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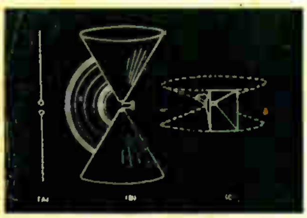
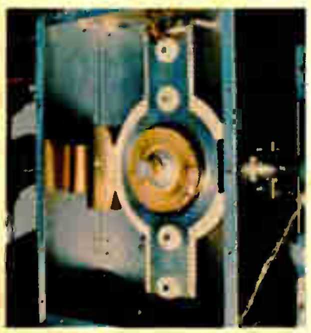
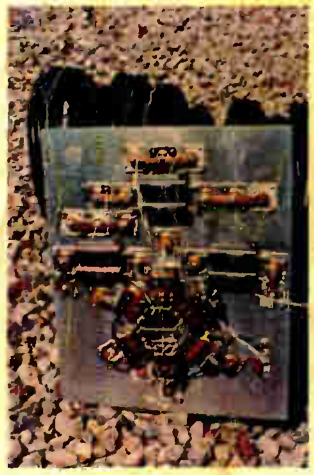
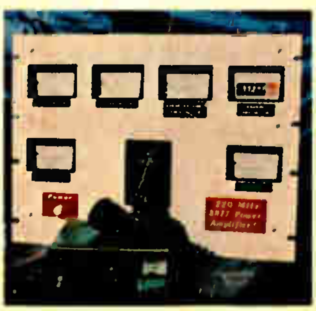
The Radio Amateur's Handbook

The Standard Manual of Amateur Radio Communication









$8.50 in the U.S.A.

1978

PUBLISHED BY THE AMERICAN RADIO RELAY LEAGUE

The Radio Amateur's Handbook

By the HEADQUARTERS STAFF

of the

AMERICAN RADIO RELAY LEAGUE

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FOREWORD

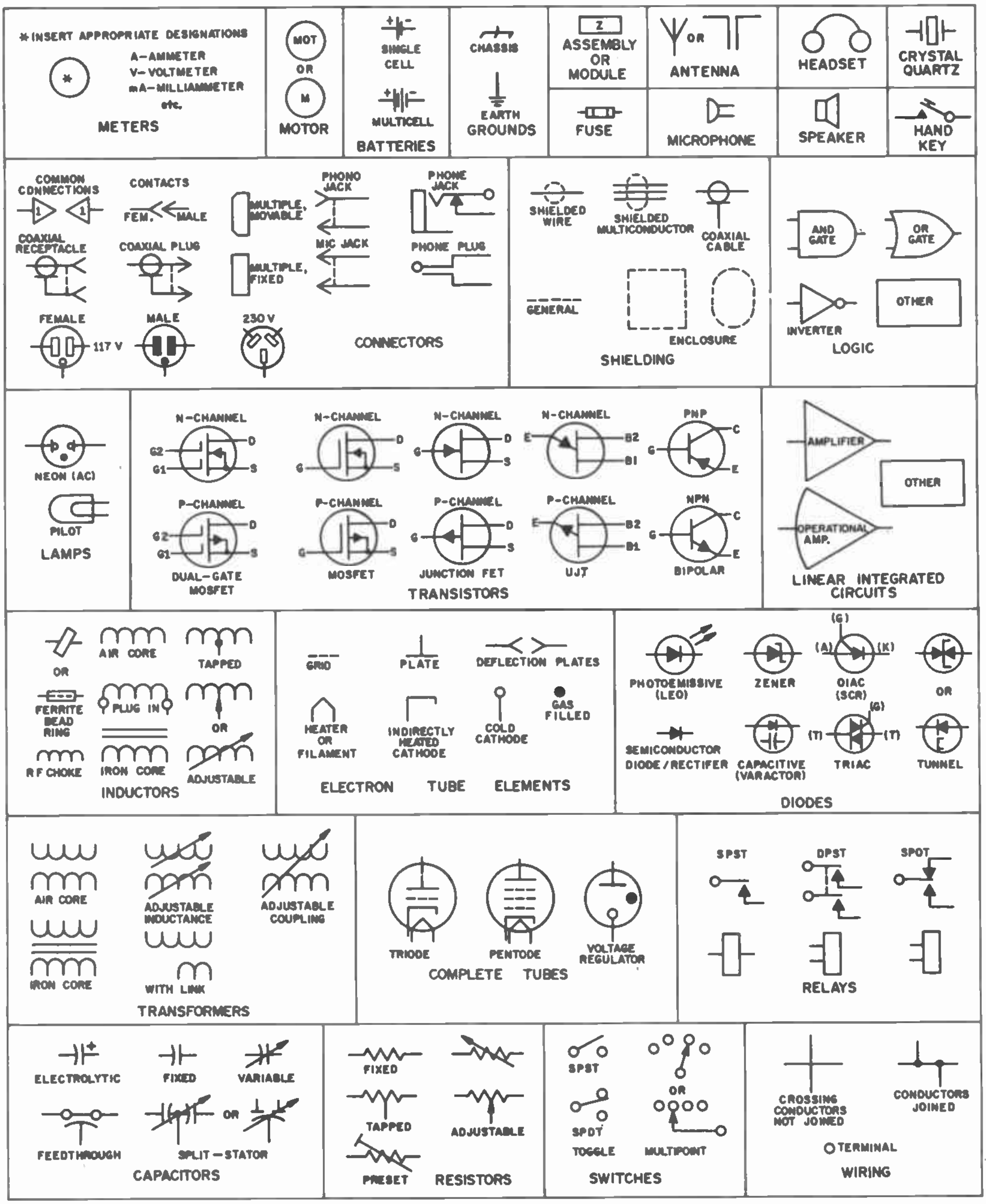
The Radio Amateur's Handbook has long been a staple of the radio amateur's library. Since it was first published in 1925, nearly five million copies have been distributed, putting it near the top of the all-time best seller list. It has achieved that distinguished record because it is a practical, useful manual. Its continuing purpose is to present the necessary fundamentals, as well as changing technology and applications, to serve the varied interests of the experimenter, the home builder, the DXer, the contester, and the ragchewer.

The present volume is the product of the efforts and the skills of many talented amateurs. We hope you will find it of value in the pursuit of your goals and your interests, particularly in these times when there seems to be a growing enthusiasm in and excitement about amateur radio.

RICHARD L. BALDWIN, W1RU
General Manager

Newington, Conn.
November, 1976

SCHEMATIC SYMBOLS USED IN CIRCUIT DIAGRAMS



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The Amateur's Code

ONE

The Amateur is considerate . . .He never knowingly uses the air in such a way as to lessen the pleasure of others.

TWO

The Amateur is Loyal . . .He offers his loyalty, encouragement and support to his fellow radio amateurs, his local club and to the American Radio Relay League, through which amateur radio is represented.

THREE

The Amateur is Progressive . . .He keeps his station abreast of science. It is well built and efficient. His operating practice is above reproach.

FOUR

The Amateur is Friendly . . .Slow and patient sending when requested, friendly advice and counsel to the beginner, kindly assistance, cooperation and consideration for the interests of others; these are marks of the amateur spirit.

FIVE

The Amateur is Balanced . . .Radio is his hobby. He never allows it to interfere with any of the duties he owes to his home, his job, his school, or his community.

SIX

The Amateur is Patriotic . . . His knowledge and his station are always ready for the service of his country and his community.

– PAUL M. SEGAL

Amateur Radio

Amateur Radio. You've heard of it. You probably know that amateur radio operators are also called "hams." (Nobody knows quite why!) But who are these people and what are they doing?

Well, every minute of every hour of every day, 365 days a year, radio amateurs all over the world are communicating with each other. It's a way of discovering new friends while experimenting with different and exciting new ways to advance the art of their hobby. Ham radio is a global fraternity of people with common and yet widely varying interests, able to exchange their ideas and learn more about each other with each new contact. Because of this, radio amateurs have the unique ability to enhance international relations as does no other hobby. How else is it possible to talk to an engineer involved in a space program, a Tokyo businessman, a U.S. legislator, a store owner in the New York Bronx, a camper in a Canadian national park, the head of state of a Mediterranean-area country, a student at a high school radio club in Wyoming or a sailor on board a ship in the middle of the Pacific? And all without leaving your home! Only with amateur radio – that's how!

The way in which communication is accomplished is just as interesting as the people you get to "meet." Signals can be sent around the world using reflective layers of the earth's ionosphere or beamed from point to point from mountaintops by relay stations. There are satellites in orbit that hams built and hams presently use to achieve communication. Still other hams bounce their signals off the moon! The possibilities are almost unlimited. Not only do radio amateurs use international Morse code and voice for communication, but they also use radioteletype, facsimile and various forms of television. Some hams even have computers hooked up to their equipment! As new techniques and modes of communication are developed, hams continue their long tradition of being among the first to use them.

What's in the future? Digital voice-encoding techniques? Three-dimensional TV? One can only guess. But if there is ever such a thing as a Star Trek transporter unit, hams will probably have them!

Once radio amateurs make sure that their gear does work, they look for things to do with the equipment and with the special skills they possess. Public service is a very large and integral part of the whole Amateur Radio Service. Hams continue this tradition by becoming involved and sponsoring various activities in their community.

Field Day, just one of many public service-type activities, is an annual event occurring every June when amateurs (using electricity generated at the site of operation) take their equipment into the great outdoors and test it for use in case of disaster. Not only do they test their equipment, but they make a contest out of the exercise and try to contact as many other hams operating emergency-type stations as possible (along with "ordinary" types). Often they make Field Day a club social event while they are operating.

Traffic nets meet on the airwaves on a schedule for the purpose of handling routine messages for people all over the country and in other countries where such third-party traffic is permitted. By doing so, amateurs stay in practice for handling messages should any real emergency or disaster occur which would require operating skill to move messages efficiently. Nets also meet because the members often have common interests: Similar jobs, interests in different languages, different hobbies (yes, some people have hobbies other than ham radio!) and a whole barrelful of other reasons. It is often a way to improve one's knowledge and to share experiences with other amateurs for the good of all involved.

DX (distance) contests are popular and awards are actively sought by many amateurs. This armchair travel is one of the more alluring activities of amateur radio. There are awards for Worked All States (WAS), Worked All Provinces (WAVE), Worked All Continents (WAC), Worked 100 Countries (DXCC) and many others.

Mobile operation (and especially on the very high frequencies) holds a special attraction to many hams. It's always fun to keep in touch with ham friends over the local repeater (devices which

HIRAM PERCY MAXIM
President ARRL, 1914–1936

A	didah	N	dahit
B	dahdididit	O	dahdahdah
C	dahdidahdit	P	didahdahdit
D	dahdidit	Q	dahdahdidah
E	dit	R	didahdit
F	dididahdit	S	dididit
G	dahdahdit	T	dah
H	didididit	U	dididah
I	didit	V	didididah
J	didahdahdah	W	didahdah
K	dahdidah	X	dahdididah
L	didahdidit	Y	dahdidahdah
M	dahdah	Z	dahdahdidit
1	didahdahdahdah	6	dahdidididit
2	dididahdahdah	7	dahdahdididit
3	didididahdah	8	dahdahdahdidit
4	dididididah	9	dahdahdahdahdit
5	dididididit	0	dahdahdahdahdah

Period: didahdidahdidah. Comma: dahdahdididahdah. Question mark: dididahdahdidit. Error: dididididididit. 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.

receive your signal and retransmit it for better coverage of the area) or finding new friends on other frequencies while driving across the country. Mobile units are often the vital link in emergency communications, too, since they are usually first on the scene of an accident or disaster.

The Oscar (Orbiting Satellite Carrying Amateur Radio) program is a new challenge to the amateur radio fraternity. Built by hams from many countries around the world, these ingenious devices hitch rides as ballast (secondary payloads) on space shots for commercial communications or weather satellites. Oscar satellites then receive signals from the ground on one frequency and convert those signals to another frequency to be sent back down to Earth. Vhf (very high frequency) and uhf (ultra-high frequency) signals normally do not have a range much greater than the horizon, but when beamed to these satellites, a vhf/uhf signal's effective range is greatly increased to make global communication a possibility. These Oscar satellites also send back telemetry signals either in Morse code or radioteleprinter (RTTY), constantly giving information on the condition of equipment aboard the satellite.

Radio amateur clubs often provide social as well as operational and technical activities. The fun provided by amateur radio is greatly enhanced when hams get together so they can "eyeball" or see each other. It's a good supplement to talking to each other over the radio. The swapping of tales (and sometimes equipment), telling of jokes and a general feeling of high spirits adds a bit of spice to club meetings along with technical matters on the agenda. Clubs offer many people their first contact with amateur radio by setting up displays in shopping centers and at events like county fairs, etc. Classes for beginners are given by many clubs on a regular basis and offer the prospective ham a chance to exchange ideas with the instructor and classmates so that learning may progress at a faster speed than by individual study.

Self-reliance has always been a trademark of the radio amateur. This is often best displayed by the strong desire by many hams to design and build their own equipment. Many others prefer to build their equipment from kits. The main point here is that hams want to know how their equipment functions, what to do with it and how to fix it if a malfunction should occur. Repair shops aren't always open during hurricanes or floods and they aren't always out in the middle of the Amazon jungle, either. Hams often come up with variations on a circuit design in common use so that they may achieve a special function, or a totally original electronic design may be brought out by a ham, all in the interest of advancing the radio art.

GETTING A LICENSE

"All of this sounds very interesting and seems to be a lot of fun, but just how do I go about getting into this hobby? Don't you almost need a degree in electronics to pass the test and get a license?"

Nothing could be farther from the truth. Although you are required to have a license to operate a station, it only takes a minimal amount of study and effort on your part to pass the basic, entry-grade exam and get on the air.

"But what about the code? Don't I have to know code to get a license?" Yes, you do. International agreements require amateur radio operators to have the ability to communicate in international Morse code. But the speed at which you are required to receive it is relatively low so you should have no difficulty passing it in the course of your exam. It's not all that hard, either. Many grade-school students have passed their tests and each month hundreds of people from 8 to 80 join the ever-growing number of amateur radio operators around the world.

Concerning the written exam, to get a license you need to know some basic electrical and radio principles and regulations governing the class of license applied for. The ARRL's basic beginner package, *Tune in the World with Ham Radio*, is available for $7 from local radio stores. It may also be purchased by mail from the address below. Quite a few radio clubs teach amateur radio theory and welcome all those interested in the hobby. Learning with a group often seems to make the study a bit easier and go more smoothly, too. You may obtain the location of the closest amateur radio club with classes simply by sending a self-addressed-stamped envelope to the American Radio Relay League, 225 Main St., Newington, CT 06111.

Don Mix, 1TS, at WNP (Wireless North Pole).

LOOKING BACK

How did amateur radio become the almost unlimited hobby it is today? The beginnings are slightly obscure, but electrical experimenters around the "turn of the century" inspired by the experiments of Marconi and others of the time, began duplicating those experiments and attempted to communicate among themselves. There were no regulatory agencies at that time and much interference was caused by these "amateur" experimenters to other stations until governments the world over stepped in and established licensing, laws and regulations to control the problems involved in this new technology. "Amateur" experimenter stations were then restricted to the "useless" wavelengths of 200 meters and below. Amateurs suddenly found that they could achieve communication over longer distances than commercial stations on the longer wavelengths. Even so, signals often had to be relayed by intermediate amateur stations to get a message to the proper destination. Because of this, the American Radio Relay League was organized to establish routes of amateur radio communication and serve the public interest through amateur radio. But the dream of eventual transcontinental and even transoceanic amateur radio contact burned hot in the minds of radio amateur experimenters.

World War I broke out and amateur radio, still in its infancy, was ordered out of existence until further notice. Many former amateur radio operators joined the armed services and served with distinction as radio operators, finding their skills to be much needed.

After the close of the "War to End All Wars," amateur radio was still banned by law; yet there were many hundreds of formerly licensed amateurs just itching to "get back on the air." The government had tasted supreme authority over the radio services and was half inclined to keep it. Hiram Percy Maxim, one of the founders of the American Radio Relay League, called the pre-war League's officers together and then contacted all the old members that could be found in an attempt to re-establish amateur radio. Maxim traveled to Washington, D.C. and after considerable effort (and untold red tape) amateur radio was opened up again on October 1, 1919.

Experiments on shorter wavelengths were then begun with encouraging results. It was found that as the wavelength dropped (i.e. frequency increased) greater distances were achieved. The commercial stations were not about to miss out on this opportunity. They moved their stations to the new shorter wavelengths while the battle raged over who had the right to transmit in this new area. Usually it turned out to be the station with the stronger signal able to blot out everyone else.

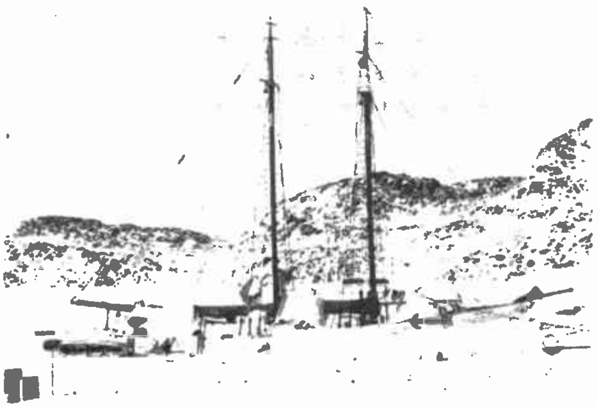
Many national and international conferences were called in the twenties to straighten out the whole mess of wavelength allocations. Through the efforts of ARRL officials, amateurs obtained frequencies on various bands similar to what we have today: One-hundred sixty through six meters. When the amateur operators moved to 20 meters, the dream of coast-to-coast and transoceanic communication without a relay station was finally realized. (A more detailed history of the early days of amateur radio is contained in the ARRL publication *Two Hundred Meters and Down* by Clinton B. DeSoto.)

PUBLIC SERVICE

Amateur radio is a grand and glorious hobby but this fact alone would hardly merit such wholehearted support as is given it by many governments at international conferences. There are other reasons. One of these is a thorough appreciation of the value of the amateur as a source of skilled radio personnel in time of war. Another asset is best described as "public service."

The "public-service" record of the amateur is a brilliant tribute to his work. These activities can be roughly divided into two classes, expeditions and emergencies. Amateur cooperation with expeditions began in 1923 when a League member, Don Mix, 1TS, accompanied MacMillan to the Arctic on the schooner *Bowdoin* with an amateur station. Amateurs in Canada and the U.S. provided the home contacts. The success of this venture was so outstanding that other explorers followed suit. During subsequent years a total of perhaps two-hundred voyages and expeditions were assisted by amateur radio, the several explorations of the Antarctic being perhaps the best known.

Since 1913 amateur radio has been the principal, and in many cases the only, means of outside communication in several hundred storm, flood and earthquake emergencies. The earthquakes which hit Alaska in 1964, Peru in 1970, California in 1971, the Guatemalan earthquake of 1976 and the 1976 Italian earthquake, the Dakota floods, the aftermath of tropical storm Agnes in 1972,



The "shack" for WNP was on the schooner *Bowdoin*.

respectively, called for the amateur's greatest emergency effort. In these disasters and many others – tornados, sleet storms, forest fires, blizzards – amateurs played a major role in the relief work and earned wide commendation for their resourcefulness in effecting communication where all other means had failed.

TECHNICAL DEVELOPMENTS

Amateurs started the hobby with spark-gap transmitters taking up great hunks of frequency space. Then they moved on to tubes when these devices came along. Much later, transistors were utilized; now integrated circuits are a part of everyday hardware in the amateur radio shack. This is because the amateur is constantly in the forefront of technical progress. His incessant curiosity and his eagerness to try anything new are two reasons. Another is that ever-growing amateur radio continually overcrowds its frequency assignments, spurring amateurs to the development and adoption of new techniques to permit the accommodation of more stations.

Amateurs have come up with ideas in their shacks while at home and then taken them to industry with surprising results. During World War II, thousands of skilled amateurs contributed their knowledge to the development of secret radio devices, both in government and private laboratories. Equally as important, the prewar technical progress by amateurs provided the keystone for the development of modern military communications equipment.

Many youngsters start out with an interest in amateur radio and then follow through with a career in a technical field. Ham engineers often come up with many of the new developments, if not actually designing a new device that is applicable not only to industry but to amateur radio use as well. This aspect of the hobby is often best demonstrated by pointing to the Oscar satellite program. Hams from around the world with technical backgrounds participate in the design, construction and launch plans for each satellite. The entire program is under the direction of AMSAT, the Amateur Radio Satellite Corporation, an affiliate of the American Radio Relay League.

THE AMERICAN RADIO RELAY LEAGUE

Ever since its establishment in 1914 by Hiram Percy Maxim and Clarence Tuska, the American Radio Relay League has been and is today not only the spokesman for amateur radio in the U.S. and Canada but the largest amateur organization in the world. It is strictly of, by and for amateurs, is noncommercial and has no stockholders. The members of the League are the owners of the ARRL and *QST*, the monthly journal of amateur radio published by the League.

The League is pledged to promote interest in two-way amateur communication and experimentation. It is interested in the relaying of messages by amateur radio. It is concerned with the advancement of the radio art. It stands for the maintenance of fraternalism and a high standard of conduct. It represents the amateur in legislative matters.

One of the League's principal purposes is to keep amateur activities so well conducted that the amateur will continue to justify his existence. Amateur radio offers its followers countless pleasures and unending satisfaction. It also calls for the shouldering of responsibilities – the maintenance of high standards, a cooperative loyalty to the traditions of amateur radio, a dedication to its ideals and principles, so that the institution of amateur radio may continue to operate "in the public interest, convenience and necessity."

In addition to publishing *QST*, the ARRL

A USAF space vehicle going into orbit with Oscar, the first amateur satellite.

maintains a model amateur radio station, W1AW, which conducts code practice and sends bulletins of interest to amateurs the world over. ARRL maintains an intruder watch so that unauthorized use of the amateur radio frequencies may be detected and appropriate action taken. At the headquarters of the League in Newington, Conn., is a well-equipped laboratory to assist staff members in preparation of technical material for *QST* and the *Radio Amateur's Handbook*. Among its other activities, the League maintains a Communications Department concerned with the operating activities of League members. A large field organization is headed by a Section Communications Manager in each of the League's seventy-four sections. There are appointments for qualified members in various fields, as outlined in Chapter 24. Special activities and contests promote operating skill. A special place is reserved each month in *QST* for amateur news from every section.

The ARRL publishes a virtual library of information on amateur radio. *Tune in the World with Ham Radio*, written for the person without previous contact with amateur radio, and the *Radio Amateur's License Manual* are two of the books designed to assist the prospective amateur to get into the hobby in the shortest possible time. *Tune in the World* comes complete with a code instruction and practice tape. They are both available from either local radio stores or from the ARRL postpaid.

Once you have studied, taken the test and have received your license, you will find that there is no other thrill quite the same as amateur radio. You may decide to operate on the lowest amateur band, 160 meters (see map). Or you may prefer to operate in the gigahertz bands (billions of cycles per second) where the entire future of communications may lie. Whatever your interest, you are sure to find it in amateur radio.

CANADIAN AMATEUR BANDS

80 meters 1) 3) 4) 5)	3.500– 3.725–	3.725 MHz 4.000 MHz	A1, F1 A1, A3, F3
40 meters 1) 3) 4) 5)	7.000– 7.150–	7.150 MHz 7.300 MHz	A1, F1, A1, A3, F3
20 meters 1) 3) 4) 5)	14.000– 14.100–	14.100 MHz 14.350 MHz	A1, F1 A1, A3, F3
15 meters 1) 3) 4) 5)	21.000– 21.100–	21.100 MHz 21.450 MHz	A1, F1 A1, A3, F3
10 meters 2) 3) 4) 5)	28.000– 28.100–	28.100 MHz 29.700 MHz	A1, F1 A1, A3, F3
6 meters 3) 4)	50.000– 50.050–	50.050 MHz 51.000 MHz	A1 A1, A2, A3, F1 F2, F3
	51.000–	54.000 MHz	AØ, A1, A2, A3 A4, F1, F2, F3, F4
2 meters 3) 4)	144.000– 144.100–	144.100 MHz 148.000 MHz	A1 AØ, A1, A2, A3 A4, F1, F2, F3 F4
3) 4)	220.000–	225.000 MHz	AØ, A1, A2, A3, A4, F1, F2, F3, F4
4) 6)	420.000– 1215.000– 2300.000– 3300.000– 5650.000– 10000.000– 24000.000-	450.000 MHz) 1300.000 MHz) 2450.000 MHz) 3500.000 MHz) 5925.000 MHz) 10500.000 MHz) 24250.000 MHz)	AØ, A1, A2, A3 A4, A5, F1, F2 F3, F4

1) Phone privileges are restricted to holders of Advanced Amateur Radio Operator Certificates, and of Commercial Certificates.

2) Phone privileges are restricted as in footnote 1), and to holders of Amateur Radio Operators Certificates, whose certificates have been endorsed for operation on phone in these bands.

3) Amplitude modulation (A2, A3, A4) shall not exceed ± 3 kHz (6A3).

4) Frequency modulation (F2, F3, F4) shall not produce a carrier deviation exceeding ±3 kHz, (6F3) except that in the 52–54 MHz and 146–148 MHz bands and higher the carrier deviation shall not exceed ±15 kHz (30F3).

5) Slow Scan television (A5), permitted by special authorization, shall not exceed a bandwidth greater than that occupied by a normal single sideband voice transmission.

6) Television (A5), permitted by special authorization, shall employ a system of standard interlace and scanning with a bandwidth of not more than 4 MHz.

Operation in frequency band 1.800–2.000 MHz shall be limited to the areas as indicated in the following table and shall be limited to the indicated maximum dc power input to the anode of the final radio frequency stage of the transmitter during day and night hours respectively; for the purpose of this table "day" means the hours between sunrise and sunset, and "night" means the hours between sunset and sunrise. A1, A3 and F3 emission are permitted.

	A	*B*	*C*	*D*	*E*	*F*	*G*	*H*
British Columbia	3 1)	3	3	1	0	0	0	0
Alberta	3 1)	3	3	3	1	0	0	1
Saskatchewan	3 1)	3	3	3	3	1	1	3
Manitoba	3 1)	2	2	2	2	2	2	3 1)
Ontario North of 50° N. Lat.	3	1	1	1	1	0	0	2
Ontario South of 50° N. Lat.	3 1)	2	1	0	0	0	0	1
Province of Quebec North of 52° N. Lat.	1	0	0	1	1	0	0	2
Province of Quebec South of 52° N. Lat.	3	2	1	0	0	0	0	0
New Brunswick	3	2	1	0	0	0	0	0
Nova Scotia	3	2	1	0	0	0	0	0
Prince Edward Island	3	2	1	0	0	0	0	0
Newfoundland (Island)	3	1	1	0	0	0	0	0
Newfoundland (Labrador)	2	0	0	0	0	0	0	0
Yukon Territory	3 1)	3	3	1	0	0	0	0
District of MacKenzie	3 1)	3	3	3	1	0	0	1
District of Keewatin	3	1	1	3	2	0	0	2
District of Franklin	0	0	0	0	1	0	0	1

1) The power levels 500 day – 100 night may be increased to 1000 day – 200 night when authorized by a Radio Inspector of the Department of Communications.

Frequency Band

A 1.800–1.825 MHz	E 1.900–1.925 MHz
B 1.825–1.850 MHz	F 1.925–1.950 MHz
C 1.850–1.875 MHz	G 1.950–1.975 MHz
D 1.875–1.900 MHz	H 1.975–2.000 MHz

Power Level – Watts

0 – Operation not permitted	
1 – 25 night	125 day
2 – 50 night	250 day
3– 100 night	500 day

U.S. AND POSSESSIONS AMATEUR BANDS

	kHz	EMISSIONS
160 m.	1800–2000	A1,A3
80 m.	3500–4000	A1
	3500–3775	F1
	3775–3890	A5[1], F5[1]
	3775–4000	A3, F3[2]
40 m.	7000–7300	A1
	7000–7150	F1
	7150–7225	A5[1], F5[1]
	7150–7300	A3, F3[2]
20 m.	14000–14350	A1
	14000–14200	F1
	14200–14275	A5[1], F5[1]
	14200–14350	A3, F3[2]
	MHz	**EMISSIONS**
15 m.	21.00–21.45	A1
	21.00–21.25	F1
	21.25–21.35	A5[1], F5[1]
	21.25–21.45	A3, F3[2]
10 m.	28.0–29.7	A1
	28.0–28.5	F1
	28.5–29.7	A3, A5[1], F3[2], F5[1]
	29.0–29.7	F3

	MHz	EMISSIONS
6 m.	50.0–54.0	A1
	50.1–54.0	A2, A3, A4, A5[3], F1, F2, F3[2], F5[1]
	51.0–54.0	AØ
	52.5–54.0	F3
2 m.	144-148	A1
	144.1–148	AØ, A2, A3, A4, A5[3], FØ, F1, F2, F3, F5[1]
	220–225	AØ, A1, A2, A3, A4, A5[3], FØ, F1, F2, F3, F4, F5[1]
	420–450[4] 1,215–1,300	AØ, A1, A2, A3, A4, A5[3], FØ, F1, F2, F3, F4, F5
	2,300–2,450 3,300–3,500 5,650–5,925 10,000–10,500[5] 24,000–24,250	AØ, A1, A2, A3, A4, A5 FØ, F1, F2, F3, F4, F5, pulse

The bands 220 through 10,500 MHz are shared with the Government Radio Positioning Service, which has priority.

NOTE: Frequencies from 3.9-4.0 MHz and 7.1-7.3 MHz are not available to amateurs on Baker, Canton, Enderbury, Guam, Howland, Jarvis, Palmyra, American Samoa, and Wake Islands.

When operating from points outside ITU Region 2 (roughly, the Western Hemisphere extended to include Hawaii), licensees of General Class and higher may operate A3 and F3 from 7075-7100 kHz; Novice licensees may operate A1 from 7050-7075 kHz.

REPEATERS:

The frequency ranges (in MHz) available for repeater inputs and outputs are as follows:

29.5 – 29.7
52.0 – 54.0
144.5 – 145.5
146.0 – 148.0
220.0 – 225.0
420.0 – 435.0
438.0 – 450.0

any amateur frequency above 1215 MHz.

[1] Slow-scan television no wider than a single-sideband voice signal may be used; on A5 if voice is simultaneously used, the total signal can be no wider than a standard a-m signal.

[2] Narrow-band frequency- or phase-modulation no wider than standard a-m voice signal.

[3] Slow-scan television no wider than a standard a-m voice signal.

[4] Input power must not exceed 50 watts in Fla., Ariz., and parts of Ga., Ala., Miss., N. Mex., Tex., Nev., and Ca. See the *License Manual* or write ARRL for further details.

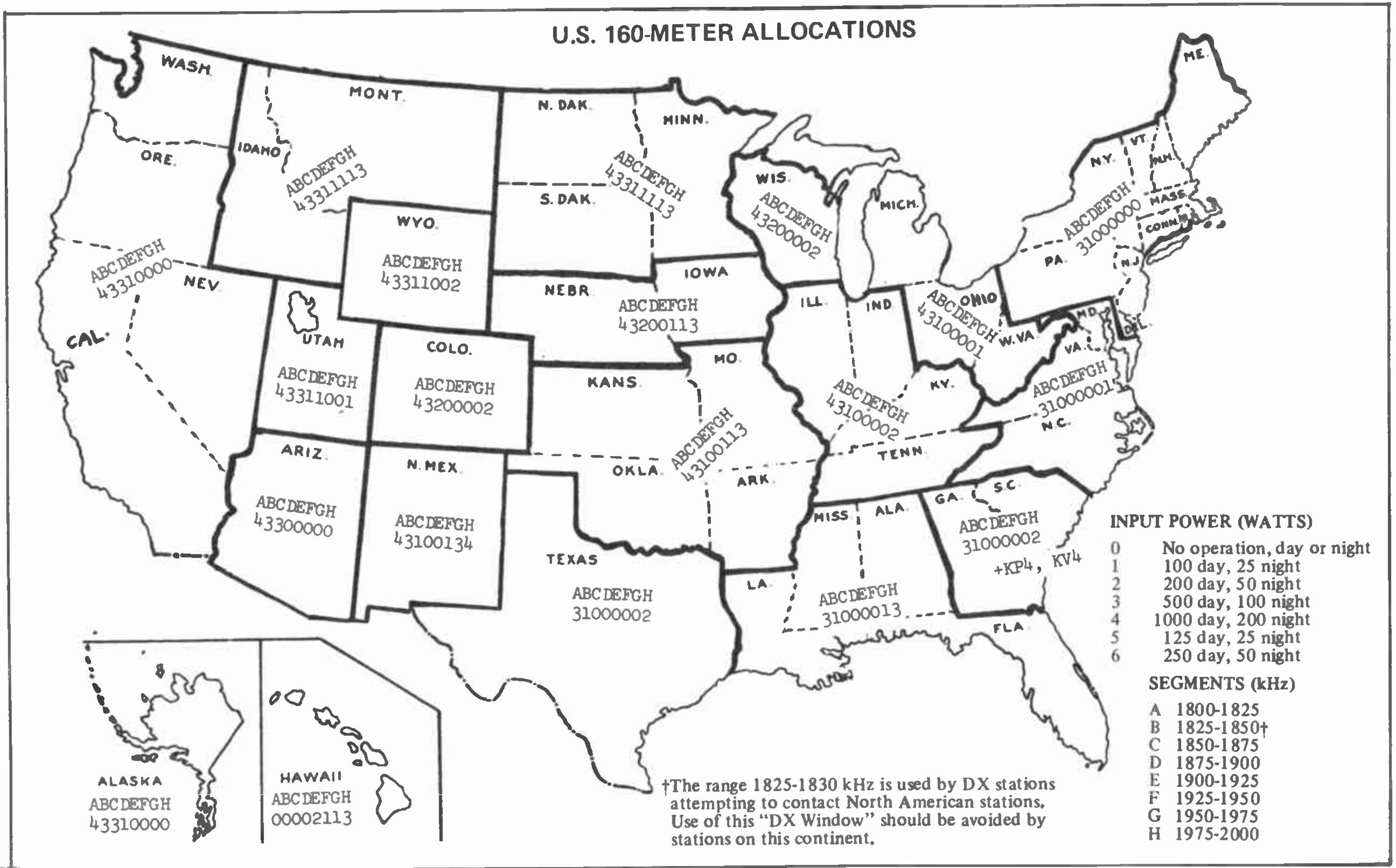
[5] No pulse permitted in this band.

Novice licensees may use A1 emission and a maximum power input of 250 watts on the following frequencies:

3.700–3.750 MHz	21.100–21.200 MHz
7.100–7.150 MHz	28.100–28.200 MHz

Technician licensees are permitted all amateur privileges in 50.1-54 MHz, 144.5-148 MHz and in the bands 220 MHz and above, and in addition have full Novice privileges.

Except as otherwise specified, the maximum amateur power input is 1000 watts.


U.S. 160-METER ALLOCATIONS
WASH
MONT
N. DAK
MINN
IDAHO
ORE
ABCDEFGH
43311113
ABCDEFGH
43311113
WIS
ABCDEFGH
43200002
MICH
ME
VT
N.H.
N.Y.
MASS
CONN
R.I.
ABCDEFGH
31000000
PA
N.J.
WYO
ABCDEFGH
43311002
S. DAK
ABCDEFGH
43310000
NEV
CAL
IOWA
NEBR
ABCDEFGH
43200113
ILL
IND
OHIO
ABCDEFGH
43100001
W.VA
MD
DEL
VA
UTAH
ABCDEFGH
43311001
COLO
ABCDEFGH
43200002
KANS
MO
KY
ABCDEFGH
43100002
ABCDEFGH
31000001
N.C.
ABCDEFGH
43100113
TENN
ARIZ
ABCDEFGH
43300000
N. MEX
ABCDEFGH
43100134
OKLA
ARK
GA
S.C.
ABCDEFGH
31000002
+KP4, KV4
MISS
ALA
TEXAS
ABCDEFGH
31000002
LA
ABCDEFGH
31000013
FLA
ALASKA
ABCDEFGH
43310000
HAWAII
ABCDEFGH
00002113
INPUT POWER (WATTS)
0 No operation, day or night
1 100 day, 25 night
2 200 day, 50 night
3 500 day, 100 night
4 1000 day, 200 night
5 125 day, 25 night
6 250 day, 50 night
SEGMENTS (kHz)
A 1800-1825
B 1825-1850†
C 1850-1875
D 1875-1900
E 1900-1925
F 1925-1950
G 1950-1975
H 1975-2000
†The range 1825-1830 kHz is used by DX stations attempting to contact North American stations. Use of this "DX Window" should be avoided by stations on this continent.

Electrical Laws and Circuits

FUNDAMENTAL PRINCIPLES

Some of the manifestations of electricity and magnetism are familiar to everyone. The effects of static electricity on a dry, wintry day, an attraction by the magnetic north pole to a compass needle, and the propagation and reception of radio waves are just a few examples. Less easily recognized as being electrical in nature perhaps, the radiation of light and even radiant heat from a stove are governed by the same physical laws that describe a signal from a TV station or an amateur transmitter. The ability to transmit electrical energy through space without any reliance on matter that might be in that space (such as in a vacuum) or the creation of a disturbance in space that can produce a force are topics that are classified under the study of **electromagnetic fields.** Knowledge of the properties and definitions of fields is important in understanding such devices as transmission lines, antennas, and circuit-construction practices such as shielding.

Once a field problem is solved, it is often possible to use the results over and over again for other purposes. The field solution can be used to derive numerical formulas for such entities as **resistance, inductance,** and **capacitance** or the latter quantities can be determined experimentally. These elements, then, form the building blocks for more complex configurations called **networks** or **circuits.** Since there is no need to describe the physical appearance of the individual elements, a pictorial representation is often used and it is called a **schematic diagram.** However, each element must be assigned a numerical value, otherwise the schematic diagram is incomplete. If the numerical values associated with the sources of energy (such as batteries or generators) are also known, it is then possible to determine the power transferred from one part of the circuit to another element by finding the numerical values of entities called **voltage** and **current.**

Finally, there is the consideration of the fundamental properties of the matter that makes up the various circuit elements or devices. It is believed that all matter is made up of complex structures called **atoms** which in turn are composed of more or less unchangeable particles called **electrons, protons,** and **neutrons.** Construction of an atom will determine the chemical and electrical properties of matter composed of like atoms. The periodic table of chemical elements is a classification of such atoms. Electrons play an important role in both the chemical and electrical properties of matter and elements where some of the electrons are relatively free to move about. These materials are called conductors. On the other hand, elements where all of the electrons are tightly bonded in the atomic structure are called insulators. Metals such as copper, aluminum, and silver are very good conductors while glass, plastics, and rubber are good insulators.

Although electrons play the principal role in the properties of both insulators and conductors, it is possible to construct matter with an apparent charge of opposite sign to that of the electron. Actually, the electron is still the charge carrier but it is the physical absence of an electron *location* that moves. However, it is convenient to consider that an actual charge carrier is present and it has been labeled a hole. Materials in which the motion of electrons and holes determine the electrical characteristics are called **semiconductors.** Transistors, integrated circuits, and similar solid-state devices are made up from semiconductors. While there are materials that fall in between the classifications of conductor and insulator, and might be labeled as semiconductors, the latter term is applied exclusively to materials where the motion of electrons and holes is important.

Electrostatic Field and Potentials

All electrical quantities can be expressed in the fundamental dimensions of *time, mass, force,* and *length.* In addition, the quantity or dimension of *charge* is also required. The metric system of units (SI system) is almost exclusively used now to specify such quantities, and the reader is urged to become familiar with this system. In the metric system, the basic unit of charge is the coulomb. The smallest known charge is that of the electron which is -1.6×10^{-19} coulombs. (The proton has the same numerical charge except the sign is positive.)

The concept of electrical charge is analogous to that of mass. It is the mass of an object that determines the force of gravitational attraction between the object and another one. A similar phenomenon occurs with two charged objects. If the charges can be considered to exist at points in space, the force of attraction (or repulsion if the charges have like signs) is given by the formula:

$$F = \frac{Q_1 Q_2}{4\pi\epsilon r^2}$$

where Q_1 is the numerical value of one charge, Q_2 is the other charge value, r is the distance in meters, ϵ is the *permittivity* of the medium surrounding the charges, and F is the force in

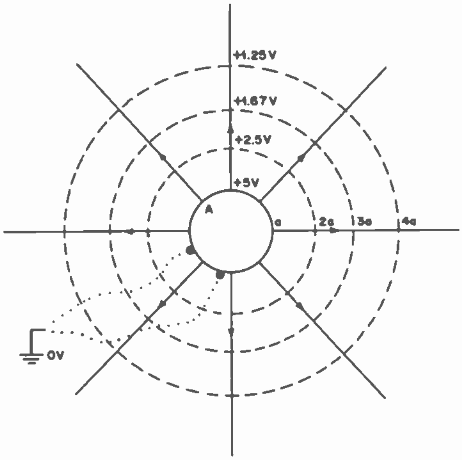


Fig. 1A — Field (solid lines) and potential (dotted lines) lines surrounding a charged sphere.

newtons. In the case of free space of a vacuum, ϵ has a value of 8.854×10^{-12} and is the permittivity of free space. The product of *relative* permittivity and ϵ_0 (the permittivity of free space) gives the permittivity for a condition where matter is present. Permittivity is also called the dielectric constant and relative dielectric constants for plastics such as polyethylene and Teflon are 2.26 and 2.1 respectively. (The relative dielectric constant is also important in transmission-line theory. The reciprocal of the square root of the dielectric constant of the material used to separate the conductors in a transmission line gives the velocity factor of the line. The effect of velocity factor will be treated in later chapters.)

If instead of just two charges, a number of charged objects are present, the force on any one member is likely to be a complicated function of the positions and magnitudes of the other charges. Consequently, the concept of electric field strength is a useful one to introduce. The field strength or field intensity is defined as the force on a given charge (concentrated at a point) divided by the numerical value of the charge. Thus, if a force of 1 newton existed on a test charge of 2 coulombs, the field intensity would be 0.5 newtons/coulomb.

Whenever a force exists on an object, it will require an expenditure of energy to move the object against that force. In some instances, the mechanical energy may be recovered (such as in a compressed spring) or it may be converted to another form of energy (such as heat produced by friction). As is the case for electric-field intensity, it is convenient to express energy ÷ charge as the potential or voltage of a charged object at a point. For instance, if it took the expenditure of 5 newton-meters (5 joules) to move a charge of 2 coulombs from a point of zero energy to a given point, the voltage or potential at that point would be 2.5 joules/coulomb. Because of the frequency of problems of this type, the dimension of joules/coulumb is given a special designation and one joule/coulomb is defined as one volt. Notice that if the voltage is divided by length (meters), the dimensions of field intensity will be obtained and a field strength of one newton/coulomb is also defined as one volt/meter. The relationship between field intensity and potential is illustrated by the following example shown in Fig. 1A.

A conducting sphere receives a charge until its surface is at a potential of 5 volts. As charges are placed on the surface of a conductor, they tend to spread out into a uniform distribution. Consequently, it will require the same amount of energy to bring a given amount of charge from a point of zero reference to *any* point on the sphere. The outside of the sphere is then said to constitute an equipotential surface. Also, the amount of energy expended will be independent of the path traveled to get to the surface. For instance, it will require 5 joules of energy to bring a charge of 1 coulomb from a point of 0 voltage to any point on the sphere (as indicated by the dotted lines in Fig. 1A). The *direction* of the force on a charged particle *at the surface* of the surface of the sphere must be perpendicular to the surface. This is because charges are able to flow about freely *on* the conductor but not *off* it. A force with a direction other than a right angle to the surface will have a component that is parallel to the conductor and will cause the charges to move about. Eventually, an equilibrium condition will be reached and any initial field component parallel to the surface will be zero. This motion of charge under the influence of an electric field is a very important concept in electricity. The *rate* at which charge flows past a reference point is defined as the current. A rate of 1 coulomb per second is defined as 1 ampere.

Because of the symmetry involved, the direction of the electric force and electric field can be represented by the solid straight lines in Fig. 1A. The arrows indicate the direction of the force on a positive charge. At points away from the sphere, less energy will be required to bring up a test charge from zero reference. Consequently, a series of concentric spherical shells indicated by the dashed lines will define the equipotential surfaces around the sphere. From mathematical consider-

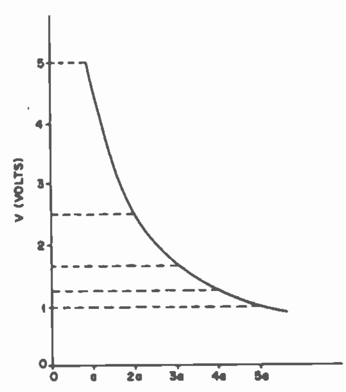


Fig. 1B — Variation of potential with distance for the charged sphere of Fig. 1A.

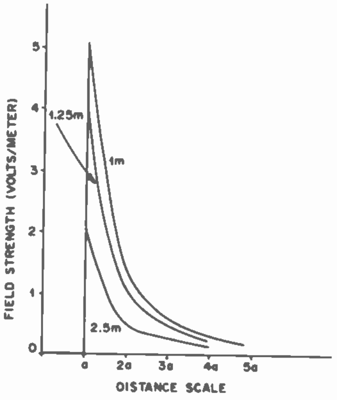


Fig. 1C – Variation of field strength with distance around a sphere charged to 5 volts for spheres of different radii.

ations (which will not be discussed here), it can be shown that the potential will vary as the inverse of the distance from the center of the sphere. This relationship is indicated by the numbers in Fig. 1A and by the graph in Fig. 1B.

While the electric field gives the direction and magnitude of a force on a charged object, it is also equal to the negative *slope* numerical value of the curve in Fig. 1B. The slope of a curve is the rate of change of some variable with distance and in this case, the variable is the potential. This is why the electric field is sometimes called the potential gradient (gradient being equivalent to slope). In the case of a curve that varies as the inverse of the distance, the slope at any point is proportional to the inverse of the distance squared.

An examination of Fig. 1 would indicate that the potential variation is only dependent upon the shape of the conductor and not its actual physical size. That is, once the value of the radius a of the sphere in Fig. 1 is specified, the potential at any other point a given distance from the sphere is also known. Thus, Fig. 1B can be used for any number of spheres with different radii. When it is changed by a certain percentage, all the other values would change by the same percentage too. However, the amount of charge required to produce a given voltage, or voltage change, does depend upon the size of the conductor, its shape, and its position in relation to other conductors and insulators. For a given conductor configuration, the voltage is related to the required charge by the formula:

$$V = \frac{Q}{C}$$

where the entity C is defined as the capacitance. Capacitance will be discussed in more detail in a later section.

Since the electric-field intensity is related to the change in potential with distance, like potential, the manner in which it changes will be uneffected by the absolute physical size of the conductor configuration. However, the exact numerical value at any point *does* depend on the dimensions of the configuration. This is illustrated in Fig. 1C for spheres with different radii. Note that for larger radii, the numerical value of the field strength at the surface of the sphere (distance equal to a) is less than it is for smaller radii. This effect is important in the design of transmission lines and capacitors. (A capacitor is a device for storing charge. In older terminology, it was sometimes called a condenser.) Even though the *same voltage* is applied across the terminals of a transmission line or capacitor, the field strength between the conductors is going to be higher for configurations of small physical size than it is for larger ones. If the field strength becomes too high, the insulating material (including air) can "break down." On the other hand, the effect can be used to advantage in spark gaps used to protect equipment connected to an antenna which is subject to atmospheric electricity. The spark-gap conductors or electrodes are filed to sharp points. Because the needlepoints appear as conductors of very small radii, the field strength is going to be higher for the same applied potential than it would be for blunt electrodes (Fig. 1D). This means the separation can be greater and the effect of the spark gap on normal circuit operation will not be as pronounced. However, a blunt electrode such as a sphere is often used on the tip of a whip antenna in order to lower the field strength under transmitting conditions.

An examination of Fig. 1C reveals that the field strength is zero for distances less than a which includes points *inside* the sphere. The implication here is that the effect of fields and charges cannot penetrate the conducting surface and disturb conditions inside the enclosure. The conducting sphere is said to form an electrostatic shield around the contents of the enclosure. However, the converse is not true. That is, charges inside the sphere will cause or induce a field on the outside surface. This is why it is very important that enclosures designed to confine the effects of charges be connected to a point of zero potential. Such a point is often called a ground.

Fields and Currents

In the last section, the motion of charged particles in the presence of an electric field was mentioned in connection with charges placed on a conducting sphere and the concept of current was

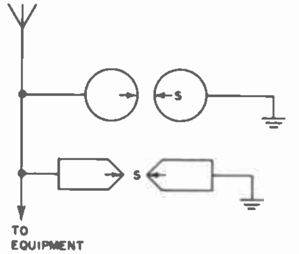


Fig. 1D – Spark gaps with sharp points break down at lower voltages than ones with blunt surfaces even though the separation s is the same.

introduced. It was assumed that charges could move around unimpeded on the surface of the sphere. In the case of actual conductors, this is not true. The charges appear to bump into atoms as they move through the conductor under the influence of the electric field. This effect depends upon the kind of material used. Silver is a conductor with the least amount of opposition to the movement of charge while carbon and certain alloys of iron are rather poor conductors of charge flow. A measure of how easily charge can flow through a conductor is defined as the conductivity and is denoted by σ.

The current density J, in a conductor is the rate of charge flow or current through a given cross-sectional area. It is related to the electric field and conductivity by the formula:

$$J = \sigma E$$

In general, the conductivity and electric field will not be constant over a large cross-sectional area, but for an important theoretical case this is assumed to be true (Fig. 2).

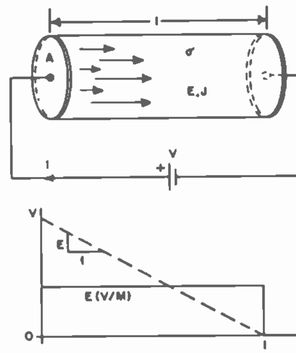


Fig. 2 — Potential and field strength along a current-carrying conductor.

A cylinder of a material with conductivity σ is inserted between two end caps of infinite conductivity. The end caps are connected to a voltage source such as a battery or generator. (A battery consists of a number of cells that convert chemical energy to electrical energy and a generator converts mechanical energy of motion to electrical energy.) The electric field is also considered to be constant along the length, l, of the cylinder and, as a consequence, the slope of the potential variation along the cylinder will also be a constant. This is indicated by the dotted lines in Fig. 2. Since the electric field is constant, the current density will also be constant. Therefore, the total current entering the end caps will just be the product of the current density and the cross-sectional area. The value of the electric field will be the quotient of the total voltage and the length of the cylinder. Combining the foregoing results and introducing two new entities gives the following set of equations:

$$J = \sigma\left(\frac{V}{l}\right) \quad \text{since } J = \sigma E \text{ and } E = \frac{V}{l}$$

$$I = J(A) = \frac{\sigma A V}{l}$$

$$\rho = \frac{1}{\sigma} \text{ and } V = I\left(\frac{\rho l}{A}\right)$$

$$R = \frac{\rho l}{A} \text{ and } V = IR$$

where ρ is the resistivity of a conducting material and R is the resistance. The final equation is a very basic one in circuit theory and is called Ohm's Law. Configurations similar to the one shown in Fig. 2 are very common ones in electrical circuits and are called resistors.

It will be shown in a later section that the power dissipated in a resistor is equal to the product of the resistance and the square of the current. Quite often resistance is an undesirable effect (such as in a wire carrying current from one location to another one) and must be reduced as much as possible. This can be accomplished by using a conductor with a low resistivity such as silver (or copper which is close to silver in resistivity but is not as expensive) with a large cross-sectional area and as short a length as possible. The current-carrying capability decreases as the diameter of a conductor size gets smaller.

Potential Drop and Electromotive Force

The application of the relations between fields, potential, and similar concepts to the physical configuration shown in Fig. 2 permitted the derivation of the formula that eliminated further consideration of the field problem. The idea of an electrical energy source was also introduced. A similar analysis involving mechanics and field theory would be required to determine the characteristics of an electrical generator and an application of chemistry would be involved in designing a chemical cell. However, it will be assumed that this problem has been solved and that the energy source can be replaced with a symbol such as that used in Fig. 2.

The term electromotive force (emf) is applied to describe a source of electrical energy, and potential drop (or voltage drop) is used for a device that consumes electrical energy. A combination of sources and resistances (or other elements) that are connected in some way is called a network or circuit. It is evident that the energy consumed in a network must be equal to the energy produced. Applying this principle to the circuit shown in Fig. 3 gives an important extension of Ohm's Law.

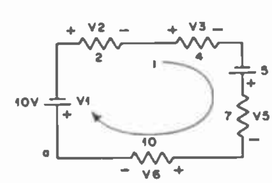


Fig. 3 — A series circuit illustrating the effects of emf and potential drop.

In Fig. 3, a number of sources and resistances are connected in tandem or *in series* to form a circuit loop. It is desired to determine the current *I*. The current can be assumed to be flowing in either a clockwise or counter-clockwise direction. If the assumption is not correct, the sign of the current will be negative when the network equations are solved and the direction can be corrected accordingly. In order to solve the problem, it is necessary to find the sum of the emfs (which is proportional to the energy produced) and to equate this sum to the sum of the potential drops (which is proportional to the energy consumed). Assuming the current is flowing in a clockwise direction, the first element encountered at point *a* is an emf, V1, but is appears to be connected "backwards." Therefore, it receives a minus sign. The next source is V4 and it appears as a voltage rise so it is considered positive. Since the current flow in all the resistors is in the same direction, all the potential drops have the same sign. The potential drop is the product of the current in amperes and the resistance in ohms. The *sums* for the emfs and potential drops and the resulting current are given by:

$$\text{sum of emf} = V_1 + V_4 = -10 + 5 = -5 \text{ volts}$$

$$\text{sum of pot. drops} = V_2 + V_3 + V_5 + V_6 =$$

$$I(2 + 4 + 7 + 10) = 23I$$

$$I = \frac{-5}{23} = -0.217 \text{ ampere}$$

Because the sign of the current is negative, it is actually flowing in a counter-clockwise direction. The physical significance of this phenomenon is that one source is being "charged." For instance, the circuit in Fig. 3 might represent a direct-current (dc) generator and a battery.

RESISTANCE

Given two conductors of the same size and shape, but of different materials, the amount of current that will flow when a given emf is applied will be found to vary with what is called the resistance of the material. The lower the resistance, the greater the current for a given value of emf.

Resistance is measured in ohms. A circuit has a resistance of one ohm when an applied emf of one volt causes a current of one ampere to flow. The resistivity of a material is the resistance, in ohms, of a cube of the material measuring one centimeter on each edge. One of the best conductors is copper, and it is frequently convenient, in making resistance calculations, to compare the resistance of the material under consideration with that of a copper conductor of the same size and shape. Table 2-I gives the ratio of the resistivity of various conductors to that of copper.

The longer the path through which the current flows the higher the resistance of that conductor.

TABLE 2-1

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

For direct current and low-frequency alternating currents (up to a few thousand cycles per second) the resistance is *inversely* proportional to the cross-sectional area of the path the current must travel; that is, given two conductors of the same material and having the same length, but differing in cross-sectional area, the one with the larger area will have the lower resistance.

Resistance of Wires

The problem of determining the resistance of a round wire of given diameter and length – or its opposite, finding a suitable size and length of wire to supply a desired amount of resistance – can be easily solved with the help of the copper-wire table given in a later chapter. This table gives the resistance, in ohms per thousand feet, of each standard wire size.

Example: Suppose a resistance of 3.5 ohms is needed and some No. 28 wire is on hand. The wire table in Chapter 18 shows that No. 28 has a resistance of 66.17 ohms per thousand feet. Since the desired resistance is 3.5 ohms, the length of wire required will be

$$\frac{3.5}{66.17} \times 1000 = 52.89 \text{ feet.}$$

Or, suppose that the resistance of the wire in the circuit must not exceed 0.05 ohm and that the length of wire required for making the connections totals 14 feet. Then

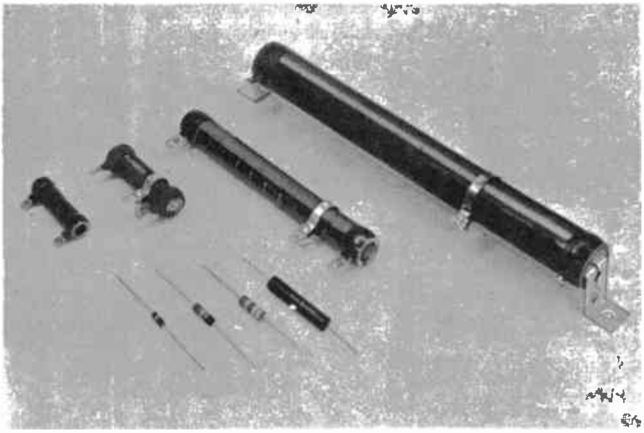
$$\frac{14}{1000} \times R = 0.05 \text{ ohm}$$

where R is the maximum allowable resistance in ohms per thousand feet. Rearranging the formula gives

$$R = \frac{0.05 \times 1000}{14} = 3.57 \text{ ohms/1000 ft.}$$

Reference to the wire table shows that No. 15 is the smallest size having a resistance less than this value.

When the wire is not copper, the resistance values given in the wire table should be multiplied by the ratios given in Table 2-1 to obtain the resistance.



Types of resistors used in radio equipment. Those in the foreground with wire leads are carbon types, ranging in size from 1/2 watt at the left to 2 watts at the right. The larger resistors use resistance wire wound on ceramic tubes; sizes shown range from 5 watts to 100 watts. Three are the adjustable type, having a sliding contact on an exposed section of the resistance winding.

Example: If the wire in the first example were nickel instead of copper the length required for 3.5 ohms would be

$$\frac{3.5}{66.17 \times 5.1} \times 1000 = 10.37 \text{ feet}$$

Temperature Effects

The resistance of a conductor changes with its temperature. Although it is seldom necessary to consider temperature in making resistance calculations for amateur work, it is well to know that the resistance of practically all metallic conductors increases with increasing temperature. Carbon, however, acts in the opposite way; its resistance *decreases* when its temperature rises. The temperature effect is important when it is necessary to maintain a constant resistance under all conditions. Special materials that have little or no change in resistance over a wide temperature range are used in that case.

Resistors

A "package" of resistance made up into a single unit is called a **resistor**. Resistors having the same resistance value may be considerably different in size and construction. The flow of current through resistance causes the conductor to become heated; the higher the resistance and the larger the current, the greater the amount of heat developed. Resistors intended for carrying large currents must be physically large so the heat can be radiated quickly to the surrounding air. If the resistor does not get rid of the heat quickly it may reach a temperature that will cause it to melt or burn.

Skin Effect

The resistance of a conductor is not the same for alternating current as it is for direct current. When the current is alternating there are internal effects that tend to force the current to flow mostly in the outer parts of the conductor. This decreases the effective cross-sectional area of the conductor, with the result that the resistance increases.

For low audio frequencies the increase in resistance is unimportant, but at radio frequencies this skin effect is so great that practically all the current flow is confined within a few thousandths of an inch of the conductor surface. The rf resistance is consequently many times the dc resistance, and increases with increasing frequency. In the rf range a conductor of thin tubing will have just as low resistance as a solid conductor of the same diameter, because material not close to the surface carries practically no current.

Conductance

The reciprocal of resistance (that is, $1/R$) is called conductance. It is usually represented by the symbol G. A circuit having large conductance has low resistance, and vice versa. In radio work the term is used chiefly in connection with vacuum-tube characteristics. The unit of conductance is the mho. A resistance of one ohm has a conductance of one mho, a resistance of 1000 ohms has a conductance of .001 mho, and so on. A unit frequently used in connection with vacuum tubes is the micromho, or one-millionth of a mho. It is the conductance of a resistance of one megohm.

OHM'S LAW

The simplest form of electric circuit is a battery with a resistance connected to its terminals, as shown by the symbols in Fig. 2-3. A complete circuit must have an unbroken path so current can flow out of the battery, through the apparatus connected to it, and back into the battery. The circuit is broken, or open, if a connection is removed at any point. A switch is a device for making and breaking connections and thereby closing or opening the circuit, either allowing current to flow or preventing it from flowing.

The values of current, voltage and resistance in a circuit are by no means independent of each

Fig. 2-3 — A simple circuit consisting of a battery and resistor.

TABLE 2-II

Conversion Factors for Fractional and Multiple Units

Change From	*To*	*Divide by*	*Multiply by*
Units	Micro-units		1,000,000
	Milli-units		1,000
	Kilo-units	1,000	
	Mega-units	1,000,000	
Micro-units	Milli-units	1,000	
	Units	1,000,000	
Milli-units	Micro-units		1,000
	Units	1,000	
Kilo-units	Units		1,000
	Mega-units	1,000	
Mega-units	Units		1,000,000
	Kilo-units		1,000

other. The relationship between them is known as Ohm's Law. It can be stated as follows: The current flowing in a circuit is directly proportional to the applied emf and inversely proportional to the resistance. Expressed as an equation, it is

$$I \text{ (amperes)} = \frac{E \text{ (volts)}}{R \text{ (ohms)}}$$

The equation above gives the value of current when the voltage and resistance are known. It may be transposed so that each of the three quantities may be found when the other two are known:

$$E = IR$$

(that is, the voltage acting is equal to the current in amperes multiplied by the resistance in ohms) and

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

(or, the resistance of the circuit is equal to the applied voltage divided by the current).

All three forms of the equation are used almost constantly in radio work. It must be remembered that the quantities are in *volts, ohms* and *amperes*; other units cannot be used in the equations without first being converted. For example, if the current is in milliamperes it must be changed to the equivalent fraction of an ampere before the value can be substituted in the equations.

Table 2-II shows how to convert between the various units in common use. The prefixes attached to the basic-unit name indicate the nature of the unit. These prefixes are;

micro – one-millionth (abbreviated μ)
milli – one-thousandth (abbreviated *m*)
kilo – one thousand (abbreviated *k*)
mega – one million (abbreviated *M*)

For example, one microvolt is one-millionth of a volt, and one megohm is 1,000,000 ohms. There are therefore 1,000,000 microvolts in one volt, and 0.000001 megohm in one ohm.

