SHAPE FACTOR	6:50 dB 1.7	6:60 dB 1.8	6:60 dB 2.2	6:60 dB 1.8	6:60 dB 1.8	6:60 dB 4.4	6:60 dB 1.8
STOP-BAND ATTEN.	> 45 dB	> 100 dB	> 90 dB	> 100 dB	> 100 dB	> 90 dB	> 90 dB

Courtesy of Spectrum International

Table 10

APPLICATION	FM	FM	FM	FM	FM	FM	FM
FILTER TYPE	XF-102	XF-107A	XF-107B	XF-107C	XF-107D	XF-107E	XM-107 SO-4
CENTER FREQ.	10.7 MHz	10.7 MHz	10.7 MHz	10.7 MHz	10.7 MHz	10.7 MHz	10.7 MHz
NO. OF CRYSTALS	2	8	8	8	8	8	4
6-dB BANDWIDTH	14 kHz	14 kHz	16 kHz	32 kHz	38 kHz	42 kHz	14 kHz
PASSBAND RIPPLE	< 2 dB	< 2 dB	< 2 dB	< 2 dB	< 2 dB	< 2 dB	< 1 dB
INSERTION LOSS	< 1.5 dB	< 3.5 dB	< 3.5 dB	< 4.5 dB	< 4.5 dB	< 4.5 dB	< 3 dB
TERM. IMPEDANCE	2500 Ω	820 Ω	910 Ω	2000 Ω	2700 Ω	3000 Ω	910 Ω
RIPPLE CAPACITORS	—	25 pF	25 pF	25 pF	25 pF	25 pF	35 pF
SHAPE FACTOR	20 dB@50 kHz 30 dB@80 kHz	6:70 dB 2.2	6:70 dB 2.2	6:70 dB 2.2	6:70 dB 2.2	6:70 dB 2.0	40 dB @ 42 kHz
ULTIMATE ATTEN.	> 30 dB	> 90 dB	> 90 dB	> 90 dB	> 90 dB	> 90 dB	> 60 dB

Courtesy of Spectrum International

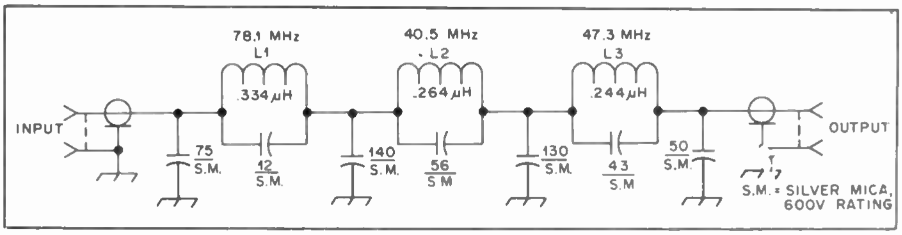


Fig. 19 – Low-pass filter suitable for use with an amateur transmitter.

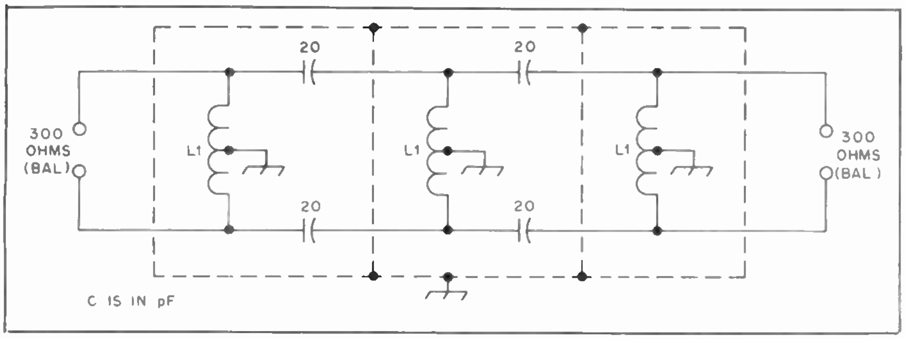


Fig. 20 – Circuit diagram of a 300-ohm balanced high-pass filter for use at the input of a TV receiver.

High-Pass TVI Filter

When a TV set and an amateur station are relatively close to one another, the strong amateur signal can overload the TV set front end (fundamental overloading). Since the nature of the interference is not harmonically related, nothing can be done at the transmitter to cure the problem. One cure is to install a trap, tuned to the amateur station frequency, at the input of the TV tuner. This was discussed earlier in the chapter.

A more universal suppressor of fundamental overloading is the high-pass TVI filter. The cutoff frequency is chosen below the lowest TV channel, but higher than 30 MHz. An ideal filter would pass all of the TV signals without attenuation, but would significantly reduce the signal energy below the filter cutoff frequency. The filter shown in Fig. 20 will do the job quite well. The circuit was designed by Bird, W3JHE. It is for use in 300-ohm TV ribbon line, and should be mounted as near the TV-tuner input terminals as possible. The filter should be contained in a

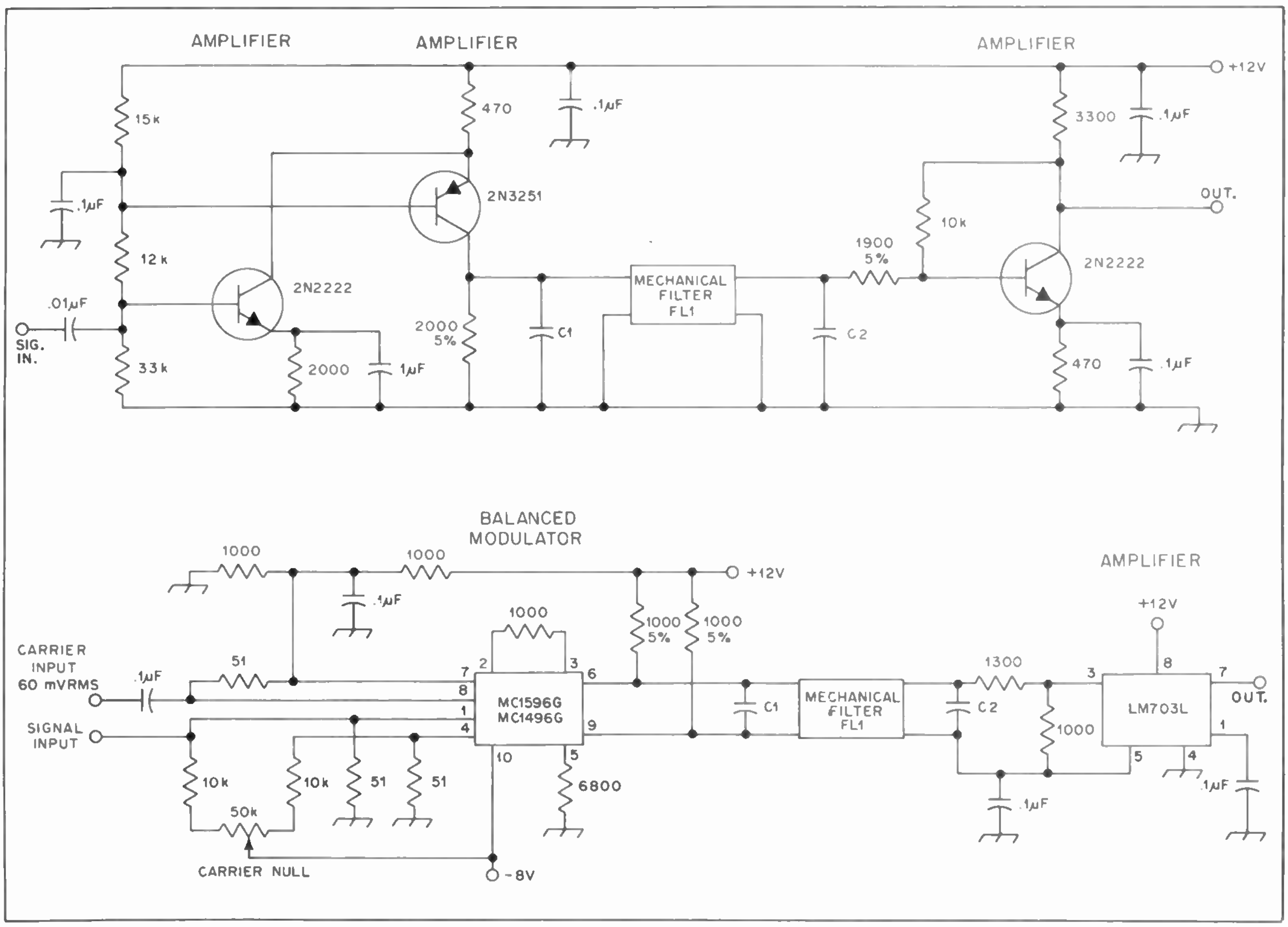


Fig. 21 – Circuit examples for mechanical filters.

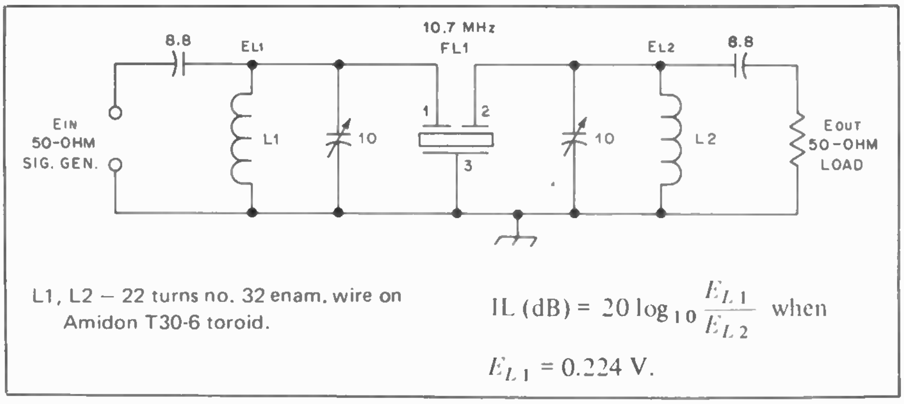


Fig. 22 — Test circuit for a monolithic filter.

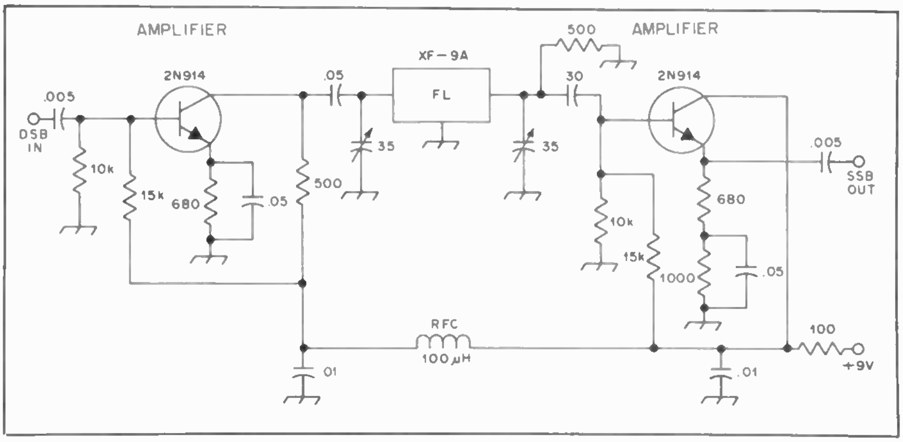


Fig. 23 — Typical circuit for using a 9-MHz crystal filter in an ssb generator. Other types of npn transistors are suitable provided f_T is 50 MHz or greater, and small-signal beta is 10 or more. Base-bias resistors may require different ohmic values if transistor substitutions are made.

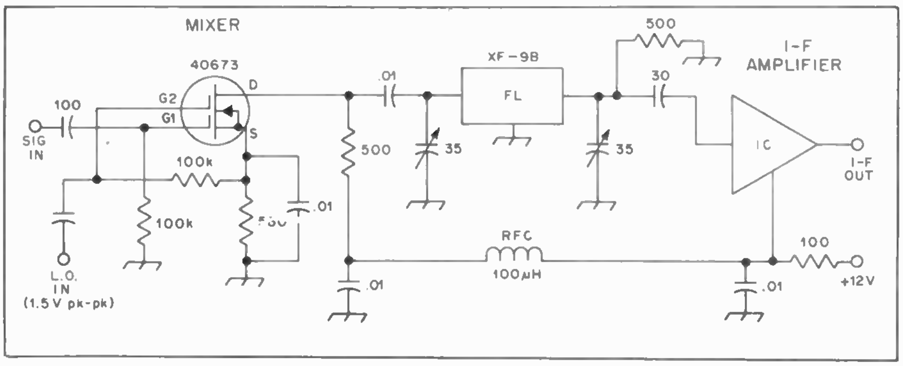


Fig. 24 — Circuit example of a 9-MHz crystal filter as used in a receiver.

shielded box, with partitions included between filter sections, as indicated by the dashed lines in Fig. 20. The shield box should be connected to an earth ground for best results. All of the capacitors are ceramic units.

L1 consists of 40 turns of no. 30 enam. wire, close-wound on a 1/8-inch dia. low-loss form. L2 has 22 turns of no. 30 wire, close-wound, on a 1/8-inch dia. form. Each coil has a center tap which is grounded. *Warning:* Do not use a direct ground on an ac-dc chassis. Complete the ground connection through a .001-μF, 1000-V disk-ceramic capacitor.

Mechanical Filters

A mechanical filter is a *mechanically* resonant device which receives an i-f signal, then converts it into mechanical vibrations, rejects unwanted frequencies, and converts the mechanical impulses back to i-f energy.

Mechanical filters are used mainly in receiver i-f systems and in ssb generators. They are available in a variety of 3-dB bandwidths and case styles.

Table 7 gives a partial listing of the filters available from Collins Radio Co. The filters should not have dc flowing through them if at all possible. If, however, dc must pass through them, dc current should not exceed 3 mA, and dc voltage must not be greater than 100. The filters listed in the table are approximately 2.53 inches long, 0.59 inch high, and 0.56 inch wide. Electrical connections are made to pins located on each end of the housing, on the bottom surface of the filter.

Fig. 21 shows typical applications for the filters in solid-state circuits.

Monolithic Crystal Filters

Monolithic filters contain two or more resonators which are formed on a single quartz plate. The resonators are coupled mechanically to form a band-pass crystal filter. This kind of construction eliminates the need for numerous separate components and interconnections as is common to the more conventional type of crystal filter. These filters can be used singly or connected in tandem. The latter configuration leads to improved stop-band attenuation, and greater attenuation of unwanted modes.

Fig. 22 shows the circuit of a monolithic filter as it would be installed in a test fixture. Filters of this variety are used extensively in amateur fm receivers.

Table 8 lists some of the filters for 10.7 MHz which are useful in amateur circuits. This information is through the courtesy of Piezo Technology, Inc., Orlando, FL.

Crystal Lattice Filters

Amateur and commercial designers use crystal filters of various pole numbers for obtaining i-f selectivity characteristics of some desired values in receiving equipment. As is true of mechanical filters, crystal filters can be employed in ssb generators to remove the unwanted sideband.

Crystal filters can take the form of a single crystal with a panel-accessed phasing control, or can be configured as simple half-lattice, full-lattice, or cascaded full-lattice types. The greater the number of resonators or poles, the more desirable the shape factor (skirt selectivity).

Among the popular crystal filters used by amateurs are those with an f_o of 9 MHz. Filters designed for that frequency are chosen by many designers to permit the building of single-conversion receivers. That is, a 9-MHz i-f is less likely to enhance image problems in a single-conversion scheme than would be the case with a lower i-f such as 455 kHz. With the latter, the usual practice is to employ a double-conversion circuit and use the 455-kHz frequency in the second i-f part of the receiver. The more conversions employed, the greater the chance for spurious responses (birdies) brought about by the various mixing circuits.

Many builders of vhf fm receivers use 10.7-MHz i-f filters of the monolithic or crystal types. In connection with the foregoing filters, some amateurs employ crystal discriminators in their fm receivers.

Table 9 lists some of the available 9-MHz crystal filters which are sold by Spectrum International, Box 1084, Concord, MA 01742. The same company provides the 10.7-MHz filters specified in Table 10.

Fig. 23 shows a typical circuit for using an XF-9A filter for generating an ssb signal. A possible application for the XF-9B filter in a receiver is illustrated in Fig. 24.

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Chapter 7

Antennas and Feed Systems

Most antenna designs are based on the free-space half-wavelength radiator equation: l (feet) = 492 ÷ f (MHz), or l (meters) = 150 ÷ f (MHz), but end effects caused by the presence of insulating hardware make it necessary to shorten the wire or tubing radiator – roughly 5 percent for wire antennas with a typical length/diameter ratio of 10,000. This is a generalized rule for frequencies up to 30 MHz where no. 12 wire is used. With the correction factor applied the equation becomes l (feet) = 492 X 0.95 ÷ f (MHz) = 467.4 ÷ f (MHz). The number 467.4 has been rounded off in most textbooks to 468. The conversion for meters is l (meters) = 150 X 0.95 ÷ f (MHz) = 142.5 ÷ f (MHz). It can be seen from this that a 40-meter half-wavelength dipole would be 65.9 feet or 20 meters long, assuming no. 12 wire was employed, and that the antenna was at least one half wavelength above ground and an equal distance or greater from nearby conductive objects. The greater the conductor diameter, the shorter the radiator must be.

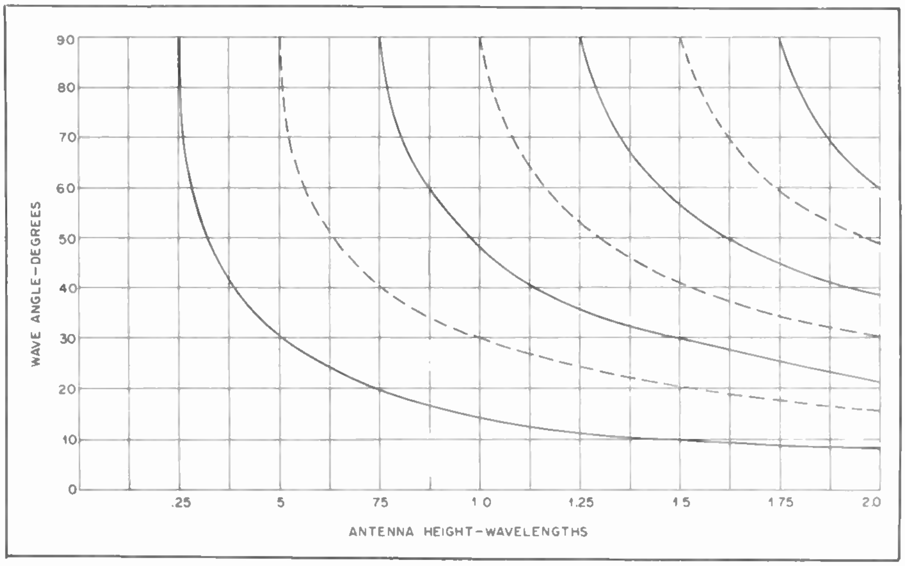


Fig. 1 – Angles at which nulls and maxima (factor = 2) in the ground-reflection factor appear for antenna heights up to two wavelengths. The solid lines are maxima, dashed lines nulls, for all horizontal antennas and for vertical antennas having a length equal to an even multiple of one-half wavelength. For vertical antennas an odd number of half waves long, the dashed lines are maxima and the solid lines nulls. For example, if it is desired to have the ground reflection give maximum reinforcement of the direct ray from a horizontal antenna at a 20-degree wave angle (angle of radiation) the antenna height should be 0.75 wavelength. The same height will give a null at 42 degrees and a second maximum at 90 degrees. Values may also be determined from the trigonometric relationship θ = arc sin $(A/4h)$, where θ is the wave angle and h is the antenna height expressed in wavelengths. For the first maxima (horizontal antennas), A has a value of 1; for the first null A has a value of 2, for the second maxima 3, for the second null 4, and so on.

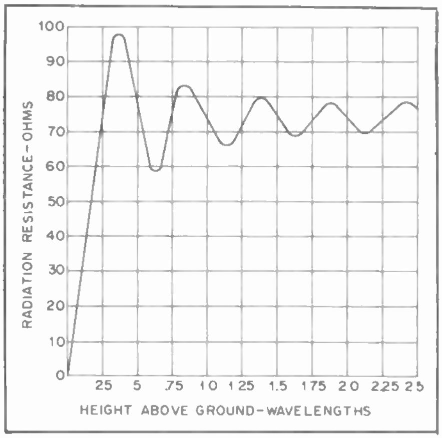


Fig. 2 – Variation in radiation resistance of a horizontal half-wave antenna with height above perfectly conducting ground.

The equation for a quarter-wavelength wire antenna is l (feet) = 234 ÷ f (MHz). Most amateur vertical quarter-wavelength antennas are made from conductors of large cross-sectional area to assure strength and rigidity. This is a departure from the 10,000:1 ratio, and calls for shortening the radiator physically to establish resonance.

Height Above Ground

For DX work the wave angle becomes a significant factor. The lower antenna wave angles are best for long-distance communications. Fig. 1 shows the wave angle versus height above a perfectly conductive earth. The earth "mirror" does not exist at the plane surface, but is usually a few feet below the surface. It is dependent upon the operating frequency and quality (conductivity) of the earth below the antenna.

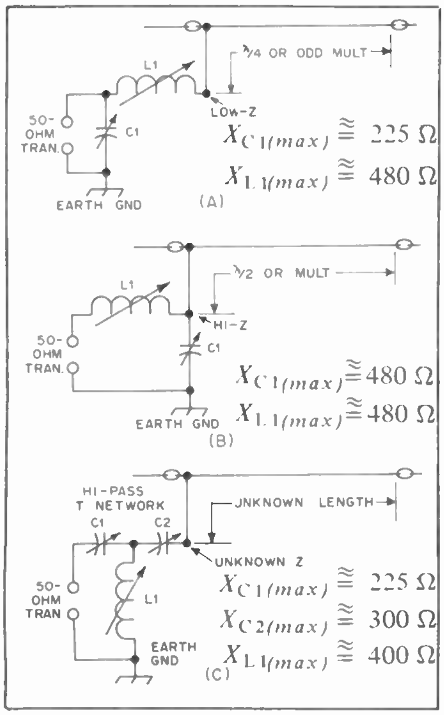


Fig. 3 — Various kinds of single-wire antennas showing methods for matching the antennas to the transmitter. Reactance values are supplied for the *L* and *C* components of the transmatches.

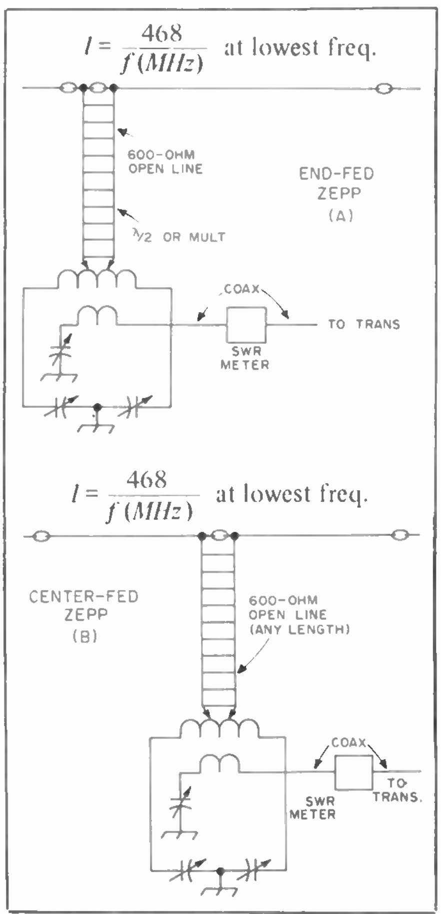


Fig. 4 — Details of end- and center-fed Zepp antennas which can be used for all-band mf and hf work.

Radiation Resistance

The radiation resistance of an antenna (in ohms) becomes lower as the antenna is brought closer to ground, assuming that the power delivered to the antenna remains constant. The current flow in an antenna increases as the system is moved closer to ground, and therefore the radiation resistance becomes lower. Thus, a center-fed half-wavelength dipole would have an impedance of approximately 70 ohms if the antenna was one half wavelength above ground. At 0.12 wavelength above ground the same dipole would exhibit a radiation resistance of roughly 30 ohms. Fig. 2 illustrates the relationship between height and radiation resistance.

Single-Wire Antennas

Single-wire end-fed antennas are popular for portable and fixed-station multiband use. A long span of wire does not constitute a classic "long-wire antenna." The latter must contain *several* wavelengths of wire to qualify for that title. The misnomer is often applied to any length of wire which is end fed. Fig. 3 shows three types of end-fed wire antennas with suitable Transmatches for feeding them from a 50- or 75-ohm transmitter. An SWR indicator is placed between the transmitter and the coaxial feeder to the Transmatch. The latter is adjusted to provide an SWR of 1.

Fig. 4A provides details for an alternative feed method which can be applied to a single-wire antenna. This antenna, the end-fed Zepp, and the center-fed version of Fig. 4B are suitable as station antennas from 160 through 10 meters.

Dipoles

Simple dipole antennas are among the more common amateur types in use. In the classic example a half-wavelength dipole would be parallel to the earth and at least one half wavelength above ground. This would assure the desired bidirectional radiation pattern and provide a predictable radiation resistance. Dipoles erected very near to ground exhibit little (if any) directivity, and are high-angle radiators, making them unsuitable for most DX work.

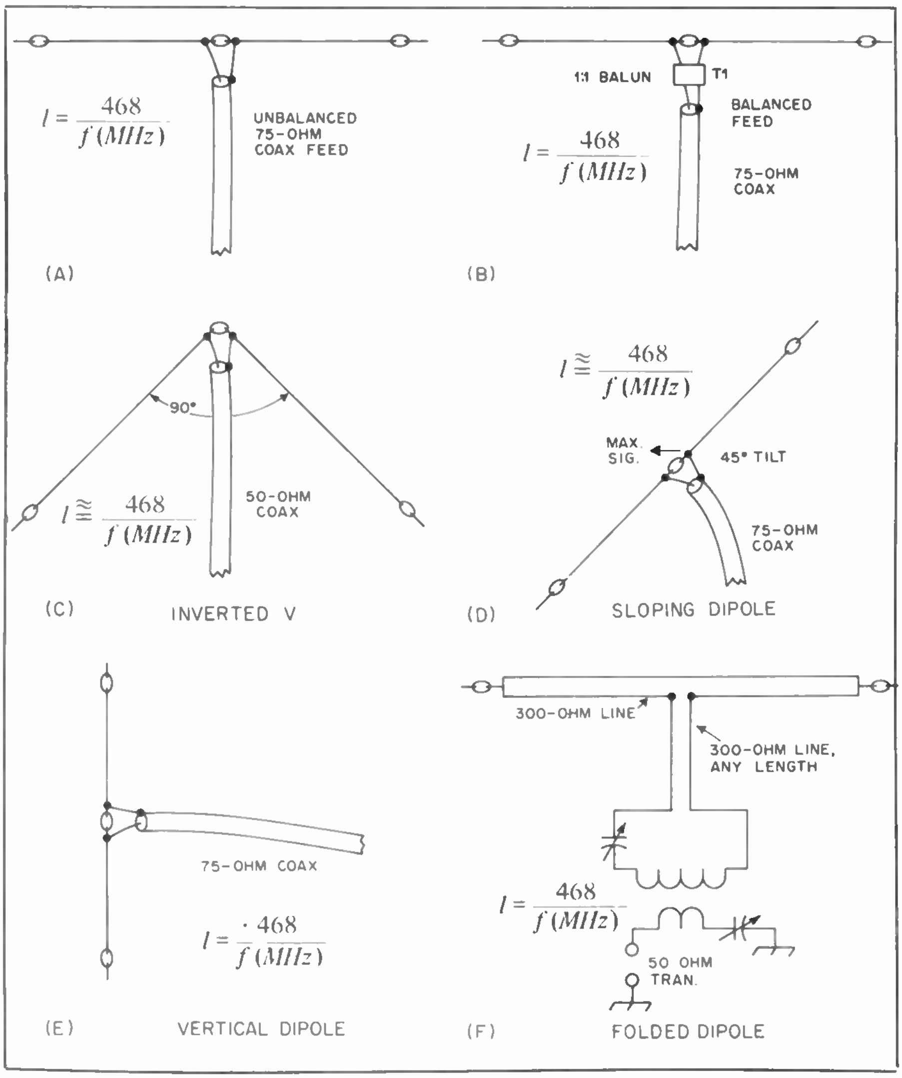


Fig. 5 — Illustrations of several types of simple dipole antennas.

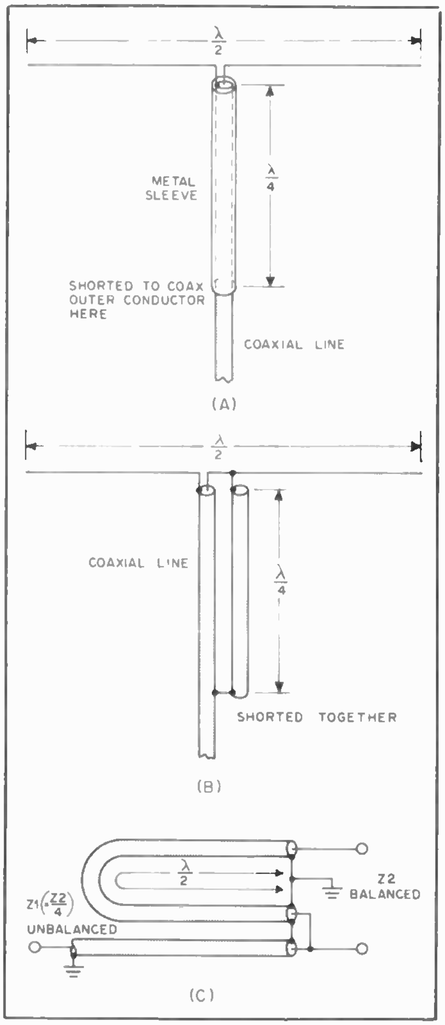


Fig. 6 — Radiator with coaxial feed (A) and methods of preventing unbalanced currents from flowing on the outside of the transmission line (B and C). The half-wave phasing section shown at D is used for coupling between an unbalanced and a balanced circuit when a 4-to-1 impedance ratio is desired or can be accepted.

Many forms of the simple dipole can be employed to make them fit into available space. Fig. 5A shows a classic dipole.

Balanced Dipole Feed

Although it is not essential to use balanced feed to a dipole, some operators prefer that method. A balun (balanced-to-unbalanced) transformer can be used to convert from coaxial cable to a balanced feed line or feed point. This is illustrated in Fig. 5B. Balanced feed is shown also at F in Fig. 5.

Other Dipole Types

Vertical and sloping dipoles are useful in DX work, and they're often employed in situations where limited antenna space exists. Fig. 5D illustrates a "sloper" whose maximum directivity is indicated by the arrow. The directional characteristic is especially significant when a metal tower or mast is used as an antenna support. Therefore, if maximum radiation in a southerly direction is desired the sloping dipole should be erected on the south side of the tower.

A vertical dipole is shown in Fig. 5E. When erected in the clear the radiation pattern is essentially omnidirectional. The results obtained will be similar to those from a ground-plane vertical antenna.

A drooping doublet — sometimes called an *inverted V* — can be erected at sites where a straight horizontal dipole is too long for the space available. The radiation pattern of such an antenna (Fig. 5C) will be omnidirectional on the frequency for which it is one half wavelength long, provided the apex angle between the wires is between 90 and 110 degrees. The feed impedance will be on the order of 50 ohms. A nonconductive support pole is recommended (wooden mast), but many amateurs have had good results when using a metal tower to support an inverted V. Multiband use of an inverted V is possible if tuned feeders are used. However, results are usually poor at more than one octave above that for which the antenna is a half wavelength. This problem results from the legs becoming electrically long as the operating frequency is increased, and high-angle radiation becomes predominant.

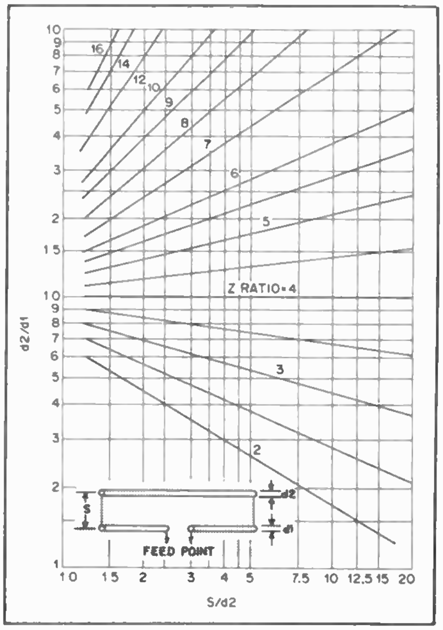


Fig. 7 — Impedance step-up ratio for the two-conductor folded dipole as a function of conductor diameters and spacing. Dimensions d1, d2 and S are shown in the inset drawing. The step-up ratio, r, may also be determined from

$$r = \left(1 + \frac{\log \frac{2S}{d1}}{\log \frac{2S}{d2}}\right)^2$$

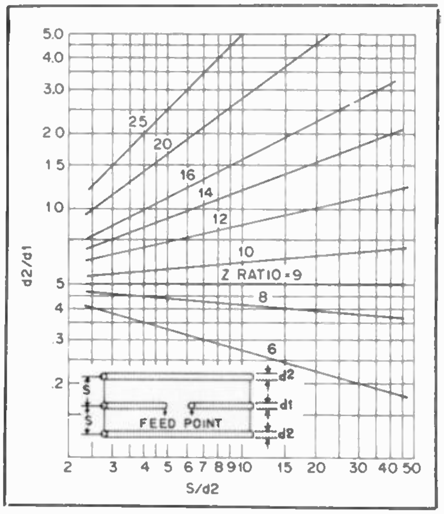


Fig. 8 — Impedance step-up ratio for the three-conductor folded dipole. The conductors that are not directly driven must have the same diameter, but this diameter need not be the same as that of the driven conductor. Dimensions are indicated in the inset.

Experimental adjustment of the dipole length is often necessary because of ground effects and cancellation characteristics caused by the dipole ends being in fairly close proximity to one another. If the voltage ends of the V are very near to ground the dipole may need to be shorter than 468 ÷ f (MHz). If it is placed high above ground the need may arise to lengthen the dipole beyond 468 ÷ f (MHz) because of the cancellation effects. Good DX results can be had on 160 and 80 meters with inverted Vs which have their centers high above ground — 60 feet or more on 80 meters, and 120 feet or greater on 160 meters.

Balancing Devices

Fig. 6 shows methods other than that of Fig. 5B to effect balanced dipole feed. The techniques at A and B prevent unbalanced currents from flowing on the outside of the transmission line. If a 4:1 impedance ratio is required the coaxial balun at C can be used, typically, to feed a 300-ohm folded dipole with 75-ohm unbalanced line, or a 200-ohm antenna can be fed with 50-ohm cable.

Multiconductor Dipoles

Three or more dipole conductors can be used to elevate the feed-point impedance of an antenna. The current is divided equally in the conductors. Therefore, the feeder looks into a higher impedance because it is delivering the

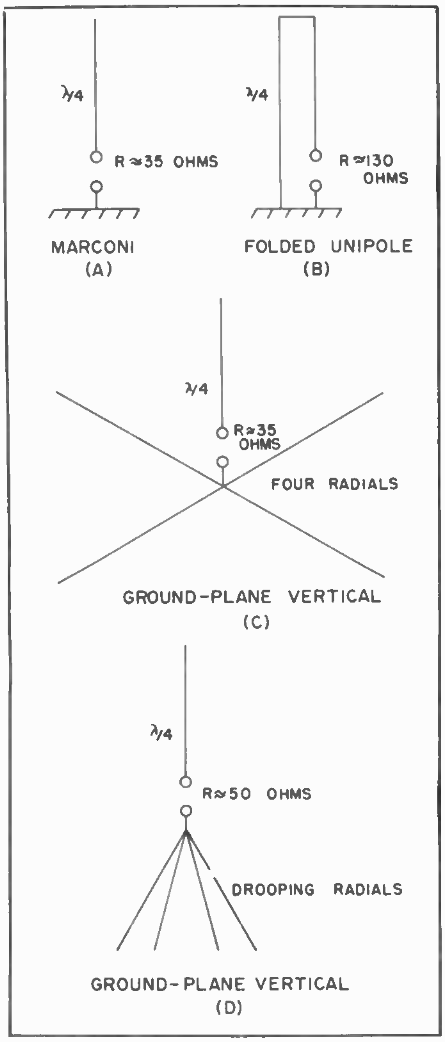


Fig. 9 — Various styles of quarter-wavelength vertical antennas showing approximate radiation resistances.

same power at lower current. Thus, a two-wire folded dipole has an impedance which is four times that of a simple two-wire doublet (Fig. 5A). As an example, assume that 500 watts of rf power are being fed to a simple dipole whose impedance is 75 ohms. The current will be

$$I = \sqrt{\frac{W}{R}} = \sqrt{\frac{500}{75}} = 2.58 \text{ A} \quad \text{(Eq. 1)}$$

where I = amperes, R = ohms, and W = watts.

If the antenna is changed to a two-wire folded dipole and is fed with 500 watts of power the current in each conductor becomes $2.58 \div 2$, which equals 1.29 A. The new feed impedance becomes:

$$R = \frac{W}{I^2} = \frac{500}{1.29^2} = \frac{500}{1.66} = 300 \text{ ohms} \quad \text{(Eq. 2)}$$

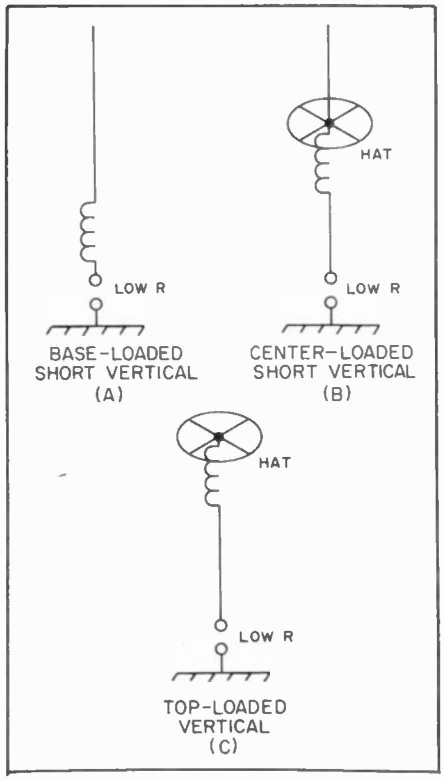


Fig. 10 — Illustrations of the three basic methods for loading short quarter-wavelength vertical antennas.

and for a 3-wire folded dipole the impedance would be $500 \div 0.739 = 677$ ohms. The foregoing examples are based on dipole conductors of equal diameter and spacing.

When conductors of unequal diameter are used the division of current is not equal. The nomographs of Figs. 7 and 8 can be used to find the feed-impedance values of two- and three-conductor folded dipoles.

Table 1 shows overall lengths for straight horizontal wire dipoles. The dimensions are founded on the premise that the antenna is at least one half wavelength above ground and spaced well away from nearby unwanted conductors. At heights less than one half wavelength it may be necessary to alter the dimensions listed in the table.

Vertical Antennas

Fig. 9 illustrates various forms for a quarter-wavelength vertical antenna. Detailed information on these antennas is given in *The ARRL Antenna Book.*

The feed impedance will be dependent upon the quality of the ground system used with the vertical radiator. The conductive properties of the earth determine how effective the image half of the antenna will be. Radial-wire networks must be used to reduce losses

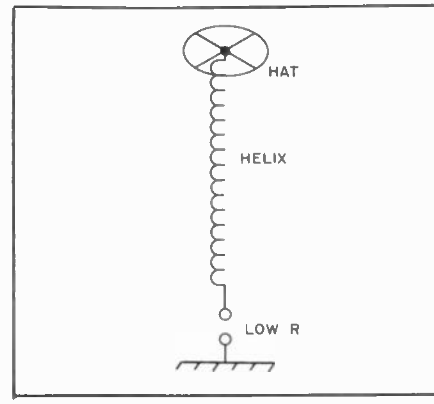


Fig. 11 — Short quarter-wavelength verticals can be wound helically and made broader in frequency response by adding a capacitance hat.

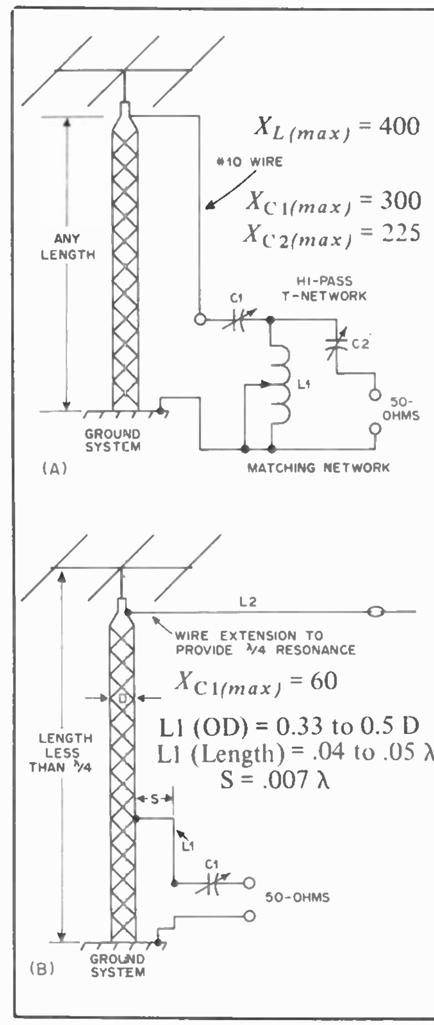


Fig. 12 — Method for matching a nonresonant tower vertical to a 50- or 75-ohm coaxial line (A). The bandwidth of the system will be narrow if the tower is less than one quarter wavelength. At B is shown a method for making a short nonresonant tower vertical one quarter wavelength long electrically. Gamma feed is used. The bandwidth of the system will be considerably greater than that of the antenna shown at A.

Table 1

FREQ. (MHZ)	FEET	+ INCHES	METERS
1.8	260	0	83.3
1.9	246	3-3/4	78.9
2.0	234	0	75.0
3.6	130	0	41.6
3.8	123	2	39.4
3.9	120	0	38.4
7.1	65	11	21.1
7.2	65	0	20.8
14.1	33	2-1/4	10.6
14.3	32	8-3/4	10.5
21.1	22	2	7.1
21.3	21	11-1/2	7.0
28.0	16	8-1/2	5.4
29.0	16	1-1/2	5.2

Dimensions in feet and inches, and also in meters, for half-wave simple or folded dipoles.

and to increase the effective radiation resistance. Radials can be laid upon the ground or buried. The constructor should use as many radials as practicable, and the length of each should be as great as possible for best results. However, in city lots it is often necessary to use very short radials, and good results can be had with the latter in many instances. A quarter-wavelength vertical antenna can be considered as one half of a dipole, with the ground system serving as the image half of the antenna.

Short Verticals

The electrical length of vertical antennas can be increased by adding series inductance to provide a resonant 90-degree (quarter wave) radiator. The loading coil should be as lossless as possible (large conductor size on a low-loss form). A large capacitance hat is recommended above the coil to reduce the amount of inductance required to effect resonance. Fig. 10 illustrates various short vertical-radiator forms.

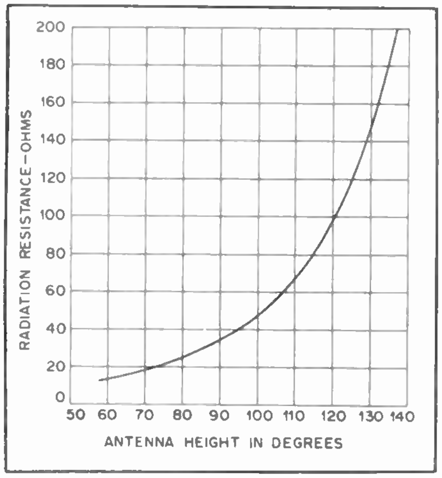


Fig. 13 – Radiation resistance vs. free-space antenna height in electrical degrees for a vertical antenna over perfectly conducting ground, or over a highly conducting ground plane. This curve also may be used for center-fed antennas (in free space) by multiplying the radiation resistance by two; the height in this case is half the actual antenna length.

Helically Wound Verticals

Continuously wound short verticals are effective as physically short radiators. It is necessary to construct a solenoid winding of approximately one half wavelength of wire to obtain a resonant quarter wavelength vertical. The wire should be wound with a uniform pitch on a low-loss form such as varnished bamboo or fiberglass. A capacitance hat is highly recommended to prevent destructive corona at the high-impedance end of the helix, and to minimize the number of coil turns needed to establish resonance. The "hat" lowers the antenna Q and prevents corona. It also increases the antenna bandwidth and radiation resistance. The latter will usually be less than 15 ohms. Fig. 11 shows a helically wound vertical.

Feeding Towers as Verticals

Towers which support beam antennas can be used as resonant or nonresonant vertical radiators (Fig. 12A). If guy wires are used they can be part of the system if desired, or they can be divorced from the tower by means of insulators.

Best results will be obtained if the tower plus the hf-band beam is made to look like a resonant quarter wavelength at the desired lower operating frequency – typically 80 or 160 meters. The beam antenna feed line and rotator cable need not be isolated, but should be dressed against the tower and routed from it at ground level. The top-mounted hf beam antenna, depending on its overall size, will have some effect as a capacitance hat. However, it is often necessary to add a long single wire extension from the top of the tower (horizontal or sloping) to establish quarter-wavelength resonance (Fig. 12B). The extender wire will do very little radiating, and polarization will be mainly vertical for the overall system. A two-band vertical system for 80 and 160 meters can be made from a tower in the 0- to 70-foot high class. A gamma section can be added to the side of the tower for each band. An 80-meter trap can be placed in the extender wire at the point where the tower and that portion of the wire comprises a resonant quarter wavelength. The wire beyond the trap is made an appropriate length to effect resonance on 160 meters. A 500-pF capacitor plus whatever inductance is necessary to provide a parallel-resonant trap will work fine for 75 or 80 meters. The trap should resonate in the part of the 75- or 80-meter band which will be

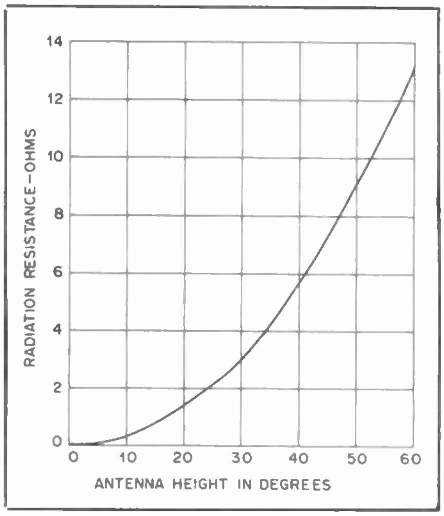


Fig. 14 – Same as Fig. 13 for heights below 60 degrees.

used (3.5-MHz resonance for the low end of 80, etc.). The 160-meter gamma section can be used for 80 meters also, provided a suitable matching network is switched in during 80-meter use.

Antennas of this variety are extremely useful for DX work, but require a good system of on-ground or buried radials, and the more the better. Tall grounded towers are not effective on 40 meters and higher, owing to the high-angle radiation which will prevail at those frequencies.

Fig. 13 shows a curve of antenna height in degrees versus radiation resistance for heights from 1/6 to 3/8 wavelength over a perfectly conducting ground. Fig. 14 gives similar data for antenna heights less than 60 degrees.

Wire-Antenna Materials

Soft-drawn copper wire can be used for dipoles and short end-fed antennas, and the usual gauge is no. 12 or 14. However, soft-drawn copper will stretch under the stress caused by wind, ice, or the sheer weight of the span of wire when long sections are erected. The stretching will change the resonant frequency, which is an undesirable event.

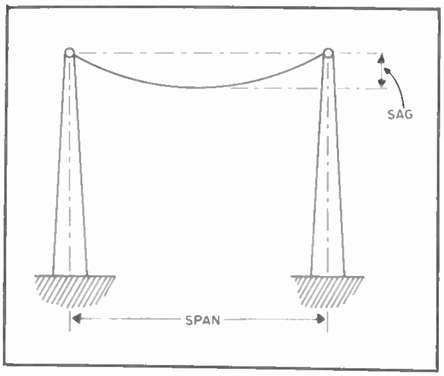


Fig. 15 – Illustration of the span and resultant sag of a long antenna made of wire.

Table 2
Stressed Antenna Wire

AMERICAN WIRE GAUGE	*RECOMMENDED TENSION[1] (POUNDS)*		*WEIGHT (POUNDS PER 1000 FEET)*	
	COPPER-CLAD STEEL[2]	*HARD-DRAWN COPPER*	*COPPER-CLAD STEEL[2]*	*HARD-DRAWN COPPER*
4	495	214	115.8	126
6	310	130	72.9	79.5
8	195	84	45.5	50
10	120	52	28.8	31.4
12	75	32	18.1	19.8
14	50	20	11.4	12.4
16	31	13	7.1	7.8
18	19	8	4.5	4.9
20	12	5	2.8	3.1

[1] Approximately one-tenth the breaking load. Might be increased 50 percent if end supports are firm and there is no danger of ice loading.

[2] "Copperweld," 40-percent copper.

Table 3
Dimensions of V and Rhombic Antennas for VHF use. Columns 1 and 2 are for V Designs. For Rhombics use 1, 3 and 4.

FREQ. (MHZ)	*SIDE A IN FT.*	*V ANGLE*	*VER-ALL B IN FT.*	*WIDTH C IN FT.*
50.5	58	60°	96.5	65.5
145	58	35°	109	39
28.7	68	70°	101.5	84
50.5	68	55°	106.5	70.6
145	68	35°	129	41
50.5	106	42°	192.5	91.5
145	106	35°	205	47.5
28.7	136	52°	237.5	133
50.5	136	37°	252.5	102

Stretching will weaken the wire and lead to eventual breaking.

It is better to employ hard-drawn copper, or copper-clad steel wire (Copperweld) when a long span of wire is anticipated. The longer the span the heavier the gauge, generally speaking. Wire sizes from no. 10 to no. 14 are suggested for use in the foregoing applications.

Whatever wire type is used, it is wise to select an insulated kind in the interest of minimizing oxidation and attendant deterioration. Most amateur antennas use enamel-coated wire, but plastic coatings are suitable also if the added weight of the insulation can be tolerated. Contrary to a misconception which exists among some amateurs, the insulation will not degrade antenna performance at mf and hf, but in the microwave region it will affect the wave propagation along the wire.

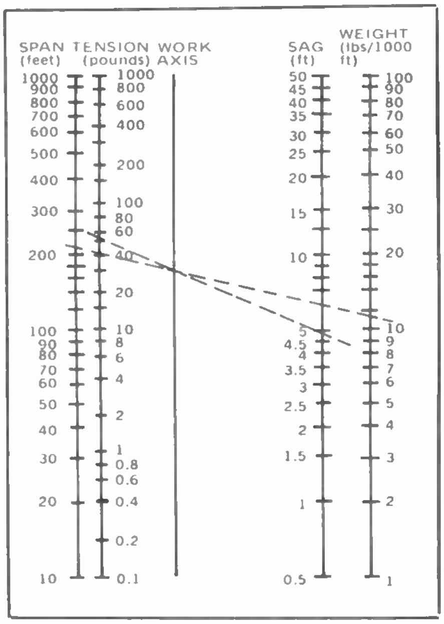


Fig. 16 — Nomograph for determining wire sag (courtesy K1AFR)

Aluminum clothesline, guy, and electric-fence wire can be utilized for some antenna designs, but it will stretch more readily from stress than copper will. Continuous flexing will break the wire quite quickly. Furthermore, aluminum corrodes more rapidly than copper, and it is difficult to effect good electrical connections without special soldering equipment.

Wire used for on-ground or buried radials should be made of copper and contain insulation. Soil acid and alkali materials will destroy metal quickly in some regions and this is particularly true when aluminum is used.

Fig. 15 shows how wire can sag when strung between two supports. Table 2 lists the maximum rated working tensions of hard-drawn and copper-clad steel wire of various sizes.

If the tension on a wire can be adjusted to a known value, the expected sag of the wire, as depicted in Fig. 15, may be determined in advance of installation with the aid of Table 2 and the nomograph of Fig. 16. Even though there may be no convenient method of determining the tension in pounds, calculation of the expected sag for practicable working tensions is often desirable. If the calculated sag is greater than allowable it may be reduced by any one or a combination of the following:

1) Providing additional supports thereby decreasing the span,

2) Increasing the tension in the wire if less than recommended,

3) Decreasing the size of the wire.

Conversely, if the sag in a wire of a particular installation is measured, the tension can be determined by reversing the procedure.

Instructions for Using the Nomograph

1) From Table 2 find the weight (pounds/1000 feet) for the particular wire size and material to be used.

2) Draw a line from the value obtained above, plotted on the weight axis, to the desired span (feet) on the span axis Fig. 16.

3) Choose an operating tension level (pounds) consistent with the values presented in Table 2 (preferably less than the recommended wire tension).

4) Construct a line from the tension value chosen, plotted on the tension axis, through the crossover point of the work axis and the original line constructed from Step 2, above, and continue this new line to the sag axis.

5) Read the sag (feet) on the sag axis.

Example:

Weight = 11 pounds/1000 feet.
Span = 210 feet.
Tension = 50 pounds.

Answer:

Sag = 4.7 feet.

Of course, these calculations do not take the weight of a feed line into account, if it is supported by the antenna wire.

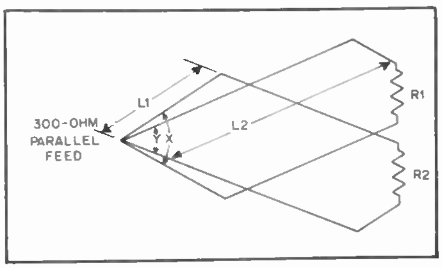


Fig. 17 — A 144-MHz rhombic with an estimated 27 dB gain over a dipole. The wires are all on the horizontal plane with the crossovers insulated.
L1 — 29.5 feet
L2 — 50.67 feet
X — 52.2°
Y — 37.7°
R1-2 — 660 ohms, total wattage should equal half the power output of the transmitter.
Height above ground — 12.29 feet
Elevation angle — 7.5°
Vertical beamwidth — 5.5°
Horizontal beamwidth — 8.5°

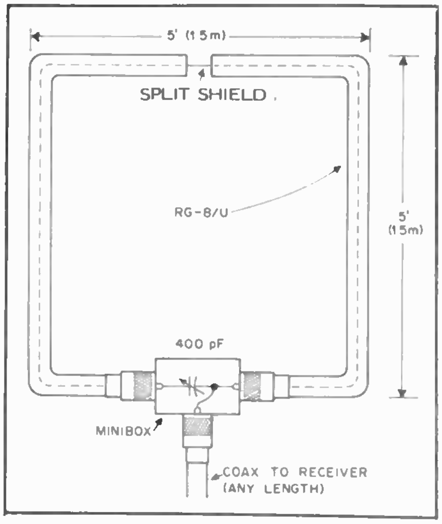


Fig. 18 — Schematic diagram of the loop antenna. The dimensions are not critical provided overall length of the loop element does not exceed approximately .04 wavelength. Small loops which are one-half or less the size of this one will prove useful where limited space is a consideration.

Large Wire Antennas

Rhomboid and V-shaped arrays of elements (rhombics and V beams in the more common vernacular) are useful for all frequencies of amateur interest from mf through uhf. The greater the number of wavelengths per leg, the greater the gain. A thorough treatment of rhombic antennas is given in *The ARRL Antenna Book* Thirteenth Edition, pages 169 through 177.

Table 3 lists dimensions for V beams beams and rhombics which can be applied in the 6- and 2-meter bands. The high-gain rhombic shown in Fig. 17 has a theoretical gain of 27 dB over a dipole, and is well suited to EME work on 144 MHz.

Low-Noise Receiving Antennas

At mf and the lower end of the hf spectrum the DX operator is frequently troubled by degradation of signal-to-noise ratio which results from high levels of man-made and atmospheric noise. The effect is particularly significant on 80 and 160 meters, where many DX transmitting antennas take the form of verticals. It is an established fact that vertical antennas respond more readily to noise than horizontal ones, so the problem is commonplace when the transmitting antenna is used also for receiving. Considerable relief from noise can be had by employment of *wave* antennas, more commonly known as *Beverage* antennas. Another solution, though not as effective, is the use of small loop antennas with which an rf preamplifier is used. An extension of the latter concept would be to include a sense antenna and phasing control to provide a cardioid response pattern, which could be used to reject noise from the direction of its source. None of the foregoing antennas are monuments to efficiency, so some preamplification is normally used ahead of the station receiver.

Beverage Antenna

The Beverage antenna is perhaps the earliest kind of effective receiving antenna in directional low-noise work. It was named after the inventor, H. H. Beverage. In its classic form this style of antenna would be regarded as a long transmission line of the open-wire kind, pointed toward the down-coming wave. It has a high amount of exposure to the horizontal component of an incoming

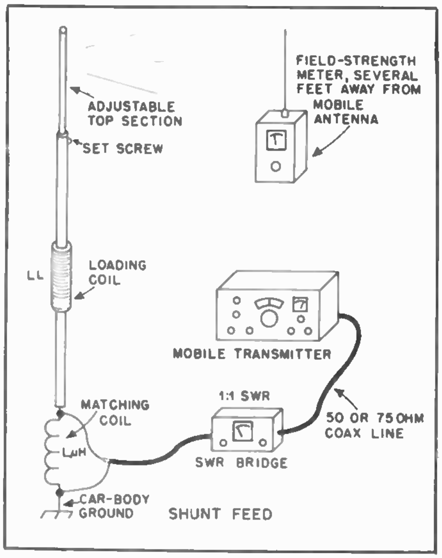


Fig. 20 — Mobile antenna with center loading. Resonance is determined by the combination of L_L and $L_{\mu H}$. Final resonance is set by pruning the turns on L_L or by changing the length of the top section of the whip. Finally, $L_{\mu H}$ is adjusted for lowest SWR.

signal, which causes a continuous voltage that is propagated along the two wires in traveling-wave fashion.

This antenna is directive in the line of its orientation. In practice, it is made unidirectional by placing a terminating resistor at the end toward the direction of receiving interest. That is, if the antenna is oriented for reception of European signals from the USA, the antenna would run NE and SW, and the termination would be at the NE end of the antenna. The terminating resistance should equal that of the characteristic impedance of the line. If a 300-ohm open-wire line was used, the resistor would be a 300-ohm value. Energy from noise and signals off the back of the antenna is dissipated in the terminating resistor.

In the more common simple form, amateur Beverage antennas are made from long single spans of wire, one wavelength or greater in dimension, and situated only a few feet above ground. The characteristic impedance is dependent upon the spacing between the wire and ground, and this is often a nebulous determination because of unknown ground-conductivity properties. In such cases the amateur may elect to choose a ball-park termination value, say, 300 or 600 ohms, or greater, and make provisions for matching the antenna to the receiver by means of a resonant network. Of course, the end of the Beverage antenna need not come directly to the operating position. It can be matched outside the building and fed to the station through 50- or 75-ohm coaxial cable.

Two Beverage antennas can be used in

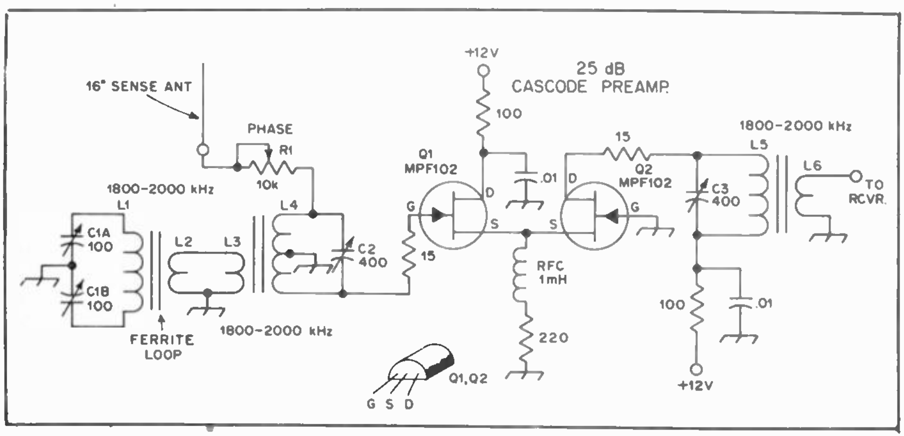


Fig. 19 — Schematic diagram of a ferrite loop antenna and preamplifier for 160-meter use. Fixed-value capacitors are disk ceramic and fixed-value resistors are 1/2-watt composition. C1 is a dual-section air variable. C2 and C3 are 400-pF mica trimmers. L1 is a modified Miller no. 2000 bc-antenna rod with 10 turns removed. L2 is a 5-turn link wound over the center of L1. L3 is a 5-turn link over L4, and L4 and L5 consist of 50 turns of no. 26 enam. wire on Amidon T80-2 toroid cores. L6 is a 5-turn link over L4. R1 is a 10,000-ohm linear-taper carbon control.

Table 4
Approximate Values for 8-ft. Mobile Whip

f KHZ	LOADING L μH	R_C (Q50) OHMS	R_C(Q300) OHMS	R_R OHMS	FEED R* OHMS	MATCHING L μH**
BASE LOADING						
1,800	345	77	13	0.1	23	3
3,800	77	37	6.1	0.35	16	1.2
7,200	20	18	3	1.35	15	0.6
14,200	4.5	7.7	1.3	5.7	12	0.28
21,250	1.25	3.4	0.5	14.8	16	0.28
29,000	–	–	–	–	36	0.23
CENTER LOADING						
1,800	700	158	23	0.2	34	3.7
3,800	150	72	12	0.8	22	1.4
7,200	40	36	6	3	19	0.7
14,200	8.6	15	2.5	11	19	0.35
21,250	2.5	6.6	1.1	27	29	0.29

R_C = Loading-coil resistance; R_R = Radiation resistance.
*Assuming loading coil Q = 300, and including estimated ground-loss resistance.
**For matching given feed resistance to 52 ohms.

a broadside configuration to obtain greater directivity. In that application the improvement of signal-to-noise ratio can be on the order of 25 to 40 dB as referenced to a simple dipole, assuming that the man-made and atmospheric noise sources do not exist in the line of antenna directivity.

Small Loop Antennas

Fractional-wavelength closed loops (known also as frame antennas) are useful in reducing noise from man-made and natural sources. Antennas of this kind exhibit a bidirectional characteristic at right angles to the plane of the loop, and the figure-eight pattern is broad. The minima of the loop were, however, very sharp, which makes it possible to orient the antenna so that noise is minimized while maximum response favors the received signal. Greater effectiveness in noise reduction can be realized by using a loop in combination with a sense antenna. The benefit comes from obtaining the resultant cardioid response pattern, which is similar to that obtained with a Beverage antenna. In the eastern part of the USA, for example, most storms move from west to east, or from south to north. A loop antenna with sense antenna can provide maximum response to the northeast for European DXing, and will discriminate against noise fronts coming in from the back side of the antenna.

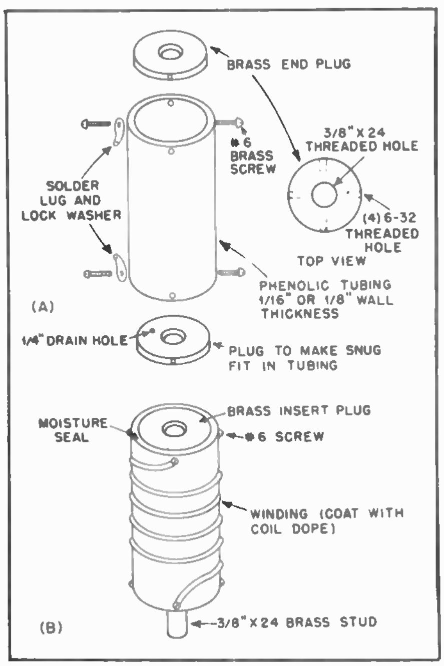


Fig. 21 — Constructional details for building a loading coil.

In order to tune a loop to resonance, it must be less than 0.1 wavelength in circumference, and its characteristic impedance will be extremely low — less than 1 ohm in most cases. Therefore, its efficiency will be quite poor, and typically below that of a Beverage antenna. A high-gain preamplifier is required for good results unless the station receiver is especially "hot" on the band of interest. If a sense antenna is used to obtain a cardioid pattern, the preamplifier should be situated at the feed point of the loop in order to provide a direct connection to the sense antenna.

The loop can take circular, rectangular, square, or diamond form, but circular and square formats are the more popular ones. It is essential to provide absolute symmetry in a loop and the latter two shapes help assure this.

Ferrite-rod loops are also suitable for low-noise directional reception. Whether the loop is wound on ferrite rod, or is one made from tubing, the height above ground is of no major importance. In fact, many amateurs report success when using loops indoors. Shielded loops made from RG-8/U or aluminum-jacketed foam-filled coaxial cable are suggested for the home constructor. The outer shield should be opened for approximately an inch exactly opposite the feed point to prevent having a shorted turn. Wood or bamboo support frames are normally used for loop antennas.

A shielded loop for 160 meters is illustrated in Fig. 18. A version using a ferrite rod and sense antenna is shown in Fig. 19. The JFET preamplifier has a gain of 25 dB, and can be used with either antenna.

Mobile Antennas

For mf and hf mobile operation the vehicular whip antenna is typically 8 feet in length and may taper in diameter from approximately 1/8 to 3/8 inch tip to base. For operation on frequencies below 28 MHz, a loading inductor is required to tune out the capacitive reactance (C_a) of the whip. The coil, L_L can be placed at the base of the whip or in the center. In theory the center-loaded antenna is more effective, but in practice it is difficult to observe that one is better or worse than the other — particularly at the low end of the hf spectrum and at mf.

To reduce losses in the system it is best to design a high-Q loading coil with large diameter wire on a low-loss form. To help assure a high Q the designer should maintain a coil-form factor of 1:1 to 2:1 (diameter-to-length ratio). At the lower frequencies this calls for a relatively large coil diameter.

The amount of capacitance represented by the whip will vary in accordance with the size of the vehicle and where the antenna is placed on the car. A close approximation can be had by using the equation for antennas shorter than 1/4 wavelength

$$C_a = \frac{17L}{\left[\left(\ln \frac{24L}{D}\right) - 1\right]\left[1 - \left(\frac{fL}{234}\right)^2\right]} \qquad \text{(Eq. 3)}$$

where C_a = antenna capacitance in pF
L = antenna height in feet
D = antenna diameter in inches
f = operating frequency in MHz
and ln = natural log

$$\ln \frac{24L}{D} = 2.3 \log_{10} \frac{24L}{D} \qquad \text{(Eq. 4)}$$

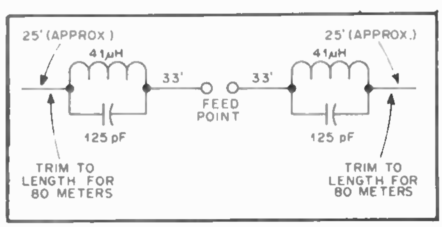


Fig. 22 — Example of a 40- and 80-meter trap dipole.

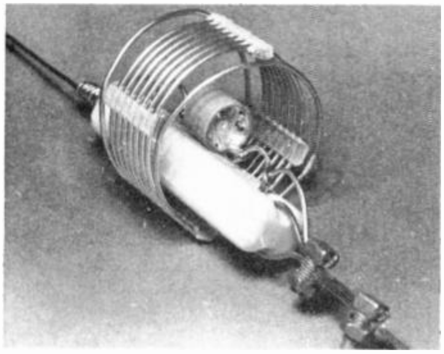


Fig. 23 – Photograph showing construction details of an antenna trap.

This formula is based on an antenna diameter which is constant rather than tapered. A closer approximation may be had by using the mean diameter in the equation. A whip that tapers from 1/8 to 3/8 inch can be considered as a 1/4-inch diameter element for the sake of the computations. As an example, let's assume an 8-foot, 1/4-inch mean-diameter whip will be used for operation at 1.9 MHz

$$2.3 \log_{10} \frac{24L}{D} = 2.3 \log_{10} \frac{192}{0.25}$$

$$= 2.3 \times 2.88 = 6.64$$

$$\therefore C_a = \frac{17L}{\left[6.64 - 1\right]\left[1 - \left(\frac{1.9 \times 8}{234}\right)^2\right]}$$

$$= \frac{136}{5.64 \times 1 - \left(\frac{15.2}{234}\right)^2}$$

$$= \frac{136}{5.64 \times 0.935^2} = 27.6 \text{ pF} \qquad \text{(Eq. 5)}$$

The capacitive reactance of the whip at 1.9 MHz is 3036 ohms. Therefore, an inductor of equal reactance (254 μH) is required to cancel the X_{Ca}. Reactance formulas are given earlier in this book. The foregoing assumes that base loading is used. For center loading divide the C_a by 2 and calculate the coil inductance.

Addition of a capacitance hat above the loading inductor will aid antenna performance by virtue of lowering the required coil inductance, which in turn reduces losses in the wire (fewer turns necessary). A nonconductive coil cover should be used to prevent rain, snow, and ice from detuning the antenna.

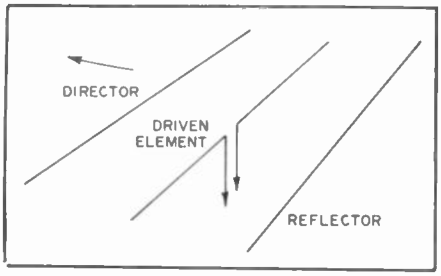


Fig. 24 – Layout of a Yagi antenna using a driven element, reflector and director.

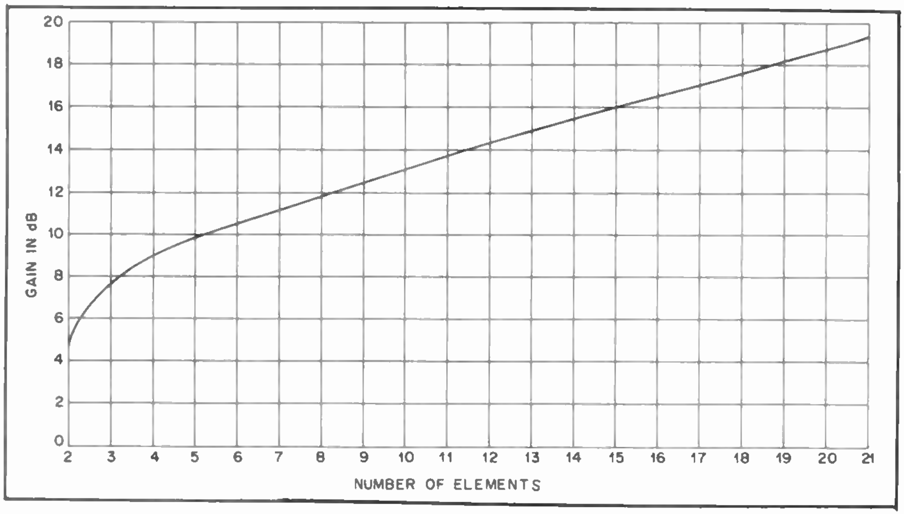


Fig. 25 – Gain in dB of a Yagi array over a dipole, assuming the array length is as given in Fig. 26.

Table 4 lists various characteristics for an 8-foot whip used as a base- or center-loaded antenna from 1.8 to 29 MHz. The values are approximate. A matching coil, $L_{\mu H}$ is listed in the table. It is shown in Fig. 20. After it is installed, the main loading coil should be pruned for minimum SWR at the desired operating frequency. It is best to use slightly more loading-coil inductance than the formula indicates, as that will allow sufficient leeway for final trimming. Fig. 21 shows construction techniques for homemade loading coils.

Antenna Traps

Multiband operation can be effected by installing antenna traps in dipole, beam, or vertical radiating systems. The trap consists of a parallel-resonant LC circuit. It is placed in the antenna at a point where the remainder of the element must be divorced from the overall system, and when placed there presents a high impedance to the signal energy. For example, an 80-meter dipole can be used also on 40 meters with a single coaxial feed line if the 40-meter traps are each located approximately 33 feet each side of the feed point. Addition of the traps will necessitate shortening the ends of the dipole somewhat in order to maintain resonance in the 80-meter band, as the trap coil adds inductance to the system. Fig. 22 illustrates the concept.

There are no set rules for the best L-C ratio to use in a trap, but the C should not be so great that it looks like a very low X_C at the trap frequency. Con-

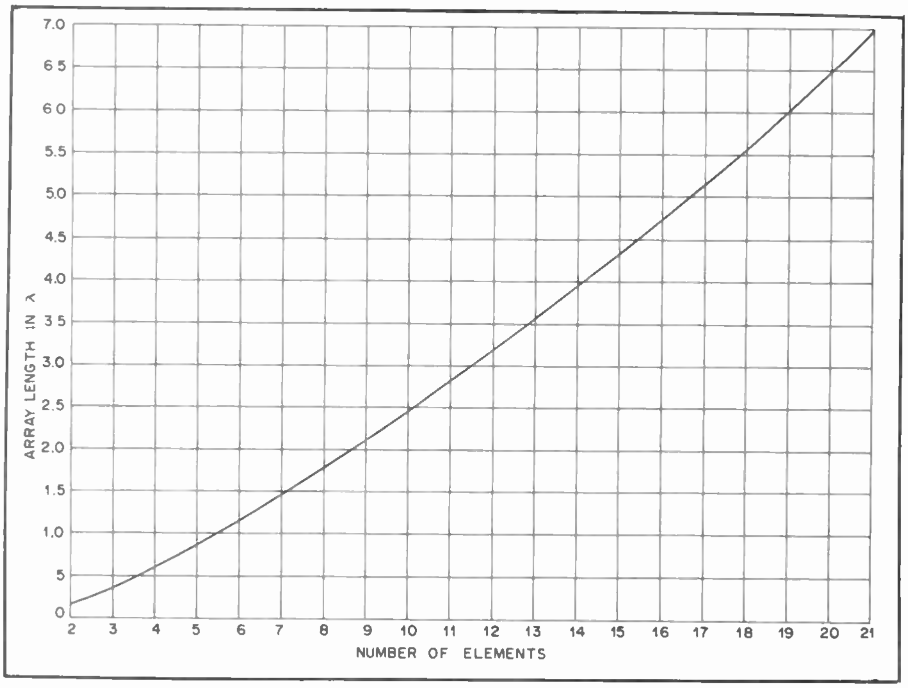


Fig. 26 – Optimum length of Yagi antennas as a function of the number of elements.

Table 5
Optimum Element Spacing for Multielement Yagi Arrays

NO. ELEMENTS	*R-DE*	*DE-D1*	*D1-D2*	*D2-D3*	*D3-D4*	*D4-D5*	*D5-D6*
2	0.15λ-0.2λ						
2		0.07λ-0.11λ					
3	0.16-0.23	0.16-0.19					
4	0.18-0.22	0.13-0.17	0.14λ-0.18λ				
5	0.18-0.22	0.14-0.17	0.15-0.20	0.17λ-0.23λ			
6	0.16-0.20	0.14-0.17	0.16-0.25	0.22-0.30	0.25λ-0.32λ		
8	0.16-0.20	0.14-0.16	0.18-0.25	0.25-0.35	0.27-0.32	0.27λ-0.33λ	0.30λ-0.40λ
8 to N	0.16-0.20	0.14-0.16	0.18-0.25	0.25-0.35	0.27-0.32	0.27-0.33	0.35-0.42

DE – Driven Element; R – Reflector; D – Director; N – any number; director spacings beyond D6 should be 0.35-0.42λ.

versely, X_c should not be so large as to degrade performance at the lowest operating frequency of the antenna. A suitable compromise can be had by making the reactances of L and C 180 ohms. Thus, for an 80-meter trap the capacitance would be 250 pF and the inductance would be 8.2 μH. The trap should be resonant at the lowest desired operating frequency. A high-Q circuit is best, and that calls for a high-quality capacitor and low-loss coil. High-voltage transmitting ceramic capacitors are excellent for the C element, and air-wound inductors are recommended for the L portion of the trap. Traps can be checked for resonance before they are installed in the antenna. A dip meter is a suitable instrument for the purpose. Fig. 23 shows how a trap can be built. Several amateur bands can be covered by installing a series of traps in an antenna. However, the cutting and testing of the various conductor lengths between and beyond the traps can be a tedious and time-consuming exercise.

Directional Gain Antennas

Most gain types of antennas are based on the classic Yagi-Uda or cubical-quad designs. Variations of the Yagi design are seen in partially driven arrays of elements, and in log-periodic configurations. Typically, a Yagi antenna is designed for one amateur band and uses parasitic elements rather than driven ones, as in Fig. 24.

Fig. 25 shows the gain in dB versus the number of Yagi elements employed – referenced to a half-wavelength dipole. The relationship between the optimum boom length and the number of Yagi elements is portrayed in Fig. 26.

Element thickness must be taken into account when determining the proper length. This is illustrated graphically in Fig. 27. Optimum element spacings for multielement Yagi antennas are listed in Table 5.

VHF and UHF Yagis

It becomes more practical at vhf and uhf to employ large numbers of parasitic Yagi elements, and the aluminum stock used can be of fairly small cross-sectional area. Considerable shortening of the elements will be required if large-diameter tubing is used for element material. For 6- and 2-meter beams it is common practice to use 3/8 and 1/4 inch stock for the elements, respectively. At 220 and 432 MHz 1/8-inch diameter solid element stock is suggested. Although solid metal construction of Yagis is common and acceptable, wooden booms may be preferred, as they will have less effect on the final element lengths. If metal booms are used, the diameter of the stock should be as small as possible, consistent with proper strength. When metal booms are used the element lengths should be increased 0.5 to 1 percent.

Fig. 28 gives dimensions for various vhf and uhf Yagis. Final adjustment of element lengths should be undertaken after the beams are assembled and matched. Adjustments should be made while the beam is a wavelength or more above ground and clear of surrounding conductive objects. Alternatively, the Yagi can be pointed toward the sky to minimize ground effects. Dimensions for various vhf and uhf Yagi antennas are given in Fig. 28.

Cubical Quads

Controversy exists concerning whether a cubical quad is superior to a Yagi antenna with respect to low-angle radiation. It is believed by many who have compared the two antennas from a given site that the quad is less dependent upon height above ground for good low-angle performance than is a Yagi. Furthermore, it is believed by some that the three- or four-element quad, when compared to a Yagi of equal element numbers each tested at an identical height above ground, will yield superior DX performance.

Quads are comprised of full wave closed loops of wire, and are supported on X-shaped frames of insulating material such as fiberglass or bamboo which has been treated to withstand weathering from rain, sun and other natural elements. Generally speaking, the cubical quad is broader in frequency response than a Yagi owing to its low-Q characteristic.

Quads can be built as square or diamond-shaped antennas. The main consideration in choosing the shape will

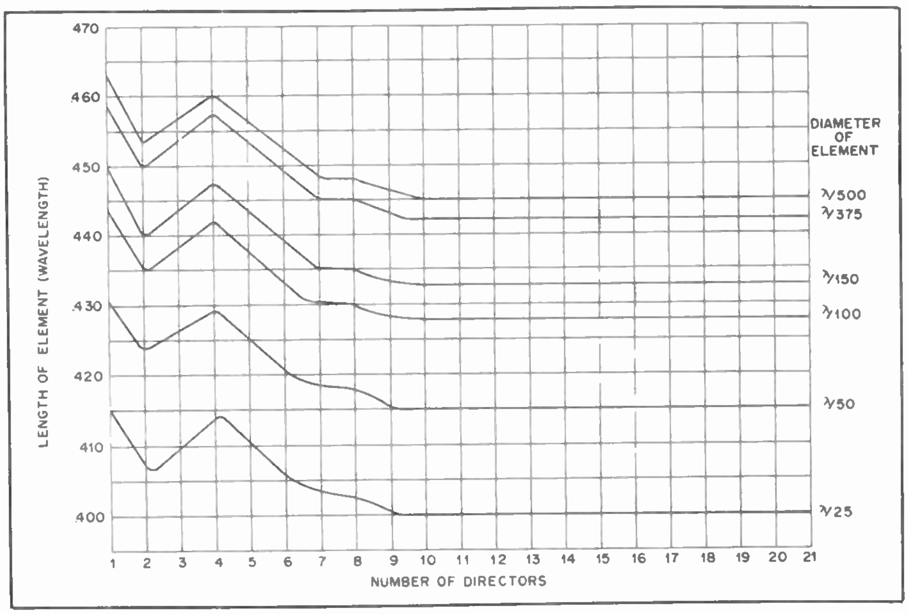


Fig. 27 – Length of director versus its position in the array, for various element thicknesses.

Table 6
Three-Band Quad Loop Dimensions

BAND	*REFLECTOR*	*DRIVEN ELEMENT*	*FIRST DIRECTOR*	*SECOND DIRECTOR*	*THIRD DIRECTOR*
20 Meters	(A) 72'8"	(B) 71'3"	(C) 69'6"	–	–
15 Meters	(D) 48'6-1/2"	(E) 47'7-1/2"	(F) 46'5"	(G) 46'5"	–
10 Meters	(H) 36'2-1/2"	(I) 35'6"	(J) 34'7"	(K) 34'7"	(L) 34'7"

Letters indicate loops identified in Fig. 29.

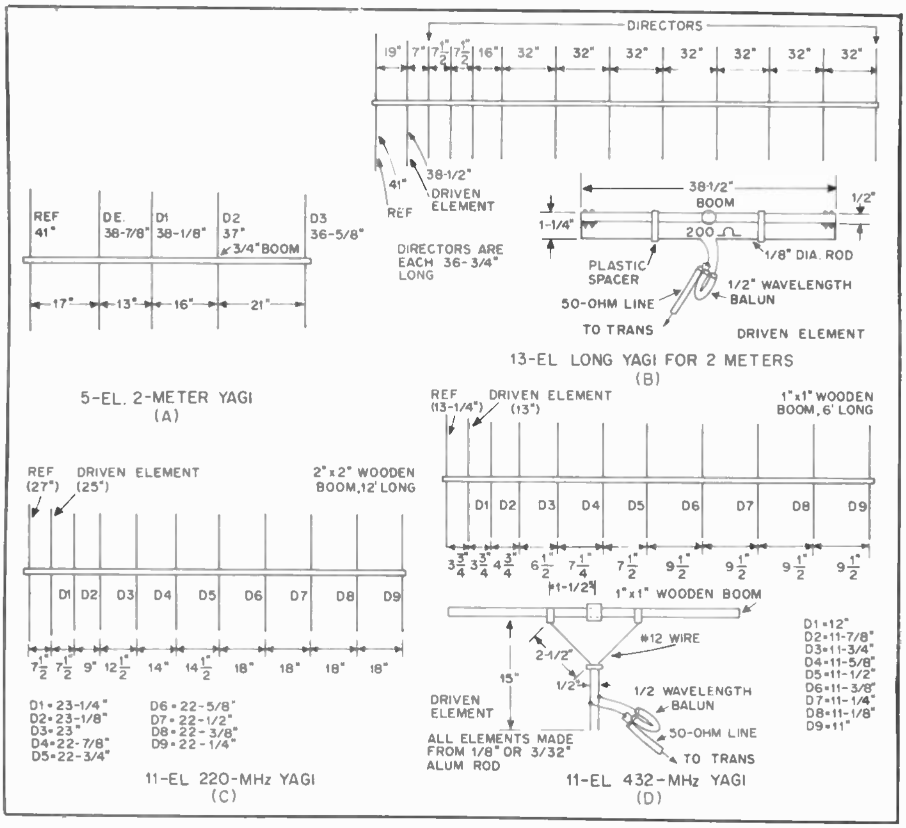


Fig. 28 – Dimensions for various vhf and uhf Yagi arrays.

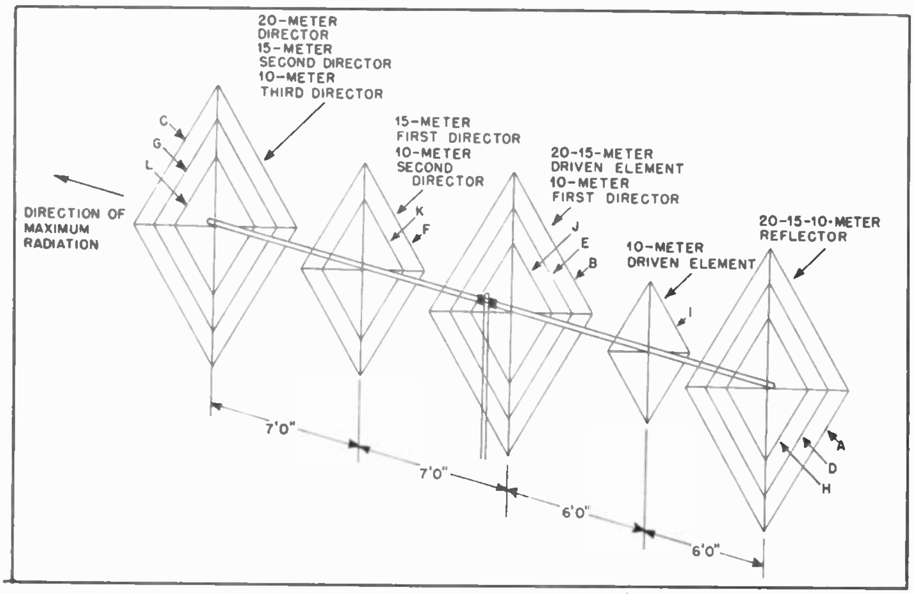


Fig. 29 – Dimensions of a three-band quad, not drawn to scale. See Table 6 for dimensions of lettered wires.

be physical ruggedness of the support structure. For horizontal polarization the feed point should be in the center of the bottom leg, or the lower point of the diamond. When the quad is fed at the side, vertical polarization results.

Fig. 29 shows how the elements of a three-band quad are laid out. Dimensions for the elements are listed in Table 6. Dimensions for a variety of quad arrays are given in Table 7. Detailed information on quad construction can be seen in *The ARRL Antenna Book.*

Transmission Lines

In the interest of minimizing transmission-line losses it is of value to utilize air-dielectric feeders, and this is especially true when long runs of line must be employed at vhf, uhf and the upper part of the hf spectrum. Fig. 30 provides equations and examples of various kinds of open-wire line for amateur use. A single-wire line above ground or some other plane-conducting surface is illustrated in Fig. 30A. The equation can be used to calculate the approximate impedance of the Beverage antenna which was described earlier in this chapter.

Fig. 30B shows the familiar two-wire open transmission line. This type of feeder is often used for feeding multiband dipoles by means of a Transmatch.