The following examples illustrate the use of ohm's law:

The current flowing in a resistance of 20,000 ohms is 150 milliamperes. What is the voltage? Since the voltage is to be found, the equation to use is $E = IR$. The current must first be converted from milliamperes to amperes, and reference to the table shows that to do so it is necessary to divide by 1000. Therefore,

$$E = \frac{150}{1000} \times 20{,}000 = 3000 \text{ volts}$$

When a voltage of 150 is applied to a circuit the current is measured at 2.5 amperes. What is the resistance of the circuit? In this case R is the unknown, so

$$R = \frac{E}{I} = \frac{150}{2.5} = 60 \text{ ohms}$$

No conversion was necessary because the voltage and current were given in volts and amperes.

How much current will flow if 250 volts is applied to a 5000-ohm resistor? Since I is unknown

$$I = \frac{E}{R} = \frac{250}{5000} = 0.05 \text{ ampere}$$

Milliampere units would be more convenient for the current, and 0.05 amp. × 1000 = 50 milliamperes.

SERIES AND PARALLEL RESISTANCES

Very few actual electric circuits are as simple as the illustration in the preceding section. Commonly, resistances are found connected in a variety of ways. The two fundamental methods of connecting resistances are shown in Fig. 2-4. In the upper drawing, the current flows from the source of emf (in the direction shown by the arrow, let us say) down through the first resistance, *R1*, then through the second, *R2*, and then back to the source. These resistors are connected in series. The current everywhere in the circuit has the same value.

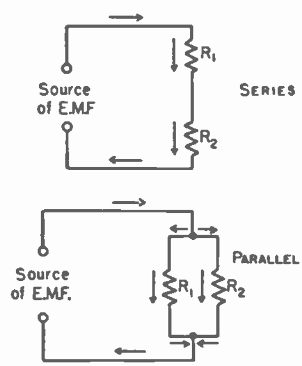


Fig. 2-4 — Resistors connected in series and in parallel.

In the lower drawing the current flows to the common connection point at the top of the two resistors and then divides, one part of it flowing through *R1* and the other through *R2*. At the lower connection point these two currents again combine; the total is the same as the current that flowed into the upper common connection. In this case the two resistors are connected in parallel.

Resistors in Series

When a circuit has a number of resistances connected in series, the total resistance of the circuit is the sum of the individual resistances. If these are numbered *R1*, *R2*, *R3*, etc., then

R (total) = R1 + R2 + R3 + R4 + ...

where the dots indicate that as many resistors as necessary may be added.

Example: Suppose that three resistors are connected to a source of emf as shown in Fig. 2-5. The emf is 250 volts. R1 is 5000 ohms, R2 is 20,000 ohms, and R3 is 8000 ohms. The total resistance is then

$$R = R1 + R2 + R3 = 5000 + 20{,}000 + 8000 = 33{,}000 \text{ ohms}$$

The current flowing in the circuit is then

$$I = \frac{E}{R} = \frac{250}{33{,}000} = 0.00757 \text{ amp.} = 7.57 \text{ mA.}$$

(We need not carry calculations beyond three significant figures, and often two will suffice because the accuracy of measurements is seldom better than a few percent.)

Voltage Drop

Ohm's Law applies to *any part* of a circuit as well as to the whole circuit. Although the current is the same in all three of the resistances in the example, the total voltage divides among them. The voltage appearing across each resistor (the voltage drop) can be found from Ohm's Law.

Example: If the voltage across R1 (Fig. 2-5) is called E1, that across R2 is called E2, and that across R3 is called E3, then

$$E1 = IR1 = 0.00757 \times 5000 = 37.9 \text{ volts}$$
$$E2 = IR2 = 0.00757 \times 20{,}000 = 151.4 \text{ volts}$$
$$E3 = IR3 = 0.00757 \times 8000 = 60.6 \text{ volts}$$

The applied voltage must equal the sum of the individual voltage drops:

$$E = E1 + E2 + E3 = 37.9 + 151.4 + 60.6 = 249.9 \text{ volts}$$

The answer would have been more nearly exact if the current had been calculated to more decimal places, but as explained above a very high order of accuracy is not necessary.

In problems such as this considerable time and trouble can be saved, when the current is small enough to be expressed in milliamperes, if the resistance is expressed in kilohms rather than ohms. When resistance in kilohms is substituted directly in Ohm's Law the current will be milliamperes if the emf is in volts.

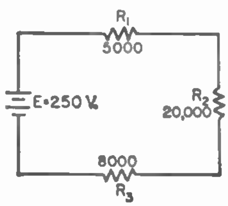


Fig. 2-5 — An example of resistors in series. The solution of the circuit is worked out in the text.

Resistors in Parallel

In a circuit with resistances in parallel, the total resistance is *less* than that of the *lowest* value of resistance present. This is because the total current is always greater than the current in any individual resistor. The formula for finding the total resistance of resistances in parallel is

$$R = \cfrac{1}{\cfrac{1}{R1} + \cfrac{1}{R2} + \cfrac{1}{R3} + \cfrac{1}{R4} + \ldots\ldots}$$

where the dots again indicate that any number of resistors can be combined by the same method. For only two resistances in parallel (a very common case) the formula becomes

$$R = \frac{R1R2}{R1 + R2}$$

Example: If a 500-ohm resistor is paralleled with one of 1200 ohms, the total resistance is

$$R = \frac{R1R2}{R1 + R2} = \frac{500 \times 1200}{500 + 1200} = \frac{600{,}000}{1700} = 353 \text{ ohms}$$

It is probably easier to solve practical problems by a different method than the "reciprocal of reciprocals" formula. Suppose the three resistors of the previous example are connected in parallel as shown in Fig. 2-6. The same emf, 250 volts, is applied to all three of the resistors. The current in each can be found from Ohm's Law as shown below, *I1*, being the current through *R1*, *I2* the current through *R2* and *I3* the current through *R3*.

For convenience, the resistance will be expressed in kilohms so the current will be in milliamperes.

$$I1 = \frac{E}{R1} = \frac{250}{5} = 50 \text{ mA}$$

$$I2 = \frac{E}{R2} = \frac{250}{20} = 12.5 \text{ mA}$$

$$I3 = \frac{E}{R3} = \frac{250}{8} = 31.25 \text{ mA}$$

The total current is

$$I = I1 + I2 + I3 = 50 + 12.5 + 31.25 = 93.75 \text{ mA}$$

The total resistance of the circuit is therefore

$$R = \frac{E}{I} = \frac{250}{93.75} = 2.66 \text{ kilohms } (= 2660 \text{ ohms})$$

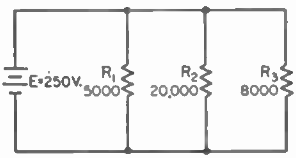


Fig. 2-6 — An example of resistors in parallel. The solution is worked out in the text.

Resistors in Series-Parallel

An actual circuit may have resistances both in parallel and in series. To illustrate, we use the same three resistances again, but now connected as in Fig. 2-7. The method of solving a circuit such as Fig. 2-7 is as follows: Consider *R2* and *R3* in parallel as through they formed a single resistor. Find their equivalent resistance. Then this resistance in series with *R1* forms a simple series circuit, as shown at the right in Fig. 2-7. An example of the arithmetic is given under the illustration.

Using the same principles, and staying within the practical limits, a value for *R2* can be computed that will provide a given voltage drop across *R3* or a given current through *R1*. Simple algebra is required.

Example: The first step is to find the equivalent resistance of R2 and R3. From the formula for two resistances in parallel,

$$R_{eq.} = \frac{R2R3}{R2 + R3} = \frac{20 \times 8}{20 + 8} = \frac{160}{28} = 5.71 \text{ kilohms}$$

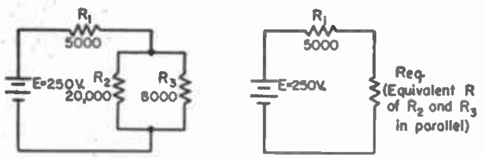


Fig. 2-7 – An example of resistors in series-parallel. The equivalent circuit is at the right. The solution is worked out in the text.

The total resistance in the circuit is then

$$R = R1 + Req. = 5 + 5.71 \text{ kilohms}$$
$$= 10.71 \text{ kilohms}$$

The current is

$$I = \frac{E}{R} = \frac{250}{10.71} = 23.3 \text{ mA}$$

The voltage drops across R1 and Req are

$$E1 = IR1 = 23.3 \times 5 = 117 \text{ volts}$$
$$E2 = IReq = 23.3 \times 5.71 = 133 \text{ volts}$$

with sufficient accuracy. These total 250 volts, thus checking the calculations so far, because the sum of the voltage drops must equal the applied voltage. Since E2 appears across both R2 and R3,

$$I2 = \frac{E2}{R2} = \frac{133}{20} = 6.65 \text{ mA}$$

$$I3 = \frac{E2}{R3} = \frac{133}{8} = 16.6 \text{ mA}$$

where I2 = Current through R2
I3 = Current through R3

The total is 23.25 mA, which checks closely enough with 23.3 mA, the current through the whole circuit.

POWER AND ENERGY

Power – the rate of doing work – is equal to voltage multiplied by current. The unit of electrical power, called the watt, is equal to one volt multiplied by one ampere. The equation for power therefore is

$$P = EI$$

where P = Power in watts
E = Emf in volts
I = Current in amperes

Common fractional and multiple units for power are the milliwatt, one one-thousandth of a watt, and the kilowatt, or one thousand watts.

Example: The plate voltage on a transmitting vacuum tube is 2000 volts and the plate current is 350 milliamperes. (The current must be changed to amperes before substitution in the formula, and so is 0.35 amp.) Then

$$P = EI = 2000 \times 0.35 = 700 \text{ watts}$$

By substituting the Ohm's Law equivalent for E and I, the following formulas are obtained for power:

$$P = \frac{E^2}{R} \qquad P = I^2R$$

These formulas are useful in power calculations when the resistance and either the current or voltage (but not both) are known.

Example: How much power will be used up in a 4000-ohm resistor if the voltage applied to it is 200 volts? From the equation

$$P = \frac{E^2}{R} = \frac{(200)^2}{4000} = \frac{40{,}000}{4000} = 10 \text{ watts}$$

Or, suppose a current of 20 milliamperes flows through a 300-ohm resistor. Then

$$P = I^2R = (0.02)^2 \times 300 = 0.0004 \times 300$$
$$= 0.12 \text{ watt}$$

Note that the current was changed from milliamperes to amperes before substitution in the formula.

Electrical power in a resistance is turned into heat. The greater the power the more rapidly the heat is generated. Resistors for radio work are made in many sizes, the smallest being rated to "dissipate" (or carry safely) about 1/8 watt. The largest resistors commonly used in amateur equipment will dissipate about 100 watts.

Generalized Definition of Resistance

Electrical power is not always turned into heat. The power used in running a motor, for example, is converted to mechanical motion. The power supplied to a radio transmitter is largely converted into radio waves. Power applied to a loudspeaker is changed into sound waves. But in every case of this kind the power is completely "used up" – it cannot be recovered. Also, for proper operation of the device the power must be supplied at a definite ratio of voltage to current. Both these features are characteristics of resistance, so it can be said that any device that dissipates power has a definite value of "resistance." This concept of resistance as something that absorbs power at a definite voltage/current ratio is very useful, since it permits substituting a simple resistance for the load or power-consuming part of the device receiving power, often with considerable simplification of calculations. Of course, every electrical device has some resistance of its own in the more narrow sense, so a part of the power supplied to it is dissipated in that resistance and hence appears as heat even though the major part of the power may be converted to another form.

Efficiency

In devices such as motors and vacuum tubes, the object is to obtain power in some other form than heat. Therefore power used in heating is considered to be a loss, because it is not the *useful* power. The efficiency of a device is the useful power output (in its converted form) divided by the power input to the device. In a vacuum-tube transmitter, for example, the object is to convert power from a dc source into ac power at some radio frequency. The ratio of the rf power output to the dc input is the efficiency of the tube. That is,

$$Eff. = \frac{P_o}{P_i}$$

where $Eff.$ = Efficiency (as a decimal)
P_o = Power output (watts)
P_i = Power input (watts)

Example: If the dc input to the tube is 100 watts, and the rf power output is 60 watts, the efficiency is

$$Eff. = \frac{P_o}{P_i} = \frac{60}{100} = 0.6$$

Efficiency is usually expressed as a percentage; that is, it tells what percent of the input power will be available as useful output. The efficiency in the above example is 60 percent.

Energy

In residences, the power company's bill is for electrical energy, not for power. What you pay for is the *work* that electricity does for you, not the *rate* at which that work is done. Electrical work is equal to power multiplied by time; the common unit is the watt-hour, which means that a power of one watt has been used for one hour. That is,

$W = PT$ where W = Energy in watt-hours
P = Power in watts
T = Time in hours

Other energy units are the kilowatt-hour and the watt-second. These units should be self-explanatory.

Energy units are seldom used in amateur practice, but it is obvious that a small amount of power used for a long time can eventually result in a "power" bill that is just as large as though a large amount of power had been used for a very short time.

CAPACITANCE

Suppose two flat metal plates are placed close to each other (but not touching) and are connected to a battery through a switch, as shown in Fig. 2-8. At the instant the switch is closed, electrons will be attracted from the upper plate to the positive terminal of the battery, and the same number will be repelled into the lower plate from the negative battery terminal. Enough electrons move into one plate and out of the other to make the emf between them the same as the emf of the battery.

If the switch is opened after the plates have been charged in this way, the top plate is left with a deficiency of electrons and the bottom plate with an excess. The plates remain charged despite the fact that the battery no longer is connected. However, if a wire is touched between the two plates (short-circuiting them) the excess electrons on the bottom plate will flow through the wire to the upper plate, thus restoring electrical neutrality. The plates have then been discharged.

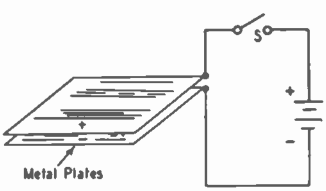


Fig. 2-8 – A simple capacitor.

The two plates constitute an electrical capacitor; a capacitor possesses the property of storing electricity. (The energy actually is stored in the electric field between the plates.) During the time the electrons are moving – that is, while the capacitor is being charged or discharged – a current is flowing in the circuit even though the circuit is "broken" by the gap between the capacitor plates. However, the current flows only during the time of charge and discharge, and this time is usually very short. There can be no continuous flow of direct current "through" a capacitor, but an alternating current can pass through easily if the frequency is high enough.

The charge or quantity of electricity that can be placed on a capacitor is proportional to the applied voltage and to the capacitance of the capacitor. The larger the plate area and the smaller the spacing between the plate the greater the capacitance. The capacitance also depends upon the kind of insulating material between the plates; it is smallest with air insulation, but substitution of other insulating materials for air may increase the capacitance many times. The ratio of the capacitance with some material other than air between the plates, to the capacitance of the same capacitor with air insulation, is called the dielectric constant of that particular insulating material. The material itself is called a dielectric. The dielectric constants of a number of materials commonly used as dielectrics in capacitors are given in Table 2-III. If a sheet of polystyrene is substituted for air between the plates of a capacitor, for example, the capacitance will be increased 2.6 times.

Units

The fundamental unit of capacitance is the farad, but this unit is much too large for practical work. Capacitance is usually measured in microfarads (abbreviated μF) or picofarads (pF). The microfarad is one-millionth of a farad, and the picofarad (formerly micromicrofarad) is one-millionth of a microfarad. Capacitors nearly always have more than two plates, the alternate plates being connected together to form two sets as shown in Fig. 2-9. This makes it possible to attain a fairly large capacitance in a small space, since several plates of smaller individual area can be

TABLE 2-III

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, fuxed	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)

Fig. 2-9 — A multiple-plate capacitor. Alternate plates are connected together.

stacked to form the equivalent of a single large plate of the same total area. Also, all plates, except the two on the ends, are exposed to plates of the other group on *both sides*, and so are twice as effective in increasing the capacitance.

The formula for calculating capacitance is:

$$C = 0.224 \frac{KA}{d} (n - 1)$$

where C = Capacitance in pF.
K = Dielectric constant of material between plates
A = Area of one side of *one* plate in square inches
d = Separation of plate surfaces in inches
n = Number of plates

If the plates in one group do not have the same area as the plates in the other, use the area of the *smaller* plates.

Capacitors in Radio

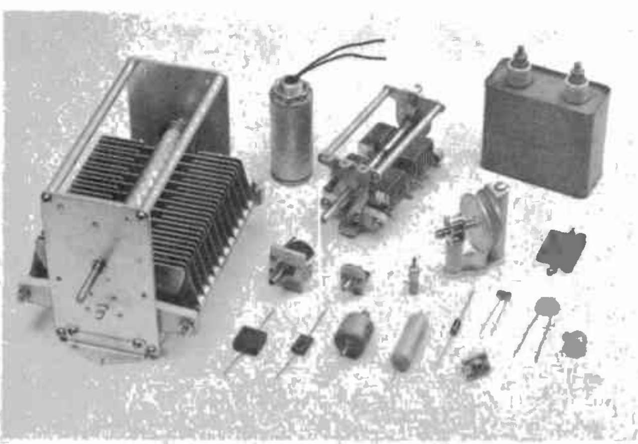
The types of capacitors used in radio work differ considerably in physical size, construction, and capacitance. Some representative types are shown in the photograph. In variable capacitors (almost always constructed with air for the dielectric) one set of plates is made movable with respect to the other set so that the capacitance can be varied. Fixed capacitors — that is, assemblies having a single, non-adjustable value of capacitance — also can be made with metal plates and with air as the dielectric, but usually are constructed from plates of metal foil with a thin solid or liquid dielectric sandwiched in between, so that a relatively large capacitance can be secured in a small unit. The solid dielectrics commonly used are mica, paper and special ceramics. An example of a liquid dielectric is mineral oil. The electrolytic capacitor uses aluminum-foil plates with a semiliquid conducting chemical compound between them; the actual dielectric is a very thin film of insulating material that forms on one set of plates through electrochemical action when a dc voltage is applied to the capacitor. The capacitance obtained with a given plate area in an electrolytic capacitor is very large, compared with capacitors having other dielectrics, because the film is so thin — much less than any thickness that is practicable with a solid dielectric.

The use of electrolytic and oil-filled capacitors is confined to power-supply filtering and audio bypass applications. Mica and ceramic capacitors are used throughout the frequency range from audio to several hundred megacycles.

Voltage Breakdown

When a high voltage is applied to the plates of a capacitor, a considerable force is exerted on the electrons and nuclei of the dielectric. Because the dielectric is an insulator the electrons do not become detached from atoms the way they do in conductors. However, if the force is great enough the dielectric will "break down"; usually it will puncture and may char (if it is solid) and permit current to flow. The breakdown voltage depends upon the kind and thickness of the dielectric, as shown in Table 2-III. It is not directly proportional to the thickness; that is, doubling the thickness does not quite double the breakdown voltage. If the dielectric is air or any other gas, breakdown is evidenced by a spark or arc between the plates, but if the voltage is removed the arc ceases and the capacitor is ready for use again. Breakdown will occur at a lower voltage between pointed or sharp-edged surfaces than between rounded and polished surfaces; consequently, the breakdown voltage between metal plates of given spacing in air can be increased by buffing the edges of the plates.

Since the dielectric must be thick to withstand high voltages, and since the thicker the dielectric the smaller the capacitance for a given plate area, a high-voltage capacitor must have more plate area than a low-voltage one of the same capacitance. High-voltage high-capacitance capacitors are physically large.



Fixed and variable capacitors. The large unit at the left is a transmitting-type variable capacitor for rf tank circuits. To its right are other air-dielectric variables of different sizes ranging from the midget "air padder" to the medium-power tank capacitor at the top center. The cased capacitors in the top row are for power-supply filters, the cylindrical-can unit being an electrolytic and the rectangular one a paper-dielectric capacitor. Various types of mica, ceramic, and paper-dielectric capacitors are in the foreground.

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. 2-10, the total capacitance of the group is equal to the sum of the individual capacitances, so

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

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,

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

and, for only two capacitors in series,

$$C\,(\text{total}) = \frac{C1C2}{C1 + C2}$$

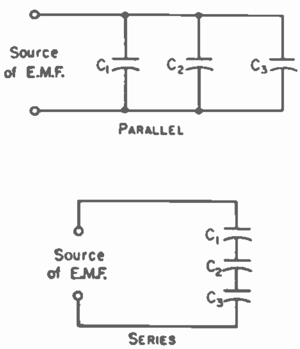


Fig. 2-10 — Capacitors in parallel and in series.

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.

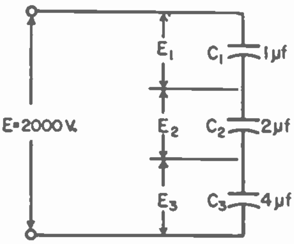


Fig. 2-11 — An example of capacitors connected in series. The solution to this arrangement is worked out in the text.

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. 2-11. The total capacitance is

$$C = \frac{1}{\frac{1}{C1} + \frac{1}{C2} + \frac{1}{C3}} = \frac{1}{\frac{1}{1} + \frac{1}{2} + \frac{1}{4}} = \frac{1}{\frac{7}{4}} = \frac{4}{7}$$

$$= 0.571\ \mu F$$

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

$$E1 = \frac{0.571}{1} \times 2000 = 1142 \text{ volts}$$

Similarly, the voltages across C2 and C3 are

$$E2 = \frac{0.571}{2} \times 2000 = 571 \text{ volts}$$

$$E3 = \frac{0.571}{4} \times 2000 = 286 \text{ volts}$$

totaling approximately 2000 volts, 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 does 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.

INDUCTANCE

It is possible to show that the flow of current through a conductor is accompanied by magnetic effects; a compass needle brought near the conductor, for example, will be deflected from its normal north-south position. The current, in other words, sets up a magnetic field.

The transfer of energy to the magnetic field represents work done by the source of emf. Power is required for doing work, and since power is equal to current multiplied by voltage, there must be a voltage drop in the circuit during the time in which energy is being stored in the field. This voltage "drop" (which has nothing to do with the voltage drop in any resistance in the circuit) is the result of an opposing voltage "induced" in the circuit while the field is building up to its final value. When the field becomes constant the induced emf or back emf disappears, since no further energy is being stored.

Since the induced emf opposes the emf of the source, it tends to prevent the current from rising rapidly when the circuit is closed. The amplitude of the induced emf is proportional to the rate at which the current is changing and to a constant associated with the circuit itself, called the inductance of the circuit.

Inductance depends on the physical characteristics of the conductor. If the conductor is formed into a coil, for example, its inductance is increased. A coil of many turns will have more inductance than one of few turns, if both coils are otherwise physically similar. Also, if a coil is placed on an iron core its inductance will be greater than it was without the magnetic core.

The polarity of an induced emf is always such as to oppose any change in the current in the circuit. This means that when the current in the circuit is increasing, work is being done against the induced emf by storing energy in the magnetic field. If the current in the circuit tends to decrease, the stored energy of the field returns to the circuit, and thus adds to the energy being supplied by the source of emf. This tends to keep the current flowing even though the applied emf may be decreasing or be removed entirely.

The unit of inductance is the henry. Values of inductance used in radio equipment vary over a wide range. Inductance of several henrys is required in power-supply circuits (see chapter on Power Supplies) and to obtain such values of inductance it is necessary to use coils of many turns wound on iron cores. In radio-frequency circuits, the inductance values used will be measured in millihenrys (a mH, one one-thousandth of a henry) at low frequencies, and in microhenrys (μH, one one-millionth of a henry) at medium frequencies and higher. Although coils for radio frequencies may be wound on special iron cores (ordinary iron is not suitable) most rf coils made and used by amateurs are of the "air-core" type; that is, wound on an insulating support consisting of nonmagnetic material.

Every conductor has inductance, even though the conductor is not formed into a coil. The inductance of a short length of straight wire is small, but it may not be negligible because if the current through it changes its intensity rapidly enough the induced voltage may be appreciable. This will be the case in even a few inches of wire when an alternating current having a frequency of the order of 100 MHz. or higher is flowing. However, at much lower frequencies the inductance of the same wire could be ignored because the induced voltage would be negligibly small.

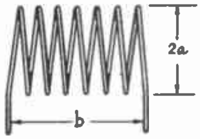


Fig. 2-12 – Coil dimensions used in the inductance formula. The wire diameter does not enter into the formula.

Calculating Inductance

The approximate inductance of single-layer air-core coils may be calculated from the simplified formula

$$L\ (\mu\text{H}) = \frac{a^2n^2}{9a + 10b}$$

where L = Inductance in microhenrys
a = Coil radius in inches
b = Coil length in inches
n = Number of turns

The notation is explained in Fig. 2-12. This formula is a close approximation for coils having a length equal to or greater than $0.8a$

Example: Assume a coil having 48 turns wound 32 turns per inch and a diameter of 3/4 inch. Thus $a = 0.75 \div 2 = 0.375$, $b = 48 \div 32 = 1.5$, and $n = 48$. Substituting,

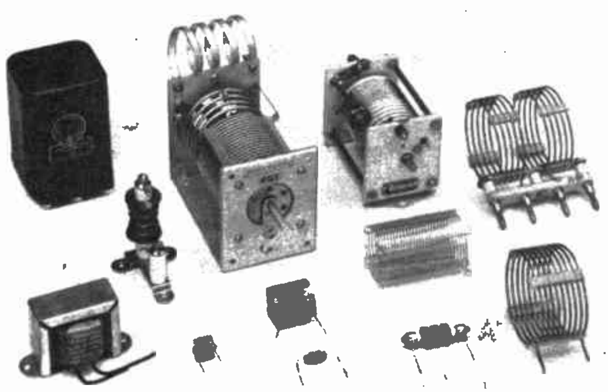
$$L = \frac{.375 \times .375 \times 48 \times 48}{(9 \times .375) + (10 \times 1.5)} = 17.6\ \mu\text{H}$$

To calculate the number of turns of a single-layer coil for a required value of inductance,

$$n = \sqrt{\frac{L\,(9a + 10b)}{a^2}}$$

Example: Suppose an inductance of 10 μH is required. The form on which the coil is to be wound has a diameter of one inch and is long enough to accommodate a coil of 1 1/4 inches. Then $a = 0.5$, $b = 1.25$, and $L = 10$. Substituting,

$$n = \sqrt{\frac{10\,(4.5 + 12.5)}{.5 \times .5}} = \sqrt{680} = 26.1 \text{ turns}$$


Inductors for power and radio frequencies. The two iron-core coils at the left are "chokes" for power-supply filters. The mounted air-core coils at the top center are adjustable inductors for transmitting tank circuits. The "pie-wound" coils at the left and in the foreground are radio-frequency choke coils. The remaining coils are typical of inductors used in rf tuned circuits, the larger sizes being used principally for transmitters.

A 26-turn coil would be close enough in practical work. Since the coil will be 1.25 inches long, the number of turns per inch will be 26.1 ÷ 1.25 = 20.8. Consulting the wire table, we find that No. 17 enameled wire (or anything smaller) can be used. The proper inductance is obtained by winding the required number of turns on the form and then adjusting the spacing between the turns to make a uniformly-spaced coil 1.25 inches long.

Inductance Charts

Most inductance formulas lose accuracy when applied to small coils (such as are used in vhf work and in low-pass filters built for reducing harmonic interference to television) because the conductor thickness is no longer negligible in comparison with the size of the coil. Fig. 2-13 shows the measured inductance of vhf coils, and may be used as a basis for circuit design. Two curves are given: curve A is for coils wound to an inside diameter of 1/2 inch; curve B is for coils of 3/4-inch inside diameter. In both curves the wire size is No. 12, winding pitch 8 turns to the inch (1/8 inch center-to-center turn spacing). The inductance values given include leads 1/2 inch long.

The charts of Figs. 2-14 and 2-15 are useful for rapid determination of the inductance of coils of the type commonly used in radio-frequency circuits in the range 3-30 MHz. They are of sufficient accuracy for most practical work. Given the coil length in inches, the curves show the multiplying factor to be applied to the inductance value given in the table below the curve for a coil of the same diameter and number of turns per inch.

Example: A coil 1 inch in diameter is 1 1/4 inches long and has 20 turns. Therefore it has 16 turns per inch, and from the table under Fig. 2-15 it is found that the reference inductance for a coil of this diameter and number of turns per inch is 16.8 μH. From curve B in the figure the multiplying factor is 0.35, so the inductance is

$$16.8 \times 0.35 = 5.9\ \mu H$$

The charts also can be used for finding suitable dimensions for a coil having a required value of inductance.

Example: A coil having an inductance of 12 μH is required. It is to be wound on a form having a diameter of 1 inch, the length available for the winding being not more than 1 1/4 inches. From Fig. 2-15, the multiplying factor for a 1-inch diameter coil (curve B) having the maximum possible length of 1 1/4 inches is 0.35. Hence the number of turns per inch must be chosen for a reference inductance of at least 12/0.35, or 34 μH. From the Table under Fig. 2-15 it is seen that 16 turns per inch (reference inductance 16.8 μH) is too small. Using 32 turns per inch, the multiplying factor is 12/68, or 0.177, and from curve B this corresponds to a coil length of 3/4 inch. There will be 24 turns in this length, since the winding "pitch" is 32 turns per inch.

Machine-wound coils with the diameters and turns per inch given in the tables are available in many radio stores, under the trade names of "B&W Miniductor" and "Illumitronic Air Dux."

IRON-CORE COILS

Permeability

Suppose that the coil in Fig. 2-16 is wound on an iron core having a cross-sectional area of 2 square inches. When a certain current is sent through the coil it is found that there are 80,000 lines of force in the core. Since the area is 2 square inches, the flux density is 40,000 lines per square inch. Now suppose that the iron core is removed and the same current is maintained in the coil, and that the flux density without the iron core is found to be 50 lines per square inch. The ratio of the flux density with the given core material to the flux density (with the same coil and same current) with an air core is called the permeability of the material. In this case the permeability of the iron is 40,000/50 = 800. The inductance of the coil is increased 800 times by inserting the iron core since, other things being equal, the inductance will be proportional to the magnetic flux through the coil.

The permeability of a magnetic material varies with the flux density. At low flux densities (or with an air core) increasing the current through the coil will cause a proportionate increase in flux, but at very high flux densities, increasing the current may cause no appreciable change in the flux. When this is so, the iron is said to be saturated. Saturation causes a rapid decrease in permeability, because it decreases the ratio of flux lines to those obtainable with the same current and an air core. Obviously, the inductance of an iron-core inductor is highly dependent upon the current flowing in the coil. In an air-core coil, the inductance is independent of current because air does not saturate.

Iron core coils such as the one sketched in Fig. 2-16 are used chiefly in power-supply equipment. They usually have direct current flowing through the winding, and the variation in inductance with current is usually undesirable. It may be overcome by keeping the flux density below the saturation point of the iron. This is done by opening the core so that there is a small "air gap," as indicated by the dashed lines. The magnetic "resistance" introduced by such a gap is so large – even though the gap is only a small fraction of an inch – compared with that of the iron that the gap, rather than the iron, controls the flux density. This reduces the inductance, but makes it practically constant regardless of the value of the current.

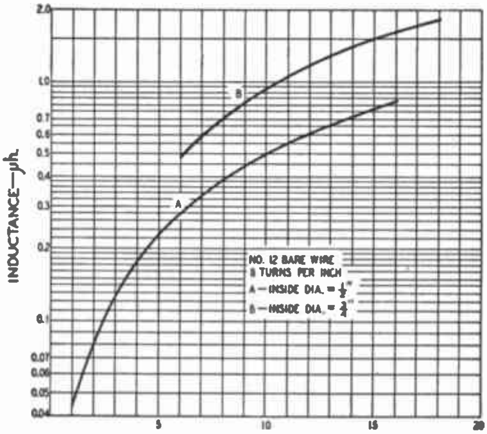


Fig. 2-13 – Measured inductance of coils wound with No. 12 bare wire, 8 turns to the inch. The values include half-inch leads.

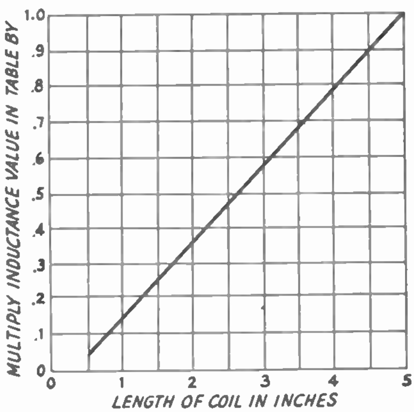


Fig. 2-14 – Factor to be applied to the inductance of coils listed in the table below, for coil lengths up to 5 inches.

Eddy Currents and Hysteresis

When alternating current flows through a coil wound on an iron core an emf will be induced, as previously explained, and since iron is a conductor a current will flow in the core. Such currents (called eddy currents) represent a waste of power because they flow through the resistance of the iron and thus cause heating. Eddy-current losses can be reduced by laminating the core; that is, by cutting it into thin strips. These strips or laminations must be insulated from each other by painting them with some insulating material such as varnish or shellac.

There is also another type of energy loss: the iron tends to resist any change in its magnetic state, so a rapidly-changing current such as ac is forced continually to supply energy to the iron to overcome this "inertia." Losses of this sort are called hysteresis losses.

Eddy-current and hysteresis losses in iron increase rapidly as the frequency of the alternating current is increased. For this reason, ordinary iron cores can be used only at power and audio frequencies – up to, say, 15,000 cycles. Even so, a very good grade of iron or steel is necessary if the core is to perform well at the higher audio frequencies. Iron cores of this type are completely useless at radio frequencies.

Coil dia, Inches	*No. of tpi*	*Inductance in μH*
1 1/4	4	2.75
	6	6.3
	8	11.2
	10	17.5
	16	42.5
1 1/2	4	3.9
	6	8.8
	8	15.6
	10	24.5
	16	63
1 3/4	4	5.2
	6	11.8
	8	21
	10	33
	16	85
2	4	6.6
	6	15
	8	26.5
	10	42
	16	108
2 1/2	4	10.2
	6	23
	8	41
	10	64
3	4	14
	6	31.5
	8	56
	10	89

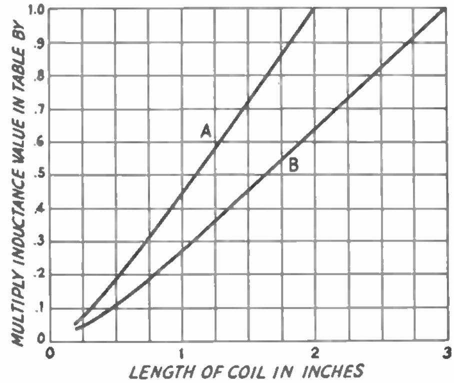


Fig. 2-15 – Factor to be applied to the inductance of coils listed in the table below, as a function of coil length. Use curve A for coils marked *A*, and curve B for coils marked *B*.

Coil dia, Inches	*No. of tpi*	*Inductance in μH*
1/2 (A)	4	0.18
	6	0.40
	8	0.72
	10	1.12
	16	2.9
	32	12
5/8 (A)	4	0.28
	6	0.62
	8	1.1
	10	1.7
	16	4.4
	32	18
3/4 (B)	4	0.6
	6	1.35
	8	2.4
	10	3.8
	16	9.9
	32	40
1 (B)	4	1.0
	6	2.3
	8	4.2
	10	6.6
	16	16.9
	32	68

For radio-frequency work, the losses in iron cores can be reduced to a satisfactory figure by grinding the iron into a powder and then mixing it with a "binder" of insulating material in such a way that the individual iron particles are insulated from each other. By this means cores can be made that will function satisfactorily even through the vhf range – that is, at frequencies up to perhaps 100 MHz. Because a large part of the magnetic path is through a nonmagnetic material, the permeability of the iron is low compared with the values

Fig. 2-16 – Typical construction of an iron-core inductor. The small air gap prevents magnetic saturation of the iron and thus maintains the inductance at high currents.

obtained at power-supply frequencies. The core is usually in the form of a "slug" or cylinder which fits inside the insulating form on which the coil is wound. Despite the fact that, with this construction, the major portion of the magnetic path for the flux is in air, the slug is quite effective in increasing the coil inductance. By pushing the slug in and out of the coil the inductance can be varied over a considerable range.

INDUCTANCES IN SERIES AND PARALLEL

When two or more inductors are connected in series (Fig. 2-17, left) the total inductance is equal to the sum of the individual inductances, *provided the coils are sufficiently separated so that no coil is in the magnetic field of another*.
That is,

$$L_{total} = L1 + L2 + L3 + L4 + \ldots\ldots\ldots\ldots$$

If inductors are connected in parallel (Fig. 2-17, right) – and the coils are separated sufficiently, the total inductance is given by

$$L_{total} = \frac{1}{\frac{1}{L1} + \frac{1}{L2} + \frac{1}{L3} + \frac{1}{L4} + \ldots\ldots}$$

and for two inductances in parallel,

$$L = \frac{L1L2}{L1 + L2}$$

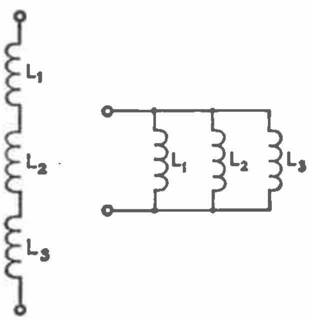


Fig. 2-17 – Inductances in series and parallel.

Thus the rules for combining inductances in series and parallel are the same for resistances, *if* the coils are far enough apart so that each is unaffected by another's magnetic field. When this is not so the formulas given above cannot be used.

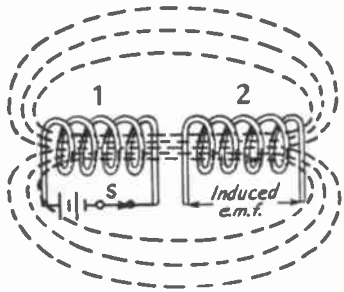


Fig. 2-18 – Mutual inductance. When the switch, *S*, is closed current flows through coil No. 1, setting up a magnetic field that induces an emf in the turns of coil No. 2.

MUTUAL INDUCTANCE

If two coils are arranged with their axes on the same line, as shown in Fig. 2-18, a current sent through Coil 1 will cause a magnetic field which "cuts" Coil 2. Consequently, an emf will be induced in Coil 2 whenever the field strength is changing. This induced emf is similar to the emf of self-induction, but since it appears in the *second* coil because of current flowing in the *first*, it is a "mutual" effect and results from the mutual inductance between the two coils.

If all the flux set up by one coil cuts all the turns of the other coil the mutual inductance has its maximum possible value. If only a small part of the flux set up by one coil cuts the turns of the other the mutual inductance is relatively small. Two coils having mutual inductance are said to be coupled.

The ratio of actual mutual inductance to the maximum possible value that could theoretically be obtained with two given coils is called the coefficient of coupling between the coils. It is frequently expressed as a percentage. Coils that have nearly the maximum possible (coefficient = 1 or 100%) mutual inductance are said to be closely, or tightly, coupled, but if the mutual inductance is relatively small the coils are said to be loosely coupled. The degree of coupling depends upon the physical spacing between the coils and how they are placed with respect to each other. Maximum coupling exists when they have a common axis and are as close together as possible (one wound over the other). The coupling is least when the coils are far apart or are placed so their axes are at right angles.

The maximum possible coefficient of coupling is closely approached only when the two coils are wound on a closed iron core. The coefficient with air-core coils may run as high as 0.6 or 0.7 if one coil is wound over the other, but will be much less if the two coils are separated.

TIME CONSTANT

Capacitance and Resistance

Connecting a source of emf to a capacitor causes the capacitor to become charged to the full emf practically instantaneously, if there is no resistance in the circuit. However, if the circuit contains resistance, as in Fig. 2-19A, the resistance limits the current flow and an appreciable length of time is required for the emf between the capacitor

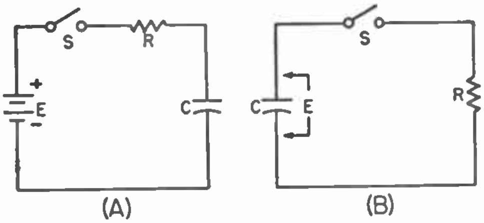


Fig. 2-19 — Illustrating the time constant of an *RC* circuit.

plates to build up to the same value as the emf of the source. During this "building-up" period the current gradually decreases from its initial value, because the increasing emf stored on the capacitor offers increasing opposition to the steady emf of the source.

Theoretically, the charging process is never really finished, but eventually the charging current drops to a value that is smaller than anything that can be measured. The time constant of such a circuit is the length of time, in seconds, required for the voltage across the capacitor to reach 63 per cent of the applied emf (this figure is chosen for mathematical reasons). The voltage across the capacitor rises with time as shown by Fig. 2-20.

The formula for time constant is

$$T = RC$$

where T = Time constant in seconds
C = Capacitance in farads
R = Resistance in ohms

Example: The time constant of a 2-μF capacitor and a 250,000-ohm (0.25 megohm) resistor is

$$T = RC = 0.25 \times 2 = 0.5 \text{ second}$$

If the applied emf is 1000 volts, the voltage between the capacitor plates will be 630 volts at the end of 1/2 second.

If C is in microfarads and R in megohms, the time constant also is in seconds. These units usually are more convenient.

If a charged capacitor is *discharged* through a resistor, as indicated in Fig. 2-19B, the same time constant applies. If there were no resistance, the capacitor would discharge instantly when S was closed. However, since R limits the current flow the capacitor voltage cannot instantly go to zero, but it will decrease just as rapidly as the capacitor can rid itself of its charge through R. When the capacitor is discharging through a resistance, the time constant (calculated in the same way as above) is the time, in seconds, that it takes for the capacitor to *lose* 63 percent of its voltage; that is, for the voltage to drop to 37 percent of its initial value.

Example: If the capacitor of the example above is charged to 1000 volts, it will discharge to 370 volts in 1/2 second through the 250,000-ohm resistor.

Inductance and Resistance

A comparable situation exists when resistance and inductance are in series. In Fig. 2-21, first consider L to have no resistance and also assume that R is zero. Then closing S would tend to send a current through the circuit. However, the instantaneous transition from no current to a finite value,

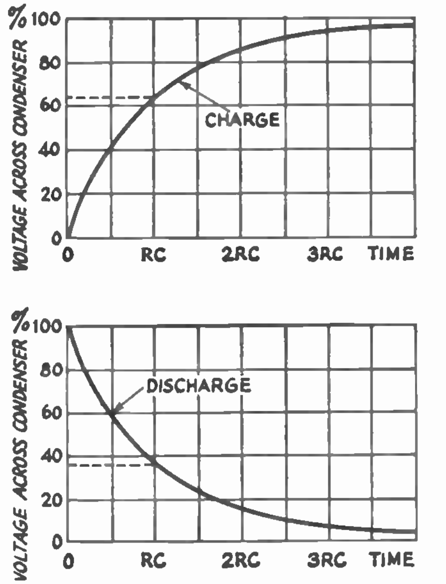


Fig. 2-20 — How the voltage across a capacitor rises, with time, when charged through a resistor. The lower curve shows the way in which the voltage decreases across the capacitor terminals on discharging through the same resistor.

however small, represents a very rapid *change* in current, and a *back emf* is developed by the self-inductance of L that is practically equal and opposite to the applied emf. The result is that the initial current is very small.

The back emf depends upon the *change* in current and would cease to offer opposition if the current did not continue to increase. With no resistance in the circuit (which would lead to an infinitely large current, by Ohm's Law) the current would increase forever, always growing just fast enough to keep the emf of self-induction equal to the applied emf.

When resistance is in series, Ohm's Law sets a limit to the value that the current can reach. The back emf generated in L has only to equal the *difference* between E and the drop across R, because that difference is the voltage actually applied to L. This difference becomes smaller as the current approaches the final Ohm's Law value. Theoretically, the back emf never quite disappears and so the current never quite reaches the Ohm's Law value, but practically the differences becomes unmeasurable after a time. The time constant of an

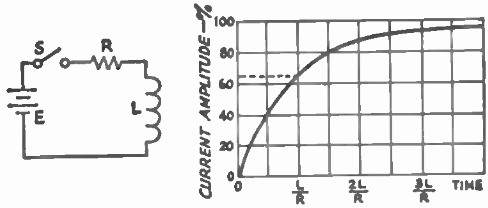


Fig. 2-21 — Time constant of an *LR* circuit.

inductive circuit is the time in seconds required for the current to reach 63 percent of its final value. The formula is

$$T = \frac{L}{R}$$

where T = Time constant in seconds
L = Inductance in Henrys
R = Resistance in ohms

The resistance of the wire in a coil acts as if it were in series with the inductance.

Example: A coil having an inductance of 20 henrys and a resistance of 100 ohms has a time constant of

$$T = \frac{L}{R} = \frac{20}{100} = 0.2 \text{ second}$$

if there is no other resistance in the circuit. If a dc emf of 10 volts is applied to such a coil, the final current, by Ohm's Law, is

$$I = \frac{E}{R} = \frac{10}{100} = 0.1 \text{ amp. or } 100 \text{ mA}$$

The current would rise from zero to 63 milliamperes in 0.2 second after closing the switch.

An inductor cannot be "discharged" in the same way as a capacitor, because the magnetic field disappears as soon as current flow ceases. Opening S does not leave the inductor "charged." The energy stored in the magnetic field instantly returns to the circuit when S is opened. The rapid disappearance of the field causes a very large voltage to be induced in the coil – ordinarily many times larger than the voltage applied, because the induced voltage is proportional to the *speed* with which the field changes. The common result of opening the switch in a circuit such as the one shown is that a spark or arc forms at the switch contacts at the instant of opening. If the inductance is large and the current in the circuit is high, a great deal of energy is released in a very short period of time. It is not at all unusual for the switch contacts to burn or melt under such circumstances. The spark or arc at the opened switch can be reduced or suppressed by connecting a suitable capacitor and resistor in series across the contacts.

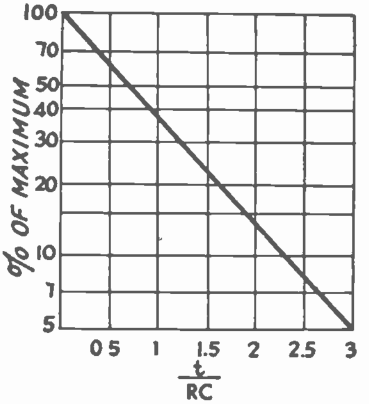


Fig. 2-22 – Voltage across capacitor terminals in a discharging *RC* circuit, in terms of the initial charged voltage. To obtain time in seconds, multiply the factor t/RC by the time constant of the circuit.

Time constants play an important part in numerous devices, such as electronic keys, timing and control circuits, and shaping of keying characteristics by vacuum tubes. The time constants of circuits are also important in such applications as automatic gain control and noise limiters. In nearly all such applications a resistance-capacitance (*RC*) time constant is involved, and it is usually necessary to know the voltage across the capacitor at some time interval larger or smaller than the actual time constant of the circuit as given by the formula above. Fig. 2-22 can be used for the solution of such problems, since the curve gives the voltage across the capacitor, in terms of percentage of the initial charge, for percentages between 5 and 100, at any time after discharge begins.

Example: A 0.01-μF capacitor is charged to 150 volts and then allowed to discharge through a 0.1-megohm resistor. How long will it take the voltage to fall to 10 volts? In percentage, 10/150 = 6.7%. From the chart, the factor corresponding to 6.7% is 2.7. The time constant of the circuit is equal to RC = 0.1 X .01 = .001. The time is therefore 2.7 X 0.001 = .0027 second, or 2.7 milliseconds.

ALTERNATING CURRENTS

PHASE

The term phase essentially means "time," or the *time interval* between the instant when one thing occurs and the instant when a second related thing takes place. The later event is said to *lag* the earlier, while the one that occurs first is said to *lead*. In ac circuits the current amplitude changes continuously, so the concept of phase or time becomes important. Phase can be measured in the ordinary time units, such as the second, but there is a more convenient method: Since each ac cycle occupies exactly the same amount of time as every other cycle of the same frequency, we can use the cycle itself as the time unit. Using the cycle as the time unit makes the specification or measurement of phase independent of the frequency of the current, so long as only one frequency is under consideration at a time. When two or more frequencies are to be considered, as in the case where harmonics are present, the phase measurements are made with respect to the lowest, or fundamental, frequency.

The time interval or "phase difference" under consideration usually will be less than one cycle. Phase difference could be measured in decimal parts of a cycle, but it is more convenient to divide the cycle into 360 parts or **degrees**. A phase degree is therefore 1/360 of a cycle. The reason for this choice is that with sine-wave alternating current the value of the current at any instant is proportional to the sine of the angle that corresponds to the number of degrees – that is, length of time – from the instant the cycle began. There is no actual "angle" associated with an alternating current. Fig. 2-23 should help make this method of measurement clear.

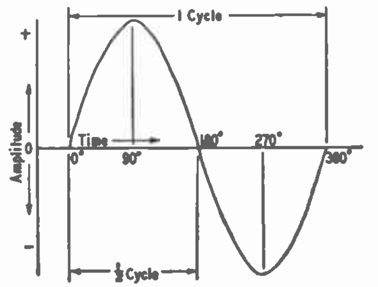


Fig. 2-23 – An ac cycle is divided off into 360 degrees that are used as a measure of time or phase.

Measuring Phase

The phase difference between two currents of the same frequency is the time or angle difference between corresponding parts of cycles of the two currents. This is shown in Fig. 2-24. The current labeled *A* leads the one marked *B* by 45 degrees, since *A*'s cycles begin 45 degrees earlier in time. It is equally correct to say that *B lags A* by 45 degrees.

Two important special cases are shown in Fig. 2-25. In the upper drawing *B* lags 90 degrees behind *A*; that is, its cycle begins just one-quarter cycle later than that of *A*. When one wave is passing through zero, the other is just at its maximum point.

In the lower drawing *A* and *B* are 180 degrees out of phase. In this case it does not matter which one is considered to lead or lag. *B* is always positive while *A* is negative, and vice versa. The two waves are thus *completely* out of phase.

The waves shown in Figs. 2-24 and 2-25 could represent current, voltage, or both. *A* and *B* might be two currents in separate circuits, or *A* might represent voltage and *B* current in the same circuit. If *A* and *B* represent two currents in the *same* circuit (or two voltages in the same circuit) the total or resultant current (or voltage) also is a sine wave, because adding any number of sine waves of the same frequency always gives a sine wave also of the same frequency.

Phase in Resistive Circuits

When an alternating voltage is applied to a resistance, the current flows exactly in step with the voltage. In other words, the voltage and current are in phase. This is true at any frequency if the resistance is "pure" – that is, is free from the reactive effects discussed in the next section. Practically, it is often difficult to obtain a purely resistive circuit at radio frequencies, because the reactive effects become more pronounced as the frequency is increased.

In a purely resistive circuit, or for purely resistive parts of circuits, Ohm's Law is just as valid for ac of any frequency as it is for dc.

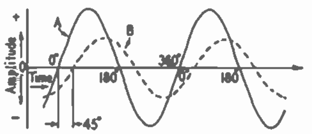


Fig. 2-24 – When two waves of the same frequency start their cycles at slightly different times, the time difference or phase difference is measured in degrees. In this drawing wave *B* starts 45 degrees (one-eighth cycle) later than wave *A*, and so lags 45 degrees behind *A*.

REACTANCE

Alternating Current in Capacitance

In Fig. 2-26 a sine-wave ac voltage having a maximum value of 100 volts is applied to a capacitor. In the period *OA*, the applied voltage increases from zero to 38 volts; at the end of this period the capacitor is charged to that voltage. In interval *AB* the voltage increases to 71 volts; that is, 33 volts additional. In this interval a *smaller* quantity of charge has been added than in *OA*, because the voltage rise during interval *AB* is smaller. Consequently the average current during

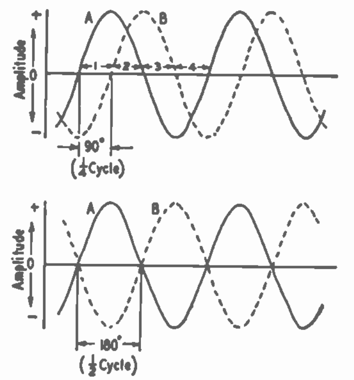


Fig. 2-25 – Two important special cases of phase difference. In the upper drawing, the phase difference between *A* and *B* is 90 degrees; in the lower drawing the phase difference is 180 degrees.

AB is smaller than during *OA*. In the third interval, *BC*, the voltage rises from 71 to 92 volts, an increase of 21 volts. This is less than the voltage increase during *AB*, so the quantity of electricity added is less; in other words, the average current during interval *BC* is still smaller. In the fourth interval, *CB*, the voltage increases only 8 volts; the charge added is smaller than in any preceding interval and therefore the current also is smaller.

By dividing the first quarter cycle into a very large number of intervals it could be shown that the current charging the capacitor has the shape of a sine wave, just as the applied voltage does. The current is largest at the beginning of the cycle and becomes zero at the maximum value of the voltage, so there is a phase difference of 90 degrees between the voltage and current. During the first quarter cycle the current is flowing in the normal direction through the circuit, since the capacitor is being charged. Hence the current is positive, as indicated by the dashed line in Fig. 2-26.

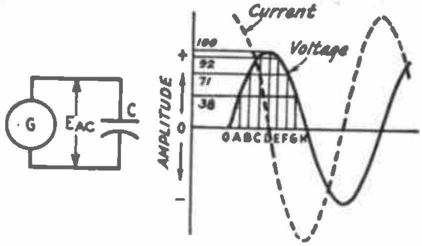


Fig. 2-26 — Voltage and current phase relationships when an alternating voltage is applied to a capacitor.

In the second quarter cycle – that is, in the time from D to H, the voltage applied to the capacitor decreases. During this time the capacitor *loses* its charge. Applying the same reasoning, it is plain that the current is small in interval DE and continues to increase during each succeeding interval. However, the current is flowing *against* the applied voltage because the capacitor is discharging into the circuit. The current flows in the *negative* direction during this quarter cycle.

The third and fourth quarter cycles repeat the events of the first and second, respectively, with this difference – the polarity of the applied voltage has reversed, and the current changes to correspond. In other words, an alternating current flows in the circuit because of the alternate charging and discharging of the capacitance. As shown by Fig. 2-26, the current starts its cycle 90 degrees before the voltage, so the current in a capacitor leads the applied voltage by 90 degrees.

Capacitive Reactance

The quantity of electric charge that can be placed on a capacitor is proportional to the applied emf and the capacitance. This amount of charge moves back and forth in the circuit once each cycle, and so the *rate* of movement of charge – that is, the current – is proportional to voltage, capacitance and frequency. If the effects of capacitance and frequency are lumped together, they form a quantity that plays a part similar to that of resistance in Ohm's Law. This quantity is called reactance, and the unit for it is the ohm, just as in the case of resistance. The formula for it is

$$X_C = \frac{1}{2\pi f C}$$

where X_C = Capacitive reactance in ohms
f = Frequency in cycles per second
C = Capacitance in farads
π = 3.14

Although the unit of reactance is the ohm, there is no power dissipation in reactance. The energy stored in the capacitor in one quarter of the cycle is simply returned to the circuit in the next.

The fundamental units (cycles per second, farads) are too large for practical use in radio circuits. However, if the capacitance is in microfarads and the frequency is in megacycles, the reactance will come out in ohms in the formula.

Example: The reactance of a capacitor of 470 pF (0.00047 μF) at a frequency of 7150 kHz (7.15 MHz) is

$$X = \frac{1}{2\pi f C} = \frac{1}{6.28 \times 7.15 \times .00047} = 47.4 \text{ ohms}$$

Inductive Reactance

When an alternating voltage is applied to a *pure* inductance (one with no resistance – all *practical* inductors have resistance) the current is again 90 degrees out of phase with the applied voltage. However, in this case the current *lags* 90 degrees behind the voltage – the opposite of the capacitor current-voltage relationship.

The primary cause for this is the *back emf* generated in the inductance, and since the amplitude of the back emf is proportional to the rate at which the current changes, and this in turn is proportional to the frequency, the amplitude of the current is inversely proportional to the applied frequency. Also, since the back emf is proportional to inductance for a given rate of current change, the current flow is inversely proportional to inductance for a given applied voltage and frequency. (Another way of saying this is that just enough current flows to generate an induced emf that equals and opposes the applied voltage.)

The combined effect of inductance and frequency is called inductive reactance, also expressed in ohms, and the formula for it is

$$X_L = 2\pi f L$$

where X_L = Inductive reactance in ohms
f = Frequency in cycles per second
L = Inductance in henrys
π = 3.14

Example: The reactance of a 15-microhenry coil at a frequency of 14 MHz is

$$X_L = 2\pi f L = 6.28 \times 14 \times 15 = 1319 \text{ ohms}$$

In radio-frequency circuits the inductance values usually are small and the frequencies are large. If the inductance is expressed in millihenrys and the frequency in kilocycles, the conversion factors for the two units cancel, and the formula for reactance may be used without first converting to fundamental units. Similarly, no conversion is necessary if the inductance is in microhenrys and the frequency is in megacycles.

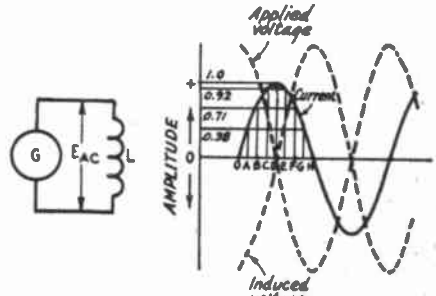


Fig. 2-27 — Phase relationships between voltage and current when an alternating voltage is applied to an inductance.

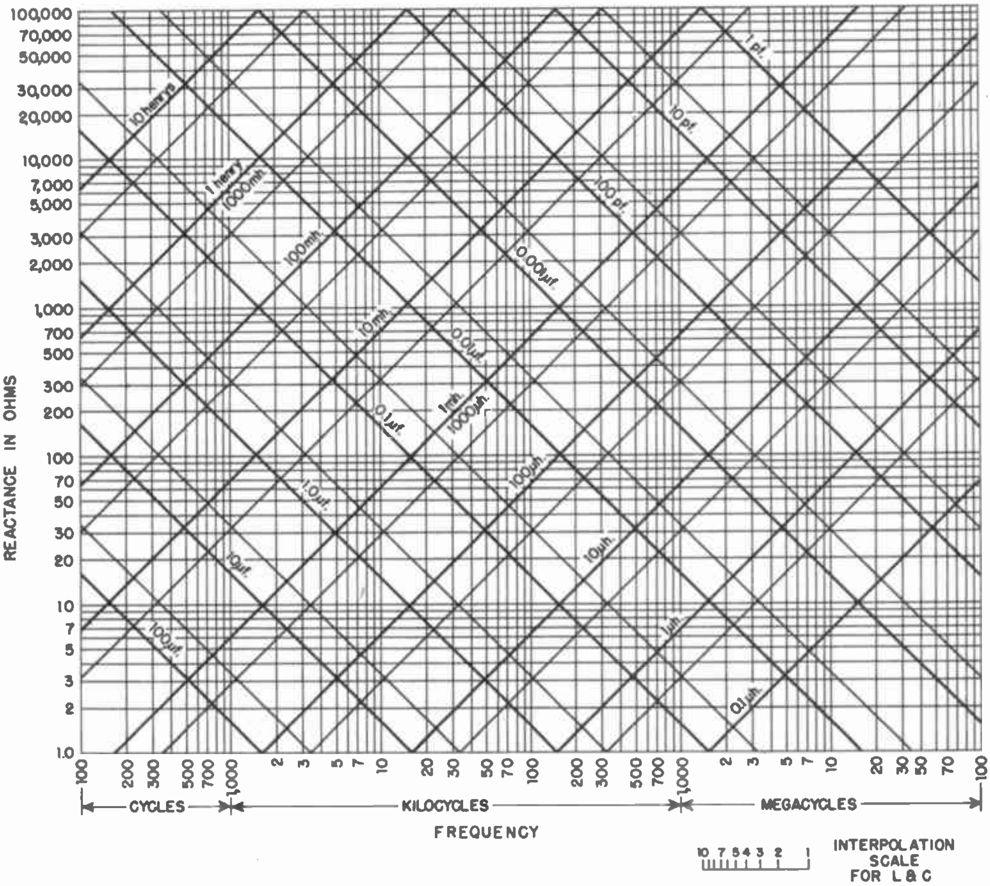


Fig. 2-28 – Inductive and capacitive reactance vs. frequency. Heavy lines represent multiples of 10, intermediate light lines multiples of 5; e.g., the light line between 10 μH and 100 μH represents 50 μH, the light line between 0.1 μF and 1 μF represents 0.5 μF, etc. Intermediate values can be estimated with the help of the interpolation scale.

Reactances outside the range of the chart may be found by applying appropriate factors to values within the chart range. For example, the reactance of 10 henrys at 60 cycles can be found by taking the reactance to 10 henrys at 600 cycles and dividing by 10 for the 10-times decrease in frequency.

Example: The reactance of a coil having an inductance of 8 henrys, at a frequency of 120 cycles, is

$$X_L = 2\pi fL = 6.28 \times 120 \times 8 = 6029 \text{ ohms}$$

The resistance of the wire of which the coil is wound has no effect on the reactance, but simply acts as though it were a separate resistor connected in series with the coil.

Ohm's Law for Reactance

Ohm's Law for an ac circuit containing *only* reactance is

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

$$E = IX$$

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

where E = Emf in volts
I = Current in amperes
X = Reactance in ohms

The reactance in the circuit may, of course, be either inductive or capacitive.

Example: If a current of 2 amperes is flowing through the capacitor of the earlier example (reactance = 47.4 ohms) at 7150 kHz, the voltage drop across the capacitor is

$$E = IX = 2 \times 47.4 = 94.8 \text{ volts}$$

If 400 volts at 120 hertz is applied to the 8-henry inductor of the earlier example, the current through the coil will be

$$I = \frac{E}{X} = \frac{420}{6029} = 0.0663 \text{ amp. (66.3 mA)}$$

Reactance Chart

The accompanying chart, Fig. 2-28, shows the reactance of capacitances from 1 pF to 100 μF, and the reactance of inductances from 0.1 μH to 10 henrys, for frequencies between 100 hertz and 100 megahertz per second. The approximate value of reactance can be read from the chart or, where more exact values are needed, the chart will serve as a check on the order of magnitude of reactances calculated from the formulas given above, and thus avoid "decimal-point errors."

Reactances in Series and Parallel

When reactances of the same kind are connected in series or parallel the resultant reactance is that of the resultant inductance or capacitance. This leads to the same rules that are used when determining the resultant resistance when resistors are combined. That is, for series reactances of the same kind the resultant reactance is

$$X = X1 + X2 + X3 + X4$$

and for reactances of the same kind in parallel the resultant is

$$X = \frac{1}{\frac{1}{X1} + \frac{1}{X2} + \frac{1}{X3} + \frac{1}{X4}}$$

or for two in parallel,

$$X = \frac{X1X2}{X1 + X2}$$

The situation is different when reactances of opposite kinds are combined. Since the current in a capacitance leads the applied voltage by 90 degrees and the current in an inductance lags the applied voltage by 90 degrees, the voltages at the terminals of opposite types of reactance are 180 degrees out of phase in a series circuit (in which the current has to be the same through all elements), and the currents in reactances of opposite types are 180 degrees out of phase in a parallel circuit (in which the same voltage is applied to all elements). The 180-degree phase relationship means that the currents or voltages are of opposite polarity, so in the series circuit of Fig. 2-29A the voltage *EL* across the inductive reactance *XL* is of opposite polarity to the voltage *EC* across the capacitive reactance *XC*. Thus if we call *XL* "positive" and *XC* "negative" (a common convention) the applied voltage *EAC* is *EL* – *EC*. In the parallel circuit at *B* the total current, *I*, is equal to *IL* – *IC*, since the currents are 180 degrees out of phase.

In the series case, therefore, the resultant reactance of *XL* and *XC* is

$$X = X_L - X_C$$

and in the parallel case

$$X = \frac{-X_L X_C}{X_L - X_C}$$

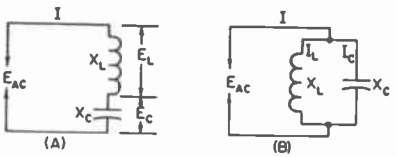


Fig. 2-29 – Series and parallel circuits containing opposite kinds of reactance.

Note that in the series circuit the total reactance is negative if *XC* is larger than *XL*; this indicates that the total reactance is capacitive in such a case. The resultant reactance in a series circuit is always smaller than the larger of the two individual reactances.

In the parallel circuit, the resultant reactance is negative (i.e., capacitive) if *XL* is larger than *XC*, and positive (inductive) if *XL* is smaller than *XC*, but in every case is always larger than the smaller of the two individual reactances.

In the special case where *XL* = *XC* the total reactance is zero in the series circuit and infinitely large in the parallel circuit.

Reactive Power

In Fig. 2-29A the voltage drop across the inductor is larger than the voltage applied to the circuit. This might seem to be an impossible condition, but it is not; the explanation is that while energy is being stored in the inductor's magnetic field, energy is being returned to the circuit from the capacitor's electric field, and vice versa. This stored energy is responsible for the fact that the voltages across reactances in series can be larger than the voltage applied to them.

In a resistance the flow of current causes heating and a power loss equal to I^2R. The power in a reactance is equal to I^2X, but is not a "loss"; it is simply power that is transferred back and forth between the field and the circuit but not used up in heating anything. To distinguish this "nondissipated" power from the power which is actually consumed, the unit of reactive power is called the volt-ampere-reactive, or var, instead of the watt. Reactive power is sometimes called "wattless" power.

IMPEDANCE

When a circuit contains both resistance and reactance the combined effect of the two is called impedance, symbolized by the letter *Z*. (Impedance is thus a more general term than either resistance or reactance, and is frequently used even for circuits that have only resistance or reactance, although usually with a qualification – such as "resistive impedance" to indicate that the circuit has only resistance, for example.)

The reactance and resistance comprising an impedance may be connected either in series or in parallel, as shown in Fig. 2-30. In these circuits the reactance is shown as a box to indicate that it may be either inductive or capacitive. In the series circuit the current is the same in both elements, with (generally) different voltages appearing across the resistance and reactance. In the parallel circuit the same voltage is applied to both elements, but different currents flow in the two branches.

Since in a resistance the current is in phase with the applied voltage while in a reactance it is 90 degrees out of phase with the voltage, the phase relationship between current and voltage in the circuit as a whole may be anything between zero and 90 degrees, depending on the relative amounts of resistance and reactance.

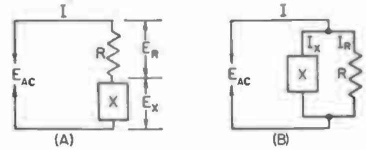


Fig. 2-30 — Series and parallel circuits containing resistance and reactance.

Series Circuits

When resistance and reactance are in series, the impedance of the circuit is

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

where Z = Impedance in ohms
R = Resistance in ohms
X = Reactance in ohms

The reactance may be either capacitive or inductive. If there are two or more reactances in the circuit they may be combined into a resultant by the rules previously given, before substitution into the formula above; similarly for resistances.

The "square root of the sum of the squares" rule for finding impedance in a series circuit arises from the fact that the voltage drops across the resistance and reactance are 90 degrees out of phase, and so combine by the same rule that applies in finding the hypothenuse of a right-angled triangle when the base and altitude are known.

Parallel Circuits

With resistance and reactance in parallel, as in Fig. 2-30B, the impedance is

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

where the symbols have the same meaning as for series circuits.

Just as in the case of series circuits, a number of reactances in parallel should be combined to find the resultant reactance before substitution into the formal above; similarly for a number of resistances in parallel.

Equivalent Series and Parallel Circuits

The two circuits shown in Fig. 2-30 are equivalent if the same current flows when a given voltage of the same frequency is applied, and if the phase angle between voltage and current is the same in both cases. It is in fact possible to "transform" any given series circuit into an equivalent parallel circuit, and vice versa.

Transformations of this type often lead to simplification in the solution of complicated circuits. However, from the standpoint of practical work the usefulness of such transformations lies in the fact that the impedance of a circuit may be modified by the addition of *either* series or parallel elements, depending on which happens to be most convenient in the particular case. Typical applications are considered later in connection with tuned circuits and transmission lines.

Ohm's Law for Impedance

Ohm's Law can be applied to circuits containing impedance just as readily as to circuits having resistance or reactance only. The formulas are

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

$$E = IZ$$

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

where E = Emf in volts
I = Current in amperes
Z = Impedance in ohms

Fig. 2-31 shows a simple circuit consisting of a resistance of 75 ohms and a reactance of 100 ohms in series. From the formula previously given, the impedance is

$$Z = \sqrt{R^2 + X_L{}^2} = \sqrt{(75)^2 + (100)^2} = 125$$

If the applied voltage is 250 volts, then

$$I = \frac{E}{Z} = \frac{250}{125} = 2 \text{ amperes}$$

This current flows through both the resistance and reactance, so the voltage drops are

$$E_R = IR = 2 \times 75 = 150 \text{ volts}$$

$$E_{XL} = IX_L = 2 \times 100 = 200 \text{ volts}$$

The simple arithmetical sum of these two drops, 350 volts, is greater than the applied voltage because the two voltages are 90 degrees out of phase. Their actual resultant, when phase is taken into account, is

$$\sqrt{(150)^2 + (200)^2} = 250 \text{ volts}$$

Power Factor

In the circuit of Fig. 2-31 an applied emf of 250 volts results in a current of 2 amperes, giving an apparent power of 250 X 2 = 500 watts. However, only the resistance actually consumes power. The power in the resistance is

$$P = I^2R = (2)^2 \times 75 = 300 \text{ watts}$$

The ratio of the power consumed to the apparent power is called the power factor of the circuit, and in this example the power factor would be 300/500 = 0.6. Power factor is frequently expressed as a percentage; in this case, it would be 60 percent.

"Real" or dissipated power is measured in watts; apparent power, to distinguish it from real power, is measured in volt-amperes. It is simply the product of volts and amperes and has no direct relationship to the power actually used up or dissipated unless the power factor of the circuit is known. The power factor of a purely resistive circuit is 100 percent or 1, while the power factor of a pure reactance is zero. In this illustration, the reactive power is $VAR = I2X$ = (2)2 X 100 = 400 volt-amperes.

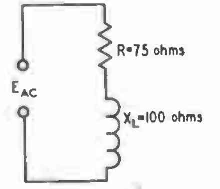


Fig. 2-31 — Circuit used as an example for impedance calculations.

Reactance and Complex Waves

It was pointed out earlier in this chapter that a complex wave (a "nonsinusoidal" wave) can be resolved into a fundamental frequency and a series of harmonic frequencies. When such a complex voltage wave is applied to a circuit containing reactance, the current through the circuit will not have the same wave shape as the applied voltage. This is because the reactance of an inductor and capacitor depend upon the applied frequency. For the second-harmonic component of a complex wave, the reactance of the inductor is twice and the reactance of the capacitor one-half their respective values at the fundamental frequency; for the third harmonic the inductor reactance is three times and the capacitor reactance one-third, and so on. Thus the circuit impedance is different for each harmonic component.

Just what happens to the current wave shape depends upon the values of resistance and reactance involved and how the circuit is arranged. In a simple circuit with resistance and inductive reactance in series, the amplitudes of the harmonic currents will be reduced because the inductive reactance increases in proportion to frequency. When capacitance and resistance are in series, the harmonic current is likely to be accentuated because the capacitive reactance becomes lower as the frequency is raised. When both inductive and capacitive reactance are present the shape of the current wave can be altered in a variety of ways, depending upon the circuit and the "constants," or the relative values of *L, C,* and *R*, selected.

This property of nonuniform behavior with respect to fundamental and harmonics is an extremely useful one. It is the basis of "filtering," or the suppression of undesired frequencies in favor of a single desired frequency or group of such frequencies.

TRANSFORMERS FOR AUDIO FREQUENCIES

Two coils having mutual inductance constitute a transformer. The coil connected to the source of energy is called the primary coil, and the other is called the secondary coil.

The usefulness of the transformer lies in the fact that electrical energy can be transferred from one circuit to another without direct connection, and in the process can be readily changed from one voltage level to another. Thus, if a device to be operated requires, for example, 115 volts ac and only a 440-volt source is available, a transformer can be used to change the source voltage to that required. A transformer can be used only with ac, since no voltage will be induced in the secondary if the magnetic field is not changing. If dc is applied to the primary of a transformer, a voltage will be induced in the secondary only at the instant of closing or opening the primary circuit, since it is only at these times that the field is changing.

THE IRON-CORE TRANSFORMER

As shown in Fig. 2-32, the primary and secondary coils of a transformer may be wound on a core of magnetic material. This increases the inductance of the coils so that a relatively small number of turns may be used to induce a given value of voltage with a small current. A closed core (one having a continuous magnetic path) such as that shown in Fig. 2-32 also tends to insure that practically all of the field set up by the current in the primary coil will cut the turns of the secondary coil. However, the core introduces a power loss because of hysteresis and eddy currents so this type of construction is normally practicable only at power and audio frequencies. The discussion in this section is confined to transformers operating at such frequencies.

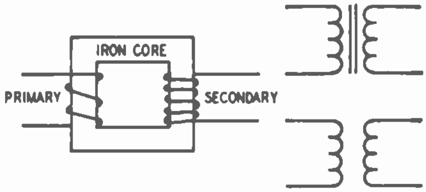


Fig. 2-32 — The transformer. Power is transferred from the primary coil to the secondary by means of the magnetic field. The upper symbol at right indicates an iron-core transformer, the lower one an air-core transformer.

Voltage and Turns Ratio

For a given varying magnetic field, the voltage induced in a coil in the field will be proportional to the number of turns in the coil. If the two coils of a transformer are in the same field (which is the case when both are wound on the same closed core) it follows that the induced voltages will be proportional to the number of turns in each coil. In the primary the induced voltage is practically equal to, and opposes, the applied voltage, as described earlier. Hence,

$$E_s = \frac{n_s}{n_p} E_p$$

where E_s = Secondary voltage
E_p = Primary applied voltage
n_s = Number of turns on secondary
n_p = Number of turns on primary

The ratio, ns/np is called the secondary-to-primary turns ratio of the transformer.

Example: A transformer has a primary of 400 turns and a secondary of 2800 turns, and an emf of 115 volts is applied to the primary.

$$E_s = \frac{n_s}{n_p} E_p = \frac{2800}{400} \times 115 = 7 \times 115$$
$$= 805 \text{ volts}$$

Also, if an emf of 805 volts is applied to the 2800-turn winding (which then becomes the primary) the output voltage from the 400-turn winding will be 115 volts.

Either winding of a transformer can be used as the primary, providing the winding has enough turns (enough inductance) to induce a voltage equal to the applied voltage without requiring an excessive current flow.

Effect of Secondary Current

The current that flows in the primary when no current is taken from the secondary is called the magnetizing current of the transformer. In any properly-designed transformer the primary inductance will be so large that the magnetizing current will be quite small. The power consumed by the transformer when the secondary is "open" – that is, not delivering power – is only the amount necessary to supply the losses in the iron core and in the resistance of the wire with which the primary is wound.

When power is taken from the secondary winding, the secondary current sets up a magnetic field that opposes the field set up by the primary current. But if the induced voltage in the primary is to equal the applied voltage, the original field must be maintained. Consequently, the primary must draw enough additional current to set up a field exactly equal and opposite to the field set up by the secondary current.

In practical calculations on transformers it may be assumed that the entire primary current is caused by the secondary "load." This is justifiable because the magnetizing current should be very small in comparison with the primary "load" current at rated power output.

If the magnetic fields set up by the primary and secondary currents are to be equal, the primary current multiplied by the primary turns must equal the secondary current multiplied by the secondary turns. From this it follows that

$$I_p = \frac{n_s}{n_p} I_s$$

where I_p = Primary current
I_s = Secondary current
n_p = Number of turns on primary
n_s = Number of turns on secondary

Example: Suppose that the secondary of the transformer in the previous example is delivering a current of 0.2 ampere to a load. Then the primary current will be

$$I_p = \frac{n_s}{n_p} I_s = \frac{2800}{400} \times 0.2 = 7 \times 0.2 = 1.4 \text{ amp.}$$

Although the secondary voltage is higher than the primary voltage, the secondary *current* is *lower* than the primary current, and by the same ratio.

Power Relationships; Efficiency

A transformer cannot create power; it can only transfer it and change the emf. Hence, the power taken from the secondary cannot exceed that taken by the primary from the source of applied emf. There is always some power loss in the resistance of the coils and in the iron core, so in all practical cases the power taken from the source will exceed that taken from the secondary. Thus,

$$P_o = nP_i$$

where P_o = Power output from secondary
P_i = Power input to primary
n = Efficiency factor

The efficiency, n, always is less than 1. It is usually expressed as a percentage; if n is 0.65, for instances, the efficiency is 65 percent.

Example: A transformer has an efficiency of 85 percent at its full-load output of 150 watts. The power input to the primary at full secondary load will be

$$P_i = \frac{P_o}{n} = \frac{150}{0.85} = 176.5 \text{ watts}$$

A transformer is usually designed to have its highest efficiency at the power output for which it is rated. The efficiency decreases with either lower or higher outputs. On the other hand, the *losses* in the transformer are relatively small at low output but increase as more power is taken. The amount of power that the transformer can handle is determined by its own losses, because these heat the wire and core. There is a limit to the temperature rise that can be tolerated, because too-high temperature either will melt the wire or cause the insulation to break down. A transformer can be operated a reduced output, even though the efficiency is low,because the actual loss will be low under such conditions.

The full-load efficiency of small power transformers such as are used in radio receivers and transmitters usually lies between about 60 and 90 percent, depending upon the size and design.

Leakage Reactance

In a practical transformer not all of the magnetic flux is common to both windings, although in well-designed transformers the amount of flux that "cuts" one coil and not the other is only a small percentage of the total flux. This leakage flux causes an emf of self-induction; consequently, there are small amounts of leakage inductance associated with both windings of the transformer. Leakage inductance acts in exactly the same way as an equivalent amount of ordinary inductance inserted in series with the circuit. It has, therefore, a certain reactance, depending upon the amount of leakage inductance and the frequency. This reactance is called leakage reactance.

Current flowing through the leakage reactance causes a voltage drop. This voltage drop increases with increasing current, hence it increases as more power is taken from the secondary. Thus, the greater the secondary current, the smaller the secondary terminal voltage becomes. The resistances of the transformer windings also cause voltage drops when current is flowing; although these voltage drops are not in phase with those caused by leakage reactance, together they result in a lower secondary voltage under load than is indicated by the turns ratio of the transformer.

At power frequencies (60 cycles) the voltage at the secondary, with a reasonably well-designed transformer, should not drop more than about 10 percent from open-circuit conditions to full load. The drop in voltage may be considerably more than this in a transformer operating at audio frequencies because the leakage reactance increases directly with the frequency.

Impedance Ratio

In an ideal transformer – one without losses or leakage reactance – the following relationship is true:

$$Z_p = Z_s\left[\frac{N_p}{N_s}\right]^2$$

where Z_p = Impedance looking into primary terminals from source of power
Z_s Impedance of load connected to secondary
N_p/N_s = Turns ratio, primary to secondary

That is, a load of any given impedance connected to the secondary of the transformer will be transformed to a different value "looking into" the primary from the source of power. The impedance transformation is proportional to the square of the primary-to-secondary turns ratio.

Example: A transformer has a primary-to-secondary turns ratio of 0.6 (primary has 6/10 as many turns as the secondary) and a load of 3000 ohms is connected to the secondary. The impedance looking into the primary then will be

$$Z_p = Z_s\left[\frac{N_p}{N_s}\right] = 3000 \times (0.6)^2 = 3000 \times 0.36$$
$$= 1080 \text{ ohms}$$

By choosing the proper turns ratio, the impedance of a fixed load can be transformed to any desired value, within practical limits. If transformer losses can be neglected, the transformed or "reflected" impedance has the same phase angle as the actual load impedance; thus if the load is a pure resistance the load presented by the primary to the source of power also will be a pure resistance.

The above relationship may be used in practical work even though it is based on an "ideal" transformer. Aside from the normal design requirements of reasonably low internal losses and low leakage reactance, the only requirement is that the primary have enough inductance to operate with low magnetizing current at the voltage applied to the primary.

The primary impedance of a transformer – *as it appears to the source of power* – is determined wholly by the load connected to the secondary and by the turns ratio. If the characteristics of the transformer have an appreciable effect on the impedance presented to the power source, the transformer is either poorly designed or is not suited to the voltage and frequency at which it is being used. Most transformers will operate quite well at voltages from slightly above to well below the design figure.

Impedance Matching

Many devices require a specific value of load resistance (or impedance) for optimum operation.

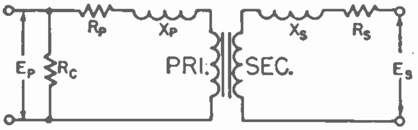


Fig. 2-33 – The equivalent circuit of a transformer includes the effects of leakage inductance and resistance of both primary and secondary windings. The resistance *Rc* is an equivalent resistance representing the core losses, which are essentially constant for any given applied voltage and frequency. Since these are comparatively small, their effect may be neglected in many approximate calculations.

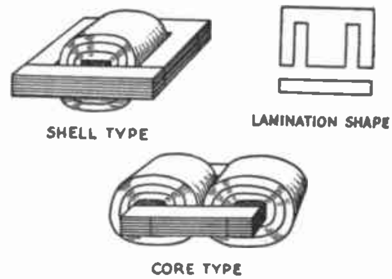


Fig. 2-34 – Two common types of transformer construction. Core pieces are interleaved to provide a continuous magnetic path.

The impedance of the actual load that is to dissipate the power may differ widely from this value, so a transformer is used to change the actual load into an impedance of the desired value. This is called impedance matching. From the preceding,

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

where N_p/N_s = Required turns ratio, primary to secondary
Z_p = Primary impedance required
Z_s = Impedance of load connected to secondary

Example: A vacuum-tube af amplifier requires a load of 5000 ohms for optimum performance, and is to be connected to a loud-speaker having an impedance of 10 ohms. The turns ratio, primary to secondary, required in the coupling transformer is

$$\frac{N_p}{N_s} = \sqrt{\frac{Z_p}{Z_s}} = \sqrt{\frac{5000}{10}} = \sqrt{500} = 22.4$$

The primary therefore must have 22.4 times as many turns as the secondary.

Impedance matching means, in general, adjusting the load impedance – by means of a transformer or otherwise – to a desired value. However, there is also another meaning. It is possible to show that any source of power will deliver its maximum possible output when the impedance of the load is equal to the internal impedance of the source. The impedance of the source is said to be "matched" under this condition. The efficiency is only 50 percent in such a case; just as much power is used up in the source as is delivered to the load. Because of the poor efficiency, this type of impedance matching is limited to cases where only a small amount of power is available and heating from power loss in the source is not important.

Transformer Construction

Transformers usually are designed so that the magnetic path around the core is as short as possible. A short magnetic path means that the transformer will operate with fewer turns, for a given applied voltage, than if the path were long. A short path also helps to reduce flux leakage and therefore minimizes leakage reactance.

Two core shapes are in common use, as shown in Fig. 2-34. In the shell type both windings are placed on the inner leg, while in the core type the

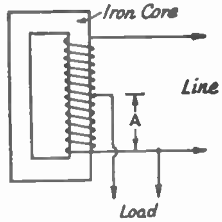


Fig. 2-35 – The autotransformer is based on the transformer principle, but uses only one winding. The line and load currents in the common winding *(A)* flow in opposite directions, so that the resultant current is the difference between them. The voltage across A is proportional to the turns ratio.

primary and secondary windings may be placed on separate legs, if desired. This is sometimes done when it is necessary to minimize capacitive effects between the primary and secondary, or when one of the windings must operate at very high voltage.

Core material for small transformers is usually silicon steel, called "transformer iron." The core is built up of laminations, insulated from each other (by a thin coating of shellac, for example) to prevent the flow of eddy currents. The laminations are interleaved at the ends to make the magnetic path as continuous as possible and thus reduce flux leakage.

The number of turns required in the primary for a given applied emf is determined by the size, shape and type of core material used, and the frequency. The number of turns required is inversely proportional to the cross-sectional area of the core. As a rough indication, windings of small power transformers frequently have about six to eight turns per volt on a core of 1-square-inch cross section and have a magnetic path 10 or 12 inches in length. A longer path or smaller cross section requires more turns per volt, and vice versa.

In most transformers the coils are wound in layers, with a thin sheet of treated-paper insulation between each layer. Thicker insulation is used between coils and between coils and core.

Autotransformers

The transformer principle can be utilized with only one winding instead of two, as shown in Fig. 2-35; the principles just discussed apply equally well. A one-winding transformer is called an autotransformer. The current in the common section (A) of the winding is the difference between the line (primary) and the load (secondary) currents, since these currents are out of phase. Hence if the line and load currents are nearly equal the common section of the winding may be wound with comparatively small wire. This will be the case only when the primary (line) and secondary (load) voltages are not very different. The autotransformer is used chiefly for boosting or reducing the power-line voltage by relatively small amounts. Continuously-variable autotransformers are commercially available under a variety of trade names; "Variac" and "Powerstat" are typical examples.

THE DECIBEL

In most radio communication the received signal is converted into sound. This being the case, it is useful to appraise signal strengths in terms of relative loudness as registered by the ear. A peculiarity of the ear is that an increase or decrease in loudness is responsive to the *ratio* of the amounts of power involved, and is practically independent of absolute value of the power. For example, if a person estimates that the signal is "twice as loud" when the transmitter power is increased from 10 watts to 40 watts, he will also estimate that a 400-watt signal is twice as loud as a 100-watt signal. In other words, the human ear has a *logarithmic* response.

This fact is the basis for the use of the relative-power unit called the decibel (abbreviated dB). A change of one decibel in the power level is just detectable as a change in loudness under ideal conditions. The number of decibels corresponding to a given power ratio is given by the following formula:

$$dB = 10 \log \frac{P_2}{P_1}$$

Common logarithms (base 10) are used.

Voltage and Current Ratios

Note that the decibel is based on *power* ratios. Voltage or current ratios can be used, but only when the impedance is the same for both values of voltage, or current. The gain of an amplifier cannot be expressed correctly in dB if it is based on the ratio of the output voltage to the input voltage unless both voltages are measured across the same value of impedance. When the impedance at both points of measurement is the same, the following formula may be used for voltage or current ratios:

$$dB = 20 \log \frac{V_2}{V_1} \quad \text{or } 20 \log \frac{I_2}{I_1}$$

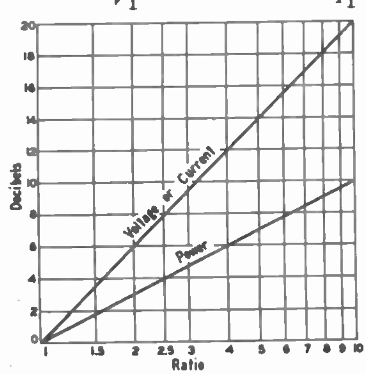


Fig. 2-36 – Decibel chart for power, voltage and current ratios for power ratios of 1:1 and 10:1. In determining decibels for current or voltage ratios the currents (or voltages) being compared must be referred to the same value of impedance.

Decibel Chart

The two formulas are shown graphically in Fig. 2-36 for ratios from 1 to 10. Gains (increases) expressed in decibels may be added arithmetically; losses (decreases) may be subtracted. A power decrease is indicated by prefixing the decibel figure with a minus sign. Thus +6 dB means that the power has been multiplied by 4, while −6 dB means that the power has been divided by 4.

The chart may be used for other ratios by adding (or subtracting, if a loss) 10 dB each time the ratio scale is multiplied by 10, for power ratios; or by adding (or subtracting) 20 dB each time the scale is multiplied by 10 for voltage or current ratios. For example, a power ratio of 2.5 is 4 dB (from the chart). A power ratio of 10 times 2.5, or 25, is 14 dB (10 + 4), and a power ratio of 100 times 2.5, or 250, is 24 dB (20 + 4). A voltage or current ratio of 4 is 12 dB, a voltage or current ratio of 40 is 32 dB (20 + 12), and one of 400 is 52 dB (40 + 12).

RADIO-FREQUENCY CIRCUITS

RESONANCE IN SERIES CIRCUITS

Fig. 2-37 shows a resistor, capacitor and inductor connected in series with a source of alternating current, the frequency of which can be varied over a wide range. At some *low* frequency the capacitive reactance will be much larger than the resistance of R, and the inductive reactance will be small compared with either the reactance of C or the resistance of R. (R is assumed to be the same at all frequencies.) On the other hand, at some very ***high*** frequency the reactance of C will be very small and the reactance of L will be very large. In either case the current will be small, because the net reactance is large.

At some intermediate frequency, the reactances of C and L will be equal and the voltage drops across the coil and capacitor will be equal and 180 degrees out of phase. Therefore they cancel each other completely and the current flow is determined wholly by the resistance, R. At that frequency the current has its largest possible value, assuming the source voltage to be constant regardless of frequency. A series circuit in which the inductive and capacitive reactances are equal is said to be resonant.

The principle of resonance finds its most extensive application in radio-frequency circuits. The reactive effects associated with even small inductances and capacitances would place drastic limitations on rf circuit operation if it were not possible to "cancel them out" by supplying the right amount of reactance of the opposite kind – in other words, "tuning the circuit to resonance."

Resonant Frequency

The frequency at which a series circuit is resonant is that for which $X_L = X_C$. Substituting the formulas for inductive and capacitive reactance gives

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

where f = Frequency in cycles per second
L = Inductance in henrys
C = Capacitance in farads
π = 3.14

These units are inconveniently large for radio-frequency circuits. A formula using more appropriate units is

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

where f = Frequency in kilohertz (kHz)
L = Inductance in microhenrys (μH)
C = Capacitance in picofarads (pF)
π = 3.14

Example: The resonant frequency of a series circuit containing a 5-μH inductor and a 35-pF capacitor is

$$f = \frac{10^6}{2\pi\sqrt{LC}} = \frac{10^6}{6.28 \times \sqrt{5 \times 35}}$$

$$= \frac{10^6}{6.28 \times 13.2} = \frac{10^6}{83} = 12{,}050 \text{ kHz}$$

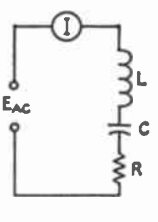


Fig. 2-37 – A series circuit containing L, C and R is "resonant" at the applied frequency when the reactance of C is equal to the reactance of L.

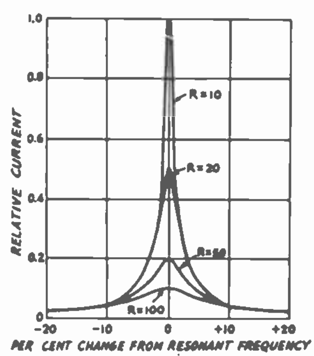


Fig. 2-38 – Current in a series-resonant circuit with various values of series resistance. The values are arbitrary and would not apply to all circuits, but represent a typical case. It is assumed that the reactances (at the resonant frequency) are 1000 ohms. Note that at frequencies more than plus or minus ten percent away from the resonant frequency the current is substantially unaffected by the resistance in the circuit.

The formula for resonant frequency is not affected by resistance in the circuit.

Resonance Curves

If a plot is drawn on the current flowing in the circuit of Fig. 2-37 as the frequency is varied (the applied voltage being constant) it would look like one of the curves in Fig. 2-38. The shape of the resonance curve at frequencies near resonance is determined by the ratio of reactance to resistance.

If the reactance of either the coil or capacitor is of the same order of magnitude as the resistance, the current decreases rather slowly as the frequency is moved in either direction away from resonance. Such a curve is said to be broad. On the other hand, if the reactance is considerably larger than the resistance the current decreases rapidly as the frequency moves away from resonance and the circuit is said to be sharp. A sharp circuit will respond a great deal more readily to the resonant frequency than to frequencies quite close to resonance; a broad circuit will respond almost equally well to a group or band of frequencies centering around the resonant frequency.

Both types of resonance curves are useful. A sharp circuit gives good selectivity – the ability to respond strongly (in terms of current amplitude) at one desired frequency and discriminate against others. A broad circuit is used when the apparatus must give about the same response over a band of frequencies rather than to a single frequency alone.

Q

Most diagrams of resonant circuits show only inductance and capacitance; no resistance is indicated. Nevertheless, resistance is always present. At frequencies up to perhaps 30 MHz this resistance is mostly in the wire of the coil. Above this frequency energy loss in the capacitor (principally in the solid dielectric which must be used to form an insulating support for the capacitor plates) also becomes a factor. This energy loss is equivalent to resistance. When maximum sharpness or selectivity is needed the object of design is to reduce the inherent resistance to the lowest possible value.

The value of the reactance of either the inductor or capacitor at the resonant frequency of a series-resonant circuit, divided by the *series* resistance in the circuit, is called the Q (quality factor) of the circuit, or

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

where Q = Quality factor
X = Reactance of either coil or capacitor in ohms
r = Series resistance in ohms

Example: The inductor and capacitor in a series circuit each have a reactance of 350 ohms at the resonant frequency. The resistance is 5 ohms. Then the Q is

$$Q = \frac{X}{r} = \frac{350}{5} = 70$$

The effect of Q on the sharpness of resonance of a circuit is shown by the curves of Fig. 2-39. In these curves the frequency change is shown in percentage above and below the resonant frequency. Qs of 10, 20, 50 and 100 are shown; these values cover much of the range commonly used in radio work. The unloaded Q of a circuit is determined by the inherent resistances associated with the components.

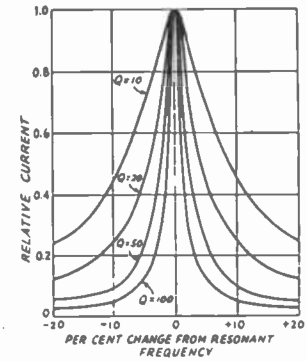


Fig. 2-39 — Current in series-resonant circuits having different Qs. In this graph the current at resonance is assumed to be the same in all cases. The lower the Q, the more slowly the current decreases as the applied frequency is moved away from resonance.

Voltage Rise at Resonance

When a voltage of the resonant frequency is inserted in series in a resonant circuit, the voltage that appears across either the inductor or capacitor is considerably higher than the applied voltage. The current in the circuit is limited only by the resistance and may have a relatively high value; however, the same current flows through the high reactances of the inductor and capacitor and causes large voltage drops. The ratio of the reactive voltage to the applied voltage is equal to the ratio of reactance to resistance. This ratio is also the Q of the circuit. Therefore, the voltage across either the inductor or capacitor is equal to QE where E is the voltage inserted in series. This fact accounts for the high voltages developed across the components of series-tuned antenna couplers (see chapter on "Transmission Lines").

RESONANCE IN PARALLEL CIRCUITS

When a variable-frequency source of constant voltage is applied to a parallel circuit of the type shown in Fig. 2-40 there is a resonance effect similar to that in a series circuit. However, in this case the "line" current (measured at the point indicated) is *smallest* at the frequency for which the inductive and capacitive reactances are equal. At that frequency the current through L is exactly canceled by the out-of-phase current through C, so that only the current taken by R flows in the line. At frequencies *below* resonance the current through L is larger than that through C, because the reactance of L is smaller and that of C higher at low frequencies; there is only partial cancellation of the two reactive currents and the line current therefore is larger than the current taken by R alone. At frequencies *above* resonance the situation is reversed and more current flows through C than

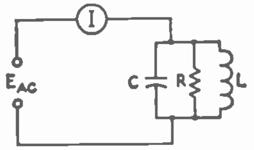


Fig. 2-40 — Circuit illustrating parallel resonance.

through L, so the line current again increases. The current at resonance, being determined wholly by R, will be small if R is large and large if R is small.

The resistance R shown in Fig. 2-40 is not necessarily an actual resistor. In many cases it will be the series resistance of the coil "transformed" to an equivalent parallel resistance (see later). It may be antenna or other load resistance coupled into the tuned circuit. In all cases it represents the total effective resistance in the circuit.

Parallel and series resonant circuits are quite alike in some respects. For instance, the circuits given at A and B in Fig. 2-41 will behave identically, when an external voltage is applied, if (1) L and C are the same in both cases; and (2) R multiplied by r, equals the square of the reactance (at resonance) of either L or C. When these conditions are met the two circuits will have the same Q. (These statements are approximate, but are quite accurate if the Q is 10 or more.) The circuit at A is a *series* circuit if it is viewed from the "inside" – that is, going around the loop formed by L, C and r – so its Q can be found from the ratio of X to r.

Thus a circuit like that of Fig. 2-41A has an equivalent parallel impedance (at resonance) of $R = \frac{X^2}{r}$; X is the reactance of either the inductor or the capacitor. Although R is not an actual resistor, to the source of voltage the parallel-resonant circuit "looks like" a pure resistance of that value. It is "pure" resistance because the inductive and capacitive currents are 180 degrees out of phase and are equal; thus there is no reactive current in the line. In a practical circuit with a high-Q capacitor, at the resonant frequency the parallel impedance is

$$Z_r = QX$$

where Z_r = Resistive impedance at resonance
Q = Quality factor of inductor
X = Reactance (in ohms) of either the inductor or capacitor

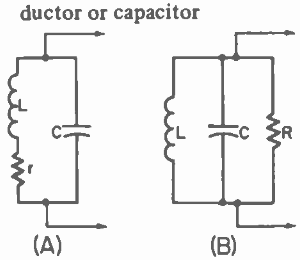


Fig. 2-41 — Series and parallel equivalents when the two circuits are resonant. The series resistance, r, in A is replaced in B by the equivalent parallel resistance $(R = X^2_C/r = X^2_L/r)$ and vice versa.

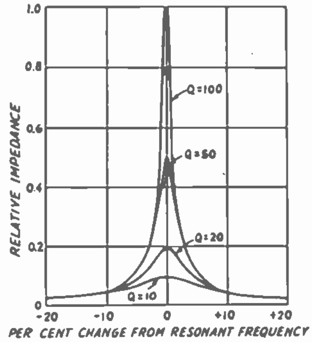


Fig. 2-42 – Relative impedance of parallel-resonant circuits with different Qs. These curves are similar to those in Fig. 2-39 for current in a series-resonant circuit. The effect of Q on impedance is most marked near the resonant frequency.

Example: The parallel impedance of a circuit with a coil Q of 50 and having inductive and capacitive reactance of 300 ohms will be

$$Z_r = QX = 50 \times 300 = 15{,}000 \text{ ohms}$$

At frequencies off resonance the impedance is no longer purely resistive because the inductive and capacitive currents are not equal. The off-resonant impedance therefore is complex, and is lower than the resonant impedance for the reasons previously outlined.

The higher the Q of the circuit, the higher the parallel impedance. Curves showing the variation of impedance (with frequency) of a parallel circuit have just the same shape as the curves showing the variation of current with frequency in a series circuit. Fig. 2-42 is a set of such curves. A set of curves showing the relative response as a function of the departure from the resonant frequency would be similar to Fig. 2-39. The −3 dB bandwidth (bandwidth at 0.707 relative response) is given by

$$\text{Bandwidth } -3 \text{ dB} = f_o/Q$$

where f_o is the resonant frequency and Q the circuit Q. It is also called the "half-power" bandwidth, for ease of recollection.

Parallel Resonance in Low-Q Circuits

The preceding discussion is accurate only for Qs of 10 or more. When the Q is below 10, resonance in a parallel circuit having resistance in series with the coil, as in Fig. 2-41A, is not so easily defined. There is a set of values for L and C that will make the parallel impedance a pure resistance, but with these values the impedance does not have its maximum possible value. Another set of values for L and C will make the parallel impedance a maximum, but this maximum value is not a pure resistance. Either condition could be called "resonance," so with low-Q circuits it is necessary to distinguish between maximum impedance and resistive impedance parallel resonance. The difference between these L and C values and the equal reactances of a series-resonant circuit is appreciable when the Q is in the vicinity of 5, and becomes more marked with still lower Q values.

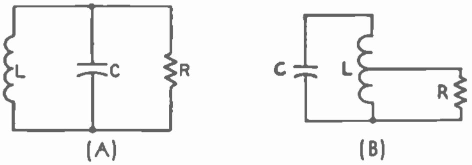


Fig. 2-43 – The equivalent circuit of a resonant circuit delivering power to a load. The resistor R represents the load resistance. At B the load is tapped across part of L, which by transformer action is equivalent to using a higher load resistance across the whole circuit.

Q of Loaded Circuits

In many applications of resonant circuits the only power lost is that dissipated in the resistance of the circuit itself. At frequencies below 30 MHz most of this resistance is in the coil. Within limits, increasing the number of turns in the coil increases the reactance faster than it raises the resistance, so coils for circuits in which the Q must be high are made with relatively large inductance for the frequency.

However, when the circuit delivers energy to a load (as in the case of the resonant circuits used in transmitters) the energy consumed in the circuit itself is usually negligible compared with that consumed by the load. The equivalent of such a circuit is shown in Fig. 2-43A, where the parallel resistor represents the load to which power is delivered. If the power dissipated in the load is at least ten times as great as the power lost in the inductor and capacitor, the parallel impedance of the resonant circuit itself will be so high compared with the resistance of the load that for all practical purposes the impedance of the combined circuit is equal to the load resistance. Under these conditions the Q of a parallel resonant circuit loaded by a resistive impedance is

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

where R = Parallel load resistance (ohms)
X = Reactance (ohms)

Example: A resistive load of 3000 ohms is connected across a resonant circuit in which the inductive and capacitive reactances are each 250 ohms. The circuit Q is then

$$Q = \frac{R}{X} = \frac{3000}{250} = 12$$

The "effective" Q of a circuit loaded by a parallel resistance becomes higher when the reactances are decreased. A circuit loaded with a relatively low resistance (a few thousand ohms) must have low-reactance elements (large capacitance and small inductance) to have reasonably high Q.

Impedance Transformation

An important application of the parallel-resonant circuit is as an impedance-matching device in the output circuit of a vacuum-tube rf power amplifier. As described in the chapter on vacuum tubes, there is an optimum value of load resistance for each type of tube and set of operating conditions. However, the resistance of the load to which the tube is to deliver power usually is considerably lower than the value required for proper tube operation. To transform the actual load resistance to the desired value the load may be tapped across part of the coil, as shown in Fig. 2-43B. This is equivalent to connecting a higher value of load resistance across the whole circuit, and is similar in principle to impedance transformation with an iron-core transformer. In high-frequency resonant circuits the impedance ratio does not vary exactly as the square of the turns ratio, because all the magnetic flux lines do not cut every turn of the coil. A desired reflected impedance usually must be obtained by experimental adjustment.

When the load resistance has a very low value (say below 100 ohms) it may be connected in series in the resonant circuit (as in Fig. 2-41A, for example), in which case it is transformed to an equivalent parallel impedance as previously described. If the Q is at least 10, the equivalent parallel impedance is

$$Z_r = \frac{X^2}{r}$$

where Z_r = Resistive parallel impedance at resonance
X = Reactance (in ohms) of either the coil or capacitor
r = Load resistance inserted in series

If the Q is lower than 10 the reactance will have to be adjusted somewhat, for the reasons given in the discussion of low-Q circuits, to obtain a resistive impedance of the desired value.

While the circuit shown in Fig. 2-43B will usually provide an impedance step-up as with an iron-core transformer, the network has some serious disadvantages for some applications. For instance, the common connection provides no dc isolation and the common ground is sometimes troublesome in regards to ground-loop currents. Consequently, a network in which only mutual magnetic coupling is employed is usually preferable. However, no impedance step-up will result unless the two coils are coupled tightly enough. The equivalent resistance seen at the input of the network will always be lower regardless of the turns ratio employed. However, such networks are still useful in impedance-transformation applications if the appropriate capacitive elements are used. A more detailed treatment of matching networks and similar devices will be taken up in the next section.

Unfortunately, networks involving reactive elements are usually narrowband in nature and it would be desirable if such elements could be eliminated in order to increase the bandwidth. With the advent of ferrites, this has become possible and it is now relatively easy to construct actual impedance transformers that are both broadband and permit operation well up into the vhf portion of the spectrum. This is also accomplished in part by tightly coupling the two (or more) coils that make up the transformer either by twisting the conductors together or winding them in a parallel fashion. The latter configuration is sometimes called a bifiliar winding.

COUPLED CIRCUITS AND FILTERS

Simple Ladder Networks

Two circuits are said to be coupled when a voltage or current in one network produces a voltage or current in the other one. The network where the energy originates is often called the **primary** circuit and the network that receives the energy is called the **secondary** circuit. Such coupling is often of a desirable nature since in the process, unwanted frequency components or noise may be rejected or isolated and power transferred from a source to a load with greatest efficiency. On the other hand, two or more circuits may be coupled inadvertently and undesirable effects produced. While a great number of coupling-circuit configurations are possible, one very important class covers so many practical applications that analysis of it will be covered in detail.

Any *two* circuits that are coupled can be drawn schematically as shown in Fig. 1A. A voltage source represented by E_{ac} with a source resistance R_p and a source reactance X_p is connected to the input of the coupling network, thus forming the primary circuit. At the output, a load reactance X_s and a load resistance R_s are connected as shown to form the secondary circuit. The circuit in the box could consist of an infinite variety of resistors, capacitors, inductors, and even transmission lines. However, it will be assumed that the network can be reduced to a combination of series and shunt elements consisting only of inductors and capacitors as indicated by the circuit shown in Fig. 1B. For obvious reasons, the circuit is often called a ladder network. In addition, if there are no resistive elements present, or if such elements can be neglected, the network is said to be dissipationless.

If a network is dissipationless, all the power delivered to the input of the network will be dissipated in the load resistance R_s. This effect leads to important simplifications in computations involved in coupled networks. The assumption of a dissipationless network is usually valid with transmitting circuits since even a small network loss (0.5 dB) will result in considerable heating at the higher power levels used in amateur applications. On the other hand, coupled circuits used in some receiving stages may have considerable loss. This is because the network may have some advantage and its high loss can be compensated by additional amplification in another stage. However, such devices form a relatively small minority of coupled networks commonly encountered and only the dissipationless case will be considered in this section.

Effective Attenuation and Insertion Loss

The most important consideration in any coupled network is the amount of power delivered to the load resistance, R_s, from the source, E_{ac}, with the network present. Rather than specify the source voltage each time, a comparison is made with the maximum available power from any source with a given primary resistance, R_p. The value of R_p might be considered as the impedance level associated with a complex combination of sources, transmission lines, coupled networks, and even antennas. Typical values of R_p are 52Ω, 75Ω, 300Ω, and 600Ω. The maximum available power is given by:

$$P_{max} = \frac{E_{ac}^2}{4R_p}$$

If the network is also dissipationless, the power delivered to the load resistance, R_s, is just the power "dissipated" in R_{in}. This power is related to the input current by:

$$P_o = I_{in}^2 R_{in}$$

and the current in terms of the other variables is:

$$I_{in} = \frac{E_{ac}}{\sqrt{(R_p + R_{in})^2 + (X_p + X_{in})^2}}$$

Combining the foregoing expressions gives a very useful formula for the ratio of power delivered to a load in terms of the maximum available power. This ratio expressed in decibels is given by:

$$\text{Attn} = -10\log\left(\frac{P_o}{P_{in}}\right) = -10\log\left[\frac{4R_{in}R_p}{(R_p + R_{in})^2 + (X_p + X_{in})^2}\right]$$

and is sometimes called the effective attenuation.

In the special case where X_p and X_s are either zero or can be combined into a coupling network, and where R_p is equal to R_s, the effective

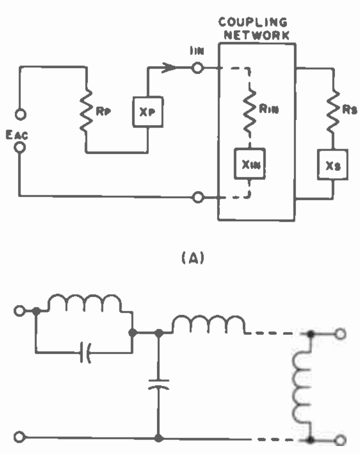


Fig. 1 — A representative coupling circuit (A) and ladder network (B).

attenuation is also equal to the insertion loss of the network. The insertion loss is the ratio of the power delivered to the load with the coupling network in the circuit to the power delivered to the load with the network absent. Unlike the effective attenuation which is always positive when defined by the previous formula, the insertion loss can take on negative values if R_p is not equal to R_s or if X_p and X_s are not zero. In effect, the insertion loss would represent a *power gain* under these conditions. The intepretation of this effect is that maximum available power does not occur with the coupling network out of the circuit because of the unequal source and load resistances and the non-zero reactances. With the network in the circuit, the resistances are now "matched" and the reactances are said to be "tuned out". The action of the coupling network in this instance is very similar to that of a transformer (which was discussed in a previous section) and networks consisting of "pure" inductors and capacitors are often used for this purpose. Such circuits are often referred to as matching networks. On the other hand, it is often desired to deliver the greatest amount of power to a load at some frequencies while rejecting energy at other frequencies. A device that accomplishes this action is called a filter. In the case of unequal source and load resistance, it is often possible to combine the processes of filtering and matching into one network.

Solving Ladder-Network Problems

From the last section, it is evident that if the values of R_{in} and X_{in} of Fig. 1A can be determined, the effective attenuation and possibly the insertion loss are also easily found. Being able to solve this problem has wide applications in rf circuits. For instance, design formulas for filters often include a simplifying assumption that the load resistance is constant with frequency. In the case of many circuits, this assumption is not true. However, if the value of R_s and X_s at any particular frequency is known, the attenuation of the filter can be determined even though it is improperly terminated.

Unfortunately, while the solution to any ladder problem is possible from a theoretical standpoint, practical difficulties are encountered as the network complexity increases. Many computations to a high degree of accuracy may be required, making the process a tedious one. Consequently, the availability of a calculator or similar computing device is recommended. The approach used here is adapted readily to any calculating method including the use of an inexpensive pocket calculator.

Susceptance and Admittance

The respective reactances of an inductor and a capacitor are given by:

$$X_L = 2\pi fL \qquad X_C = \frac{-1}{2\pi fC}$$

In a simple series circuit, the total resistance is just the sum of the individual resistances in the network

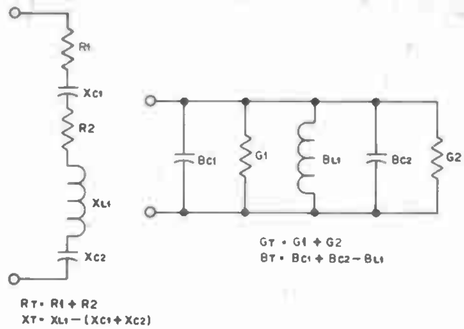


Fig. 2 — Resistances and reactances add in series circuits while conductances and susceptances add in parallel circuits. (Formulas shown are for *numerical* values of X and B.)

and the total reactance is the sum of the reactances. However, it is important to note the sign of the reactance. Since capacitive reactance is negative and inductive reactance is positive, it is possible that the sum of the reactances might be zero even though the individual reactances are not zero. In a series circuit, it will be recalled that the network is said to be resonant at the frequency where the reactances cancel.

A complementary condition exists in a parallel combination of circuit elements and it is convenient to introduce the concepts of admittance, conductance, and susceptance. In the case of a simple resistance, the conductance is just the reciprocal. That is, the conductance of a 50-Ω resistance is 1/50 or 2×10^{-2}. The reciprocal unit of the ohm is the mho. For simple inductances and capacitances, the formulas for the respective reciprocal entities are:

$$B_L = \frac{-1}{2\pi fL} \qquad B_C = 2\pi fC$$

and are defined as susceptances. In a parallel combination of conductances and susceptances, the total conductance is the sum of the individual conductances and the total susceptances is the sum of the individual susceptances taking the respective signs of the latter into account. A comparison between the way resistance and reactance add and the manner in which conductance and susceptance add is shown in the example of Fig. 2. An entity called admittance can be defined in terms of the total conductance and total susceptance by the formula:

$$Y = \sqrt{G_T{}^2 + B_T{}^2}$$

and is often denoted by the symbol ***Y***. If the impedance of a circuit is known, the admittance is just the reciprocal. Likewise, if the admittance of a circuit is known, the impedance is the reciprocal of the admittance. However, conductance, reactance, resistance, and susceptance are not so simply related. If the total resistance and total reactance of a series circuit are known, the conductance and

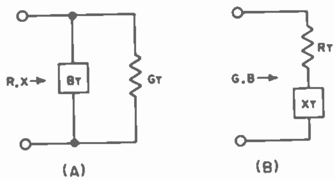


Fig. 3 — Application of conversion formulas can be used to transform a shunt conductance and susceptance to a series equivalent circuit. The converse is illustrated at (B).

susceptance of the circuit are related to the latter by the formulas:

$$G = \frac{R_T}{R_T^2 + X_T^2}, \quad B = \frac{-X_T}{R_T^2 + X_T^2}$$

On the other hand, if the total conductance and total susceptance of a parallel combination are known, the equivalent resistance and reactance can be found from the formulas:

$$R = \frac{G_T}{G_T^2 + B_T^2}, \quad X = \frac{-B_T}{G_T^2 + B_T^2}$$

These relations are illustrated in Fig. 3A and Fig. 3B respectively. While the derivation of the mathematical expressions will not be given, *the importance of the change of sign cannot be stressed too highly.* Solving network problems with a calculator is merely a matter of bookkeeping, and failure to take the sign change associated with the transformed reactance and susceptance is the most common source of error.

A Sample Problem

The following example illustrates the manner in which the foregoing theory can be applied to a practical problem. A filter with the schematic diagram shown in Fig. 4A is *supposed* to have an insertion loss at 6 MHz of 3 decibels when connected between a 52-Ω load and a source with a 52-Ω primary resistance (both X_p and X_s are zero). Since this is a case where the effective attenuation is equal to the insertion loss, the previous formula for effective attenuation applies. Therefore, it is required to find R_{in} and X_{in}.

Starting at the output, the values for the conductance and susceptance of the parallel *RC* circuit must be determined first. The conductance is just the reciprocal of 52Ω and the previous formula for capacitive susceptance gives the value shown in parentheses in Fig. 4A. (The upside-down Ω is the symbol for mho.) The next step is to apply the formulas for resistance and reactance in terms of the conductance and susceptance and the results give a 26-Ω resistance in series with a −26-Ω capacitive reactance as indicated in Fig. 4B. The reactance of the inductor can now be added to give a total reactance of 78.01Ω. The conductance and susceptance formulas can now be applied and the results of both of these operations is shown in Fig. 4C. Finally, adding the susceptance of the 510.1-pF capacitor (Fig. 4D) gives the circuit at Fig. 4E and applying the formulas once more gives the value of R_{in} and X_{in} (Fig. 4F). If the latter values are substituted into the effective attenuation formula, the insertion loss and effective attenuation are 3.01 dB, which is very closs to the value specified. The reader might verify that the insertion loss is 0.167, 0.37, and 5.5 dB at 3.5, 4.0, and 7.0 MHz respectively. If a plot of insertion loss versus frequency was constructed this would give the frequency response of the filter.

Frequency Sealing and Normalized Impedance

Quite often, it is desirable to be able to change a coupling network at one frequency and impedance level to another one. For example, suppose it was desired to move the 3-dB point of the filter in the preceding illustration from 6 to 7 MHz. An examination of the reactance and susceptance formulas reveals that multiplying the frequency by some constant *k* and dividing both the inductance and capacitance by the same value of *k* leaves the equations unchanged. Thus, if the capacitances and inductance in Fig. 4A are multiplied by 6/7, all the reactances and susceptances in the new circuit will now have the same value at 7 MHz that the old one had at 6 MHz.

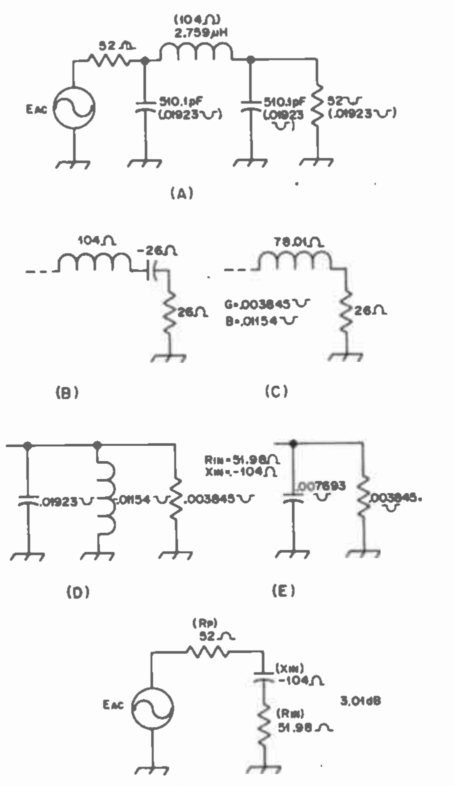


Fig. 4 — Problem illustrating network reduction to find insertion loss.

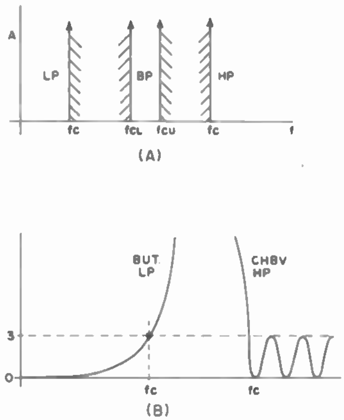


Fig. 5 — Ideal filter response curves are shown at (A) and characteristics of practical filters are shown at (B).

It is common practice with many filter tables especially, to present all the circuit components for a number of designs at some convenient frequency. Translating the design to some desired frequency is simply accomplished by multiplying all the components by some constant factor. The most common frequency used is the value of f such that $2\pi f$ is equal to 1.0. This is sometimes called a radian frequency of 1.0 and corresponds to 0.1592 Hz. To change a "one-radian" filter to a new frequency f_0 (in Hz), all that is necessary is to multiply the inductances and capacitances by $0.1592/f_0$.

In a similar manner, if one resistance (or conductance) is multiplied by some factor n, all the other resistances (or conductances) and reactances (or susceptances) must be multiplied by the same factor in order to preserve the network characteristics. For instance, if the secondary resistance, R_s is multiplied by n, all circuit inductances must be multiplied by n and the circuit capacitances *divided* by n (since capacitive reactance varies as the inverse of C). If, in addition to converting the filter of Fig. 4A to 7 MHz from 6 MHz, it was also desired to change the impedance level from 52 to 600Ω, the inductance would have to be multiplied by (6/7)(600/52) and the capacitances by (6/7)(52/600).

Using Filter Tables

In a previous example, it was indicated that the frequency response of a filter could be derived by solving for the insertion loss of the ladder network for a number of frequencies. The question might be asked if the converse is possible. That is, given a desired frequency response, could a network be found that would have this response? The answer is a qualified yes and the technical nomenclature for this sort of process is network synthesis. Frequency responses can be "cataloged" and, if a suitable one can be found, the corresponding network elements can be determined from an associated table. Filters derived by network synthesis and similar methods (such as optimized computer designs) are often referred to as "modern filters" even though the theory has been in existence for years. The term is useful in distinguishing such designs from those of an older approximate method called "image-parameter" theory.

Butterworth Filters

Filters can be grouped into four general categories as illustrated in Fig. 5A. Low-pass filters have zero insertion loss up to some critical frequency (f_c) or cutoff frequency and then provide high rejection above this frequency. (The latter condition is indicated by the shaded lines in Fig. 5A.) Band-pass filters have zero insertion loss between two cutoff frequencies with high rejection outside of the prescribed "bandwidth." (Band-stop filters reject a band of frequencies while passing all others.) And high-pass filters reject all frequencies below some cutoff frequency.

The attenuation shapes shown in Fig. 5A are ideal and can only be approached or approximated in practice. For instance, if the filter in the preceding problem was used for low-pass purposes in an 80-meter transmitter to reject harmonics on 40 meters, its performance would leave a lot to be desired. While insertion loss at 3.5 MHz was acceptable, it would likely be too high at 4.0 MHz and rejection would probably be inadequate at 7.0 MHz.

Fortunately, design formulas exist for this type of network and form a class called **Butterworth** filters. The name is derived from the shape of the curve for insertion-loss *vs.* frequency and is sometimes called a **maximally flat response.** A formula for the frequency response curve is given by:

$$A = 10\log_{10}\left[1 + \left(\frac{f}{f_c}\right)^{2k}\right]$$

where f_c is the frequency for an insertion loss of 3.01 dB, and k is the number of circuit elements.

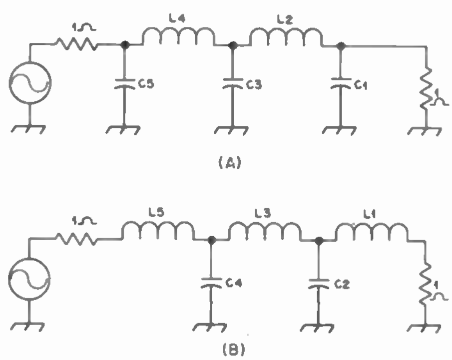


Fig. 6 — Schematic diagram of a Butterworth low-pass filter. (See Table I for element values.)

Table I

Fig. 6A	C1	L2	C3	L4	C5	L6	C7	L8	C9	L10
Fig. 6B	L1	C2	L3	C4	L5	C6	L7	C8	L9	C10
k										
1	2.0000									
2	1.4142	1.4142								
3	1.0000	2.0000	1.0000							
4	0.7654	1.8478	1.8478	0.7654						
5	0.6180	1.6180	2.0000	1.6180	0.6180					
6	0.5176	1.4142	1.9319	1.9319	1.4142	0.5176				
7	0.4450	1.2470	1.8019	2.0000	1.8019	1.2470	0.4450			
8	0.3902	1.1111	1.6629	1.9616	1.9616	1.6629	1.1111	0.3902		
9	0.3473	1.0000	1.5321	1.8794	2.0000	1.8794	1.5321	1.0000	0.3473	
10	0.3129	0.9080	1.4142	1.7820	1.9754	1.9754	1.7820	1.4142	0.9080	0.3129

The shape of a Butterworth low-pass filter is shown in the left-hand portion of Fig. 5B. (Another type that is similar in nature, only one that allows some "ripple" in the passband, is also shown in Fig. 5B. Here, a high-pass characteristic illustrates a Chebyshev response.)

As can be seen from the formula, increasing the number of elements will result in a filter that approaches the "ideal" low-pass shape. For instance, a 20-element filter designed for a 3.01-dB cutoff frequency of 4.3 MHz, would have an insertion loss at 4 MHz of 0.23 dB and 84.7 dB at 7 MHz. However, practical difficulties would make such a filter very hard to construct. Therefore, some compromises are always required between a theoretically perfect frequency response and ease of construction.

Element Values

Once the number of elements, *k*, is determined, the next step is to find the network configuration corresponding to *k*. (Filter tables sometimes have sets of curves that enable the user to select the desired frequency response curve rather than use a formula. Once the curve with the fewest number of elements for the specified passband and stop-band insertion loss is found, the filter is then fabricated around the corresponding value of *k*.) Table 1 gives normalized element values for values of *k* from 1 to 10. This table is for a 1-Ω source and load resistance (reactance zero) and a 3.01-dB cutoff frequency of 1 radian/second (0.1592 Hz). There are two possible circuit configurations, and these are shown in Fig. 6. Here, a 5-element filter is given as an example with either a shunt element next to the load (Fig. 6A) or a series element next to the load (Fig. 6B). Either filter will have the same response.

After the values for the 1-Ω, 1-radian/second "prototype" filter are found, the corresponding values for the actual frequency/impedance level can be determined (see the section on frequency and impedance scaling). The prototype inductance and capacitance values are multiplied by the ratio ($0.1592/f_c$) where f_c is the actual 3.01-dB cutoff frequency. Next, this number is multiplied by the load resistance in the case of an inductor and divided by the load resistance if the element is a capacitance. For instance, the filter in the preceding example is for a 3-element design (k equal to 3) and the reader might verify the values for the components for an f_c of 6 MHz and load resistance of 52Ω.

High-Pass Butterworth Filters

The formulas for change of impedance and frequency from the 1-Ω, 1-radian/second prototype to some desired level can also be conveniently written as:

$$L = \frac{R}{2\pi f_c} L_{\text{prototype}} \qquad C = \frac{1}{2\pi f_c R} C_{\text{prototype}}$$

where R is the load resistance in ohms, f_c is the desired 3.01-dB frequency in Hz. Then, L and C give the actual circuit-element values in henrys and farads in terms of the prototype element values from table 1.

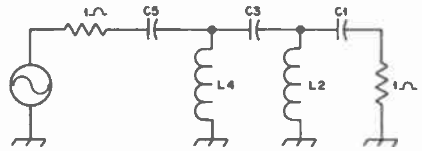
However, the usefulness of the low-pass prototype does not end here. If the following set of equations is applied to the prototype values, circuit elements for a high-pass filter can be obtained. The filter is shown in Fig. 7A and Fig. 7B which correspond to Fig. 6A and Fig. 6B in table 1. The equations for the actual high-pass circuit values in terms of the low-pass prototype are given by

$$C = \frac{1}{R\, 2\pi f_c C_{\text{prot.}}} \qquad L = \frac{R}{2\pi f_c L_{\text{prot.}}}$$

and the frequency response curve can be obtained from:

$$A = 10 \log \left[1 + \left(\frac{f_c}{f} \right)^{2k} \right]$$

For instance, a high-pass filter with 3 elements, a 3.01-dB f_c of 6 MHz and 52Ω, has a C1 and C3 of 510 pF and an L2 of 0.6897 μH. The insertion loss at 3.5 and 7 MHz would be 14.21 and 1.45 dB respectively.