Where extremely long runs of coaxial line are necessary, but when the expense of air-dielectric copper line is beyond the means of the operator, a four-wire line of conductors can be used. This is illustrated in Fig. 30C. Some a-m broadcast systems use four-wire line to feed station antennas which are located a great distance from the transmitter – 100 yards away or farther. The four-wire line feeder can be supported on short poles with wooden crossarms which are fitted with insulators.

Fig. 31 shows a section of air-dielectric coaxial line and gives the equation for determining the characteristic impedance of the line. An example of the solution to an impedance problem is included.

A chart of impedance values for two-wire line using various popular sizes of wire and tubing is given in Fig. 32A. For conductor sizes and spacings not shown in the chart, use the equation of Fig. 30B. Data for coaxial line are shown in Fig. 32B.

Aluminum Tubing

Table 8 lists the physical characteristics of standard U.S. aluminum tubing which can be used for beam antennas and transmission-line sections. The dimensions are those for tubing which comes in 12-foot lengths.

Table 9 contains a list of electrical characteristics for various types of manufactured coaxial cable. Specific data

Table 7
Quad Dimensions

2-element quad (W7ZQ)
Spacing (given below)
Boom length (given below)

BAND	*40 M.*	*20 M.*	*15 M.*	*10 M.*
Reflector	144'11-1/2"	72'4"	48'8"	35'7"
Driven Element	140'11-1/2"	70'2"	47'4"	34'7"
Spacing	30'	13'	10'	6'6"
Boom length	30'	13'	10'	6'6"
Feed method	Directly with 23' of RG11, then any length of RG8 coax	Directly with 11'7" RG11, then any length RG8 coax	Directly with 7'8½" RG11, then any length RG8 coax	Directly with 5'8" RG11, then any length RG8 coax

Note that a spider or boomless quad arrangement could be used for the 10/15/20 meter parts of the above dimensions yielding a triband antenna.

4-element quad* (WØAIW-20 M./W7ZQ**/KØKKU/KØEZH/W6FXB)
Spacing: equal; 10 ft.
Boom length: 30 ft.

BAND	*20 M. PHONE*	*CW*	*15 M.*	*10 M.*
Reflector	72'1-1/2"	72'5"	48'8"	35'8-1/2"
Driven Element	70'1-1/2"	70'5"	47'4"	34'8-1/2"
Director 1	69'1"	69'1"	46'4"	33'7-1/4"
Director 2	69'1"	69'1"	46'4"	33'7-1/4"
Feed method	Directly with 50-ohm coax		Directly with 50-ohm coax	Directly with 5'9" RG11, then any length RG8 coax

*Common boom used to form a triband array.
**The 2-element 40-meter quad given above is added to form a four-band quad array.

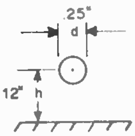
4-element quad (W7ZQ /K8DYZ*/K8YIB*/W7EPA*)
Spacing: equal; 13'4"
Boom length: 40 ft.

BAND	*20 M.*	*15 M.*	*10 M.*
Reflector	72'5"	48'4"	35'8-1/2"
Driven Element	70'5"	47'0"	34'8-1/2"*
Director 1	69'1"	46'1"	(Directors 1-3 all 33'7")*
Director 2	69'1"	46'1"	
Feed method	Directly with 50-ohm coax	Directly with 7'9" RG11, then any length 50-ohm coax	Directly with 50-ohm coax

*For the 10-meter band the driven element is placed between the 20/15 reflector and 20/15 driven element. The 10-meter reflector is placed on the same frame as the 20/15-meter reflectors and the remaining 10-meter directors are placed on the remaining 20/15-meter frames. The 10-meter portion is then a 5-element quad.

6 element Quad (WØYDM, W7UMJ)
Spacing: equal; 12 ft.
Boom length: 60 ft.

BAND	*20 M.*
Reflector	72'1-1/2"
Driven Element	70'1-1/2"
Directors 1, 2 and 3	69'1"
Director 4	69'4"
Feed method	Directly with 50-ohm coax

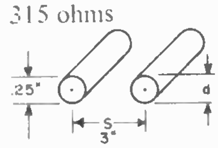


(A)

$$Z_o = 138 \log \frac{4h}{d} \text{ ohms}$$

$$\therefore Z_o = 138 \log \frac{48}{0.25} = 138 \times 2.28$$

$$= 315 \text{ ohms}$$

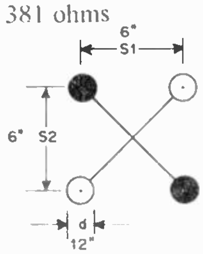


(B)

$$Z_o = 276 \log \frac{2S}{d} \text{ ohms}$$

$$\therefore Z_o = 276 \log \frac{6}{0.25} = 276 \times 1.38$$

$$= 381 \text{ ohms}$$



(C)

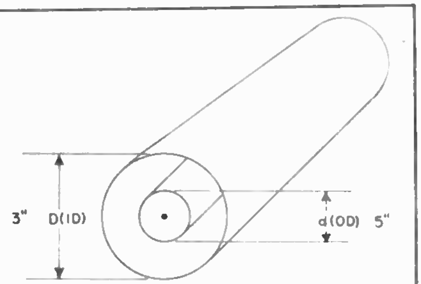
$$Z_o = 138 \log \frac{2S2}{d\sqrt{1 + (S2/S1)^2}}$$

$$\therefore Z_o = 138 \log \frac{12}{0.12\sqrt{(1+1)^2}}$$

$$= 138 \log \frac{12}{0.12 \times 2} = 138 \log 50$$

$$= 138 \times 1.69 = 234 \text{ ohms}$$

Fig. 30 — Equations and illustrations for determining the characteristic impedances of various air-dielectric transmission lines. An example is worked out in each case.



$$Z_o = 138 \log \frac{D}{d} \text{ ohms}$$

$$\therefore Z_o = 138 \log \frac{3}{0.5} = 138 \log 6$$

$$= 138 \times 0.778 = 107 \text{ ohms}$$

Fig. 31 — Example of coaxial transmission line with equation and example for determining characteristic impedance.

Table 8
6061-T6 (61S-T6) Round Aluminum Tube

O D INCHES	WALL THICKNESS INCHES	STUBS GA.	I D INCHES	APPROX. WEIGHT PER FOOT	APPROX. WEIGHT PER LENGTH
3/16"	.035	(No. 20)	.117	.019 lbs.	.228 lbs.
	.049	(No. 18)	.089	.025 lbs.	.330 lbs.
1/4"	.035	(No. 20)	.180	.027 lbs.	.324 lbs.
	.049	(No. 18)	.152	.036 lbs.	.432 lbs.
	.058	(No. 17)	.134	.041 lbs.	.492 lbs.
5/16"	.035	(No. 20)	.242	.036 lbs.	.432 lbs.
	.049	(No. 18)	.214	.047 lbs.	.564 lbs.
	.058	(No. 17)	.196	.055 lbs.	.660 lbs.
3/8"	.035	(No. 20)	.305	.043 lbs.	.516 lbs.
	.049	(No. 18)	.277	.060 lbs.	.720 lbs.
	.058	(No. 17)	.259	.068 lbs.	.816 lbs.
	.065	(No. 16)	.245	.074 lbs.	.888 lbs.
7/16"	.035	(No. 20)	.367	.051 lbs.	.612 lbs.
	.049	(No. 18)	.339	.070 lbs.	.840 lbs.
	.065	(No. 16)	.307	.089 lbs.	1.068 lbs.
1/2"	.028	(No. 22)	.444	.049 lbs.	.588 lbs.
	.035	(No. 20)	.430	.059 lbs.	.708 lbs.
	.049	(No. 18)	.402	.082 lbs.	.984 lbs.
	.058	(No. 17)	.384	.095 lbs.	1.040 lbs.
	.065	(No. 16)	.370	.107 lbs.	1.284 lbs.
5/8"	.028	(No. 22)	.569	.061 lbs.	.732 lbs.
	.035	(No. 20)	.555	.075 lbs.	.900 lbs.
	.049	(No. 18)	.527	.106 lbs.	1.272 lbs.
	.058	(No. 17)	.509	.121 lbs.	1.452 lbs.
	.065	(No. 16)	.495	.137 lbs.	1.644 lbs.
3/4"	.035	(No. 20)	.680	.091 lbs.	1.092 lbs.
	.049	(No. 18)	.652	.125 lbs.	1.500 lbs.
	.058	(No. 17)	.634	.148 lbs.	1.776 lbs.
	.065	(No. 16)	.620	.160 lbs.	1.920 lbs.
	.083	(No. 14)	.584	.204 lbs.	2.448 lbs.
7/8"	.035	(No. 20)	.805	.108 lbs.	1.308 lbs.
	.049	(No. 18)	.777	.151 lbs.	1.810 lbs.
	.058	(No. 17)	.759	.175 lbs.	2.100 lbs.
	.065	(No. 16)	.745	.199 lbs.	2.399 lbs.
1"	.035	(No. 20)	.930	.123 lbs.	1.476 lbs.
	.049	(No. 18)	.902	.170 lbs.	2.040 lbs.
	.058	(No. 17)	.884	.202 lbs.	2.424 lbs.
	.065	(No. 16)	.870	.220 lbs.	2.640 lbs.

O D INCHES	WALL THICKNESS INCHES	STUBS GA.	I D INCHES	APPROX. WEIGHT PER FOOT	APPROX. WEIGHT PER LENGTH
1"	.083	(No. 14)	.834	.281 lbs.	3.372 lbs.
1-1/8"	.035	(No. 20)	1.055	.139 lbs.	1.668 lbs.
	.058	(No. 17)	1.009	.228 lbs.	2.736 lbs.
1-1/4"	.035	(No. 20)	1.180	.155 lbs.	1.860 lbs.
	.049	(No. 18)	1.152	.210 lbs.	2.520 lbs.
	.058	(No. 17)	1.134	.256 lbs.	3.072 lbs.
	.065	(No. 16)	1.120	.284 lbs.	3.408 lbs.
	.083	(No. 14)	1.084	.357 lbs.	4.284 lbs.
1-3/8"	.035	(No. 20)	1.305	.173 lbs.	2.076 lbs.
	.058	(No. 17)	1.259	.282 lbs.	3.384 lbs.
1-1/2"	.035	(No. 20)	1.430	.180 lbs.	2.160 lbs.
	.049	(No. 18)	1.402	.260 lbs.	3.120 lbs.
	.058	(No. 17)	1.384	.309 lbs.	3.708 lbs.
	.065	(No. 16)	1.370	.344 lbs.	4.128 lbs.
	.083	(No. 14)	1.334	.434 lbs.	5.208 lbs.
	*.125	1/8"	1.250	.630 lbs.	7.416 lbs.
	*.250	1/4"	1.000	1.150 lbs.	14.832 lbs.
1-5/8"	.035	(No. 20)	1.555	.206 lbs.	2.472 lbs.
	.058	(No. 17)	1.509	.336 lbs.	4.032 lbs.
1-3/4"	.058	(No. 17)	1.634	.363 lbs.	4.356 lbs.
	.083	(No. 14)	1.584	.510 lbs.	6.120 lbs.
1-7/8"	.058	(No. 17)	1.759	.389 lbs.	4.668 lbs.
2"	.049	(No. 18)	1.902	.350 lbs.	4.200 lbs.
	.065	(No. 16)	1.870	.450 lbs.	5.400 lbs.
	.083	(No. 14)	1.834	.590 lbs.	7.080 lbs.
	*.125	1/8"	1.750	.870 lbs.	9.960 lbs.
	*.250	1/4"	1.500	1.620 lbs.	19.920 lbs.
2-1/4"	.049	(No. 18)	2.152	.398 lbs.	4.776 lbs.
	.065	(No. 16)	2.120	.520 lbs.	6.240 lbs.
	.083	(No. 14)	2.084	660 lbs.	7.920 lbs.
2-1/2"	.065	(No. 16)	2.370	.587 lbs.	7.044 lbs.
	.083	(No. 14)	2.334	.740 lbs.	8.880 lbs.
	*.125	1/8"	2.250	1.100 lbs.	12.720 lbs.
	*.250	1/4"	2.000	2.080 lbs.	25.440 lbs.
3"	.065	(No. 16)	2.870	.710 lbs.	8.520 lbs.
	*.125	1/8"	2.700	1.330 lbs.	15.600 lbs.
	*.250	1/4"	2.500	2.540 lbs.	31.200 lbs.

*These sizes are extruded. All other sizes are drawn tubes.

which are not found in the table can be obtained by contacting the manufacturers directly, or by consulting their catalogs.

SWR and Line Loss

The power lost in a transmission line is not directly proportional to the line length, but varies logarithmically with the length. Thus, if 10 percent of the power is lost in a section of line of a given length, 10 percent of the remaining power will be lost in the next section of the same length, and so on. The decibel is a logarithmic unit of measurement, and is therefore applied when measuring line losses. Line loss is normally expressed in dB per 100 feet of line. The loss varies with frequency, and this can be seen in Fig. 33.

An increase in line loss occurs because of SWR. Total loss is based on the characteristic loss of the line versus

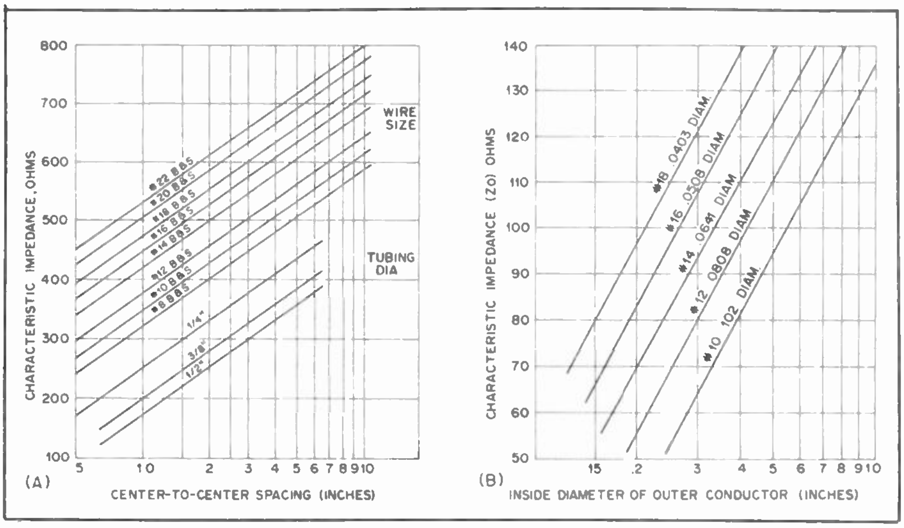


Fig. 32 – Chart showing characteristic impedance of spaced-conductor parallel transmission lines with air dielectric, at A. Tubing sizes are given in OD. At B, characteristic impedance of air-dielectric coaxial lines.

Table 9

	1	2	3	4	5	6	7	8	9	10	11		12
TYPE OF LINE	NOMINAL IMP., OHMS	RG/U TYPE	MFRS. NO.	OUTSIDE DIA. IN.	JACKET	INNER COND. SIZE	DIELECTRIC	CAP. PER FT. pF	VELOCITY FACTOR	MAX. RMS V	POWER RATING, W UP TO 30 MHz	400 MHz	CONNECTOR SERIES
FLEXIBLE COAXIAL MEDIUM	52	8		.405	I	7 21	SP	29.5	66	5000	1720	465	UHF, N
	52	8A		.405	IIA	7/21	SP	29.5	.66	5000	1720	465	UHF, N
	50		621-111	.405	I	7/19	FP	24.5	80	1500			UHF
	50		T-4 50	.407	X	10	FP						UHF
	75	11		.405	I	7 26	SP	20.5	66	5000	1400	340	UHF, N
	75	11A		.405	IIA	7/26	SP	20.5	.66	5000	1400	340	UHF, N
	75		621-100	.405	I	14	FP	16.5	.80	3000			UHF
	75		JT-204	.407	X	14	FP						UHF
SMALL	53.5	58		.195	I	20	SP	28.5	.66	1900	580	135	UHF, BNC, N
	50	58A		.195	I	19/.0071	SP	30	.66	1900	550	105	UHF, BNC, N
	53.5	58B		.195	IIIA	20	SP	28.5	66	1900	580	135	UHF, BNC, N
	50	58C		.195	IIA	19 .0071	SP	30	66	1900	550	105	UHD, BNC, N
	73	59		.242	I	22 cw	SP	21 5	.66	2300	720	185	UHF, BNC, N
	75	59B		.242	IIA	023 cw	SP	21	66	2300	720	185	UHF, BNC, N
	73		621 186	242	P	20 cw	FP	17 3	80	1000			UHF, BNC
	93	62		.242	I	22 cw	SSP	13.5	84	750	850	230	UHF, BNC, N
PARALLEL CONDUCTOR FLAT OR OVAL	75		214-023			7 21	SP	20	71		1000		
	300		214-056			7 28	SP	5 8	82				
	300		214 022			16 cw	SP	3 0	.82				
TUBULAR	300		214 271			7 28	PA		.82		500		
	300		214-076			7 26	PA	3 9	82		1000		
	300		214 103			7/28	FP*						

Column 3 T-4-50 and JT 204 are manufactured by Times Wire & Cable, Wallingford, CT. Other numbers are types made by Amphenol, Chicago, IL.

Column 5: 1 – Polyvinylchloride (PVC), black. IIA – Noncontaminating PVC, black or gray. IIA – Polyethylene, black. Noncontaminating and abrasion-resistant. Recommended when cable is to be buried underground. P – Polyethylene, X – Xelon.

Column 6: Conductors are copper unless followed by CW (copper-weld). Decimal numbers give wire diameter in inches; others are standard copper-wire gauge except when preceding a virgule, when the figure indicates number of strands: e.g., 7/21 means 7 strands of no. 21 copper wire.

Column 7: SP – Solid polyethylene, SSP – Polyethylene strand wound around inner conductor; enclosed in solid tube of same material. FP – Foamed polyethylene. FP* – Foamed polyethylene surrounding each conductor; outer enclosure solid polyethylene. Type 214-103 is intended for use under adverse moisture and salt-spray conditions. PA – Polyethylene tube with air core.

Column 9: Open parallel-conductor line has a velocity factor of 0.95 to 0.975, depending on number of spacers and dielectric material of which they are made. Polyethylene spacers used in types listed.

Column 12: Only connectors in common use by amateurs are included.

Fig. 33 – Nomograph of line loss per 100 feet versus operating frequency.

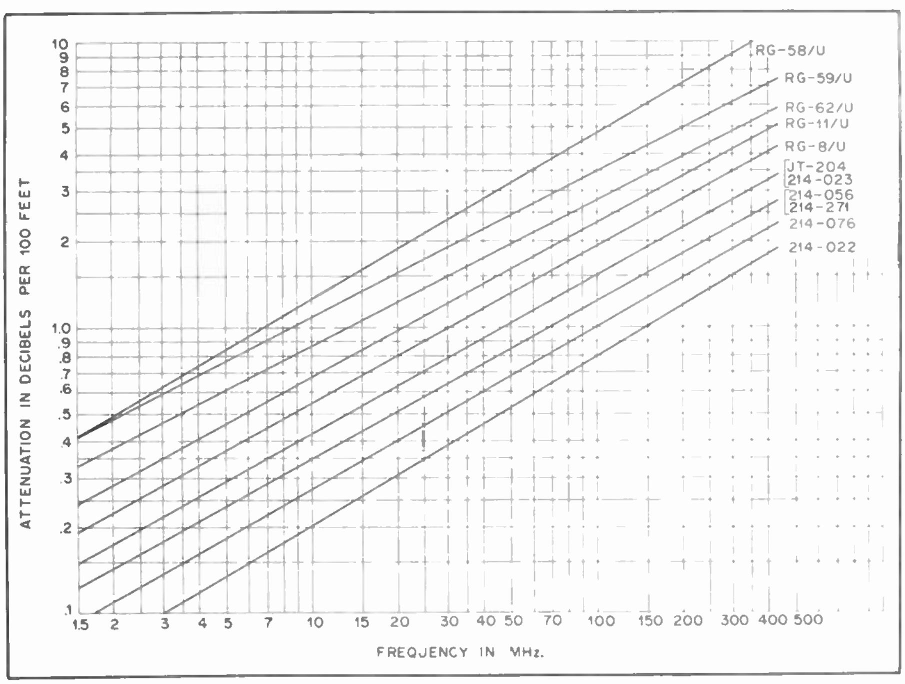
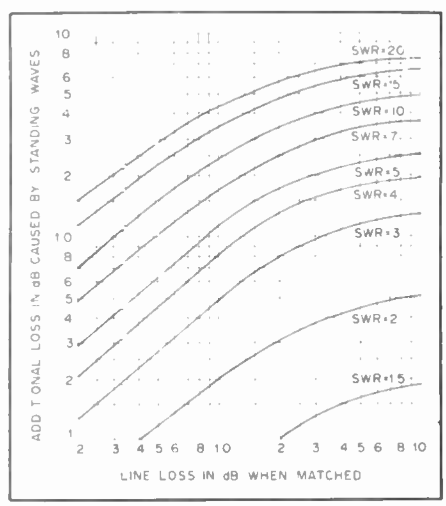


Fig. 34 – Nomograph of line loss versus SWR as compared to line loss in dB for a matched condition.



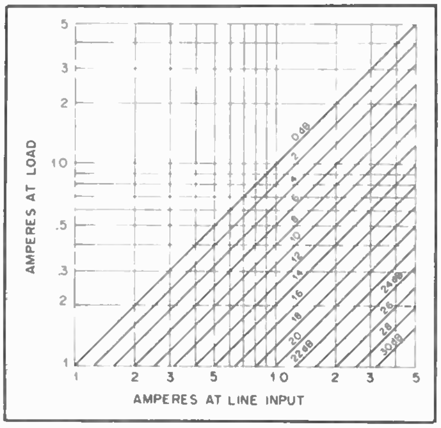


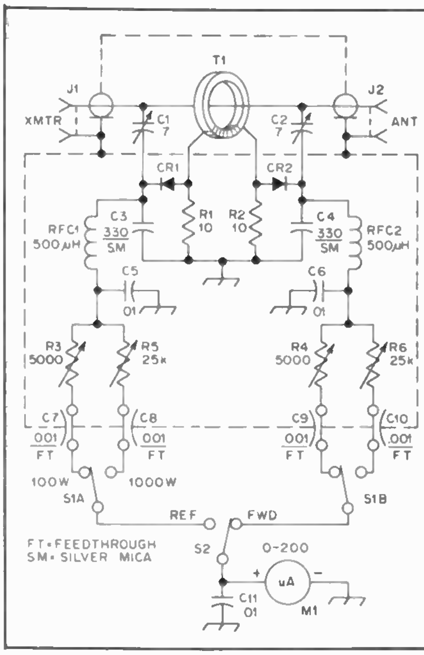
Fig. 35 — Graph for determining losses in lines being tested for quality factor.

frequency with an SWR of 1, plus the loss incurred as a result of standing waves. The relationship between SWR and line loss is illustrated in Fig. 34.

Additional losses can occur as a result of contaminated coaxial cable. The contamination results from age and exposure to the natural elements. If the chemical properties of the insulating material become degraded sufficiently, considerable line loss will result. It is wise, therefore, to check the condition of used or surplus coaxial cable before installing it in an antenna system. Fig. 35 shows how the dB loss can be determined by means of an rf ammeter and a dummy load which provides an SWR of 1. The rf ammeter is inserted first at the transmitter end of the line, then placed at the dummy-load end of the line. The readings are compared and checked against the chart of Fig. 35.

SWR and RF Power Measurements

An instrument suitable for measuring rf power and SWR is shown schematically in Fig. 36. The center conductor of the feed line passes through the center of toroidal transformer T1 to function as the primary winding. The multiturn winding is the secondary. Current flowing through the primary induces a voltage in the secondary which causes current to flow through R1 and R2. The voltage drops across R1 and R2 are equal in amplitude but of opposite phase with respect to ground. Capacitive voltage dividers C1/C3 and C2/C4 are connected across the line to obtain equal-amplitude voltages *in phase* with the line voltage, the division ratio being adjusted so that these voltages match the voltage drops across R1 and R2 in amplitude. The current/voltage ratio in the line depends on the load. Therefore, the bridge must be standardized in a pure resistance (50 or 75 ohms, depending on the desired line impedance) during initial adjustment, and later use.



WATTS	*MI*	*WATTS*
100	200	1000
90	180	900
80	170	800
70	155	700
60	145	600
50	125	500
40	105	400
30	85	300
20	65	200
10	40	100
5	20	50

Fig. 36 — Schematic diagram of the rf wattmeter. A calibration scale for *MI* is shown also. Fixed-value resistors are 1/2-watt composition. Fixed-value capacitors are disk ceramic unless otherwise noted. Decimal-value capacitances are in µF. Others are pF. Resistance is in ohms; k = 1000.

C1, C2 — 1.3 to 6.7-pF miniature trimmer (E.F. Johnson 189-502-4, available from Newark Electronics, Chicago, IL.).
C3-C11, incl. — Numbered for circuit-board identification.
CR1, CR2 — Matched small-signal germanium diodes, 1N34A, etc. (see text).
J1, J2 — Chassis-mount coax connector of builder's choice. Type SO-239 used here.
M1 — 0 to 200-µA meter.
R1, R2 — Matched 10-ohm resistors (see text).
R3, R4 — 5000-ohm printed-circuit carbon control (IRC R502-B).
R5, R6 — 25,000-ohm printed-circuit carbon control (IRC R252-B).
RFC1, RFC2 — 500-µH rf choke (Millen 34300-500 or similar).
S1 — Dpdt single-section phenolic wafer switch (Mallory 3222J).
S2 — Spdt phenolic wafer switch (Centralab 1460).
T1 — Toroidal transformer; 35 turns of no. 26 enam. wire to cover entire core of Amidon T-68-2 toroid (Amidon Assoc., 12033 Otsego St., N. Hollywood, CA 91607).

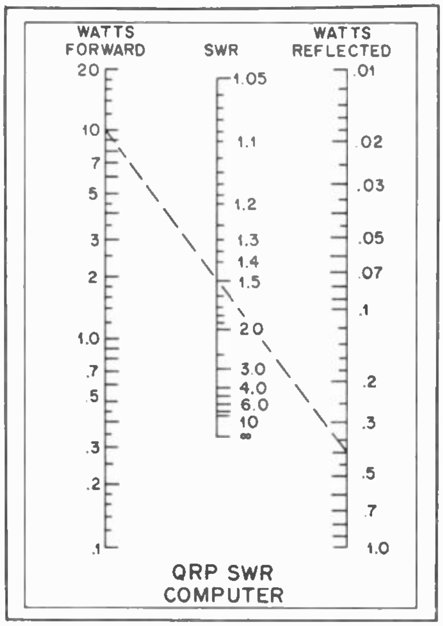


Fig. 37 — Nomograph of SWR versus forward and reflected power for levels up to 20 watts. Dashed line shows an SWR of 1.5:1 for 10 W forward and 0.4 W reflected.

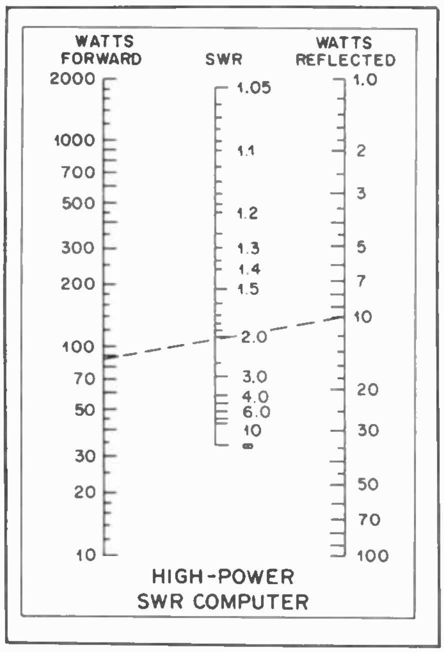


Fig. 38 — Nomograph of SWR versus forward and reflected power for levels up to 2000 watts. Dashed line shows an SWR of 2:1 for 90 W forward and 10 W reflected.

Adjustment is carried out by terminating first one port then the other in 50 or 75 ohms. C1 and C2 are adjusted for a null in reflected-power reading, respectively, as the bridge is reversed.

Controls R3/R4 and R5/R6 are set to provide full-scale sensitivities of 200 and 1000 watts, respectively. An rf ammeter can be used as the standard while feeding power into a resistive dummy load. Alternatively, an rf probe and VTVM can be used to determine rms voltage across a dummy load. The voltage/resistance readings can then be converted to watts.

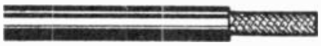
Fig. 37 contains a handy nomograph for computing SWR when forward and reflected power amounts are known. The scales are for power levels up to 20 watts (QRP). A similar nomograph is given in Fig. 38 for a maximum power of 2000 watts.

Coaxial-Line Connectors

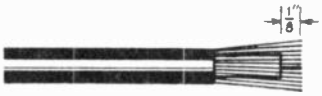
The following illustrations show how to prepare coaxial cable for attachment of coaxial connectors of various types.

BNC (UG-88/U) CONNECTORS

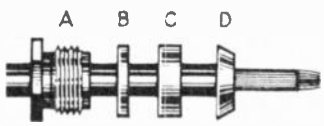
Connectors bearing suffix letters (UG-88C/U, etc.) differ slightly in internal construction; assembly and dimensions must be varied accordingly.



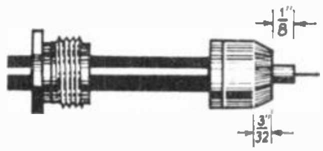
Cut end square and trim jacket 5/16" for RG-58/U.



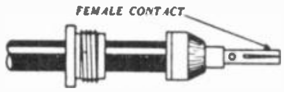
Fray shield and strip inner dielectric 1/8". Tin center conductor.



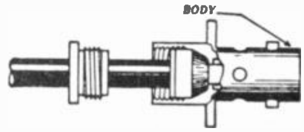
Taper braid and slide nut (A), washer (B), gasket (C), and clamp (D), over braid. Clamp is inserted so that its inner shoulder fits squarely against end of cable jacket.



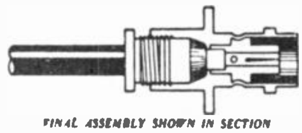
With clamp in place, comb out braid, fold back smooth as shown, and trim 3/32" from end.



Tin center conductor of cable. Slip female contact in place and solder. Remove excess solder. Be sure cable dielectric is not heated excessively and swollen so as to prevent dielectric entering body.



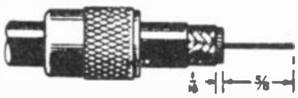
Push into body as far as it will go. Slide nut into body and screw into place with wrench until tight. Hold cable and shell rigidly and rotate nut.



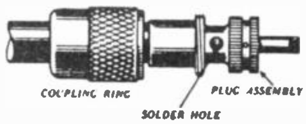
This assembly procedure applies to BNC jacks. The assembly for plugs is the same except for the use of male contacts and a plug body.

83-1SP (PL-259) PLUG

Cut end of cable even. Remove vinyl jacket 1-1/8" — don't nick braid.

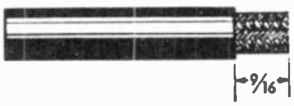


Bare 5/8" of center conductor — don't nick conductor. Trim braided shield 9/16" and tin. Slide coupling ring on cable.

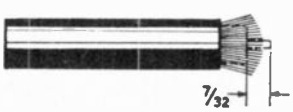


Screw the plug assembly on cable. Solder plug assembly to braid through solder holes. Solder conductor to contact sleeve. Screw coupling ring on assembly.

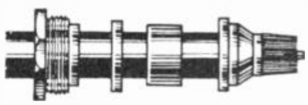
N (UG-21/U) CONNECTORS



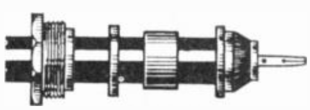
Remove 9/16" of vinyl jacket. When using double-shielded cable remove 5/8".



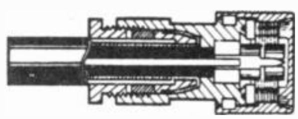
Comb out copper braid as shown. Cut off dielectric 7/32" from end. Tin center conductor.



Taper braid as shown. Slide nut, washer and gasket over vinyl jacket. Slide clamp over braid with internal shoulder of clamp flush against end of vinyl jacket. When assembling connectors with gland, be sure knife-edge is toward end of cable and groove in gasket is toward the gland.



Smooth braid back over clamp and trim. Soft-solder contact to center conductor. Avoid use of excessive heat and solder. See that end of dielectric is clean. Contact must be flush against dielectric. Outside of contact must be free of solder. Female contact is shown; procedure is similar for male contact.

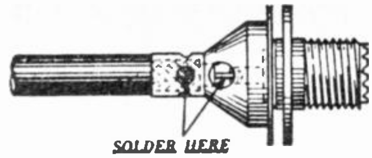


Slide body into place carefully so that contact enters hole in insulator (male contact shown). Face of dielectric must be flush against insulator. Slide completed assembly into body by pushing nut. When nut is in place, tighten with wrenches. In connectors with gland, knife edge should cut gasket in half by tightening sufficiently.

83 SERIES (SO-239) WITH HOODS

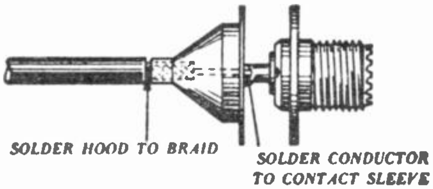
Cut end of cable even. Remove vinyl jacket to dimension appropriate for type of hood. Tin exposed braid.

Remove braid to expose dielectric to appropriate dimension. Tin center conductor. Soldering and assembly depends on the hood used, as illustrated.

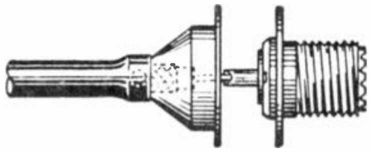


Slide hood over braid. Bring receptacle flush against hood. Solder hood to braid and conductor to contact sleeve through solder holes as illustrated. Tape junction if necessary. (For UG-372/U.)

Remove braid and dielectric to expose center conductor. Do not nick conductor.

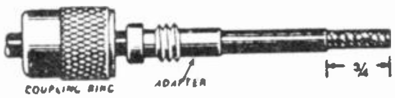


Slide hood over braid. Solder conductor to contact. Slide hood flush against receptacle and bolt both to chassis. Solder hood to braid as illustrated. Tape junction if necessary. (For UG-177/U.)



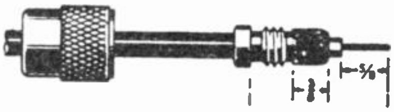
Slide hood over braid and force under vinyl. Place inner conductor in contact sleeve and solder. Push hood flush against receptacle. Solder hood to braid through solder holes. Tape junction if necessary. (For UG-106/U.)

83-1SP (PL-259) PLUG WITH ADAPTERS (UG-176/U) OR UG-175/U)

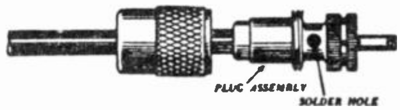


Cut end of cable even. Remove vinyl jacket 3/4" – don't nick braid. Slide coupling ring and adapter on cable.

Fan braid slightly and fold back over cable.



Position adapter to dimension shown. Press braid down over body of adapter and trim to 3/8". Bare 5/8" of conductor. Tin exposed center conductor.



Screw the plug assembly on adapter. Solder braid to shell through solder holes. Solder conductor to contact sleeve.

Screw coupling ring on plug assembly.

Chapter 8

Catalog of Solid-State Circuits

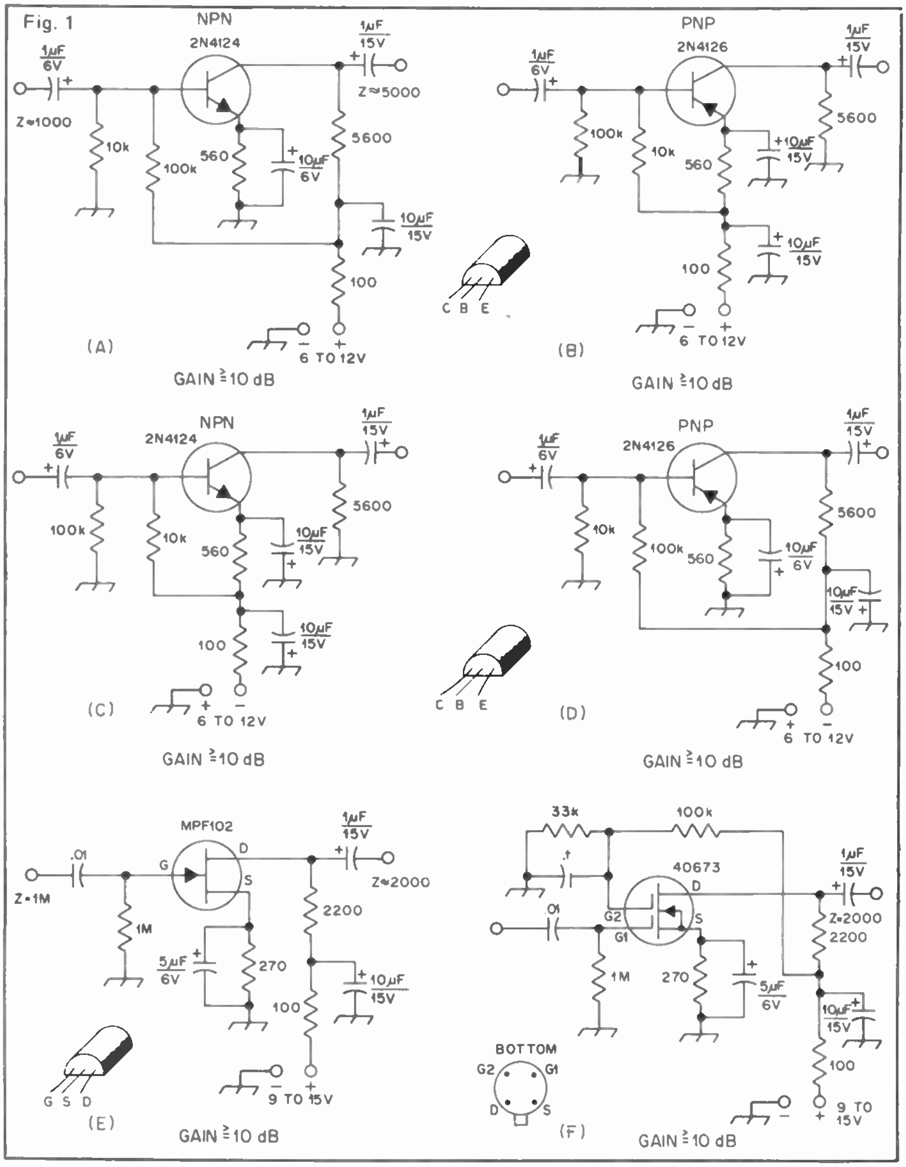
In the course of building amateur equipment it is not unusual to spend considerable time arriving at the circuit configurations and parts values necessary for good performance. It would be convenient at times to have on hand a collection of workable circuit diagrams for various common individual stages of solid-state receivers and transmitters. The purpose of this chapter is to provide a catalog of proven circuits for this application.

In some instances it will be necessary to change resistance and/or capacitance values slightly to compensate for differences in solid-state devices, but the values given in each diagram are adequate as starting points in developing a properly working stage. Although specific types of transistors, diodes, and ICs are suggested, others of similar characteristics can be substituted in most instances. The important considerations will be the beta, f_T, maximum voltage and maximum current ratings. When a substitution is made, the device characteristics should match as closely as possible those of the semiconductor specified in the diagram.

Most of the circuits offered here can be "married" to one another by following common-sense practices. Final optimization can be effected by adjusting bias and signal-input levels to assure good performance.

Audio Amplifiers

Fig. 1 contains a collection of single-stage audio amplifiers. Examples are given in illustrations A to D which show the use of npn and pnp transistors with positive or negative ground polarity. It can be seen that either kind of device can be used with either ground polarity



by changing the positions of some resistors and capacitors. The rule applies to any circuit, dc, rf or audio.

Practical examples are given for FET audio stages of the junction (JFET) and insulated-gate (MOSFET) types of transistors. The choice of device for a circuit will depend mainly upon what is available in the parts bin, and whether or not a low- or high-impedance input circuit is needed.

Fig. 2 provides circuits for direct-coupled audio amplifiers. These audio blocks provide higher gain than can be obtained from the single stages of Fig. 1. It is important to remember that the higher the gain the greater the occasion for instability. Therefore, it is imperative that the layout be orderly, with isolation assured between the input and output ports of the blocks. Short leads and proper bypassing are essential. Approximate input and output impedances are given for each amplifier circuit.

Fig. 3 illustrates a circuit which contains an af preamplifier and IC PA stage. Such a module is suitable for receivers that must draw minimum dc current from the power supply.

Fig. 4 shows a direct-coupled three-stage audio amplifier block which uses dc feedback for stabilization. Q3 operates in Class A to assure low distortion. However, the idling current of this amplifier is relatively high – approximately 450 mA, which does not make it ideal for battery-operated equipment in the field. It is fine for mobile work from the car, however.

Fig. 5 shows a circuit which will deliver 3.5 watts of audio output. A preamplifier supplies af energy to an op amp which serves as a driver for the complementary-symmetry output transistors. This is a low-distortion amplifier which is suitable for home-station amateur receivers.' The power-output transistors must be attached to a heat sink. A metal chassis will be adequate as the heat sink.

For applications calling for high audio power and low distortion the circuit of Fig. 6 is excellent. The design was done by RCA (RCA transistors are specified). Q5 and Q6 must be installed on heat sinks to ensure damage-free operation.

RF and I-F Amplifiers

A group of individual rf amplifiers is shown in Fig. 7. The frequency range listed on the drawings is 1.8 to 30 MHz. It should be stated that all of the circuits can be made to operate anywhere from a few kHz to as high as 50 MHz, and they can be used as rf or i-f amplifiers. The major consideration in changing any of them for use on some other frequency of interest is that of the tuned circuits (*L* and *C* values). Furthermore, the coupling and bypass capacitors may need to be changed in value

Fig. 2

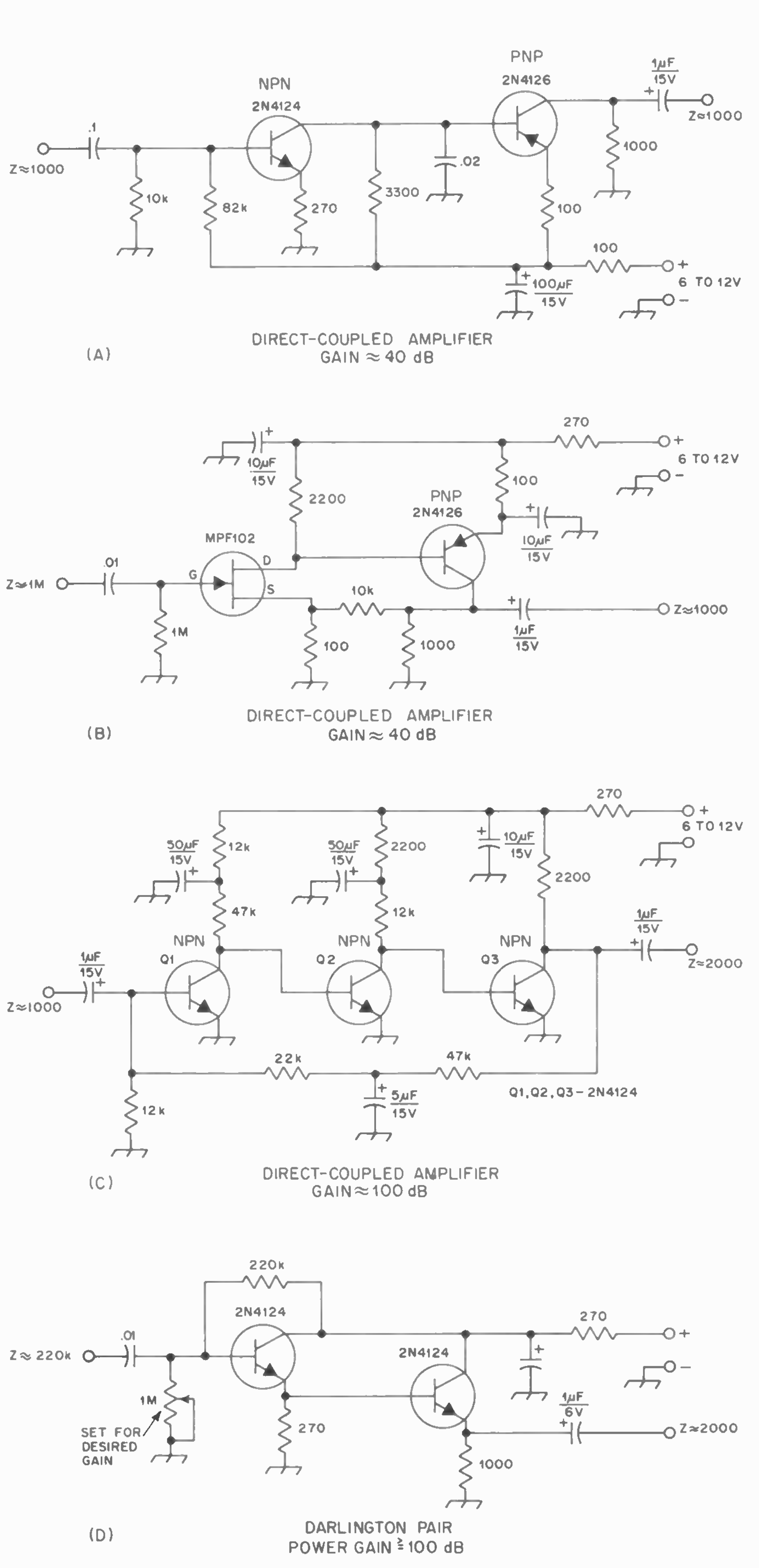


Fig. 3

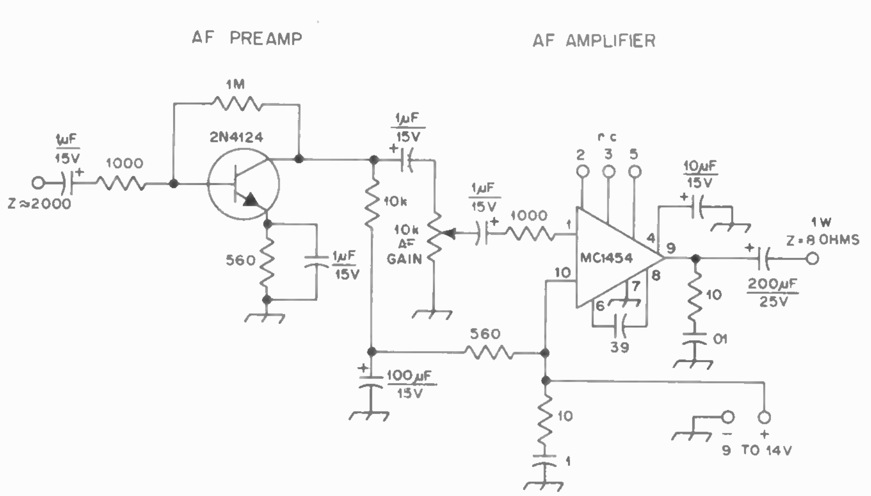
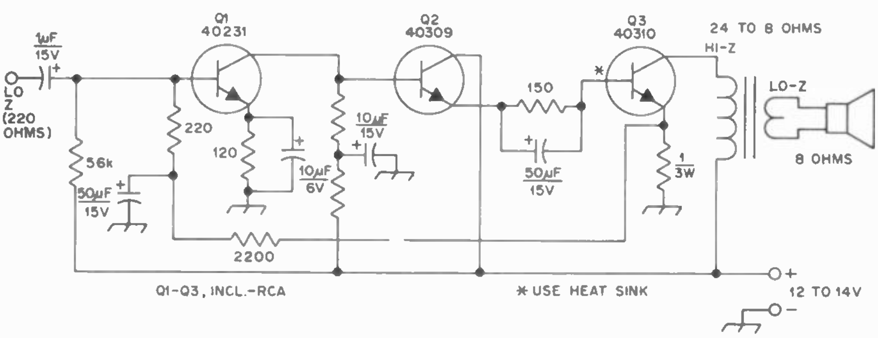
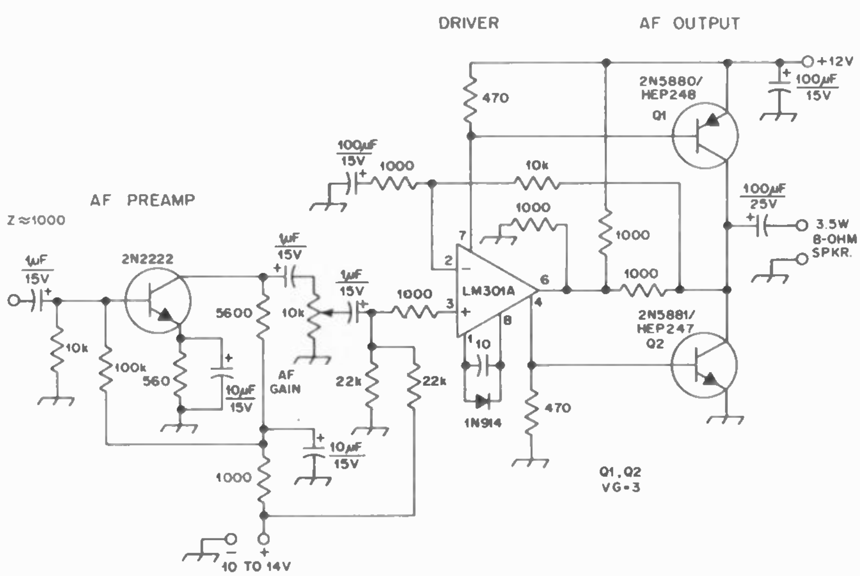


Fig. 4



NEGATIVE DC FEEDBACK
2-W CLASS A AMPLIFIER

Fig. 5



to assure effectiveness at the chosen operating frequency.

Although there are numerous ICs which lend themselves admirably to rf and i-f amplification jobs, only two of the more popular ones are shown in Fig. 8. Both have excellent agc characteristics, but the circuit at B has slightly greater dynamic range and gain than that at A. Information on other brands and types of rf ICs can be obtained from the manufacturer's data sheets, books, and application notes.

Mixers

There are a number of amateur applications for which simple mixers can be used. Notable among the uses is that of a mixer in a simple receiver for field work, or in a heterodyne type of transmitter. Bipolar-transistor mixers exhibit good conversion gain at low oscillator-voltage injection levels, but they are more prone to IMD and overloading at high signal levels than is the case with diode or FET mixers. The JFET and MOSFET devices make excellent unbalanced mixers and they are quite simple to wire into a circuit. A collection of discrete mixers is offered in Fig. 9. As was true of the rf amplifiers shown in Fig. 7, the mixers can be used for any frequency from a few kHz to 50 MHz and higher. Tuned-circuit constants, coupling, and the bypass capacitors must be selected according to the frequency of operation.

Fig. 10 contains an assortment of balanced mixers, all of which will offer better performance than their discrete counterparts. The principal feature of a balanced mixer is the reduction it affords in leakthrough to the i-f of even-order harmonics of one input – usually the local oscillator. The doubly balanced mixer reduces spurious output caused by the signals supplied to both input ports. The balanced JFET mixer of Fig. 10B is superb in its large dynamic range – approximately 130 dB. Dc balance is effected by means of the control in the source leads of the FETs. Conversion gain from any of these balanced mixers is quite good, whereas with the passive mixer of Fig. 11A there is a significant insertion loss which must be compensated for by an additional stage of gain.

The CA3028A balanced mixer of Fig. 10C needs no external balancing controls if a symmetrical layout is used. This results from the nearly identical characteristics of the differential pair of transistors on the chip substrate, brought about through the devices being formed at the same time from the same silicon crystal, and under the same conditions.

Two doubly balanced mixers are illustrated in Fig. 11. At A, four hot-

Fig. 6

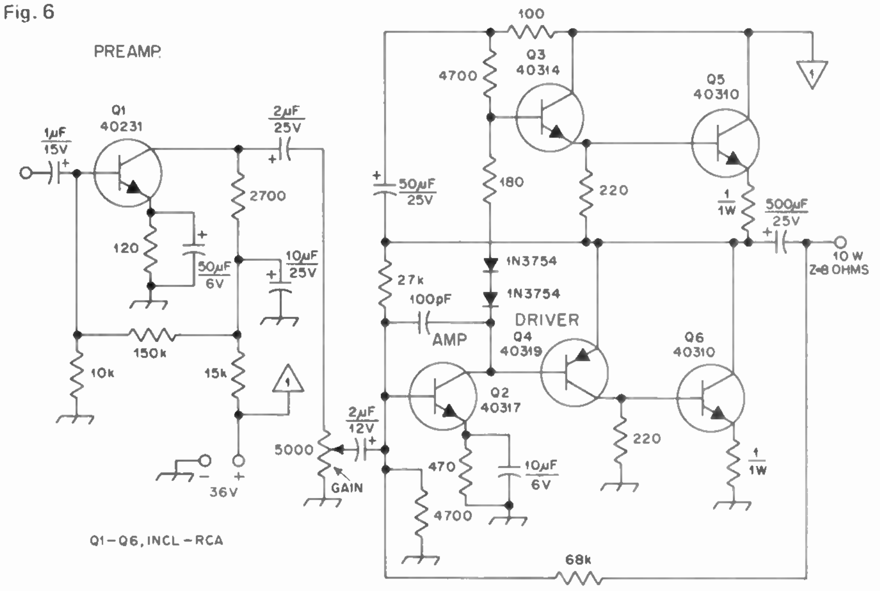
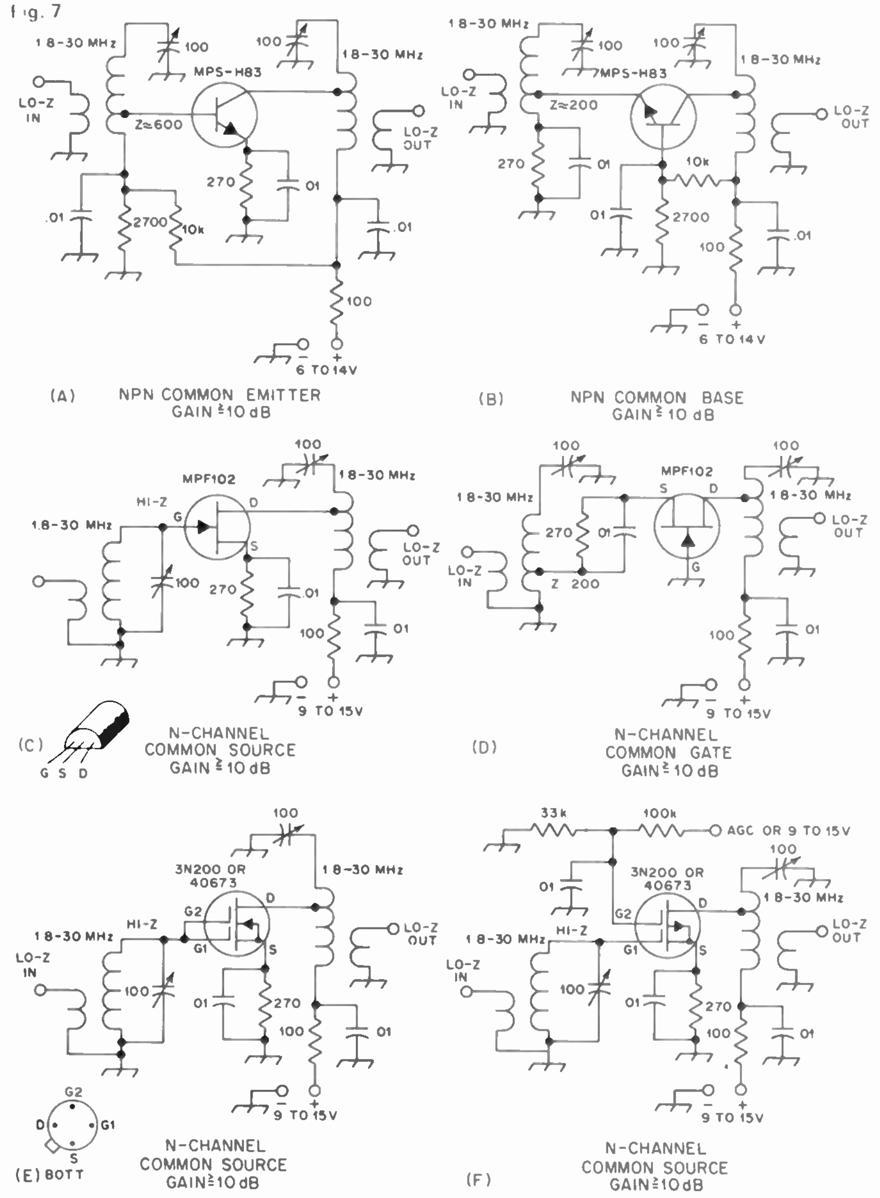


Fig. 7



carrier diodes are used in a broadband mixer configuration. A matched set of 1N914 diodes can be used also, but hot-carrier diodes offer superior performance. Mixers of the type shown at A are suitable for operation from 500 kHz to 500 MHz, provided the broadband transformers (T1 and T2) are wound properly to assure correct frequency characteristics and symmetry. Generally, ferrite toroid cores with a permeability of 125 are used for the transformers. Insertion loss with the diode-quad mixer will be approximately 8 dB. The inherent balance of this mixer will provide some 50 dB of reduction of the local-oscillator signal at the output ports.

A doubly balanced active mixer is presented in Fig. 11B. It contains an IC which provides two bipolar-transistor differential pairs, and each pair has its own current-sink transistor in the emitter return leg. The four transistors of the differential pairs are cross-connected externally to form a doubly balanced mixer configuration. Push-pull oscillator injection voltage is fed to the bases of the current sinks. An internal shield separates the differential pairs, and the IC is rated to 500 MHz.

Detectors

A-m, ssb/cw and fm detectors are cataloged in Figs. 12 and 13. Detectors are nearly identical to mixers in hookup and performance. The essential difference lies in the fact that detectors usually provide audio-frequency output instead of rf energy. The passive detectors which contain diodes will not exhibit gain as a result of the conversion process. Therefore, they are best suited to circuits which have a reasonable amount of rf and i-f amplification ahead of them. In most instances it is necessary to follow a diode detector with a low-noise af preamplifier.

Active product detectors are shown in Fig. 13 at B and C. With this type of detector it is possible to eliminate an audio-frequency post-detector amplifier, as the detector has considerable conversion gain. Furthermore, the BFO injection voltage requirements are much lower than for a diode detector.

A collection of diode fm detectors is presented in Fig. 14, A through C. They follow the timeless classic discriminator and ratio-detector formats. At D is an active quadrature detector for fm. It uses an IC which contains a limiter and squelch circuit. When using the CA3089E chip it is necessary to keep all circuit leads as short as possible, and bypassing (where needed) must be done as close to the IC pins as practicable. If these precautions are not taken there will be instability, and that will spoil the performance of all three chip functions.

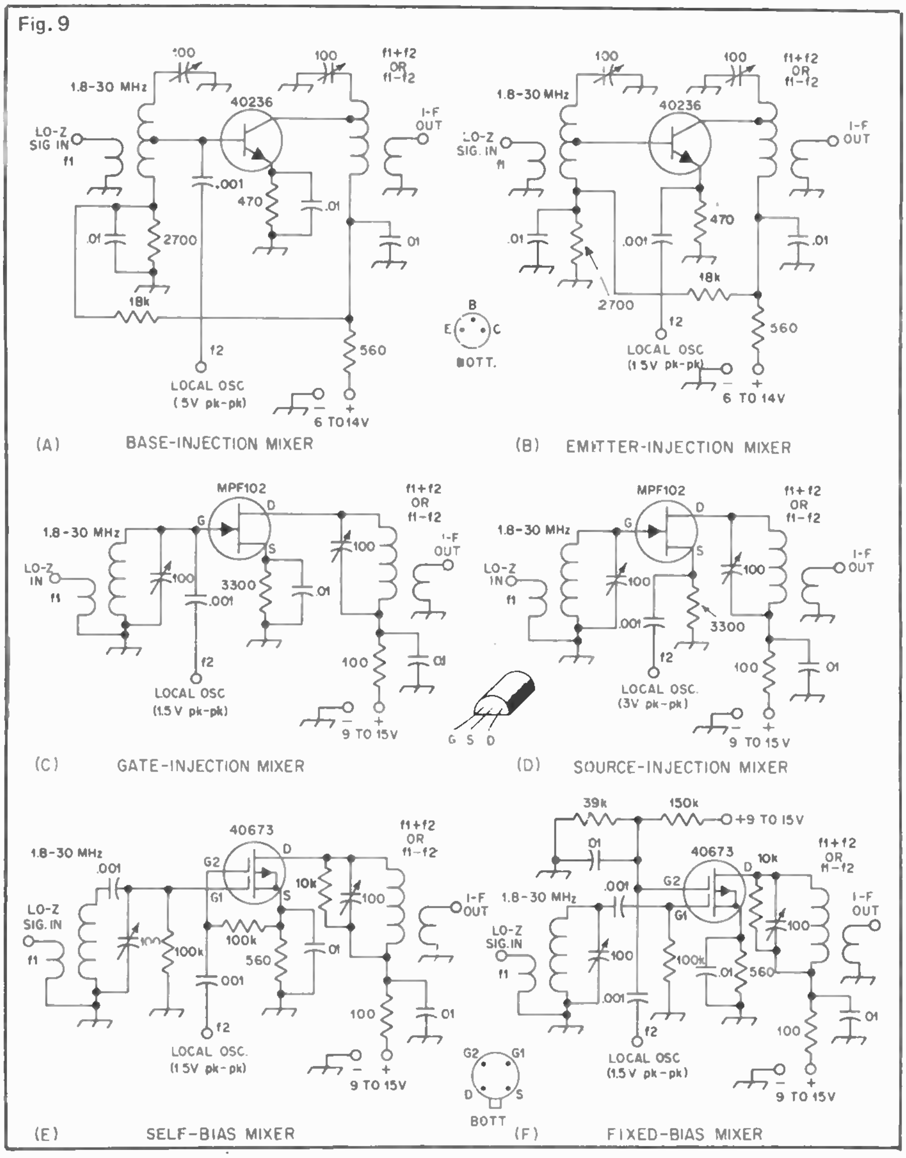
Oscillators

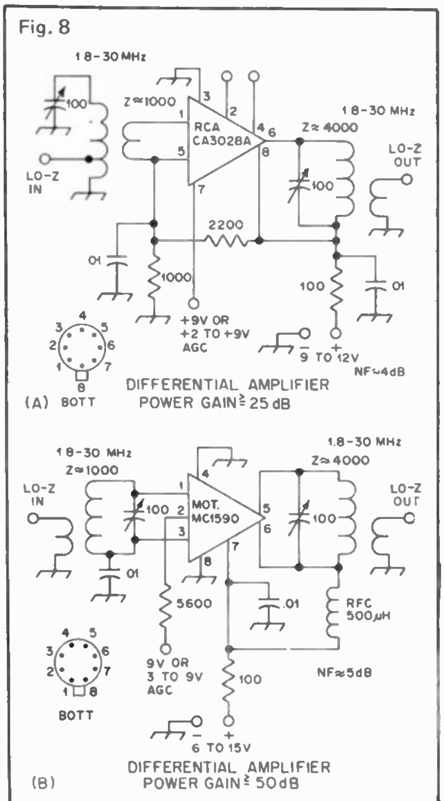
Fig. 15 lists numerous crystal-controlled oscillators. Although specific frequencies of operation are suggested, most of the circuits can be modified for use on other frequencies. Changes will have to be made in feedback capacitor values, and the tuned circuits will need modification for frequencies other than those suggested. In all examples, the size of the output-coupling capacitor should be kept as small in capacitance as possible to avoid frequency "pulling." It is better to take small amounts of energy from the oscillators, and then build up the power level by means of subsequent amplifiers.

Various self-excited oscillators are presented in Fig.16. Polystyrene capacitors are specified in the frequency-determining portions of the circuits. These capacitors exhibit excellent temperature stability, and surpass dipped silver micas in most oscillator applications. Excellent frequency stability can be obtained from all of the circuits shown, but the oscillator at E has been proved the most stable of the assortment during ARRL laboratory tests. As is true of crystal-controlled oscillators, the output-coupling capacitance amount should be kept as low as possible to assure good stability.

Buffers for Oscillators

Fig. 17 contains workable examples of active buffers for use after crystal or self-excited oscillators. The circuits at A and B are broadband types with only





slight gain. They serve mainly to isolate the oscillator from the load which is connected to it. Changes in load (transmitter stage being keyed, for example) will often be reflected to the oscillator, thereby shifting its frequency. Buffer stages help to prevent this condition.

Designing the output buffer stage for 50 ohms at the signal take-off point is useful. Even though the buffer output connects to a high-impedance external circuit, the low output impedance of the last buffer will be capable of permitting ample driving voltage or power to the driven stage, provided the last buffer is operated as an amplifier. An example of this is seen in Fig. 17C. A 2N2222 or similar transistor is used to build up the signal from the oscillator so that at least 5 volts pk-pk are available across the 50-ohm output port. The low output impedance greatly reduces frequency pulling which can be caused by load changes in higher-impedance external circuits. The collector tank of the buffer/amplifier is a pi network. It aids in reduction of harmonic currents, assuring a 35-dB or greater reduction in the second harmonic, and 45 dB or more reduction of third-harmonic energy. Broadbanding is enhanced by inclusion of a 3300-ohm resistor across the pi-section coil.

Even greater oscillator isolation can be had by operating the VFO or crystal-controlled stage one octave lower than the desired operating frequency. By placing a frequency multiplier after the VFO considerable immunity to pulling will result. An example of a practical circuit is given in Fig. 18. A source-follower FET is located immediately after the oscillator. It has a tuned transformer in the source lead. This enhances the signal level slightly while providing push-pull drive to the push-push frequency doubler. This type of doubler is operated in Class C, and is nearly as efficient as a straight-through Class C amplifier. Some forward bias is

Fig. 10

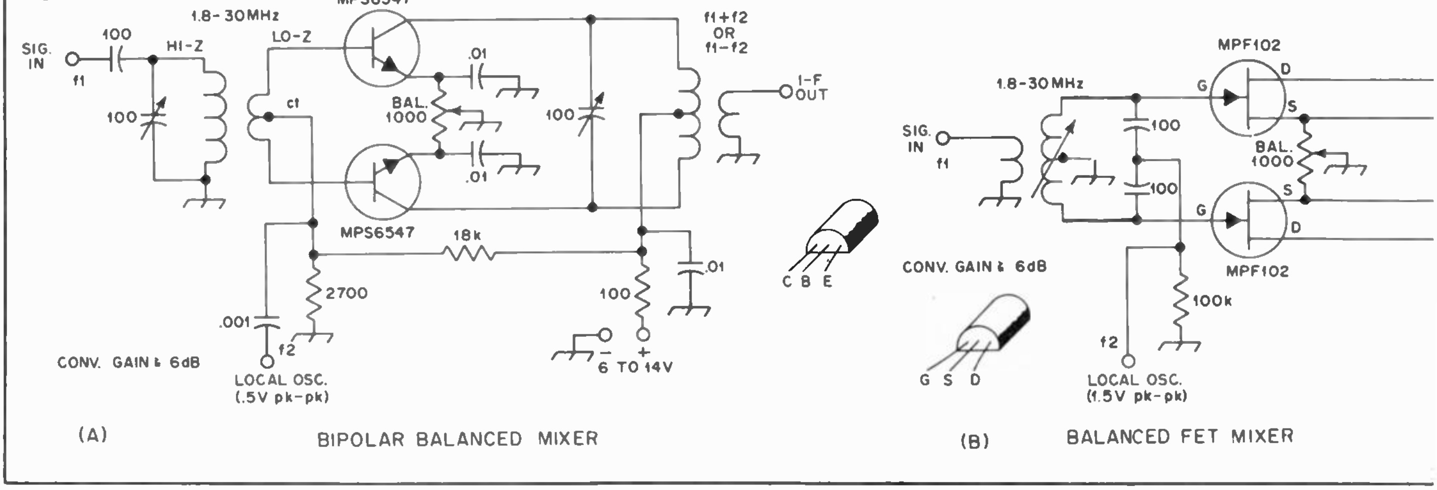


Fig. 11

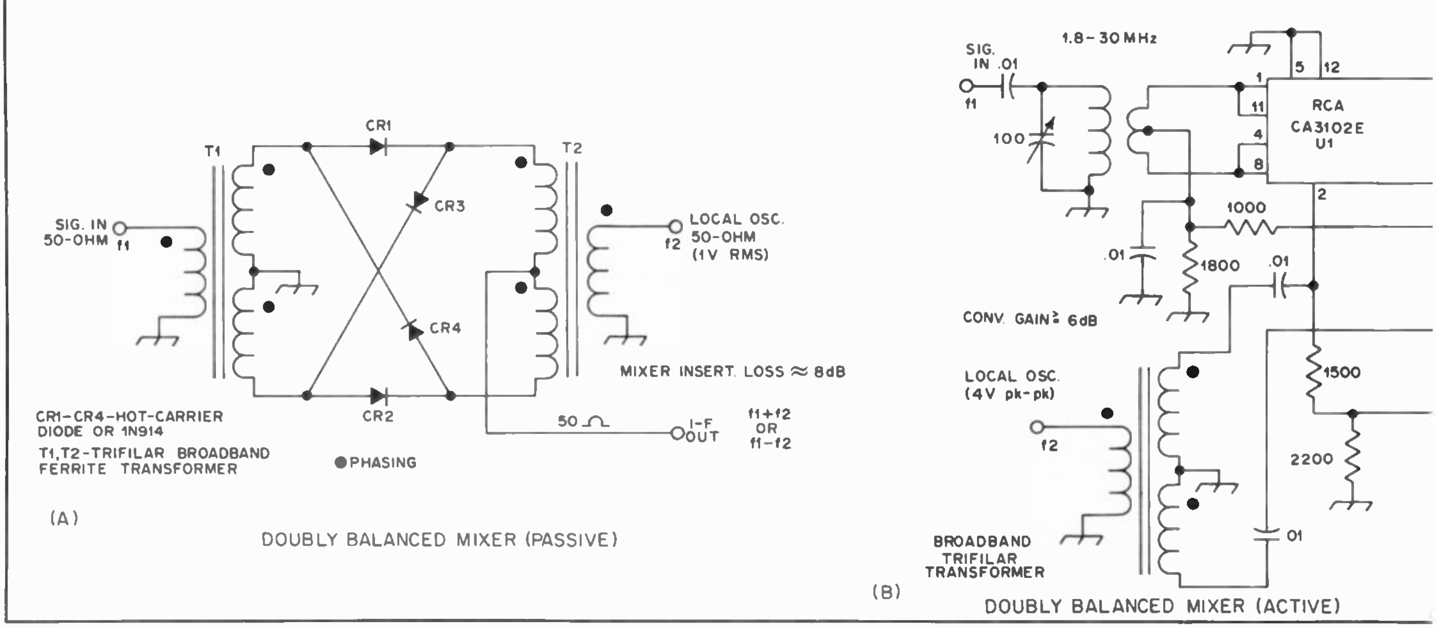
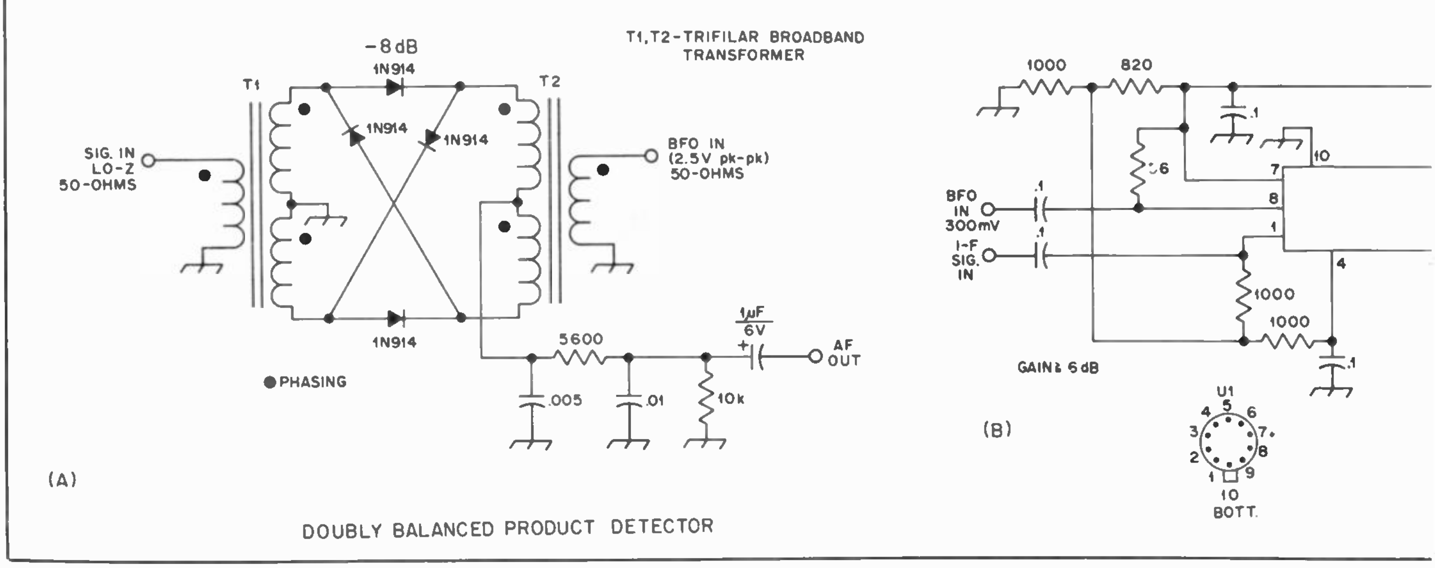
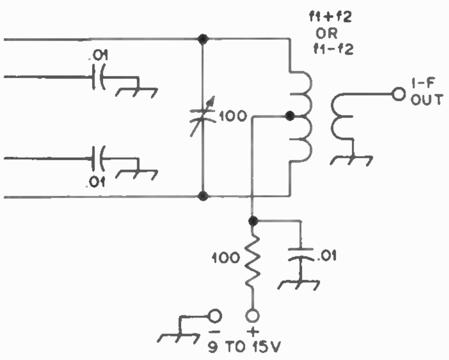
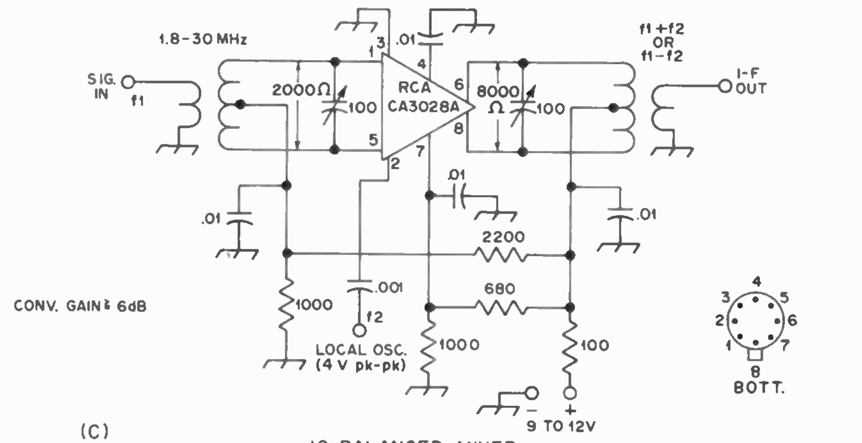


Fig. 13







(C)

IC BALANCED MIXER

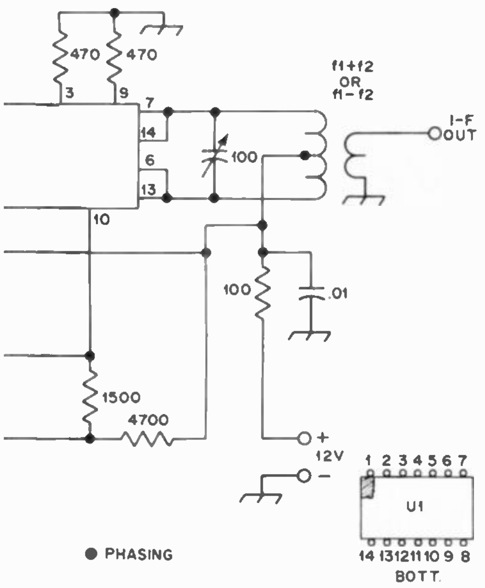
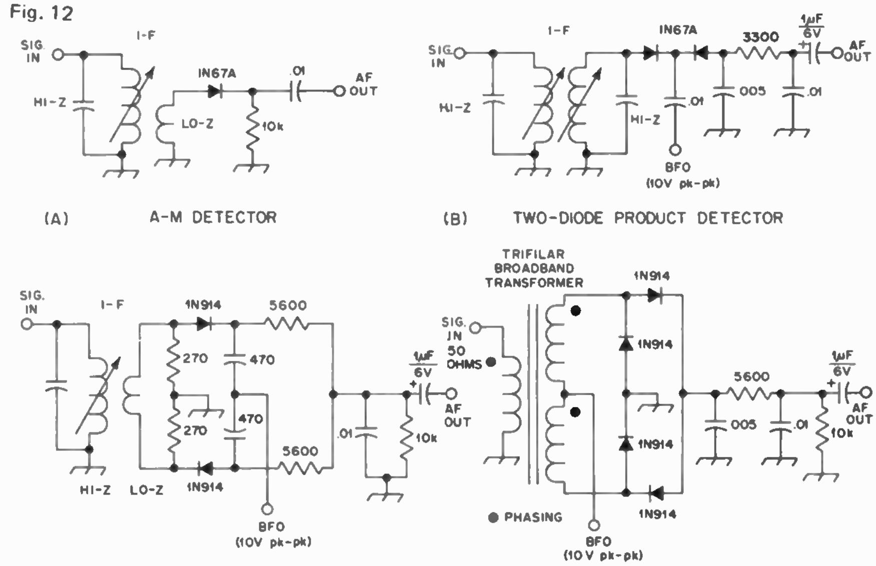


Fig. 12

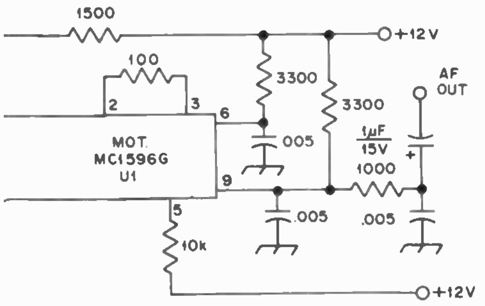


(A) A-M DETECTOR

(B) TWO-DIODE PRODUCT DETECTOR

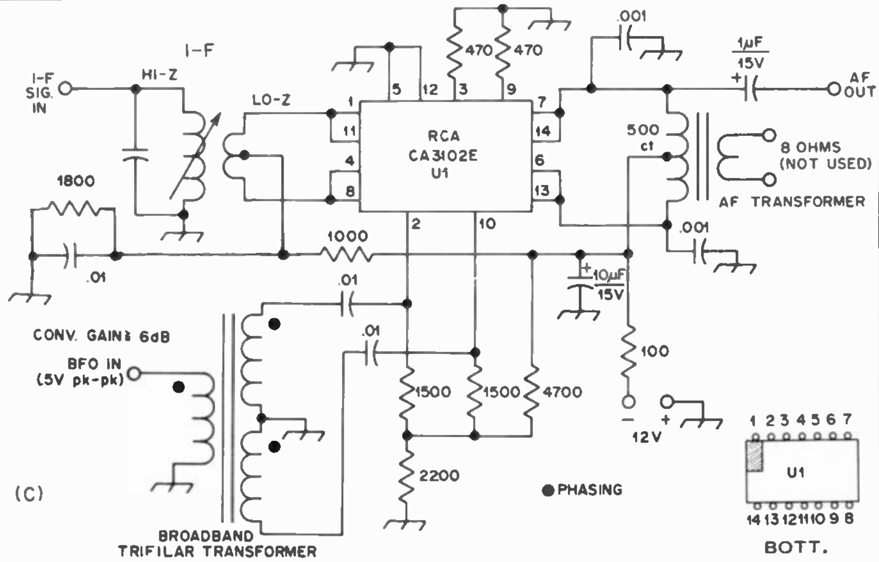
(C) BALANCED DIODE PRODUCT DETECTOR

(D) BALANCED DIODE PRODUCT DETECTOR



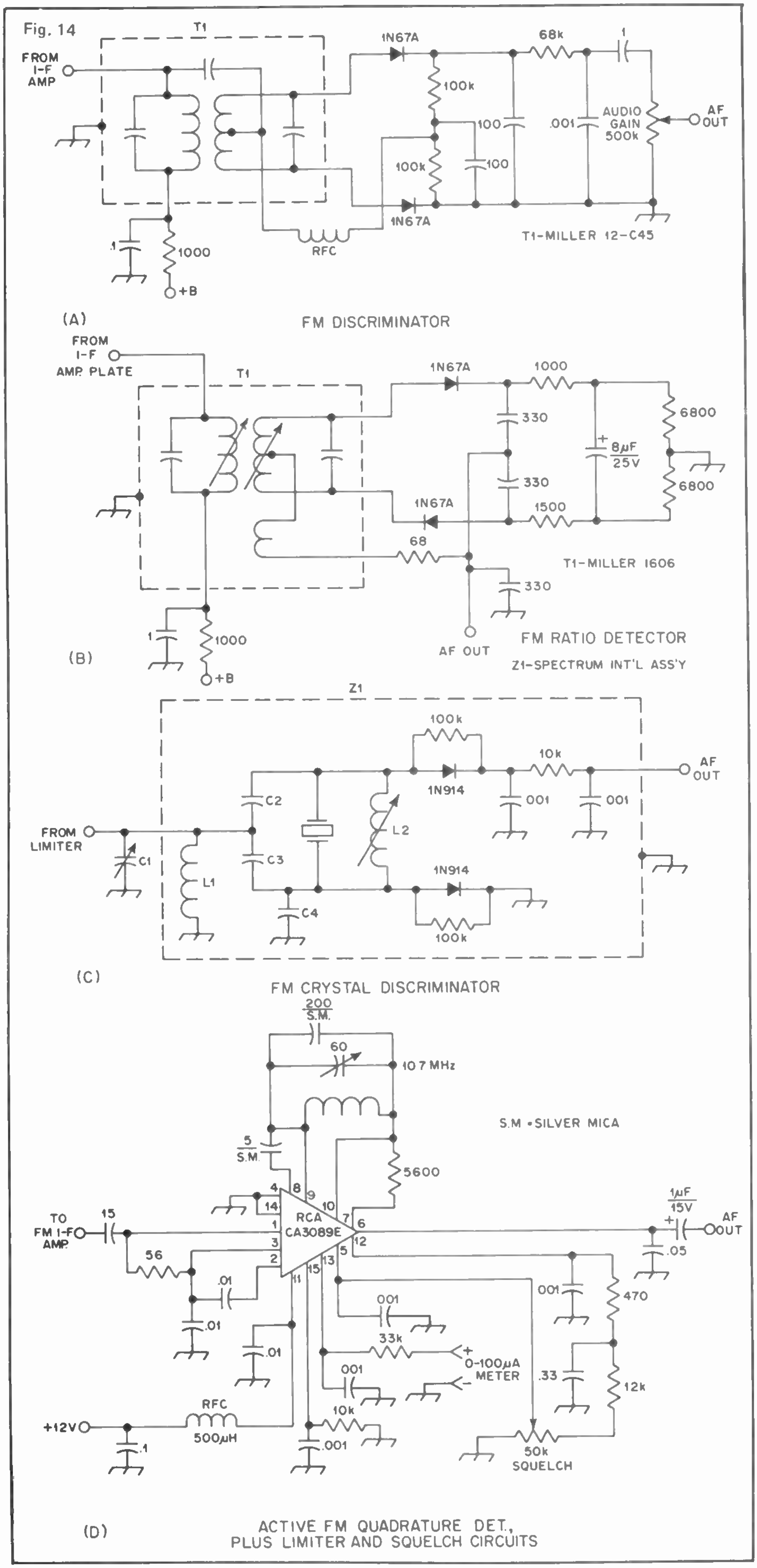
DOUBLY BALANCED ACTIVE

PRODUCT DETECTOR



(C)

DOUBLY BALANCED PRODUCT DETECTOR



applied to the 2N2222s to assure adequate stage gain, but the drive from Q1 is great enough to push the doubler into the Class C region. When separate devices are used in a push-push doubler it is necessary to provide dynamic balance. This is accomplished by means of R1, which is adjusted for best waveform purity at the output of the doubler. A scope can be used during this adjustment, or a probe wire can be placed near the doubler tank, connected to a receiver, and R1 adjusted for minimum response at the *oscillator* frequency. A CA3028A IC would make a good push-push doubler because it contains a pair of bipolar npn transistors of identical characteristics. The source terminal of the current sink should be grounded, and the current sink transistor should be fully saturated by feeding +12 volts to the current sink base through a 2200-ohm resistor. In effect, these measures will have the equivalent effect of removing the current-sink transistor from the IC circuit, leaving only the differential pair to function as a push-push doubler. Numerous transistor-array ICs exist, and many of them are suitable for use as push-push doublers. A properly balanced push-push doubler is capable of providing a nearly pure sine-wave output, even without a tuned tank.

Audio Oscillators

In amateur work it is often necessary to use some form of audio generator to provide a side-tone facility in transmitters. Audio oscillators find common use in code-practice equipment, and there are other amateur circuits which call for the use of audio oscillators. Fig. 19 shows three kinds of audio oscillators. The example at A employs a unijunction transistor, which delivers a saw-tooth wave. From the outlook of simplicity, the circuit is a good one for use in side-tone circuits of portable, compact gear. The circuit at B is simple too, but the audio transformer makes it less compact than the oscillator at A.

Better waveform purity can be had from the twin-T oscillator of Fig. 19C. Changes in operating frequency can be effected by altering the *R* and *C* values in the T network.

Varicap-Tuned BFO

An extremely stable 455-kHz BFO is shown in Fig. 20. A Motorola MV104 Varicap tuning diode (voltage-variable capacitor) is used to change the oscillator frequency. Its capacitance is varied by means of a change in dc voltage applied through a panel-mounted 100,000-ohm control. The circuit costs considerably less than would one in which three 455-kHz crystals were used for cw, upper sideband, and lower-sideband reception in a receiver. Tuning diodes can be used well into the vhf spectrum, but the diode must be selected *for the application* to assure the desired capacitance change and Q.