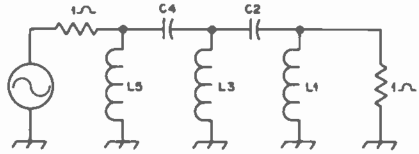


Fig. 7 — Network configuration of a Butterworth high-pass filter. The low-pass prototype can be transformed as described in the text.

Butterworth Band-Pass Filters

Band-pass filters can also be designed through the use of Table I. Unfortunately, the process is not as straightforward as it is for low-pass and high-pass filters if a practical design is to be obtained. In essence, a low-pass filter is resonated to some "center frequency" with the 3.01-dB cutoff frequency being replaced by the filter bandwidth. The ratio of the bandwidth to center frequency must be relatively large otherwise component values tend to become unmanageable.

While there are many variations of specifying such filters, a most useful approach is to determine an upper and lower frequency for a given attenuation. The center frequency and bandwidth are then given by:

$$f_o = \sqrt{f_1 f_2} \qquad BW = f_2 - f_1$$

If the bandwidth specified *is not* the 3.01-dB bandwidth (BW_c), the latter can be determined from:

$$BW_c = \frac{BW}{\left[10^{\frac{A}{10}} - 1\right]^{\frac{1}{2K}}}$$

in the case of a Butterworth response or from tables of curves. A is the required attenuation at the cutoff frequencies. The upper and lower cutoff frequencies (f_{cu} and f_{cl}) are then given by:

$$f_{cl} = \frac{-BW_c + \sqrt{(BW_c)^2 + 4f_o^2}}{2}$$

$$f_{cu} = f_{cl} + BW_c$$

A somewhat more convenient method is to pick a 3.01-dB bandwidth (the wider the better) around some center frequency and compute the attenuation at other frequencies of interest by using the transformation:

$$\frac{f}{f_c} = \left|\left(\frac{f}{f_o} - \frac{f_o}{f}\right) \frac{f_o}{BW_c}\right|$$

which can be substituted into the insertion-loss formula or table of curves.

As an example, suppose it is desired to build a band-pass filter for the 15-meter Novice band in order to eliminate the possibility of radiation on the 14- and 28-MHz bands. For a starting choice, 16 and 25 MHz will be picked as the 3.01-dB points giving a 3-dB bandwidth of 9 MHz. For these two points, f_o will be 20 MHz. It is common practice to equate the number of branch elements or filter resonators to certain mathematical entities called "poles" and the number of poles is just the value of k for purposes of discussion here. For a 3-pole filter (k of 3), the insertion loss will be 12.79 and 11.3 dB at 14 and 28 MHz respectively.

C1, C3 and L2 are then calculated for a 9-MHz low-pass filter and the elements for this filter are resonated to 20 MHz as shown in Fig. 8A. The response shape is plotted in Fig. 8B and it appears to be unsymmetrical about f_o. In spite of this fact, such filters are called symmetrical band-pass filters and f_o is the "center frequency."

If the response is plotted against a logarithmic frequency scale, the symmetry will become apparent. Consequently, using a logarithmic plot is helpful in designing filters of this type.

Examination of the component values reveals that while the filter is practical, it is a bit untidy from a construction standpoint. Rather than using a single 340.1-pF capacitor, paralleling a number of smaller valued units would be advisable. Encountering difficulty of this sort is typical of most filter designs, consequently, some tradeoffs between performance, complexity, and ease of construction are usually required.

Coupled Resonators

A problem frequently encountered in rf circuits is that of a coupled resonator. Applications include simple filters, oscillator tuned circuits, and even antennas. The circuit shown in Fig. 9A is illustrative of the basic principles involved. A series *RLC* circuit and the external terminals *ab* are "coupled" through a common capacitance, *C*m. Applying the formulas for conductance and susceptance in terms of series reactance and resistance gives the following set of formulas:

$$G_{ab} = \frac{R_r}{R_r^2 + X^2}$$

$$B_{ab} = B_{cm} - \frac{X}{R_r^2 + X^2}$$

The significance of these equations can be seen with the aid of Fig. 9B. At some point, the series inductive reactance will cancel the series capacitive reactance (at a point slightly below f_o where the conductance curve reaches a peak). Depending

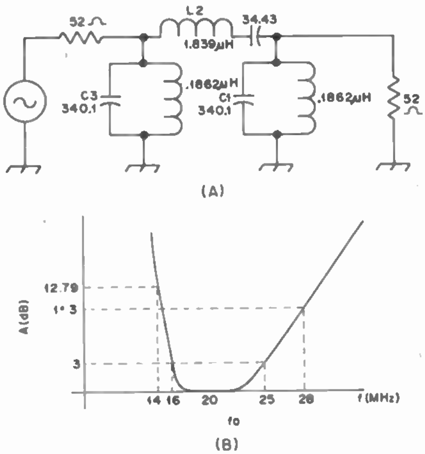


Fig. 8 — A Butterworth Bandpass filter. (Capacitance values are in picofarads.)

upon the value of the coupling susceptance, Bm, it is possible that another point can be found where the total input susceptance is zero. The input conductance at this frequency, f_o, is then G_o.

Since G_o is less than the conductance at the peak of the curve, $1/G_o$ or R_o is going to be greater than R_r. This effect can be applied when it is desired to match a low-value load resistance (such as found in a mobile whip antenna) to a more practical value. Suppose R_r and C_r in Fig. 9A are 10Ω and 21 pF respectively, and represent the equivalent circuit of a mobile antenna. Find the value of L_r and C_m which will match this antenna to a 52-Ω feed line at a frequency of 3900 kHz. Substituting the foregoing values into the formulas for input conductance gives:

$$\frac{1}{52} = \frac{10}{10^2 + X^2}$$

Solving for X (which is the *total* series reactance) gives a value of 20.49Ω. The reactance of a 21-pF capacitor at 3900 kHz is 1943.3Ω so the inductive reactance must be 1963.7Ω. (While either a positive or negative reactance will satisfy the equation for G_{ab}, a positive value is required to tune out B_{cm}. If the coupling element was a shunt inductor, the total reactance would have to be capacitive or negative in value.) Thus, the required inductance value for L_r will be 80.1 μH. In order to obtain a perfect match, the input susceptance must be zero and the value of B_{cm} can be found from the equation:

$$0 = B_{cm} - \frac{20.49}{10^2 + (20.49)^2}$$

giving a susceptance value of .04 mhos which corresponds to a capacitance of 1608 pF.

Piezoelectric Crystals

A somewhat different form of resonator consists of a quartz crystal between two conducting plates. If a voltage is applied to the plates, the resultant electric field causes a mechanical stress in the crystal. Depending upon the size and "cut" of the crystal, a frequency will exist at which the crystal begins to vibrate. The effect of this mechanical vibration is to simulate a series *RLC* circuit as in Fig. 9A. There is a capacitance associated with the crystal plates which appears across the terminals (C_m in Fig. 9A). Consequently, this circuit can also be analyzed with the aid of Fig. 9B. At some frequency (f_1 in Fig. 10), the series reactance is zero and G_{ab} in the preceding formula will just be $1/R_r$. Typical values for R_r range from 10 kΩ and higher. However, the equivalent inductance of the mechanical circuit is normally extremely high (over 10,000 *henries* in the case of some low-frequency units) which results in a very high circuit Q (30,000). Above f_1, the reactance is "inductive" and at f_2, the susceptance of the series resonator is just equal to the susceptance of the crystal holder, B_{cm}. Here, the total susceptance is zero. Since B_{cm} is usually very small, the equivalent series susceptance is also small. This means the value for X in the susceptance formula will be very large and consequently G_{ab} will be small, which corresponds to a high input resistance. A plot of the magnitude of the impedance is shown in Fig. 10. The dip at f_1 is called the series-resonant mode and the peak at f_2 is referred to as the parallel-resonant or "antiresonant" mode. When specifying crystals for oscillator applications, the type of mode must be given along with external capacitance across the holder or type oscillator circuit to be used. Otherwise, considerable difference in actual oscil-

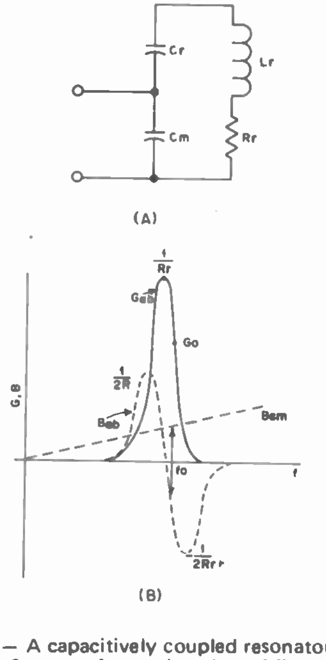


Fig. 9 — A capacitively coupled resonator is shown at (A). See text for explanation of figure shown at (B).

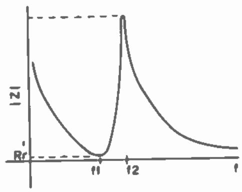


Fig. 10 – Frequency response of a quartz-crystal resonator. The minimum value is only approximate since holder capacitance is neglected.

lator frequency will be observed. The effect can be used to advantage and the frequency of a crystal oscillator can be "pulled" with an external reactive element or even frequency modulated with a device that converts voltage or current fluctuations into changes in reactance.

Coefficient of Coupling

If the solution to the mobile whip-antenna problem is examined, it can be seen that for a given frequency, R_r, L_r, and C_r, only one value of C_m results in an input load that appears as a pure resistance. While such a condition might be defined as resonance, the resistance value obtained is not necessarily the one required for maximum transfer of power.

A definition that is helpful in determining how to vary the circuit elements in order to obtain the desired input resistance is called the *coefficient of coupling*. The coefficient of coupling is defined as the ratio of the common or mutual reactance and the square root of the product of two specially defined reactances. If the mutual reactance is capacitive, one of the special reactances is the sum of the series capacitive reactances of the primary mesh (with the resonator disconnected) and the other one is the sum of the series capacitive reactances of the resonator (with the primary disconnected). Applying this definition to the circuit of Fig. 9A, the coefficient of coupling, k, is given by:

$$k = \sqrt{\frac{C_r}{C_r + C_m}}$$

How meaningful the coefficient of coupling will be depends upon the particular circuit configuration under consideration and which elements are being varied. For example, suppose the value of L_r in the mobile-whip antenna problem was fixed at 100 μH and C_m and C_r were allowed to vary. (It will be recalled that C_r is 21 pF and represents the antenna capacitance. However, the total resonator capacitance could be changed by adding a series capacitor between C_m and the antenna. Thus, C_r could be varied from 21 pF to some lower value but not a higher one.)

A calculated plot of k versus input resistance, R_{in}, is shown in Fig. 11. Note the unusually high change in k when going from resistance values near 10Ω to slightly higher ones.

Similar networks can be designed to work with any ratio of input resistance and load resistance but it is evident small ratios are going to pose difficulties. For larger ratios, component tolerances are more relaxed. For instance, C_m might consist of switchable fixed capacitors with C_r being variable. With a given load resistance, C_m essentially sets the value of the reactance and thus the input resistance while C_r and L_r provide the required reactance for the conductance formula. However, if L_r is varied, k varies also. Generally speaking, higher values of L_r (and consequently circuit Q) require lower values of k.

At this point, the question arises as to the significance and even the merit of such definitions as coefficient of coupling and Q. If the circuit element values are known, and if the configuration can be resolved into a ladder network, important properties such as input impedance and attenuation can be computed directly for any frequency. On the other hand, circuit information might be obscured or even lost by attempting to attach too much importance to an arbitrary definition. For example, the plot in Fig. 11 merely indicates C_m and C_r are changing with respect to one another. But it doesn't illustrate how they are changing. Such information is important in practical applications and even a simple table of C_m and C_r vs. R_{in} for a particular R_r would be much more valuable than a plot of k.

Similar precautions have to be taken with the interpretation of circuit Q. Selectivity and Q are simply related for single resonators and circuit components but the situation rapidly deteriorates with complex configurations. For instance, *adding* loss or resistance to circuit elements would seem to contradict the idea that low-loss or high-Q circuits provide the best selectivity. However, this is actually done in some filter designs to *improve* frequency response. In fact, the filter with the added loss has identical characteristics to one with "pure" elements. The method is called *predistortion* and is very useful in designing filters where practical considerations require the use of circuit elements with parasitic or undesired resistance.

As the frequency of operation is increased, discrete components become smaller until a point is reached where other forms of networks have to be used. Here, entities such as k and Q are sometimes the only means of describing such networks. Another definition of Q that is quite useful in this instance is that it is equal to the ratio

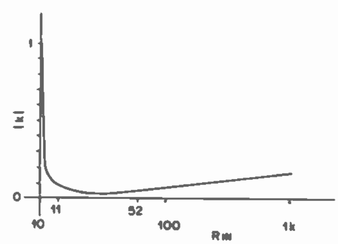
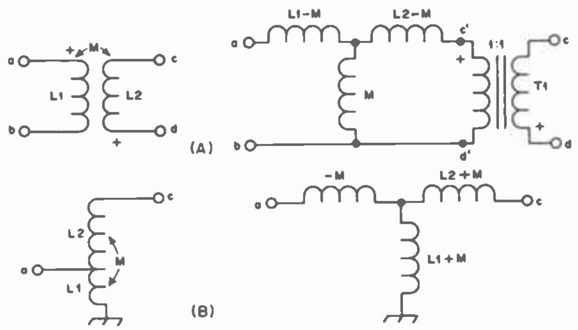


Fig. 11 – Variation of k with input resistance for circuit of Fig. 9.

Fig. 12 — Two types of magnetically coupled circuits. At (A), only mutual magnetic coupling exists while the circuit at (B) contains a common inductance also. Equivalents of both circuits are shown at the right which permit the application of the ladder-network analysis discussed in this section. (If the sign of voltage is unimportant, T1 can be eliminated.)



of 2π *(energy stored per rf cycle)/(energy lost per rf cycle).*

Mutually Coupled Inductors

A number of very useful rf networks involve coupled inductors. In a previous section, there was some discussion on iron-core transformers which represent a special case of the coupled-inductance problem. The formulas presented apply to instances where the coefficient of coupling is very close to 1.0. While it is possible to approach this condition at frequencies in the rf range, many practical circuits work at values of k that are considerably less than 1.0. A general solution is rather complex but many practical applications can often be simplified and solved through use of the ladder-network method. In particular, the sign of the mutual inductance must be taken into account if there are a number of coupled circuits or if the phase of the voltage between two coupled circuits is important.

The latter consideration can be illustrated with the aid of Fig. 12A. An exact circuit for the two mutually coupled coils on the left is shown on the right. T1 is an "ideal" transformer that provides the "isolation" between terminals *ab* and *cd*. If the polarity of the voltages between these terminals can be neglected, the transformer can be eliminated and just the circuit before terminals *c'd'* substituted. A second circuit is shown in Fig. 12B. Here, it is assumed that the winding sense doesn't change between L1 and L2. If so, then the circuit on the right of Fig. 12B can be substituted for the tapped coil shown at the left.

Coefficients of coupling for the circuits in Figs. 12A and 12B are given by:

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

$$k = \frac{L_1 + M}{\sqrt{L_1(L_1 + L_2 + 2M)}}$$

If L1 and L2 do not have the same value, an interesting phenonemon takes place as the coupling is increased. A point is reached where the mutual inductance exceeds the inductance of the smaller coil. The interpretation of this effect can be illustrated with the aid of Fig. 13. While all of the flux lines (as indicated by the dashed lines) associated with L1 also encircle turns of L2, there are additional ones that encircle extra turns of L2, also. Thus, there are more flux lines for M than there are for L1. Consequently, M becomes larger than L1. Normally, this condition is difficult to obtain with air-wound coils but the addition of ferrite material greatly increases the coupling. As k increases so that M is larger than L1 (Fig. 13), the network begins to behave more like a transformer and for a k of 1, the equivalent circuit of Fig. 12A yields the transformer equations of a previous section. On the other hand, for small values of k, the network becomes merely three coils arranged in a "T" fashion. One advantage of the circuit of Fig. 12A is that there is no direct connection between the two coils. This property is important from an isolation standpoint and can be used to suppress unwanted currents that are aften responsible for RFI difficulties.

Matching Networks

In addition to filters, ladder networks are frequently used to match one impedance value to another one. While there are many such circuits, a few of them offer particular advantages such as simplicity of design formulas or minimum number of elements. Some of the more popular ones are shown in Fig. 14. Shown at Fig. 14A and 14B, are two variations of an "L" network. These networks are relatively simple to design.

The situation is somewhat more complicated for the circuits shown at 14C and 14D. For a given value of input and output resistance, there are many networks that satisfy the conditions for a perfect match. The difficulty can be resolved by introducing the "dummy variable" labeled N.

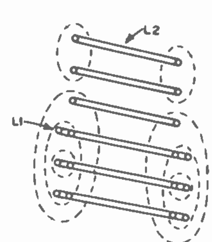
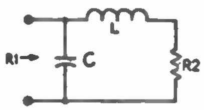


Fig. 13 — Diagram illustrating how M can be larger than one of the self inductances. This represents the transition from lightly coupled circuits to conventional transformers since an impedance step up is possible without the addition of capacitive elements.

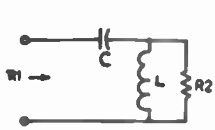


$R_1 > R_2$

$X_L = \sqrt{R_1 R_2 - R_2^2}$

$X_C = \frac{R_1 R_2}{X_L}$

(A)

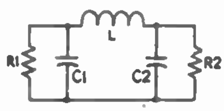


$R_2 > R_1$

$X_L = R_2 \sqrt{\frac{R_1}{R_2 - R_1}}$

$X_C = \frac{R_1 R_2}{X_L}$

(B)



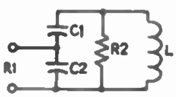
$R_1 > R_2,\ N > \sqrt{R_1/R_2 - 1}$

$X_{C1} = \frac{R_1}{N}$

$X_{C2} = \frac{R_2}{\sqrt{R_2/R_1\,(1 + N^2) - 1}}$

$X_L = R_1 \frac{N + R_2/X_{C2}}{N^2 + 1}$

(C)



$R_1 < R_2,\ N > \sqrt{R_2/R_1 - 1}$

$X_L = \frac{R_2}{N}$

$X_{C2} = \frac{R_1}{\sqrt{\frac{R_1\,(N^2 + 1)}{R_2} - 1}}$

$X_{C1} = \frac{R_2 N}{N^2 + 1} \left(1 - \frac{R_1}{N X_{C2}}\right)$

(D)

Fig. 14 — Four matching networks that can be used to couple a source and load with different resistance values. (Although networks are drawn with R1 appearing as the source resistance, all can be applied with R2 at the source end. Also, all formulas with capacitive reactance are for the numerical or absolute value.)

From a practical standpoint, N should be selected in order to optimize circuit component values. Either values of N that are too low or too high result in networks that are hard to construct.

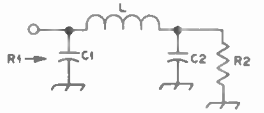
The reason for this complication is as follows. Only two reactive elements are required to match any two resistances. Consequently, adding a third element introduces a redundancy. This means one element can be assigned a value arbitrarily and the other two components can then be found. For instance, suppose C_2 in Fig. 14C is set to some particular value. The parallel combination of C_2 and R_2 can then be transformed to a series equivalent (see Fig. 15). Then, L could be found by breaking it down into two components, L' and L''. One component (L'') would tune out the remaining capacitive reactance of the output series equivalent circuit. The network is then reduced to the one shown in Fig. 14A and the other component (L') of L along with the value for C_1 could be determined from formulas (Fig. 14A). Adding the two inductive components would give the actual inductive reactance required for match in the circuit of Fig. 14C.

As mentioned before, it is evident an infinite number of networks of the form shown in Fig. 14C exist since C_2 can be assigned any value. Either a set of tables or a family of curves for C_1 and L in terms of C_2 could then be determined from the foregoing method and as illustrated in Fig. 15. However, similar data along with other information can be obtained by approaching the problem somewhat differently. Instead of setting one of the element values arbitrarily and finding the other two, a third variable is contrived and in the case of Fig. 14C and Fig. 14D is labeled N. All three reactances are then expressed in terms of the variable N.

The manner in which the reactances change with variation in N for two representative circuits of the type shown in Fig. 14C is shown in Fig. 16. The solid curve is for an R_1 of 3000Ω and R_2 equal to 52Ω. The dashed curve is for the same R_2 (52Ω) but with R_1 equal to 75Ω. For values of N very close to the minimum specified by the inequality (Fig. 14C), X_{C2} becomes infinite which means C_2 approaches zero. As might be expected, the values of X_L and X_{C1} at this point are approximately those of an L network (Fig. 14A) and could be determined by means of the formulas in Fig. 14A for the corresponding values of R_1 and R_2.

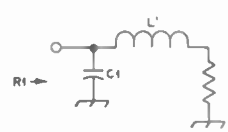
The plots shown in Fig. 16 should give a general idea of the optimum range of component values. The region close to the left-hand portion should be avoided since there is little advantage to be gained over an L, network while an extra component is required. For very high values of N, the capacitance values become large without producing any particular advantage either. A good design choice is an N a few percent above the minimum specified by the inequality.

Quite often, one of the elements is fixed with either one or the other or both of the remaining two elements variable. In many amateur transmitters, it is the inductor that remains fixed (at



$$X_L'' = X_{Ceq} = \frac{X_{C2}}{1 + (X_{C2}/R_2)^2}$$

$$\frac{R_2}{1 + (R_2/X_{C2})^2}$$



$$\frac{R_2}{1 + (R_2/X_{C2})^2}$$

Fig. 15 — Illustration of the manner in which the network of Fig. 14C can be reduced to the one of Fig. 14A assuming C_2 is assigned some arbitrary value. (The formulas shown are for *numerical* reactance values.)

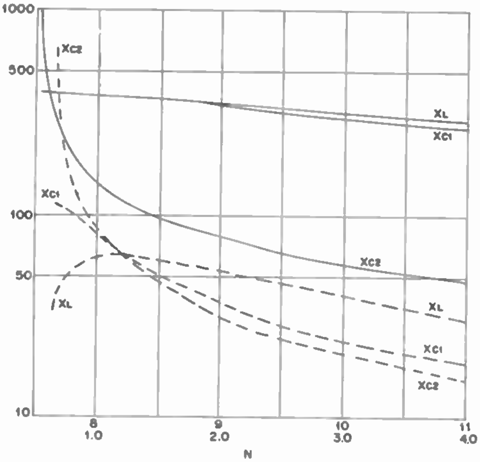


Fig. 16 — Network reactance variation as a function of dummy variable N. Solid curves and values of N from 8 to 11 are for an input resistance of 3000Ω and an output resistance of 52Ω. The dashed curves are for a similar network with an input and output resistance of 75 and 52Ω, respectively. Values of N from 1 to 4 are for the latter curves.

least for a given band) while C_1 and C_2 (Fig. 14C) are made variable. While this system limits the bandwidth and matching capability somewhat, it is still a very useful approach. For instance, the plot shown in Fig. 17 indicates the range of input resistance values that can be matched for an R_2 of 52Ω. The graph is for an inductive reactance of 219Ω. X_{C1} varies from 196 to 206Ω over the entire range of R_1 (or approximately 20 percent). However, X_{C2} varies from 15 to almost 100Ω as can be seen from the graph.

Since C_2 more or less sets the transformed resistance, it is often referred to as the "loading" control on transmitters using the network of Fig. 14C, with C_1 usually labeled "Tune." While the meaning of the latter term should be clear, the idea of loading in a matching application perhaps needs some explanation. For small values of X_{C2} (very large C_2), the transformed resistance is very high. Consequently, a source that was designed for a much lower resistance would deliver relatively little power. However, as the resistance is lowered, increasing amounts of current will flow resulting in more power output. Then, the source is said to be "loaded" more heavily.

Similar considerations such as those discussed for the network of Fig. 14C also exist for the circuit of Fig. 14D. Only the limiting L network for the latter is the one shown in Fig. 14B. The circuit of Fig. 14C is usually called a pi network and as pointed out, it is used extensively in the output stage of transmitters. The circuit of Fig. 14D has never been given any special name but it is quite popular in both antenna and transistor matching applications.

The plot shown in Fig. 16 is for fixed input and output resistances with the reactances variable. Similar figures can be plotted for other combinations of fixed and variable elements. An interesting case is for X_L and R_1 fixed with R_2, X_{C1}, and X_{C2} variable. A lower limit for N also exists for this plot only instead of an L network, the limiting circuit is a network of three *equal* reactances. A feature of this circuit is that the output resistance is the ratio of the square of the reactance and the input resistance. An analogous situation exists with a quarter-wavelength transmission-line transformer. The output resistance is the ratio of the square of the characteristic impedance of the line and the input resistance. Consequently, the special case where all the reactances are equal in the circuit of Fig. 14C is the lumped-constant analog of the quarter-wavelength transformer. It has identical phase shift (90 degrees) along with the same impedance-transforming properties.

Frequency Response

In many instances, a matching network performs a dual role in transforming a resistance value while providing frequency rejection. Usually, matching versatility, component values, and number of elements are the most important considerations. But a matching network might also be able to provide sufficient selectivity for some application, thus eliminating the need for a separate circuit such as a filter.

It will be recalled that Q and selectivity are closely related for simple RLC series and parallel circuits. Bandwidth and the parameter N of Fig. 14 are approximately related in this manner. For values of N much greater than the minimum specified by the inequality N and Q can be considered to mean the same thing for all practical purposes. However, the frequency response of networks that are more complex than simple RLC types is usually more complicated also. Consequently, some care is required in the interpretation of N or Q in regard

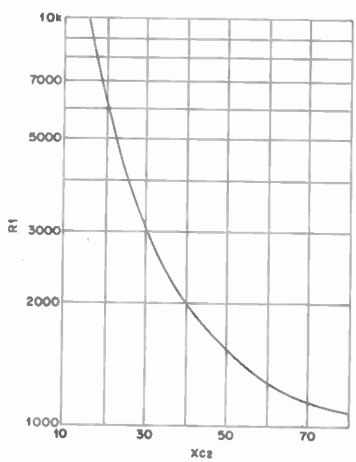


Fig. 17 — Input resistance *vs* output reactance for an output resistance of 52Ω. The curve is for a fixed inductor of 219Ω (Fig. 14C). X_{C1} varies from 196 to 206Ω.

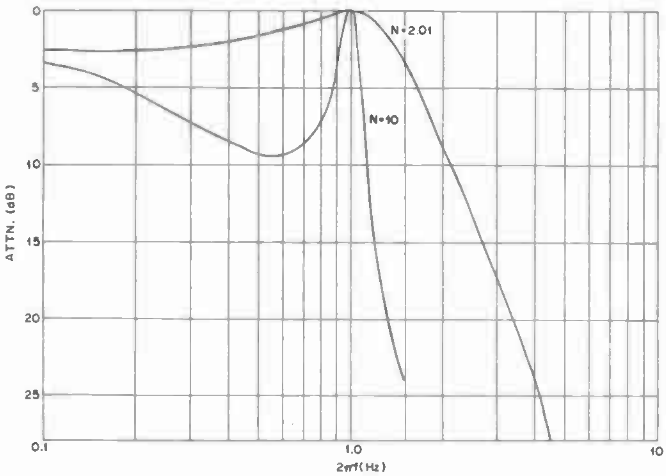


Fig. 18 – Frequency response of the network of Fig. 14C for two values of N.

to frequency rejection. For instance, a simple circuit has a frequency response that results in increasing attenuation for increasing excursions from resonance. This is not true for the pi network as can be seen from Fig. 18. For slight frequency changes below resonance, the attenuation increases as in the case of a simple *RLC* network. At lower frequencies, the attenuation decreases and approaches 2.55 dB. This plot is for a resistance ratio of 5:1 and the low-frequency loss is just caused by the mismatch in source and load resistance. Thus, while increasing N improves the selectivity near resonance, it has little effect on response for frequencies much farther away.

A somewhat different situation exists for the circuit of Fig. 14D. At frequencies far from resonance, either a series capacitance provides decoupling at the lower frequencies or a shunt capacitance causes additional mismatch at the higher ones. This circuit then, has a response resembling those of simple circuits unlike the pi network. Curves a and b are for a resistance ratio of 5:1 with N equal to 2.01 for curve a. Curve b is for an N of 10. Curves c and d are for a resistance ratio of 50:1 with N equal to 7.04 and 10 respectively.

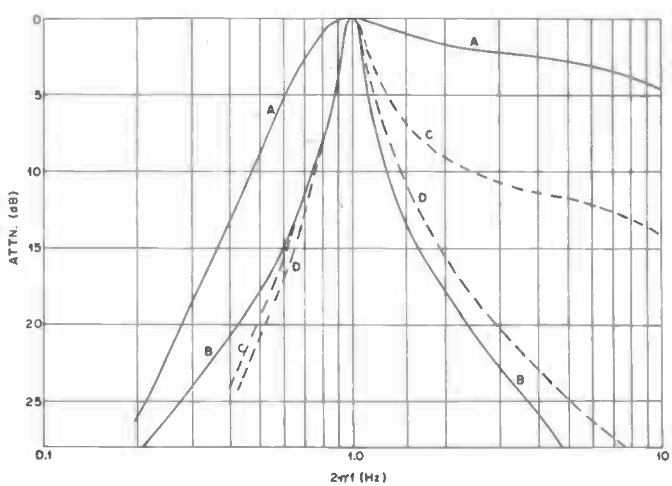


Fig. 19 – Frequency response of the circuit of Fig. 14D (see text).

Chapter 3

Radio Design Technique and Methods

Many amateurs desire to construct their own radio equipment and some knowledge of certain design procedures becomes important. Even when commercially manufactured equipment is mostly used, these techniques may still be required in setting up peripheral equipment such as antennas. Also, an applicant for an amateur radio license might be tested on material in this area.

"PURE" VS "UNPURE" COMPONENTS

In the chapter on Electrical Laws and Circuits, it was usually assumed that the components in an electrical circuit consisted solely of elements that could be reduced to a resistance, capacitance, or inductance. However, such elements do not exist in nature. An inductor always has some resistance associated with its windings and a carbon-composition resistor becomes a complicated circuit as the frequency of operation is increased. Even conductor resistance must be taken into account if long runs of cable are required.

In many instances, the effects of these parasitic components can be neglected and the actual device can be approximated by a "pure" element such as a resistor, capacitor, inductor, or a short circuit in the case of an interconnecting conductor. In other cases, the unwanted component must be taken into account. However, it may be possible to break the element down into a simple circuit consisting of single elements alone. Then, the actual circuit may be analyzed by means of the basic laws discussed in the previous chapter. It may be also possible to make a selection such that the effects of the residual element are negligible.

However, there are other parasitic elements that may not only be very difficult to remove but will affect circuit operation adversely as well. In fact, such considerations often set a limit on how

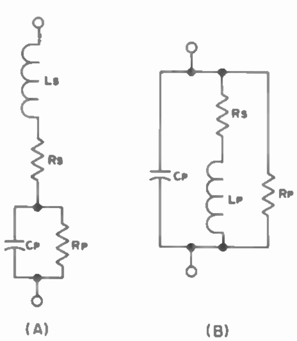


Fig. 1 – Equivalent circuit of a capacitor is shown at A, and for an inductor in B.

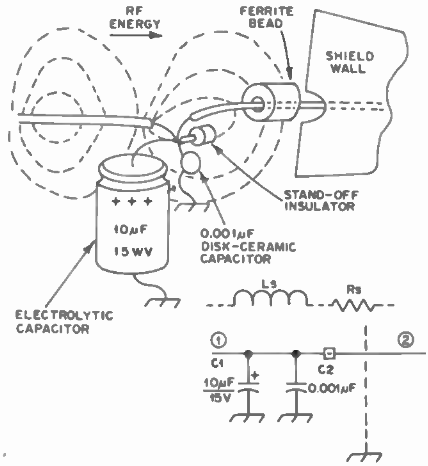


Fig. 2 – A bypassing arrangement that affords some measure of isolation (with the equivalent circuit shown in the inset). Dashed lines indicate a mode of wave travel that permits rf energy to leak past the bypass circuitry and should be taken into account when more stringent suppression requirements are necessary. (L_s and R_s in the inset represent the equivalent circuit of the ferrite bead.

stringent a design criteria can be tolerated. For instance, it is a common practice to connect small-valued capacitors in various parts of a complex circuit such as a transmitter or receiver for bypassing purposes. A bypass capacitor permits energy below some specified frequency to pass a given point while providing rejection to energy at higher frequencies. In essence, the capacitor is used in a crude form of filter. In more complicated filter designs, capacitors may be required for complex functions (such as matching) in addition to providing a low reactance to ground.

An equivalent circuit of a capacitor is shown in Fig. 1A. Normally, the series resistance, R_s, can be neglected. On the other hand, the upper frequency limit of the capacitor is limited by the series inductance, L_s. In fact, above the point where L_s and C_p form a resonant circuit, the capacitor actually appears as an inductor at the external terminals. As a result, it becomes useless for bypassing purposes. This is why it is common practice to use two capacitors in parallel for bypassing as shown in Fig. 2. At first inspection, this might appear as superfluous duplication. But the "self-resonant" frequency of a capacitor is lower for high-capacitance units than it is for smaller valued ones. Thus, C_1 in Fig. 2 provides a low reactance for low frequencies such as those in the audio range while C_2 acts as a bypass for frequencies above the self-resonant frequency of C_1.

Rf Leakage

Although the capacitor combination shown in Fig. 2 provides a low-impedance path to ground, it may not be very effective in preventing rf energy from reaching point 2 that travels along the conductor from point 1. At dc and low-frequency applications, a circuit must always form a closed path in order for a current to flow. Consequently, two conductors are required if power is to be delivered from a source to a load. In many instances, one of the conductors may be common to several other circuits and constitutes a local ground.

However, as the frequency of operation is increased, a second type of coupling mechanism is possible. Power may be transmitted *along a single conductor.* (Although the same effect is possible at low frequencies, unless circuit dimensions are extremely large, such transmission effects can be neglected.) The conductor acts as a waveguide in much the same manner as a large conducting surface, such as the earth, will permit propagation of a radio wave close to its boundary with the air. This latter type of propagation is often called a ground wave and is important up to and slightly above the standard a-m broadcast band. At higher frequencies, the conductivity of the earth is such to attenuate ground-wave propagation.

A mode similar to ground-wave propagation that can travel along the boundary of a single conductor is illustrated by the dashed lines in Fig. 2. As with the wave traveling close to the earth, a poor conducting bóundary will cause attenuation. This is why a ferrite bead is often inserted over the exit point of a conductor from an area where rf energy is to be contained or excluded. In addition to loss (particularly in the vhf range), the high permeability of the ferrite introduces a series-inductive reactance as well. Finally, the shield wall provides further isolation.

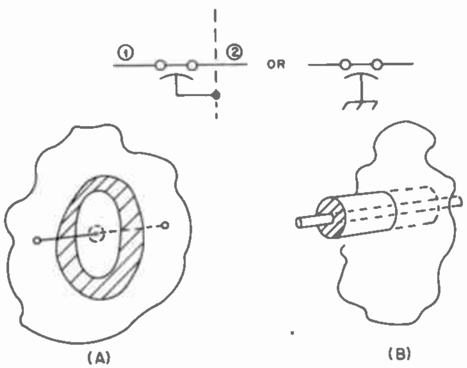


Fig. 3 — A superior type of bypassing arrangement to that shown in Fig. 2. Concentric conductors provide a low-inductance path to ground and better rejection of unwanted single-wire wave modes.

While the techniques shown in Fig. 2 get around some of the deficiencies of capacitors that are used for bypass purposes, the resulting suppression is inadequate for a number of applications. Examples would be protection of a VFO to surrounding rf energy, a low-frequency receiver with a digital display, and suppression of radiated harmonic energy from a transmitter. In each of these cases, a very high degree of isolation is required. For instance, a VFO is sensitive to voltages that appear on dc power supply lines and a transmitter output with a note that sounds "fuzzy" or rough may result. Digital displays usually generate copious rf energy in the lf spectrum. Consequently, a receiver designed for this range presents a situation where a strong source of emission is in close proximity to very sensitive receiving circuits. A similar case exists with transmitters operating on a frequency that is a submultiple of a fringe-area TV station. In the latter two instances, the problem is not so severe if the desired signal is strong enough to "override" the unwanted energy. Unfortunately, this is not the case normally and stringent measures are required to isolate the sensitive circuits from the strong source.

A different type of bypass-capacitor configuration is often used with associated shielding for such applications, as shown in Fig. 3. In order to reduce the series inductance of the capacitor, and to provide better isolation between points 1 and 2, either a disk-type (Fig. 3A) or a coaxial configuration (Fig. 3B) is employed. The circuit diagram for either configuration is shown in the inset. While such "feedthrough" capacitors are always connected to ground through the shield, this connection is often omitted on drawings. Only a connection to ground is shown as in the inset at the right in Fig. 3B.

Dielectric Loss

Even though capacitors are usually high-Q devices, the effect of internal loss can be more severe than the case of a coil. This is because good insulators of electricity are usually good insulators of heat also. Therefore, heat generated in a capacitor must be conducted to the outside via the conducting plates to the capacitor leads. In addition, most capacitors are covered with an insulating coating that further impedes heat conduction. The problem is less severe with capacitors using air as a dielectric for two reasons. The first advantage of air over other dielectrics is that the loss in the presence of an alternating electric field is extremely small. Secondly, any heat generated by currents on the surface of the conducting plates is either conducted away by air currents or through the mass of the metal.

The dielectric loss in a capacitor can be represented by R_p as shown in Fig. 1A. However, if a dc ohmmeter was placed across the terminals of the capacitor, the reading would be infinite. This is because dielectric loss is an ac effect. Whenever an alternating electric field is applied to an insulator, there is a local motion of the electrons in the individual atoms that make up the material. Even though the electrons are not displaced as they would be in a conductor, this local motion requires the expenditure of energy and results in a power loss.

Consequently, some care is required in the application of capacitors in moderate to high-power circuits. The applied voltage should be such that rf-current ratings are not exceeded for the particular frequency of operation. This is illustrated in Fig. 4. A parallel-resonant circuit consisting of L_p, C_p, and R_L is connected to a voltage source, V_s, through a coupling capacitor, C_c. It is also assumed that R_L is much greater than either the inductive or capacitive reactance taken alone. This condition would be typical of that found in most rf-power amplifier circuits employing vacuum tubes.

Since the inductive and capacitive reactance of L_p and C_p cancel at resonance, the load presented to the source would be just R_L. This would mean the current through C_c would be much less than the current through either C_p or L_p. The effect of such "current rise" is similar to the voltage rise at resonance discussed in the previous chapter. Even though the current at the input of the parallel-resonant circuit is small, the currents that flow in the elements that make up the circuit can be quite large.

The requirements for C_c then, would be rather easy to satisfy in regards to current rating and power dissipation. On the other hand, C_p would ordinarily be restricted to air-variable types although some experiments have been successful using Teflon as a dielectric.[1] Generally speaking, the coupling capacitor should have a low reactance (at the lowest frequency of operation) in comparison to the load presented by the tuned circuit. The effect of the coupling-capacitor reactance could then be compensated by slightly retuning the parallel-resonant circuit.

Inductors

Similar considerations to those discussed in the previous sections exist with inductors also, as shown in Fig. 1B. Since an inductor usually consists of a coil of wire, there will be a resistance associated with the wire material and this component is represented by R_s (Fig. 1B). In addition, there is always a capacitance associated with conductors in proximity as illustrated in Fig. 5. While such capacitance is distributed throughout the coil, it is a convenient approximation to consider an equivalent capacitance, C_p, exists between the terminals (Fig. 1B). Finally, inductors are often wound on materials that have high permeability in order to increase the inductance. Thus, it is possible to build an inductor with fewer turns and smaller in size than an equivalent coil with an "air core."

Unfortunately, high-permeability materials

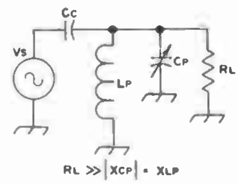


Fig. 4 — Consideration of capacitor voltage and current ratings should be kept in mind in moderate-power applications.

presently available have considerable loss in the presence of an rf field. It will be recalled a similar condition existed with the dielectric in a capacitor. Consequently, in addition to the wire resistance, a loss resistance is associated with the core and represented by R_p. (See Fig. 1B.) Since this loss is more or less independent of the current through the coil but dependent upon the applied voltage, it is represented by a parallel resistance.

RF TRANSFORMERS

Although the term transformer might be applied to any network that "transformed" a voltage or an impedance from one level to another one, the term is usually reserved for circuits incorporating mutual magnetic coupling. Examples would be i-f transformers, baluns, broadband transformers, and certain antenna matching networks. Of course, many devices used at audio and power frequencies are also transformers in the sense used here and have been covered in a previous chapter.

Networks that use mutual magnetic coupling exclusively have attractive advantages over other types in many common applications. A principal advantage is that there is no direct connection between the input and output terminals. Consequently, dc and ac components of current are separated easily thus eliminating the need for coupling capacitors. Perhaps even more importantly, it is also possible to isolate rf currents because of the lack of a common conductor. Quite often, an hf receiver in an area where strong local broadcast stations are present will suffer from "broadcast harmonics" and possibly even rectified audio signals getting into sensitive af circuits. In such cases, complicated filters sometimes prove ineffective while a simple tuned rf transformer clears up the problem completely. This is because the unwanted bc components are prevented from

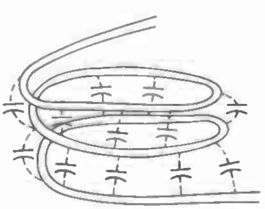


Fig. 5 — Distributed capacitance (indicated by dashed lines) affects the operation of a coil at high frequencies.

[1]DeMaw and Dorbuck, "Transmitting Variables," *QST* for February, 1975, p. 37.

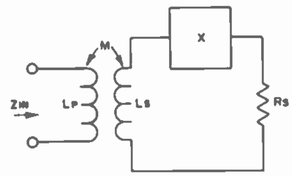


Fig. 6 — Basic magnetically coupled circuit.

flowing on the receiver chassis along with being rejected by the tuned-transformer filter characteristic.

A second advantage of coupled circuits using mutual magnetic coupling exclusively is that analysis is relatively simple compared to other forms of coupling although exact synthesis is somewhat complicated. That is, finding a network with some desired frequency response would be quite difficult in the general case.

However, circuits using mutual-magnetic coupling usually have very good out-of-band rejection characteristics when compared to networks incorporating other forms. (A term sometimes applied to transformer or mutual-magnetic coupling is indirect coupling. Circuits with a single resistive or reactive element for the common impedance are called direct-coupled networks. Two or more elements in the common impedance are said to comprise complex coupling.) For instance, relatively simple band-pass filters are possible with mutual-magnetic coupling and are highly recommended for vhf-transmitter multiplier chains. For receiving, such filters are often the main source of selectivity. Standard a-m and fm broadcast receivers would be examples where intermediate-frequency (i-f) transformers derive their band-pass characteristics from mutually coupled inductors.

A third advantage of mutually coupled networks is that practical circuits with great flexibility particularly in regard to matching capabilities are possible. For this reason, variable-coupling matching networks or those using "link coupling" have been popular for many years. In addition to matching flexibility, these circuits are good band-pass filters and can also provide isolation between antenna circuits and those of the transmitter.

Design Formulas

A basic two-mesh circuit with mutual magnetic coupling is shown in Fig. 6. The reactance, X, is arbitrary and could be either inductive or capacitive. However, it is convenient to combine it with the secondary reactance (X_{LS}) since this makes the equations somewhat more compact. Hence, the total secondary reactance is defined by

$$X_s = 2\pi f L_s + X$$

The primary reactance and mutual reactance are also defined respectively as

$$X_p = 2\pi f L_p, \quad X_m = 2\pi f M$$

A set of equations for the input resistance and reactance is given by

$$R_{in} = \frac{R_s X_m^2}{R_s^2 + X_s^2}$$

$$X_{in} = X_p - \frac{X_m^2 X_s}{R_s^2 + X_s^2}$$

This permits reducing the two-mesh circuit of Fig. 6 to the single-mesh circuit of Fig. 7.

Double-Tuned Circuits

A special case occurs if the value of X_s is zero. This could be accomplished easily by tuning out the inductive reactance of the secondary with an appropriate capacitor or by varying the frequency until a fixed capacitor and the secondary inductance resonated. Under these conditions, the input resistance and reactance would be

$$R_{in} = \frac{X_m^2}{R_s}, \quad X_{in} = X_p$$

Then, in order to make the input impedance purely resistive, a second series capacitor could be used to cancel the reactance of X_p. The completed network is shown in Fig. 8 with C_1 and C_2 being the primary and secondary series capacitors.

If X_m could be varied, it is evident that the secondary resistance could be transformed to almost any value of input resistance. Usually, the desired resistance would be made equal to the generator resistance, R_g, for maximum power transfer. It might also be selected to satisfy some

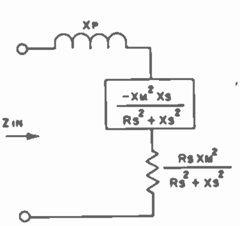


Fig. 7 — Equivalent single-mesh network of the two-mesh circuit of Fig. 6.

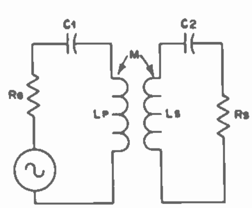


Fig. 8 — Double-tuned series circuits with magnetic coupling.

design goal, not necessarily related to maximum power transfer. This brings up a minor point but one that can cause considerable confusion. Normally, in transmitting circuits, the "unloaded Q" of the reactive components would be very high and the series parasitic resistances (discussed in a previous section) could be neglected. However, if it is not desired to do so, how should these resistances be taken into account? If maximum power transfer is the goal, the series resistance of the primary coil would be added to the generator resistance, R_g, and the transformed secondary resistance would be made equal to this sum.

On the other hand, a more common case requires the *total* input resistance to be equal to some desired value. For instance, an amplifier might provide optimum efficiency or harmonic suppression when terminated in a particular load resistance. Transmission lines also require a given load resistance in order to be "matched." In such cases, the series resistance of the primary coil would be *subtracted* from the actual resistance desired and the transformed resistance made equal to this difference. As an example, suppose an amplifier required a load resistance of 3000 ohms, and the primary-coil resistance was 100 ohms. Then, the transformed resistance must be equal to 2900 ohms. (In either case, the secondary coil resistance is merely added to the secondary load resistance and the sum substituted for R_s.)

Coefficient of Coupling

Although the equations for the input impedance can be solved in terms of the mutual reactance, the transforming mechanism involved becomes somewhat clearer if the coefficient of coupling is used instead. The coefficient of coupling, k, in terms of the corresponding reactances or inductances is

$$k = \frac{X_m}{\sqrt{X_p X_s}} = \frac{L_m}{\sqrt{L_p L_s}}$$

Then, the input resistance becomes

$$R_{in} = \frac{k^2 X_p X_s}{R_s}$$

The primary and secondary Qs are defined as

$$Q_p = \frac{X_p}{R_g}\ , \quad Q_s = \frac{X_s}{R_s}$$

where a "loaded" Q is assumed. This would mean R_s included any secondary-coil loss. For maximum-power transfer, R_g would be the total primary resistance which consists of the generator and coil resistance.

The coefficient of coupling under these conditions reduces to a rather simple formula

$$k_c = \frac{1}{\sqrt{Q_p Q_s}}$$

However, if it is desired to make the input resistance some particular value (as in the case of the previous example), the coefficient of coupling is then

$$k_c = \sqrt{\frac{R_{in} - R_p}{X_p Q_s}}$$

If the primary "loss" resistance is zero, both formulas are identical.

At values of k less than k_c, the input resistance is lower than either the prescribed value or for conditions of maximum power transfer. Higher values of k result in a higher input resistance. For this reason, k_c is called the critical coefficient of coupling. If k is less than k_c, the circuit is said to be undercoupled and for k greater than k_c, an overcoupled condition results. A plot of attenuation *vs* frequency for the three cases is shown in Fig. 9. Critical coupling gives the flatest response although greater bandwidth can be obtained by increasing k to approximately $1.5k_c$. At higher values, a pronounced dip occurs at the center or resonant frequency.

In the undercoupled case, a peak occurs at the resonant frequency of the primary and secondary circuit but the transformed resistance is too low and results in a mismatch. As the coupling is decreased still further, very little power is transferred to the secondary circuit and most of it is dissipated in the primary-loss and generator-source resistances. On the other hand, an interesting phenomenon occurs with the overcoupled case. It will be recalled that the transformed resistance is too high at resonance because the coefficient of coupling is greater than the critical value. However, a special case occurs if the primary and secondary circuits are identical which also means the transformed resistance, R_{in}, must equal R_s.

The behavior of the circuit under these conditions can be analyzed with the aid of Fig. 7. Assuming the Q of both circuits is high enough, the reactance, X_s, increases very rapidly on either side of resonance. If this variation is much greater than the variation of X_m with frequency, a frequency exists on each side of resonance where the ratio of $X_m{}^2$ and $R_s{}^2 + X_s{}^2$ is 1.0. Consequently, R_{in} is equal to R_s and the transformed reactance is $-X_s$. Since the primary and secondary resonators are identical, the reactances cancel because of the minus sign. The frequency plot for a k of 0.2 (k_c is 0.1) is shown in Fig. 9. If the primary and secondary circuits are not identical, a double-hump response still occurs but the points where the transformed resistance is equal to the desired value, the reactances are not the same numerically. Consequently, there is attenuation at peaks unlike the curve of Fig. 9.

Other Circuit Forms

While the coupled network shown in Fig. 8 is the easiest to analyze, it is not commonly encountered in actual circuits. As the resistance levels are increased, the corresponding reactances become very large also. In transmitting circuits, extremely

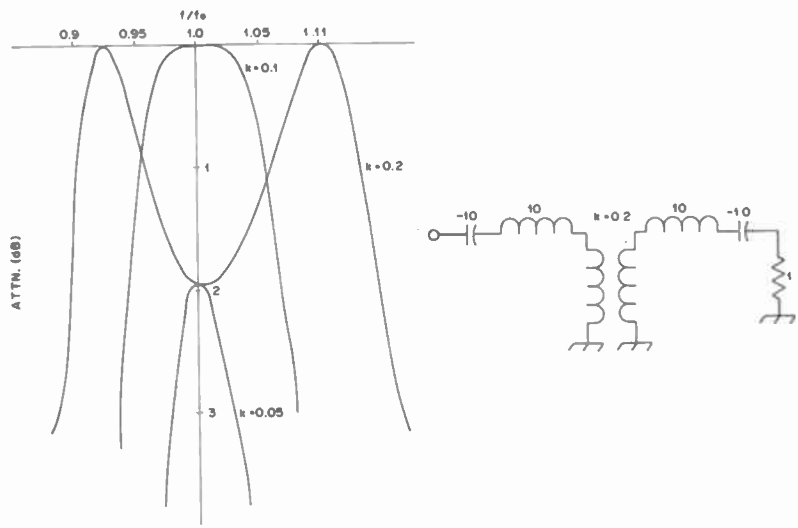


Fig. 9 — Response curves for various degrees of coupling coefficient k. The critical coefficient of coupling for the network shown in the inset is 0.1. Lower values give a single response peak (but less than maximum power transfer) while "tighter" coupling results in a double-peak response.

high voltages are then developed across the coils and capacitors. For high-impedance circuits, the circuit shown in Fig. 10 is often used. Although the frequency response is somewhat different than the circuit of Fig. 8 (in fact, the out-of-band rejection is greater), a matching network can be designed based upon the previous analysis for the series circuit.

This is accomplished by changing the parallel primary and secondary circuits to series equivalents. (It should be emphasized that this transformation is good at one frequency only.) The equivalent circuit of the one shown in Fig. 10 is illustrated in Fig. 11 where the new resistance and reactance of the secondary are given by

$$R_{eq}(S) = \frac{R_s}{1+\gamma^2}$$

$$X_{eq}(C_s) = \frac{-R_s\gamma}{1+\gamma^2}$$

$$\gamma = R_s/X_c$$

A similar set of transformations exists for the primary circuit also. In most instances, where one high-impedance load is matched to another one, R_s in Fig. 10 is much greater than the reactance of C_s and C_p. This simplifies the transformations and approximate relations are given by

$$R_{eq}(S) \cong X_c^2/R_s$$

$$X_{eq}(C_s) \cong X_c$$

As an example, suppose it was desired to match a 3-kΩ load to a 5-kΩ source using a coupled inductor with a 250-Ω (reactance) primary and secondary coil. Assume the coupling can be varied. Determine the circuit configuration and the critical coefficient of coupling.

Since the load and source resistance have a much higher numerical value than the reactance of the inductors, a parallel-tuned configuration must be used. In order to tune out the inductive reactance, the equivalent series capacitive reactance must be -250Ω. Since both R_s and R_p are known, the exact formulas could be solved for γ and R_{eq}. However, because the respective resistances are much greater than the reactance, the simplified approximate formulas can be used. This means the primary and secondary equivalent capaci-

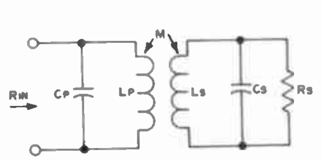


Fig. 10 — Coupled network with parallel-tuned circuits or "i-f" transformer.

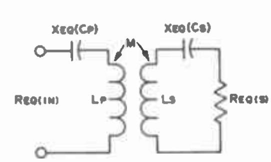


Fig. 11 — Equivalent series circuit of the parallel network shown in Fig. 10. This transformation is only valid at single frequencies and must be revalued if the frequency is changed.

tive reactances are -250Ω. The equivalent secondary resistance is $(250)^2/3000$ or $20.83\,\Omega$ resulting in a secondary Q of 250/20.83 or 12. (A formula could be derived directly for the Q from the approximate equations.) The equivalent primary resistance and Q are 12.5Ω and 20, respectively. Substituting the values for Q into the formula for the critical coefficient of coupling gives $1/\sqrt{(20)(12)}$ or .065.

Double-tuned coupled circuits of the type shown in Fig. 10 are widely used in radio circuits. Perhaps the most common example is the i-f transformer found in a-m and fm bc sets. Many communications receivers have similar transformers although the trend has been toward somewhat different circuits. Instead of achieving selectivity by means of i-f transformers (which may require a number of stages), a single filter with quartz-crystal resonators is used instead. (The subject of receivers will be taken up in greater detail in a later chapter.)

Single-Tuned Circuits

In the case of double-tuned circuits, separate capacitors are used to tune out the inductive components of the primary and secondary windings. However, examination of the equivalent circuit of the coupled coil shown in Fig. 7 suggests an alternative. Instead of a separate capacitor, why not "detune" a resonant circuit slightly and "reflect" a reactance of the proper sign into the primary in order to tune out the primary inductance. Since the transformation function (shown in the box in Fig. 7) reverses the sign of the secondary reactance, it is evident X_s must be *inductive* in order to tune out the primary inductance.

This might seem to be a strange result but it can be explained with the following reasoning. From a mathematical point of view, the choice of the algebraic sign of the transformed reactance is perfectly arbitrary. That is, a set of solutions to the equations governing the coupled circuit is possible assuming either a positive or negative sign for the transformed reactance. However, if the positive sign is chosen, the transformed resistance *would be negative.* But from a physical point of view, this is a violation of the conservation of energy since it would imply the secondary resistance acts as a source of energy rather than an energy "sink." Consequently, the solution with the negative resistance does not result in a physically realizable network.

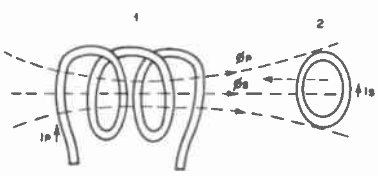


Fig. 12 — A coil coupled magnetically to a "shorted" turn provides insight to coils near solid shield walls.

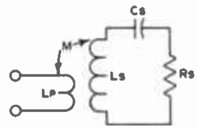


Fig. 13 — "Link" coupling can be used to analyze a number of important circuits.

The foregoing phenomenon has implications for circuits one might not normally expect to be related to coupled networks. For instance, consider coil 1 (Fig. 12) in proximity to the one-turn "shorted" coil 2. A time-varying current in coil 1 will induce a current in coil 2. In turn, the induced current will set up a magnetic field of its own. The question is will the induced field aid or oppose the primary field. Since the energy in a magnetic field is proportional to the square of the flux, the induced field must oppose the primary field, otherwise the principle of the conservation of energy would be violated as it was with the "negative" resistance. Consequently, the induced current must always be in a direction such that the induced field opposes changes in the generating field. This result is often referred to as Lenz's Law.

If, instead of a one-turn loop, a solid shield wall was substituted, a similar phenomenon would occur. Since the total flux (for a given current) would be less with the shield present than it would be in the absence of the shield, the equivalent coil inductance is decreased. That is why it is important to use a shield around a coil that is big enough to reduce the effect of such coupling. Also, a shield made from a metal with a high conductivity such as copper or aluminum is advisable, otherwise a loss resistance will be coupled into the coil as well.

Link Coupling

An example of a very important class of single-tuned circuits is shown in Fig. 13. The primary inductor consists of a small coil either in close proximity or wound over one end of a larger coil. Two resonators can be coupled in this manner although there may be considerable separation (and no mutual coupling between the larger coils) hence the term "link" coupling. While this particular method is seldom used nowadays, the term is still applied to the basic configuration shown in Fig. 13. Applications would be antenna-matching networks, output stages for amplifiers and, especially important, many circuits used at vhf that have no direct hf equivalent.

The cavity resonators used in repeater duplexers are one form of vhf circuit that uses link coupling. A cross-sectional view of a representative type is shown in Fig. 14. Instead of ordinary coils and capacitors, a section of coaxial transmission line comprises the resonant circuit. The frequency of the resonator may be varied by adjusting the tuning screw which changes the value of the capacitor. Energy is coupled into and out of the resonator by means of two small one-turn loops. Current in the input loop causes a magnetic field

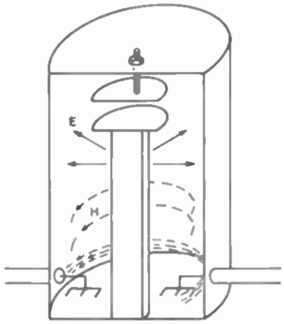


Fig. 14 – A vhf/uhf circuit which can be approximated by a link-coupled network using "conventional" components.

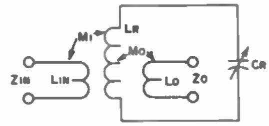


Fig. 15 – Equivalent low-frequency analog of the circuit shown in Fig. 14.

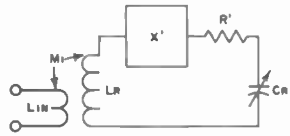


Fig. 16 – The network of Fig. 15 can be reduced with the transformation shown in Fig. 7.

(shown by dashed lines). If the frequency of the generating field is near one of the resonant "modes" of the configuration, an electric field will also be generated (shown by solid lines). Finally, energy may then be coupled out of the resonator by means of a second loop.

A low-frequency equivalent circuit of the resonator is shown in Fig. 15. However, the circuit can only be used to give an approximate idea of the actual frequency response of the cavity. At frequencies not close to the resonant frequency, the mathematical laws governing resonant circuits are different from those of "discrete" components used at hf. Over a limited frequency range, the resonator can be approximated by the series *LC* circuit shown in Fig. 15.

Applying the formulas for coupled networks shown in Fig. 7 to the two-link circuit of Fig. 15, the output link and load can be transformed to an equivalent series resistance and reactance as shown in Fig. 16. In most instances, the reactance, X_s, in the formula is just the reactance of the output link. Since the two-link network has been reduced to a single coupled circuit, the formulas can be applied again to find the input resistance and reactance.

Analysis of Single-Tuned Circuits

Single-tuned circuits are very easy to construct and adjust experimentally. If desired, the tuned circuit consisting of L_s, C_s, and perhaps the load, R_s, can be constructed first and tuned to the "natural" resonant frequency

$$f_o = \frac{1}{2\pi\sqrt{L_c C_s}}$$

Then, the primary inductor, which may be a link or a larger coil, is brought into proximity of the resonant circuit. The resonant frequency will usually shift upward. For instance, a coil and capacitor combination was tuned to resonance by means of a grid-dip oscillator (see the chapter on measurements) at a frequency of 1.8 MHz. When a two-turn link was wound over the coil, and coupled to the GDO the resonant frequency had increased to 1.9 MHz. A three-turn link caused a change to 2 MHz.

Quite often an actual load may be an unknown quantity, such as an antenna, and some insight into the effects of the various elements is helpful in predicting single-tuned circuit operation. Usually, as in the case of most matching networks, R_s (Fig. 7) and the input resistance are specified with the reactive components being the variables. Unfortunately, the variables in the case of mutually coupled networks are not independent of each other which complicates matters somewhat.

Examination of the equivalent circuit shown in Fig. 7 would indicate the first condition is that the reactance reflected from the secondary into the primary be sufficient to tune out the primary reactance. Otherwise, even though the proper resistance transformation is obtainable, a reactive component would always be present. A plot of the reflected reactance as a function of X_s is shown in Fig. 17. From mathematical considerations (which will not be discussed) it can be shown that the maximum and minimum of the curve have a value equal to $X_m{}^2/2R_s$. Consequently, this value must be greater than or equal to X_p in order that a value of X_s exists such that the reflected reactance will

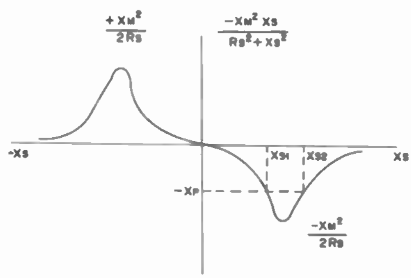


Fig. 17 – "Reflected" reactance into the primary of a single-tuned circuit places restraints on resistances that can be matched. This gives rise to a general rule that high-*Q* secondary circuits require a lower coefficient of coupling than low-*Q* ones.

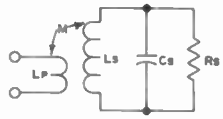


Fig. 18 — Single-tuned circuit with a parallel *RC* secondary.

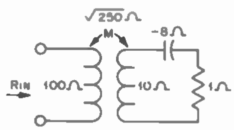


Fig. 19 — Text example.

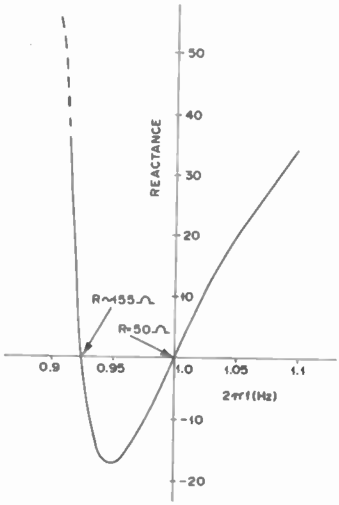


Fig. 21 — Input reactance of the network of Fig. 19. Note two "resonant" frequencies (where reactance is zero).

cancel X_p. In the usual case where $X_m^2/2R_s$ is greater than X_p, it is interesting to note that *two values* of X_s exist where X_p and the reflected reactance cancel. This means there are two cases where the input impedance is purely resistive and R_s could be matched to either one of two source resistances if so desired. The value of X_s at these points is designated as X_{s1} and X_{s2}.

On the other hand, a high value of R_s requires X_m to be large also. This could be accomplished by increasing the coefficient of coupling or by increasing the turns on the secondary coil. Increasing the turns on the primary also will cause X_m to be higher but X_p will increase also. This is somewhat self defeating since X_m^2 is proportional to X_p.

An alternate approach is to use the parallel configuration of Fig. 18. The approximate equivalent series resistance of the parallel combination is then $X(C_s)^2/R_s$ and the reactance is approximately $X(C_s)$. (See diagram and text for Fig. 11.) This approach is often used in multiband antenna systems. On some frequencies, the impedance at the input of the feed line is high so the circuit of Fig. 18 is employed. This is referred to as parallel tuning. If the impedance is very low, the circuit of Fig. 13 is used and is called series tuning.

As an example, suppose a single-tuned circuit is to be used to match a 1-Ω load to a 50-Ω source as shown in Fig. 19. It might be pointed out at this juncture that coupling networks using mutual magnetic coupling can be scaled in the same manner that filter networks are scaled (as discussed in Chapter 2). For instance, the circuit of Fig. 19 could be scaled in order to match a 50-Ω load to a 2500-Ω source merely by multiplying all the reactances by a factor of 50.

The input resistance and reactance of the circuit of Fig. 19 are plotted in Figs. 20 and 21, respectively. As pointed out earlier, there are two possible points where the reactance is zero and this circuit could be used to match the 1-Ω load to either a 50-Ω or 155-Ω source. Assuming a 50-Ω source was being used, the attenuation plot as a function of frequency would be given by the solid curve in Fig. 22.

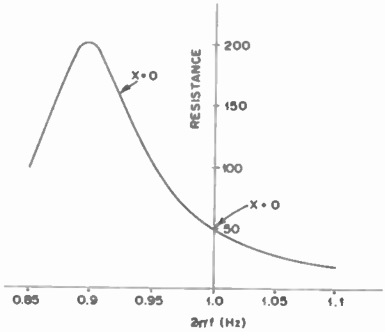


Fig. 20 — Input resistance of the circuit of Fig. 19 as a function of frequency.

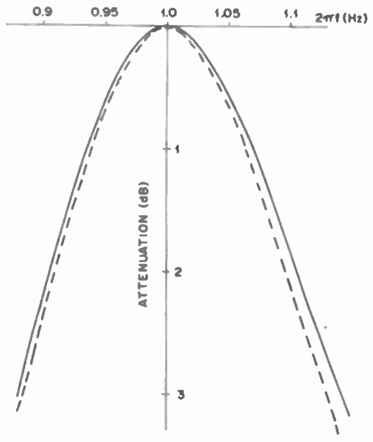


Fig. 22 — Response of the circuit shown in Fig. 19.

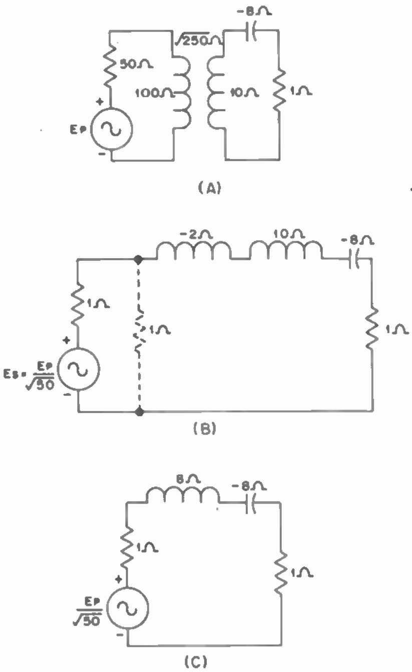


Fig. 23 – The transformation of Fig. 7 applied to the primary side of the circuit of Fig. 19.

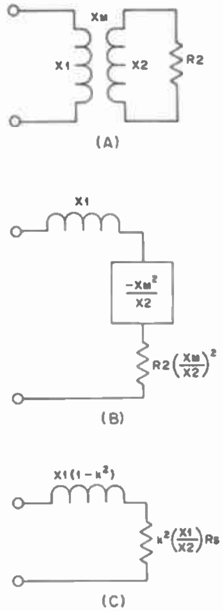


Fig. 24 – Equivalent-circuit approximation of two coupled coils.

With slight modification to include the effect of the source, the transformation of Fig. 7 can be applied to the primary side of the coupled circuit shown in Fig. 19. This is illustrated in Fig. 23. The complete circuit is shown at Fig. 23A and the network with the transformed primary resistance and reactance is shown in Fig. 23B.

In a lossless transformer, the maximum available power at the secondary must be the same as that of the original source on the primary side neglecting the effects of reactance. That is, the power delivered to a 1-Ω resistance (shown as a dashed line in Fig. 23B) must be the same as that delivered to a 50-Ω load in Fig. 23A. This assumes that the rest of the circuit has been disconnected in either case. In order to fulfill this requirement, the original source voltage must be multiplied by the square root of the ratio of the new and old source resistance.

The single-mesh transformed network is shown in Fig. 23C and it is interesting to compare the response of an *RLC* series circuit that actually possessed these element values at resonance with the circuit of Fig. 19. For comparison, the response of such a circuit is shown in Fig. 22 as a dashed curve and it can be seen that it differs only slightly from the coupled-circuit curve. The reason for the similarity is that even though the transformation of the primary resistance and reactance also changes with frequency, the effect is not that great in the present case.

BROADBAND RF TRANSFORMERS

The "sensitivity" of the frequency characteristic of the transformation shown in Fig. 7 depends mostly on the ratio of X_s to R_s. However, if X_s is much greater than R_s, the transformed reactance can be approximated by

$$\frac{-X_m{}^2 X_s}{R_s{}^2 + X_s{}^2} \cong \frac{-X_m{}^2}{X_s}$$

and the resistance becomes

$$\frac{R_s X_m{}^2}{R_s{}^2 + X_s{}^2} \cong R_s \frac{X_m{}^2}{X_s{}^2}$$

Applying this approximation to the general coupled circuit shown in Fig. 24A results in the transformed network of Fig. 24B. The coefficient of coupling for the circuit of Fig. 24A is

$$k = \frac{X_m}{\sqrt{X_1 X_2}}$$

and the network shown in Fig. 24B in terms of the coefficient of coupling is illustrated in Fig. 24C. For k equal to 1.0, the input reactance is zero and the input resistance is given by

$$R_{in} = \left(\frac{X_1}{X_2}\right) R_2 = \left(\frac{L_1}{L_2}\right) R_2 \cong \left(\frac{N_1}{N_2}\right)^2 R_2$$

where N_1 and N_2 are the number of turns on coil 1 and 2, respectively. From maximum-power transfer considerations, such as those discussed for the circuit of Fig. 23, the voltage transfer ratio becomes

$$e_2 = \frac{N_2}{N_1} e_1$$

It will be recalled that the foregoing equations occurred in the discussion of the "ideal transformer" approximation in Chapter 2. It was

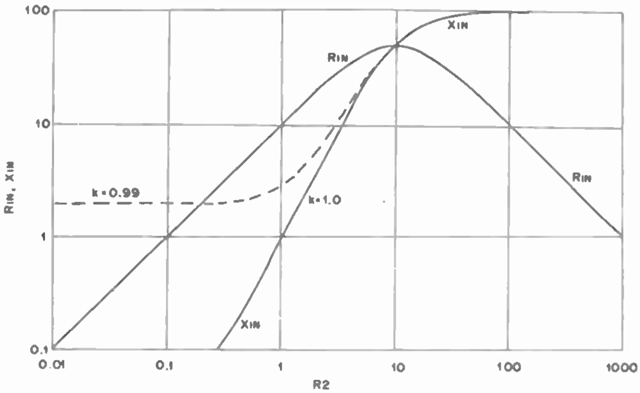


Fig. 25 — Input resistance and reactance as a function of output load resistance for X_1 and X_2 equal to 100Ω and 10Ω respectively (Fig. 24).

assumed then that the leakage reactance and magnetizing current were negligible. The effects on circuit operation of these variables are shown in Fig. 25. The curves were computed for various load resistances (R_s) using the exact equations shown in Fig. 7.

X_1 and X_2 are assumed to be 100 and 10Ω, respectively, with the solid curves for a k of 1.0 and the dashed reactance curve for k equal to 0.99 (the resistance curve for the latter value is the same as the one for k equal to 1.0). The ideal-transformer representation can be modified slightly to approximate the curve of Fig. 25 as shown in Fig. 26. The shunt reactance, X_{mag} is called the magnetizing reactance and X_L is referred to as the leakage reactance.

Unfortunately, the two reactances are not independent of each other. That is, attempts to change one reactance so that its effect is suppressed causes difficulties in eliminating the effects of the other reactance. For instance, increasing X_1, X_m, and X_2 will increase X_{mag} which is desirable. However, examination of Fig. 24C reveals that the coefficient of coupling, k, will have to be made closer to 1.0. Otherwise, the leakage reactance increases since it is proportional to X_1.

High-Permeability Cores

As a consequence of the interaction between the leakage reactance and the magnetizing reactance, transformers that approach ideal conditions are extremely difficult (if not impossible) to build using techniques common in air-wound or low-permeability construction. In order to build a network that will match one resistance level to another one over a wide range of frequencies, ideal-transformer conditions have to be approached quite closely. Otherwise, considerable inductive reactance will exist along with the resistive component as shown in Fig. 25.

One approach is to use a core with a higher permeability than air. Familiar examples would be power transformers and similar types common to the af range. However, when an inductor configuration contains materials of more than one permeability, the analysis relating to Fig. 24C has to be modified somewhat. The manner in which the core affects the circuit is a bit complicated although even a qualitative idea of how such transformers work is very useful.

First, consider the coupled coils shown in Fig. 27. For a given current, I_1, a number of "flux lines" are generated that link both coil 1 and coil 2. Note that in coil 1, not all of the flux lines are enclosed by all the turns. The inductance of a coil is equal to the ratio of the sum of flux lines linking each turn and the generating current or

$$L_1 = \frac{\Lambda_{TOTAL}}{I_1}$$

where for the example shown in Fig. 27, Λ_{TOTAL} is given by

$$\Lambda_{TOTAL} = \Lambda_1 + \Lambda_2 + \Lambda_3 + \Lambda_4 + \Lambda_5$$

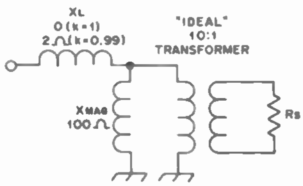


Fig. 26 — Approximate network for the curves of Fig. 25.

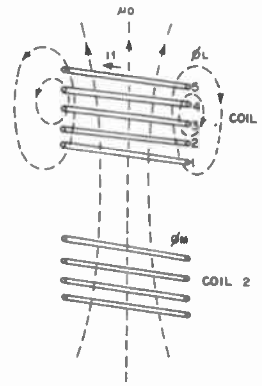


Fig. 27 — Coupled coils showing magnetic flux lines.

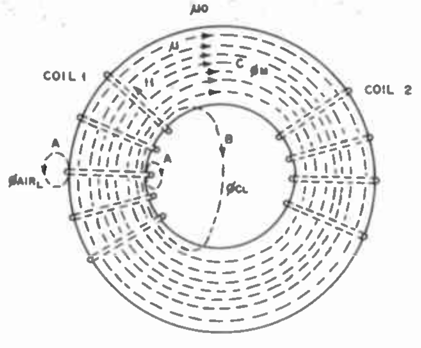


Fig. 28 – Toroidal Transformer.

Counting up the number of flux linkages in coil 1 gives

$$\Lambda_{TOTAL} = 5 + 5 + 7 + 7 + 5 = 29$$

If all the flux lines linked all the turns, Λ_{TOTAL} would be 35 so L_1 is 29/35 or 83 percent of its maximum possible value. Likewise, if all the flux (7 lines) generated in coil one linked all the turns of coil 2, the maximum number of flux linkages would be the number of turns on coil 2 times 7 or 28. Since only three lines link coil 2, the mutual inductance is 3 × 4/28 or 43 percent of maximum.

Assuming both coils are "perfect," if a current I_1 produced 7 flux lines in a five-turn coil, then the same current in a four-turn coil would produce (4/5)(7) flux lines, since the flux is proportional to the magnetizing current times the number of turns. Consequently, the maximum flux linkages in coil 1 from a current of the same value as I_1 but in coil 2 instead would be (4/5)(7)(5) or 28. Therefore, it can be seen that the mutual inductance is independent of the choice of coil used for the primary or secondary. That is, a voltage produced in one coil by a current in the other one would be the same if the coils were merely interchanged. (This result has been used implicitly on a number of previous occasions without proof.) In addition, the maximum flux linkages in coil 2 produced by a current, I_1, would be (4/5)(7)(4). As an exercise, substitute the maximum inductance values into the formula for the coefficient of coupling and show that k is 1.0.

The next step is to consider the effect of winding coils on a form with a magnetic permeability much higher than that of air. An example is illustrated in Fig. 28 and the configuration shown is called a toroidal transformer. Since the flux is proportional to the product of the permeability and the magnetizing current, the flux in the core shown in Fig. 28 will be much greater than the coil configuration of Fig. 27. However, not all of the flux is confined to the core. As can be seen in Fig. 28, some of the flux lines never penetrate the core (see lines marked a in Fig. 28) while others enclose all the windings of coil 1 but not coil 2 (see line marked b). The significance of these effects is as follows. The total flux linkage produced by the current, I_1, is

$$\Lambda_{TOTAL} = \Lambda_{air} + \Lambda_{core}$$

and dividing both sides of the equation by I_1 gives

$$L_T = L_{air} + L_{core}$$

Consequently, the circuit of Fig. 24 can be represented as shown in Fig. 29A. For X_2 much greater than the load resistance, the approximate network of Fig. 29B can replace the one of Fig. 29A.

At first sight, it might seem as though little advantage has been gained by introducing the core since the formulas are much the same as those of Fig. 24C. However, the reactances associated with the core can be made very high by using a material with a high permeability. Also, even though there may be some "leakage" from the core as indicated by line b in Fig. 28, it is ordinarily low and the coefficient of coupling in the core can be considered 1.0 for all practical purposes. This is especially true at af and power frequencies with transformers using iron cores where the permeability is extremely high. This means the magnetizing reactance can be made very high without increasing the leakage reactance accordingly as is the case with the circuit in Fig. 24C. Therefore, ideal transformer conditions are considered to exist in the core and the final circuit can be approximated by the one shown in Fig. 29C.

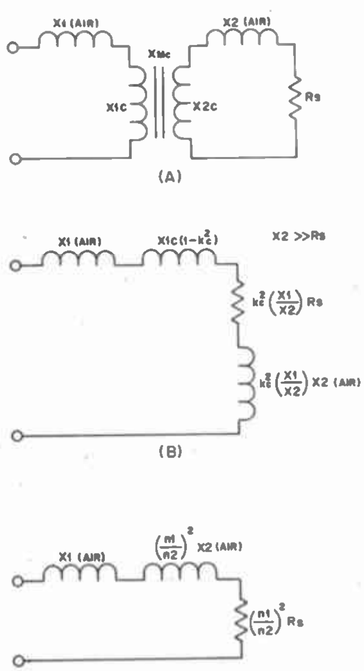


Fig. 29 – Effect of a high-permeability core on transformer equivalent circuit.

Bifilar and Twisted-Pair Windings

Although the core helps alleviate some of the problems with leakage and magnetizing reactance, the residual parasitic elements must still be made as low as possible. This is especially important in matching applications as the following example illustrates. A transformer has a primary and secondary leakage reactance of 1Ω and 0.1Ω, respectively, with a coefficient of coupling of 1.0 in the core. X_1 and X_2 are 1000Ω and 100Ω.

A plot similar to the one of Fig. 25 is shown in Fig. 30 along with a curve for voltage-standing-wave ratio (*VSWR*). These results are based on the exact equations and it can be seen that the approximate relations shown in Fig. 29C are valid up to 1Ω or so. Curve *A* (Fig. 30) only includes the effect of the secondary reactance and illustrates the manner in which the reactance is transformed. Curve *B* is the total input reactance which merely requires the addition of 1Ω. The *VSWR* curve includes the effect of the latter. Useful range of the transformer is between 1 and 10Ω with rapid deterioration in *VSWR* outside of these values. (The *VSWR* curve is for a characteristic impedance equal to 10 times the secondary resistance. For instance, the transformer would be useful in matching a 5-Ω load to a 50-Ω line.)

As mentioned previously, these difficulties are less pronounced at audio frequencies since the permeabilities normally encountered in iron-core transformers are so high, the actual inductance of the winding itself is small in comparison to the component represented by the core. That is, a small number of turns of wire wound on a core may actually be the equivalent of a very large coil. However, materials suitable for rf applications have much lower permeabilities and a narrower range of matching values is likely to be the result (such as in the example of Fig. 30). Therefore, other means are required in keeping the parasitic elements as low as possible. Either that, or less conventional transformer designs are used.

One approach is shown in Fig. 31. Instead of separating the windings on the core as shown in Fig. 28, they are wound in parallel fashion. This is called a bifilar winding although a more common approach to achieve the same purpose is to twist the wires together. Either way, there are a number of advantages (and some disadvantages) to be gained. Referring to Fig. 27, the fact that not all the flux lines linked all of the turns of a particular coil meant the self inductance was lower than if all the turns were linked. Since the separation between turns of a particular coil is quite large in the configuration of Fig. 31, the flux linkage between turns is quite low. This means the corresponding leakage inductance is reduced accordingly. However, the coupling between both coils is increased because of the bifilar winding (flux line *A*) in Fig. 31) which also tends to reduce the leakage inductance of either coil.

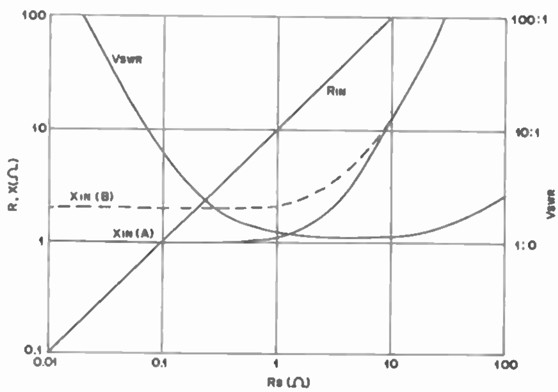


Fig. 30 — Curve for transformer problem discussed in the text.

On the other hand, the capacitance between windings is increased considerably as indicated by *B* in Fig. 31. As a result, the coupling between windings is both electrical and magnetic in nature. Generally speaking, analysis of the problem is quite complicated. However, a phenomenon usually associated with such coupling is that it tends to be directional. That is, energy transferred from one winding to another one propagates in a preferred direction rather than splitting equally.

Directional Coupling

Two conductors are oriented side by side over a conducting plane as shown in Fig. 32. A current *I* in conductor 1 will induce a current I_m in conductor 2 because of magnetic coupling. The actual value of the current will depend upon the external circuitry attached to the conductors but it

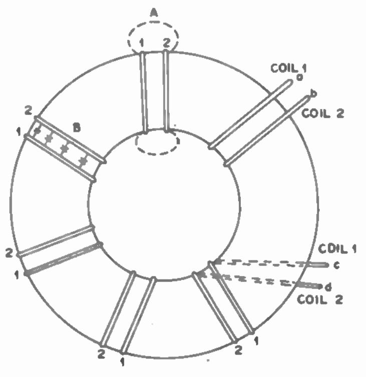


Fig. 31 — Bifilar-wound transformer on toroidal core.

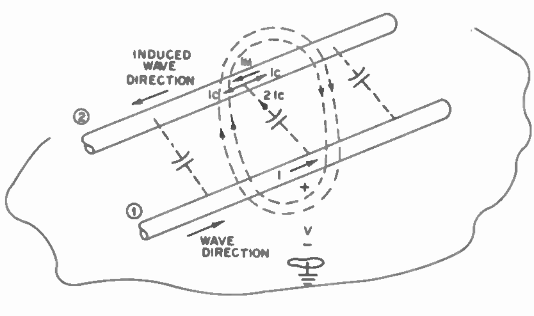


Fig. 32 — Effect of distributed capacitance on transformer action.

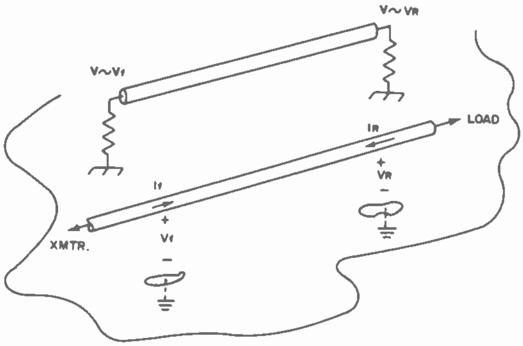


Fig. 33 – Basic configuration for a directional-coupler type *VSWR* detector.

will be assumed that the two of them extend to infinity in both directions.

Since capacitive coupling exists also, a second set of current components denoted by I_c will also flow. The result is that a wave traveling toward the *right* in conductor 1 will produce a wave traveling toward the *left* in conductor 2. Such coupling is called contradirectional coupling since the induced wave travels in the opposite direction to the generating wave.

This is the principle behind many practical devices and ones that are quite common in amateur applications. In adjusting a load such as an antenna, it is desirable to insure that energy is not reflected back to the transmitter. Otherwise, the impedance presented to the transmitter output may not be within range of permissable values. A directional coupler is useful in determining how much power is reflected as indicated in Fig. 33. Energy originating from the transmitter and flowing to the right causes a voltage to be produced across the resistor at the left. On the other hand, a wave traveling from the right to the left produces a voltage across the right-hand resistor. If both of these voltages are sampled, some idea of the amount of power reflected can be determined. (The subject of reflected-power is taken up in more detail in the chapter on transmission lines.)

In some situations, the coupling described can be very undesirable. For instance, the lines shown in Fig. 33 might be conductors on a circuit board in a piece of equipment. As a result, the coupling between lines can cause "feedback" and because of its directional nature, it can be very difficult to suppress with conventional methods. Therefore, it is good design practice to use "doubled-sided" board (board with conductive foil on both sides) so that a ground plane of metal is in close proximity to the conductors. This tends to confine the fields to the region in the immediate vicinity of wires.

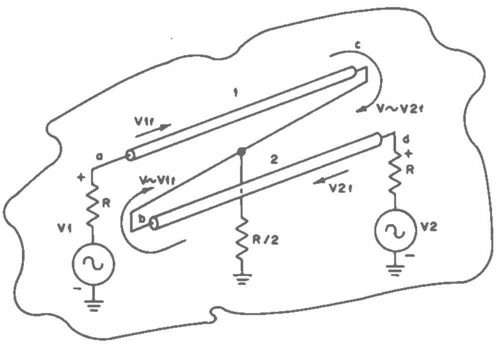


Fig. 34 – Directional-coupler hybrid combiner.

Transmission-Line Transformers

In effect, sections of transmission line in close proximity act as transformers with the unique feature that the coupling is directional. For instance, if only magnetic coupling was present in the configuration of Fig. 33, power would be divided equally between the resistors at either end of the "secondary" section of transmission line. As another example of directional effects, the network shown in Fig. 34 can be used to couple two sources to a common load without "cross coupling" of power from one source to the other one. (This assumes the sources have the same frequency and phase. Otherwise, a resistance of value $2R$ must be connected from points *a* to *d*.) Such a configuration is called a hybrid combiner and is often used to combine the outputs of two solid-state amplifiers in order to increase the power-handling capability. This permits the use of less expensive low-power devices rather than very expensive high-power ones. Even though more devices are required, it is still simpler since the difficulties in producing a high-power transistor increase in a greater proportion as the power level is raised.

The manner in which the circuit shown in Fig. 34 operates is as follows. A wave from the generator on the left end of line 1 travels toward the right and induces a wave in line 2 that travels toward the left and on into the load. No wave is induced in line 2 that travels toward the right except for a small fraction of power.

A similar situation exists with the second generator connected at the right end of line 2. A wave is induced in line 1 that travels toward the right. Since the load is also connected to the right end of line 1, power in the induced wave will be dissipated here with little energy reaching the generator at the left end of line 1. In order to "simulate" a single load (since there are two generators involved), the value of the load resistance must be half of the generator resistance. Assuming that two separate resistors of value R were connected to the ends of the line, it would be possible to connect them together without affecting circuit operation. This is because the voltage across both resistors is of the same phase and amplitude. Consequently, no additional current would flow if the two resistors were paralleled or combined into a single resistor of $R/2$.

Extending the Low-Frequency Range

As might be expected, the coupling mechanism illustrated in Figs. 32 through 34 is highly dependent on dimensions such as conductor spacing and

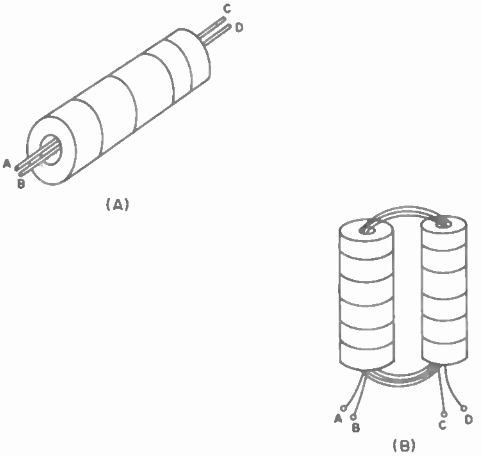


Fig. 35 — Transmission-line transformers with ferrite cores.

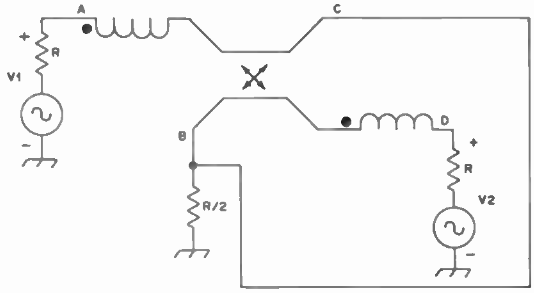


Fig. 36 — Equivalent circuit of transmission-line transformer in the presence of the core. Dots indicate winding sense of coils. A positive current into a dotted end of one coil will produce a voltage in the other coil because of mutual coupling. The polarity of this voltage will be such that dotted end of the "secondary" coil will be positive. (See text for crossed-arrow symbol in the middle of the parallel lines.)

line length. For instance, maximum coupling of power from the primary wave to the induced wave occurs when the "secondary" line is a quarter-wavelength long* or some odd multiple of a quarter wavelength. This would normally make such couplers impractical for frequencies in the hf range. However, by running the leads through a ferrite core as shown in Fig. 35, lower-frequency operation is possible. Although the transformer of Fig. 35A is seldom used, it illustrates the manner in which the conductors are employed electrically in the more complicated configurations of Fig. 31 and Fig. 35B. Also, the relationship between the parallel-line coupler in Fig. 34 and the "loaded" version of Fig. 35A is easier to visualize.

Recalling an earlier problem discussion (Fig. 28), a set of coupled coils wound on a high-permeability core can be broken down into combinations of two series inductances. One inductance represents the path in air while the other one includes the effects of the flux in the core. As before, it is assumed that the coefficient of coupling in the core is 1.0.

If the hybrid combiner of Fig. 34 is wound on a core (such as those of Fig. 31 or Fig. 35), the low-frequency range of the entire system is increased considerably. The equivalent circuit showing the effect of the core on the air-wound coupler is illustrated in Fig. 36. (The symbol in the middle of the parallel lines is the standard one for a directional coupler.) At the higher frequencies, most core materials decrease in permeability so the operation approaches that of the original air-wound coupler and the inductance produced by the core can be neglected. At the low end of the frequency range, the line lengths are usually too short to provide much coupling or isolation. Therefore, the circuit can be represented by the set of coupled coils shown in Figs. 37 and 38.

For a current I_{12} flowing from source 1 over to the mesh that includes source 2, the mutual-reactance components *add* to the self inductance of each coil. Consequently, a large reactance appears in series between the two sources which effectively isolates them. On the other hand, currents from both sources that flow through the load resistor $R/2$ produce fluxes that cancel and the voltages produced by the self- and mutual-reactance terms *subtract*. If both sources have the same amplitude and phase, currents I_1 and I_2 must be identical because of the symmetry involved. However, if the coefficient of coupling is 1.0, the

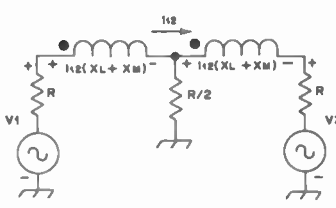


Fig. 37 — Low-frequency equivalent circuit of hybrid combiner showing isolation of sources.

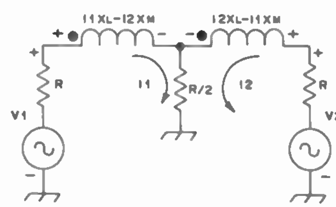


Fig. 38 — Desired coupling mode of hybrid combiner.

*Oliver, "Directional Electromagnetic Couplers," *Proceedings of the I.R.E.*, Vol. 42, p. 1686-1692; November, 1954.

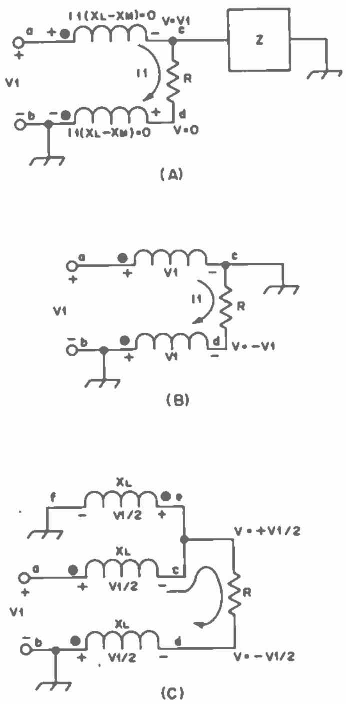


Fig. 39 — Other applications of transmission-line transformers.

self and mutual reactance must be equal. Therefore, the voltage across either coil is zero since the terms subtract and a low-impedance path exists between both sources and the load.

Other Transformer Types†

The hybrid combiner is only one application of a combination transmission-line or directional-coupler transformer and conventional coupled-coil arrangement. With other variations, the low-frequency isolation is accomplished in the same manner. Mutual-reactance terms add to the self reactance to provide isolation for some purpose with cancellation of reactive components in the path for the desired coupling. Very good bandwidth is possible with a range from bc frequencies to uhf in the more esoteric designs. Models that cover all the amateur hf bands can be constructed easily.

Unfortunately, there is also a tendency to expect too much from such devices on occasion. Misapplication or poor design often results in inferior performance. For instance, as indicated in an earlier example (Fig. 30), actual impedance levels were important along with the desired transforming ratio. Using a transformer for an impedance level that it was not intended for resulted in undesirable reactive components and improper transforming ratio. However, when applied properly, the transformers discussed in the previous sections can provide bandwidth characteristics that are obtainable in no other way.

Another transformer type is shown in Fig. 39A. The windings of the coils are such that the voltages across the inductors caused by the desired current are zero. This is because the induced voltages produced by the current in the mutual-reactance terms just cancel the voltage drop caused by the current flowing in the self reactances of either coil. (Assuming that the coefficient of coupling is 1.0.) However, an impedance connected to ground at point *c* would be in series with the self reactance (X_L) of the coil connected between points *a* and *c*. But there would be no induced voltage to counter the voltage drop across this coil. Therefore, if X_L is large, very little current would flow in the impedance Z and it would effectively be isolated from the source.

In fact, terminal *c* could be grounded as shown in Fig. 39B. The voltage drop across the coil from *a* to *c* would then be equal to V_1. However, the induced voltage in the coil connected between points *b* and *d* would also be V_1 assuming unity coupling (k equal to 1.0). Although the voltage drop produced by the inductors around the mesh through which I_1 flows is still zero, point *d* is now at potential $-V_1$ and a phase reversal has taken place. For this reason, the configuration shown in Fig. 39B is called a phase-reversal transformer.

Baluns

The circuit shown in Fig. 39A is useful in isolating a load from a grounded source. This is often required in many applications and the device that accomplishes this goal is called a balun (*bal*anced to *un*balanced) transformer. Baluns may also be used in impedance transforming applications along with the function of isolation and a "1:1 balun" such as the one shown in Fig. 39A means the impedance at the input terminals *ab* will be the same as the load connected across terminals *cd*. Other transforming ratios are possible such as 4:1 with the appropriate circuit connections.

One disadvantage of the network of Fig. 39A is that although the load is isolated from the source, the voltages at the output are not balanced. This is important in some applications such as diode-ring mixers where a "push-pull" input is required and so the circuit of Fig. 39C is used. A third coil connected between points *e* and *f* is wound on the same core as the original transformer (Fig. 39A). This coil is connected so that a voltage across it produces a flux that adds to that produced by the coil between *a* and *c*. Assuming that both coils are identical, the voltage drop across either one must be the same or half the applied voltage. However, since the coil between *b* and *d* is also coupled to this combination (and is an identical coil), the induced voltage must also be $V_1/2$. Consequently, the end of the load connected to points *c* and *e* is at a potential of $+V_1/2$ with respect to ground while point *d* is $-V_1/2$ with respect to ground when the input voltage has the polarity shown. Therefore, this circuit not only isolates the load

†Ruthroff, "Some Broad-Band Transformers," *Proceedings of the I.R.E.*, Vol. 97, pp. 1337-1342; August, 1959.

from the source but provides a balanced voltage also.

Either the circuit of Fig. 39A or Fig. 39C can be used if only isolation is desired. However, the network shown in Fig. 39C is more difficult to design and construct since the reactance of the coils between points *a* and *f* must be very high throughout the frequency range of the transformer. With both transformers, the coefficient of coupling must also be very close to 1.0 in order to prevent undesirable reactance in series with the load. This problem can be offset somewhat by reducing X_L slightly (by using fewer turns) but this is counter to the requirement of large X_L in the circuit of Fig. 39C. Isolation is reduced in both cases although no detrimental effect on input impedance results in the transformer of Fig. 39A by reducing X_L.

NONLINEAR AND ACTIVE NETWORKS

Almost all the theory in previous sections has dealt with so-called passive components. Passive networks and components can be represented solely by combinations of resistors, capacitors and inductors. As a consequence, the power output at one set of terminals in a passive network cannot exceed the total power input from sources connected to other terminals in the circuit. This assumes all the sources are at one frequency. Similar considerations hold true for any network, however, it is possible for energy to be converted from one frequency (including dc) to other ones. While the total power input must still equal the total power output, it is convenient to consider certain elements as controllable sources of power. Such devices are called amplifiers and are part of a more general class of circuits called active networks. An active network generally possesses characteristics that are different than those of simple *RLC* circuits although the goal in many instances is to attempt to represent them in terms of passive elements and generators.

NONLINEARITY

Two other important attributes of passive *RLC* elements are that they are linear and bilateral. A two-terminal element such as a resistor is said to be bilateral since it doesn't matter which way it is connected in a circuit. Semiconductor and vacuum-tube devices such as triodes, diodes, transistors and integrated circuits (ICs) are all examples where the concept of a bilateral element breaks down. (For readers with limited backgrounds in the basic operation of vacuum tubes, recommended study would be *The Radio Amateur's License Manual* and *Understanding Amateur Radio*. Both publications contain fundamental treatments of vacuum-tube principles and are available from The American Radio Relay League.) The manner in which the device is connected in a circuit and the polarity of the voltages involved are very important.

An implication of the failure to satisfy the bilateral requirements is that such devices are nonlinear in the strictest sense. Linearity means that the amplitude of a voltage or current is related to other voltages and currents in a circuit by a single proportionality constant. For instance, if all the voltages and currents in a circuit were doubled, a single remaining voltage or current would be doubled also. That is, it couldn't change by a factor of 1/2 or 3 no matter how complex the network might be. Likewise, if all the polarities of the currents and voltages in a circuit are reversed, the polarity of a remaining voltage or current must be reversed also. Finally, if all the generators or sources in a linear network are sine waves at a single frequency, any voltage or current produced by these sources must also be a sine wave at the same frequency too.

Consequently, if a device is sensitive to the polarity of the voltage applied to its terminals, it doesn't meet the requirements of a bilateral element or a linear one either. However, because of the extreme simplicity of the mathematics of linear circuits as compared to the general nonlinear case, there is tremendous motivation in being able to represent a nonlinear circuit by a linear approximation. Many devices exhibit linear properties over part of their operating range or may satisfy some but not all of the requirements of linear circuits. Such devices in these categories are sometimes termed piece-wise linear. Either that, or they are just referred to simply as linear. For instance, a linear mixer doesn't satisfy the rule that a voltage or current must be at the same frequency as the generating source(s). However, since the desired output voltage (or current) varies in direct proportion to the input voltage (or current), the term linear is applied to distinguish the mixer from types without this "quasi-linear" property.

HARMONIC-FREQUENCY GENERATION

In a circuit with only linear components, the only frequencies present are those generated by the sources themselves. However, this is not true with nonlinear elements. One of the properties of nonlinear networks mentioned earlier is that energy at one frequency (including dc) may become converted to other frequencies. In effect, this is how devices such as transistors and vacuum tubes are able to amplify radio signals. Energy from the dc power supply is converted to energy at the desired signal frequency. Therefore, a greater amount of signal power is available at the output of the network of an active device than at the input.

On the other hand, such frequency generation may be undesirable. For instance, the output of a transmitter may have energy at frequencies that could cause interference to nearby receiving equipment. Filters and similar devices must be used to suppress this energy as much as possible.

The manner in which this energy is produced is

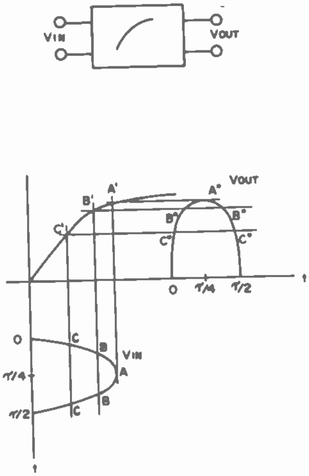


Fig. 40 – Nonlinear transfer characteristic (see text discussion).

shown in Fig. 40. A sine-wave at the input of a nonlinear network (V_{in}) is "transformed" into the output voltage waveform (V_{out}) illustrated. If the actual device characteristic is known, the waveform could be constructed graphically. It could also be tabulated if the output voltage as a function of input voltage was available in either tabular or equation form. (Only one-half of the period of a sine wave is shown in Fig. 40 for clarity.)

Although the new waveform retains many of the characteristics of the original sine wave, some transformations have taken place. It has zero value when t is either 0 or T/2 and attains a maximum at T/4. However, the fact that the curve is flattened somewhat means energy at the original sine-wave frequency has been converted to other frequencies. It will be recalled that the sum of a number of sine waves at one frequency result in another sine wave at the same frequency. Therefore, it must be concluded that the waveform of Fig. 40 has more than one frequency component present since it is no longer a sine wave.

One possible "model" for the new waveform is shown in Fig. 41A. Instead of one sine wave at a single frequency, there are two generators in series with one generator at three times the "fundamental" frequency, ω where ω is $2\pi f$ (Hz). If the two sine waves are plotted point by point, the dashed curve of Fig. 41B results. While this curve doesn't resemble the one of Fig. 40 very closely, the general symmetry is the same. It would take an infinite number of generators to represent the desired curve exactly, but it is evident all the frequencies must be odd multiples of the fundamental. Even multiples would produce a lopsided curve which might be useful for representing other types of waveforms.

In either case, the multiples have a specific name and are called harmonics. There is no "first" harmonic (by definition) with the second, third and fourth multiples designated as the second, third and fourth harmonics. Thus the dashed curve of Fig. 41 is the sum of the fundamental and third harmonic.

Analyzing waveforms such as those of Fig. 40 is a very important subject. A plot of harmonic amplitude such as that shown in Fig. 41C is called the spectrum of the waveform and can be displayed on an instrument called a spectrum analyzer. If the mathematical equation or other data for the curve is known, the harmonics can also be determined by means of a process called Fourier Analysis.

Linear Approximations of Nonlinear Devices

Nonlinear circuits may have to be analyzed graphically as in the previous example. There are many other instances where only a graphical method may be practical such as in power-amplifier problems. However, a wide variety of applications permit a different approach. A model is derived from the nonlinear characteristics using linear elements to approximate the more difficult nonlinear problem. This model is then used in more complicated networks instead of the non-linear characteristics which simplifies analysis considerably.

The following example illustrates how this is accomplished and although a vacuum-tube application is considered, a similar process is employed in solving semiconductor problems as well. However,

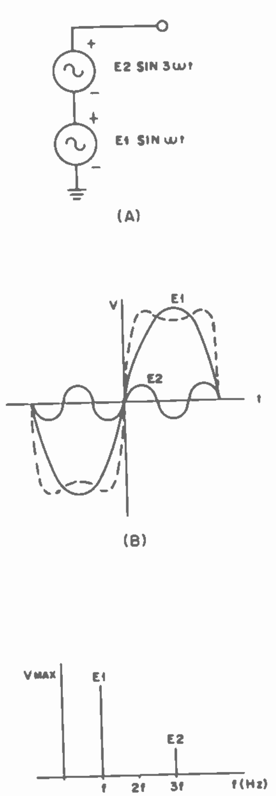


Fig. 41 – Harmonic analysis and spectrum.

there are some additional factors involved in semiconductor design that do not apply to vacuum tubes. Device characteristics of early transistors were less uniform than those of transistors although this is much less of a problem than it was formerly. In fact, much of the analysis required with vacuum tubes is unnecessary with modern solid-state components since many of the problems have already been "solved" before the device leaves the counter at the radio store. That is, amplifiers such as those in integrated circuits have the peripheral elements built in and there is no need to determine the gain or other parameters such as the values of bias resistors.

THE TRIODE AMPLIFIER

A simple network using a triode vacuum type is shown in Fig. 42A and a typical set of characteristic curves is illustrated in Fig. 43A. The first chore in finding a suitable linear approximation for the triode is to determine an optimum operating point. Generally speaking, a point in the center of the set of curves is desirable and is indicated by point Q in Fig. 43A. (Other areas are often picked for power-amplifier operation but the goal here is to find a point where the maximum voltage swing is possible without entering regions where the nonlinearities affect the linear approximation.)

In the particular operating point chosen, the cathode-to-grid voltage is −3, the cathode-to-plate voltage is 280, and the plate current is 10 mA. It is assumed that the input-signal source in Fig. 42A is a "short circuit" at dc and a 3-V battery connected as shown results in a dc voltage of −3 being applied to the grid at all times. Such a battery is called a bias battery or bias supply.

The next step is to determine how the plate voltage varies with grid voltage (e_g) for a constant plate current. Assuming that the characteristic curves were completely linear, this would permit evaluation of an equivalent ac voltage generator as shown in Fig. 42B. For a constant plate current of 10 mA, the plate voltage changes from 325 (point b) to 230 (point a) when the grid voltage is changed from −4 to −2 (Fig. 43A).

These numbers can be used to compute the amplification factor (μ) of the triode which is

$$\mu = \frac{325 - 230}{(-4) - (-2)} = -47.5$$

Quite often, a set of characteristics will not be published for a triode and only the amplification factor will be given along with a typical operating point. However, note that the amplification factor is ***negative***. This means that for an increase in the signal voltage (e_{in}), the controlled generator decreases in voltage. Consequently, there is a 180-

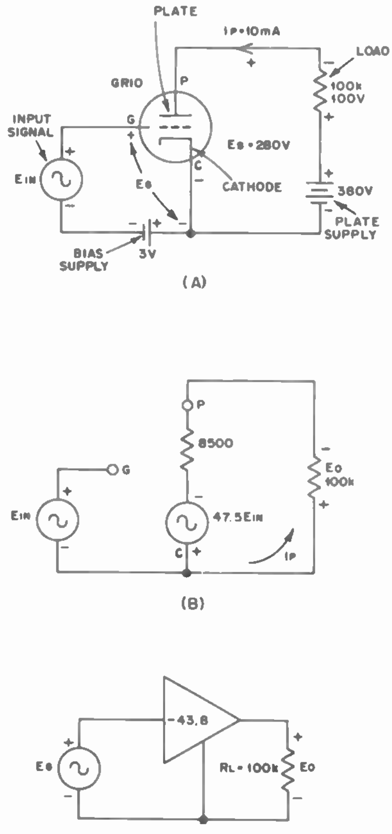


Fig. 42 — Basic triode amplifier and equivalent circuit.

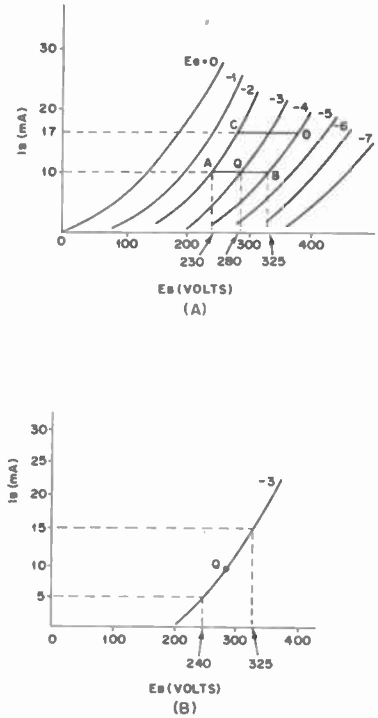


Fig. 43 — Triode characteristics and derivation of small-signal parameters.

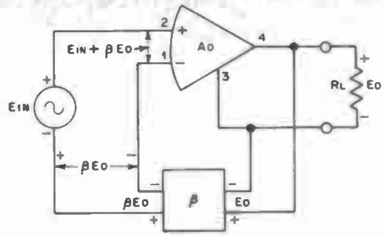


Fig. 44 — Network illustrating voltage feedback.

degree phase shift between the input voltage and the controlled source. (Note the polarity of the generator shown in Fig. 42B.)

In order to complete the equivalent generator circuit, the source "impedance" must be computed. This is accomplished by determining how the plate voltage varies with plate current at constant grid voltage as shown in Fig. 43B. The plate resistance is then

$$r_p = \frac{325 - 240}{(15 - 5) \times 10^{-3}} = 8500\Omega$$

which must be considered to be in series with the controlled source of Fig. 42B.

It should be pointed out at this juncture that the reasoning *why* the foregoing procedure is valid has not been presented. That is, why was the amplification factor defined as the ratio of a change in plate voltage to change in grid voltage at constant current? Unfortunately, the mathematics involved although not difficult is somewhat sophisticated. Some knowledge of the subject of partial differential equations is required for the theoretical derivation of these parameters. However, an intuitive idea can be obtained from the following.

If the characteristics were completely linear, instead of being nonlinear as shown, the equivalent generator would be unaffected by *changes* in plate current but only by changes in grid voltage. For instance, if the plate current was increased from 10 to 17 mA (Fig. 43A), the amplification factor would be the equivalent of the change in voltage represented by the line *cd* divided by −2. However, since the length of *cd* is almost the same as that of *ab* (the difference in plate voltage for a −2-V change at 10 mA), it can be concluded μ doesn't change very much. Not at least in the center region of the characteristics.

Similar considerations hold for the plate resistance, r_p. It wouldn't matter if the curve for −4 or −2 V was picked (Fig. 43B), since the change in plate voltage *vs* plate current would be approximately the same. Entities such as μ and r_p are often called *incremental* or *small-signal* parameters. This means they are valid for small ac voltages or currents around some operating point but less so for large variations in signal or for regions removed from the specified operating point. Also, such parameters are not closely related to *dc voltage* characteristics. For instance, a "static" plate resistance could be defined as the ratio of plate voltage to plate current. For the −3-V operating point chosen, the static plate resistance would be 280 divided by 10×10^{-3} or 280 kΩ. This is considerably different from the small-signal plate resistance determined previously which was 8500 Ω.

Amplifier Gain

The ratio of the variation in voltage across the load resistance to change in input voltage is defined as the gain of the amplifier. For the equivalent circuit shown in Fig. 42B, this ratio would be

$$A = \frac{e_o}{e_{in}}$$

In order to solve for the gain, the first step is to determine the incremental plate current. This is just the source voltage divided by the total resistance of the circuit mesh or

$$i_p = \frac{47.5 e_{in}}{100 + 8.5}\ mA$$

The output voltage is then

$$e_o = i_p\ 100$$

and combining the two foregoing equations gives

$$A = \frac{e_o}{e_{in}} = \frac{(47.5)(100)}{100 + 8.5} = 43.8$$

It is somewhat inconvenient to have the input and output voltages defined with opposite polarities as shown in Fig. 42B. Therefore, the gain becomes negative as illustrated in the triangle in Fig. 42C. A triangle is the standard way of representing an amplifier stage in "block-diagram" form. The amplifier gain depends of course on the load resistance, R_L, and a general formula for the gain of the circuit of Fig. 42B is

$$A = \frac{-\mu R_L}{r_p + R_L}$$

FEEDBACK

Being able to eliminate the equivalent circuit and use only one parameter such as the gain permits analysis of more complicated networks. A very important application occurs when part of the output energy of an amplifier is returned to the input circuit and gets amplified again. Since energy is being "fed back" into the input, the general phenomenon is called feedback. The manner in which feedback problems are analyzed is illustrated in Fig. 44. The output voltage is "sampled" by a network in the box marked β and multiplied by this term. This transformed voltage then appears in series with the input voltage, e_{in} which is applied to the input terminals of the amplifier (triangle with A_o). A_o is defined as the open-loop gain. It is the ratio of the voltage that appears between terminals 3 and 4 when a voltage is applied to terminals 1 and 2. The circuit of Fig. 44 is an example of voltage feedback and a similar analysis holds for networks incorporating current feedback.

The closed-loop gain, A_c, can then be found by inspection of Fig. 44. From the diagram, the output voltage must be

$$e_o = A_o(e_{in} + \beta e_o)$$

rearranging terms gives

$$e_o(1 - \beta A_o) = A_o e_{in}$$

and the closed-loop gain is defined by

$$A_c = \frac{A_o}{1 - \beta A_o} = \frac{e_o}{e_{in}}$$

Cathode Bias

As an application of the feedback concept, consider the amplifier circuit shown in Fig. 45. It will be recalled that a bias battery was required in the previous example and a method of eliminating this extra source is to insert a small-valued resistor in series with the cathode lead to ground (Fig. 45A). In terms of the amplifier block diagram, the circuit of Fig. 45B results. The next task is to evaluate the open-loop gain and the value of β.

With the exception of the cathode resistor, the circuit of Fig. 45 is the same as that of Fig. 42. Consequently, the ac plate current must be

$$i_p = \frac{-\mu e_{12}}{r_p + R_L + R_c}$$

The open-loop gain can then be determined and is

$$\frac{e_o}{e_{12}} = A_o = \frac{-\mu R_L}{r_p + R_L + R_c}$$

Next, β is determined from the expression for output voltage

$$e_o = i_p R_L$$

and the feedback voltage which is

$$e_f = i_p R_c$$

β is then

$$\beta = \frac{e_f}{e_o} = \frac{i_p R_c}{i_p R_L} = \frac{R_c}{R_L}$$

Note that β is positive since if the path 1 to 2 is considered, the feedback voltage is *added* to the input signal. Substituting the values of β and A_o into the feedback equation gives

$$A_c = \frac{A_o}{1 - \frac{R_c}{R_L} A_o}$$

which after some manipulation becomes

$$A_c = \frac{-\mu R_L}{r_p + R_L + (1 + \mu)R_c}$$

Comparison of this equation with the one for the previous circuit with no cathode resistor reveals that the gain has *decreased* because of the term $(1 + \mu)R_c$ in the denominator. Such an effect is called *negative* or *degenerative* feedback.

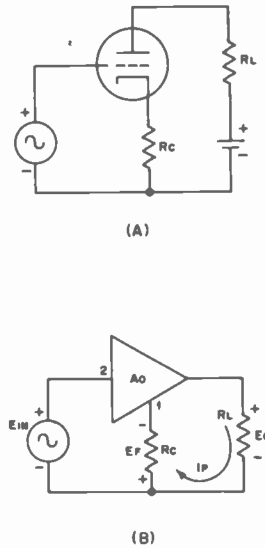


Fig. 45 – Feedback example of an amplifier with cathode bias.

On the other hand, if the feedback was such that the gain *increased*, *regenerative* or *positive* feedback would result. Positive feedback can be either beneficial or detrimental in nature and the study of feedback is an important one in electronics. For instance, frequency generation is possible in a circuit called an oscillator. But on the other hand, unwanted oscillation or *instability* in an amplifier is very undesirable.

OSCILLATORS

A special case of feedback occurs if the term

$$1 - \beta A_o$$

becomes zero. This would mean the closed-loop gain would become infinite. An implication of this effect is that a very small input signal would be amplified and fed back and amplified again until the output voltage became infinite. Either that, or amplifier output would exist with *no* signal input. Random noise could "trigger" the input into producing output.

Of course, an infinite output voltage is a physical impossibility and circuit limitations such as the nonlinearities of the active device would alter the feedback equation. For instance, at high output voltage swings, the amplifier would either "saturate" (be unable to supply more current) or "limit" (be cutoff because the grid was too negative) and A_o would decrease.

Tuned-Plate Tuned-Grid Oscillator

It should be stressed that it is the *product* of β A_o that must be 1.0 for oscillations to occur. In the general case, both β and A_o may be complex

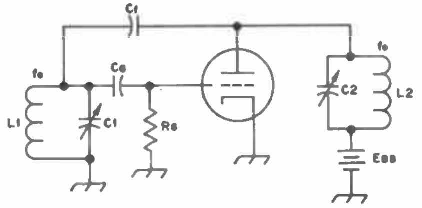
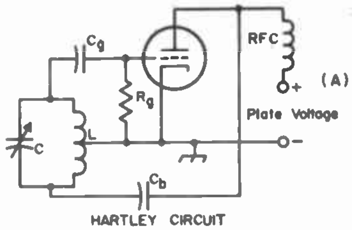


Fig. 46 – Tuned-plate tuned-grid oscillator.



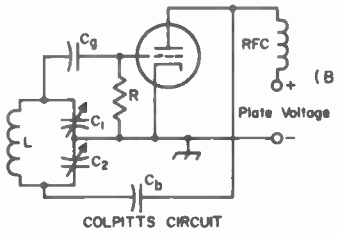


Fig. 47 – Hartley and Colpitts oscillators.

numbers unlike those of the cathode-bias problem just discussed. That is, there is a phase shift associated with A_o and β with the phase shift of the product being equal to the sum of the individual phase shifts associated with each entity.

Therefore, if the total phase shift is 180 degrees and if the amplitude of the product is 1.0, oscillations will occur. At low frequencies, these conditions normally are the result of the effects of reactive components. A typical example is shown in Fig. 46 and the configuration is called a tuned-plate tuned-grid oscillator. If the input circuit consisting of L_1 and C_1 is tuned to a frequency f_o, with the output circuit (L_2, C_2) tuned to the same frequency, a high impedance to ground will exist at the input and output of the amplifier. Consequently, a small capacitance value represented by C_f is capable of supplying sufficient voltage feedback from the plate to the grid.

At other frequencies, or if either circuit is detuned, oscillations may not occur. For instance, off-resonant conditions in the output tank will reduce the output voltage and in effect, reduce the open-loop gain to the point where oscillations will cease. On the other hand, if the input circuit is detuned far from f_o, it will present a low impedance in series with the relatively high reactance of C_f. The voltage divider thus formed will result in a small-valued β and the conditions for oscillations will not be fulfilled. However, for conditions near f_o, both the amplitude and phase of the βA_o product will be correct for oscillations to occur.

Under some conditions, the voltage across the tank circuit may be sufficient to cause the grid to be driven positive with respect to the cathode and grid current will flow through C_g. During the rest of the rf cycle, C_g will discharge through R_g causing a negative bias voltage to be applied to the grid. This bias voltage sets the operating point of the oscillator and prevents excessive current flow.

Miscellaneous Oscillator Circuits

Two other common type of oscillators are shown in Fig. 47. In Fig. 47A, feedback voltage is applied across a tapped inductor while in Fig. 47B, the voltage is applied across a capacitor instead. Quite often, a tuned plate circuit is not employed and an rf choke coil provides a high impedance load instead.

So-called "conventional" components such as tubes, transistors, ICs, resistors, inductors and capacitors are suitable up to and including the uhf range. However, at higher frequencies and for higher power levels in the uhf range, physical restrictions on the size of such components makes them impractical. Consequently, a different approach is required. All the components necessary for a particular application may be included in the active device itself. This is true in the klystron oscillator shown in Fig. 48. Here the feedback action takes place inside of the tube and in the electron stream. Electrons emitted from the cathode are accelerated and "modulated" on the first pass through the cavity resonator (which replaces the conventional tuned circuit used at lower frequencies). The electrons are then turned around by the repeller electrode and pass through

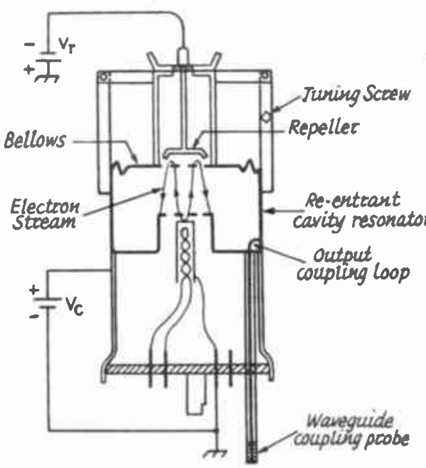


Fig. 48 – Cross-sectional view of a typical reflex klystron oscillator. Such types as the 732 may still be available on occasion in surplus sales.

the cavity again. On entering the cavity, the phase of the ac field there is such that the stream is retarded. However, this means that energy must be given up to the cavity and on out to the external circuit. As a result, the oscillations in the cavity are sustained.

Similar effects are employed in other microwave oscillators and amplifiers. Motional energy in the electron stream is transferred to a desired ac field. In doing so, dc energy in the power supply is converted to useful ac energy at the microwave frequency.

Solid-State Fundamentals

The electrical characteristics of solid-state devices such as diodes and transistors are dependent upon phenomena that take place at the atomic level. While semiconductors can be employed without a complete knowledge of these effects, some understanding is helpful in various applications. Electrons, which are the principal charge carriers in both vacuum tubes and semiconductors, behave much differently in either of the two circumstances. In free space, an electron can be considered as a small charged solid particle. On the other hand, the presence of matter affects this picture greatly. For instance, an electron attached to an atom has many properties similar to those of rf energy in tuned circuits. It has a frequency and wavelength that depend upon atomic parameters just as the frequency associated with electrical energy in a tuned circuit depends upon the values of inductance and capacitance.

A relation between the energy of an electron in an atomic "orbit" and its associated frequency is given by

$$f(\text{Hz}) = \frac{E(\text{joules})}{6.626 \times 10^{-34}}$$

where the constant in the denominator is called Planck's constant. This equation is quite important when an electron is either raised or falls between two different energy "states." For instance, when an electron drops from one level to a lower one, energy is emitted in the form of electromagnetic radiation. This is the effect that gives the characteristic glow to neon tubes, mercury-vapor rectifiers, and even light-emitting diodes. The frequency of the emitted radiation is given by the foregoing formula where E is the *difference* in energy.

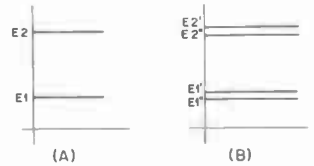


Fig. 1 — Energy-level diagram of a single atom is shown at A. In Fig. 1B, the levels split when two atoms are in close proximity.

However, if an electron receives enough energy such that it is torn from an atom, a process called ionization is said to occur (although the term is also loosely applied to transitions between any two levels). If the energy is divided by the charge of the electron (-1.6×10^{-19} coulombs), the equivalent in voltage is obtained.

A common way of illustrating these energy transistions is by means of the energy-level diagram shown in Fig. 1A. It should be noted that unlike ordinary graphical data, there is no significance to the horizontal axis. In the case of a single atom, the permitted energy can only exist at discrete levels (this would be characteristic of a gas at low pressure where the atoms are far apart). However, if a single atom is brought within close proximity of another one of similar type, the single energy levels split into pairs of two that are very close together (Fig. 1B). The analogy between tuned circuits and electron energy levels can be carried even further in this case.

Consider the two identical circuits that are coupled magnetically as shown in Fig. 2A. Normally, energy initially stored in C_1 would oscillate back and forth between L_1 and C_1 at a single

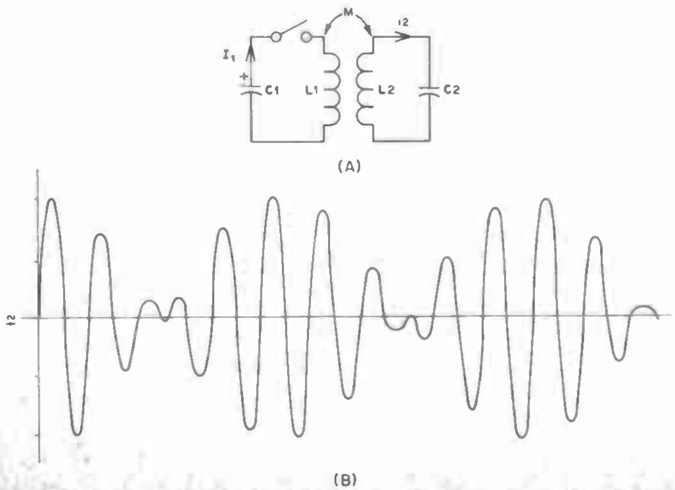


Fig. 2 — Electrical-circuit analog of coupled atoms.