Bipolar Switches

Dc switching can be done with bipolar transistors. Two circuits are given in Fig. 21. Examples are provided for npn and pnp devices. When the base circuit is open, no current flows through the transistor junction. When the base circuit is closed, the transistor is fully saturated and functions as a closed switch. The transistor dc beta should be high to assure proper switching at high currents. Furthermore, the power and maximum voltage ratings of the transistor used should be well beyond the voltage and current values which must pass through the bipolar switch. Bipolar switches find many applications in amateur work, but are best known for their roles in keying circuits of transmitters – grid-block and cathode keying.

Break-In Delay Circuit

Fig. 22 illustrates a workable cw break-in delay circuit with antenna changeover relay. Two pnp bipolar switches are used at the input. Q1 handles more current than Q2, so a huskier transistor is used in that part of the circuit. Q2 merely supplies dc voltage at low current through its junction, then through a 1N914 diode gate, which enables the dc to charge the variable-delay *RC* network at the base of Q3. When the time-constant network is charged it drives Q3 into saturation and permits current to flow through the field of K1, thereby closing contacts K1B. As the voltage charge decays, Q3 drops out of saturation and K1 opens. The Q1 switch keys a low-level transmitter stage.

The principle of operation is similar to that of a VOX. In fact, rectified audio voltage (minus polarity) could be fed to the bases of Q1 and Q2 for ssb operation, and the circuit would perform as a VOX. There would be no anti-trip facility, however.

Op-Amp Circuits

Perhaps the most universal solid-state device at the builder's disposal is the op amp. The applications for it are so abundant that it would require a separate book to cover the subject adequately. In fact, Jung did just that in his superb 591-page book, *IC Op-Amp Cookbook*, published by Howard Sams.

Some practical amateur kinds of circuits are provided in Figs. 23 and 24. One significant feature of the op amp for audio work is the high gain possible with a single IC. The circuit of Fig. 24 is ideal for use in audio amplifiers which employ push-pull driver or output stages. Most phase inverters have unity gain or less, whereas this circuit can

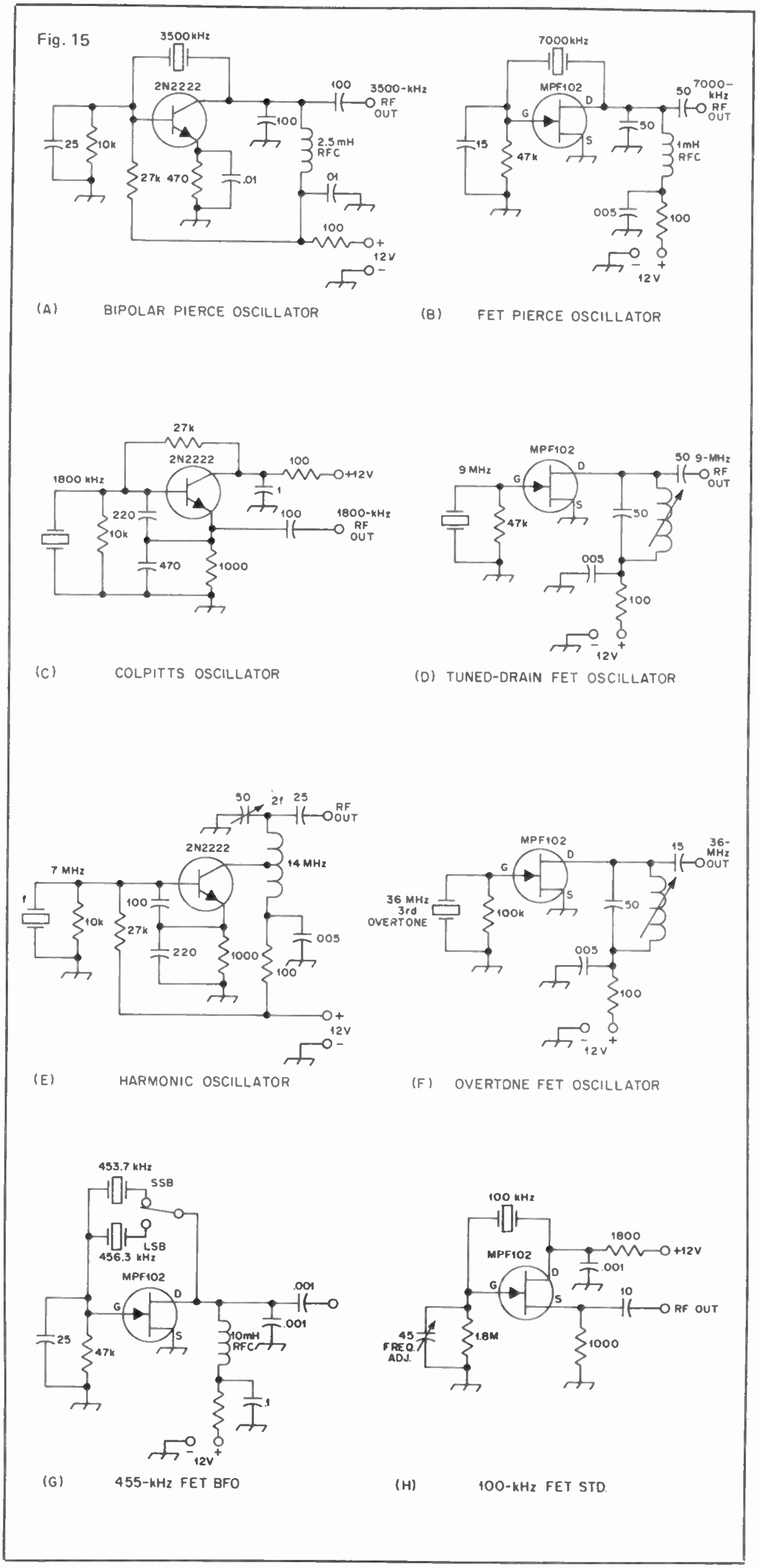
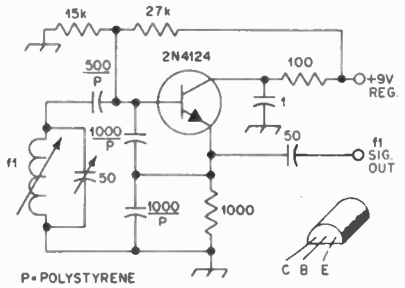
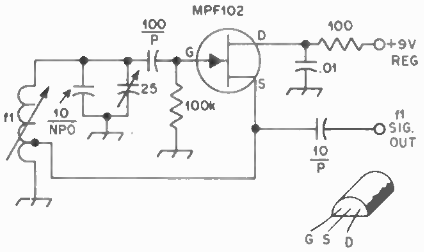


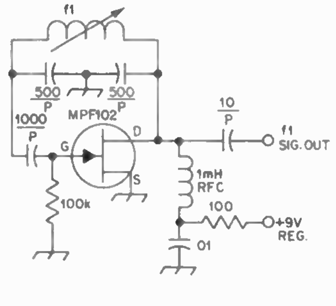
Fig. 16



COMMON-COLLECTOR COLPITTS VFO
(A) (1.8 OR 3.5 MHz)

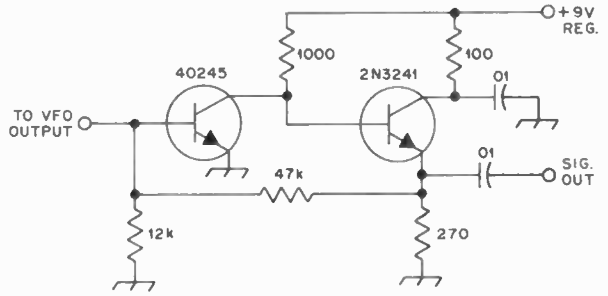


COMMON-DRAIN HARTLEY VFO
(B) (5 OR 7 MHz)

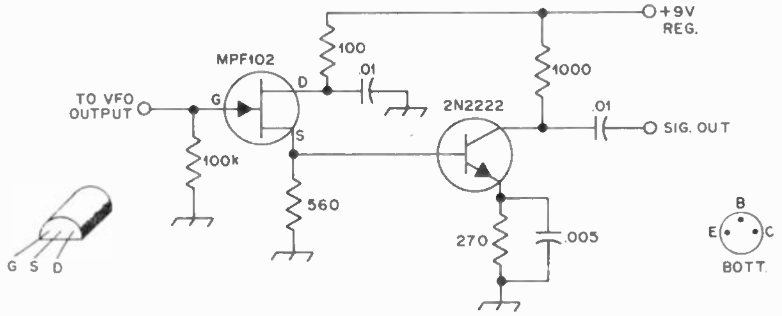


COMMON-SOURCE HARTLEY
TUNED-FIXED OSCILLATOR
(C) (1.8, 3.5, 5.0, OR 7 MHz)

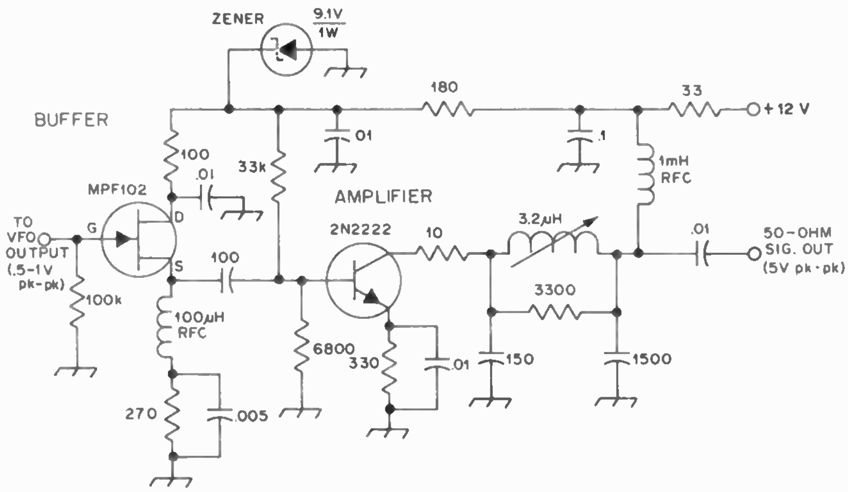
Fig. 17



(A) DIRECT-COUPLED VFO
BUFFER WITH DC FEEDBACK

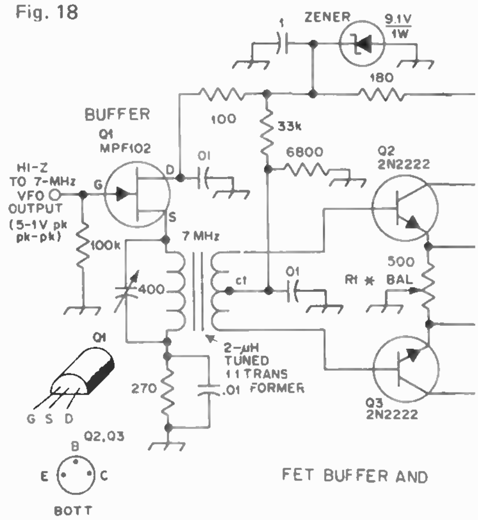


(B) DIRECT-COUPLED
HI-Z INPUT VFO BUFFER



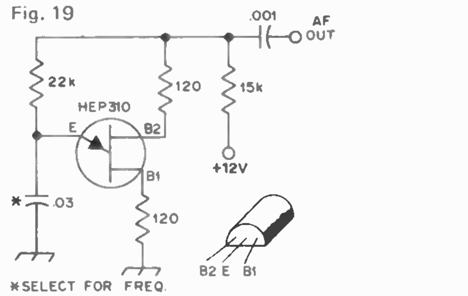
(C) BUFFER/AMPLIFIER FOR VFO
(7-MHz VALUES SHOWN)

Fig. 18

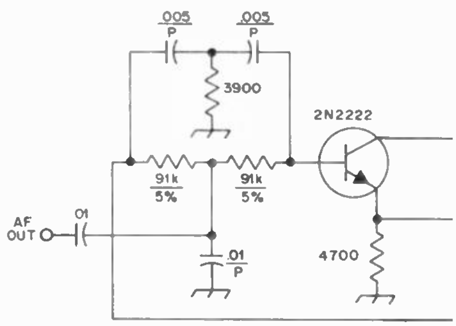


FET BUFFER AND

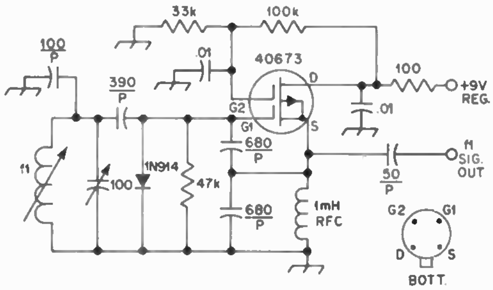
Fig. 19



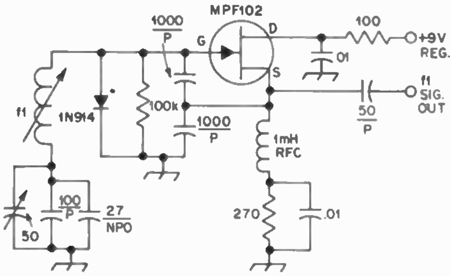
(A) 1-kHz UNIJUNCTION
AF OSCILLATOR



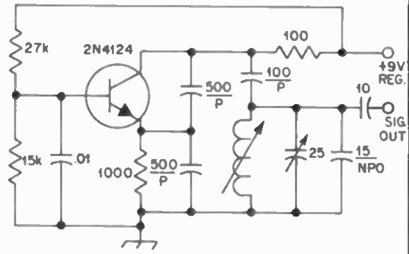
(C)



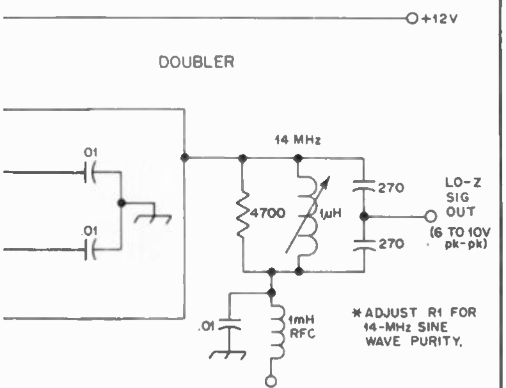
(D) COMMON-DRAIN COLPITTS (3.5-4.0 MHz)



(E) SERIES-TUNED COLPITTS VFO (1.8 MHz)

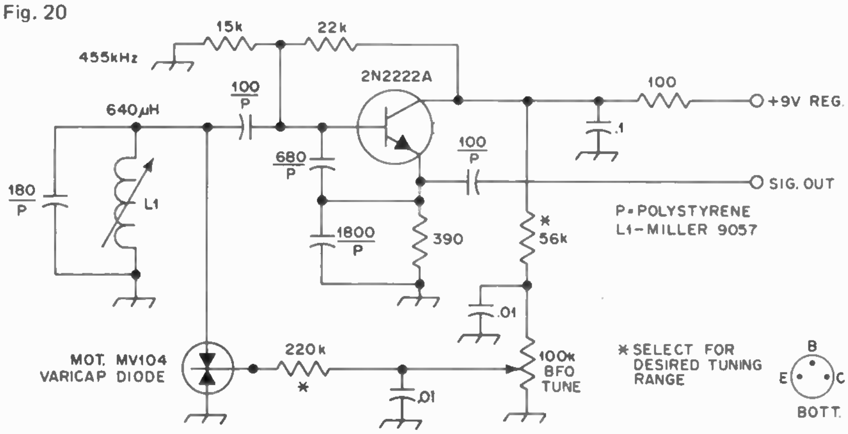


(F) COMMON-BASE COLPITTS VFO (5 OR 7 MHz)



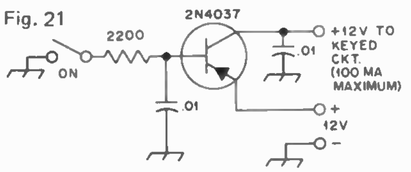
PUSH-PUSH DBLR/AMP FOR VFO

Fig. 20

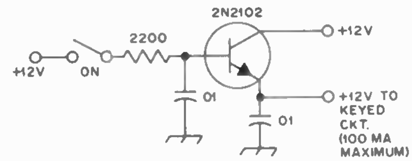


VARICAP-TUNED 453 TO 457-kHz VFO

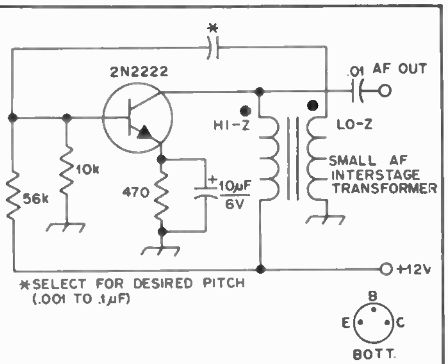
Fig. 21



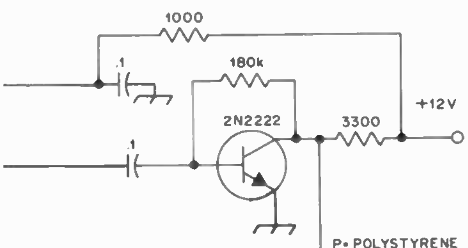
(A) PNP BIPOLAR DC SWITCH



(B) NPN BIPOLAR DC SWITCH

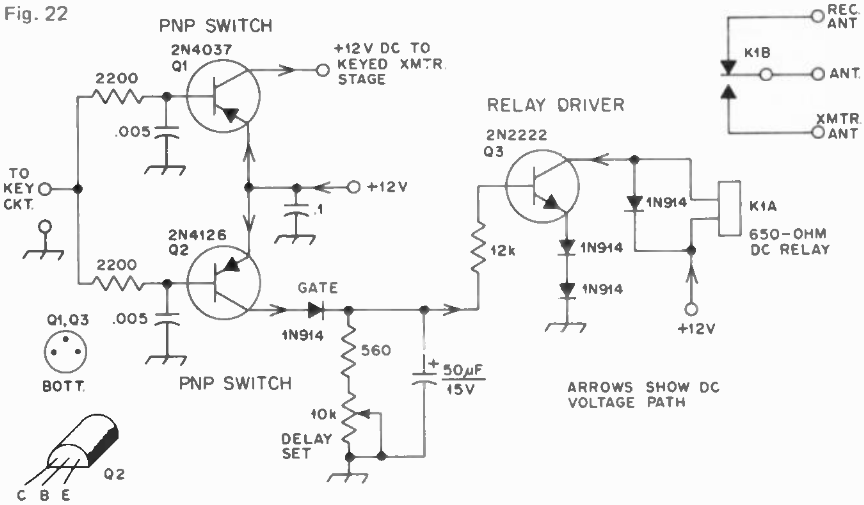


(B) AUDIO OSCILLATOR



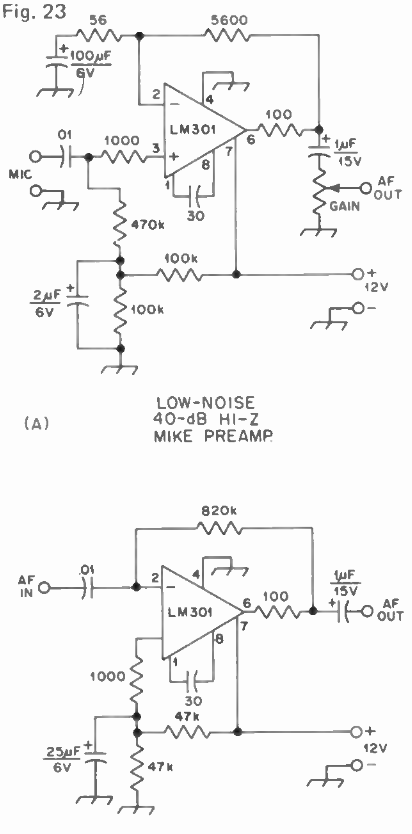
STABLE 700-Hz TWIN-T AF OSCILLATOR

Fig. 22



CW BREAK-IN DELAY CKT

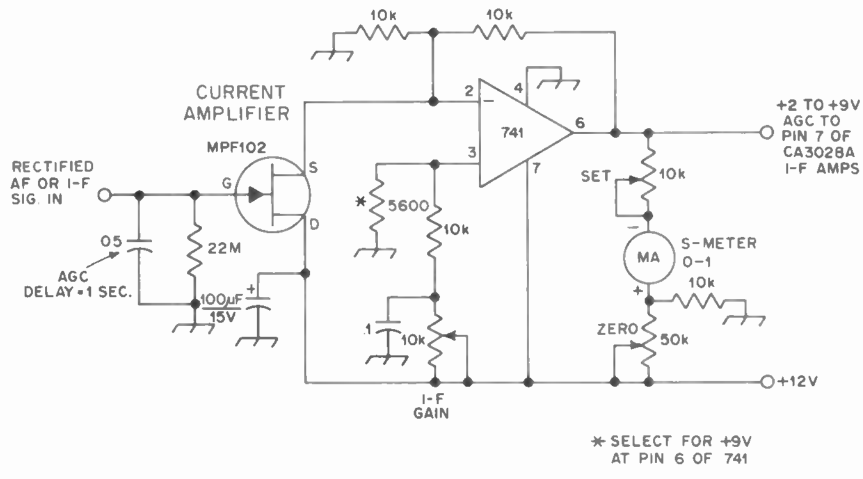
Fig. 23



(A) LOW-NOISE 40-dB HI-Z MIKE PREAMP

(B) LOW-NOISE 40-dB AUDIO AMPLIFIER

Fig. 25



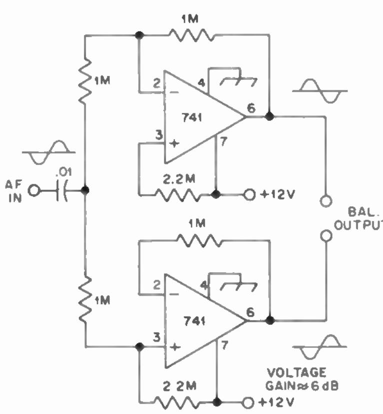
AGC AND S-METER AMPLIFIER FOR CA3028A I-F AMPLIFIER STRIP

effect a 6-dB signal increase while supplying two output af voltages which are 180 degrees out of phase with one another.

Fig. 25 illustrates a practical circuit for controlling the agc of a CA3028A i-f amplifier strip. The op amp responds to changes in voltage across the FET source resistor (10,000 ohms) at pin 2 of the 741. The changes in current of the FET are brought about by shifting the level of the positive dc voltage at the FET gate. This circuit requires a gate-voltage change of 0 to +2 volts for full agc and S-meter range. The agc can be audio or rf derived. A 30-dB preamplifier should be used between the i-f or af signal takeoff point to assure ample dc voltage at the gate of the FET. A 1N914 half-wave rectifier will suffice between the preamplifier and control circuit of Fig. 25.

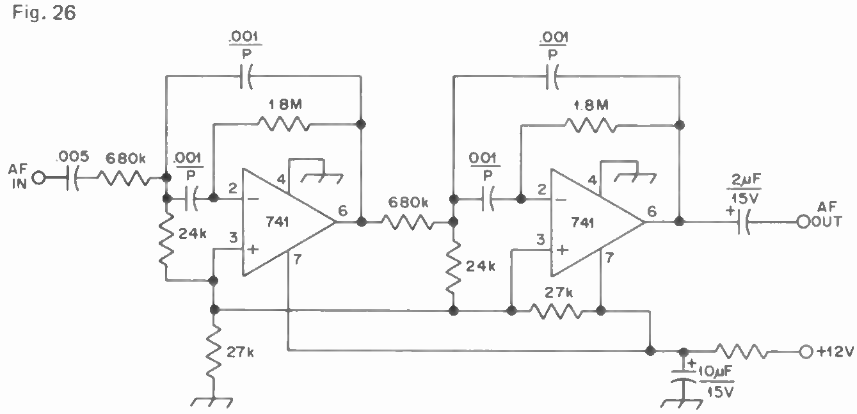
An *RC* active peak audio filter is shown in Fig. 26. Two bandpass-filter poles are employed to give a relatively sharp skirt response. Polystyrene capacitors are specified in the frequency-determining part of the circuit because they are high-*Q* types (desirable) and because they are temperature stable. It is important that the capacitors be closely matched in value, as is true of the resistors in the frequency-determining network. This will assure proper nose response of the filter curve. If the values are not closely matched there will be some offsetting of the peak responses. This will broaden the peak and cause passband ripple (dips between peaks). Generally speaking, 5-percent resistors will be adequate. Filter performance will be best if the circuit is inserted after a low-level audio stage of a receiver. It should be situated just after the detector or first af amplifier.

Fig. 24



OP AMPLIFIER AF PHASE SPLITTER

Fig. 26



P = POLYSTYRENE

840-Hz RC ACTIVE CW AUDIO BANDPASS FILTER

Chapter 9

Construction and Testing Data

Many significant achievements carried out by radio amateurs are products of the workshop effort. This chapter contains tables and other data which should be useful to any amateur who devotes his talents to the testing and building of equipment. Various charts and diagrams are supplied in this part of the book so that a quick reference is available for those who engage in construction work. The last section of this chapter contains numerous tables which will be useful as handy references of material characteristics. Elsewhere in this section there are detailed data on techniques used in the workshop – silver plating, pc-board fabrication and heat-sink assembly.

Winding Toroids

Small toroid cores can be hand-wound with ease, looping the turns of wire through the core and drawing them up snugly as the process continues. However, when a great many turns must be placed on a large core the task can become a tedious one if manageable lengths of wire are fed through the core by hand, then spliced to the next length that will be added. A bobbin is useful for threading the wire through the core, as a large number of coil turns can be stored on the bobbin prior to winding.

Fig. 5 shows the details for making a bobbin. A long, slender piece of thin material (plastic or pc board) is notched at each end. The bobbin is wound with the required amount of wire to fulfill the needs of the toroid winding job. The bobbin is passed through the center of the toroid, brought out, then passed through the core again and again until the required number of turns have been placed on the core.

Cable Neatness

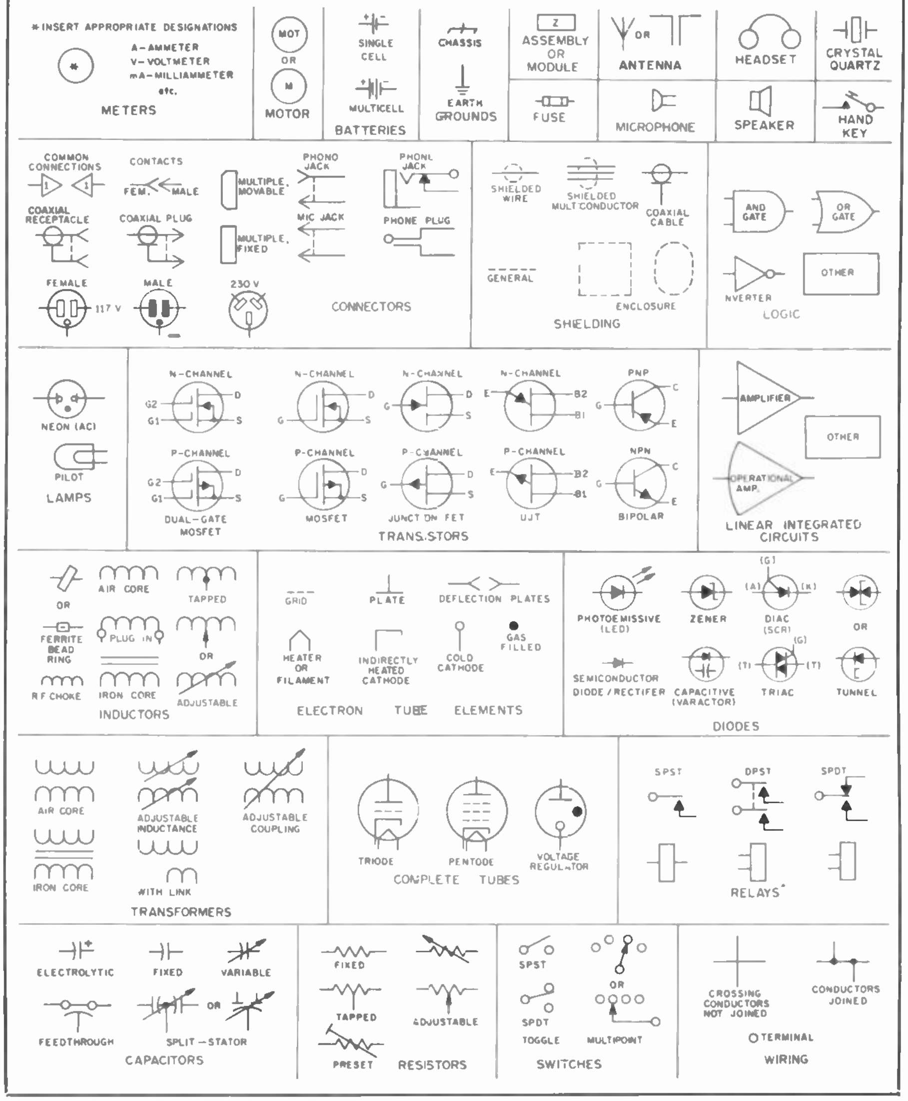
Many items of amateur equipment require a large number of circuit wires which are routed in a group around a chassis. Also, cables which are used between equipment and power sources can contain many individual wires. It is convenient and desirable to wrap or lace these bundles of wire to assure neatness and minimize stress on any one wire of the group. Fig. 6 shows the right and wrong ways to lace a cable. The example at A will loosen up under stress and become ineffective. The method shown at B is better, and will suffice for all but the most rugged of requirements. The most secure of the techniques is that of Fig. 6C. Waxed linen cord is available for lacing cables, but nylon or silk fishing line can be used as a substitute.

Heat Sinks

It is not difficult to make one's own heat sinks from pieces of scrap copper, brass, or aluminum. Fig. 7 shows three kinds of heat sinks that can be fashioned by hand in the amateur workshop. Illustrations A through D show the progression for forming a high-density sink suitable for use with high-power transistors. A coating of silicone grease should be applied to the mating surfaces, and between the transistor body and the heat sink. This will assure efficient transfer of heat from one metallic part to another. The heat-sink sections should be drawn tightly together.

In Fig. 7 at E and F are drawings

Fig. 1



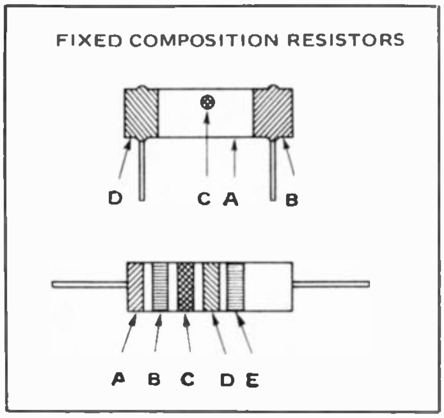


Fig. 2 — Color coding of fixed composition resistors. The color code is given in Table 5. The colored areas have the following significance:

A — First significant figure of resistance in ohms.
B — Second significant figure.
C — Decimal multiplier.
D — Resistance tolerance in percent. If no color is shown the tolerance is ±20 percent.
E — Relative percent change in value per 1000 hours of operation: brown, 1 percent; red 0.1 percent; orange, .01 percent; yellow .001 percent.

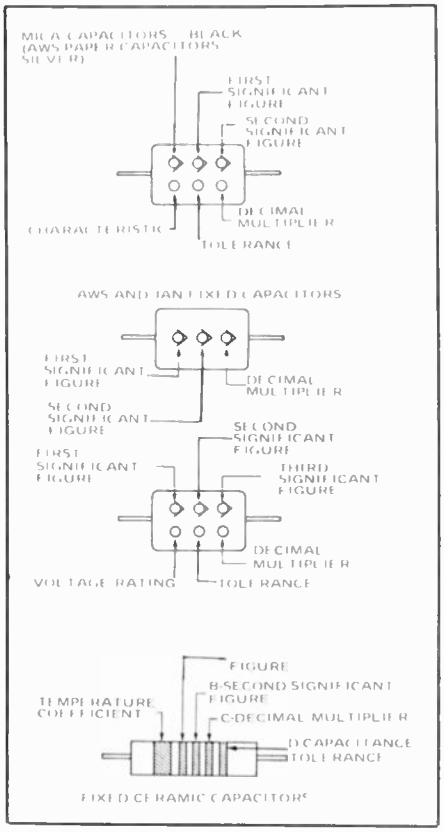


Fig. 3 — Color coding of fixed mica, molded paper and tubular ceramic capacitors. The color code for mica and paper capacitors is given in Table 5. Table 5 gives the color code for tubular ceramic capacitors.

which illustrate how a small heat sink can be made for use with TO-5 types of transistors. The sinks can be formed over a drill bit of the appropriate OD to match the OD of the transistor case. Pinching the metal tightly at the flanges will form the material around the drill bit. The pinching can be done easily in the jaws of a vise.

Aluminum angle stock serves well as heat sink material. An example is given at G in Fig. 7. A small metal plate is used to hold the transistor tightly against the heat sink.

Making PC Boards

Simple one-of-a-kind pc boards can be fabricated in the home workshop with relative ease. When several boards of a given pattern are required, the best method is the photo-etch one. Kits of materials for making negatives are available from Kepro Company, listed in the last chapter of this book. The kits also contain photo-sensitized pc-board material.

For one-shot boards it is quite simple to place the circuit pattern on paper, then trace the pattern onto a section of board material which has a layer of masking tape on the metal side of the board. Carbon paper is placed between the masking tape and the pattern to be traced, then the transfer is made. The next step is to cut away the masking tape in those areas where the copper must be etched. A sharp hobby knife with a pointed blade is excellent for the job.

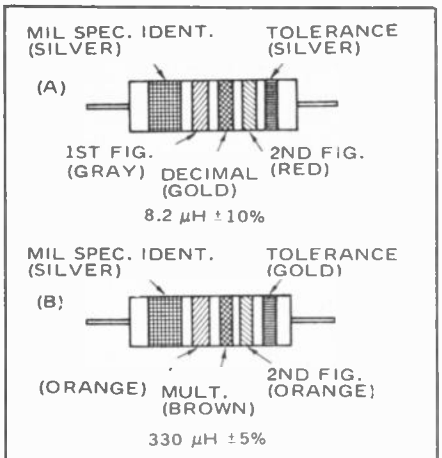
Ferric-chloride etchant solution is recommended for amateur work. It is available from Kepro in quart and gallon amounts. It is among the safest of the etchants an amateur can use, but the user should avoid getting it on his person. If bodily contact is made, the chemical should be washed off immediately. In the interest of safety, *the chemical should be stored out of reach of children.*

Ferric chloride can be used again and again until it is exhausted. When the chemical becomes old enough to throw out, it will be ineffective as an etchant as evidenced by inordinately long time periods being required to etch a pc board. Also, the solution will become almost black in color when it is no longer useful.

The etchant solution should be kept between 90 and 110 degrees F during the etching process. Fresh solutions at some temperature in the prescribed range will permit etching a board in approximately 15 minutes (thin copper) to 30 minutes (thick copper). The chemical should be agitated continuously for best results. A homemade etching stand with provisions for agitation is shown in Fig. 8.

Another method for making a circuit board is to cut numerous squares or rectangles on the copper side of the board. A hacksaw can be used to make isolated pads on the board, or a moto-tool with a cutting bit can be used to grind away the copper material around each pad.

Still another technique for making a simple circuit board is to cut squares or circles from a piece of pc board, retaining both the insulating material and copper foil. The individual pads can then be affixed to an unaltered piece of board material by means of epoxy cement or hot-melt glue. The metal



COLOR	*FIGURE*	*MULTIPLIER*	*TOLERANCE*
Black	0	1	
Brown	1	10	
Red	2	100	
Orange	3	1000	
Yellow	4		
Green	5		
Blue	6		
Violet	7		
Gray	8		
White	9		
None			20%
Silver			10%
Gold			5%

Multiplier is the factor by which the two color figures are multiplied to obtain the inductance value of the choke coil.

Fig. 4 — Color coding for tubular encapsulated rf chokes. At A, an example of the coding for an 8.2-μH choke is given. At B, the color bands for a 330-μH inductor are illustrated.

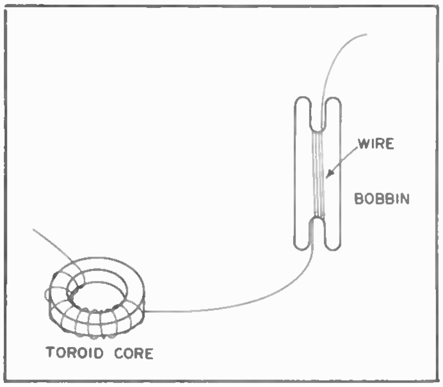


Fig. 5 — Method for using a bobbin to wind wire on a toroid core.

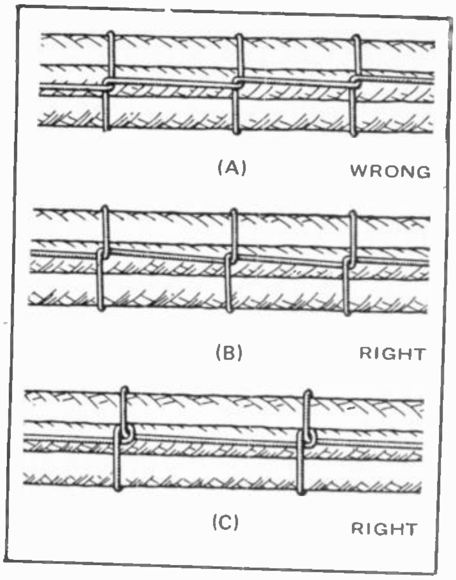


Fig. 6 – Methods of lacing cables. The method shown at C is more secure, but takes more time than the method of B. The latter is usually adequate for most amateur requirements.

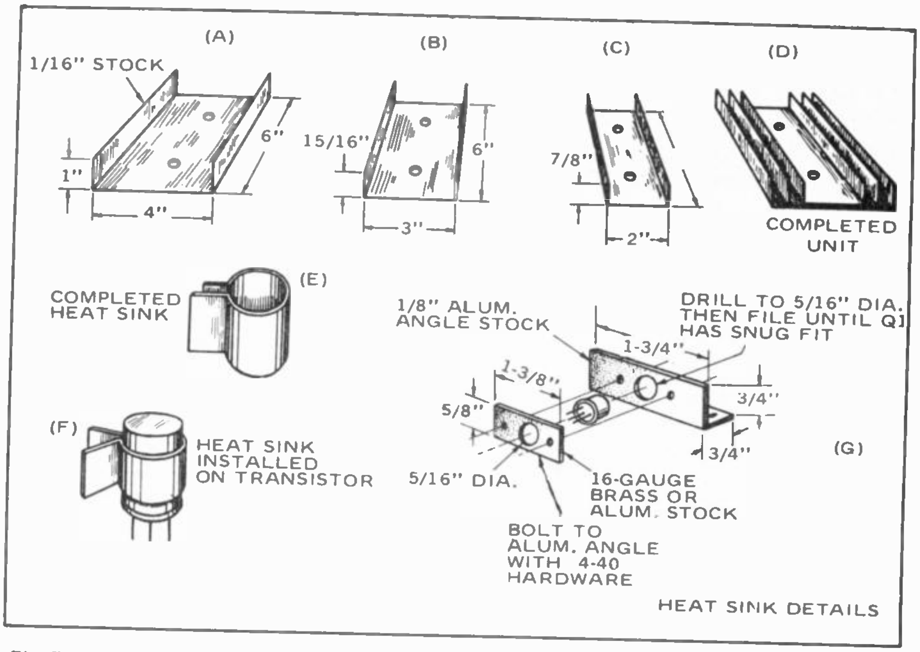


Fig. 7 – Layout and assembly details of another homemade heat sink. The completed assembly can be insulated from the main chassis of the transmitter by using insulating washers.

surface of the main board can be used as the circuit ground.

Silver Plating

Plating can be done in several ways. First, you can take your parts to a plating shop. This costs money, but assures a good job. There are at least three do-it-yourself methods now available, including a home version of the process the plating shops use.

For this you need a silver anode and a quart of concentrated plating solution. Both are available from distributors of plating materials. They can be obtained from Hoover & Strong Co., Tupper Bldg., Buffalo, NY. Other items required before you set up in the plating business are a voltage source, 1 to 3 volts dc, a 2-quart plastic dish, a 5-quart rinsing bucket, degreasing solvent, a pair of clip leads, and some fine steel wool. The plating solution will enable you to plate with other metals as well as silver.

Preparing the Work

Copper, brass and bronze are most suitable for silver plating. Steel can be plated, if it is first plated with copper. Whatever the metal, it should be cleaned and polished before immersion in the plating bath. Rub it down with fine steel wool, and clean it in a degreasing solution. Chemical houses supply degreasers, or you can boil the work in a mild solution of laundry detergent. Rinse thoroughly in clean hot water. Handle only with rubber gloves; finger oils and acids will prevent the metal from plating properly.

Plating

Use distilled water to dilute the plating solution, usually 3 quarts of water to 1 of solution. This must be at room temperature. Too warm a bath will cause discoloration, and too cold will make for spotty plating. Connect the metal to be plated to the negative side of a 1-1/2-volt cell, and slide it into one end of the plating tank. Connect the silver anode to the positive terminal, and submerge it at the opposite end. Maintain

Fig. 8 – A homemade stand for processing etched-circuit boards. The heat lamp maintains the etchant-bath temperature between 90 and 110 degrees F and is mounted on an adjustable arm. The tray for the bath is raised and lowered at one end by the action of a motor-driven eccentric disk, providing the necessary agitation of the chemical solution. A darkroom thermometer monitors the temperature of the bath.

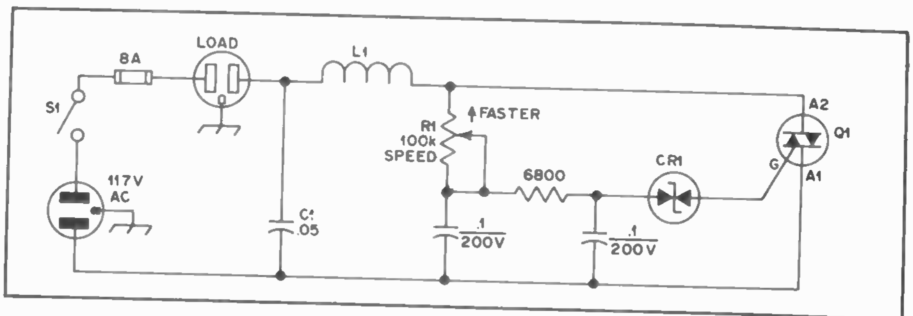


Fig. 9 – Schematic diagram of motor-speed control. Resistances are in ohms (k = 1000) and capacitances are in microfarads. Important note: The basing diagram for Q1 is correct as shown here. Some early literature accompanying the packaging of the HEP device appears to be in error.

C1 – .05-μF, 600-V paper.
CR1 – Diac (silicon bilateral trigger), 2-A, 300-mW (Motorola MPT28 or HEP311 or equiv.).
L1 – Approx. 70 μH; made with 18 ft, no. 18 enam. wire scramble-wound on body of C1, or on a 1-1/2-inch length of 1/2-inch dia. rod.
Q1 – Triac (silicon bidirectional thyristor), 8-A, 200-V (Motorola MAC2-4 or HEP340 or equiv.).
R1 – Linear-taper composition control, 2-W.
S1 – Spst toggle.

COIL RANGE	A (INCHES)	B (INCHES)	WIRE SIZE	TURNS	TAP
1.6-4 MHz	3/8	7/8	No. 30	125	32 turns
3.2-7.4 MHz	1/8	1/4	No. 30	35	11
6-14 MHz	3/8	3/4	No. 20	27	8
12-29 MHz	1/8	1/4	No. 20	10	3
30-90 MHz	4 turns of no. 20, turns spaced to cover 1 inch; tap is 1-1/2 turns from ground end.				
95–300 MHz	Hair pin of no. 14 tinned wire, 1/2-inch spacing, 2 inches long including coil pins, tapped 1-1/2 inch from ground.				