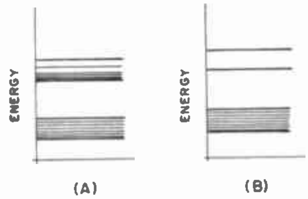


Fig. 3 – Energy-level diagram of a conductor is shown at A and a similar one for an insulator is shown in Fig. 3B.

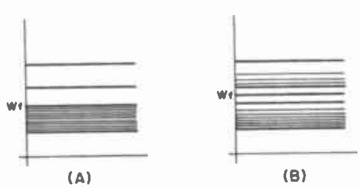
frequency after the switch was closed. However, the presence of the second circuit consisting of L_2 and C_2 (assume L_1 equals L_2 and C_1 equals C_2) results in the waveform shown in Fig. 2B. Energy also oscillates back and forth between the two circuits and the current then consists of components at two slightly different frequencies. The effect is similar to the splitting of electron energy levels when two atoms are close enough to interact.

CONDUCTORS, INSULATORS, AND SEMICONDUCTORS

Solids are examples of large numbers of atoms in close proximity. As might be expected, the splitting of energy levels continues until a band structure is reached. Depending upon the type of atom, and the physical arrangement of the component atoms in the solid, three basic conditions can exist. In Fig. 3A, the two discrete energy levels have split into two bands. All the states in the lower band are "occupied" by electrons while the ones in the higher energy band are only partially filled.

In order to impart motion to an electron, the expenditure of energy is required. This means an electron must then be raised from one energy state to a higher one. Since there are many permitted states in upper level of Fig. 3A that are both unoccupied and close together, electrons in this level are relatively free to move about. Consequently, the material is a conductor. In Fig. 3B, all the states in the lower level are occupied, there is a big gap between this level and the next higher one, and the upper level is empty. This means if motion is to be imparted to an electron, it must be raised from the lower level to the upper one. Since this requires considerable energy, the material is an insulator. (The energy-level representation gives an insight into the phenomena of breakdown. If the force on an electron in an insulator becomes high

Fig. 4 – Energy-level diagram of a semiconductor.



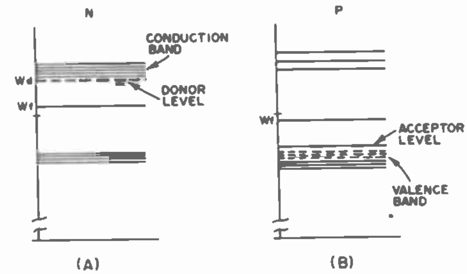


Fig. 5 – The effects on the energy-level diagram if impurity atoms are introduced.

enough because of an applied field, it can acquire enough energy to be raised to the upper level. When this happens, the material goes into a conducting state.)

A third condition is shown in Fig. 4. In the material associated with this diagram, the upper level is unoccupied but is very close to the occupied one. Hence, under conditions where the random electron motion is low (low temperature), the material acts as an insulator (Fig. 4A). However, as the random or thermal motion increases, some electrons acquire enough energy to move up to states in the upper level. Consequently, both levels are partially occupied as shown in Fig. 4B. The line marked W_f represents a statistical entity related to the "average" energy of electrons in the material and is called the Fermi Level. At absolute zero (no thermal motion), W_f is just at the top of the lower energy level. As electrons attain enough energy to move to the upper level, W_f is approximately halfway between the two levels.

THE PN JUNCTION

The material for the diagram shown in Fig. 4 is called an intrinsic semiconductor and examples are the elements germanium and silicon. As such, the materials do not have any rectifying properties by themselves. However, if certain elements are mixed into the intrinsic semiconductor in trace amounts, a mechanism for rectification exists. This is shown in Fig. 5A. If an element with an occupied energy level such as arsenic is introduced into germanium, a transformation in conductivity takes place. Electrons in the new occupied level are very close to the upper partially filled band of the intrinsic germanium. Consequently, there are many extra charge carriers available when thermal energy is sufficient to raise some of the electrons in the new level to the partially filled one. Germanium with an excess of mobile electrons is called an n-type semiconductor.

By introducing an element with an empty or unoccupied energy level near the lower partially filled level (such as boron), a somewhat different transformation in conductivity occurs. This is shown in Fig. 5B. Electrons from the lower level can move into the new unoccupied level if the thermal energy is sufficient. This means there is an excess of *unoccupied* states in the germanium

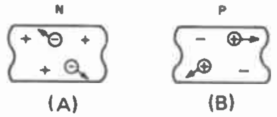


Fig. 6 — N- and p-type semiconductors.

lower energy level. Germanium treated this way is called a p-type semiconductor.

A physical picture of both effects is shown in Fig. 6. The trace elements or impurities are spread throughout the intrinsic crystal. Since the distance of separation is much greater for atoms of the trace elements than it is for ones of the intrinsic crystal, there is little interaction between the former. Because of this lack of "coupling," the distribution of energy states is a single level rather than a band. In Fig. 6A, atoms of the trace element are represented by the + signs since they have lost an electron to the higher energy level. Consequently, such elements are called donors. In Fig. 6B, the impurity atoms that have "trapped" an electron in the new state are indicated by the – signs. Atoms of this type are called acceptor impurities.

While it is easy to picture the extra free electrons by the circled "minus" charges in Fig. 6A, a conceptual difficulty exists with the freed "positive" charges shown in Fig. 6B. In either case, it is the motion of electrons that is actually taking place and the factor that is responsible for any current. However, it is convenient to consider that a positive charge carrier exists called a hole. It would seem as though a dislocation in the crystal-lattice structure was moving about and contributing to the total current.

If a section of n-type material is joined to another section made from p-type, a one-way current flow results. This is shown in Fig. 7. A positive potential applied to the p-type electrode attracts any electrons that diffuse in from the n-type end. Likewise, holes migrating from the

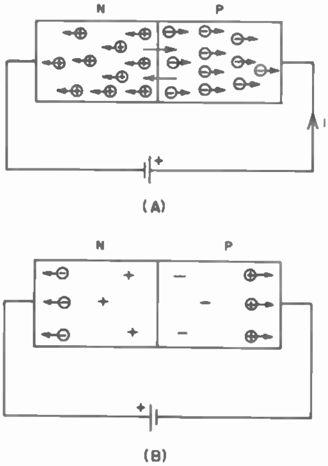


Fig. 7 — Elementary illustration of current flow in a semiconductor diode.

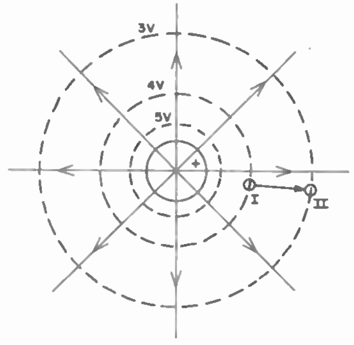


Fig. 8 — Potential diagram of an electron in an atomic orbit.

p-type end into the n-type electrode are attracted to the negative terminal. Note that the diagram indicates not all the carriers reach the terminals. This is because some carriers combine with ones of the opposite sign while enroute. In the case of a diode, this effect doesn't present much of a problem since the total current remains the same. Other carriers take the place of those originally injected from the opposite regions. However, such recombination degrades the performance of transistors considerably and will be discussed shortly.

If a voltage of the opposite polarity to that of Fig. 7A is applied to the terminals, the condition in Fig. 7B results. The mobile charge carriers migrate to each end as shown leaving only the fixed charges in the center near the junction. Consequently, little current flows and the pn junction is "back biased." It can be seen that the pn junction constitutes a diode since current can flow readily only in one direction. While this simple picture suffices for introductory purposes, proper treatment of many important effects in semiconductors requires a more advanced analysis than the elementary model affords. Returning to Figs. 3, 4 and 5, it would be convenient if the diagrams were in terms of voltage rather than energy. As pointed out earlier, the relation between energy and voltage associated with an electron is given by

$$W = eV = (-1.6 \times 10^{-19})\,V$$

Because the electron has been assigned a minus charge, a somewhat upside-down world results. However, if it is kept in mind that it requires the *expenditure* of energy to move an electron from a point of *higher* potential to one at a *lower* value,

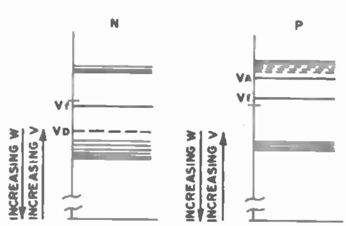


Fig. 9 — Energy-level diagrams in terms of potential.

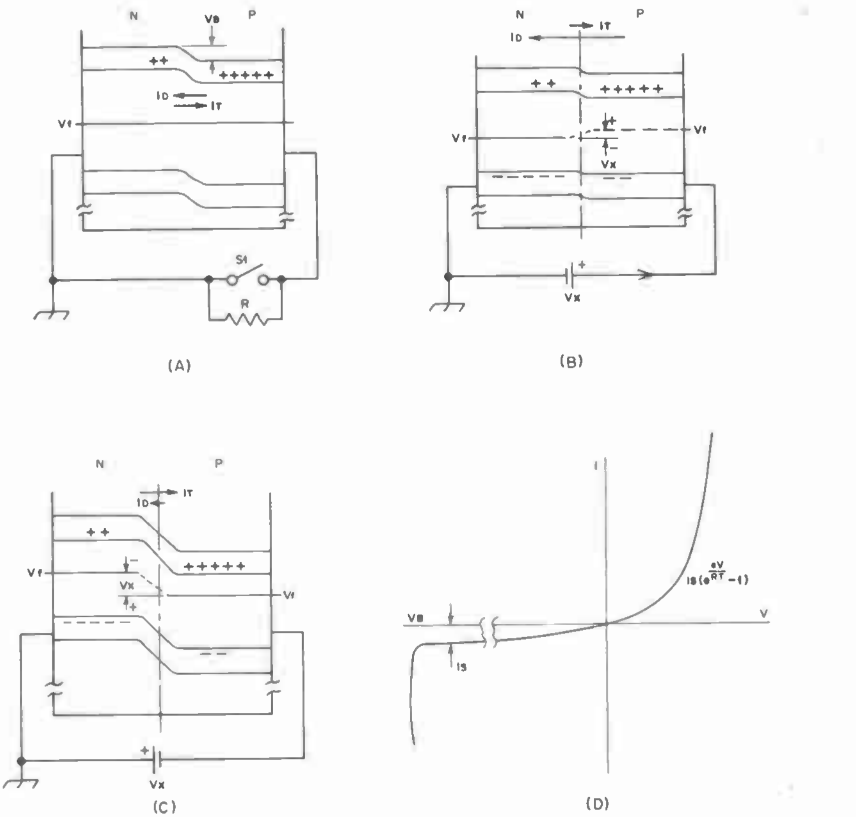


Fig. 10 — Energy-level diagrams for unbiased (A), forward-biased (B), and reversed-biased (C) diode. Resultant diode characteristics are shown in Fig. 10D.

this confusion can be avoided. As an illustration, suppose an electron is moved from an atomic orbit indicated by I in Fig. 8 to orbit II. This would mean the electron would have had to been moved against the force of attraction caused by the positive nucleus resulting in an *increase* in potential energy. (In other words, orbit II is at a higher energy level than orbit I.) However, note that the electrostatic potential around the nucleus decreases with distance and that orbit II is at a lower potential than orbit I.

Consequently, the energy-level diagram in terms of voltage becomes inverted as shown in Fig. 9. It is now possible to approach the problem of the pn junction diode in terms of the energy-level diagrams presented previously. If a section of n-type and p-type material is considered separately, the respective energy (or voltage) levels would be the same. However, if the two sections were joined together and connected by an external conductor as shown in Fig. 10, a current would flow initially. This is because the voltage corresponding to the statistical entity referred to previously (Fermi Level) is not the same for p- and n-type materials at the same temperature. At the Fermi Level, the probability that a particular energy state is occupied is one half. For n-type material, the Fermi Level is shifted upward toward the "conduction band" (Fig. 5A). In a p-type material, it is shifted downward toward the "valence band." Although the theory behind the Fermi Level and definitions concerning the conduction and valence bands won't be dealt with here, it is sufficient to know that the band structure shifts so that the Fermi Levels are the same in both parts of the joined sections (Fig. 10).

The reasoning behind this effect is as follows. Consider conditions for hole flow only for the moment. Since there is an excess of holes in the p region (Fig. 10), there is a tendency for them to move over into the adjacent n region because of diffusion. The process of diffusion is demonstrated easily. If a small amount of dye is dropped into some water, it is concentrated in a small area at first. However, after a period of time has passed, it spreads out completely through the entire volume.

Once the holes diffuse into the n region, they recombine with the electrons present and produce a current in the external terminals denoted by I_D (Fig. 10). But a paradox results because of this current. If S_1 is opened so that I_D flows through R, where does the energy that is transferred (irreversibly) to this resistance come from? In effect, it represents a perpetual-motion dilemma or else the semiconductor will cool down since the diffusion process is the result of a form of thermal motion. Both conclusions are against the laws of physics so a third alternative is necessary. It is then assumed that the Fermi Levels align so that the potential across the terminals becomes zero and no current will flow in the external circuit.

However, if the Fermi Levels are the same, the conduction and valence bands in either section will no longer align. As a consequence, a difference in potential between the two levels exists and is indicated by V_B in Fig. 10. *The formation of this junction or barrier voltage is of prime importance in the operation of pn-junction devices.* Note that holes in the p region must overcome the barrier voltage which impedes the flow of the diffusion current. It will also be recalled that both holes and electrons were generated in the intrinsic semiconductor because of thermal effects (Fig. 4B). The addition of either donor or acceptor atoms modifies this effect somewhat. If donor atoms are present (n-type material), fewer holes are generated. On the other hand, if acceptor atoms represent the impurities, fewer electrons are generated in comparison to conditions in an intrinsic semiconductor. In the case of p-type material, holes predominate and are termed the majority carriers. Since there are fewer electrons in p-type material, they are termed the minority carriers.

Referring to Fig. 10A, there are some holes in the n region (indicated by the + signs) because of the foregoing thermal effects. Those near the junction will experience a force caused by the electric field associated with the barrier voltage. This field will produce a flow of holes into the p region and the current is denoted by I_T. Such a current is called a drift current as compared to the diffusion current, I_D. Under equilibrium conditions, the two currents are equal and just cancel each other. This is consistent with the assumption that no current flows in the external circuit because of the fact that the Fermi Levels are the same and no voltage is produced.

So far, only conditions for the holes in the upper (or conduction) band have been considered but identical effects take place with the motion of electrons in the lower energy band (valence band). Since the flow of charge carriers is in opposition, but because holes and electrons have opposite signs, the currents *add.*

The Forward-Biased Diode

If an external emf is applied to the diode terminals as shown in Fig. 10B, the equilibrium conditions no longer exist and the Fermi-Level voltage in the right-hand region is shifted upward. This means the barrier voltage is decreased and considerable numbers of carriers may now *diffuse* across the junction. Consequently, I_D becomes very large while I_T decreases in value because of the decrease in barrier voltage. The total current under "forward-bias" conditions then becomes

$$I = I_s\,(e^{\frac{eVx}{kT}} - 1)$$

At room temperature, the ratio kT/e is approximately .026 volts and so for an external voltage, V_x that exceeds this value, the current increases very rapidly.

The Reversed-Biased Diode

If the source, V_x, is reversed as shown in Fig. 10C, the barrier voltage is *increased.* Consequently, charge carriers must overcome a large "potential hill" and the diffusion current becomes very small. However, the drift current caused by the thermally generated carriers returns to the value it had under equilibrium conditions. For large values of V_x, the current approaches I_s, defined as the reverse saturation current, I_s is the sum of I_T and its counterpart in the lower or "valence" band. Finally, the characteristic curves of the forward- and reversed-bias diode can be constructed and are shown in Fig. 10D.

It is obvious that I_s should be as small as possible in a practical diode since it would only degrade rectifier action. Also, since it is the result of the generation of thermal carriers, it is quite temperature sensitive which is important when the diode is part of a transistor. If the reverse voltage is increased further, an effect called avalanche breakdown occurs as indicated by the sudden increase in current at V_b. In such an instance, the diode might be damaged by excessive current. However, the effect is also useful for regulator purposes and devices used for this purpose are called Zener diodes.

BARRIER RECTIFICATION

The pn junction is involved in just about every modern semiconductor device that is likely to be encountered, such as bipolar transistors, field-effect transistors (FETs), integrated circuits (ICs), and various other components. Although such junctions are common, and the theory behind them is well understood, they are difficult to construct. Very pure germanium or silicon is required and special equipment must be used to diffuse in the impurity elements that result in either p- or n-type material. It is not likely there will ever be a "do-it-yourself" repair kit for the experimenter that will allow him to rebuild faulty pn-junction devices.

On the other hand, rectification of another form exists and components such as the "cat's-whisker" detector, copper-oxide rectifier, and more recently, Schottky barrier diodes operate on principles that are somewhat different from those involved in the pn junction. Except for the latter, these devices are very easy to construct (sometimes undesirably so as in the case of unwanted rectification in downspouts and other structures that may cause TVI), but the theory is rather complicated and not completely understood in regard to some aspects.

Fermi Level in Conductors

It will be recalled from previous sections that the Fermi Level represented a sort of statistical average separating energy levels or states populated by electrons from empty states. This entity also played an important role in the theory of the pn junction. It is also involved in the rectifiers under discussion.

Before proceeding, some comment in regard to the units associated with electron-energy considerations would be helpful. Because of its extremely small charge (-1.6×10^{-19} coulombs), an electron doesn't acquire much energy when it is moved through a potential of one volt under the influence of an electric field. The energy so acquired would be 1.6×10^{-19} joules and it is convenient to define this quantity as one electron volt.

In terms of thermal energy, an electron volt represents an extremely high temperature. For instance, room temperature is approximately 273 degrees on the Kelvin scale while 1 electron volt would be equivalent to 11,606 degrees Kelvin! In copper, electrons occupy orbits out to 7 electron volts (7 eV) which would correspond to a temperature of 82,000 degrees Kelvin. As might be expected, variations at room temperature are not likely to have much effect on the electron distribution which tends to remain as it was at 0 degrees K. At zero degrees Kelvin (or "absolute" zero), the Fermi Level (W_f) forms a sharp dividing line between the occupied and unoccupied states in the upper end of the conduction band as illustrated in Fig. 11. For all practical purposes, the Fermi Level in a conductor is independent of temperature.

As a comparison, the difference in energy levels between an arsenic donor impurity in germanium and the bottom of the conduction band is .0127 eV. This corresponds to a temperature of 147°K or −126°C. Consequently, many of the donor atoms are "ionized" and contribute electrons to the conduction band. On the other hand, the gap between the valance and conduction bands in germanium is 0.7 eV. One might expect that carriers generated in the intrinsic crystal do not have a significant effect until much higher temperatures are reached than normally encountered since this gap represents a temperature of over 7800 degrees. However, the thermally generated carriers in the intrinsic semiconductor are still important as indicated in the discussion on the pn junction. This is because even small numbers of thermally generated carriers affect operation of the device.

One might think of the Fermi Level as corresponding to the surface of the water in the ocean. On calm days, the surface is smooth and the *average* and *actual* water levels are almost the same. On the other hand, during rough weather, peaks and valleys formed by the waves may differ considerably from the average. The former condition is representative of the energy "surface" of a conductor while the latter one is typical of the distribution of energy states in a semiconductor caused by the "thermal wind."

Work Function

While the Fermi Level more or less defines the upper limit of the occupied bands in a conductor (Fig. 11) in the *interior* regions, conditions on the surface are of interest. An important consideration

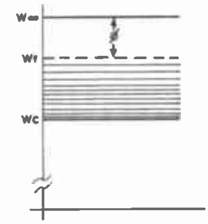


Fig. 11 — Energy-level diagram of metal showing work function.

is how much energy is required to remove completely an electron from the conductor. This energy is indicated by the symbol ϕ in Fig. 11 and is called the work function of the material. Factors that contribute to the work function are unconnected atomic bonds (because of the discontinuity at the surface) which attract the electron as it is pulled away from the boundary. Also, the "image" of the electron itself can be considered to be a positive charge moving into the metal thus causing an attractive force.

The work function is of importance in many devices, the most common example being the vacuum tube. A formula for the emission of electrons from the cathode in a vacuum tube is given by

$$j = AT^2 e^{\frac{-\phi}{kT}}$$

known as the Richardson-Dushman equation. A is a constant, T is the temperature in degrees Kelvin, k is Boltzman's constant (1.38×10^{-23} joules/degree) and ϕ is the work function in joules (or ϕ in electron volts multiplied by 1.6×10^{-19}). Typical work functions range from 1 to 5 eV. For instance, thoriated tungsten has a work function of 2.7 eV with a value of .04 A/cm^2 (°K)^2 for A. Operating at 1600°C, this would result in a current density of 7.7 mA for every square centimeter of cathode area. However, applying a field at the cathode that tends to pull the electrons away (such as that caused by the positive voltage of the plate) also reduces the work function. This is known as the Schottky effect and the resultant current density increases as the plate voltage is increased.

THE BARRIER DIODE

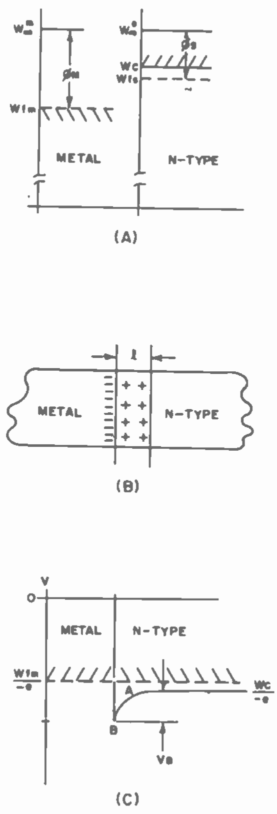
One possible explanation of the rectification effects found naturally in many devices is illustrated in Fig. 12A.* A metal with a work function ϕ_m is brought into contact with an n-type semiconductor with a work function ϕ_s. Much of the difficulty between theory and experiment is concerned with the latter entity. Conditions on the surface of the semiconductor affect the work function greatly. In the present discussion, it is assumed these effects can be neglected.

*C. Kittel, *Introduction to Solid State Physics*, John Wiley & Sons, New York, 1956.

After contact, some of the electrons from donor atoms in the semiconductor move into the metal along with those that move up into the conduction band of the semiconductor. This leaves a positive charge near the junction in the semiconductor (Fig. 12B). The electrons migrate until an equilibrium is reached and the Fermi Levels align. A corresponding negative charge exists on the junction with the metal but because of the high conductivity, it is confined to a very narrow region. A charge distribution in the interior of a metal implies the existence of an electric field. Such a field is inconsistent with high conductivity since a high current would flow. Consequently, all the charge must reside on the surface of the metal.

On the other hand, because of the lower conductivity of the semiconductor, the charge distribution can penetrate into the n-type region for a considerable distance. This distribution is known as a barrier layer. It is also sometimes referred to as the space-charge region. (The conditions in the pn junction are somewhat more complicated because of the presence of two kinds of charge carriers. For instance, in a region in the vicinity of the middle of the junction, the charge is neutralized and the region behaves much like an intrinsic semiconductor. This area is called the depletion layer.)

Fig. 12 — Energy-level diagrams of a metal and a semiconductor before contact are shown at A. After the metal and semiconductor are joined, a charged layer is formed as shown at B. The energy-level diagram (in terms of the voltage) of the metal/semiconductor junction is shown at C.



The resultant energy-level diagram in terms of the voltage is shown in Fig. 12C. Note that since the charge on the electron is negative, the diagram becomes inverted because of considerations discussed earlier (Figs. 8 and 9). Electrons to the right of the barrier that might tend to diffuse over to the metal now experience an energy barrier that must be surmounted. The barrier voltage is denoted by V_B (Fig. 12C) and is the difference between the values of the two work functions divided by the electronic charge (-1.6×10^{-19} coulombs). It will also be recalled from an earlier discussion that moving an electron from a point to one that is more negative required a positive expenditure of energy. Therefore, an electron "falls up" a potential "hill" but must "climb down" a potential "valley." This means expenditure of energy in moving an electron from point *a* to *b* (Fig. 12C) which would be returned in going from *b* to *a*.

Diode Operation

If a voltage is applied to the terminals of the configuration shown in Fig. 13A, the rectifying properties can be illustrated in the following manner. A negative voltage at the semiconductor end will shift the Fermi "voltage" downward and the barrier will decrease. Consequently, electrons will migrate easily across the barrier and the current will increase exponentially with the applied voltage as was the case with the forward-biased pn junction. An interesting feature of this diode action is that electrons entering the metal do so at energy levels that exceed the Fermi Level. They remain at higher levels than the rest of the electrons in the metal for some time and consequently are called hot carriers.

Another difference between this type of diode and the pn junction is that the current is caused by majority carriers. It will be recalled in the pn junction that holes were injected into the n region and electrons injected into the p region where they either recombined or migrated to the opposite electrode. Although an electron is the majority carrier in the n region, it becomes a *minority* carrier in the adjacent p region. Only dc effects have been considered so far, but slow recombination of the injected carriers degrades the high-frequency characteristics of the pn-junction diodes. The problem arises when the diode is "switched" from the forward conducting state into the reverse-biased condition. Injected minority carriers are indistinguishable from those generated thermally. Consequently, instead of the "normal" reverse saturation current, a much larger current flows because of this "stored charge." The effect is to increase the junction capacitance of the diode when it is in a reverse-biased condition. (In some pn-junction diodes, this action is used to advantage. A reverse current flows until all the stored charge is used up. At this time, the diode goes from a conducting state into nonconduction during an extremely short period. Such diodes are called "snap" diodes or step-recovery diodes and are useful in frequency-multiplier applications because of the short pulse generated when the switching action occurs.)

If the n-type region is made positive with respect to the metal, the condition in Fig. 13C results. Because of its lower conductivity, most of the voltage drop appears across the semiconductor. The effect is to widen the barrier layer and increase the barrier voltage as shown. This means fewer electrons to the right of the barrier will have sufficient energy to reach the metal and the current is reduced considerably.

Electrons in the metal with sufficient energy to overcome the barrier in the metal side will migrate toward the y terminal and a reverse current, I_o will flow (Fig. 13C). However, few electrons will have sufficient energy and the reverse current of the barrier diode is similar to that of the pn-junction diode. Note that barrier voltage on the metal side of the metal-semiconductor boundary stays the same for either forward, reverse, or equilibrium conditions. The validity of this assumption is illustrated in Fig. 13D. Assume that the Fermi Level in the semiconductor remained constant while the Fermi Level in the metal was allowed to vary as indicated in Fig. 13D. An immediate implication is that the barrier-layer voltage, V_B, and the voltage distribution would also remain the same under all conditions. However, this is a contradiction to the assumption that the greatest percentage of the voltage drop appeared in the semiconductor region because of its lower conductivity. Consequently, the conditions illustrated in Fig. 12C, and Figs. 13B and 13C must be correct while the condition of Fig. 13D cannot exist for a metal with a much higher conductivity than the adjacent semiconductor.

This leads to an important conclusion that two dissimilar metals joined together do not result in rectification even though the work functions are different and a "contact potential" exists. On the other hand, many metal oxides have low conductivity and exhibit semiconducting properties which should be kept in mind in the construction of antennas and similar hardware exposed to the elements. Corroded connections or intermittent contacts are often the cause of harmonic problems generated by stray rectification.

THE FIELD-EFFECT TRANSISTOR

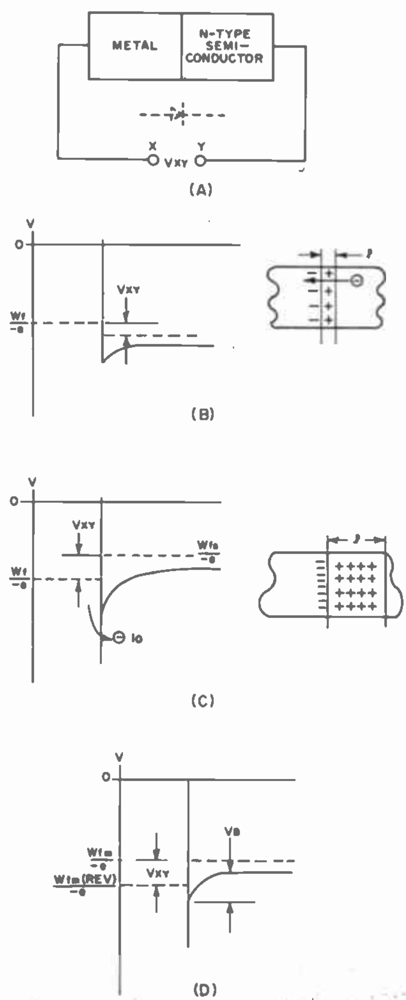
It was indicated in the previous section that the width of the barrier layer in the metal-semiconductor diode was dependent upon the applied voltage. Similar effects also occur with the pn junction. In the case of a reverse-biased diode, the width is proportional to the square root of the voltage for either case.

This effect is used in a device known as the field-effect transistor (FET). Historically, FETs came much later than the bipolar transistor (which will be discussed presently), but the mechanism for amplification is somewhat easier to visualize with the FET. A simplified diagram of the FET is shown in Fig. 14A. A wafer of n-type semiconducting material is sandwiched between two p-type electrodes. The combination of the latter two elements is called the gate with the top of the n-type wafer designated as the drain. The bottom electrode is called the source.

A positive voltage applied to the drain-source terminals results in a current I_{DS} (Fig. 14A). If the gate-to-source voltage is zero, this current is limited only by the resistance of the n region. However, if a bias is applied (V_{GS} not zero), the junctions are reverse-biased and a space-charge extends into the n wafer as shown in Fig. 14C. The crosshatched area indicates the "intrinsic" or neutral region.

Because this area is one of low conductivity, the current is restricted to the middle or is "channeled" between the two p-type electrodes. Finally, if the gate voltage is made great enough, the channel width is zero and little current flows. Such a condition is referred to as a pinch off.

Fig. 13 — Energy-level diagrams (voltage) for forward and reversed-biased metal/semiconductor junction are shown at B and C.



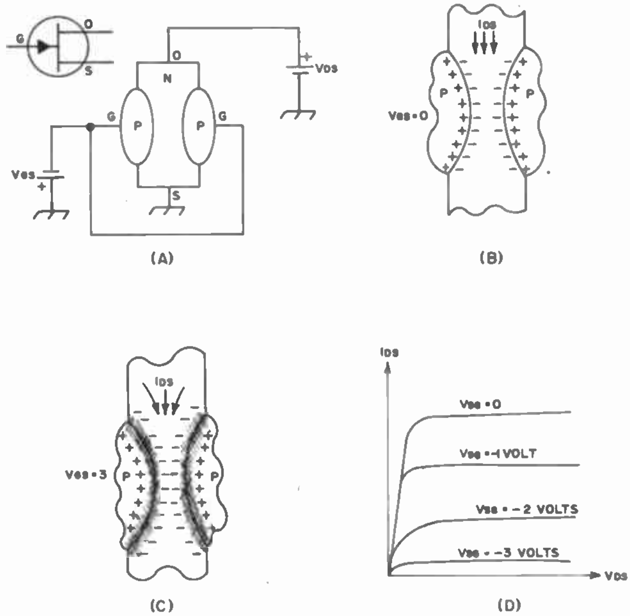


Fig. 14 – The field-effect transistor.

Typical characteristic curves for a field-effect transistor are illustrated in Fig. 14D.

Since the input of the FET is basically a reverse-biased diode, the gate-to-source impedance is very high. A similar condition exists with the vacuum tube and both devices might be described as being voltage controlled. Except for the voltage levels, an FET and a pentode vacuum tube have almost identical characteristics.

Devices such as the FET that operate with reverse-biased junctions are less common than those with diodes biased in the forward-conducting state. However, the variation in junction width as a function of voltage of the reverse-biased diode is used in another device. The semiconducting regions outside of the junction act as capacitor plates and the diode can be used as a voltage-variable capacitor. Such diodes (sometimes called Varicaps) are used in voltage-controlled oscillators and similar applications. Either special diodes or ordinary "switching" diodes can be used for this purpose.

THE BIPOLAR TRANSISTOR

Next, consider the configuration illustrated in Fig. 15. Two pn junctions are connected back to back and voltages with the polarities shown are applied. Note that the junction on the left is forward biased and electrons migrate easily into the adjoining p region from the n region. Holes also migrate from the center p region into the left-hand n region where they either recombine or reach the terminal marked "E."

Some of the electrons injected into the p region also recombine and thus cause a current to flow in terminal "B" (in addition to that caused by the migration of holes into the left-hand n region). Others reach the right-hand n region and contribute to the current at terminal "C." Since terminals "B" and "C" form a back-biased diode, practically all of the current i_c results from electrons injected into the middle p region from the left-hand n region.

Since the power gain of any device is defined as the ratio of output power across some load to input power consumed, it can be shown that the device of Fig. 8 is capable of power amplification. The input power is the square of the input current times the input resistance. However, the pn junction between terminals "E" and "B" is forward biased and hence the input resistance will be very low. On the other hand, there are no restrictions on the value of R_L (Fig. 15) and it can be made much higher than the equivalent input resistance. Since the current i_e is only slightly higher than i_c by an amount equal to i_b, the input and output currents are approximately the same. Consequently, the power gain is approximately

$$P \cong 10 \log_{10} \frac{R_L}{R_e}$$

where R_e is the equivalent input resistance. Since R_L is assumed to be much larger than R_e, the device has a positive power gain. Note that the

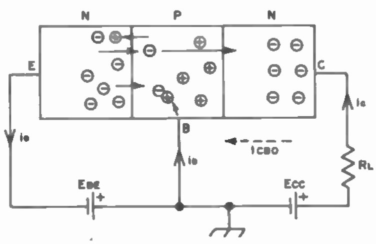


Fig. 15 – The bipolar transistor.

power gain is accomplished by means of a *trans*ferred res*istance* hence the name transistor.

TRANSISTOR NOMENCLATURE

As pointed out, it is possible to fabricate a device such that almost all of the current injected into the middle p region (i_e) reaches the load (i_c). Because of the similarity of the left-hand n region to the cathode of a vacuum tube (which emits electrons), the terminal marked "E" is referred to as the emitter. The electrode at the right-hand end is somewhat analogous to the plate of a vacuum tube (which collects electrons) and is called the collector. In transistor construction, the collector and emitter electrodes may be "cat-whisker" wires attached to base of n-type germanium (point-contact transistor) or small pellets of metal fused onto a base of n- or p-type semiconducting material (fused- or alloy-junction transistor). In either case, since the electrode marked "B" is a block or base of semiconductor material, it is labeled the base.

Point-contact transistors were the first type to be made but are seldom used anymore. Most transistors encountered are fabricated by more complex processes. The general method is to take a block of either p- or n-type semiconducting material which forms the base and diffuse in impurities of the opposite type to form the collector and emitter electrodes. Such devices are usually referred to as junction transistors regardless of the actual process used. Both point-contact and junction transistor come under a more general category called bipolar transistors.

Transistor Parameters

Since the power gain depends upon the efficiency with which current is transferred from the emitter to collector, the common-base forward current-transfer ratio is an important parameter in determining transistor-amplifier performance. This ratio is defined by

$$\alpha = \frac{i_c}{i_e}$$

and in high-quality transistors, attains a value very close to 1.0.

The difference between the emitter current and the collector current of the transistor "model" shown in Fig. 15 is caused only by recombination of carriers in the base region. Therefore, by making the base region very narrow, more carriers have a chance of getting across resulting in higher values of alpha. However, another factor is present that causes an additional contribution to collector current.

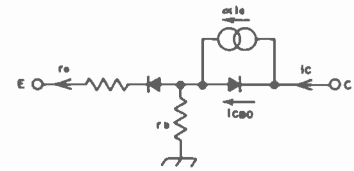


Fig. 17 — Equivalent circuit of a bipolar transistor.

It will be recalled from previous discussions that even in an intrinsic semiconductor, some electrons attain enough energy from thermal motion to go from the lower energy level to the higher one. The electrons and holes generated thermally in the intrinsic semiconductor are indistinguishable from those generated thermally by the impurity atoms. However, the latter far outnumber the former even for very low impurity concentrations. Also, at room temperature, almost all of the impurity holes and electrons have been generated. This is not true for the ones from the intrinsic semiconductor.

In a p-type semiconductor, the holes are called the majority carriers since they far outnumber the electrons which in this instance would be the minority carriers. The presence of minority carriers may be caused purposely as in the case of electrons injected into the base p region of the transistor shown in Fig. 15. But there are also thermally generated holes from the intrinsic semiconductor in the collector n region and likewise for electrons in the base p region. Consequently, there is an additional current component labeled I_{CBO} with the direction indicated by the dashed arrow in Fig. 15. Thus, even with the emitter open-circuited, the collector current will not be zero as shown on the set of somewhat idealized transistor curves of Fig. 16. Such a current is called a reverse-saturation current and is typical of any back-biased pn junction. A more accurate model of the transistor can be derived by including a reverse-biased diode as in the equivalent circuit of Fig. 17.

Since the reverse-saturation current is caused by the generation of minority carriers in the intrinsic semiconductor and since the latter effect is highly dependent upon temperature, the current is sensitive to temperature also. An equation for this current is given by

$$I_{CBO} = Ce^{k(T - T_o)}$$

where C is a constant measured at some arbitrary temperature, T_o. With silicon transistors, the value of C is usually quite low and such units are superior to germanium types in this respect. However, with either variety, caution must be taken to insure temperature effects do not interfere with the operation of the device (or even cause damage) as illustrated in Fig. 18.

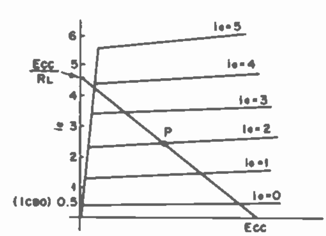


Fig. 16 — Bipolar-transistor characteristics.

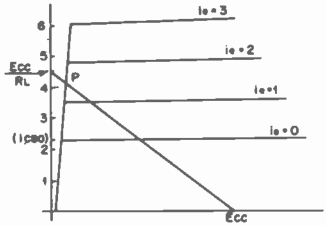


Fig. 18 – Shift in operating point of a bipolar transistor as a function of temperature.

In Fig. 16, the operating point *P* of a transistor is shown for a forward-biased emitter current of 2 units and a collector load R_L. Note that I_{CBO} is 0.5 units originally. However, because of a temperature rise, I_{CBO} has increased to a value slightly over 5 units as shown in Fig. 18. This means the entire set of curves will be shifted up and, as a consequence, the quiescent current of the operating point will also be increased. In the case of a power device, R_L is usually very low in value and the new current is likely to be outside of safe operating limits. Consequent permanent damage will result and the effect is called thermal runaway.

There are two possible solutions to the problem. One would be to employ a temperature-sensitive or current-sensitive feedback system that lowers the forward bias as the temperature and current increases. However, a better approach is to conduct heat away by means of heat sinks or similar cooling devices. This is why solid-state equipment must be provided with adequate ventilation especially in high-power applications such as power inverters or rf amplifiers. Since heat rise is proportional to the time that power is consumed, strict adherence to duty cycles is required also.

Transistor parameters are also sensitive to frequency since it takes time for charge carriers to traverse the base region. Rf-*power* devices are now frequently optimized for a certain frequency range. Thus, while a vhf transistor would work at hf, it is better to employ devices designed for this range. This represents a departure from older vacuum-tube conditions where there was only an upper frequency limit to be concerned about.

So-called small-signal (low-power) transistors can be usually used at low frequencies and up to a limit called the alpha-cutoff frequency which is related to a by

$$a = \frac{a_o}{\sqrt{1 + \left(\frac{f}{f_{ab}}\right)^2}}$$

However, it is wise to restrict operation to a range considerably below this value. (In the foregoing equation, a_o is the current gain at low frequencies. Note that when f approaches f_{ab} the a is reduced by a factor $1/\sqrt{2}$ which corresponds to a decrease of 3 dB. Consequently, f_{ab} is sometimes called the 3-dB cutoff frequency.)

COMMON–EMITTER CONNECTION

Although the configuration of Fig. 19 is capable of power gain, it is somewhat inconvenient to use in many applications. In addition to the transistor type shown in Fig. 19, a second type is possible and is illustrated in Fig. 19A. Instead of a middle p region, an n-type semiconductor wafer is used while the emitter and collector are made from p-type material. The transistor shown in Fig. 19A is called a pnp type while the illustration and schematic diagram shown in Fig. 19B is for an npn transistor.

The schematic diagram shown in Fig. 20 is for the configuration illustrated in Fig. 19 and is referred to as the common-base connection (and is for an npn transistor). A corresponding common-emitter connection is shown in Fig. 20B. It will be recalled that a, the common-base current gain is given by

$$i_c = ai_e$$

and from current-flow conservation considerations

$$i_e = i_c + i_b$$

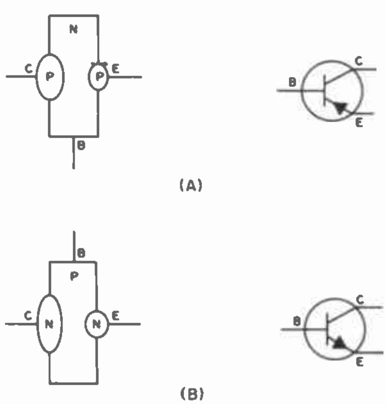


Fig. 19 – Pnp and npn bipolar transistors.

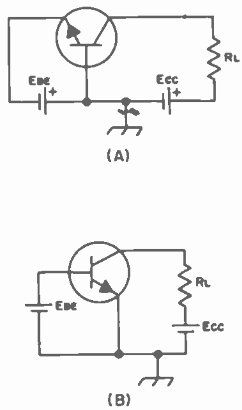


Fig. 20 – Grounded-base and grounded-emitter circuit arrangements.

Combining the two equations gives

$$ai_c = i_c + i_b$$

and with some rearranging, the common-emitter current gain β, is then determined by

$$\beta = \frac{i_c}{i_b} = \frac{1}{1-a}$$

Since a is very close to 1.0 in a good transistor, β is greater than 1.0 and in fact, is usually quite high. Frequency considerations such as those for a exist for the common-emitter current gain also. The cutoff frequency f_c is the frequency where the value of β is down by 3 dB from its low-frequency value. A somewhat more useful term is the gain-bandwidth product, f_T. This is the frequency where the value of β is equal to 1.0. Consequently, above f_T the transistor gain is less than 1.0.

CHARACTERISTIC CURVES

The application of semiconductors differs somewhat from methods used with vacuum tubes. Only a few parameters are required to characterize a vacuum tube and testing is relatively simple. While many semiconductor devices can be similarly employed using only the data supplied in specification tables, more information is desirable on occasion. For instance, a transistor *may supposed* to be a substitute for another one. However, experience indicates considerable differences often exist and some circuit "tailoring" required. If the exact characteristics are known, matters are simplified considerably.

A number of transistor parameters can be determined simultaneously from a set of characteristic curves as shown in Fig. 21. This information can be obtained from a data sheet, or better still, by means of a curve tracer. A curve tracer applies a fixed control current or voltage to the device to be tested and displays another parameter (such as collector current) on the screen of an oscilloscope. Then the control parameter is incremented or "stepped" to a new value. The entire process is performed quickly enough so that the display appears as a family of curves as shown in Fig. 21. Curve tracers are available commercially. However, there are also models in kit form[1] and designs have been published in *QST.*[2]

Interpretation of the Curves

Some of the constants associated with a particular transistor that can be obtained from sets of curves are: dc beta (h_{FE}), ac beta (h_{fe}), reverse-saturation or "leakage" current (I_{CEO}), linearity, and breakdown voltage. Other characteristics such as f_T or input/output impedances have to be determined by different means.

[1]See *QST* for November, 1974, pp. 44-47.
[2]Ulrich, "A Semiconductor Curve Tracer for the Amateur," *QST* for August, 1971 (pp. 24-28) and September, 1971 (pp. 14-18, 31).

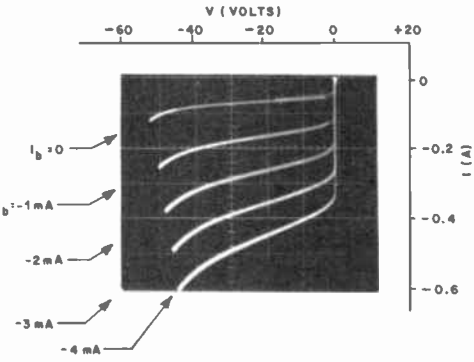


Fig. 21 — Curve-tracer plot of a medium-power pnp switching transistor.

Referring to Fig. 21, h_{FE} can be determined as follows. At −10 V, the collector current is −150 mA for a base current of −1 mA giving a dc beta (h_{FE}) of 150. However, if the base current is increased to −4 mA, the collector current is 400 mA and the ratio of collector current to base current (h_{FE}) is 100. On the other hand, at a collector voltage of −40, h_{FE} for a base current of 1 mA is 200.

The variations in h_{FE} illustrate one of the difficulties mentioned earlier. It would be impossible to specify any single value of h_{FE} for this particular transistor and some sort of "average" might be listed. However, a substitute model would likely have entirely different characteristics even though the specified h_{FE} might be the same. Similar considerations hold for the ac beta (h_{fe}) which is the *change* in collector current *vs* a corresponding change in base current. However, the variations are not as great as for h_{FE}. At −30 V, the collector current changes approximately 100 mA when the base current is changed from −1 to −2 mA. The corresponding value of h_{fe} would then be 100. Since the curves are more or less parallel over the useful operating range of the transistor, this means the value of h_{fe} remains the same.

Note that for zero base current, the collector current is not zero. The collector current for the zero base current curve is then equal to the leakage or reverse-saturation current. For the transistor under test, I_{CEO} varies from 50 to over 100 mA. However, during this test, the transistor was allowed to *heat up* and I_{CEO} was much smaller initially. This is why proper heat sinks are required with transistors designed to provide moderate to large amounts of power.

For values of collector voltage above 40, the collector current begins to increase rapidly. This indicates the breakdown voltage for the transistor is being approached and operation should be kept to within 30 volts or so. Although the transistor curves shown in Fig. 21 are for a switching type, the linearity would be acceptable for power-amplifier operation also. The linearity can be evaluated by noting the spacing between curves

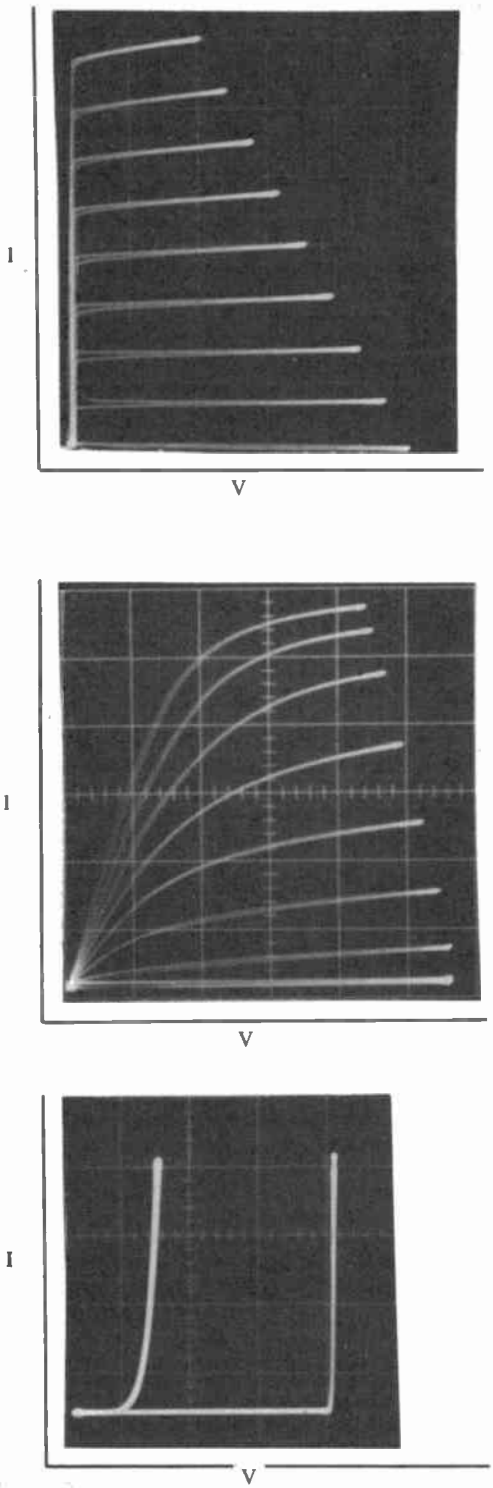


Fig. 22 — Curve-tracer plots for an npn small-signal bipolar transistor (A), a junction FET (B), and two Zener diodes (C).

and their straightness. Curves that change a great deal for equal increments in base current or ones that are not straight indicate poor linearity.

Other Devices

The curves shown in Fig. 21 are for a pnp power switching transistor. A similar curve is shown in Fig. 22A for a small-signal npn transistor of the 2N222 class. Judging from the curves for the latter, linearity should be very good for this particular model.

In Fig. 22B, a set of curves for a field-effect transistor (FET) is illustrated. While the horizontal and vertical axes are the same as for the bipolar types (current *vs* voltage), each curve is for a corresponding value of gate *voltage*.

The curve of Fig. 22B is for a n-channel junction FET. This type of FET has a characteristic curve very similar to the pentode vacuum tube. The curve at the top of the display is for zero gate voltage and the ones below represent the current-voltage characteristics for corresponding *negative* values of gate voltage. In the case of a pentode, a similar set of curves would be obtained for negative values of grid voltage.

Another common type of field-effect transistor is the MOSFET (*M*etal-*O*xidized *S*emiconductor FET) and a schematic diagram of a typical model is shown in Fig. 24. The version shown is a dual-gate MOSFET with gate protection in the form of back-to-back Zener diodes. (Since the oxide film is so thin, very small static charges are capable of puncturing this insulator. The diodes limit the voltage buildup by conducting if a rating value is exceeded. Below this value, the diodes do not affect circuit operation.) Under normal operating conditions, one gate is used as a gain-control electrode with the other one constituting the signal input.

Diodes can also be checked on a curve tracer and a double-exposure display for two different Zener diodes is shown in Fig. 22C. These curves illustrate a point often overlooked with Zener diodes used in regulator applications. The curve at the left is for a diode in the 4-volt class. The "soft" knee indicates considerable current must flow before the diode approaches a constant-voltage characteristic. On the other hand, the curve at the right has a very sharp turn-on and is for a 10-volt Zener diode. As the voltage rating on such diodes increases, the device characteristic approaches one that is ideal for regulation purposes.

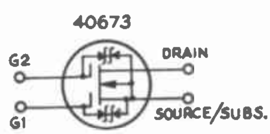
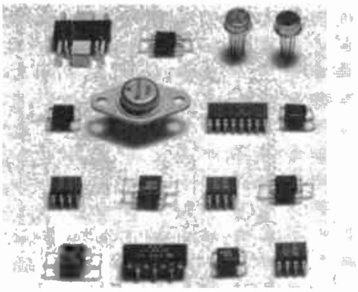


Fig. 23 — Schematic diagram of a gate-protected MOSFET. Back-to-back Zener diodes are bridged internally from gates 1 and 2 to the source/substrate element.



A collection of modern ICs. Various case styles of metal and epoxy materials are illustrated.

INTEGRATED CIRCUITS

Just as the term implies, integrated circuits (ICs) contain numerous components which are manufactured in such a way as to be suitably interconnected for a particular application, and on one piece of semiconductor base material. The various elements of the IC are comprised of bi-polar transistors, MOSFETs, diodes, resistances, and capacitances. There are often as many as ten or more transistors on a single IC chip, and frequently their respective bias resistors are formed on the chip. Generally speaking, ICs fall into four basic categories – **differential amplifiers, operational** amplifiers, diode or transistor arrays, and logic ICs.

IC Structures

The basic IC is formed on a uniform chip of n-type or p-type silicon. Impurities are introduced into the chip, their depth into it being determined by the diffusion temperature and time. The geometry of the plane surface of the chip is determined by masking off certain areas, applying photochemical techniques, and applying a coating of insulating oxide. Certain areas of the oxide coating are then opened up to allow the formation of interconnecting leads between sections of the IC. When capacitors are formed on the chip, the oxide serves as the dielectric material. Fig. 4-23 shows a representative three-component IC in both pictorial and schematic form. Most integrated circuits are housed in TO-5 type cases, or in flat-pack epoxy blocks. ICs may have as many as 12 or more leads which connect to the various elements on the chip.

Types of IC Amplifiers

Some ICs are called differential amplifiers and others are known as operational amplifiers. The basic differential-amplifier IC consists of a pair of transistors that have similar input circuits. The inputs can be connected so as to enable the transistors to respond to the difference between two voltages or currents. During this function, the circuit effectively suppresses like voltages or currents. For the sake of simplicity we may think of the differential pair of transistors as a push-pull amplifier stage. Ordinarily, the differential pair of transistors are fed from a controlled, constant-current source (Q3 in Fig. 4-24A. Q1 and Q2 are the differential pair in this instance). Q3 is commonly called a transistor current sink. Excellent balance exists between the input terminals of differential amplifiers because the base-to-emitter voltages and current gains (beta) of

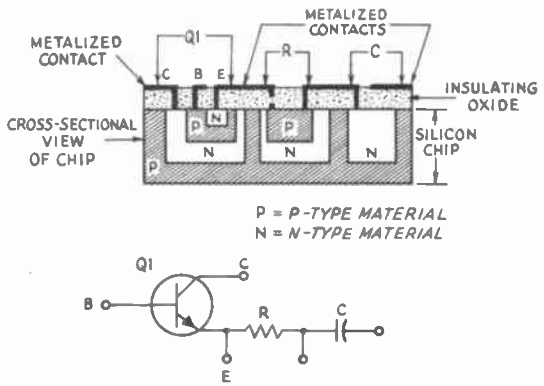


Fig. 4-23 – Pictorial and schematic illustrations of a simple IC device.

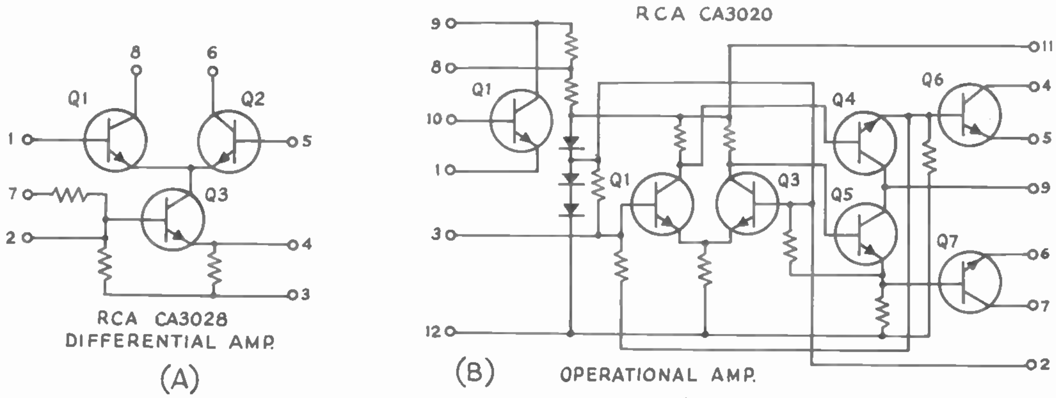
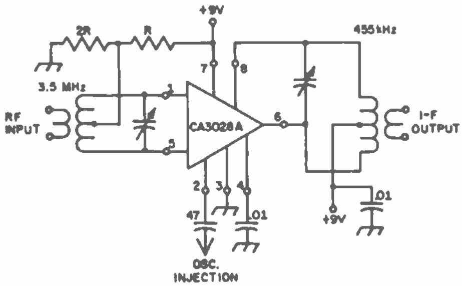
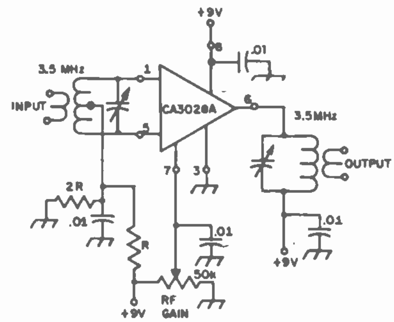


Fig. 4-24 – At A, a representative circuit for a typical differential IC. An Operational Amplifier IC is illustrated at B, also in representative form.



BALANCED MIXER

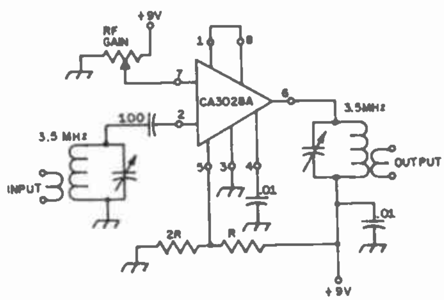
(A)



BALANCED DIFFERENTIAL AMPLIFIER

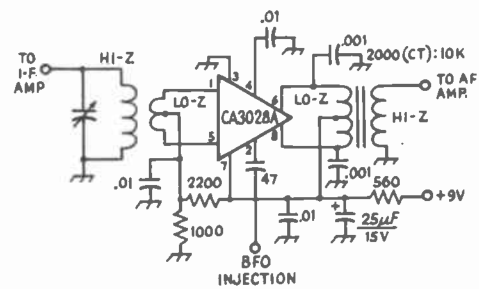
(B)

Fig. 4-25 — Some typical circuit applications for a differential amplifier IC. The internal circuit of the CA3028A IC is given in Fig. 4-24 at A.



CASCODE AMP.

(C)



PRODUCT DETECTOR

(D)

the two transistors are closely matched. The match results from the fact that the transistors are formed next to one another on the same silicon chip.

Differential ICs are useful as linear amplifiers from dc to the vhf spectrum, and can be employed in such circuits as limiters, product detectors, frequency multipliers, mixers, amplitude modulators, squelch, rf and i-f amplifiers, and even in signal-generating applications. Although they are designed to be used as differential amplifiers, they can be used in other types of circuits as well, treating the various IC components as discrete units.

Operational-amplifier ICs are basically very-high-gain direct-coupled amplifiers that rely on feedback for control of their response characteristics. They contain cascaded differential amplifiers of the type shown in Fig. 4-24A. A separate output stage, Q6–Q7, Fig. 4-24B, is contained on the chip. Although operational ICs can be successfully operated under open-loop conditions, they are generally controlled by externally applied negative feedback. Operational amplifiers are most often used for audio amplification, as frequency-shaping (peaking, notching, or bandpass) amplifiers, or as integrator, differentiator, or comparator amplifiers.

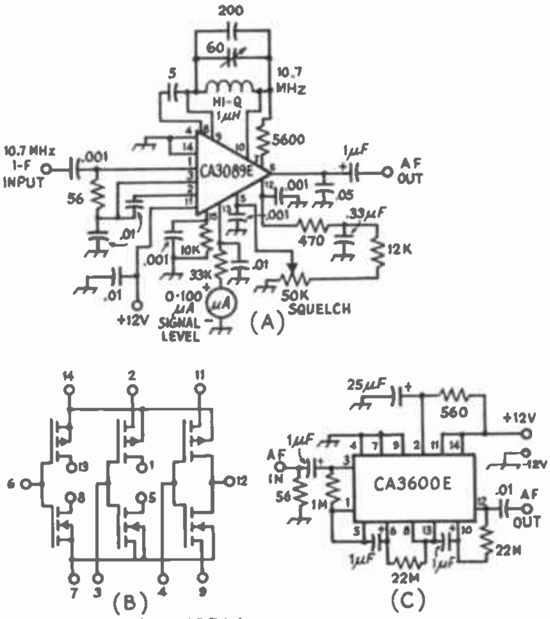
Diode-ICs are also being manufactured in the same manner as outlined in the foregoing section. Several diodes can be contained on a single silicon wafer to provide a near-perfect match between diode characteristics. The diode arrangement can take the form of a bridge circuit, series-connected groups, or as separate components. Diode ICs of this kind are useful in balanced-modulator circuits, or to any application requiring closely matched diodes.

Fig. 4-25 demonstrates the versatility of just one type of IC, an RCA CA3028A differential amplifier. Its internal workings are shown in Fig. 4-24A, permitting a comparison of the schematic diagram and the block representations of Fig. 4-25. The circuit at B in Fig. 4-25 is characterized by its high input and output impedances (several thousand ohms), its high gain, and its stability. This circuit can be adapted as an audio amplifier by using transformer or RC coupling. In the circuits of B and C terminal 7 is used to manually control the rf gain, but agc can be applied to that terminal instead. In the circuit at D the CA3028A provides low-noise operation and exhibits good conversion gain as a product detector. The CA3028A offers good performance from dc to 100 MHz.

PRACTICAL CONSIDERATIONS

Some modern-day ICs are designed to replace nearly all of the discrete components used in earlier composite equipment. One example can be seen in the RCA CA3089E flat-pack IC which contains nearly the entire circuit for an fm receiver. The IC contains 63 bipolar transistors, 16 diodes, and 32 resistors. The CA3089E is designed for an i-f of 10.7 MHz and requires but one outboard tuned circuit. The chip consists of an i-f

Fig. 4-26 — The circuit at A shows practical component values for use with the CA3089E fm subsystem IC. A COS/MOS array IC is illustrated at B in schematic-diagram form. It consists of three complementary-symmetry MOSFET pairs. The illustration at C shows how the CA3600E can be connected in cascade to provide at least 100 dB of audio amplification.



RCA 3600E COS/MOS ARRAY (GOOD TO 5 MHz)

CASCADE 100-dB AF AMPLIFIER

EXCEPT AS INDICATED, DECIMAL VALUES OF CAPACITANCE ARE IN MICROFARADS (μF); OTHERS ARE IN PICOFARADS (pF OR μμF); RESISTANCES ARE IN OHMS; k = 1000, M = 1000 000.

amplifier, quadrature detector, audio preamplifier, agc, afc, squelch, and a tuning-meter circuit. Limiting of -3 dB takes place at the 12-μV input level. When using an IC of this kind it is necessary only to provide a front-end converter for the desired frequency of reception, an audio amplifier, and a power supply (plus speaker, level controls, and meter).

There are two IC subsystem units designed for a-m receiver use. Each is similar in complexity to the CA3089E illustrated in Fig: 4-26. These components are identified as CA3088E and CA3123E. The latter is described in RCA Data File No. 631. Both ICs are readily adaptable to communications receiver use and should become popular building blocks for amateurs who desire compact, portable receiving equipment.