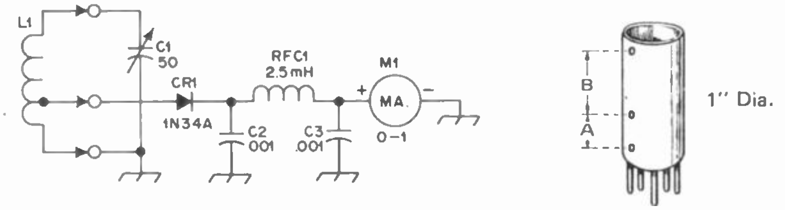


Fig. 10 — Circuit diagram of the frequency meter. Coil-form is 1 X 1-1/2 inch.
C1 — 50-pF variable (Millen 20050).
C2, C3 — .001-μF disk ceramic.
CR1 — 1N34A germanium diode.
L1 — See coil table.
M1 — 0-1 milliammeter.
RFC1 — 2.5-mH rf choke (Millen 34000-2500).

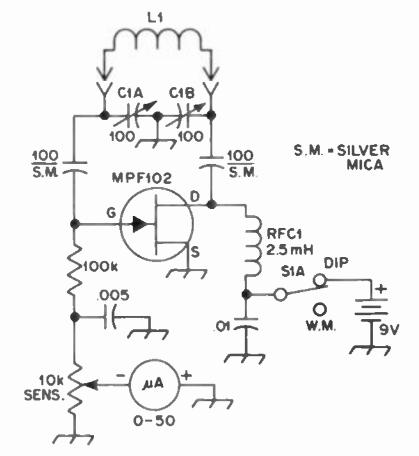


Fig. 11 — An FET gate-dipper circuit suitable for use from 1.5 to 50 MHz. For operation at vhf and uhf the value of C1 should be made smaller, RFC1 would be a vhf type, and the bypass capacitors would be smaller in value. For uhf use Q1 would be changed to a uhf-type FET, a 2N4416 or similar.

a spacing of at least 6 inches between anode and work. Too close spacing causes excessive current flow and discoloration. Agitate the work frequently to prevent bubble formation on it.

Immersion time is usually 5 to 10 minutes. Longer will give heavier coating, and it is best to err on this side, as far as the rf quality of the plating is concerned. The higher the voltage the rougher the finish. Something between 0.5 and 1-1/2 volts is best.

After plating is completed rinse immediately in fresh clean water, preferably lukewarm. Do not touch with the bare hands if you want a clean surface. To preserve the finish, spray with clear lacquer after the work is thoroughly dry. A lacquer spray does not affect the ability of the surface to take solder. If incomplete plating is found near solder areas, it is probably due to the presence of flux. Such areas can be scrubbed with a stiff brush and xylol or alcohol. Replating can be done as needed, in the manner already outlined.

Caution: Silver plating solutions contain cyanide. Avoid breathing the vapors from the bath. In mixing, pour the plating solution into the water, *not* vice versa. Wash hands thoroughly after any contact with the fluid. Do the plating in a well-ventilated room. Store the chemicals in clearly marked containers, out of the reach of children.

An Alternative

Another method, very simple to use, involves a plating powder. It is applied with a damp cloth dipped in the powder, and then rubbed onto the surface to be plated. Because some rubbing is required, the resulting surface comes out nice and smooth. The material, called COOL-AMP, is made by a company of that name, 8603 S.W. 17th Ave., Portland, Oregon. The powder is sold only in jars, minimum order 1 pound, but a little goes a long way. Several would-be platers could do quite a bit of work each with one pound which covers about 6000 square inches!

The above method is best used with rubber gloves. The plating materials are a little rough on the skin otherwise, and neater work is possible if the fingers are kept from direct contact with the work or the plating substances.

Motor-Speed Control

It is often necessary to exercise caution when drilling a chassis or box which contains component parts, lest the drill bit pass through the metal too quickly and destroy a component. A motor-speed control for electric hand drills is a handy workshop aid when low-speed

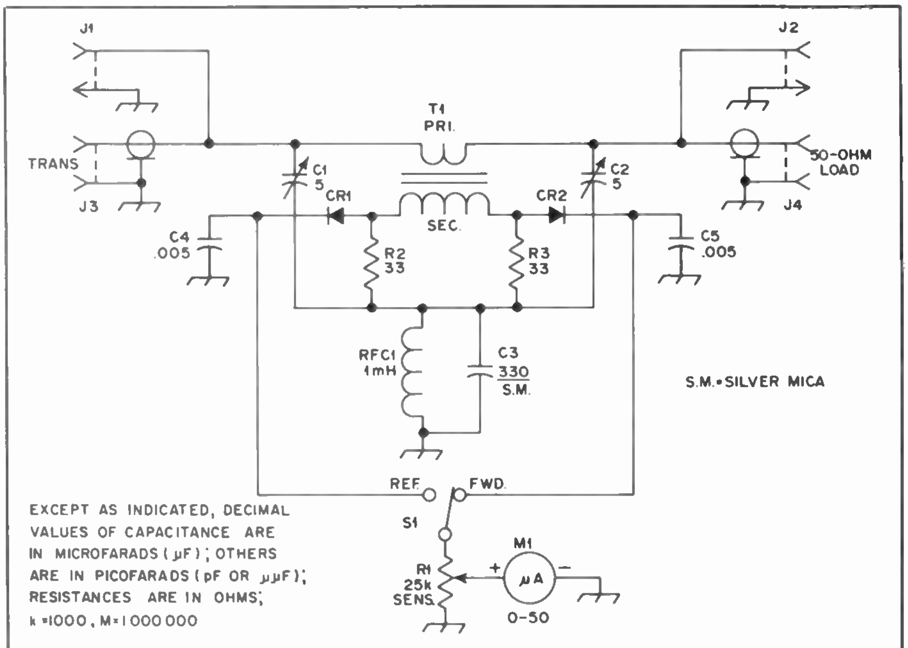


Fig. 12 — Schematic diagram of the wattmeter.
C1, C2 — 0.5- to 5-pF trimmer.
CR1, CR2 — 1N34A or equivalent.
M1 — 50-μA panel meter.
R1 — Linear-taper, 1/4- or 1/2 watt, 25,000 ohm.
R2, R3 — 33-ohm, 1/2-W composition resistor (matched pair recommended)
RFC1 — 1-mH rf choke.
S1 — Spdt toggle.
T1 — 60 turns no. 28 enam. wire, close wound on Amidon T-68-2 toroid core (secondary). Primary is 2 turns of small-diameter hookup wire over T1 secondary.

drilling is desired. Furthermore, the drill motor can be slowed down for coil-winding jobs when the drill is used to turn the coil form. Fig. 9 provides a circuit for use in controlling the speed of a drill motor.

Wavemeter for Testing Use

A wavemeter is an essential instrument for the ham workshop. When equipped with a calibrated dial it can be invaluable in sampling various oscillator or rf power circuits to peak them, or to determine at which frequency they are delivering output. Wavemeters are useful in determining the relative level of harmonic energy in a tuned circuit when checking out a transmitter. A simple circuit for general-purpose bench use is given in Fig. 10. The unit should be housed in a metal box, but the coil should be outside the box.

Gate-Dip Meter

Another useful workshop instrument is the dip meter. It can be used to determine the resonant frequency of a tuned circuit, or for finding the inductance value of the unknown coil. In the latter situation, a capacitor of known value is connected in parallel with the coil and a dip is found with the instrument. Then, the frequency of resonance and capacitor values are compared to determine inductance. Generally speaking, the deeper the dip obtained when checking a tuned circuit, the higher the tuned-circuit *Q*. This technique can be applied when trying to find the relative *Q* of an unknown surplus slug-tuned coil.

Fig. 11 contains a circuit for an FET dip meter. The coils can be wound on plug-in forms, and the remainder of the circuit should be contained in a small metal box. Coils can be wound to cover the broadcast band through 50 MHz. The number of turns used will be determined by the frequency spread per coil desired, and in accordance with the capacitor value used at C1.

RF Power and SWR Meter

Rf power measurements must be made frequently in the amateur's workshop. Also, the SWR of an amplifier input port must be known and compensated for before certain exciters can be connected to the amplifier. Antenna SWR is but another factor of importance to amateurs.

An instrument suitable for rf power and SWR measuring is shown schematically in Fig. 12. For high-power rf work the primary of T1 should have the center conductor of the feeder pass through it from J3 to J4. For low power testing the primary of T1 should consist of a two-turn link around the core. R1 can be calibrated for various power ranges by determining the rf power at a specific setting while measuring rms rf voltage across a 50-ohm noninductive dummy load, then converting the data to watts.

The bridge is nulled by terminating it in 50 ohms at one port, while a 50-ohm signal source is attached to the remaining port. The source and load are reversed several times while nulling is effected by alternate adjustment of C1 and C2. The null is sought in the reflected power mode, whichever way the instrument is connected to signal source and load. A complete treatment of this subject was given in *QST* for December, 1969, page 11.

Testing Transistors

No modern amateur workshop is complete without a transistor tester. Such an instrument enables the builder to determine if unknown faults exist in a transistor. Also, the equipment is handy for finding out if surplus or unmarked transistors are defective, and whether they are npn or pnp types. A general idea of their useful frequency range can be obtained by inserting them in an oscillator circuit of known frequency.

Fig. 13 shows the circuit of a general-purpose tester. The builder can test MOSFETs, JFETs, and bipolar transistors of the small-signal variety. This is somewhat a "go-no-go" tester, in that a defective transistor will not oscillate in the circuit. The strength of oscillation is noted on the meter. The higher the reading, relatively speaking, the greater the transistor gain at the crystal frequency. This concept could be carried further by the addition of multiplier stages which would permit checking transistors as amplifiers well into the uhf region. An rf-sampling circuit would have to be switched from each stage to the meter.

Sockets for bipolars, FETs, and MOSFETs are mounted on the instrument case to facilitate plugging in the device to be tested. S1 is used to reverse the battery polarity when checking npn or pnp transistors. The FET/Bipolar switch, S3, removes forward bias from the line while testing FETs.

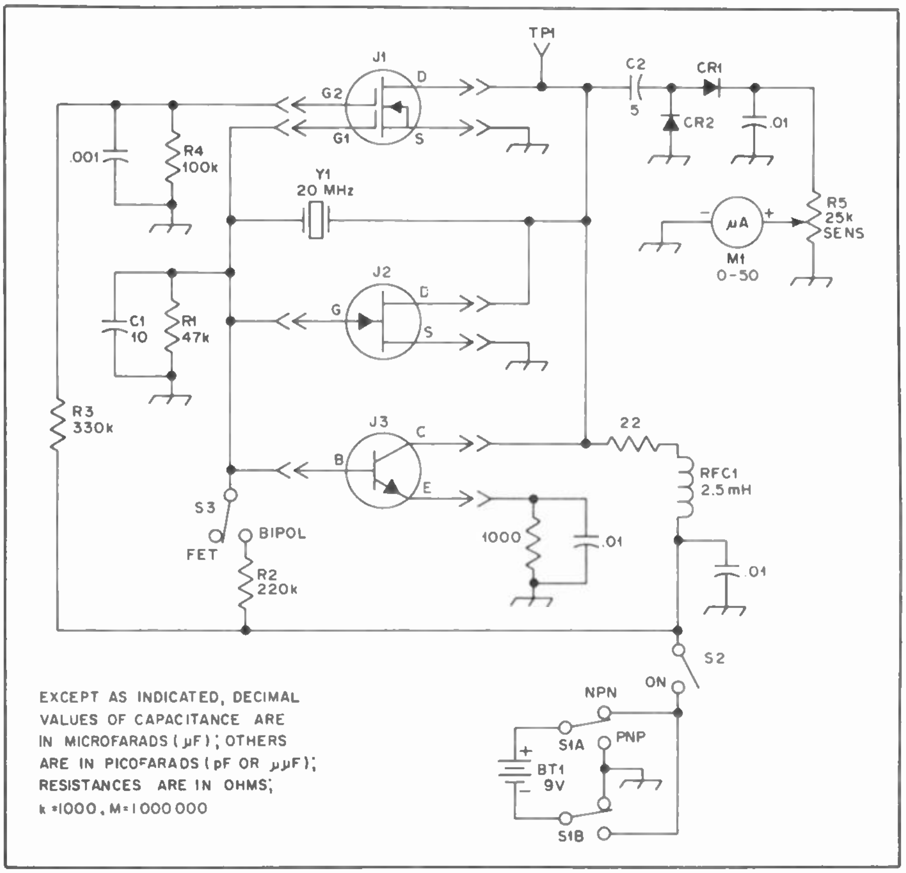


Fig. 13 — Schematic diagram of the transistor tester. Capacitors are disk ceramic or mica. Resistors are 1/2- or 1/4-watt composition except for R5. Numbered components not appearing in parts list are so designated for text discussion.

BT1 — Small 9-V transistor-radio battery.
CR1, CR2 — 1N34A, germanium diode or equiv.
J1 — Four-terminal transistor socket.
J2, J3 — Three-terminal transistor socket.
M1 — Microampere meter. Calectro D1-910 used here.
R5 — 25,000-ohm linear-taper composition control with switch.
RFC1 — 2.5-mH rf choke
S1 — Two-pole double-throw miniature toggle.
S2 — Part of R5.
S3 — Spst miniature toggle.
Y1 — Surplus crystal.

Table 1

Fractions of an Inch with Metric Equivalents

FRACTIONS OF AN INCH	*DECIMALS OF AN INCH*	*MILLIMETERS*	*FRACTIONS OF AN INCH*	*DECIMALS OF AN INCH*	*MILLIMETERS*
1/64	0.0156	0.397	33/64	0.5156	13.097
1/32	0.0313	0.794	17/32	0.5313	13.494
3/64	0.0469	1.191	35/64	0.5469	13.891
1/16	0.0625	1.588	9/16	0.5625	14.288
5/64	0.0781	1.984	37/64	0.5781	14.684
3/32	0.0938	2.381	19/32	0.5938	15.081
7/64	0.1094	2.778	39/64	0.6094	15.478
1/8	0.1250	3.175	5/8	0.6250	15.875
9/64	0.1406	3.572	41/64	0.6406	16.272
5/32	0.1563	3.969	21/32	0.6563	16.669
11/64	0.1719	4.366	43/64	0.6719	17.066
3/16	0.1875	4.763	11/16	0.6875	17.463
13/64	0.2031	5.159	45/64	0.7031	17.859
7/32	0.2188	5.556	23/32	0.7188	18.256
15/64	0.2344	5.953	47/64	0.7344	18.653
1/4	0.2500	6.350	3/4	0.7500	19.050
17/64	0.2656	6.747	49/64	0.7656	19.447
9/32	0.2813	7.144	25/32	0.7813	19.844
19/64	0.2969	7.541	51/64	0.7969	20.241
5/16	0.3125	7.938	13/16	0.8125	20.638
21/64	0.3281	8.334	53/64	0.8281	21.034
11/32	0.3438	8.731	27/32	0.8438	21.431
23/64	0.3594	9.128	55/64	0.8594	21.828
3/8	0.3750	9.525	7/8	0.8750	22.225
25/64	0.3906	9.922	57/64	0.8906	22.622
13/32	0.4063	10.319	29/32	0.9063	23.019
27/64	0.4219	10.716	59/64	0.9219	23.416
7/16	0.4375	11.113	15/16	0.9375	23.813
29/64	0.4531	11.509	61/64	0.9531	24.209
15/32	0.4688	11.906	31/32	0.9688	24.606
31/64	0.4844	12.303	63/64	0.9844	25.003
1/2	0.5000	12.700		1.0000	25.400

Table 2

Conversion — Decimal Feet to Inches (Nearest 16th)

	0	*1*	*2*	*3*	*4*	*5*	*6*	*7*	*8*	*9*
0.0	0-0	0-1/8	0-1/4	0-3/8	0-1/2	0-5/8	0-3/4	0-13/16	0-15/16	1-1/16
0.1	1-3/16	1-5/16	1-7/16	1-9/16	1-11/16	1-13/16	1-15/16	2-1/16	2-3/16	2-1/4
0.2	2-3/8	2-1/2	2-5/8	2-3/4	2-7/8	3-0	3-1/8	3-1/4	3-3/8	3-1/2
0.3	3-5/8	3-3/4	3-13/16	3-15/16	4-1/16	4-3/16	4-5/16	4-7/16	4-9/16	4-11/16
0.4	4-13/16	4-15/16	5-1/16	5-3/16	5-1/4	5-3/8	5-1/2	5-5/8	5-3/4	5-7/8
0.5	6-0	6-1/8	6-1/4	6-3/8	6-1/2	6-5/8	6-3/4	6-13/16	6-15/16	7-1/16
0.6	7-3/16	7-5/16	7-7/16	7-9/16	7-13/16	7-11/16	7-15/16	8-3/16	8-1/16	8-1/4
0.7	8-3/8	8-1/2	8-5/8	8-3/4	8-7/8	9-0	9-1/8	9-1/4	9-3/8	9-1/2
0.8	9-5/8	9-3/4	9-13/16	9-15/16	10-1/16	10-3/16	10-5/16	10-7/16	10-9/16	10-11/16
0.9	10-13/16	10-15/16	11-1/16	11-3/16	11-1/4	11-3/8	11-1/2	11-5/8	11-3/4	11-7/8

Conversion — Inches and Fractions to Decimal Feet

	0	*1/8*	*1/4*	*3/8*	*1/2*	*5/8*	*3/4*	*7/8*
0	000	010	021	031	042	052	063	073
1	083	094	104	115	125	135	146	156
2	167	177	188	198	208	219	229	240
3	250	260	271	281	292	302	313	323
4	333	344	354	365	375	385	396	406
5	417	427	438	448	458	469	479	490
6	500	510	521	531	542	552	563	573
7	583	594	604	615	625	635	646	656
8	667	677	688	698	708	719	729	740
9	750	760	771	781	792	802	813	823
10	833	844	854	865	875	885	896	906
11	917	927	938	948	958	969	979	990

Table 3

Numbered Drill Sizes

NUMBER	DIAMETER (MILS)	WILL CLEAR SCREW	DRILLED FOR TAPPING FROM 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	140.0	6-32	—
29	136.0	—	8-32
30	128.5	—	—
31	120.0	—	—
32	116.0	—	—
33	113.0	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	—	4-40
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 for tapping bakelite and phenolics.

Table 4

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	.006640626	—
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.

Table 5

Resistor-Capacitor Color Code

COLOR	SIGNIFICANT FIGURE	DECIMAL MULTIPLIER	TOLERANCE (%)	VOLTAGE RATING*
Black	0	1	—	—
Brown	1	10	1*	100
Red	2	100	2*	200
Orange	3	1,000	3*	300
Yellow	4	10,000	4*	400
Green	5	100,000	5*	500
Blue	6	1,000,000	6*	600
Violet	7	10,000,000	7*	700
Gray	8	100,000,000	8*	800
White	9	1,000,000,000	9*	900
Gold	—	0.1	5	1000
Silver	—	0.01	10	2000
No color	—		20	500

*Applies to capacitors only.

Color Code for Ceramic Capacitors

COLOR	SIGNIFICANT FIGURE	DECIMAL MULTIPLIER	TOLERANCE MORE THAN 10 pF (IN %)	TOLERANCE LESS THAN 10 pF (IN %)	TEMP. COEFF. PPM/ DEG. C
Black	0	1	±20	2.0	0
Brown	1	10	± 1		−30
Red	2	100	± 2		−80
Orange	3	1000			−150
Yellow	4				−220
Green	5				−330
Blue	6		± 5	0.5	−470
Violet	7				−750
Gray	8	0.01		0.25	30
White	9	0.1	±10	1.0	500

Capacitor Characteristic Code

COLOR SIXTH DOT	TEMPERATURE COEFFICIENT PPM/DEG. C	CAPACITANCE DRIFT
Black	±1000	±5% + 1 pF
Brown	±500	±3% + 1 pF
Red	±200	±0.5%
Orange	±100	±0.3%
Yellow	−20 to +100	±0.1% + 0.1 pF
Green	0 to +70	±0.05% + 0.1 pF

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 transformer is center-tapped, the second diode plate lead is green-and-black striped, and black is used for the center-tap lead.

Audio 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.

Power Transformers

1) Primary Leads — Black
   If tapped:
   Common — Black
   Tap — Black and Yellow Striped
   Finish — Black and Red Striped
2) High-Voltage Place Winding — Red
   Center-Tap — Red and Yellow Striped
3) Rectifier Filament Winding — Yellow
   Center-Tap — Yellow and Blue Striped
4) Filament Winding No. 1 — Green
   Center-Tap — Green and Yellow Striped
5) Filament Winding No. 2 — Brown
   Center-Tap — Brown and Yellow Striped
6) Filament Winding No. 3 — Slate
   Center-Tap — Slate and Yellow Striped

Table 6

Approximate Series-Resonance Frequencies of Disc Ceramic Bypass Capacitors

CAPACITANCE	FREQ.[1]	FREQ.[2]
.01 µF	13 MHz	15 MHz
.0047	18	22
.002	31	38
.001	46	55
.0005	65	80
.0001	135	165

[1] Total lead length of 1 inch.
[2] Total lead length of 1/2-inch.

Table 7

Dielectric Constants and Breakdown Voltages

MATERIAL	DIELECTRIC CONSTANT*	PUNCTURE VOLTAGE**
Air	1.0	
Alsimag 196	5.7	240
Bakelite	4.4-5.4	300
Bakelite, mica-filled	4.7	325-375
Cellulose acetate	3.3-3.9	250-600
Fiber	5-7.5	150-180
Formica	4.6-4.9	450
Glass, window	7.6-8	200-250
Glass, Pyrex	4.8	335
Mica, ruby	5.4	3800-5600
Mycalex	7.4	250
Paper, Royalgrey	3.0	200
Plexiglass	2.8	990
Polyethylene	2.3	1200
Polystyrene	2.6	500-700
Porcelain	5.1-5.9	40-100
Quartz, fixed	3.8	1000
Steatite, low-loss	5.8	150-315
Teflon	2.1	1000-2000

*At 1 MHz
**In volts per mil (0.001 inch)

Table 8

Relative Resistivity of Metals

MATERIALS	RESISTIVITY COMPARED TO COPPER
Aluminum (pure)	1.6
Brass	3.7-4.9
Cadmium	4.4
Chromium	1.8
Copper (hard-drawn)	1.03
Copper (annealed)	1.00
Gold	1.4
Iron (pure)	5.68
Lead	12.8
Nickel	5.1
Phosphor Bronze	2.8-5.4
Silver	0.94
Steel	7.6-12.7
Tin	6.7
Zinc	3.4

Table 9

Color Code for Hookup Wire

WIRE COLOR	*TYPE OF CIRCUIT*
Black	Grounds, grounded elements, and returns
Brown	Heaters or filaments, off ground
Red	Power supply B plus
Orange	Screen grids and Base 2 of transistors
Yellow	Cathodes and transistor emitters
Green	Control grids, diode plates, and Base 1 of transistors
Blue	Plates and transistor collectors
Violet	Power supply, minus leads
Gray	Ac power line leads
White	Bias supply, B or C minus, agc

Wires with tracers are coded in the same manner as solid-color wires, allowing additional circuit identification over solid-color wiring. The body of the wire is white and the color band spirals around the wire lead. When more than one color band is used, the widest band represents the first color.

Table 10

Pilot-Lamp Data

LAMP NO.	*BEAD COLOR*	*BASE (MINIATURE)*	*BULB TYPE*	*RATING V*	*A*
40	Brown	Screw	T-3-1/4	6-8	0.15
40A[1]	Brown	Bayonet	T-3-1/4	6-8	0.15
41	White	Screw	T-3-1/4	2.5	0.5
42	Green	Screw	T-3-1/4	3.2	**
43	White	Bayonet	T-3-1/4	2.5	0.5
44	Blue	Bayonet	T-3-1/4	6-8	0.25
45	*	Bayonet	T-3-1/4	3.2	**
46[2]	Blue	Screw	T-3-1/4	6-8	0.25
47[1]	Brown	Bayonet	T-3-1/4	6-9	0.15
48	Pink	Screw	T-3-1/4	2.0	0.06
49[3]	Pink	Bayonet	T-3-1/4	2.0	0.06
49A[3]	White	Bayonet	T-3-1/4	2.1	0.12
50	White	Screw	G-3-1/2	6-8	0.2
51[2]	White	Bayonet	G-3-1/2	6-8	0.2
53	—	Bayonet	G-3-1/2	14.4	0.12
55	White	Bayonet	G-4-1/2	6-8	0.4
292[4]	White	Screw	T-3-1/4	2.9	0.17
292A[4]	White	Bayonet	T-3-1/4	2.9	0.17
1455	Brown	Screw	G-5	18.0	0.25
1455A	Brown	Bayonet	G-5	18.0	0.25
1487	—	Screw	T-3-1/4	12-16	0.20
1488	—	Bayonet	T-3-1/4	14	0.15
1813	—	Bayonet	T-3-1/4	14.4	0.10
1815	—	Bayonet	T-3-1/4	12-16	0.20

[1]40A and 47 are interchangeable.
[2]Have frosted bulbs.
[3]49 and 49A are interchangeable.
[4]Use in 2.5-volt sets where regular bulb burns out too frequently.
*White in G. E. and Sylvania; green in National Union, Raytheon and Tung-Sol.
**0.35 in G. E. and Sylvania; 0.5 in National Union, Raytheon and Tung-Sol.

Table 11

Rf Chokes for 50-, 144- and 220-MHz Service

FREQUENCY	*INDUCTANCE*	*DESCRIPTION*
50 MHz	7.8 to 9.5 μH	B & W Miniductor no. 3004, 1-3/8 to 1-9/16 inch long.*
50 MHz	8.3 μH	No. 28 d.s.c., spacewound on 1/2-inch Teflon rod. Winding 1-3/4 inch long. See text.
50 MHz	7.2 μH	No. 28 d.s.c., closewound on 1/4-inch Teflon rod. Winding 1-7/16-inch long.
144 MHz	2.15 μH	No. 22 Nyclad, closewound 1-3/16-inch on 1/4-inch Teflon rod.
144 MHz	1.42 μH	31 turns no. 28 d.s.c., spacewound on 1/4-inch Teflon rod. Winding 1-1/16-inch long.
144 MHz	1.3 μH	29 turns no. 22 Nyclad 1-1/8 inch long 1/4 inch diameter self-supporting. (Above 144-MHz chokes works well on 220 MHz)
220 MHz	0.6 μH	13 turns no. 22 Nyclad on 1/4-inch Teflon rod.
220 MHz	0.75 μH	17 turns no. 28 d.s.c., spacewound on 1/4-inch Teflon rod winding 5/8-inch long.
220 MHz	0.52 μH	22 turns no. 22 Nyclad closewound on no. 24 drill, self-supporting.

*Excellent for use except where high temperatures are involved.

Chapter 10

Data Potpourri

This section of the book contains assorted bits of useful information which did not have a logical place in the foregoing chapters of the data file. The late entries range from power-supply information through station operating aids. Maximum emphasis is placed on tables and diagrams rather than text.

Power-Supply Data

Fig. 1 shows the relationship between half-wave, full-wave, and full-wave bridge rectifiers. Comparisons are given for voltage output levels and ripple amounts. The advantage of the circuit at C is that no center tap is required on the transformer secondary winding.

Fig. 2 illustrates the circuits of a voltage doubler (A), voltage tripler (B), and a voltage quadrupler (C). Other configurations are possible, but those shown are recommended for safety purposes as the secondary windings have one leg grounded.

At Fig. 3 is a presentation of output voltages from a full-wave voltage-doubling circuit as a function of the filter capacitances and load resistance. The graph illustrates how the voltage depends upon the ratio of the series resistance to the load resistance, and the product of the load resistance times the filter capacitance.

Zener-Diode Applications

Perhaps one of the least understood and most used solid-state devices in amateur work is the Zener diode. Today's manufacturing technology has caused the Zener diode to outrank its ancestor, the gaseous regulator tube. Furthermore, Zener diodes enable the builder to regulate voltages at more precise values than was possible with tube regulators. That is, a wide variety of diode voltages are available as opposed to the relatively high and limited quantities of voltages covered by regulator tubes.

Fig. 4 presents some typical uses to which Zener diodes can be put. Illustration A shows a simple shunt regulator. The output voltage is governed by the rated voltage of the diode, VR1. Some flexibility is possible by adding various Zener diodes in series, as seen at B. For example, 6.8 and 10-volt diodes can be connected in series to obtain 16.8 volts regulated. This can be carried even further by using several diodes in series as shown at D. In this example regulated voltages of various amounts can be picked off at the points where the diodes are joined.

Fig. 4C shows how a Zener diode is used as a reference element in a series-pass regulator circuit. Output voltage from Q1 will be the Zener-diode voltage minus the barrier voltage of the transistor, Q1. Typically, the drop across a silicon-transistor junction will range from 0.5 to 0.7 volts.

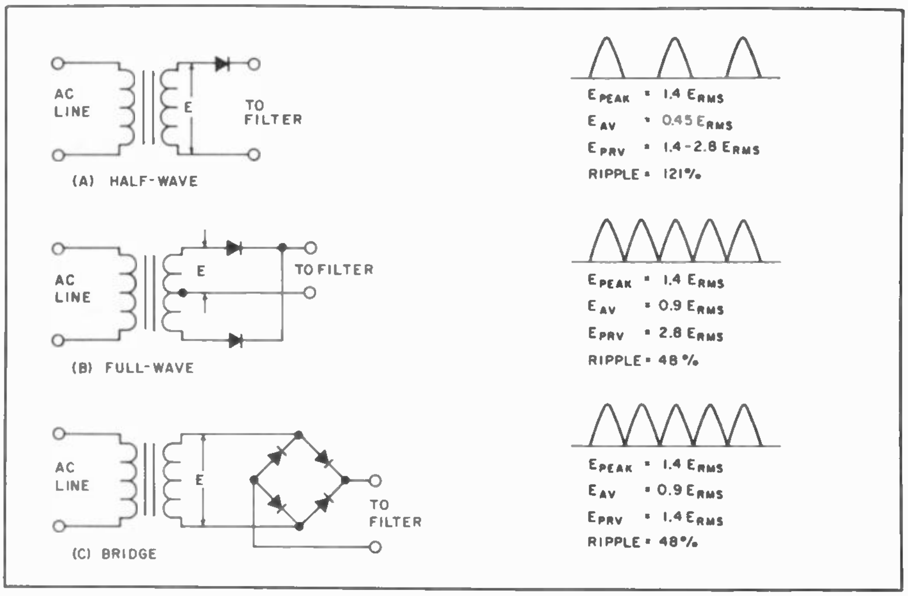


Fig. 1 — Fundamental rectifier circuits. A — Half-wave (E_{PRV} = 1.4 E_{rms} with resistive load, = 2.8 E_{rms} with capacitor-input filter). B — Full-wave. C — Full-wave bridge. Output voltage values do not include rectifier voltage drops.

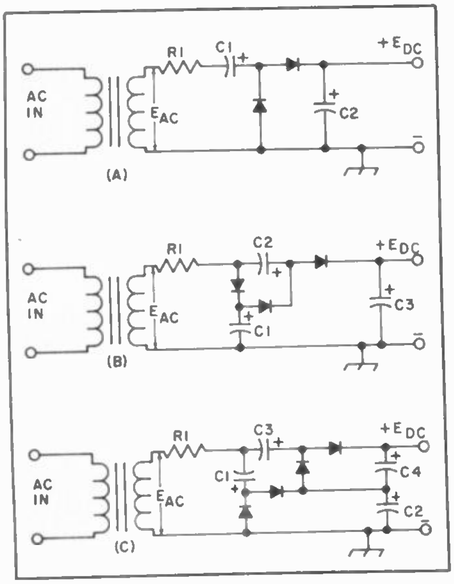


Fig. 2 — Voltage-multiplying circuits with one-side of transformer secondary grounded. (A) — Voltage doubler. (B) Voltage tripler. (C) Voltage-quadrupler.

Capacitances are typically 20 to 50 µF depending upon output current demand. Dc ratings of capacitors are related to E_{peak} (1.4 E_{ac})

C1 — Greater than E_{peak}
C2 — Greater than $2E_{peak}$
C3 — Greater than $3E_{peak}$
C4 — Greater than $4E_{peak}$

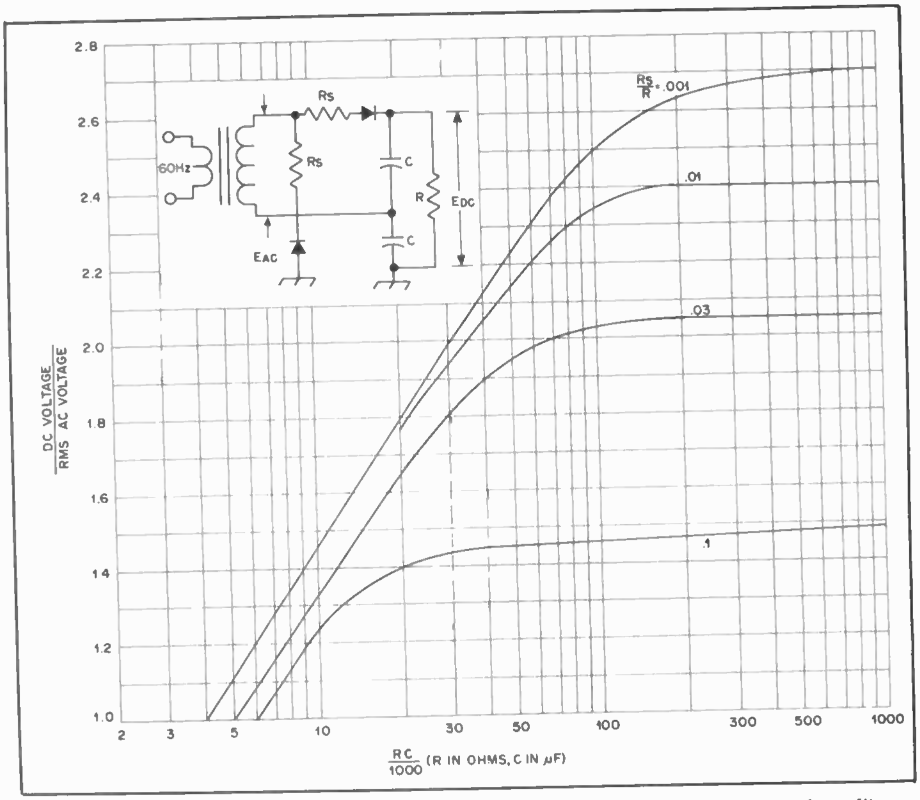


Fig. 3 — Dc output voltages from a full-wave voltage-doubling circuit as a function of the filter capacitances and load resistance. For the ratio R_s/R and for the RC product, resistances are in ohms and capacitance is in microfarads. Equal resistance values for R_s and equal capacitance values for C are assumed.

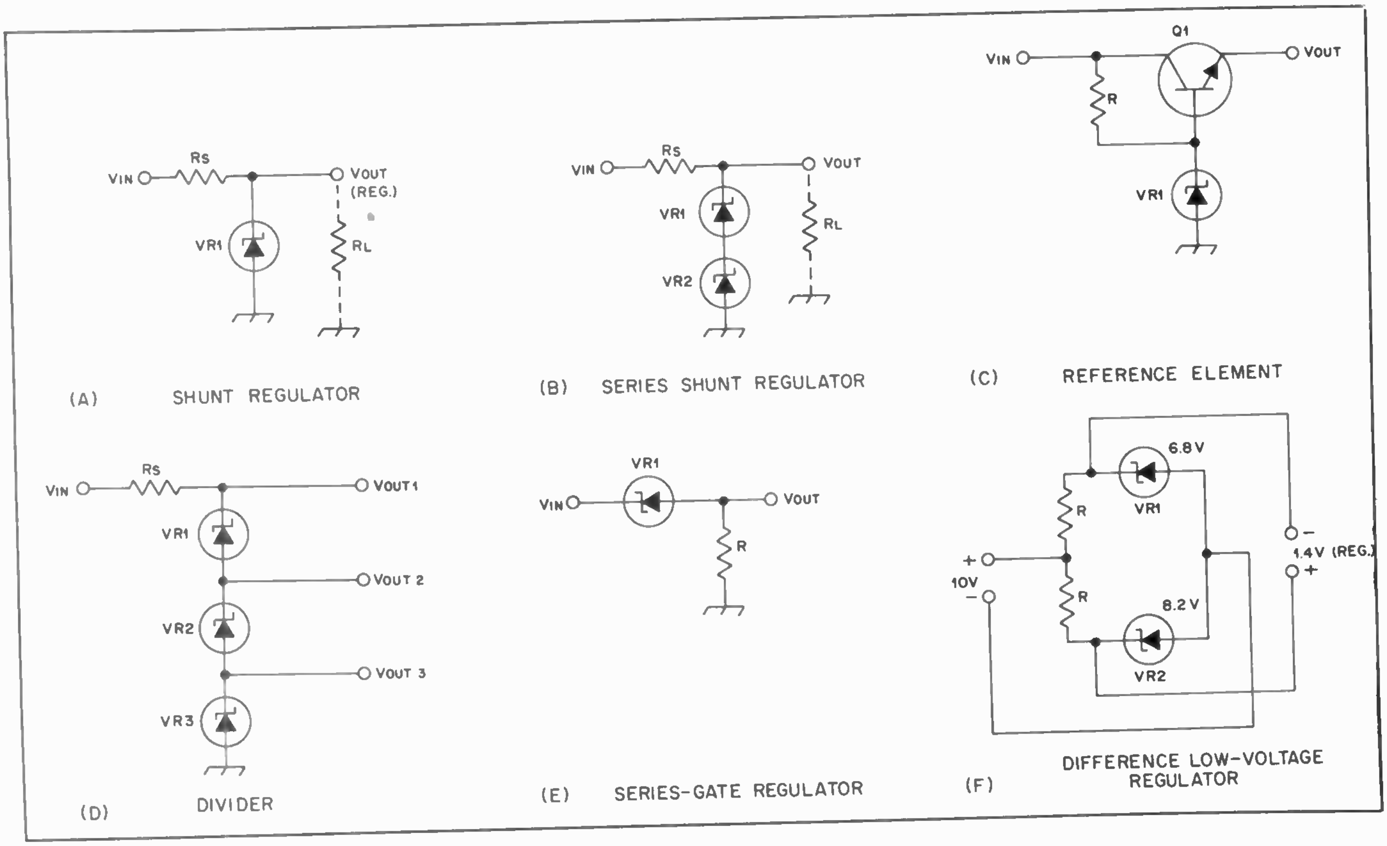


Fig. 4 — Dc applications for Zener-diode regulators.

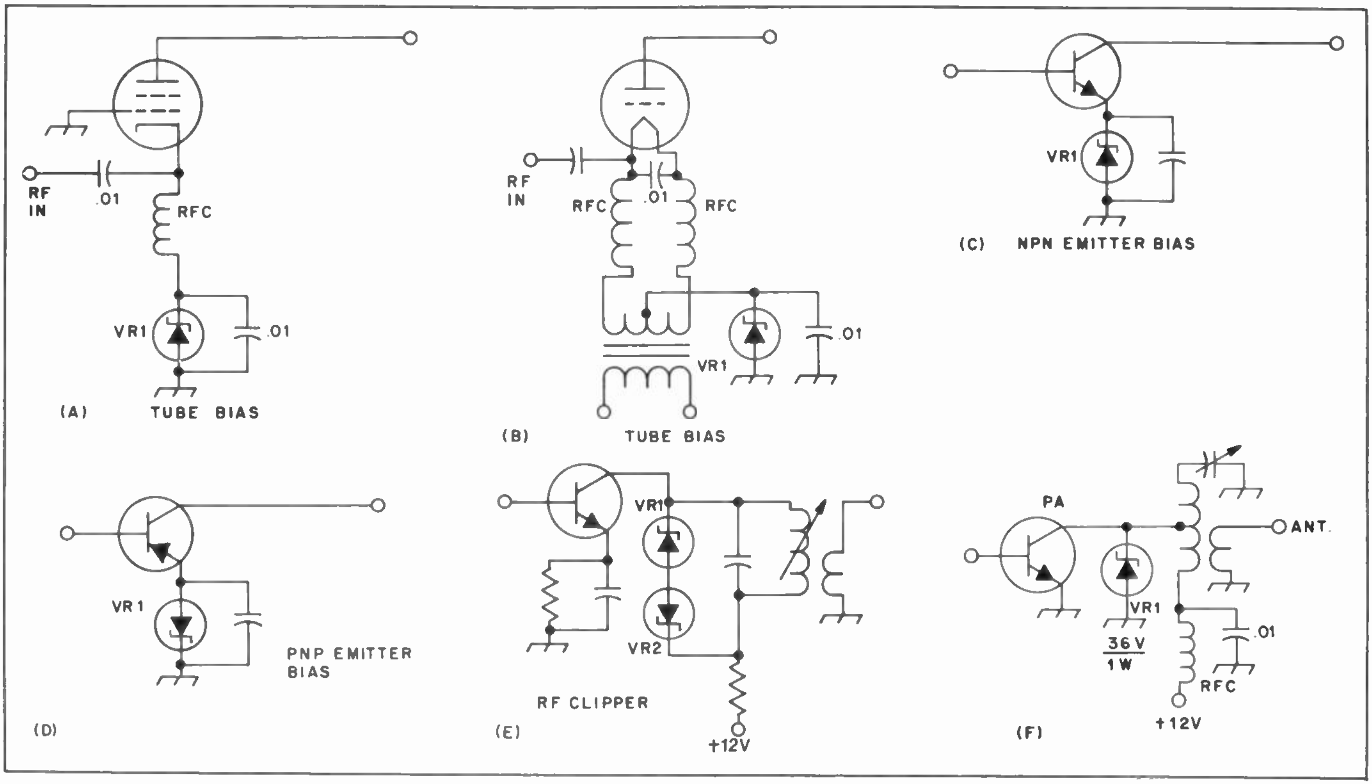


Fig. 5 – Dc and rf applications for Zener diodes.

A series-gate regulator is seen at Fig. 4E. Output voltage will not appear until the input voltage reaches and exceeds the barrier voltage of VR1. A novel method for obtaining very low regulated voltages is seen at F. This circuit is useful for replacing a standard flashlight battery.

More examples of how Zener diodes can be utilized are shown in Fig. 5. Illustrations A and B are useful for providing bias for vacuum-tube amplifiers. The Zener diode must be capable of safely passing the total cathode current of the tube or tubes with which it is used. Biasing for transistors can be done in a like manner, as seen at C and D.

Fig. 5E illustrates the hookup for rf clipping with Zener diodes. The positive and negative rf-voltage peaks cannot exceed the barrier voltages of the diodes. For low-level signal work it is more common to use small silicon or germanium diodes for rf clipping.