TRANSISTOR ARRAYS

Amateur designers should not overlook the usefulness of transistor- and diode-array ICs. These devices contain numerous bipolar or MOSFET transistors on a common substrate. In most instances the transistors can be employed as one would treat discrete npn devices. An entire receiver can be made from one transistor-array IC if one wishes to construct a compact assembly. The CA3049 is a dual independent differential rf/i-f amplifier chip with an f_T of 1.3 GHz. It is especially well suited to applications which call for double-balanced mixers, detectors, and modulators. Another device of similar usefulness is the CA3018A. The CA3045 should also be of interest to the amateur. Matched electrical characteristics of the transistors in these ICs offer many advantages not available when using discrete transistors. Fig. 4-27 shows the internal workings of the CA3018A and CA3045 ICs.

COS/MOS (complementary-symmetry metal-oxide silicon) ICs are becoming increasingly popular, and one RCA part, the CA3600E, contains an array of complementary-symmetry MOSFET pairs (three) which can be used individually or in cascade, as shown in Fig. 4-26 at B. Detailed information is given in RCA File No. 619. The CA3600E is a high-input impedance, micropower component which is suitable for use as a preamplifier, differential amplifier, op amp, comparator, timer, mixer, chopper, or oscillator.

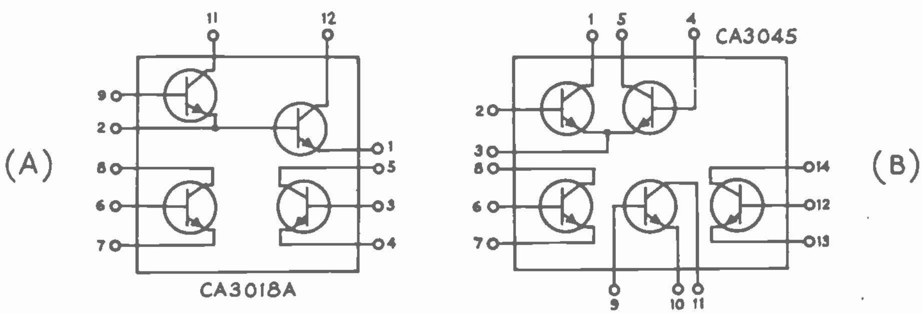


Fig. 4-27 — Transistor arrays offer unlimited application because several circuit combinations are possible. The CA3018A IC at A has a Darlington-connected pair plus two separate transistors. At B, two transistors are internally connected in a differential amplifier fashion. Three separate transistors are available for use in other functions. The arrays shown here are useful into the vhf spectrum.

One of the more interesting and useful array of ICs is the RCA CA3102E. It contains two differential pairs and two current-source transistors. The device is ideally suited for use in doubly balanced mixers, modulators, and product detectors. The CA3102E is excellent for use in the following additional applications: vhf amplifiers, vhf mixers, rf amp./mixer/oscillator combinations, i-f amplifiers (differential and/or cascode mode), synchronous detectors, and sense amplifiers. This IC is similar in configuration to the CA3049T array, but has an independent substrate connection which is common to an internal shield that separates the two differential amplifiers. The shield helps assure good isolation in applications where that feature is required.

F_T for CA3102E is in excess of 1000 MHz. Noise figure at 100 MHz single transistor is 1.5 dB, R_s = 500 ohms. Noise figure at 200 MHz cascode mode is 4.6 dB. Additional specifications can be found in *RCA Data File* No. 611. The CA3102E offers almost limitless possibilities for applications in amateur radio design work. The chip is manufactured in a 14-lead DIP package. The CA3049T comes encased in a standard TO-5 package.

DIGITAL-LOGIC INTEGRATED CIRCUITS

Digital logic is the term used to describe an overall design procedure for electronic systems in which "on" and "off" are the important words, not "amplification," "detection," and other terms commonly applied to most amateur equipment. It is "digital" because it deals with discrete events that can be characterized by digits or integers, in contrast with linear systems in which an infinite number of levels may be encountered. It is "logic" because it follows mathematical laws, in which "effect" predictably follows "cause."

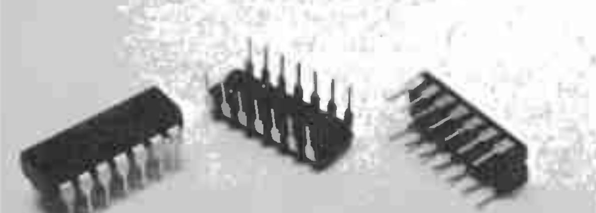
Just like linear integrated circuits, digital ICs are manufactured in such a way that the internal components are interconnected for particular applications. Packaging of the digital ICs is the same for their linear counterparts, with the full package range pictured earlier being used. From outward appearances, it would be impossible to tell the difference between the two types of ICs except from the identification numbers.

Linear ICs are constructed to respond to continuously variable or analog signals, such as in an amplifier. Digital devices, on the other hand, generally have active components operating only in either of two conditions – cutoff or saturation. Digital ICs find much application in on-off switching circuits, as well as in counting, computation, memory-storage, and display circuits. Operation of these circuits is based on binary mathematics, so words such as "one" and "zero" have come into frequent use in digital-logic terminology. These terms refer to specific voltage levels, and vary between manufacturers and devices. Nearly always, a "0" means a voltage near ground, while "1" means whatever the manufacturer specifies. One must distinguish between "positive logic" and "negative logic." In positive logic, a 1 is more positive than a 0, though both may be negative voltages. In negative logic, the reverse is true. Often the terms "high" and "low" are used in reference to these voltage levels. The definitions of these terms are the same for both positive and negative logic. A "high" is the most positive or least negative potential, while a "low" is the least positive or most negative.

For practical use in some applications it is desirable to convert binary data into decimal equivalents, such as in electronic counting and display systems. In other applications, such as for the graphic recording or metering of summations or products of integers, it is convenient to convert the digital data into analog equivalents. Specialized integrated circuits designed to perform these functions are also considered to be included in the digital-IC category.

LOGIC SYMBOLS

With modern microcircuit technology, hundreds of components can be packaged in a single case. Rather than showing a forest of transistors, resistors, and diodes, logic diagrams show symbols based on the four distinctive shapes given in Fig. 4-28 at A through D. These shapes may be "modified" or altered slightly, according to



From outward appearances, these three ICs appear to be identical. Although each is a *J-K* flip-flop, there are differences in their characteristics. Pictured at the left is a Texas Instruments SN74H72N integrated circuit, called a *J–K* master-slave flip-flop. Shown in the center is a Motorola MC1927P IC, which is a 120-MHz ac-coupled *J-K* flip-flop. Both of these ICs might be considered "universal" flip-flops, for they may be used in a variety of ways. Shown at the right is a Motorola MC726P, a simple *J-K* flip-flop.

specific functions performed. Examples are shown at E through H of Fig. 4-28.

The square, Fig. 4-28D and H, may appear on logic diagrams as a rectangle. This symbol is a somewhat universal one, and thus must be identified with supplemental information to indicate the exact function. Internal labels are usually used. Common identification labels are:

FF – Flip-flop
FL – Flip-flop latch
SS – Single shot
ST – Schmitt trigger.

Other logic functions may also be represented by the square or rectangle, and the label should adequately identify the function performed. Unique identifying shapes are used for gates and inverters, so these need no labels to identify the *function*. Hardware- or package-identification information may appear inside any of the symbols on logic diagrams.

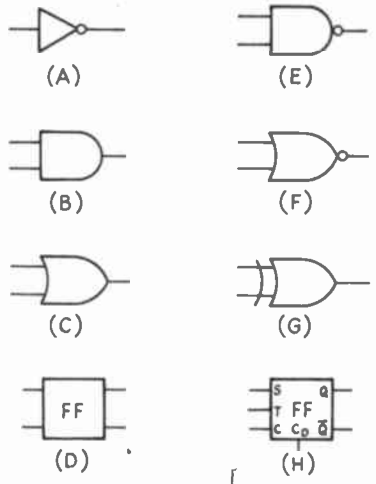


Fig. 4-28 – Distinctive symbols for digital logic diagrams. At A is shown an inverter, at B an AND gate, at C an OR gate, and at D a flip-flop. Additions to these basic symbols indicate specific functions performed. A small circle, for example, placed at the output point of the symbol, denotes that inversion occurs at the output of the device. Shown at E is an inverting AND or NAND gate, and at F is an inverting OR or NOR gate. At G is the symbol for an exclusive OR gate. The symbol at H represents a J-K flip-flop.

TYPES OF DIGITAL ICs

Digital integrated circuits perform a variety of functions, but these functions can generally be cataloged into just a few categories: gates, inverters, flip-flops, drivers and buffers, adders and subtractors, registers, and memories, plus the special-purpose ICs as mentioned earlier – decoders and converters. Some of these types, such as adders and subtractors, registers, and memories, find use primarily in computer systems. More universally used types of ICs are the inverters, gates and flip-flops.

Inverters

A single chip in one IC package may be designed to perform several functions, and these functions can be independent of each other. One example of an IC of this type is Motorola's MC789P, which bears the name, "hex inverter." This IC contains six identical inverter sections. The schematic diagram of one section is shown in Fig. 4-29A. In operation, 3.0 to 3.6 volts are applied between +Vcc and ground. For this device in positive-logic applications, a 0 is defined as any potential less than approximately 0.6 volt, and a 1 is any voltage greater than about 0.8. With a logic 0 applied at the input, the transistor will be at or near cutoff. Its output will be a potential near +Vcc, or a logic 1. If the 0 at the input is replaced by a 1, the transistor goes into saturation and its output drops nearly to ground potential; a 0 appears at the output. The output of this device is always the opposite or *complement* of the input logic level. This is sometimes called a NOT gate, because the input and output logic levels are *not* the same, under any conditions of operation.

Shown at the right in Fig. 4-29A is the logic symbol for the inverter. In all logic symbols, the connections for +Vcc and the ground return are omitted, although they are understood to be made. The proper connections are given in the manufacturer's data sheets, and, of course, must be made before the device will operate properly. In the case of all multiple-function ICs, such as the hex inverter, a single ground connection and a single +Vcc connection suffice for all sections contained in the package.

Gates

Another example of an IC containing several independent functions in one package is Motorola's MC724P, a quad 2-input gate. Four gates are contained in one chip. The schematic diagram and logic symbols for a gate section are shown at B in Fig. 4-29. As with the MC789P, a supply of 3.0 to 3.6 volts is used; for positive logic a 0 is a potential less than 0.6 volt, and a 1 is a potential greater than 0.8 volt. It may be seen from the schematic diagram that the two transistors have an independent input to each base, but they share a common collector resistor. Either transistor will be saturated with a logic 1 applied at its input, and a 0 output will result. A 0 at the input of either transistor will cause that transistor to be cut off, but a 1 at the opposite input will hold the output at 0. Thus, a 1 at either Input 1 *or* Input 2 will cause a 0 (or a NOT 1) to appear at the output. The NOT functions are usually written with a bar over them, so $\bar{1}$ means the same thing as NOT 1,

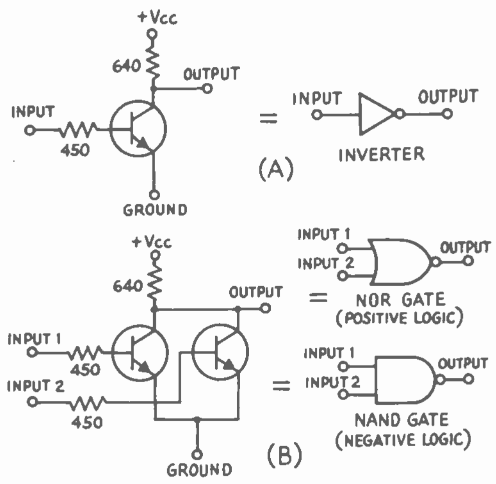


Fig. 4-29 — Digital circuits and their equivalent logic symbols. See text. Indicated resistor values are typical.

and is expressed as NOT 1 when reading the term. Logic-circuit operation can be expressed with equations. Boolean algebra, a form of binary mathematics, is used. These equations should not be confused with ordinary algebraic equations. The logic equation for the operation of the circuit in Fig. 4-28A is $1 \vee 1 = \overline{1}$. The little ν means OR. Sometimes "+" is used instead of "ν." In plain words, the equation says that a 1 at Input 1 *or* Input 2 will yield a NOT 1 at the output. This is equivalent of saying the circuit is an inverting OR gate, or a NOT OR gate. This latter name is usually contracted to NOR gate, the name by which the circuit is known.

If the circuit of Fig. 4-29B is used with negative logic, circuit operation remains the same; only the definitions of terms are changed. A logic 1, now, is a voltage level less than 0.6, and a 0 is a level greater than 0.8 volt. If a logic 1 is applied at both inputs, 1 *and* 2, both transistors will be cut off. The output is near +Vcc, which is a logic 0 or NOT 1. The equation for this operation is $1 \cdot 1 = \overline{1}$, where the dot means AND. In this way, with negative logic, the circuit becomes an inverting AND gate, or a NOT AND gate or, more commonly, a NAND gate. Manufacturers' literature frequently refers to this type of device as a NAND/NOR gate, because it performs either function.

Flip-Flops

It is not necessary for the various functions on a single chip to be identical. Motorola's MC780P IC, a decade up-counter, contains four flip-flops, an inverter, and a 2-input gate. These functions are interconnected to provide divide-by-10 operation, with ten input pulses required for every output pulse which appears. Intermediate outputs are also provided (in binary-coded form) so that the number of pulses which have entered the input can be determined at all times. These binary-coded decimal (BCD) outputs, after decoding, may be used to operate decimal-readout indicators.

The term, medium-scale integration (MSI) is frequently applied to ICs such as this decade up-counter, which contains the equivalent of 15 or more gates on a single chip. Large-scale integration (LSI) describes ICs containing the equivalent of 100 or more gates on a single chip. These terms, when applied to a particular IC, convey an idea of the complexity of the circuitry.

A flip-flop is a device which has two outputs that can be placed in various 1 and 0 combinations by various input schemes. Basically, one output is a 1 when the other is a 0, although situations do occur (sometimes on purpose) where both outputs are alike. One output is called the Q output, or "set" output, while the other is the $\overline{Q}$ (NOT Q) or "reset" output. If $Q = 1$ and $\overline{Q} = 0$, the flip-flop is said to be "set" or in the "1 state," while for the reverse, the flip-flop is "reset," or "cleared," or in the "0 state." A variety of inputs exist, from which the flip-flops derive their names.

The *R-S* flip-flop is the simplest type. Its outputs change directly as a result of changes at its inputs. The type *T* flip-flop "toggles," "flips," or changes its state during the occurrence of a *T* pulse, called a clock pulse. The *T* flip-flop can be considered as a special case of the *J-K* flip-flop described later. The type *D* flip-flop acts as a storage element. When a clock pulse occurs, the complementary status of the *D* input is transferred to the *Q* output. The flip-flop remains in this state even though the input may change, as it can change states only when a clock pulse occurs.

Although there is some disagreement in the nomenclature, a *J-K* flip-flop is generally considered to be a toggled or clocked *R-S* flip-flop. It may also be used as a storage element. The *J* input is frequently called the "set" or *S* input; the *K* is called the "clear" or *C* input (not to be confused with the clock input). The clock input is called *T*, as in the type *T* flip-flop. A clear-direct or C_D input which overrides all other inputs to clear the flip-flop to 0 is provided in most *J-K* flip-flop packages. The logic symbol for the *J-K* flip-flop is shown in Fig. 4-28H. A simple *J-K* flip-flop circuit contains 13 or 14 transistors and 16 or 18 resistors.

There are essentially two types of flip-flop inputs, the dc or level-sensitive type, and the "ac" or transistion-sensitive type. It should not be concluded that an ac input is capacitively coupled. This was true for the discrete-component flip-flops, but capacitors just do not fit into microcircuit dimensions. The construction of an ac input uses the "master-slave" principle, where the actions of a master flip-flop driving a slave flip-flop are combined to produce a shift in the output level during a transit of the input.

DIGITAL-LOGIC IC FAMILIES

There are seven categories or families of which nearly all semiconductor digital ICs are members.

Each family has its own inherent advantages and disadvantages. Each is geared to its own particular market, meeting a specific set of needs.

Resistor-Transistor Logic – RTL

RTL is known primarily for its economy. It is well named, since it contains resistors and transistors exclusively. The circuits of Fig. 4-29 are RTL. Advantages of the RTL family are economy, ease of use in system designs, ease of interface with discrete components, and high speed–power product. There are a wide number of functions available in this family. Disadvantages are low immunity to voltage noise (transients, rf pickup, and the like), and relatively low fanout (the number of loads that may be connected to an output before performance is degraded). The RTL family requires a supply of 3.0 to 3.6 volts.

Diode-Transistor Logic – DTL

DTL ICs contain diodes, as well as resistors and transistors. Early DTL ICs used design criteria carried over from the use of discrete components, where diodes were inexpensive compared to transistors. These ICs required negative and positive voltage sources. Later DTL ICs are of a modified design which lends itself more easily to IC processing. Performance characteristics are also enhanced, with less input current being required, and only a single voltage source needed. Members of the DTL family are limited generally to gates. Advantages of this family are low power dissipation, compatibility with TTL (see later section), low cost, ease of use in system design, ease of interface with discrete circuits, and relatively high fanout. DTL disadvantages are low noise immunity, especially in the high state where the input impedance is relatively high, rapid change in voltage thresholds with temperature, speed slowdown with capacitive loading, and lower speed capabilities than some other families. The DTL family requires a supply voltage of 5.

High-Threshold Logic – HTL

HTL devices are designed for high noise immunity. The circuit form is the same as DTL except that breakdown (Zener) diodes are used at the inputs. Higher supply voltages and higher power dissipations accompany the HTL family. These ICs find applications in industrial environments and locations likely to have high electrical noise levels. Advantages are high noise immunity, stable operation over very large temperature ranges, interfaces easily with discrete components, electromechanical components, and linear functions (operational amplifiers and multipliers), and a constant threshold-versus-temperature characteristic. Disadvantages are higher cost than other families, and relatively high power dissipation. The HTL family requires a supply voltage of 15.

Transistor-Transistor Logic –TTL

TTL has characteristics that are similar to DTL, and is noted for many complex functions and the highest available speed of any saturated logic. TTL may be thought of as a DTL modification that results in higher speed and driving capability. It is noted for better noise immunity than that offered by DTL, and is more effective for driving high-capacitance loads because of its low output impedance in both logic states. TTL ICs fall into two major categories – medium speed and high speed. Various manufacturing techniques are used to increase the speed, including gold doping and incorporation of high-speed Schottky diodes on the chip. Another advantage of TTL is that it is compatible with various other families. Multiple sources and extensive competition have resulted in low prices for TTL devices. Disadvantages are that more care is required in the layout and mechanical design of systems because of its high speed, and additional capacitors are required for bypassing because of switching transients. The TTL family requires a supply of 5 volts.

Emitter-Coupled Logic – ECL

ECL has the highest speed of any of the logic forms. It is sometimes called current-mode logic. This family is different from standard saturating logic in that circuit operation is analogous to that of some linear devices. In this case, the transistors do not saturate and the logic swings are reduced in amplitude. Very high speeds can be attained because of the small voltage swings and the use of nonsaturating transistors. The input circuitry of ECL devices is of the nature of a differential amplifier, resulting in much higher input impedances than saturated-logic devices. Emitter-follower outputs are of low impedance with high fanout capabilities, and are suited for driving 50-ohm transmission lines directly. Disadvantages are higher power dissipation, less noise immunity than some saturated logic, translators are required for interfacing with saturated logic, and slowed-down operation with heavy capacitive loading. The ECL family requires a supply of -5.2 volts.

Metal-Oxide Semiconductor (MOS)

Digital MOS devices are gaining significance in industrial applications, with p-channel or P-MOS ICs being the most popular. Large, complex repetitive functions, such as long shift registers and high-capacity memories, have proved very practical. Gates and basic logic circuits have not become as popular, because they exhibit lower drive capability than other IC families. Input impedances to these devices are essentially capacitive (an open circuit for dc). This feature allows very high fanout where speed is not a consideration. Bidirectional devices give more flexibility to the circuit designer. P-MOS technology results in the lowest cost per bit for memories and long shift registers, because many more functions can be contained on a given chip size than in bipolar devices. Disadvantages are that devices must be handled more carefully than bipolar ICs because excessive static electricity can destroy the narrow gate oxide, even with internal breakdown-diode input protection. Drive capability is limited because of the high output impedances characteristic of these devices. Two power supplies are usually required. The P-MOS family requires supplies of -13 and -27 volts.

Complementary Metal-Oxide Semiconductor – CMOS

CMOS technology employs both p-channel and n-channel devices on the same silicon substrate. Both types are enhancement-mode devices; that is, gate voltage must be increased in the direction that inverts the surface in order for the device to conduct. Only one of the two complementary devices of a circuit section is turned on at a time, resulting in extremely low power dissipation. Dissipation is primarily from the switching of devices through the active region and the charging and discharging of capacitances. Advantages are low power dissipation, good noise immunity, very wide power supply voltage variations allowed, high fanout to other CMOS devices, and full temperature-range capabilities. Disadvantages are restricted interfacing capabilities because of high output impedance, and medium to high cost. The CMOS family requires a supply of 1.5 to 16 volts, 10 volts being nominal.

IC Family Groups

The popular digital-logic families have several groups where basic designs have been modified for medium speed, high speed, or low power consumption. The TTL family ICs have single-letter designators added to the part number to identify the group: S – Schottky high speed, H – medium speed, L – low power. ECL logic, as yet, has no such simple identification system. Manufacturers group their ECL products by propagation delay, an expression of the maximum speed at which the logic device will operate. Motorola, for example, calls the ECL group with 8-ns delay MECL. MECL II has a speed of 4 ns; MECL 10,000, 2 ns; and MECL III, 1.1 ns. With a propagation delay of 1 ns, operation at 300 MHz is possible.

Special Digital ICs

In addition to the logic families, many special-purpose digital ICs are available to accomplish specific tasks. A divide-by-10 circuit, such as the Fairchild U6B95H9059X, operates up to 320 MHz and is used as a prescaler to extend the range of a frequency counter. This IC has been designed to operate with low-level input signals, typically 100 mV at 150 MHz.

Large MOS arrays are being used for a number of applications which require the storage of logic instructions. These ICs are called memories. Instructions are stored in the memory by a process named programming. Some memories can be programmed only once; they are called ROMs (Read-Only Memory). ROMs must be read in sequence, but another group of devices called RAMs (Random-Access Memory) can be used a section at a time. Both ROMs and RAMs are also made in reprogrammable versions, where the information stored in the memory can be changed as desired. These models are named PROMs and PRAMs, respectively.

Large memory arrays are often used for the generation and conversion of information codes. One IC can be programmed to convert the 5-level RTTY code to the 8-level ASCII code popular in computer devices. National Semiconductor manufactures a single IC which generates the entire 56-character 8-level code. Several ICs are now available for character generation where letters and numerals are produced for display on an oscillograph screen.

OTHER DEVICES

THE UNIJUNCTION TRANSISTOR

Unijunction transistors (UJT) are being used by amateurs for such applications as side-tone oscillators, sawtooth generators, pulse generators, and timers.

Structurally, the UJT is built on an n-type silicon bar which has ohmic contacts – base one (B1) and base two (B2) – at opposite ends of the bar. A rectifying contact, the emitter, is attached between B1 and B2 on the bar. During normal operation B1 is grounded and a positive bias is supplied to B2. When the emitter is forward biased, emitter current will flow and the device will conduct. The symbol for a UJT is given in Fig. 4-30 at C. A circuit showing a typical application in which a UJT is employed is shown in Fig. 4-30.

SILICON CONTROLLED RECTIFIERS

The silicon controlled rectifier, also known as a Thyristor, is a four-layer (p-n-p-n or n-p-n-p) three-electrode semiconductor rectifier. The three terminals are called anode, cathode and gate, Fig. 4-28B.

The SCR differs from the silicon rectifier in that it will not conduct until the voltage exceeds the *forward breakover* voltage. The value of this voltage can be controlled by the gate current. As the gate current is increased, the value of the forward breakover voltage is decreased. Once the rectifier conducts in the forward direction, the gate current no longer has any control, and the rectifier behaves as a low-forward-resistance diode. The gate regains controls when the current through the rectifier is cut off, as during the other half cycle.

The SCR finds wide use in power-control applications and in time-delay circuits. SCRs are available in various voltage and wattage ratings.

TRIACS

The triac, similar to the SCR, has three electrodes – the main terminal (No. 1), another

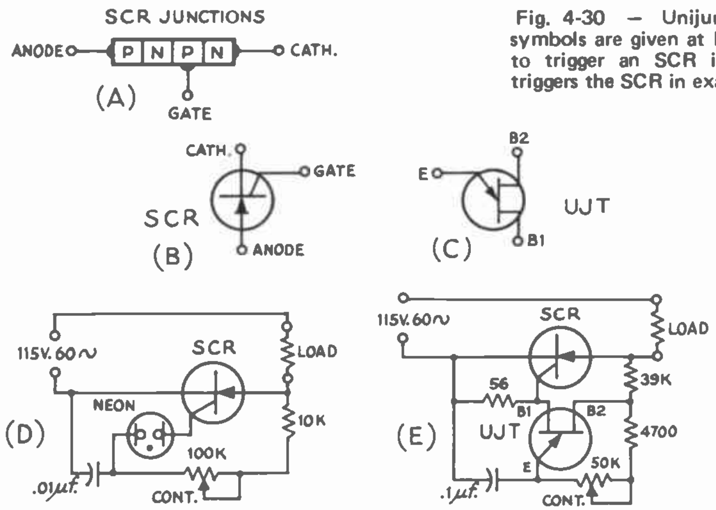


Fig. 4-30 – Unijunction transistor and SCR symbols are given at B and C. A neon lamp is used to trigger an SCR in the circuit at D. A UJT triggers the SCR in example E.

main terminal (No. 2), and a gate. The triac performs in the same manner as the SCR, but for either polarity of voltage applied to its main terminals. The SCR, as mentioned in the foregoing, conducts only during one half the sine-wave cycle. When an SCR is used in a motor-speed control, therefore, the motor cannot be brought up to full speed. The triac, however, does trigger on both halves of the cycle. Therefore, triacs are preferred to SCRs in many control circuits. The triac can be regarded as a device in which two SCRs are employed in parallel and oriented in opposite directions as shown in the drawing of Fig. 4-30. An example of a motor-speed control which uses a triac is given in the construction chapter of this book.

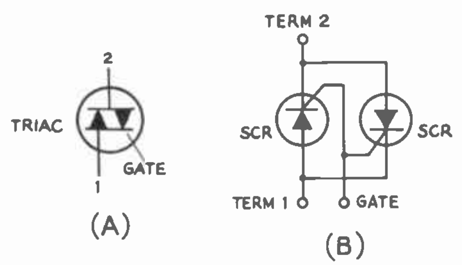


Fig. 4-31 – The symbol for a triac is given at A. The illustration at B shows how a triac compares to two SCRs connected for the same performance offered by a triac, thus permitting conduction during both halves of the cycle.

OPERATIONAL AMPLIFIERS

Early analog computers used amplifier blocks which became known as operational amplifiers, or simply op amps. Operational amplifiers can be constructed using tubes or transistors, and as hybrid or monolithic integrated circuits. The monolithic IC has become the most popular type of op amp. Today op-amp ICs cost approximately one dollar for the preferred types. They are used as building blocks in many circuit applications.

The op amp is a dc-coupled multistage linear amplifier which, in an ideal device, would have infinite input impedance and infinite gain. While the ideal op amp remains an unobtainable goal, voltage gains of 100,000 or more can be achieved. FET-input op amps have sufficiently high input impedance that the current required from the driving source is measured in pA (µµA).

Gain and Feedback

The internal circuit of a popular op-amp IC, the Fairchild µA741 (also produced by most other semiconductor manufacturers) is shown in Fig. 4-32. Two inputs are provided, one the complement or inverse of the other. An amplifier with two such inputs is known as a differential amplifier. If a small positive voltage is applied to the noninverting (+) terminal, it will produce a positive output. The same positive voltage applied to the inverting (–) terminal will result in a negative output. If the same voltage was applied to both terminals, the output would be zero. Both inputs can be used, called the differential connection, or one can be returned to ground for single-ended operation. In practical ICs, the output may not be exactly zero when both inputs are at zero potential. Any output under these conditions is called offset – some op amps have provision for connections to an external control which compensates for any offset voltage by applying bias current to the input transistors. The offset connections for the µA741 are shown in Fig. 4-32. Op amps are available in all of the popular IC packages; consult

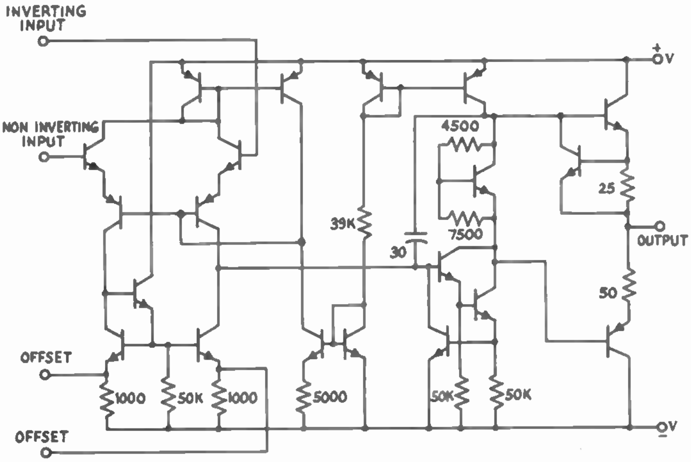


Fig. 4-32 — Internal circuit of a μA741 operational amplifier.

the manufacturer's literature for pin connections. Usually the pin connections are not the same for a particular device when it is made up in different package styles.

For most applications the full gain of the op amp is not used. Feedback is employed, as shown in Figs. 4-34A and B.. The addition of a resistive divider network, Ro-Ri, causes negative feedback by allowing part of the output voltage to be applied to the inverting input. The gain of the device will be equal to the sum of Ro and Ri, divided by the value of Ri. Feedback can be applied in a similar manner for a noninverting amplifier, Fig. 4-33B.. The voltage summer, Fig. 4-33C, provides an output voltage which is the sum of all input voltages multiplied by the gain of the operational amplifier. This circuit is often employed as an audio mixer. Fig. 4-33D shows the voltage-follower connections. The load at the output of this circuit can draw a large current while the input draws almost no current. The output voltage follows the level of the input potential almost exactly. The output of the differentiator (Fig. 4-33E) is proportional to the rate of change of the input voltage, while the integrator (Fig. 4-33F) averages the level of a voltage that varies over a short period of time. A differential connection of a single op amp is shown at G.

Stability

Because op amps are high-gain devices with frequency response from dc to several megahertz,

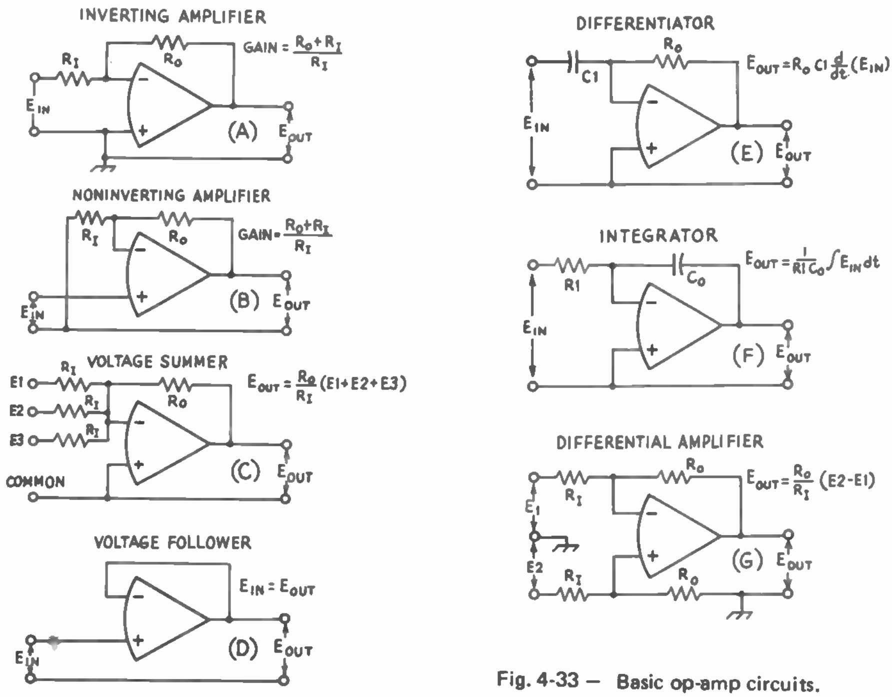
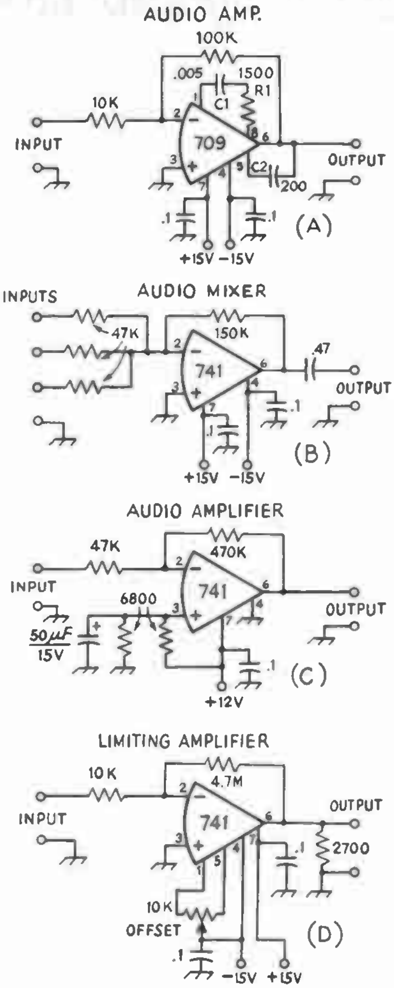


Fig. 4-33 — Basic op-amp circuits.



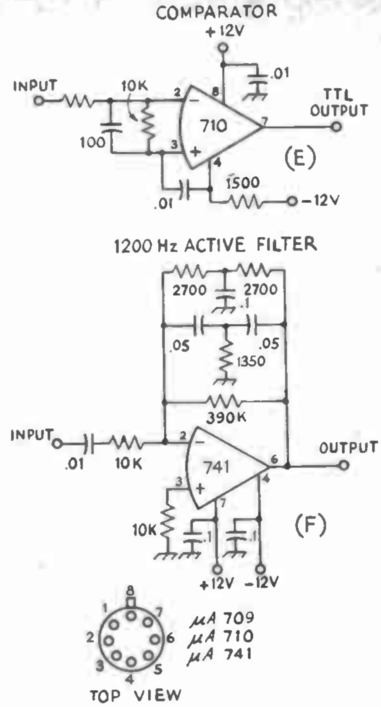


Fig. 4-34 – Some typical applications of operational amplifiers. The pin numbers shown are all for the metal can (TO-99) package.

oscillation can occur. In any op-amp circuit layout, the inputs should be well isolated from the output. Input leads should be kept as short as possible. Supply-voltage terminals should be bypassed with 0.1- or .01-µF capacitors. As the frequency is increased, the stages within an op amp will introduce phase shift. If the phase shift in the amplifier reaches 180 degrees before the gain has decreased to unity, the amplifier will be unstable. Some op amps, such as the µA709 of Fig. 4-34A. require an external compensation network, R1-C1, to reduce the gain of the device at hf. Others, the µA741 of Fig. 4-34B for example, contain internal compensation and, thus, require no additional components to assure stability.

Applications

Most monolithic op-amp ICs require supply voltages of plus 5 to 15 and minus 5 to 15. Practical examples of an audio amplifier and audio mixer are given in Fig. 4-33A and B,, respectively. In some amateur applications, the dual-polarity requirement can be eliminated by using a resistive divider to bias the noninverting input as indicated in Fig. 4-34C. If the amplifier is intended to be used as a limiting device (the input stage of an RTTY demodulator is an example) an offset control should be added to allow adjustment for equal clipping of the negative and positive peaks (Fig. 4-34D).

Another popular use for the op amp is as a comparator – see Fig. 4-34E. A comparator is used to indicate when a difference exists between a reference voltage and an input voltage. The output of the comparator will swing from its maximum positive voltage to maximum negative when the input exceeds the reference (zero voltage if the reference is zero). A number of op amps optimized for comparator service are available; they are often used as interface devices between linear and digital circuits. The operational amplifier is often employed in *active filters*, which use *RC* components to provide low-pass, high-pass, and bandpass characteristics. A simple illustration, an *RC* filter network tuned to 1200 Hz connected in parallel with the feedback resistor, is given in Fig. 4-34F. This design is for low *Q* giving a characteristic suitable for a cw receiver. The gain at resonance is approximately 40. Additional information about active filters and other op-amp circuits is available in the publications listed in the bibliography at the end of this chapter.

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

FIELD-EFFECT TRANSFER 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

Semiconductor Bibliography

Garrett, "Integrated-Circuit Digital Logic Families," in three parts, *IEEE Spectrum*, October, November, and December, 1970.

Heilweil, *Introduction to Boolean Algebra and Logic Design*, McGraw-Hill, 1964.

Maley, *Manual of Logic Circuits*, Prentice-Hall, 1970.

Pike, "The Operational Amplifier," Parts I and II, *QST*, August and September, 1970.

Pos, "Digital Logic Devices," *QST*, July, 1968.

Pos, "Integrated-Circuit Flip-Flops," *QST*, February, 1971.

RCA Transistor, Thyristor, and Diode Manual, Series SC-14, RCA, Harrison, NJ 07029.

RCA Power Circuits, DC to Microwaves, Series SP-51, RCA, Harrison, NJ 07029.

RCA Linear Integrated Circuits, Series IC-42, RCA, Harrison, NJ 07029.

RCA Hobby Circuits Manual, Series HM-91, RCA, Harrison, NJ 07029.

Solid-State Communications McGraw-Hill.

Transistor Circuit Design, McGraw-Hill.

Malmstadt and Enke, *Digital Electronics for Scientists*, W. A. Benjamin, Inc., New York, NY 10016.

Malmstadt and Enke, *A Laboratory Workbook* (computer logic). W. A. Benjamin, Inc., New York, NY 10016.

Chapter 5

AC-Operated Power Supplies

Power-line voltages have been "standardized" throughout the U.S. at 115 – 230 V in residential areas where a single voltage phase is supplied. These figures represent nominal voltages, however. "Normal" line voltage in a particular area may be between approximately 110 and 125 volts, but generally will be above 115 volts. In many states, the service is governed by the state's public utilities commission. The voltage average across the country is approximately 117 volts. Source of information: Edison Electric Company (an association of power companies), New York, NY.

The electrical power required to operate amateur radio equipment is usually taken from the ac lines when the equipment is operated where this power is available; in mobile operation the prime source of power is usually the storage battery.

The high-voltage dc for the plates of vacuum tubes used in receivers and transmitters is derived from the commercial ac lines by the use of a transformer-rectifier-filter system. The transformer changes the voltage of the ac to a suitable value, and the rectifier converts it to pulsating dc. The filter reduces the pulsations to a suitably low level, and may have either a capacitor input or a choke input, depending on whether a shunt capacitor or a series inductor is the first filter element. Essentially pure direct current is required to prevent hum in the output of receivers, speech amplifiers, modulators and transmitters. In the case of transmitters, a pure dc plate supply is also dictated by government regulations. If a constant supply voltage is required under conditions of changing load or ac line voltage, a regulator is used following the filter.

When the prime power source is dc (a battery), the dc is first changed to ac, and is then followed by the transformer-rectifier-filter system. Additional information on this type of supply is contained in Chapter 10.

The cathode-heating power can be ac or dc in the case of indirectly heated cathode tubes, and ac or dc for filament-type tubes if the tubes are operated at a high power level (high-powered audio and rf applications). Low-level operation of filament-type tubes generally requires dc on the filaments if undue hum is to be avoided.

Occasionally transformerless power supplies are used in some applications (notably in the ac-dc type of broadcast receiver). Such supplies operate directly from the power line, and it is necessary to connect the chassis or common-return point of the circuit directly to one side of the ac line. This type of power supply represents an extreme shock hazard when the equipment is interconnected with other apparatus in the amateur station, or when the chassis is exposed. For safety reasons, an isolation transformer should be used with such equipment when it is present in an amateur station.

POWER-LINE CONSIDERATIONS

POWER LINE CONNECTIONS

In most residential systems, three wires are brought in from the outside to the distribution board, while in other systems there are only two wires. In the three-wire system, the third wire is the neutral which is grounded. The voltage between the other two wires normally is 230, while half of this voltage (115) appears between each of these wires and neutral, as indicated in Fig. 5-1A. In systems of this type, usually it will be found that the 115-volt household load is divided as evenly as possible between the two sides of the circuit, half of the load being connected between one wire and the neutral, while the other half of the load is connected between the other wire and neutral. Heavy appliances, such as electric stoves and heaters, normally are designed for 230-volt operation and therefore are connected across the two ungrounded wires. While both ungrounded wires should be fused, a fuse should never be used in the wire to the neutral, nor should a switch be used in this side of the line. The reason for this is that opening the neutral wire does not disconnect the equipment. It simply leaves the equipment on one side of the 230-volt circuit in series with whatever load may be across the other side of the

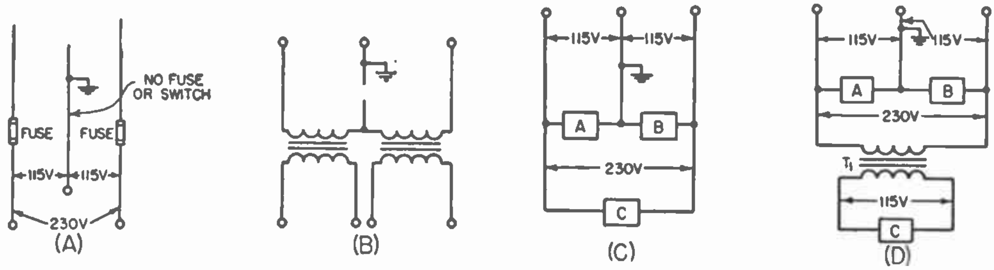


Fig. 5-1 – Three-wire power-line circuits. A – Normal 3-wire-line termination. No fuse should be used in the grounded (neutral) line. B – Showing that a switch in the neutral does not remove voltage from either side of the line. C – Connections for both 115- and 230-volt transformers. D – Operating a 115-volt plate transformer from the 230-volt line to avoid light blinking. T1 is a 2-to-1 step-down transformer.

circuit, as shown in Fig. 5-1B. Furthermore, with the neutral open, the voltage will then be divided between the two sides in inverse proportion to the load resistance, the voltage on one side dropping below normal, while it soars on the other side, unless the loads happen to be equal.

The usual line running to baseboard outlets is rated at 15 amperes. Considering the power consumed by filaments, lamps, transmitter, receiver and other auxiliary equipment, it is not unusual to find this 15-A rating exceeded by the requirements of a station of only moderate power. It must also be kept in mind that the same branch may be in use for other household purposes through another outlet. For this reason, and to minimize light blinking when keying or modulating the transmitter, a separate heavier line should be run from the distribution board to the station whenever possible. (A three-volt drop in line voltage will cause noticeable light blinking.)

If the system is of the three-wire 230-V type, the three wires should be brought into the station so that the load can be distributed to keep the line balanced. The voltage across a fixed load on one side of the circuit will increase as the load current on the other side is increased. The rate of increase will depend upon the resistance introduced by the neutral wire. If the resistance of the neutral is low, the increase will be correspondingly small. When the currents in the two circuits are balanced, no current flows in the neutral wire and the system is operating at maximum efficiency.

Light blinking can be minimized by using transformers with 230-volt primaries in the power supplies for the keyed or intermittent part of the load, connecting them across the two ungrounded wires with no connection to the neutral, as shown in Fig. 5-1C. The same can be accomplished by the insertion of a step-down transformer with its primary operating at 230 volts and secondary delivering 115 volts. Conventional 115-volt transformers may be operated from the secondary of the step-down transformer (see Fig. 5-1D).

When a special heavy-duty line is to be installed, the local power company should be consulted as to local requirements. In some localities it is necessary to have such a job done by a licensed electrician, and there may be special requirements to be met. Some amateurs terminate the special line to the station at a switch box, while others may use electric-stove receptacles as the termination. The power is then distributed around the station by means of conventional outlets at convenient points. All circuits should be properly fused.

Three-Wire 115-V Power Cords

To meet the requirements of state and national codes, electrical tools, appliances and many items of electronic equipment now being manufactured to operate from the 115-volt line must be equipped with a 3-conductor power cord. Two of the conductors carry power to the device in the usual fashion, while the third conductor is connected to the case or frame.

When plugged into a properly wired mating receptacle, the 3-contact polarized plug connects this third conductor to an earth ground, thereby grounding the chassis or frame of the appliance and preventing the possibility of electrical shock to the user. All commercially manufactured items of electronic test equipment and most ac-operated amateur equipments are being supplied with these 3-wire cords. Adapters are available for use where older electrical installations do not have mating receptacles. For proper grounding, the lug of the green wire protruding from the adapter must be attached underneath the screw securing the cover plate of the outlet box where connection is made, and the outlet box itself must be grounded.

Fusing

All transformer primary circuits should be properly fused. To determine the approximate current rating of the fuse to be used, multiply each current being drawn from the supply in amperes by the voltage at which the current is being drawn. Include the current taken by bleeder resistances and voltage dividers. In the case of series resistors, use the source voltage, not the voltage at the equipment end of the resistor. Include filament power if the transformer is supplying filaments. After multiplying the various voltages and currents, add the individual products. Then divide by the line voltage and add 10 or 20 percent. Use a fuse with the nearest larger current rating.

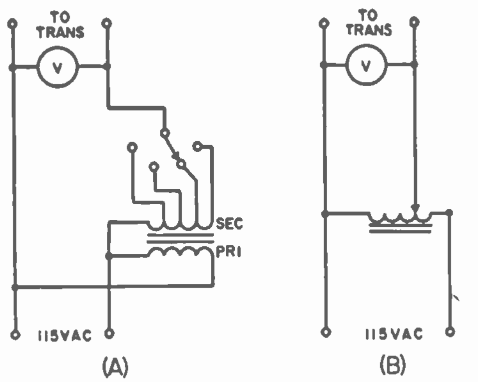


Fig. 5-2 — Two methods of transformer primary control. At A is a tapped toy transformer which may be connected so as to boost or buck the line voltage as required. At B is indicated a variable transformer or autotransformer (Variac) which feeds the transformer primaries.

LINE-VOLTAGE ADJUSTMENT

In certain communities trouble is sometimes experienced from fluctuations in line voltage. Usually these fluctuations are caused by a variation in the load on the line. Since most of the variation comes at certain fixed times of the day or night, such as the times when lights are turned on at evening, they may be taken care of by the use of a manually operated compensating device. A simple arrangement is shown in Fig. 5-2A. A toy transformer is used to boost or buck the line voltage as required. The transformer should have a tapped secondary varying between 6 and 20 volts in steps of 2 or 3 volts and its secondary should be capable of carrying the full load current.

The secondary is connected in series with the line voltage and, if the phasing of the windings is correct, the voltage applied to the primaries of the transmitter transformers can be brought up to the rated 115 volts by setting the toy-transformer tap switch on the right tap. If the phasing of the two windings of the toy transformer happens to be reversed, the voltage will be reduced instead of increased. This connection may be used in cases where the line voltage may be above 115 volts. This method is preferable to using a resistor in the primary of a power transformer since it does not affect the voltage regulation as seriously. The circuit of 5-2B illustrates the use of a variable autotransformer (Variac) for adjusting line voltage.

Constant-Voltage Transformers

Although comparatively expensive, special transformers called constant-voltage transformers are available for use in cases where it is necessary to hold line voltage and/or filament voltage constant with fluctuating supply-line voltage. These are static-magnetic voltage regulating transformers operating on principles of ferroresonance. They have no tubes or moving parts, and require no manual adjustments. These transformers are rated over a range of less than one VA at 5 volts output up to several thousand VA at 115 or 230 volts. On the average they will hold their output voltages within one percent under an input voltage variation of ±15 percent.

SAFETY PRECAUTIONS

All power supplies in an installation should be fed through a single main power-line switch so that all power may be cut off quickly, either before working on the equipment, or in case of an accident. Spring-operated switches or relays are not sufficiently reliable for this important service. Foolproof devices for cutting off all power to the transmitter and other equipment are shown in Fig. 5-3. The arrangements shown in Fig. 5-3A and B are similar circuits for two-wire (115-volt) and three-wire (230-volt) systems. S is an enclosed double-throw switch of the sort usually used as the entrance switch in house installations. J is a standard ac outlet and P a shorted plug to fit the outlet. The switch should be located prominently in plain sight, and members of the household should be instructed in its location and use. I is a red lamp located alongside the switch. Its purpose is not so much to serve as a warning that the power is on as it is to help in identifying and quickly

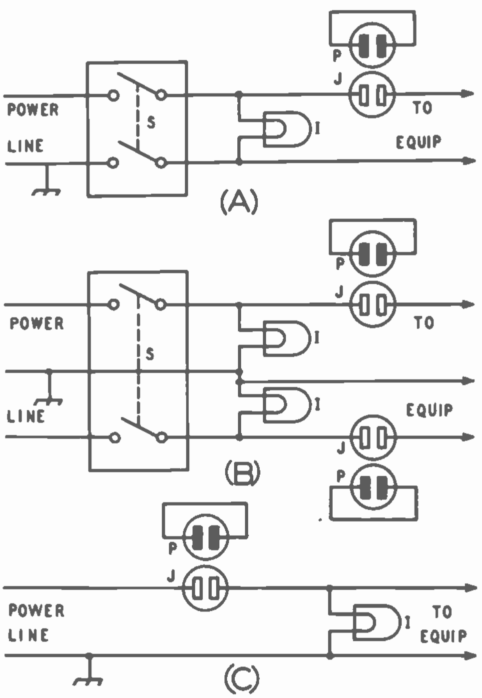


Fig. 5-3 — Reliable arrangements for cutting off all power to the transmitter. S is an enclosed double-pole power switch, J a standard ac outlet, P a shorted plug to fit the outlet and I a red lamp.

A is for a two-wire 115-volt line, B for a three-wire 230-volt system, and C a simplified arrangement for low-power stations.

locating the switch should it become necessary for someone else to cut the power off in an emergency.

The outlet J should be placed in some corner out of sight where it will not be a temptation for children or others to play with. The shorting plug can be removed to open the power circuit if there are others around who might inadvertently throw the switch while the operator is working on the rig. If the operator takes the plug with him, it will prevent someone from turning on the power in his absence and either hurting themselves or the equipment or perhaps starting a fire. Of utmost importance is the fact that the outlet J *must* be placed in the *ungrounded* side of the line.

Those who are operating low power and feel that the expense or complication of the switch isn't warranted can use the shorted-plug idea as the main power switch. In this case, the outlet should be located prominently and identified by a signal light, as shown in Fig. 5-3C.

The test bench should be fed through the main power switch, or a similar arrangement at the bench, if the bench is located remotely from the transmitter.

A bleeder resistor with a power rating which gives a considerable margin of safety should be used across the output of all transmitter power supplies, so that the filter capacitors will be discharged when the high-voltage is turned off.

PLATE AND FILAMENT TRANSFORMERS

Output Voltage

The output voltage which the plate transformer must deliver depends upon the required dc load voltage and the type of filter circuit.

With a choke-input filter (see Fig. 5-4), the required rms secondary voltage (each side of center-tap for a center-tap rectifier) can be calculated by the equation:

$$E_t = 1.1\,[E_o + I\,(R_1 + R_2 + R_s)]$$

where E_o is the required dc output voltage, I is the load current (including bleeder current) in amperes, R1 and R2 are the dc resistances of the chokes, and R_s is the series resistance (transformer and rectifier). E_t is the open-circuit rms voltage.

With a capacitive-input filter system, the approximate transformer output voltage required to give a desired dc output voltage with a given load can be calculated with the aid of Fig. 5-5.

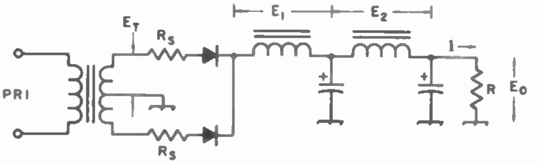


Fig. 5-4 — Diagram showing various voltage drops that must be taken into consideration in determining the required transformer voltage to deliver the desired output voltage.

Example:

Required dc output volts – 25
Load current to be drawn – 500 mA (0.5 ampere)
Input capacitor – 1000 µF

Series resistance – 5 ohms

Load resistance = $\frac{25}{0.5}$ = 50 ohms

RC = 50 × 1000 = 50,000

R_s/R = 5/50 = 0.1

Fig. 5-5 shows that the ratio of dc volts to the required transformer rms voltage is 1.07.

The required transformer terminal voltage under load is

$$E_{AC} = \frac{E_{DC} + I \times R_s}{1.07}$$

where I is the load current in amperes.

$$E_{AC} = \frac{25 + 0.5 \times 5}{1.07} = \frac{27.5}{1.07} = 25.7 \text{ volts}$$

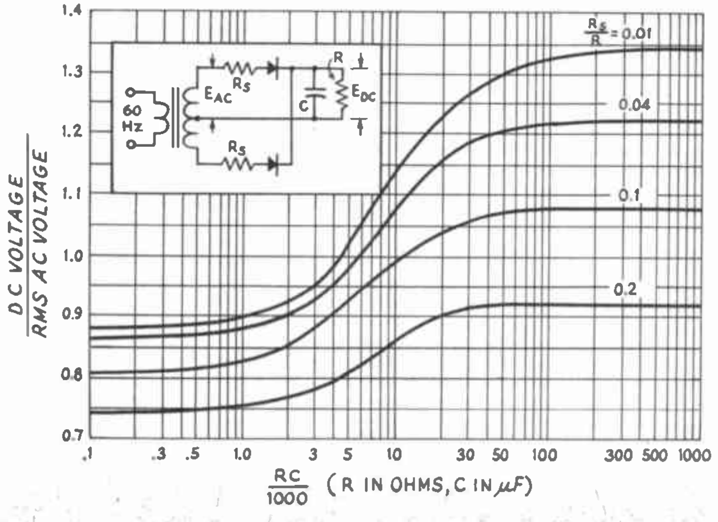


Fig. 5-5 — Dc output voltages from a full-wave rectifier circuit as a function of the filter capacitance and load resistance. R_s includes transformer winding resistance and rectifier forward resistance. For the ratio R_s/R, both resistances are in ohms; for the RC product, R is in ohms and C is in µF.

The required transformer is one having a 51.4-V center-tapped secondary. A 50- or 55-V secondary would be entirely satisfactory. Should the filter section contain one or more filter chokes connected between the input capacitor and the load, the dc-resistance values of the chokes are added to the value of R_s in the equation before multiplying by the load-current value.

Volt-Ampere Rating

The number of volt-amperes delivered by a transformer depends upon the type of filter (capacitor or choke input) used, and upon the type of rectifier used (full-wave center tap, or full-wave bridge). With a capacitive-input filter the heating effect in the secondary is higher because of the high ratio of peak-to-average current. The volt-amperes handled by the transformer may be several times the watts delivered to the load. With a choke-input filter, provided the input choke has at least the critical inductance, the secondary volt-amperes can be calculated quite closely by the equation:

$$\text{(Full-wave ct)}\ Sec.\ VA = \frac{.707\,EI}{1000}$$

$$\text{(Full-wave bridge)}\ Sec.\ VA = \frac{EI}{1000}$$

where E is the *total* rms voltage of the secondary (between the outside ends in the case of a center-tapped winding) and I is the dc output current in milliamperes (load current plus bleeder current). The primary volt-amperes will be somewhat higher because of transformer losses.

BROADCAST & TELEVISION REPLACEMENT TRANSFORMERS

Small power transformers of the type sold for replacement in broadcast and television receivers are usually designed for service in terms of use for several hours continuously with capacitor-input filters. In the usual type of amateur transmitter service, where most of the power is drawn intermittently for periods of several minutes with equivalent intervals in between, the published ratings can be exceeded without excessive transformer heating.

With a capacitor-input filter, it should be safe to draw 20 to 30 percent more current than the rated value. With a choke-input filter, an increase in current of about 50 percent is permissible. If a bridge rectifier is used, the output voltage will be approximately doubled. In this case, it should be possible in amateur transmitter service to draw the rated current, thus obtaining about twice the rated output power from the transformer.

This does not apply, of course, to amateur transmitter plate transformers, which usually are already rated for intermittent service.

REWINDING POWER TRANSFORMERS

Although the home winding of power transformers is a task that few amateurs undertake, the

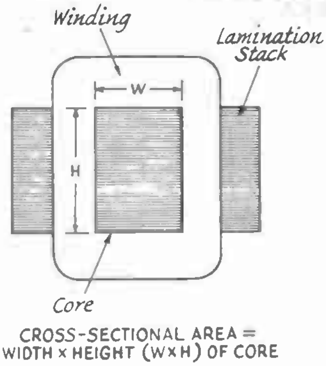


Fig. 5-6 — Cross-sectional drawing of a typical power transformer. Multiplying the height (or thickness of the laminations) times the width of the central core area in inches gives the value to be applied to Fig. 5-7.

rewinding of a transformer secondary to give some desired voltage for powering filaments or a solid-state device is not difficult. It involves a matter of only a small number of turns and the wire is large enough to be handled easily. Often a receiver power transformer with a burned-out high-voltage winding or the power transformer from a discarded TV set can be converted into an entirely satisfactory transformer without great effort and with little expense. The average TV power transformer for a 17-inch or larger set is capable of delivering from 350 to 450 watts, continuous duty. If an amateur transmitter is being powered, the service is not continuous, so the ratings can be increased by a factor of 40 or 50 percent without danger of overloading the transformer.

The primary volt-ampere rating of the transformer to be rewound, if known, can be used to determine its power-handling capability. The secondary volt-ampere rating will be ten to twenty percent less than the primary rating. The power rating may also be determined approximately from the cross-sectional area of the core which is *inside* the windings. Fig. 5-6 shows the method of determining the area, and Fig. 5-7 may be used to convert this information into a power rating.

Before disconnecting the winding leads from their terminals, each should be marked for identification. In removing the core laminations, care should be taken to note the manner in which the core is assembled, so that the reassembling will be done in the same manner. Most transformers have secondaries wound over the primary, while in some the order is reversed. In case the secondaries are on the inside, the turns can be pulled out from the center after slitting and removing the fiber core.

The turns removed from one of the original filament windings of known voltage should be carefully counted as the winding is removed. This will give the number of turns per volt and the same figure should be used in determining the number of turns for the new secondary. For instance, if the

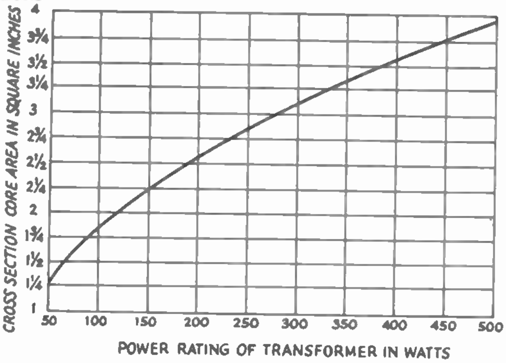


Fig. 5-7 – Power-handling capability of a transformer versus cross-sectional area of core.

old filament winding was rated at 5 volts and had 15 turns, this is 15/5 = 3 turns per volt. If the new secondary is to deliver 18 volts, the required number of turns on the new winding will be 18 X 3 = 54 turns.

In winding a transformer, the size of wire is an important factor in the heat developed in operation. A cross-sectional area of 1000 circular mils per ampere is conservative. A value commonly used in amateur-service transformers is 700 cmil/A. The larger the cmil/A figure, the cooler the transformer will run. The current rating in amperes of various wire sizes is shown in the copper-wire table in another chapter. If the transformer being rewound is a filament transformer, it may be necessary to choose the wire size carefully to fit the small available space. On the other hand, if the transformer is a power unit with the high-voltage winding removed, there should be plenty of room for a size of wire that will conservatively handle the required current.

After the first layer of turns is put on during rewinding, secure the ends with cellulose tape. Each layer should be insulated from the next; ordinary household waxed paper can be used for the purpose, a single layer being adequate. Sheets cut to size beforehand may be secured over each layer with tape. Be sure to bring all leads out the same side of the core so the covers will go in place when the unit is completed. When the last layer of the winding is put on, use two sheets of waxed paper, and then cover those with vinyl electrical tape, keeping the tape as taut as possible. This will add mechanical strength to the assembly.

The laminations and housing are assembled in just the opposite sequence to that followed in disassembly. Use a light coating of shellac between each lamination. During reassembly, the lamination stack may be compressed by clamping in a vise. If the last few lamination strips cannot be replaced, it is better to omit them than to force the unit together.

RECTIFIER CIRCUITS

Half-Wave Rectifier

Fig. 5-8 shows three rectifier circuits covering most of the common applications in amateur equipment. Fig. 5-8A is the circuit of a half-wave rectifier. The rectifier is a device that will conduct current in one direction but not in the other. During one half of the ac cycle the rectifier will conduct and current will flow through the rectifier to the load. During the other half of the cycle the rectifier does not conduct and no current flows to the load. The shape of the output wave is shown in (A) at the right. It shows that the current always flows in the same direction but that the flow of current is not continuous and is pulsating in amplitude.

The average output voltage – the voltage read by the usual dc voltmeter – with this circuit (no filter connected) is 0.45 times the rms value of the ac voltage delivered by the transformer secondary. Because the frequency of the pulses is relatively low (one pulsation per cycle), considerable filtering is required to provide adequately smooth dc output, and for this reason this circuit is usually limited to applications where the current involved is small, such as supplies for cathode-ray tubes and for protective bias in a transmitter.

The peak reverse voltage (PRV), the voltage the rectifier must withstand when it isn't conducting, varies with the load. With a resistive load it is the peak ac voltage (1.4 E_{RMS}) but with a capacitor load drawing little or no current it can rise to 2.8 E_{RMS}.

Another disadvantage of the half-wave rectifier circuit is that the transformer must have a considerably higher primary volt-ampere rating (approximately 40 percent greater), for the same dc power output, than in other rectifier circuits.