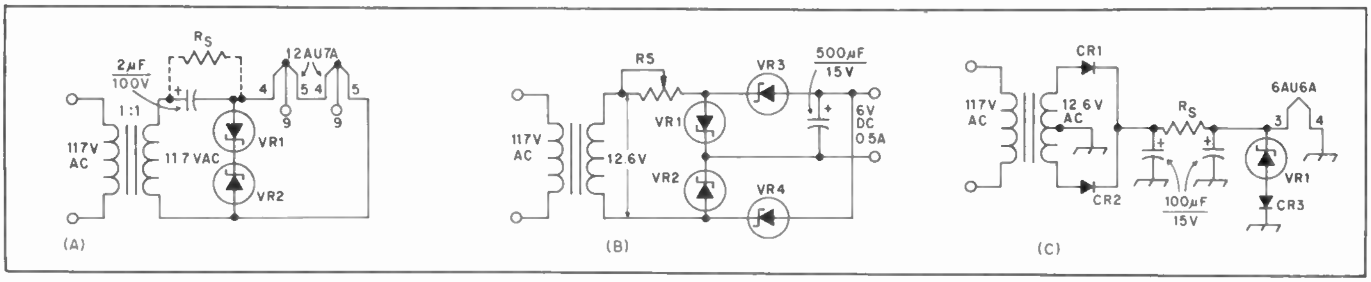
A protective circuit for power transistors is seen at Fig. 5F. The example shows an npn rf transistor with the collector shunted by means of a 36-volt Zener diode. The diode has little effect during normal operation because the peak voltage swing of the collector does not exceed 24 volts. However, should the stage break into self-oscillation, or should a severe mismatch occur at the output, the peak voltage can soar quite high – possibly destroying the transistor. VR1, in such an event, will clamp on peaks at and above 36 volts positive, thereby protecting the transistor. If a pnp transistor is used the Zener diode must be reversed in polarity to limit negative-going peaks.

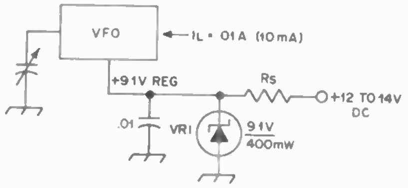
Special Applications

Some less ordinary uses for Zener diodes are covered in Fig. 6. At A it can be seen that a regulated filament voltage can be obtained for two 12AU7A tubes from a 117-volt source. Either C1 or R_s can be used to provide the needed voltage drop. If C1 is used, its reactance at 60 Hz should equal the calculated resistance of R_s.

Fig. 6B illustrates how four Zener diodes can be employed to serve as rectifiers and regulators for a simple 6-volt, 500-mA power supply. The regulation of this type of supply will not be

Fig. 6 – Some unique uses to which Zener diodes can be put.





$$R_s = \frac{E_{in(min)} - E_{VR1}}{I_L + 0.1\, I_L} \;\therefore$$

$$R_s = \frac{12 - 9.1}{.01\,\text{A} + .001\,\text{A}} = \frac{2.9}{.011}$$

$$= 264 \text{ ohms}$$

and

$$P_{D(max)} = \left(\frac{E_{in(max)} - E_{VR1}}{R_s} - I_L\right) E_{VR1} \,..$$

$$P_{D(max)} = \left(\frac{14 - 9.1}{264} - .01\right) 9.1$$

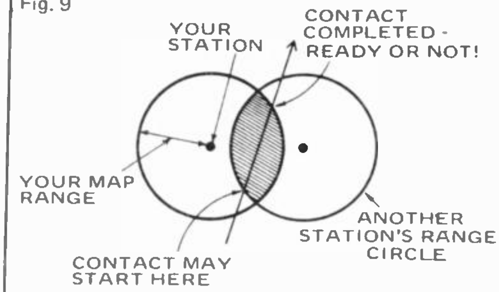
$$= \left(\frac{4.9}{264} - .01\right) 9.1$$

$$= .0085 \times 9.1 = .078 \text{ W}$$

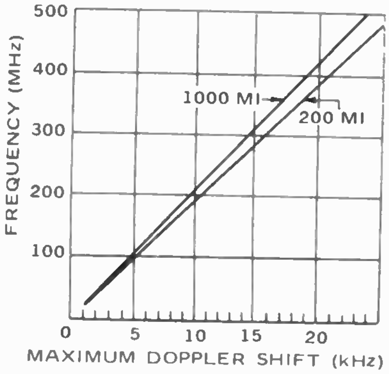
R_s = series resistance
I_L = current of load in amperes
E_{VR1} = Zener-diode voltage
$P_{D(max)}$ = total VR1 dissipation in watts

Fig. 7 — Example of how a shunt Zener-diode regulator is used. Equations are provided for calculating the series-resistor ohmic value, and the dissipation of the Zener diode.

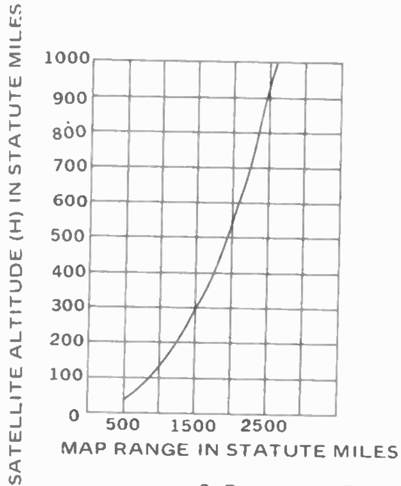
Fig. 9



Satellite passes through the range of two stations, enabling contact.



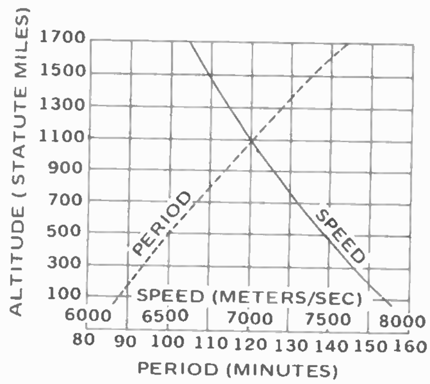
Satellite transmitter frequency versus Doppler shift for satellite in 200- or 1000-statute-mile orbits. For a translator, use the difference between uplink and downlink frequencies as the "frequency."



$$\text{MAP RANGE} = \frac{2\pi R}{360} \cos^{-1} \frac{R}{R + H}$$

R = EARTH RADIUS (3960 STAT MI)
H = SATELLITE ALT (STAT MI)

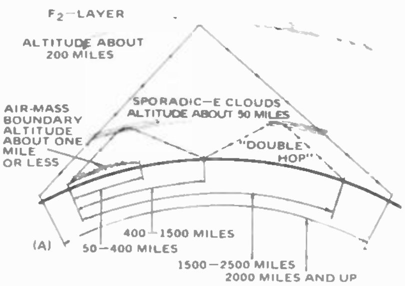
Satellite altitude above earth versus ground station map range (statute miles).



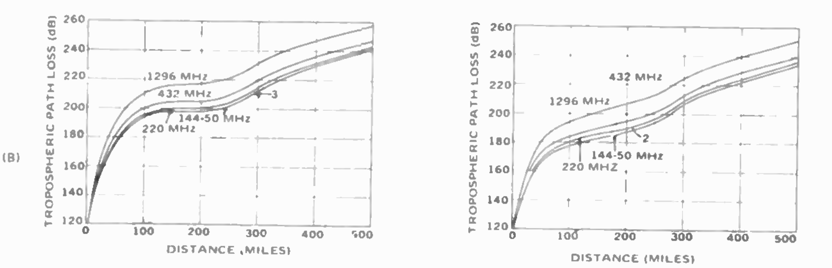
Satellite altitude versus its period (time for one revolution) and speed.

Fig. 8

Approximate effective communications distances.



The principal means by which vhf signals may be returned to earth. The $F2$ layer may provide 50 MHz DX at the peak of the 11-year sunspot cycle. Such communication may be world-wide in scope. Sporadic ionization of the E region produces the familiar "short skip" on 28 and 50 MHz. It is most common in early summer but may occur at any time and regardless of the sunspot cycle. Refraction of vhf waves also takes place at air-mass boundaries, making possible communication over distances of several hundred miles on all vhf bands. Normally it exhibits no skip zone.



Path loss vs. distance for amateur frequencies above 50 MHz. Curves at the left are for 50 percent of the time; those at the right for 99 percent. The former is the more representative of amateur radio requirements.

as precise as that of a supply using the method shown in Fig. 4C, but will be ample for many amateur applications.

Still another technique is presented at C in Fig. 6. Here a standard-value 5.6-volt Zener diode is hooked in series with a silicon power diode to take advantage of the barrier voltage of CR3. This provides regulation at 6.3 volts. A circuit such as the one at C is useful for regulating the filament voltage and reducing ac hum of oscillator stages in VFOs and receivers. Some amateurs keep the voltage applied to oscillators around the clock to reduce warm-up drift. This circuit is ideal for the purpose.

Resistance and Power Calculations

It is necessary to determine the value of the series resistor when using a Zener-diode regulator. One must also know how much power will be dissipated in the diode junction in order to select a diode of proper wattage rating. It is essential also to know the total circuit dissipation (Zener-diode load and

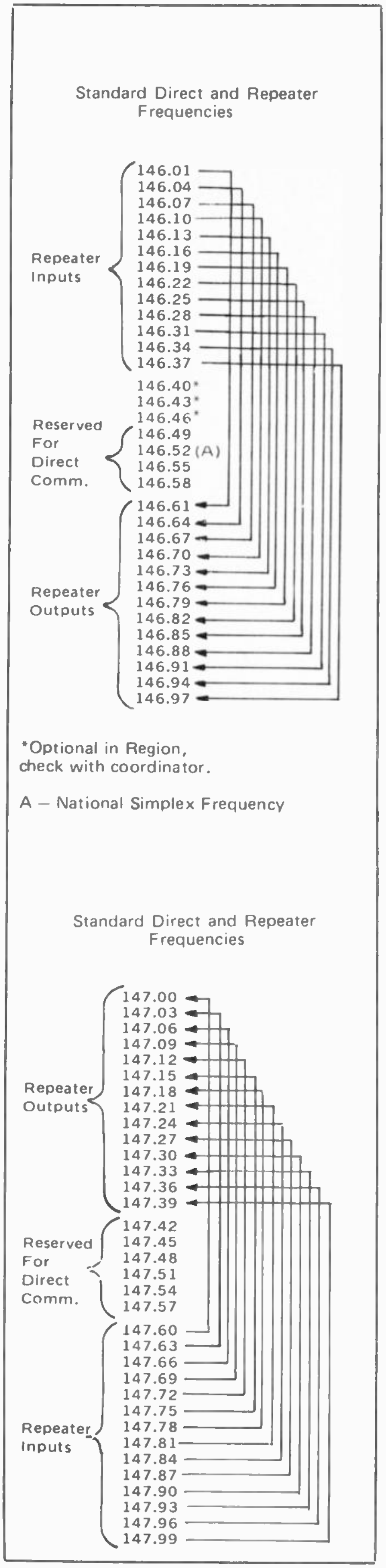


Fig. 10 – Frequency arrangement for 2-meter repeaters.

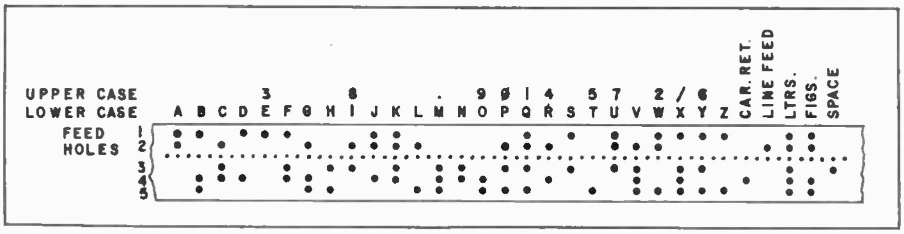


Fig. 11 – Teleprinter letter code as it appears on perforated tape; start and stop elements do not appear. Elements are numbered from top to bottom; dots indicate marking pulses. Numerals, punctuation, and other arbitrary symbols are secured by carriage shift. There are no lower-case letters on a teletypewriter using this 5-unit code.

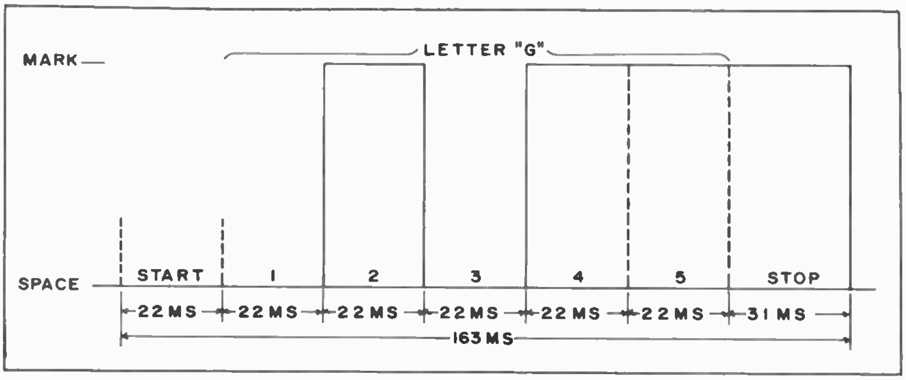


Fig. 12 – Pulse sequence in the teleprinter code. Each character begins with a start pulse, always a "space," and ends with a "stop" pulse, always a "mark." The distribution of marks and spaces in the five elements between start and stop determines the particular character transmitted.

circuit load in watts) if a correct wattage rating is selected for the series resistor. These calculations are simple, and basic algebra is all one needs to know for the work.

There are three conditions related to regulation. One is where the input voltage to the regulator is assumed to be constant, but where the load current will vary. The method for determining the right value for R_s in this case is

$$R_s = \frac{E_{in} - E_{VR}}{I_{L\,(max)} + 0.1\, I_{L\,(max)}} \quad \text{(Eq. 1)}$$

where R_s = series resistance, E_{in} = input voltage, E_{VR} = Zener-diode voltage, and I_L = load current in amperes.

The maximum power taken by the Zener diode can be found from

$$P_{D\,(max)} = \left(\frac{E_{in} - E_{VR}}{R_s} - I_{L\,(min)}\right) E_{VR} \quad \text{(Eq. 2)}$$

where P_D = diode dissipation in watts, E_{in} = input voltage, E_{VR} = Zener-diode voltage, R_s = series resistance in ohms, and I_L = load current in amperes.

In situations where E_{in} is variable and I_L is constant, R_s can be found from

$$R_s = \frac{E_{in\,(min)} - E_{VR}}{I_L + 0.1\, I_L} \quad \text{(Eq. 3)}$$

In this case one can obtain P_D from

$$P_{D\,(max)} = \left(\frac{E_{in\,(max)} - E_{VR}}{R_s} - I_{L\,(min)}\right) E_{VR} \quad \text{(Eq. 4)}$$

A third possibility exists where both E_{in} and I_L are variable. In that situation one can calculate R_s from

$$R_s = \frac{E_{in\,(min)} - E_{VR}}{I_{L\,(max)} + 0.1\, I_{L\,(max)}} \quad \text{(Eq. 5)}$$

and P_D can be obtained from

$$P_{D\,(max)} = \left(\frac{E_{in\,(max)} - E_{VR}}{R_s} - I_{L\,(min)}\right) E_{VR} \quad \text{(Eq. 6)}$$

A practical example of some calculations is found in Fig. 7, where a 9.1-volt regulated feed is desired for a VFO which draws 10 mA. The supply voltage to the regulator is one which varies between 12 and 14 volts during mobile service. R_s comes out as 264 ohms. Since this is not a standard value one may safely use a 220-ohm value. It is better to use the next lower resistance value, for if a 270-ohm resistor was selected the regulation at 12 volts might suffer.

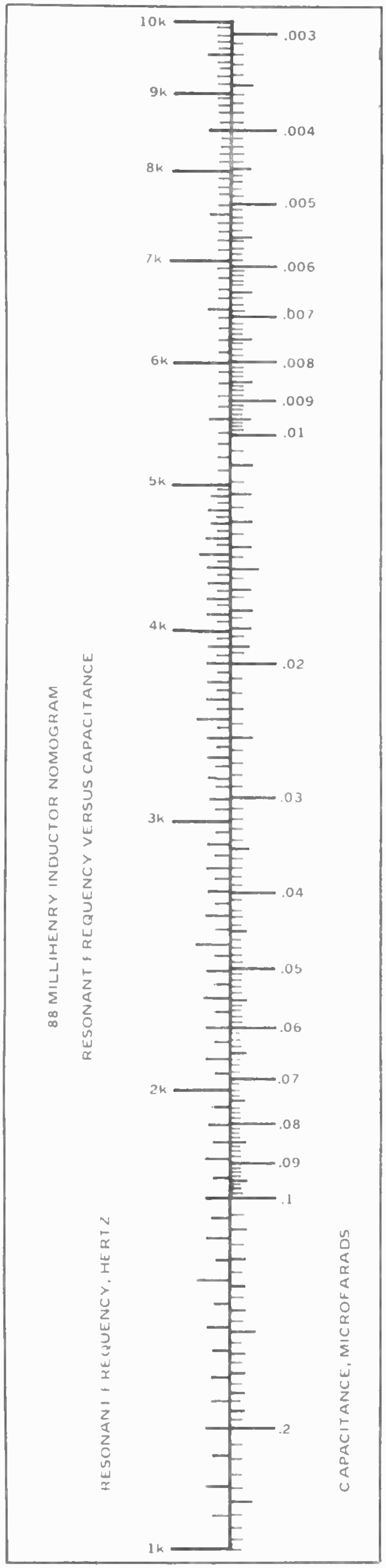


Fig. 13 — Nomograph of capacitor values versus resonant frequency for 88-mH toroidal inductors.

A matter of importance is the wattage rating of the Zener diode. It was found by the formula for P_D that VR1 will consume approximately 78 milliwatts (.078 W). A Zener diode should operate as cool as possible to assure longevity and proper performance. Therefore, one should choose a diode wattage well in excess of 78 mW. A safe rule of thumb is to select a diode wattage 5 times or more the diode dissipation amount. Based on this rule a 400-mW Zener diode is suitable. A 1-W diode would be a good choice also.

Finally, the safe wattage rating of the series resistor, R_s, must be determined. This is a relatively simply calculation, and the information required can be obtained from the data relating to Fig. 7. The following equation will be suitable for finding the power dissipated in R_s

$$P_D\ (R_s) = \frac{E_{drop}{}^2}{R_s(\text{ohms})} = \frac{4.9\ \text{V}^2}{220}$$

$$= 0.109\ \text{W} \qquad \text{(Eq. 7)}$$

where, E_{drop} = the voltage drop across R_s (14 V – 9.1 V), and R_s = the value of the series resistor in ohms. To assure ample power rating in the resistor, the five-times power rule specified earlier is applied. Therefore, a rating of 0.545 watt would suffice. Since that is very close to a standard 1/2-watt rating, we can employ a half-watt resistor. However, any wattage rating greater than 0.5 would also be acceptable, provided space for the large component was available.

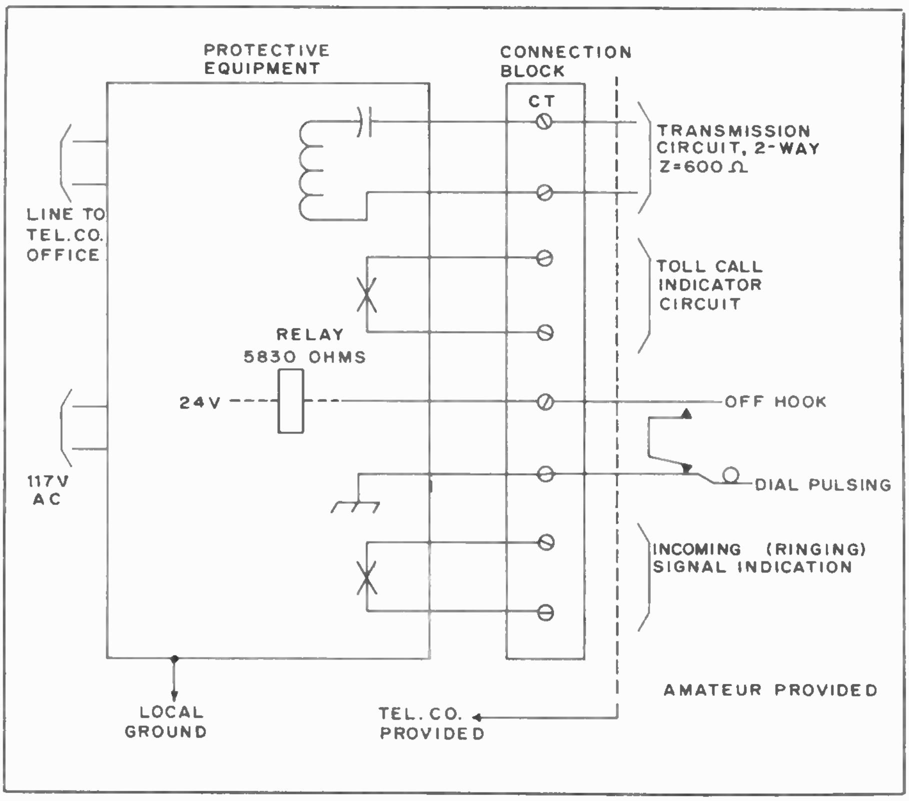
Table 1 lists several Zener diodes with voltage ratings from 3.9 to 180. Power ratings run from 400 mW to 50 W. The stud-mounted diodes should be affixed to a heat sink, and a coating of silicone heat-sink compound should be used between the Zener-diode body and the heat sink to aid effective heat transfer.

In the interest of practicality, the tolerance of the resistor or resistors used is not a matter of prime significance. The results obtained will be entirely adequate if 10-percent resistors are employed.

Protective Diodes

Selenium or Zener diodes can be used to clip excess ac or peak dc voltages in circuits where some safe maximum

Fig. 14 — Interconnection diagram for a Bell CD8 coupler, representative of connections to unattended interface devices.



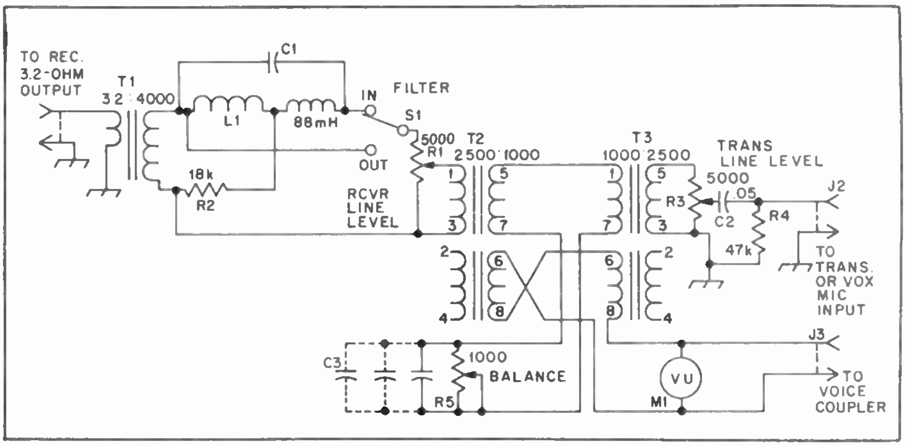
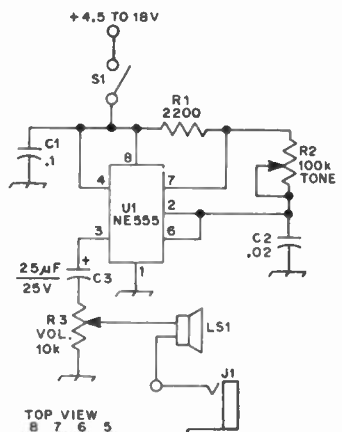


Fig. 15 – Schematic diagram of the phone-patch circuit. Resistances are in ohms; k = 1000. Fixed resistors may be 1/2 watt, 10-percent tolerance. Capacitance is in microfarads.

C1 – Capacitors in parallel to give required value of .0427 μF; low-voltage metallized paper or Mylar are suitable.
C3 – Typical value, .04 μF.
J1, J2, J3 – Phono jack, J3 should be insulated from chassis.
L1 – Surplus 88-mH toroidal inductor, connected with half-windings in series aiding.
M1 – Calectro DI-930 VU meter, modified.[1]
R1, R3 – 5000-ohm audio-taper control (Mallory U4 or equiv.).
T1 – Audio transformer, 4 or 8 ohms to 4000 ohms (UTC SO-10 or equiv.).
T2, T3 – Audio transformer, 2500-ohm split primary, 1000-ohm split secondary (UTC 0-19 or equiv.).

[1] M1 is a modified Calectro model DI-930 "VU" meter, as shown in Fig. 2. In early models, the existing 7,000-ohm multiplier resistor must be replaced by a 365-ohm 1-percent precision resistor. Later models, which may be identified by the letter A appearing in a circle near the bottom of the meter-scale card, are supplied with a 300-ohm resistor which need not be changed. Damping capacitors must be added across the meter coil, observing proper polarity. For early models of the DI-940 meter, the correct capacitance value is 300 μF; for later models with the circled A appearing on the meter-scale card, the required value is 400 μF. These values apply only to this particular make and models of meter. The meter, as modified, has a 1-kHz impedance of approximately 6500 ohms. It should be mounted only on a nonferrous panel.

Schematic diagram for the code oscillator. Resistance is in ohms, k = 1000. The 0.1-μF capacitor is a disk ceramic. U1 is a Signetics NE555 IC timer.



A	didah	**S**	dididit
B	dahdididit	**T**	dah
C	dahdidahdit	**U**	dididah
D	dahdidit	**V**	didididah
E	dit	**W**	didahdah
F	dididahdit	**X**	dahdididah
G	dahdahdit	**Y**	dahdidahdah
H	didididit	**Z**	dahdahdidit
I	didit	**1**	didahdahdahdah
J	didahdahdah	**2**	dididahdahdah
K	dahdidah	**3**	didididahdah
L	didahdidit	**4**	dididididah
M	dahdah	**5**	dididididit
N	dahdit	**6**	dahdididídit
O	dahdahdah	**7**	dahdahdididit
P	didahdahdit	**8**	dahdahdahdidit
Q	dahdahdidah	**9**	dahdahdahdahdit
R	didahdit	**0**	dahdahdahdahdah

Period: didahdidahdidah. Comma: dahdahdididahdah. Question mark: dididahdahdidit. Error: didididididididit. Double dash: dahdidididah. Colon: dahdahdahdididit. Semicolon: dahdidahdidahdit. Parenthesis: dahdidahdahdidah. Fraction bar: dahdididahdit. Wait: didahdididit. End of message: didahdidahdit. Invitation to transmit: dahdidah. End of work: didididahdidah.
The Continental (International Morse) Code.

Fig. 16 – Listing of Continental Morse Code characters. A schematic diagram is provided for those wishing to build a simple code-practice oscillator.

peak-voltage amount must be assured. A typical case is one in which the primary of a transformer in an ac power supply is protected from line transients which might lead to the demise of the silicon diodes in the rectifier circuit. A Klip-Sel (International Rectifier) or Thyrector (G.E.) protective-diode assembly can be bridged across the transformer primary (117 or 220 volts). The ampere rating of the diode unit is based on the power magnitude of the expected voltage peaks. The diodes with low ampere ratings could be burned out by transients of high current. Table 2 lists a variety of Klip-Sel (selenium) protective diodes for the application just described.

Another area where transient-suppression is of concern is seen in switch and relay circuits, where inductive loads can cause high transient voltages which may lead to the gradual or instant destruction of switch or relay contacts. This kind of event takes place mainly in ac or dc power-switching circuits. Table 3 contains a list of suitable suppressors. The voltage rating of the suppressor should be higher than the voltage being switched.

Small-Signal Diodes

Table 4 lists a number of general-purpose small-signal diodes which are useful for many amateur applications. The 1N82AG is suitable for vhf and uhf work. For use in circuits which call for matched diodes, the 1N542 is a good choice (discriminators, ratio detectors, balanced mixers, etc.).

Table 5 contains a listing of various types of silcion power diodes for use as rectifiers in power supplies. Details for making the proper selection for a particular application are given in *The Radio Amateur's Handbook*, Power Supply chapter.

Chapter 9 of this book included the description of a motor-speed control circuit which uses a Triac. Table 6 lists some general-purpose SCRs and Triacs for applications such as motor-speed controls, light dimmers, etc. At the bottom of the table is a recommended unijunction transistor for general-purpose applications.

Table 7 provides information on estimated hours of service for batteries. Size AA, C and D cells are treated.

Specialized-Techniques Data

Vhf and uhf operators may find the information of Fig. 8 useful in determining the kind of coverage that can be expected in that part of the amateur spectrum. Ionospheric characteristics are depicted at A, and path-loss information is provided at B and C. Additional information is available in *The Radio*

UNITED STATES RADIO DISTRICTS

Address the District FCC Engineer in Charge

1) India & State Sts.; Boston, MA 02109
2) 201 Varick St., New York, NY 10014
3) 601 Market St.; Philadelphia, PA 19106
4) 823 Geo. M. Fallon Federal Bldg., 31 Hopkins Plaza, Baltimore, MD 21201
5) Military Circle, 870 North Military Hwy. Norfolk, VA 23502
6) 1365 Peachtree St., N.E., Rm. 440, Atlanta, GA 30309
6S) Bull & State Sts., P. O. Box 8004, Savannah, GA 31402
7) 51 S.W. First Ave., Miami, FL 33130
7T) 500 Zack St., Tampa, FL 33602
8) 600 South St., Rm. 829, New Orleans, LA 70130
8M) 113 St. Joseph St., Mobile AL 36602
9) 515 Rusk Ave., Houston, TX 77002
9B) Rm. 323, Fed. Bldg., 300 Willow St., Beaumont, TX 77701
10) Rm. 13E7, Fed. Bldg., 1100 Commerce St., Dallas, TX 75202
11) Suite 501, Long Beach Blvd., Long Beach, CA 90807
11SD) 1245 Seventh Ave., San Diego, CA 92101
12) 555 Battery St., San Francisco, CA 94111
13) 1782 Fed. Office Bldg., 1220 S.W. 3rd Ave., Portland, OR 97204
14) 3256 Fed. Office Bldg., 915 2nd Ave., Seattle, WA 98174
15) Suite 2925, The Executive Tower, 1405 Curtis St., Denver, CO 80202
16) 316 N. Robert St., St. Paul, MN 55101
17) 601 E. 12th St., 1703 Fed. Bldg., Kansas City, MO 64106
18) Rm. 3935, 230 S. Dearborn St., Chicago, IL 60604
19) 1054 Fed. Bldg., 231 W. Lafayette St., Detroit, MI 48225
20) 1307 Fed. Bldg., 111 W. Huron St., Buffalo, NY 14204
21) 502 Fed. Bldg., Honolulu, HI 96808
22) 323 U.S. Post Office and Court House, P. O. Box 2987, San Juan, PR 00903
23) U.S. Post Office Bldg., Rm. G-63, 4th and G Sts., P. O. Box 644, Anchorage, AL 99510
24) 1919 M St., N.W., Rm 411, Washington, DC 20554

Fig. 17 – U.S. map showing the various FCC radio districts.

Amateur's VHF Manual and *The ARRL Antenna Book*.

Those interested in satellite work will find the data of Fig. 9 useful. The graphic illustrations cover doppler shift, map range, and altitude versus period.

Fig. 10 contains a suggested frequency setup for amateur 2-meter repeaters. The latest information is available from local fm/repeater frequency coordinators.

Meteor-shower enthusiasts can refer to the detailed listing of Table 8 for data on when to expect meteor showers. The notes at the bottom of the table explain the frequency of the occurrences.

Table 9 provides amateur slow-scan TV standards. The ARRL *Specialized Techniques Book* treats the subject of SSTV in considerable depth, and is recommended to those who are interested in the mode.

RTTY

RTTY operators find frequent use for perforated tapes. Fig. 11 shows how the perforations compare to the alphabet. Mark and space intervals are illustrated in Fig. 12.

Table 10 contains pertinent data on the USASCII 8-unit RTTY code. It is hoped that the 8-unit code will eventually be permitted in amateur service. At present the FCC requires that U.S. amateurs employ the 5-unit code.

A handy nomograph is offered in Fig. 13 for those who use 88-mH telephone toroids in RTTY work. It permits the builder to compare resonant frequency with the capacitance values on the right side of the scale. Additional data on 88-mH toroidal inductors are given in Table 11.

Phone Patching

Table 12 shows the maximum permissible energy levels at the input of a voice interconnection patch setup. Details for a Bell System CD8 coupler are given in Fig. 14.

Fig. 15 contains the diagram of an amateur phone patch. Phone patches are used to provide service to overseas persons and amateurs, and are put into service also during times of emergency. Many amateur repeaters feature patches for phone communications from the vehicle to fixed points. Table 13 provides detailed data concerning signal and circuit conditions in the telephone system.

Table 1

ZENER VOLTAGE	LEAD MOUNTED PART NO. 400 mW	PART NO. 1 WATT	PART NO. 1.5 WATT	PART NO. 5 WATT	STUD MOUNTED (HARDWARE INCLUDED) PART NO. 10 WATT	PART NO. 50 WATT
3.9	Z1002	Z1100	Z1200	Z1500	Z1300†*	–
4.7	Z1004	Z1102	Z1202	Z1502	Z1302†*	–
5.1	–	Z1103	Z1203	Z1504	–	–
5.6	Z1006	Z1104	Z1204	Z1506	Z1304 *	–
6.8	Z1008	Z1106	Z1206	Z1507	Z1306 *	Z3305
7.5	–	Z1107	Z1207	Z1508	–	–
8.2	Z1010	Z1108	Z1208	Z1510	Z1308 *	Z3307
9.1	–	Z1109	Z1209	Z1512	–	–
10	Z1012	Z1110	Z1210	Z1514	Z1310 *	Z3309
12	Z1014	Z1112	Z1212	Z1516	Z1312 *	Z3311
15	Z1016	Z1114	Z1214	Z1518	Z1314 *	Z3314
18	Z1018	Z1116	Z1216	Z1520	Z1316 *	Z3317
22	Z1020	Z1118	Z1218	Z1522	Z1318 *	Z3320
27	Z1022	Z1120	Z1220	Z1524	Z1320 *	Z3323
33	Z1024	Z1122	Z1222	Z1526	Z1322	Z3325
39	–	Z1124	–	–	Z1324	Z3327
47	–	Z1126	–	Z1528	Z1326	–
50	–	–	Z1226	–	–	–
51	–	–	–	Z1530	–	–
56	–	Z1128	–	–	Z1328	Z3334
68	–	Z1130	–	–	Z1330	–
82	–	Z1132	–	–	Z1332	–
100	–	Z1134	–	Z1532	Z1334	Z3340
120	–	Z1136	–	–	Z1336	–
150	–	Z1138	Z1238	–	Z1338	–
180	–	Z1140	–	–	Z1340 *	–
Case Types	DO-7	DO-13	DO-41	C-12	DO-4	C-8

*Units have cathode stud. All others have anode stud. Courtesy International Rectifier

Table 2
Klip-Sel Protective Diodes

DEVICE VOLTAGE RATING (V)	MAX. RMS OPERATING VOLTS (V)	MAX. CLAMPING VOLTS @ PEAK CURRENT	CAT. NO. FOR INDIVIDUAL RECTIFIER RATED THRU 3 AMP	CAT. NO. FOR INDIVIDUAL RECTIFIER RATED THRU 16 AMP	CAT. NO. FOR INDIVIDUAL RECTIFIER RATED THRU 60 AMP (S-6A CASE)
50	26	70	KY1DPF	KZ1DPF	KSA1DAF
100	52	140	KY2DPF	KZ2DPF	KSA2DAF
150	78	210	KY3DPF	KZ3DPF	KSA3DAF
200	104	280	KY4DPF	KZ4DPF	KSA4DAF
300	156	420	KY6DPF	KZ6DPF	KSA6DBF
500	234	630	KY9DPF	KZ9DPF	KSA9DBF
600	286	770	KY11DPF	KZ11DPF	KSA11DBF
800	364	980	KY14DPF	KZ14DPF	KSA14DBF
900	416	1120	KY16DPF	KZ16DPF	KSA16DBF

Two each required. Courtesy International Rectifier

Table 3
Relay and Switch Protectors

CAT. NO.	MAX. WORKING VOLTAGE AC	DC	MAX. COIL CURRENT (AMPS)	CAT. NO.	MAX. WORKING VOLTAGE AC	MAX. COIL CURRENT (AMPS)
S1V1P	26		.20	S5V2P	110	.25
S5V5P	130		.20	S6V2P	132	.25
S1Z1P	26		.60	S7V2P	154	.25
S5Z5P	130		.60	S1Z1P	22	.90
S2W2P	52		1.20	S2Z1P	44	.90
S3W3P	78		1.20	S3Z2P	66	.90
S4W4P	104		1.20	S4Z2P	88	.90
S5W5P	130		1.20	S5Z2P	110	.90
S6W6P	156		1.20	S6Z2P	132	.90
S1V1P		22	.25	S7Z2P	154	.90
S2V1P		44	.25	S1W1P	22	2.00
S3V2P		66	.25	S2W1P	42	2.00
S4V2P		88	.25			

Courtesy International Rectifier

Table 4
Germanium Point-Contact Diodes

Miniature, hermetically sealed point-contact construction provides optimum operation over a wide range of humidity, temperature, and other environmental extremes. The color bands indicate cathode (positive). Clear glass case within DO-7 limits. Color code indicates catalog number.

PART NO.	PEAK REVERSE VOLTAGE	FORWARD CURRENT (mA)
1N34A	60	50
1N48	70	50
1N60	40	50
1N64	25	50
1N67A	80	30
1N82AG*	25	25
1N87A	30	50
1N294	70	60
1N295	50	35
1N541	45	35
1N542**	–	–
1N636A	60	–

*Silicon uhf diode. Courtesy Int'l. Rectifier
**This is a matched pair of 1N541s.

Table 5
Silicon Power Rectifiers

Hermetically sealed high-reliability power rectifiers for commercial and replacement use. Rated from 3 to 40 amperes.

PART NO.	CURRENT RATING (A)	PEAK REVERSE VOLTAGE	RMS INPUT VOLTAGE RESISTANCE LOAD	DC OUTPUT (A)	CASE STYLE
3F20D	3	200	140	3	DO-4
3F40D	3	400	280	3	DO-4
3F60D	3	600	420	3	DO-4
3F80D	3	800	560	3	DO-4
3F100D	3	1000	700	3	DO-4
6F20D	6	200	140	6	DO-4
6F40D	6	400	280	6	DO-4
6F60D	6	600	420	6	DO-4
6F80D	6	800	560	6	DO-4
6F100D	6	1000	700	6	DO-4
12F10*	12	100	70	12	DO-4
12F20*	12	200	140	12	DO-4
12F40*	12	400	280	12	DO-4
12F60*	12	600	420	12	DO-4
12F80*	12	800	560	12	DO-4
12F100*	12	1000	700	12	DO-4
40HF5*	40	50	35	40	DO-5
40HF10*	40	100	70	40	DO-5
40HF20*	40	200	140	40	DO-5
40HF30*	40	300	210	40	DO-5
40HF40*	40	400	280	40	DO-5
40HF60*	40	600	420	40	DO-5
40HF80*	40	800	560	40	DO-5
40AF100*	40	1000	700	40	DO-5
70H10A*	70	100	70	70	DO-5
70H20A*	70	200	140	70	DO-5
70H30A*	70	300	210	70	DO-5
70H40A*	70	400	280	70	DO-5
70H60A*	70	600	420	70	DO-5
70H80A*	70	800	560	70	DO-5
70H100A*	70	1000	700	70	DO-5
150L10A*	150	100	70	150	DO-8
150L20A*	150	200	140	150	DO-8
150L30A*	150	300	210	150	DO-8
150L40A*	150	400	280	150	DO-8
150L60A*	150	600	420	150	DO-8
150L80A*	150	800	560	150	DO-8
150L100A*	150	1000	700	150	DO-8
150L120A*	150	1200	840	150	DO-8

*Reverse polarity types available. Designated by adding "R" to IR Cat. No. e.g., "12FR80, 40HFR40, 70HR40, 150KR20"

Table 6
General-Application SCRs

Selected 9-A power SCRs to meet most amateur needs. Units are suited for a wide range of circuits including light dimmers, motor-speed controls, solid-state relays, model-railroad circuits, science projects and more. Units have 10-32 inch threaded stud base with mounting hardware included. Minimum gate excitation requirements are 90 mA at 3 volts. Case style E-64.

PART NO.	REVERSE VOLTAGE
SCR03	50
SCR04	200

Triacs and Diacs

Triac IRT82

An ac switch for motor controls, heater controls, etc. 200 V max., 8 A max. current. 100 A surge capability. Lead 3 is gate terminal. Mounting tab is isolated.

Diac IRD54

For triggering IRs Triac. 32 volts – typical V(BO) Breakover voltage. 500 mW power rating. 50 nA(BO) Breakover current.

PART NO.	CASE STYLE
IRT82	E-25
IRD54	C-10

Unijunction Transistor

IR2160

PN Type Unijunction Transistor similar to JEDEC 2N2160 for timing and triggering SCRs. V_{B2E} is rated 30 volts; V_{B2B1}, 35 volts. Case style T092-E3.

Courtesy International Rectifier

Table 7
Estimated Hours of Service at 70° F

AA CELL

SCHEDULE HRS./DAY	DRAIN mA	MINIMUM PERMISSIBLE VOLTAGE 0.8 V	1.0 V	1.2 V
2	2	350	300	250
2	10	54	43	32
2	50	5.5	3.5	1.1
4	2	310	270	220
4	10	49	40	28
4	30	10	7.8	3.4
12	2	320	260	200
12	10	48	37	24
12	30	9	5.9	2.9

C CELL

SCHEDULE HRS./DAY	DRAIN mA	MINIMUM PERMISSIBLE VOLTAGE 0.8 V	1.0 V	1.2 V
2	10	275	220	140
2	30	100	74	60
2	100	23	17	7.8
4	10	310	240	
4	30	96	68	
4	100	20	10	
12	10	330	250	180
12	30	90	64	30
12	100	18	8	3.5

D CELL

SCHEDULE HRS./DAY	DRAIN mA	MINIMUM PERMISSIBLE VOLTAGE 0.8 V	1.0 V	1.2 V
2	30	210	175	135
2	100	57	45	29
2	300	11	7	2
4	30	220	185	125
4	100	50	36	22
4	300	8	3.5	2
24	30	200	115	65
24	100	32	18.5	9.6
24	300	6	3.5	2

Table 8

Meteor-Shower Data for VHF Use

SHOWER	*TIME VISIBLE*		*OPTIMUM PATHS AND TIMES*				*HOURLY RATE*		*VELOCITY*	*PERIOD*	*NEXT*
AND DATE	*RISE*	*SET*	*N-S*	*NW-SE*	*E-W*	*SW-NE*	*VISUAL*	*RADIO*	*KM/SEC.*	*YEARS*	*MAX.*
*January 3-5 Quadrantids	2300	1800	—	0300-0800 SW	0800-0900 S	0900-1400 SE	35	45	45	7	Note 1
January 17 Cygnids	0230	2130	—	0600-1100 SW	1100-1300 S	1300-1800 SE	—	—	—	—	—
February 5-10 Aurigids	1200	0330	—	1400-1730 SW	—	2130-0100 SE	—	—	—	—	—
March 10-12 Bootids	2200	0830	2330-0030 W 0530-0630 E	0330-0530 NE	0230-0330 N	0030-0230 NW	—	—	—	—	—
March 20 Coma Berenices	1800	0630	2130-2300 W 0100-0300 E	2000-2130 SW	—	0300-0430 SE	—	—	—	—	—
*April 19-23 Lyrids	2100	1100	0230 W 0530 E	2330-0100 SW	—	0700-0830 SE	8	12	51	415	Note 1
*May 1-6 Aquarids	0300	1200	—	0830-1000 NE	0630-0830 N	0500-0630 NW	12	12	66	76	Note 1
May 11-24 Herculids	1800	0630	2130-2300 W 0100-0300 E	2000-2130 SW	—	0300-0430 SE	—	—	—	—	—
May 30 Pegasids	2300	1200	0300-0430 W 0630-0800 E	0130-0300 SW	—	0800-0930 SE	—	—	—	—	—
June 2-17 Scorpiids	2000	0300	—	0100 NE	2300-2400 N	2200 NW	—	—	—	—	—
June 27-30 Pons Winnecke	Does not set; min. at 0900		—	1500-1830 SW	1830-2330 S	2330-0300 SE	—	—	—	—	—
July 14 Cygnids	1800	1000	—	2100-2330 SW	0130 S	0330-0600 SE	—	—	—	—	—
July 18-30 Capricornids	2030	0400	—	0100-0200 NE	2300-0100 N	2200-2300 NW	—	—	—	—	—
*July 26-31 Aquarids	2200	0600	—	0300-0500 NE	0100-0300 N	0000-0100 NW	10	22	50	3.6	Note 1
*July 27-Aug. 14 Perseids	Does not set; min. at 1730		—	2330-0300 SW	0300-0800 S	0800-1130 SE	50	50	61	120	Note 1
August 10-20 Cygnids	1200	0700	—	1700-1930 SW	2130 S	2330-0200 SE	—	—	—	—	—
August 21-23 Draconids	Does not set min. at 0900		—	1500-1830 SW	1830-2330 S	2330-0300 SE	—	—	—	—	—
Aug. 21-31 Draconids	Does not set; min. at 0700		—	1300-1630 SW	1630-2130 S	2130-0100 SE	—	—	—	—	—
September 7-15 Perseids	2130	1200	—	0030-0200 SW	—	0700-0830 SE	—	—	—	—	—
September 22 Aurigids	2100	1230	—	0030-0200 SW	—	0700-0830 SE	—	—	—	—	—

Table 9

Amateur Slow-Scan Standards

	60-Hz AREAS	*50-Hz AREAS*
Sweep Rates:		
Horizontal	15 Hz (60 Hz/4)	16-2/3 Hz (50 Hz/3)
Vertical	8 s	7.2 sec.
No. of Scanning Lines	120	120
Aspect Ratio	1:1	1:1
Direction of Scan:		
Horizontal	Left to Right	Left to Right
Vertical	Top to Bottom	Top to Bottom
Sync Pulse Duration:		
Horizontal	5 ms	5 ms
Vertical	30 ms	30 ms
Subcarrier Freq.:		
Sync	1200 Hz	1200 Hz
Black	1500 Hz	1500 Hz
White	2300 Hz	2300 Hz
Req. Trans.:		
Bandwidth	1.0 to 2.5 kHz	1.0 to 2.5 kHz

Table 10

U.S. American Standard Code for Information Interchange

The USA Standard Code for Information Interchange (USASCII, or, more commonly, ASCII) is an 8-unit code as shown here, used largely with computers. Some Teletype Corp. machines, such as Models 33 and 35, use this

NUL	Null, or all zeros
SOH	Start of heading
STX	Start of text
ETX	End of text
EOT	End of transmission
ENQ	Enquiry
ACK	Acknowledge
BEL	Bell, or alarm
BS	Backspace
HT	Horizontal tabulation
LF	Line feed
VT	Vertical tabulation
FF	Form feed
CR	Carriage return
SO	Shift out
SI	Shift in
DLE	Data link escape

SHOWER	TIME VISIBLE		OPTIMUM PATHS AND TIMES				HOURLY RATE		VELOCITY	PERIOD	NEXT
AND DATE	RISE	SET	N-S	NW-SE	E-W	SW-NE	VISUAL	RADIO	KM/SEC.	YEARS	MAX.
October 2 Quadrantids	0500	0000	–	0900-1400 SW	1400-1500 S	1500-2000 SE	–	–	–	–	–
October 9 Giacobinids	0600	0300	–	1100-1600 SW	1600-1700 S	1700-2200 SE	Note 2		20	6.6	1979
October 12-23 Arietids	1900	0700	2130-2330 W 0230-0430 E	–	–	–	–	–	–	–	–
*October 18-23 Orionids	2230	0930	0000-0200 W 0600-0800 E	0430-0600 NE	0330-0430 N	0200-0330 NW	15	30	68	76	Note 1
*Oct. 26-Nov. 16 Taurids	1900	0630	2100-2300 W 0300-0500 E	0130-0300 NE	0030-0130 N	2300-0030 NW	10	16	27	3.3	Note 1
*November 14-18 Leonids	0000	1230	0300-0500 W 0800-1000 E	–	–	–	12	Note 3	72	33.2	1999
November 22-30 Andromedids	1300	0600	–	1600-2000 SW	–	2300-0300 SE	Note 4	–	22	6.7	1978
*December 10-14 Geminids	1900	0900	0030 W 0330 E	2130-2300 SW	–	0500-0630 SE	60	70	35	1.6	Note 1
*December 22 Ursids	Does not set; min. at 2030		–	–	0130-1530 S	–	13 Note 5	13	38	13.5	1984 1984
*May 19-21 Cetids	0530	1430	–	1100-1230 NE	0900-1100 N	0730-0900 NW	–	–	20	37	–
*June 4-6 Perseids	0500	1730	0800-1000 W 1300-1500 E	–	–	–	–	–	40	29	–
*June 8 Arietids	0330	1530	0600-0800 W 1100-1300 E	–	–	–	Note 6	–	70	38	–
*June 30-July 2 Taurids	0500	1700	0700-0900 W 1300-1500 E	1130-1300 NE	1030-1130 N	0900-1030 NW	–	–	30	31	–

*Major showers — Last four are daylight showers.

NOTES

1) These streams are evenly distributed and little year-to-year variation is to be expected.

2) Very concentrated stream. Peak years give up to 400 meteors per minute, but with duration of only 6 hours. 1946 peak was most concentrated shower in amateur radio experience up to that time (see December, 1946, *QST*, page 43) but 1959 recurrence was deflected and was hardly observable.

3) Peak years give 60/hour visual. In the peak years of the 1800s, prior to being deflected by Jupiter and Saturn, this shower gave 1200 per minute. Spectacular results in 1965 and 1966 are reported in Jan. 1966 *QST*, page 80, and Jan. 1967, page 83.

4) Before being deflected by Jupiter this stream gave peak year rates of 100/minute. No notable rates have been observed since, though the stream could return.

5) Short duration shower. Peak years the radio rate is 165/hour.

6) This intense daylight shower begins June 2 and runs to June 14 with radio rates from 25 to 70/hour.

code. Because FCC regulations require that U.S. amateurs use a 5-unit code for RTTY operation, operation of these machines is presently not permitted for amateur communications.

DC1	Device control 1
DC2	Device control 2
DC3	Device control 3
DC4	Device control 4
NAK	Negative acknowledge
SYN	Synchronous idle
ETB	End of transmission block
CAN	Cancel
EM	End of medium
SUB	Substitute
ESC	Escape
FS	File separator
GS	Group separator
RS	Record separator
US	Unit separator
SP	Space
DEL	Delete

American Standard Code for Information Interchange (ASCII)

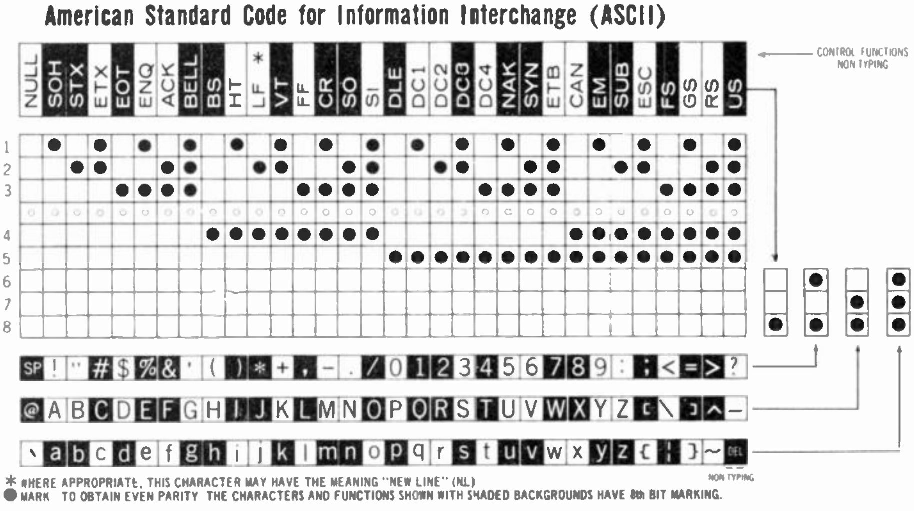


Table 11

Calibration Chart for 88-mH Toroid and Decade Box. Shift is Measured in Hertz with Respect to 2125 Hz.

These represent 500-pF steps on the decade capacitance box. The small figures between steps are the hertz for each 100 pF. For example, suppose the decade box reads 0.0321. This is close to 0.0320, so subtract 5 Hz from 874 for resultant final shift of 869. The decade capacitance box has 1-percent capacitors, and the practical results are usually within 2 to 3 Hz of these figures with an unmodified 88-mH toroid.

µF	*SHIFT*		*FREQ.*
.0290	1026	5	3151
.0295	999	5	3124
.0300	973	5	3098
.0305	947	5	3072
.0310	922	5	3047
.0315	898	5	3023
.0320	874	5	2999
.0325	850	5	2975
.0330	828	4	2953
.0335	806	4	2931
.0340	785	4	2910
.0345	763	4	2888
.0350	743	4	2868
.0355	723	4	2848
.0360	703	4	2828
.0365	683	4	2808
.0370	664	4	2789
.0375	646	4	2771
.0380	627	4	2752
.0385	609	4	2734
.0390	592	4	2717
.0395	574	3	2699
.0400	558	3	2683
.0405	541	3	2666
.0410	525	3	2650
.0415	509	3	2634
.0420	493	3	2618
.0425	477	3	2602
.0430	462	3	2587
.0435	447	3	2572
.0440	433	3	2558
.0445	418	3	2543
.0450	404	3	2529
.0455	390	3	2515
.0460	376	3	2501
.0465	363	3	2488
.0470	350	3	2475
.0475	337	3	2462
.0480	324	3	2449
.0485	311	2	2436
.0490	299	2	2424
.0495	286	2	2411
.0500	274	2	2399
.0505	262	2	2387
.0510	251	2	2376
.0515	239	2	2364
.0520	228	2	2353
.0525	217	2	2342
.0530	205	2	2330
.0535	195	2	2320
.0540	184	2	2309
.0546	170	2	2295
.0550	163	2	2288
.0555	152	2	2277
.0560	142	2	2267
.0565	132	2	2257
.0570	122	2	2247
.0575	112	2	2237
.0580	103	2	2228
.0585	93	2	2218
.0590	84	2	2209
.0595	74	2	2199
.0600	65	2	2190
.0605	56	2	2181
.0610	47	2	2172
.0615	38	2	2163
.0620	30	2	2155
.0625	21	2	2146
.0630	13	2	2138
.0635	4	2	2129
.0637	0		2125

Table 12

Maximum Permissible Energy Levels at the Input of a Voice Interconnection Arrangement

FREQ. BAND	*MAXIMUM LEVEL*
Direct current	0.5 milliampere
Voice range (nominally 300 to 3000 Hz)	Voice coupler: −3 dBm. Other arrangements: 9 dB below 1 mW (levels averaged over 3 seconds, see note)
2450 to 2750 Hz	Preferably no energy; in no case greater than the level present simultaneously in the 800- to 2450-Hz band.
3995 to 4005 Hz	18 dB below the voice-band level.
4.0 to 10.0 kHz	16 dB below one milliwatt (−16 dBm).
10.0 to 25.0 kHz	−24 dBm
25.0 to 40.0 kHz	−36 dBm
Above 40.0 kHz	−50 dBm

NOTE: The above limits should be met with amateur-provided equipment having an internal impedance of 900 ohms if it is to work into a voice coupler, or 600 ohms if other arrangements are to be used.

Table 13

Signals and Circuit Conditions Used in the Telephone System

1) The status of a local telephone line (idle or busy) is indicated by on-hook or off-hook signals as follows:

On-Hook — Minimum dc resistance between tip and ring conductors of 30,000 ohms.
Off-Hook — Maximum dc resistance between tip and ring conductors of 200 ohms.

Telephone sets give an off-hook condition at all times from the answer or origination of a call to its completion. The only exception to this is during dial pulsing.

2) Dial pulses consist of momentary opens in the loop; dial pulses should meet the following standards:

Pulsing rate	10 pulses/second ±10%
Pulse shape	58% to 64% break (open)
Interdigital time	600 milliseconds minimum

Note: Two pulses indicate the digit "2," three pulses indicate the digit "3," and so on, up to ten, indicating the digit "0."

3) The standards for tone "dialing" are as follows:

a) Each digit is represented by a unique pair of tones as shown below.

DIGIT	*LOW TONE*		*HIGH TONE*
1	697	and	1209 Hz
2	697	and	1336 Hz
3	697	and	1477 Hz
4	770	and	1209 Hz
5	770	and	1336 Hz
6	770	and	1477 Hz
7	852	and	1209 Hz
8	852	and	1336 Hz
9	852	and	1477 Hz
0	941	and	1336 Hz
*	941	and	1209 Hz
#	941	and	1477 Hz

b) In order for the central-office receiver to register the digit properly, the tone-address signals must meet the following requirements:

1) Signal levels:
Nominal level per frequency: −6 to −4 dBm. Minimum level per frequency: Low Group, −10 dBm; High Group, −8 dBm. Max. level per frequency pair; +2 dBm. Max. difference in levels between frequencies: 4 dB.

2) Frequency deviation: ±1.5 percent of the values given above.

3) Extraneous frequency components: The total power of all extraneous frequencies accompanying the signal should be at least 20 dB below the signal power, in the voice band above 500 Hz.

4) Voice Suppression: Voice energy from any source should be suppressed at least 45 dB during tone signal transmission. In the case of automatic dialing the suppression should be maintained continuously until pulsing is completed.

5) Rise Time: Each of the two frequencies of the signal should attain at least 90 percent of full amplitude within 5 ms, and preferably within 3 ms for automatic dialers, from the time that the frequency begins.

6) Pulsing Rate: Minimum duration of two-frequency tone signal: 50 ms normally; 90 ms if transmitted by radio. Minimum interdigital time: 45 ms.

7) Tone leak during signal off time should be less than −55 dBm.

8) Transient Voltages: Peak transient voltages generated during tone signaling should be no greater than 12 dB above the zero-to-peak voltage of the composite two-frequency tone signal.

4) Audible tones will be used in the telephone system to indicate the progress or disposition of a call. These include:

a) Dial tone: 350 and 440 Hz
b) Line busy: 480 and 620 Hz, interrupted at 60 interruptions per minute (I/min).
c) Reorder (all trunks busy); 480 and 620 Hz, interrupted at 120 I/min.
d) Audible ringing: 440 and 480 Hz, 2 seconds on, 4 seconds off.
e) Reserved high tone: 1633 Hz.
f) Invalid dialing code: Voice announcement.

Table 14

Conversion: Inches and Fractions to Nearest Millimeters

+	0	1/8	1/4	3/8	1/2	5/8	3/4	7/8
0	0	3	6	10	13	16	19	22
1	25	29	32	35	38	41	44	48
2	51	54	57	60	64	67	70	73
3	76	79	83	86	89	92	95	98
4	104	105	108	111	114	117	121	124
5	127	130	133	137	140	143	146	149
6	152	156	159	162	165	168	171	175
7	178	181	184	187	191	194	197	200
8	203	206	210	213	216	219	222	225
9	229	232	235	238	241	244	248	251
10	254	257	260	264	267	270	273	276
11	279	283	286	289	292	295	298	302

Conversion: Feet to Meters*

X	x1	x10	x100	x1000
1	.305	3.048	30.480	304.801
2	.610	6.096	60.960	609.601
3	.914	9.144	91.440	914.402
4	1.219	12.192	121.920	1219.202
5	1.524	15.240	152.400	1524.003
6	1.829	18.288	182.880	1828.804
7	2.134	21.336	213.360	2133.604
8	2.438	24.384	243.840	2438.405
9	2.743	27.432	274.321	2743.205

*Note: For most purposes these values may be rounded to the nearest tenth. If dimensions in feet, inches and a fraction are to be converted, then the three decimal places may be justified.

Table 15

Some Abbreviations for CW Work

Abbreviations help to cut down unnecessary transmission. However, make it a rule not to abbreviate unnecessarily when working an operator of unknown experience.

AA	All after
AB	All before
ABT	About
ADR	Address
AGN	Again
ANT	Antenna
BCI	Broadcast interference
BCL	Broadcast listener
BK	Break; break me; break in
BN	All between; been
BUG	Semi-automatic key
B4	Before
C	Yes
CFM	Confirm; I confirm
CK	Check
CL	I am closing my station; call
CLD-CLG	Called; calling
CQ	Calling any station
CUD	Could
CUL	See you later
CUM	Come
CW	Continuous wave (i.e., radiotelegraph)
DLD-DLVD	Delivered
DR	Dear
DX	Distance, foreign countries
ES	And, &
FB	Fine business, excellent
FM	Frequency modulation
GA	Go ahead (or resume sending)
GB	Good-by
GBA	Give better address
GE	Good evening
GG	Going
GM	Good morning
GN	Good night
GND	Ground
GUD	Good
HI	The telegraphic laugh; high
HR	Here, hear
HV	Have
HW	How
LID	A poor operator
MA, MILS	Milliamperes
MSG	Message; prefix to radiogram
N	No
NCS	Net control station
ND	Nothing doing
NIL	Nothing; I have nothing for you
NM	No more
NR	Number
NW	Now; I resume transmission
OB	Old boy
OC	Old chap
OM	Old man
OP-OPR	Operator
OT	Old timer; old top
PBL	Preamble
PSE	Please
PWR	Power
PX	Press
R	Received as transmitted; are
RCD	Received
RCVR (RX)	Receiver
REF	Refer to; referring to; reference
RFI	Radio frequency interference
RIG	Station equipment
RPT	Repeat; I repeat
RTTY	Radioteletype
RX	Receiver
SASE	Self-addressed, stamped envelope
SED	Said
SIG	Signature; signal
SINE	Operator's personal initials or nickname
SKED	Schedule
SRI	Sorry
SSB	Single sideband
SVC	Service; prefix to service message
T	Zero
TFC	Traffic
TMW	Tomorrow
TNX-TKS	Thanks
TT	That
TU	Thank you
TVI	Television interference
TX	Transmitter
TXT	Text
UR-URS	Your; you're; yours
VFO	Variable-frequency oscillator
VY	Very
WA	Word after
WB	Word before
WD-WDS	Word; words
WKD-WKG	Worked; working
WL	Well; will
WUD	Would
WX	Weather
XCVR	Transceiver
XMTR (TX)	Transmitter
XTAL	Crystal
XYL(YF)	Wife
YL	Young lady
73	Best regards
88	Love and kisses

Table 16

Phonetics

A – ALFA	N – NOVEMBER
B – BRAVO	O – OSCAR
C – CHARLIE	P – PAPA
D – DELTA	Q – QUEBEC
E – ECHO	R – ROMEO
F – FOXTROT	S – SIERRA
G – GOLF	T – TANGO
H – HOTEL	U – UNIFORM
I – INDIA	V – VICTOR
J – JULIETT	W – WHISKEY
K – KILO	X – X-RAY
L – LIMA	Y – YANKEE
M – MIKE	Z – ZULU

Example: W1AW . . . W 1 ALFA WHISKEY . . . W1AW

Table 17

Q Signals

Given below are a number of Q signals whose meanings most often need to be expressed with brevity and clearness in amateur work. (Q abbreviations take the form of questions only when each is sent followed by a question mark.)

- **QRG** Will you tell me my exact frequency (or that of . . .)? Your exact frequency (or that of . . .) is . . . kHz.
- **QRH** Does my frequency vary? Your frequency varies.
- **QRI** How is the tone of my transmission? The tone of your transmission is . . . (1. Good; 2. Variable; 3. Bad).
- **QRK** What is the intelligibility of my signals (or those of . . .)? The intelligibility of your signals (or those of . . .) is . . . (1. Bad; 2. Poor; 3. Fair; 4. Good; 5. Excellent).
- **QRL** Are you busy? I am busy (or I am busy with . . .). Please do not interfere.
- **QRM** Is my transmission being interfered with? Your transmission is being interfered with . . . (1. Nil; 2. Slightly; 3. Moderately; 4. Severely; 5. Extremely.
- **QRN** Are you troubled by static? I am troubled by static . . . (1-5 as under QRM).
- **QRO** Shall I increase power? Increase power.
- **QRP** Shall I decrease power? Decrease power
- **QRQ** Shall I sendfaster? Send faster (. . . wpm).
- **QRS** Shall I send more slowly? Send more slowly (. . . wpm).
- **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 inform . . . that you are calling him on . . . kHz? Please inform . . . that I am calling on . . . kHz.
- **QRX** When will you call me again? I will call you again at . . . hours (on . . . kHz).
- **QRY** What is my turn? Your turn is number . . . .
- **QRZ** Who is calling me? You are being called by . . . (on . . . kHz).
- **QSA** What is the strength of my signals (or those of . . .)? The strength of your signals (or those of . . .) is . . . (1. Scarcely perceptible; 2. Weak; 3. Fairly good; 4. Good; 5. Very good).
- **QSB** Are my signals fading? Your signals are fading.
- **QSD** Are my signals mutilated? Your signals are mutilated.
- **QSG** Shall I send . . . messages at a time? Send . . . messages at a time.
- **QSK** Can you hear me between your signals and if so can I break in on your transmission? I can hear you between my signals; break in on my transmission.
- **QSL** Can you acknowledge receipt? I am acknowledging receipt.
- **QSM** Shall I repeat the last message which I sent you, or some previous message? Repeat the last message which you sent me (or message(s) number(s) . . .).
- **QSN** Did you hear me (or . . .) on . . . kHz? I did hear you (or . . .) on . . . kHz.
- **QSO** Can you communicate with . . . direct or by relay? I can communicate with . . . direct (or by relay through . . .).
- **QSP** Will you relay to . . .? I will relay to . . .
- **QSU** Shall I send or reply on this frequency (or on . . . kHz)? Send or reply on this frequency (or on . . . kHz)?
- **QSV** Shall I send a series of Vs on this frequency (or . . . kHz); Send a series of Vs on this frequency (or . . . kHz).
- **QSW** Will you send on this frequency (or on . . . kHz)? I am going to send on this frequency (or on . . . kHz).
- **QSX** Will you listen to . . . on . . . kHz? I am listening to . . . on . . . kHz..

Table 18
International Prefixes

AAA-ALZ	United States of America
AMA-AOZ	Spain
APA-ASZ	Pakistan
ATA-AWZ	India
AXA-AXZ	Commonwealth of Australia
AYA-AZZ	Argentine Republic
BAA-BZZ	China
CAA-CEZ	Chile
CFA-CKZ	Canada
CLA-CMZ	Cuba
CNA-CNZ	Morocco
COA-COZ	Cuba
CPA-CPZ	Bolivia
CQA-CRZ	Portuguese Overseas Provinces
CSA-CUZ	Portugal
CVA-CXZ	Uruguay
CYA-CZZ	Canada
DAA-DTZ	Germany
DUA-DZZ	Republic of the Philippines
EAA-EHZ	Spain
EIA-EJZ	Ireland
EKA-EKZ	Union of Soviet Socialist Rep.
ELA-ELZ	Liberia
EMA-EOZ	Union of Soviet Socialist Rep.
EPA-EQZ	Iran
ERA-ERZ	Union of Soviet Socialist Rep.
ESA-ESZ	Estonia
ETA-ETZ	Ethiopia
EUA-EWZ	Bielorussian Soviet Socialist Rep.
EXA-EZZ	Union of Soviet Socialist Rep.
FAA-FZZ	France and French Community
GAA-GZZ	United Kingdon
HAA-HAZ	Hungarian People's Republic
HBA-HBZ	Switzerland
HCA-HDZ	Ecuador
HEA-HEZ	Switzerland
HFA-HFZ	People's Republic of Poland
HGA-HGZ	Hungarian People's Republic
HHA-HHZ	Republic of Haiti
HIA-HIZ	Dominican Republic
HJA-HKZ	Republic of Colombia
HLA-HMZ	Korea
HNA-HNZ	Iraq
HOA-HPZ	Republic of Panama
HQA-HRZ	Republic of Honduras
HSA-HSZ	Thailand
HTA-HTZ	Nicaragua
HUA-HUZ	Republic of El Salvador
HVA-HVZ	Vatican City State
HWA-HYZ	France and French Community
HZA-HZZ	Saudi Arabia
IAA-IZZ	Italy
JAA-JSZ	Japan
JTA-JVZ	Mongolian People's Republic
JWA-JXZ	Norway
JYA-JYZ	Jordan
JZA-JZZ	Western New Guinea
KAA-KZZ	United States of America
LAA-LNZ	Norway
LOA-LWZ	Argentine Republic
LXA-LXZ	Luxembourg
LYA-LYZ	Lithuania
LZA-LZZ	People's Republic of Bulgaria
MAA-MZZ	United Kingdom
NAA-NZZ	United States of America
OAA-OCZ	Peru
ODA-ODZ	Lebanon
OEA-OEZ	Austria
OFA-OJZ	Finland
OKA-OMZ	Czechoslovakia
ONA-OTZ	Belgium
OUA-OZZ	Denmark
PAA-PIZ	Netherlands
PJA-PJZ	Netherlands Antilles
PKA-POZ	Republic of Indonesia
PPA-PYZ	Brazil
PZA-PZZ	Surinam
QAA-QZZ	(Service abbreviations)
RAA-RZZ	Union of Soviet Socialist Rep.
SAA-SMZ	Sweden
SNA-SRZ	People's Republic of Poland
SSA-SSM	United Arab Republic
SUA-SUZ	Arab Republic of Egypt
SVA-SZZ	Greece
TAA-TCZ	Turkey
TDA-TDZ	Guatemala
TEA-TEZ	Costa Rica
TFA-TFZ	Iceland
TGA-TGZ	Guatemala
THA-THZ	France and French Community
TIA-TIZ	Costa Rica
TJA-TJZ	Republic of Cameroon
TKA-TKZ	France and French Community
TLA-TLZ	Central African Republic
TMA-TMZ	France and French Community
TNA-TNZ	Republic of Congo (Brazzaville)
TOA-TQZ	France, French Community
TRA-TRZ	Republic of Gabon
TSA-TSZ	Tunisia
TTA TTZ	Republic of Chad
TUA-TUZ	Republic of the Ivory Coast
TVA-TXZ	France and French Community
TYA-TYZ	Republic of Dahomey
TZA-TZZ	Republic of Mali
UAA-UQZ	Union of Soviet Socialist Reps.
URA-UTZ	Ukrainian Soviet Socialist Reps.
UUA-UZZ	Union of Soviet Socialist Reps.
VAA-VGZ	Canada
VHA-VNZ	Commonwealth of Australia
VOA-VOZ	Canada
VPA-VSZ	British Overseas Territories
VTA-VWZ	India
VXA-VYZ	Canada
VZA-VZZ	Commonwealth of Australia
WAA-WZZ	United States of America
XAA-XIZ	Mexico
XJA-XOZ	Canada
XPA-XPZ	Denmark
XQA-XRZ	Chile
XSA-XSZ	China
XTA-XTZ	Republic of the Upper Volta
XUA-XUZ	Khmer Republic
XVA-XVZ	Vietnam
XWA-XWZ	Laos
XXA-XXZ	Portuguese Overseas Provinces
XYA-XZZ	Burma

QSY Shall I change to transmission on another frequency? Change to transmission on another frequency (or on . . . kHz).
QSZ Shall I send each word or group more than once? Send each word or group twice (or . . . times).
QTA Shall I cancel message number . . .? Cancel message number . . . .
QTB Do you agree with my counting of words? I do not agree with your counting of words? I will repeat the first letter or digit of each word or group.
QTC How many messages have you to send? I have . . . messages for you (or for . . .).
QTH What is your location? My location is . . . .
QTR What is the correct time? The time is . . . .

Special abbreviations adopted by ARRL:

QST General call preceding a message addressed to all amateurs and ARRL members. This is in effect "CQ ARRL."

The RST System

READABILITY

1- Unreadable.
2- Barely readable, occasional words distinguishable.
3- Readable with considerable difficulty.
4- Readable with practically no difficulty.
5- Perfectly readable.

SIGNAL STRENGTH

1- Faint signals barely perceptible.
2- Very weak signals.
3- Weak signals.
4- Fair signals.
5- Fairly good signals.
6- Good signals.
7- Moderately strong signals.
8- Strong signals.
9- Extremely strong signals.

TONE

1- Sixty cycle ac or less, very rough and broad.
2- Very rough ac, very harsh and broad.
3- Rough ac tone, rectified but not filtered.
4- Rough note, some trace of filtering.
5- Filtered rectified ac but strongly ripple-modulated.
6- Filtered tone, definite trace of ripple modulation.
7- Near pure tone, trace of ripple modulation.
8- Near perfect tone, slight trace of modulation.
9- Perfect tone, no trace of ripple or modulation of any kind.

The "tone" report refers only to the purity of the signal, and has no connection with its stability or freedom from clicks or chirps. If the signal has the characteristic steadiness of crystal control, add X to the report (e.g., RST 469X). If it has a chirp or "tail" (either on "make" or "break") add C (e.g., 469K). If it has clicks or noticeable other keying transients, add K (e.g., 469D). Of course a signal could have both chirps and clicks, in which case both C and K could be used (e.g., RST 469CK).

YAA-YAZ	Afghanistan
YBA-YHA	Republic of Indonesia
YIA-YIZ	Iraq
YJA-YJZ	New Hebrides
YKA-YKZ	Syria
YLA-YLZ	Latvia
YMA-YMZ	Turkey
YNA-YNZ	Nicaragua
YOA-YRZ	Roumanian People's Republic
YSA-YSZ	Republic of El Salvador
YTA-YUZ	Yugoslavia
YVA-YYZ	Venezuela
YZA-YZZ	Yugoslavia
ZAA-ZAZ	Albania
ZBA-ZJZ	British Overseas Territories
ZKA-ZMZ	New Zealand
ZNA-ZOZ	British Overseas Territories
ZPA-ZPZ	Paraguay
ZQA-ZQZ	British Overseas Territories
ZRA-ZUZ	Republic of South Africa
ZVA-ZZZ	Brazil
2AA-2ZZ	Great Britain
3AA-3AZ	Monaco
3BA-3BZ	Mauritius
3CA-3CZ	Equatorial Guinea
3DA-3DM	Swaziland
3DN-3DZ	Fiji
3DN-3FZ	Panama
3GA-3GZ	Chile
3HA-3UZ	China
3VA-3VZ	Tunisia
3WA-3WZ	Vietnam
3XA-3XZ	Guinea
3YA-3YZ	Norway
3ZA-3ZZ	People's Republic of Poland
4AA-4CZ	Mexico
4DA-4IZ	Republic of the Philippines
4JA-4LZ	Union of Soviet Socialist Rep.
4MA-4MZ	Venezuela
4NA-4OZ	Yugoslavia
4PA-4SZ	Ceylon
4UA-4UZ	United Nations
4VA-4VZ	Republic of Haiti
4WA-4WZ	Yemen
4XA-4XZ	State of Israel
4YA-4YZ	International Civil Aviation Org.
4ZA-4ZZ	State of Israel
5AA-5AZ	Libya
5BA-5BZ	Republic of Cyprus
5CA-5GZ	Morocco
5HA-5IZ	Tanzania
5JA-5KZ	Colombia
5LA-5MZ	Liberia
5NA-5OZ	Nigeria
5PA-5QZ	Denmark
5RA-5SZ	Malagasy Republic
5TA-5TZ	Islamic Republic of Mauritania
5UA-5UZ	Republic of the Niger
5VA-5VZ	Togolese Republic
5WA-5WZ	Western Samoa
5YA-5ZZ	Uganda
5XA-5XZ	Kenya
6AA-6BZ	Arab Republic of Egypt
6CA-6CZ	Syria
6DA-6JZ	Mexico
6KA-6NZ	Korea
6OA-6OZ	Somalia
6PA-6SZ	Pakistan
6TA-6UZ	Sudan
6VA-6WZ	Republic of the Senegal
6XA-6XZ	Malagasy Republic
6YA-6YZ	Jamaica
6ZA-6ZZ	Liberia
7AA-7IZ	Indonesia
7JA-7NZ	Japan
7OA-7OZ	South Yemen Popular Republic
7PA-7PZ	Lesotho
7QA-7QZ	Malawi
7RA-7RZ	Algeria
7SA-7SZ	Sweden
7TA-7YZ	Algeria
7ZA-7ZZ	Saudi Arabia
8AA-8IZ	Indonesia
8JA-8NZ	Japan
8OA-8OZ	Botswana
8PA-8PZ	Barbados
8QA-8QZ	Maldive Islands
8RA-8RZ	Guyana
8SA-8SZ	Sweden
8TA-8YZ	India
8ZA-8ZZ	Saudi Arabia
9AA-9AZ	San Marino
9BA-9DZ	Iran
9EA-9FZ	Ethiopia
9GA-9GZ	Ghana
9HA-9HZ	Malta
9IA-9JZ	Zambia
9KA-9KZ	Kuwait
9LA-9LZ	Sierra Leone
9MA-9MZ	Malaysia
9NA-9NZ	Nepal
9OA-9TZ	Republic of Zaire
9UA-9UZ	Burundi
9VA-9VZ	Singapore
9WA-9WZ	Malaysia
9XA-9XZ	Rwanda
9YA-9ZZ	Trinidad and Tobago
A2A-A2Z	Republic of Botswana
A3A-A3Z	Kingdom of Tonga
A4A-A4Z	Oman
A5A-A5Z	Bhutan
A6A-A6Z	United Arab Emirates
C2A-C2Z	Republic of Nauru
C3A-C3Z	Principality of Andorra
L2A-L9Z	Argentina
S2A-S3Z	Bangladesh

Table 19
Some Abbreviations Used in Text and Drawings

A — ampere
ac — alternating current
A/D — analog-to-digital
af — audio frequency
afc — automatic frequency control
afsk — audio frequency-shift keying
agc — automatic gain control
alc — automatic load (or level) control
a-m — amplitude modulation
anl — automatic noise limiter
ARC — amateur radio club
AREC — Amateur Radio Emergency Corps
ARPSC — Amateur Radio Public Service Corps
ATV — amateur television
avc — automatic volume control

bc — broadcast
BCD — binary-coded decimal
bci — broadcast interference
bcl — broadcast listener
BFO — beat-frequency oscillator
BPL — Brass Pounders League

CB — citizens band
CCIR — International Radio Consultative Committee
ccw — counterclockwise
c.d. — civil defense
CD — Communications Department (ARRL)
CMOS or COSMOS — complimentary-symmetry metal-oxide semiconductor
coax — coaxial cable, connector
COR — carrier-operated relay
CP — Code Proficiency (award)
CR — cathode ray
CRT — cathode-ray tube
ct — center tap
CTCSS — continuous tone-controlled squelch system
cw — continuous wave (code), clockwise

D/A — digital-to-analog
dB — decibel
dc — direct current
DF — direction finder
DOC — Department of Communications (Canadian)
dpdt — double-pole double-throw
dpst — double-pole single-throw
dsb — double sideband
DTL — diode-transistor logic
DX — long distance
DXCC — DX Century Club

EC — Emergency Coordinator
ECO — electron-coupled oscillator
ECL — emitter-coupled logic
EME — earth-moon-earth
emf — electromotive force (voltage)

FAX — facsimile
FCC — Federal Communications Commission
FD — Field Day
FET — field-effect transistor
FF — flip-flop
fm — frequency modulation
FMT — frequency measuring test
fsk — frequency-shift keying

GDO — grid-dip oscillator
GHz — gigahertz
GMT — Greenwich Mean Time
gnd — ground

H — henry
hf — high frequency
HFO — heterodyne frequency oscillator
Hz — hertz

IARU — International Amateur Radio Union
IC — integrated circuit
ID — inside diameter
i-f — intermediate frequency
in./s — inch per second
IRC — International Reply Coupon
ITU — International Telecommunication Union
IW — Intruder Watch

JFET — junction field-effect transistor

k — kilo
kc — kilocycle
kHz — kilohertz
kW — kilowatt

LED — light-emitting diode
lf — low frequency
LMO — linear master oscillator
LO — local oscillator
lsb — lower sideband
LSB — least-significant bit
LSD — least-significant digit
LSI — large-scale integration
luf — lowest usable frequency

mA — milliampere
MARS — Military Affiliate Radio System
Mc — Megacycle
mf — medium frequency
MG — motor-generator
mH — milihenry
MHz — Megahertz
mic — microphone
mix — mixer
MO — master oscillator
MOSFET — metal-oxide semiconductor field-effect transistor
MOX — manually operated switching
ms — millisecond
m.s. — meteor scatter
MSB — most-significant bit
MSD — most-significant digit
MSI — medium-scale integration
muf — maximum usable frequency
MUX — multiplex
mV — millivolt
mW — milliwatt

nbfm — narrow-band frequency modulation
n.c. — no connection
NC — normally closed
NCS — net control station
NO — normally open
npn — negative-positive-negative
NTS — National Traffic System (ARRL)

OBS — Official Bulletin Station
OD — outside diameter
OO — Official Observer
op amp — operational amplifier
OPS — Official Phone Station
ORS — Official Relay Station
osc — oscillator
OVS — Official VHF Station
oz — ounce

PA — power amplifier
pc — printed or etched circuit board
PEP — peak-envelope power
PEV — peak-envelope voltage
pF — picofarad
PIV — peak-inverse voltage
pk — peak
pk-pk — peak-to-peak
PL — private line
PLL — phase-locked loop
pm — phase modulation
pnp — positive-negative-positive
pot — potentiometer
PRV — peak-reverse voltage
PSHR — Public Service Honor Roll
PTO — permeability-tuned oscillator
PTT — push-to-talk

RACES — Radio Amateur Civil Emergency Service
RCC — Rag Chewers Club
rcvr — receiver
rf — radio frequency
rfc — radio-frequency choke
RFI — radio-frequency interference
RM — Route Manager
RM-(number) — FCC rulemaking
rms — root-mean-square
RO — Radio Officer (c.d.)
RST — readability-strength-tone
RTL — resistor-transistor logic
RTTY — radio teletype

s.a.e. — self-addressed envelope
s.a.s.e. — stamped s.a.e.
SCM — Section Communications Manager
SCR — silicon-controlled rectifier
SEC — Section Emergency Coordinator
SET — simulated emergency test
S.M. — silver mica (capacitor)
SNR — signal-to-noise ratio
spdt — single-pole double-throw
spst — single-pole single-throw
SS — Sweepstakes (contest)
ssb — single sideband
SSTV — slow-scan TV
SWL — short-wave listener
SWR — standing wave ratio
sync — synchronous, synchronizing

TCC — Transcontinental Corps
TD — transmitting distributor
TE — transequatorial (propagation)
tfc — traffic
tpi — turns per inch
T-R — transmit-receive
TTL or T^2L — transistor-transistor logic
TTY — Teletype
TV — television
TVI — television interference

UJT — unijunction transistor
usb — upper sideband
uhf — ultra-high frequency

V — volt
VCO — voltage-controlled oscillator
VCXO — voltage-controlled crystal oscillator
VFO — variable frequency oscillator
vhf — very high frequency
vlf — very low frequency
VOM — volt-ohm-milliammeter
VOX — voice-operated break-in
VR — voltage regulator
VTVM — vacuum-tube voltmeter
VXO — variable crystal oscillator

W — watt
WAC — Worked All Continents
WAS — Worked All States
wbfm — wide-band fm
wpm — words per minute
ww — wire wound
wv — working voltage
xtal — crystal
μ — micro (10^{-6})

Table 20

Abbreviated Semiconductor Symbol List

BIPOLAR TRANSISTOR SYMBOLS

C_{ibo}	– Input capacitance, open circuit (common base)
C_{ieo}	– Input capacitance, open circuit (common emitter)
C_{obo}	– Output capacitance, open circuit (common base)
C_{oeo}	– Output capacitance, open circuit (common emitter)
f_{c}	– Cutoff frequency
f_{T}	– Gain-bandwidth product (frequency at which small-signal forward current-transfer ratio common emitter, is unity, or 1)
g_{me}	– Small-signal transconductance (common emitter)
h_{FB}	– Static forward-current transfer ratio (common base)
h_{fb}	– Small-signal forward-current transfer ratio, short circuit (common base)
h_{FE}	– Static forward-current transfer ratio (common emitter)
h_{fe}	– Small-signal forward-current transfer ratio, short circuit (common emitter)
h_{IE}	– Static input resistance (common emitter)
h_{ie}	– Small-signal input impedance, short circuit (common emitter)
I_{b}	– Base current
I_{c}	– Collector current
I_{CBO}	– Collector-cutoff current, emitter open
I_{CEO}	– Collector-cutoff current, base open
I_{E}	– Emitter current
MAG	– Maximum available amplifier gain
P_{CE}	– Total dc or average power input to collector (common emitter)
P_{OE}	– Large-signal output power (common emitter)
R_{L}	– Load resistance
R_{s}	– Source resistance
V_{BB}	– Base-supply voltage
V_{BC}	– Base-to-collector voltage
V_{BE}	– Base-to-emitter voltage
V_{CB}	– Collector-to-base voltage
V_{CBO}	– Collector-to-base voltage (emitter open)
V_{CC}	– Collector-supply voltage
V_{CE}	– Collector-to-emitter voltage
V_{CEO}	– Collector-to-emitter voltage (base open)
$V_{CE\,(sat)}$	– Collector-to-emitter saturation voltage
V_{EB}	– Emitter-to-base voltage
V_{EBO}	– Emitter-to-base voltage (collector open)
V_{EE}	– Emitter-supply voltage
Y_{fe}	– Forward transconductance
Y_{ie}	– Input admittance
Y_{oe}	– Output admittance

FET SYMBOLS

A	– Voltage amplification
C_{c}	– Intrinsic channel capacitance
C_{ds}	– Drain-to-source capacitance (includes approximately 1-pF drain-to-case and interlead capacitance)
C_{gd}	– Gate-to-drain capacitance (includes 0.1-pF interlead capacitance)
C_{gs}	– Gate-to-source interlead and case capacitance
C_{iss}	– Small-signal input capacitance, short circuit
C_{rss}	– Small-signal reverse transfer capacitance, short circuit
g_{fs}	– Forward transconductance
g_{is}	– Input conductance
g_{os}	– Output conductance
I_{D}	– Dc drain current
$I_{DS\,(OFF)}$	– Drain-to-source OFF current
I_{GSS}	– Gate leakage current
r_{c}	– Effective gate series resistance
$r_{DS\,(ON)}$	– Drain-to-source ON resistance
r_{gd}	– Gate-to-drain leakage resistance
r_{gs}	– Gate-to-source leakage resistance
V_{DB}	– Drain-to-substrate voltage
V_{DS}	– Drain-to-source voltage
V_{GB}	– Dc gate-to-substrate voltage
V_{GB}	– Peak gate-to-substrate voltage
V_{GS}	– Dc gate-to-source voltage
V_{GS}	– Peak gate-to-source voltage
$V_{GS\,(OFF)}$	– Gate-to-source cutoff voltage
Y_{fs}	– Forward transadmittance $\approx g_{fs}$
Y_{os}	– Output admittance
Y_{L}	– Load admittance

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