Full-Wave Center-Tap Rectifier

A commonly used rectifier circuit is shown in Fig. 5-8B. Essentially an arrangement in which the outputs of two half-wave rectifiers are combined, it makes use of both halves of the ac cycle. A transformer with a center-tapped secondary is required with the circuit.

The average output voltage is 0.9 times the rms voltage of half the transformer secondary; this is the maximum voltage that can be obtained with a suitable choke-input filter. The peak output voltage is 1.4 times the rms voltage of half the transformer secondary; this is the maximum voltage that can be obtained from a capacitor-input filter (at little or no load).

The peak reverse voltage across a rectifier unit is 2.8 times the rms voltage of half the transformer secondary.

As can be seen from the sketches of the output wave form in (B) to the right, the frequency of the output pulses is twice that of the half-wave rectifier. Therefore much less filtering is required.

Since the rectifiers work alternately, each handles half of the load current, and the load-current rating of each rectifier need be only half the total load current drawn from the supply.

Two separate transformers, with their primaries connected in parallel and secondaries connected in series (with the proper polarity) may be used in this circuit. However, if this substitution is made, the primary volt-ampere rating must be reduced to about 40 percent less than twice the rating of one transformer.

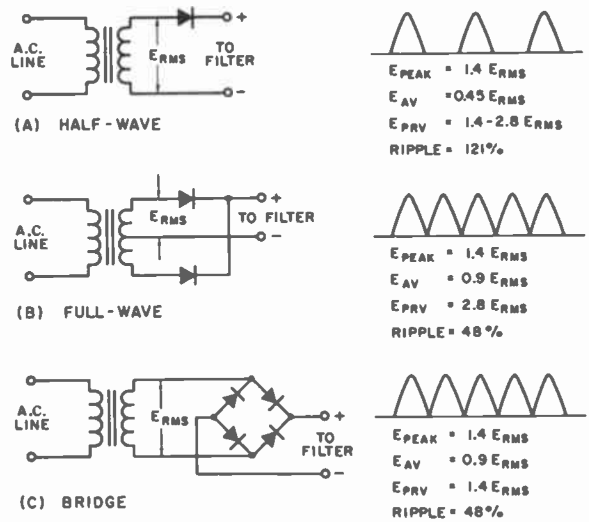


Fig. 5-8 — Fundamental rectifier circuits. A — Half-wave (E_{PRV} = 1.4 E_{RMS} with resistive load, = 2.8 E_{RMS} with capacitor-input filter). B — Full-wave. C — Full-wave bridge. Output voltage values do not include rectifier voltage drops.

Full-Wave Bridge Rectifier

Another full-wave rectifier circuit is shown in Fig. 5-8C. In this arrangement, two rectifiers operate in series on each half of the cycle, one rectifier being in the lead to the load, the other being in the return lead. The current flows through two rectifiers during one half of the cycle and through the other two rectifiers during the other half of the cycle. The output wave shape (C), to the right, is the same as that from the simple center-tap rectifier circuit. The maximum output voltage into a resistive load or a properly designed choke-input filter is 0.9 times the rms voltage delivered by the transformer secondary; with a capacitor-input filter and a very light load the output voltage is 1.4 times the secondary rms voltage. The peak reverse voltage per rectifier is 1.4 times the secondary rms voltage. Each rectifier in a bridge circuit should have a minimum load-current rating of one-half the total load current to be drawn from the supply.

RECTIFIER RATINGS

All rectifiers are subject to limitations as to breakdown voltage and current-handling capability. Some tube types are rated in terms of the maximum rms voltage that should be applied to the rectifier plate. This is sometimes dependent on whether a choke- or capacitive-input filter is used. Others, particularly mercury-vapor and semiconductor types, are rated according to maximum peak reverse voltage.

Rectifiers are rated also as to maximum dc load current, and some may carry peak-current ratings in addition. To assure normal life, all ratings should be carefully observed.

HIGH-VACUUM RECTIFIERS

High-vacuum rectifiers depend entirely upon the thermionic emission from a heated filament and are characterized by a relatively high internal resistance. For this reason, their application usually is limited to low power, although there are a few types designed for medium and high power in cases where the relatively high internal voltage drop may be tolerated. This high internal resistance makes them less susceptible to damage from temporary overload and they are free from the bothersome electrical noise sometimes associated with other types of rectifiers.

Some rectifiers of the high-vacuum full-wave type in the so-called receiver-tube class will handle up to 275 mA at 400- to 500-volts dc output. Those in the higher power class can be used to handle up to 500 mA at 2000 volts dc in full-wave circuits. Most low-power high-vacuum rectifiers are produced in the full-wave type, while those for greater power are invariably of the half-wave type, two tubes being required for a full-wave rectifier circuit. A few of the lower voltage types have indirectly heated cathodes, but are limited in heater-to-cathode voltage rating.

SEMICONDUCTOR RECTIFIERS

Silicon rectifiers are being used almost exclusively in power supplies for amateur equipment. Types are available to replace high-vacuum and mercury-vapor rectifiers. The semiconductors have the advantages of compactness, low internal voltage drop, low operating temperature and high current-handling capability. Also, no filament transformers are required.

Silicon rectifiers are available in a wide range of voltage and current ratings. In peak reverse voltage ratings of 600 or less, silicon rectifiers carry current ratings as high as 400 amperes, and at 1000 PRV the current ratings may be 1.5 amperes or so. The extreme compactness of silicon types makes feasible the stacking of several units in series for higher voltages. Standard stacks are available that will handle up to 10,000 PRV at a dc load current of 500 mA, although the amateur can do much better, economically, by stacking the rectifiers himself.

PROTECTION OF SILICON POWER DIODES

The important specifications of a silicon diode are:

1) PRV (or PIV), the peak reverse (or peak inverse) voltage,

2) I_o, the average dc current rating.

3) I_{REP}, the peak repetitive forward current, and

4) I_{SURGE}, the peak one-cycle surge current. The first two specifications appear in most catalogs. The last two often do not, but they are very important.

Since the rectifier never allows current to flow more than half the time, when it does conduct it has to pass at least twice the average direct current. With a capacitor-input filter, the rectifier conducts much less than half the time, so that when it does conduct, it may pass as much as ten to twenty times the average dc current, under certain conditions. This peak current is I_{REP}, the peak repetitive forward current.

Also, when the supply is first turned on, the discharged input capacitor looks like a dead short, and the rectifier passes a very heavy current. This is I_{SURGE}. The maximum I_{SURGE} rating is usually for a duration of one cycle (at 60 Hz), or about 16.7 milliseconds.

If a manufacturer's data sheet is not available, an educated guess about a diode's capability can be made by using these rules of thumb for silicon diodes of the type commonly used in amateur power supplies:

Rule 1) The maximum I_{REP} rating can be assumed to be approximately four times the maximum I_o rating.

Rule 2) The maximum I_{SURGE} rating can be assumed to be approximately twelve times the maximum I_o rating. (This should provide a reasonable safety factor. Silicon rectifiers with 750-mA dc ratings, as an example, seldom have 1-cycle surge ratings of less than 15 amperes; some are rated up to 35 amperes or more.) From this then, it can be seen that the rectifier should be selected on the basis of I_{SURGE} and not on I_o ratings.

Thermal Protection

The junction of a diode is quite small, hence it must operate at a high current density. The heat-handling capability is, therefore, quite small. Normally, this is not a prime consideration in high-voltage, low-current supplies. When using high-current rectifiers at or near their maximum ratings (usually 2-ampere or larger stud-mount rectifiers), some form of heat sinking is necessary. Frequently, mounting the rectifier on the main chassis – directly, or by means of thin mica insulating washers – will suffice. If insulated from the chassis, a thin layer of silicone grease should be used between the diode and the insulator, and between the insulator and the chassis to assure good heat conduction. Large high-current rectifiers often require special heat sinks to maintain a safe operating temperature. Forced-air cooling is sometimes used as a further aid. Safe case temperatures are usually given in the manufacturer's data sheets and should be observed if the maximum capabilities of the diode are to be realized.

Surge Protection

Each time the power supply is activated, assuming the input filter capacitor has been discharged, the rectifiers must look into what represents a dead short. Some form of surge protection is usually necessary to protect the diodes until the input capacitor becomes nearly charged. Although the dc resistance of the transformer secondary can be relied upon in some instances to provide ample surge-current limiting, it is seldom enough on high-voltage power supplies to be suitable. Series resistors can be installed between the secondary and the rectifier strings as illustrated in Fig. 5-4, but are a deterrent to good

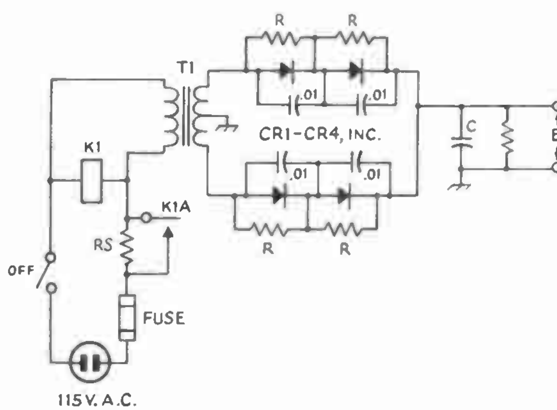
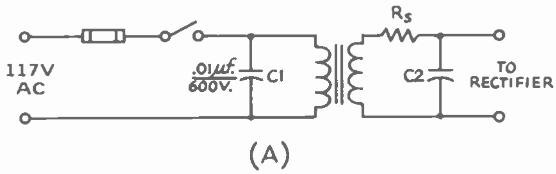


Fig. 5-9 – The primary circuit of T1 shows how a 115-volt ac relay and a series dropping resistor, R_S, can provide surge protection while C charges. When silicon rectifiers are connected in series for high-voltage operation, the inverse voltage does not divide equally. The reverse voltage drops can be equalized by using equalizing resistors, as shown in the secondary circuit. To protect against voltage "spikes" that may damage an individual rectifier, each rectifier should be bypassed by a .01-μF capacitor. Connected as shown, two 400-PRV silicon rectifiers can be used as an 800-PRV rectifier, although it is preferable to include a safety factor and call it a "750-PRV" rectifier. The rectifiers, CR1 through CR4, should be the same type (same type number and ratings).



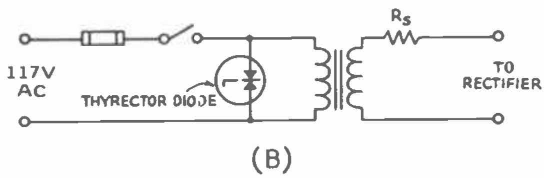


Fig. 5-10 – Methods of suppressing line transients. See text.

voltage regulation. By installing a surge-limiting device in the primary circuit of the plate transformer, the need for series resistors in the secondary circuit can be avoided. A practical method for primary-circuit surge control is shown in Fig. 5-9. The resistor, R_s introduces a voltage drop in the primary feed to T1 until C is nearly charged. Then, after C becomes partially charged, the voltage drop across R_s lessens and allows K1 to pull in, thus applying full primary power to T1 as K1A shorts out R_s. R_s is usually a 25-watt resistor whose resistance is somewhere between 15 and 50 ohms, depending upon the power supply characteristics.

Transient Problems

A common cause of trouble is transient voltages on the ac power line. These are short spikes, mostly, that can temporarily increase the voltage seen by the rectifier to values much higher than the normal transformer voltage. They come from distant lightning strokes, electric motors turning on and off, and so on. Transients cause unexpected, and often unexplained, loss of silicon rectifiers.

It's always wise to suppress line transients, and it can be easily done. Fig. 5-10A shows one way. C1 looks like 280,000 ohms at 60 Hz, but to a sharp transient (which has only high-frequency components), it is an effective bypass. C2 provides additional protection on the secondary side of the transformer. It should be .01µF for transformer voltages of 100 or less, and .001µF for high-voltage transformers.

Fig. 5-10B shows another transient-suppression method using selenium suppressor diodes. The diodes do not conduct unless the peak voltage becomes abnormally high. Then they clip the transient peaks. General Electric sells protective diodes under the trade name, "Thyrector." Sarkes-Tarzian uses the descriptive name, "Klip-volt."

Transient voltages can go as high as twice the normal line voltage before the suppressor diodes clip the peaks. Capacitors cannot give perfect suppression either. Thus, it is a good idea to use power-supply rectifiers rated at about twice the expected PRV.

Diodes in Series

Where the PRV rating of a single diode is not sufficient for the application, similar diodes may be used in series. (Two 500-PRV diodes in series will withstand 1000 PRV, and so on.) When this is done, a resistor and a capacitor should be placed across each diode in the string to equalize the PRV drops and to guard against transient voltage spikes, as shown in Fig. 5-9. Even though the diodes are of the same type and have the same PRV rating, they may have widely different back resistances when they are cut off. The reverse voltage divides according to Ohm's Law, and the diode with the higher back resistance will have the higher voltage developed across it. The diode may break down.

If we put a swamping resistor across each diode, R as shown in Fig. 5-9, the resultant resistance across each diode will be almost the same, and the back voltage will divide almost equally. A good rule of thumb for resistor size is this: Multiply the PRV rating of the diode by 500 ohms. For example, a 500-PRV diode should be shunted by 500 X 500, or 250,000 ohms.

The shift from forward conduction to high back resistance does not take place instantly in a silicon diode. Some diodes take longer than others to develop high back resistance. To protect the "fast" diodes in a series string until all the diodes are properly cut off, a .01-µF capacitor should be placed across each diode. Fig. 5-9 shows the complete series-diode circuit. The capacitors should be noninductive, ceramic disk, for example, and should be well matched. Use 10-percent-tolerance capacitors if possible.

Diodes in Parallel

Diodes can be placed in parallel to increase current-handling capability. Equalizing resistors should be added as shown in Fig. 5-11. Without the resistors, one diode may take most of the current. The resistors should be selected to have about a 1-volt drop at the expected peak current.

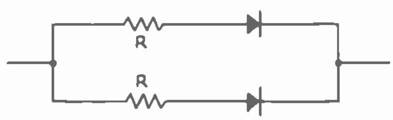


Fig. 5-11 – Diodes in parallel should have equalizing resistors. See text for appropriate value.

FILTERING

The pulsating dc waves from the rectifiers are not sufficiently constant in amplitude to prevent hum corresponding to the pulsations. Filters are required between the rectifier and the load to smooth out the pulsations into an essentially constant dc voltage. Also, upon the design of the filter depends to a large extent the dc voltage output, the voltage regulation of the power supply, and the maximum load current that can be drawn from the supply without exceeding the peak-current rating of the rectifier. Power supply filters are low-pass devices using series inductors and shunt capacitors.

Load Resistance

In discussing the performance of power-supply filters, it is sometimes convenient to express the load connected to the output terminals of the supply in terms of resistance. The load resistance is equal to the output voltage divided by the total current drawn, including the current drawn by the bleeder resistor.

Voltage Regulation

The output voltage of a power supply always decreases as more current is drawn, not only because of increased voltage drops on the transformer, filter chokes and the rectifier (if high-vacuum rectifiers are used) but also because the output voltage at light loads tends to soar to the peak value of the transformer voltage as a result of charging the first capacitor. By proper filter design the latter effect can be eliminated. The change in output voltage with load is called *voltage regulation* and is expressed as a percentage.

$$\textit{Percent regulation} = \frac{100\,(E1 - E2)}{E2}$$

Example: No-load voltage = E1 = 1550 volts.
Full-load voltage = E2 = 1230 volts.

$$\text{Percentage regulation} = \frac{100\,(1550 - 1230)}{1230}$$

$$= \frac{32{,}000}{1230} = 26 \text{ percent}$$

A steady load, such as that represented by a receiver, speech amplifier or unkeyed stages of a transmitter, does not require good (low) regulation as long as the proper voltage is obtained under load conditions. However, the filter capacitors must have a voltage rating safe for the highest value to which the voltage will soar when the external load is removed.

A power supply will show more (higher) regulation with long-term changes in load resistance than with short temporary changes. The regulation with long-term changes is often called the **static** regulation, to distinguish it from the **dynamic** regulation (short temporary load changes). A load that varies at a syllabic or keyed rated, as represented by some audio and rf amplifiers, usually requires good dynamic regulation (15 percent or less) if distortion products are to be held to a low level. The dynamic regulation of a power supply is improved by increasing the value of the output capacitor.

When essentially constant voltage regardless of current variation is required (for stabilizing an oscillator, for example), special voltage-regulating circuits described elsewhere in this chapter are used.

Bleeder

A bleeder resistor is a resistance connected across the output terminals of the power supply. Its functions are to discharge the filter capacitors as a safety measure when the power is turned off and to improve voltage regulation by providing a minimum load resistance. When voltage regulation is not of importance, the resistance may be as high as 100 ohms per volt. The resistance value to be used for voltage-regulating purposes is discussed in later sections. From the consideration of safety, the power rating of the resistor should be as conservative as possible, since a burned-out bleeder resistor is more dangerous than none at all!

Ripple Frequency and Voltage

The pulsations in the output of the rectifier can be considered to be the resultant of an alternating current superimposed upon a steady direct current. From this viewpoint, the filter may be considered to consist of shunting capacitors which short-circuit the ac component while not interfering with the flow of the dc component, and series chokes which pass dc readily but which impede the flow of the ac component.

The alternating component is called the ripple. The effectiveness of the filter can be expressed in terms of percent ripple, which is the ratio of the rms value of the ripple to the dc value in terms of percentage. Any multiplier or amplifier supply in a code transmitter should have less than 5 percent ripple. A linear amplifier can tolerate about 3 percent ripple on the plate voltage. Bias supplies for linear amplifiers, and modulator and modulated-amplifier plate supplies, should have less than 1 percent ripple. VFOs, speech amplifiers and receivers may require a ripple reduction to .01 percent.

Ripple frequency is the frequency of the pulsations in the rectifier output wave – the number of pulsations per second. The frequency of the ripple with half-wave rectifiers is the same as the frequency of the line supply – 60 Hz with 60-Hz supply. Since the output pulses are doubled with a full-wave rectifier, the ripple frequency is doubled – to 120 Hz with a 60-Hz supply.

The amount of filtering (values of inductance and capacitance) required to give adequate smoothing depends upon the ripple frequency, with more filtering being required as the ripple frequency is lowered.

Type of Filter

Power-supply filters fall into two classifications, capacitor input and choke input. Capacitor-input filters are characterized by relatively high output voltage in respect to the transformer voltage. Advantage of this can be taken when silicon rectifiers are used or with any rectifier when the load resistance is high. Silicon rectifiers have a higher allowable peak-to-dc ratio than do thermionic rectifiers. This permits the use of capacitor-input filters at ratios of input capacitor to load resistance that would seriously shorten the life of a thermionic rectifier system. When the series resistance through a rectifier and filter system is appreciable, as when high-vacuum rectifiers are used, the voltage regulation of a capacitor-input power supply is poor.

The output voltage of a properly designed choke-input power supply is less than would be obtained with a capacitor-input filter from the same transformer.

CAPACITIVE-INPUT FILTERS

Capacitive-input filter systems are shown in Fig. 5-12. Disregarding voltage drops in the chokes, all have the same characteristics except in respect to ripple. Better ripple reduction will be obtained when LC sections are added, as shown in Figs. 5-12B and C.

Output Voltage

To determine the approximate dc voltage output when a capacitive-input filter is used, reference should be made to the graph of Fig. 5-5.

Example:
Transformer rms voltage – 350
Load resistance – 2000 ohms

Series resistance – 200 ohms
$200 \div 2000 = 0.1$
Input capacitor $C = 20\ \mu F$

$$\frac{RC}{1000} = \frac{2000 \times 20}{1000} = 40$$

From curve 0.1 and $RC = 40$, dc voltage $= 350 \times 1.06 = 370$.

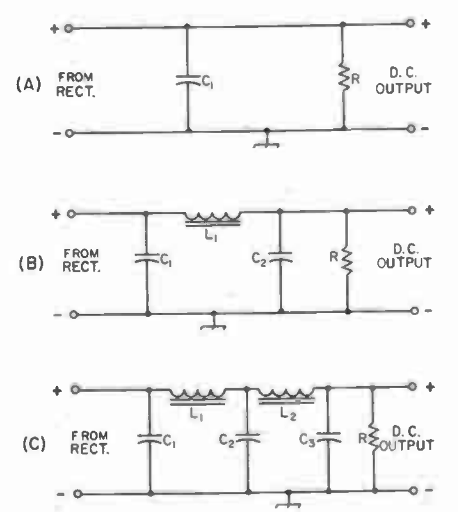


Fig. 5-12 — Capacitive-input filter circuits. A — Simple capacitive. B — Single-section. C — Double-section.

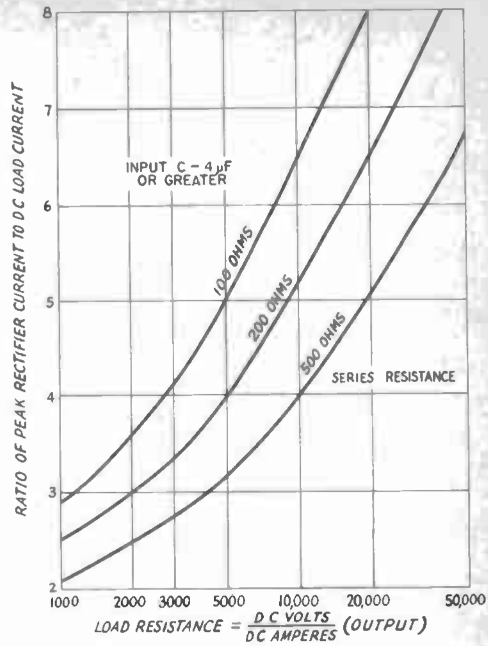


Fig. 5-13 — Graph showing the relationship between the dc load current and the rectifier peak current with capacitive input for various values of load and input resistance.

Regulation

If a bleeder resistance of 20,000 ohms is used in the example above, when the load is removed and R becomes 20,000, the dc voltage will rise to 470. For minimum regulation with a capacitor-input filter, the bleeder resistance should be as high as possible, or the series resistance should be low and the filter capacitance high, without exceeding the transformer or rectifier ratings.

Maximum Rectifier Current

The maximum current that can be drawn from a supply with a capacitive-input filter without exceeding the peak-current rating of the rectifier may be estimated from the graph of Fig. 5-13. Using values from the preceding example, the ratio of peak rectifier current to dc load current for 2000 ohms, as shown in Fig. 5-13, is 3. Therefore, the maximum load current that can be drawn without exceeding the rectifier rating is 1/3 the peak rating of the rectifier. For a load current of 185 mA, as above (370 V $\div$ 2000 Ω), the rectifier peak current rating should be at least $3 \times 185 = 555$ mA.

With bleeder current only, Fig. 5-13 shows that the ratio will increase to 6.5. But since the bleeder draws 23.5 mA dc, the rectifier peak current will be only 153 mA.

CHOKE-INPUT FILTERS

With thermionic rectifiers better voltage regulations results when a choke-input filter, as shown in Fig. 5-4, is used. Choke input permits better utilization of the thermionic rectifier, since a higher load current usually can be drawn without exceeding the peak current rating of the rectifier.

Minimum Choke Inductance

A choke-input filter will tend to act as a capacitive-input filter unless the input choke has at least a certain minimum value of inductance called the critical value. This critical value is given by

$$L_{crit}\text{ (henrys)} = \frac{E\text{ (volts)}}{I\text{ (mA)}}$$

where E is the output voltage of the supply, and I is the current being drawn through the filter.

If the choke has at least the critical value, the output voltage will be limited to the average value of the rectified wave at the input to the choke (see Fig. 5-8) when the current drawn from the supply is small. This is in contrast to the capacitive-input filter in which the output voltage tends to soar toward the peak value of the rectified wave at light loads.

Minimum-Load – Bleeder Resistance

From the formula above for critical inductance, it is obvious that if no current is drawn from the supply, the critical inductance will be infinite. So that a practical value of inductance may be used, some current must be drawn from the supply at all times the supply is in use. From the formula we find that this minimum value of current is

$$I\text{ (mA)} = \frac{E\text{ (volts)}}{L_{crit}}$$

In the majority of cases it will be most convenient to adjust the bleeder resistance so that the bleeder will draw the required minimum current. From the formula, it may be seen that the value of critical inductance becomes smaller as the load current increases.

Swinging Chokes

Less costly chokes are available that will maintain at least the critical value of inductance over the range of current likely to be drawn from practical supplies. These chokes are called swinging chokes. As an example, a swinging choke may have an inductance rating of 5/25 H and a current rating of 200 mA. If the supply delivers 1000 volts, the minimum load current should be 1000/25 = 40 mA. When the full load current of 200 mA is drawn from the supply, the inductance will drop to 5 H. The critical inductance for 200 mA at 1000 volts is 1000/200 = 5 H. Therefore the 5/25 H choke maintains the critical inductance at the full current rating of 200 mA. At all load currents between 40 mA and 200 mA, the choke will adjust its inductance to the approximate critical value.

Output Voltage

Provided the input-choke inductance is at least the critical value, the output voltage may be calculated quite close by the following equation:

$$E_o = 0.9E_t - (I_B + I_L)(R1 + R2) - E_r$$

where E_o is the output voltage; E_t is the rms voltage applied to the rectifier (rms voltage between center-tap and one end of the secondary in the case of the center-tap rectifier); I_B and I_L are the bleeder and load currents, respectively, in amperes; R_1 and R_2 are the resistances of the first and second filter chokes; and E_r is the voltage drop across the rectifier. The various voltage drops are shown in Fig. 5-4. At no load I_L is zero; hence the no-load voltage may be calculated on the basis of bleeder current only. The voltage regulation may be determined from the no-load and full-load voltages using the formula previously given.

OUTPUT CAPACITOR

Whether the supply has a choke- or capacitor-input filter, if it is intended for use with a Class A af amplifier, the reactance of the output capacitor should be low for the lowest audio frequency; 16 μF or more is usually adequate. When the supply is used with a Class B amplifier (for modulation or for ssb amplification) or a cw transmitter, increasing the output capacitance will result in improved dynamic regulation of the supply. However, a region of diminishing returns can be reached, and 20 to 30 μF will usually suffice for any supply subjected to large changes at a syllabic (or keying) rate.

RESONANCE

Resonance effects in the series circuit across the output of the rectifier, formed by the first choke and first filter capacitor, must be avoided, since the ripple voltage would build up to large values. This not only is the opposite action to that for which the filter is intended, but may also cause excessive rectifier peak currents and abnormally high peak-reverse voltages. For full-wave rectification the ripple frequency will be 120 Hz for a 60-Hz supply, and resonance will occur when the product of choke inductance in henrys times capacitor capacitance in microfarads is equal to 1.77. At least twice this product of inductance and capacitance should be used to ensure against resonance effects. With a swinging choke, the minimum rated inductance of the choke should be used.

RATINGS OF FILTER COMPONENTS

In a power supply using a choke-input filter and properly designed choke and bleeder resistor, the no-load voltage across the filter capacitors will be about nine-tenths of the ac rms voltage. Nevertheless, it is advisable to use capacitors rated for the *peak* transformer voltage. This large safety factor is suggested because the voltage across the capacitors can reach this peak value if the bleeder should burn out and there is no load on the supply.

In a capacitive-input filter, the capacitors should have a working-voltage rating at least as high, and preferably somewhat higher, than the peak voltage from the transformer. Thus, in the case of a center-tap rectifier having a transformer delivering 550 volts each side of the center tap, the minimum safe capacitor voltage rating will be 550 X 1.41 or 775 volts. An 800-volt capacitor should be used, or preferably a 1000-volt unit.

Filter Capacitors in Series

Filter capacitors are made in several different types. Electrolytic capacitors, which are available for peak voltages up to about 800, combine high capacitance with small size, since the dielectric is an extremely thin film of oxide on aluminum foil. Capacitors of this type may be connected in series for higher voltages, although the filtering capacitance will be reduced to the resultant of the two capacitances in series. If this arrangement is used, it is important that *each* of the capacitors be shunted with a resistor of about 100 ohms per volt of supply voltage applied to the individual capacitors, with an adequate power rating. These resistors may serve as all or part of the bleeder resistance. Capacitors with higher voltage ratings usually are made with a dielectric of thin paper impregnated with oil. The working voltage of a capacitor is the voltage that it will withstand continuously.

Filter Chokes

Filter chokes or inductances are wound on iron cores, with a small gap in the core to prevent magnetic saturation of the iron at high currents. When the iron becomes saturated its permeability decreases, and consequently the inductance also decreases. Despite the air gap, the inductance of a choke usually varies to some extent with the direct current flowing in the winding; hence it is necessary to specify the inductance at the current which the choke is intended to carry. Its inductance with little or no direct current flowing in the winding will usually be considerably higher than the value when full load current is flowing.

NEGATIVE-LEAD FILTERING

For many years it has been almost universal practice to place filter chokes in the positive leads of plate power supplies. This means that the insulation between the choke winding and its core (which should be grounded to chassis as a safety measure) must be adequate to withstand the output voltage of the supply. This voltage requirement is removed if the chokes are placed in the negative lead as shown in Fig. 5-14. With this connection, the capacitance of the transformer secondary to ground appears in parallel with the filter chokes tending to bypass the chokes. However, this effect will be negligible in practical application except in cases where the output ripple must be reduced to a very low figure. Such applications are usually limited to low-voltage devices such as receivers, speech amplifiers and VFOs where insulation is no problem and the chokes may be placed in the positive side in the conventional manner. In higher voltage applications, there is no reason why the filter chokes should not be placed in the negative lead to reduce insulation requirements. Choke terminals, negative capacitor terminals and the transformer center-tap terminal should be well protected against accidental contact, since these will assume full supply voltage to chassis should a choke burn out or the chassis connection fail.

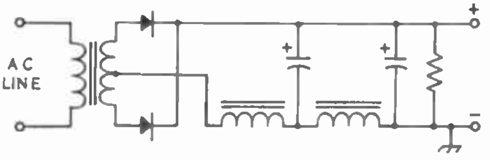


Fig. 5-14 – In most applications, the filter chokes may be placed in the negative instead of the positive side of the circuit. This reduces the danger of a voltage breakdown between the choke winding and core.

THE "ECONOMY" POWER SUPPLY

In many transmitters of the 100-watt class, an excellent method for obtaining plate and screen voltages without wasting power in resistors is by the use of the "economy" power-supply circuit. Shown in Fig. 5-15, it is a combination of the full-wave and bridge-rectifier circuits. The voltage at E1 is the normal voltage obtained with the full-wave circuit, and the voltage at E2 is that obtained with the bridge circuit (see Fig. 5-8). The *total* dc power obtained from the transformer is, of course, the same as when the transformer is used in its normal manner. In cw and ssb applications, additional power can usually be drawn without excessive heating, especially if the transformer has a rectifier filament winding that isn't being used.

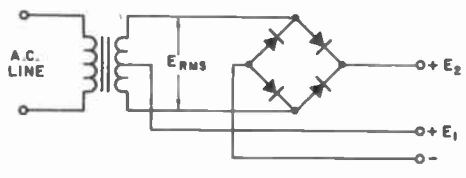


Fig. 5-15 – The "economy" power supply circuit is a combination of the full-wave and bridge-rectifier circuits.

VOLTAGE-MULTIPLYING CIRCUITS

Although vacuum-tube rectifiers can be used in voltage-multiplying circuits, semiconductor rectifiers are recommended.

A simple half-wave rectifier circuit is shown in Fig. 5-16. Strictly speaking this is not a voltage-multiplying circuit. However, if the current demand is low (a milliampere or less), the dc output voltage will be close to the peak voltage of the source, or $1.4E_{rms}$. A typical application of the circuit would be to obtain a low bias voltage from a heater winding; the + side of the output can be grounded by reversing the polarity of the rectifier and capacitor. As with all half-wave rectifiers, the output voltage drops quickly with increased current demand.

The resistor R1 in Fig. 5-16 is included to limit the current through the rectifier, in accordance with the manufacturer's rating for the diode. If the resistance of the transformer winding is sufficient, R1 can be omitted.

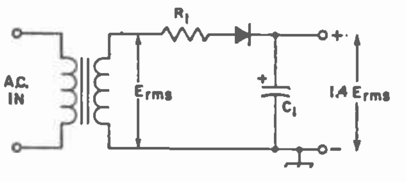


Fig. 5-16 — If the current demand is low, a simple half-wave rectifier will deliver a voltage increase. Typical values, for E_{RMS} = 117 and a load current of 1 mA:
C1 — 50-µF, 250-V electrolytic.
E_{output} — 160 volts.
R1 — 22 ohms.

VOLTAGE DOUBLERS

Several types of voltage-doubling circuits are in common use. Where it is not necessary that one side of the transformer secondary be at ground potential, the voltage-doubling circuit of Fig. 5-17 is used. This circuit has several advantages over the voltage-doubling circuit to be described later. For a given output voltage, compared to the full-wave rectifier circuit (Fig. 5-8B), this full-wave doubler circuit requires rectifiers having only half the PRV rating. Again for a given output voltage, compared to a full-wave bridge circuit (Fig. 5-8C) only half as many rectifiers (of the same PRV rating) are required.

Resistors R1 in Fig. 5-17 are used to limit the surge currents through the rectifiers. Their values are based on the transformer voltage and the rectifier surge-current rating, since at the instant the power supply is turned on the filter capacitors look like a short-circuited load. Provided the limiting resistors can withstand the surge current, their current-handling capacity is based on the maximum load current from the supply.

Output voltages approaching twice the peak voltage of the transformer can be obtained with the voltage-doubling circuit of Fig. 5-17. Fig. 5-18 shows 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.

When one side of the transformer secondary must be at ground potential, as when the ac is derived from a heater winding, the voltage-multiplying circuits of Fig. 5-19 can be used. In the voltage-doubling circuit at A, C1 charges through the left-hand rectifier during one half of the ac cyle; the other rectifier is nonconductive during this time. During the other half of the cycle the right-hand rectifier conducts and C2 becomes charged; they see as the source the transformer plus the voltage in C1. By reversing the polarities of the capacitors and rectifiers, the + side of the output can be grounded.

VOLTAGE TRIPLING AND QUADRUPLING

A voltage-tripling circuit is shown in Fig. 5-19B. On one half of the ac cycle C1 is charged to the source voltage through the left-hand rectifier. On the opposite half of the cycle the middle rectifier conducts and C2 is charged to twice the source voltage, because it sees the transformer plus the charge in C1 as its source. (The left-hand rectifier is cut off on this half cycle.) At the same time the right-hand rectifier conducts and, with the transformer and the charge in C2 as the source, C3 is charged to three times the transformer voltage. The + side of the output can be grounded if the polarities of all of the capacitors and rectifiers are reversed.

The voltage-quadrupling circuit of Fig. 5-19C works in substantially similar fashion.

In any of the circuits of Fig. 5-19, the output voltage will approach an exact multiple (2, 3 or 4, depending upon the circuit) of the peak ac voltage when the output current drain is low and the capacitance values are high.

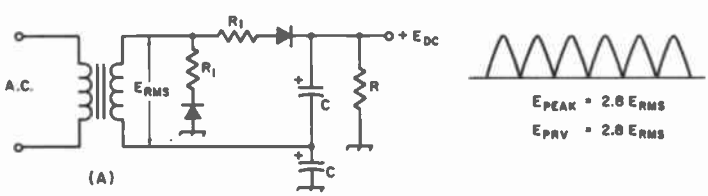


Fig. 5-17 — Full-wave voltage-doubling circuit. Values of limiting resistors, R1, depend upon allowable surge currents of rectifiers.

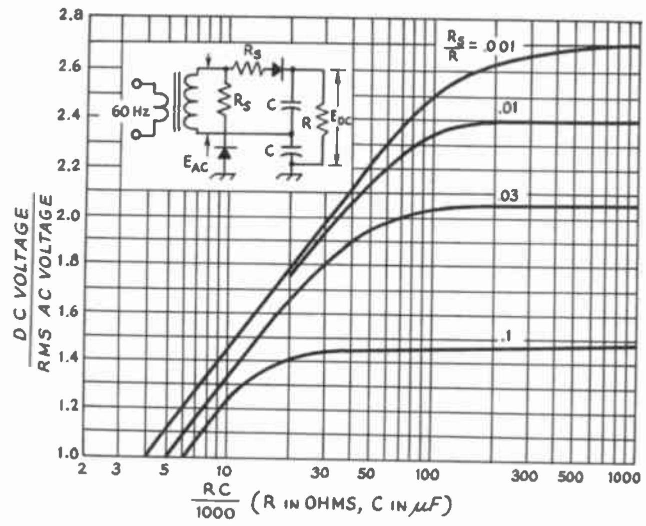
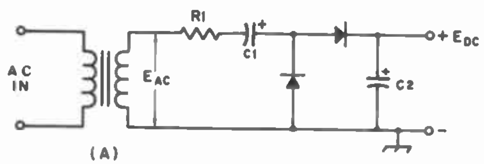
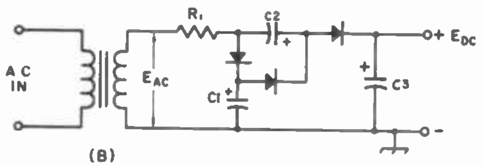


Fig. 5-18 – Dc output voltages from a full-wave voltage-doubling circuit as a function of the filter capacitances and load resistance. For the ratio R_s/R and for the RC product, resistances are in ohms and capacitance is in microfarads. Equal resistance values for R_s and equal capacitance values for C are assumed.





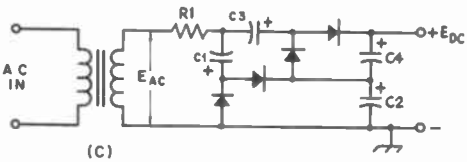


Fig. 5-19 – Voltage-multiplying circuits with one side of transformer secondary grounded. (A) Voltage doubler (B) Voltage tripler (C) Voltage quadrupler.

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 $4E_{peak}$

VOLTAGE DROPPING

Series Voltage-Dropping Resistor

Certain plates and screens of the various tubes in a transmitter or receiver often require a variety of operating voltages differing from the output voltage of an available power supply. In most cases, it is not economically feasible to provide a separate power supply for each of the required voltages. If the current drawn by an electrode (or combination of electrodes operating at the same voltage) is reasonably constant under normal operating conditions, the required voltage may be obtained from a supply of higher voltage by means of a voltage-dropping resistor in series, as shown in Fig. 5-20A. The value of the series, resistor, R1, may be obtained from Ohm's Law,

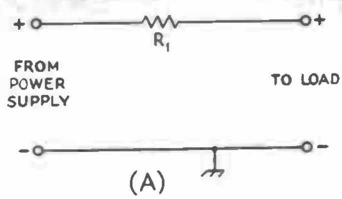
$$R = \frac{E_d}{I}$$

where E_d is the voltage *drop* required from the supply voltage to the desired voltage and I is the total rated current of the load.

Example: The plate of the tube in one stage and the screens of the tubes in two other stages require an operating voltage of 250. The nearest available supply voltage is 400 and the total of the rated plate and screen currents is 75 mA. The required resistance is

$$R = \frac{400 - 250}{.075} = \frac{150}{.075} = 2000 \text{ ohms}$$

The power rating of the resistor is obtained from P (watts) $= I^2R = (0.075)^2 \times (2000) = 11.2$ watts. A 20-watt resistor is the nearest safe rating to be used.



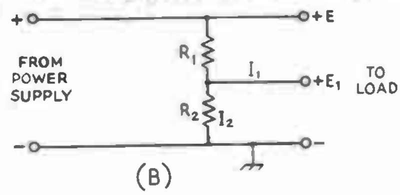
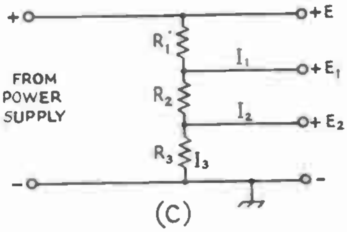


Fig. 5–20 — A — Series voltage-dropping resistor.
B — Simple voltage divider.

$$R2 = \frac{E1}{I2}; \; R1 = \frac{E - E1}{I1 + I2}$$

$I2$ must be assumed.



C — Multiple divider circuit.

$$R3 = \frac{E2}{I3}; \; R2 = \frac{E1 - E2}{I2 + I3}; \; R1 = \frac{E - E1}{I1 + I2 + I3}$$

$I3$ must be assumed.

Voltage Dividers

The regulation of the voltage obtained in this manner obviously is poor, since any change in current through the resistor will cause a directly proportional change in the voltage drop across the resistor. The regulation can be improved somewhat by connecting a second resistor from the low-voltage end of the first to the negative power-supply terminal, as shown in Fig. 5-20B. Such an arrangement constitutes a voltage divider. The second resistor, R2, acts as a constant load for the first, R1, so that any variation in current from the tap becomes a smaller percentage of the total current through R1. The heavier the current drawn by the resistors when they alone are connected across the supply, the better will be the voltage regulation at the tap.

Such a voltage divider may have more than a single tap for the purpose of obtaining more than one value of voltage. A typical arrangement is shown in Fig. 5-20C. The terminal voltage is E, and two taps are provided to give lower voltages, E1 and E2, at currents I1 and I2 respectively. The smaller the resistance between taps in proportion to the total resistance, the lower is the voltage between the taps. The voltage divider in the figure is made up of separate resistances, R1, R2 and R3. R3 carries only the bleeder current, I3; R2 carries I2 in addition to I3; R1 carries I1, I2 and I3. To calculate the resistances required, a bleeder current, I3, must be assumed; generally it is low compared with the total load current (10 percent or so). Then the required values can be calculated as shown in the caption of Fig. 5-20, I being in decimal parts of an ampere.

The method may be extended to any desired number of taps, each resistance section being calculated by Ohm's Law using the needed voltage drop across it and the total current through it. The power dissipated by each section may be calculated by multiplying I and E or I^2 and R.

VOLTAGE STABILIZATION

Gaseous Regulator Tubes

There is frequent need for maintaining the voltage applied to a low-voltage low-current circuit at a practically constant value, regardless of the voltage regulation of the power supply or variations in load current. In such applications, gaseous regulator tubes (0B2/VR105, 0A2/VR150, etc.) can be used to good advantage. The voltage drop across such tubes is constant over a moderately wide current range. Tubes are available for regulated voltages near 150, 105, 90 and 75 volts.

The fundamental circuit for a gaseous regulator is shown in Fig. 5-21. The tube is connected in series with a limiting resistor, R1, across a source of voltage that must be higher than the starting voltage. The starting voltage is about 30 to 40 percent higher than the operating voltage. The load is connected in parallel with the tube. For stable operation, a minimum tube current of 5 to 10 mA is required. The maximum permissible current with most types is 40 mA; consequently, the load current cannot exceed 30 to 35 mA if the voltage is to be stabilized over a range from zero to maximum load. A single VR tube may also be used to regulate the voltage to a load current of almost any value as long as the *variation* in the current does not exceed 30 to 35 mA. If, for example, the average load current is 100 mA, a VR tube may be used to hold the voltage constant provided the current does not fall below 85 mA or rise above 115 mA.

The value of the limiting resistor must lie between that which just permits minimum tube current to flow and that which just passes the maximum permissible tube current when there is no load current. The latter value is generally used. It is given by the equation:

$$R = \frac{(E_s - E_r)}{I}$$

where R is the limiting resistance in ohms, E_s is the voltage of the source across which the tube and resistor are connected, E_r is the rated voltage drop across the regulator tube, and I is the maximum tube current in amperes (usually 40 mA, or .04 A).

Two tubes may be used in series to give a higher regulated voltage than is obtainable with one, and also to give two values of regulated voltage. Regulation of the order of 1 percent can be obtained with these regulator tubes when they are operated within their proper current range. The capacitance in shunt with a VR tube should be limited to 0.1 μF or less. Larger values may cause the tube drop to oscillate between the operating and starting voltages.

ZENER DIODE REGULATION

A Zener diode (named after Dr. Carl Zener) can be used to stabilize a voltage source in much the same way as when the gaseous regulator tube is used. The typical circuit is shown in Fig. 5-22A. Note that the cathode side of the diode is connected to the positive side of the supply. The electrical characteristics of a Zener diode under conditions of forward and reverse voltage are given in Chapter 4.

Zener diodes are available in a wide variety of voltages and power ratings. The voltages range from less than 2 to a few hundred, while the power ratings (power the diode can dissipate) run from less than 0.25 watt to 50 watts. The ability of the Zener diode to stabilize a voltage is dependent upon the conducting impedance of the diode, which can be as low as one ohm or less in a low-voltage high-power diode to as high as a thousand ohms in a low-power high-voltage diode.

Diode Power Dissipation

Unlike gaseous regulator tubes, Zener diodes of a particular voltage rating have varied maximum current capabilities, depending upon the power ratings of each of the diodes. The power dissipated in a diode is the product of the voltage across it and the current through it. Conversely, the maximum current a particular diode may safely conduct equals its power rating divided by its voltage rating. Thus, a 10-V 50-W Zener diode, if

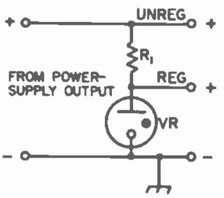


Fig. 5-21 — Voltage stabilization circuit using a VR tube. A negative-supply output may be regulated by reversing the polarity of the power-supply connections and the VR-tube connections from those shown here.

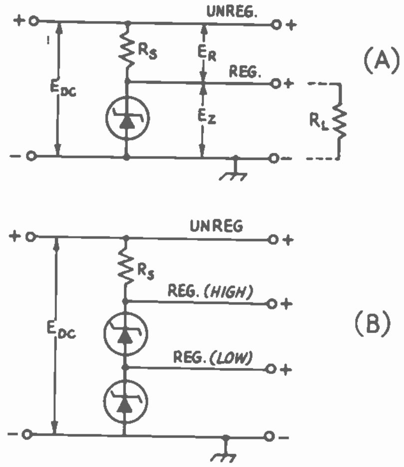


Fig. 5-22 — Zener-diode voltage regulation. The voltage from a negative supply may be regulated by reversing the power-supply connections and the diode polarities.

operated at its maximum dissipation rating, would conduct 5 amperes of current. A 10-V 1-W diode, on the other hand, could safely conduct no more than 0.1 A, or 100 mA. The conducting impedance of a diode is its voltage rating divided by the current flowing through it, and in the above examples would be 2 ohms for the 50-W diode, and 100 ohms for the 1-W diode. Disregarding small voltage changes which may occur, the conducting impedance of a given diode is a function of the current flowing through it, varying in inverse proportion.

The power-handling capability of most Zener diodes is rated at 25 degrees C, or approximately room temperature. If the diode is operated in a higher ambient temperature, its power capability must be derated. A typical 1-watt diode can safely dissipate only 1/2 watt at 100 degrees C.

Limiting Resistance

The value of R_s in Fig. 5-22 is determined by the load requirements. If R_s is too large the diode will be unable to regulate at large values of I_L, the current through R_L. If R_s is too small, the diode dissipation rating may be exceeded at low values of I_L. The optimum value for R_s can be calculated by:

$$R_S = \frac{E_{DC}\,(\text{min}) - E_Z}{1.1\, I_L\,(\text{max})}$$

When R_S is known, the maximum dissipation of the diode, P_D, may be determined by:

$$P_D = \left[\frac{E_{DC}\,(\text{max}) - E_Z}{R_S} - I_L\,(\text{min})\right] E_Z$$

In the first equation, conditions are set up for the Zener diode to draw 1/10 the maximum load current. This assures diode regulation under maximum load.

Example: A 12-volt source is to supply a circuit requiring 9 volts. The load current varies between 200 and 350 mA.

E_Z = 9.1 V (nearest available value).

$$R_S = \frac{12 - 9.1}{1.1 \times 0.35} = \frac{2.9}{0.385} = 7.5 \text{ ohms}$$

$$P_D = \left[\frac{12 - 9.1}{7.5} - 0.2\right] 9.1 = .185 \times 9.1 = 1.7 \text{ W}$$

The nearest available dissipation rating above 1.7 W is 5; therefore, a 9.1-V 5-W Zener diode should be used. Such a rating, it may be noted, will cause the diode to be in the safe dissipation range even though the load is completely disconnected [I_L (min) = 0].

Obtaining Other Voltages

Fig. 5-22B shows how two Zener diodes may be used in series to obtain regulated voltages not normally obtainable from a single Zener diode, and also to give two values of regulated voltage. The diodes need not have equal breakdown voltages, because the arrangement is self equalizing. However, the current-handling capability of each diode should be taken into account. The limiting resistor may be calculated as above, taking the sum of the diode voltages as E_Z, and the sum of the load currents as I_L.

ELECTRONIC VOLTAGE REGULATION

Several circuits have been developed for regulating the voltage output of a power supply electronically. While more complicated than the VR-tube and Zener-diode circuits, they will handle higher voltage and current variations, and the output voltage may be varied continuously over a wide range.

Voltage regulators fall into two basic types. In the type most commonly used by amateurs, the dc supply delivers a voltage higher than that which is available at the output of the regulator, and the regulated voltage is obtained by dropping the voltage down to a lower value through a dropping "resistor." Regulation is accomplished by varying either the current through a fixed dropping resistance as changes in input voltage or load currents occur (as in the VR-tube and Zener-diode regulator circuits), or by varying the equivalent resistive value of the dropping element with such changes. This latter technique is used in electronic regulators where the voltage-dropping element is a vacuum tube or a transistor, rather than an actual resistor. By varying the dc voltage at the grid or current at the base of these elements, the conductivity of the device may be varied as necessary to hold the output voltage constant. In solid-state regulators the series-dropping element is called a pass transistor. Power transistors are available which will handle several amperes of current at several hundred volts, but solid-state regulators of this type are usually operated at potentials below 100 volts.

The second type of regulator is a switching type, where the voltage from the dc source is rapidly switched on and off (electronically). The average dc voltage available from the regulator is proportional to the duty cycle of the switching wave form, or the ratio of the ON time to the total period of the switching cycle. Switching frequencies of several kilohertz are normally used to avoid the need for extensive filtering to smooth the switching frequency from the dc output.

The above information pertains essentially to voltage regulators. A circuit can also be constructed to provide current regulation. Such regulation is usually obtained in the form of current limitation – to a maximum value which is either preset or adjustable, depending on the circuit. Relatively simple circuits, such as described later,

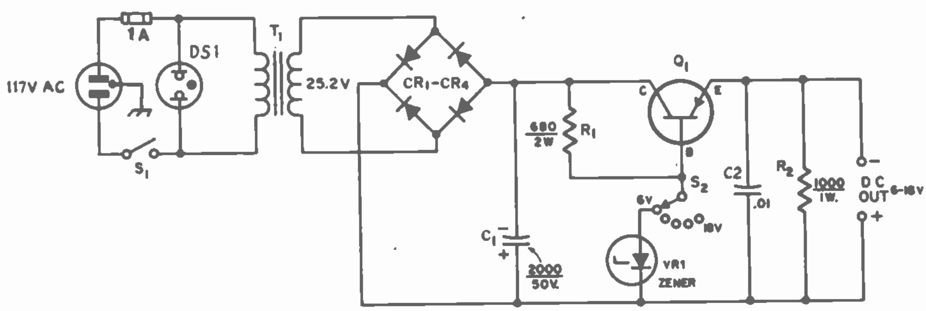


Fig. 5-23 – Schematic diagram of the power supply. Capacitances are in μF; capacitors marked with a polarity are electrolytic. Resistances are in ohms; R1 and R2 are composition.

C1 – 2000-μF 50 volts dc electrolytic (Mallory CG23U50C1).
C2 – .01-μF disk ceramic.
CR1-CR4, incl. – 50 PRV 3-A silicon diode (Motorola 1N4719).
DS1 – Neon lamp assembly with resistor (Leecraft 32-2111).
Q1 – 2N1970.
S1 – Spst toggle switch.
S2 – Phenolic rotary, 1 section, 2-pole (1 used), 6-position, shorting (Mallory 3126J).
T1 – Filament transformer, 25.2 V, 2 A (Knight 54 D 4140 or similar).
VR1 – Voltage regulator diode.

can be used to provide current limiting only. Current limiting circuitry may also be used in conjunction with voltage regulators.

Solid-State Regulators

One of the simplest forms of solid-state regulation is shown at Fig. 5-23. A bridge rectifier supplies 25 volts dc to a series regulator transistor, Q1, whose base bias is established by means of a Zener diode, VR1, providing a voltage reference of a fixed level. C1 is the input capacitor for the filter. R1 is chosen to establish a safe Zener-diode current, which is dependent upon the wattage rating of the diode. A 1-watt Zener diode is adequate for the circuit of Fig. 5-23. R2 is a bleeder resistor and C2 is an rf bypass. If several output voltages are desired, say from 6 to 18 volts, Zener diodes from 6 to 18 volts can be wired to S2 as shown. When a 2N1970 is used at Q1, the value of R1 will be 680 ohms. This value offers a compromise for the 5 reference diodes used (6,9,12,15, and 18 volts).

The output of the supply is equal to the Zener voltage minus the emitter-to-base bias voltage of Q1. Both the Zener voltage and bias voltage will be approximately zero with only R2 as a load, but will rise to roughly 0.3 volt with a 1-A load connected to the output. An increase in load current lowers the unregulated dc input voltage which appears across VR1 and R1. Zener current is reduced, decreasing the voltage at which the diode regulates. How much the voltage drops depends upon the characteristics of the particular Zener employed.

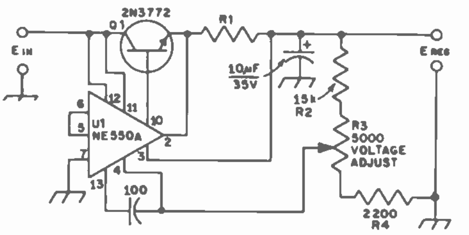
This power supply has very low output ripple. The main limitation of the circuit is the possibility of destroying Q1, the series-regulator transistor, when a dead short or heavy overload is connected across the output of the supply. To protect Q1 during normal operation, it should be mounted on a fairly large heat sink which is thermally coupled to the main chassis of the supply. The transistor should be insulated from the sink by means of a mica spacer and a thin layer of silicone grease. The sink can then be bolted directly to the chassis.

IC Regulators

The solid-state regulator described above provides only fixed voltages. Regulator circuits with the output voltage continuously variable over a wide range and with a very high degree of regulation can be built, but the number of circuit components is comparatively large when discrete components are used. Integrated-circuit devices can be used in a solid-state regulator circuit to replace many or all of the discrete components, depending on the output requirements. The voltage reference, control, shut-down (for current limiting) and pass-transistor driver elements are contained on a single silicon chip. The construction of a regulated power supply is simplified to a few interconnections if an IC regulator is used.

Fig. 5-24 is the diagram of a regulator using an IC and a single pass transistor. With a dc potential of 24 to 30 volts applied at E_{IN} the circuit as shown will provide an adjustable output voltage between 5 and 15. The circuit will handle up to 5 amperes of current, provided, of course, that the dc source will deliver this amount. If the load requires no more than 150 mA of current the pass transistor may be eliminated from the circuit altogether; in this case pins 2 and 10 of the IC should be interconnected.

The NE550 regulator will safely accept input voltages as high as 50, and output voltages may be adjusted by appropriate resistance values for R2, R3, and R4 from 2 to 40 volts. The value of R1 determines the shut-down current (maximum cur-



EXCEPT AS INDICATED, DECIMAL VALUES OF CAPACITANCE ARE IN MICROFARADS (µF); OTHERS ARE IN PICOFARADS (pF OR µµF); RESISTANCES ARE IN OHMS; k = 1000, M = 1000 000.

Fig. 5-24 — Schematic diagram of 15-V 5-A regulator (W1KLK, *QST* for November, 1971).

Q1 — Motorola power transistor; 30-cubic-inch heat sink required (Delco 7281366 radiator or equiv.).

R1 — 0.1-ohm resistor, made from 8 feet of No. 22 enam. copper wire.

R2, R4 — For text reference.

R3 — Linear taper.

U1 — Signetics IC.

Table 5-1

Voltage Divider			Current Limit	
V_{OUT}	R_A	R_B	I_{MAX}	R_1
3.6	6135	2967	.05	12
5	4417	3654	0.1	6
9	11,043	2442	0.5	1.2
12	14,724	2314	1.0	0.6
13.6	16,687	2272	1.5	0.4
15	18,405	2243	2	0.3
20	24,540	2177	2.5	0.24
28	34,356	2122	3.0	0.20
			5	0.12
			10	.006

Table 5-1 — Resistance values for various voltage and current outputs from the regulator of Fig. 5-24. These values were determined by mathematical calculation and are not necessarily available from stock supplies. The figures given do indicate the practical values which may be used along with an appropriate-value control for R2 in the circuit of Fig. 5-24.

R_A — R2 plus top portion of R3.

R_B — R4 plus bottom portion of R3.

rent which the circuit will deliver into a short circuit) and is usually selected to protect either the pass transistor or the power supply transformer, whichever has the lower current rating. Table 5-1 gives resistance values for various levels of voltage and current from the regulator.

The use of a high-gain pass device improves the output regulation, and a Darlington-connected pair is frequently employed. Of course it is easy to purchase a ready-made Darlington transistor, but the enterprising amateur can make his own, as shown in Fig. 5-25A. However, some of the IC regulators which are available on the market have so much internal gain that it is difficult to avoid oscillation with a high-gain pass transistor.

High-Current-Output Regulators

When a single pass transistor is not available to handle the current which may be required from a regulator, the current-handling capability may be increased by connecting two or more pass transistors in parallel. The circuits at B and C of Fig. 5-25 show the method of connection. The resistances in the emitter leads of each transistor are necessary to equalize the currents.

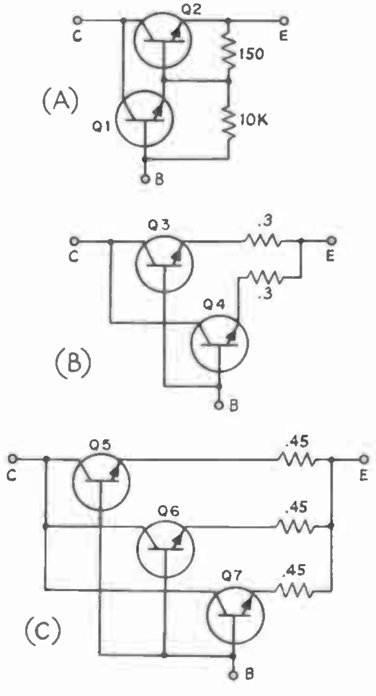


Fig. 5-25 — At A, a Darlington-connected pair for use as the pass element in a series-regulating circuit. At B and C, the method of connecting two or more transistors in parallel for high current output. Resistances are in ohms. The circuit at A may be used for load currents from 100 mA to 5 A, at B for currents from 6 to 10 A, and at C for currents from 9 to 15 A.

Q1 — Motorola MJE 340 or equivalent.
Q2-Q7, incl. — Power transistor such as 2N3055 or 2N3772.

Fixed-Voltage IC Regulators

IC regulators with all circuitry contained on a single silicon chip are becoming available for different values of fixed-voltage outputs. The LM309 five-volt regulator, manufactured by National Semiconductor and others, is one type of such ICs. These regulators are three-terminal devices, for making connections to the positive unregulated input, positive regulated output, and ground. They are designed for local regulation on digital-logic circuit-board cards to eliminate the distribution problems associated with single-point regulation. For this reason they are frequently called on-card regulators.

The LM309 is available in two common transistor packages. The LM309H in a TO-5 package can deliver output currents in excess of 200 mA if adequate heat sinking is provided, and the LM309K in the TO-3 power package can provide an output current greater than 1 A. The regulator is essentially blow-out proof, with current limiting included in the circuit. In addition, thermal shutdown is provided to keep the IC from overheating. If internal dissipation becomes too great, the regulator will shut down to prevent excessive heating.

It is not necessary to bypass the output of the LM309, although bypassing does improve immunity from transient responses. Input bypassing is needed, however, if the regulator is located very far from the filter capacitor of the power supply. Typical values of input bypass capacitance are 0.15 and 0.22 μF. Although designed primarily as a fixed-voltage regulator, the LM309 can be used to obtain a regulated output at voltages higher than five. This is done by returning the "ground" connection of the IC to a tap point on a voltage divider which is connected between the regulated output and a true circuit ground. An adjustable output regulator for voltages above five can be had if the "ground" pin is connected to the junction of a 300-ohm fixed resistor and one end of a 1000-ohm linear control. The opposite end of the 300-ohm resistor should be connected to the output pin, and the wiper contact and third lug of the control to a true circuit ground.

Switching Regulator

Switching regulators are used when it is necessary or desired to minimize power losses which would otherwise occur in the series pass transistor (or transistors) with large variations in input or output voltages. The basic operation of the switching regulator, known as the flyback type, may be understood by referring to Fig. 5-26A. Assume that the switch is closed and the circuit has been in operation long enough to stabilize. The voltage across the load, R_L, is zero, and the current through L is limited only by R_I, the internal resistance of the inductor. At the instant the switch is opened, the voltage across the load goes to a value higher than the source voltage, E, because of the series-aiding or "flyback" effect of the inductor. When the magnetic lines of flux about the inductor collapse completely, the voltage across R_L will be equal to that of the source (minus the small voltage drop across R_I). Each time the switch is closed and then opened, the process is repeater. By opening and closing the switch rapidly, voltage pulses may be applied across R_L which are higher than the dc input voltage. A capacitor may be connected across R_L to produce a dc output voltage. To keep the capacitor from discharging when the switch is closed, a diode can be connected in series with the load and its parallel-connected capacitor.

In a practical switching-regulator circuit the switching is performed by a transistor, as shown at B of Fig. 5-26. The transistor may be driven by any number of circuits. In the practical circuit shown later (Fig. 5-27) four sections make up the driving circuit, as shown in block diagram form in Fig. 5-26B. The oscillator triggers the monostable multivibrator and determines the frequency of operation. The sensor measures the output voltage and controls the pulse width of the multivibrator accordingly. The monostable multivibrator combines the signals from the oscillator and sensor to produce the correct pulse width. The driver receives the multivibrator output and drives the power transistor, Q1.

The voltage step-up capability of the inductor has been mentioned briefly. However, in choosing the value of the inductor, energy is an important consideration. During the time the transistor is turned on, the inductor stores energy. This energy is added to the supply and delivered to the load when the transistor turns off. The total energy must be enough to supply the load and maintain output voltage. As the load is increased, the transistor must remain on longer in order to store more energy in the inductor. The required value of inductance depends on frequency of operation, duty cycle, and load. A linear change in current through the inductor is a desirable condition and indicates operation is over a small segment of the inductor's charging and discharging curve. A powdered-iron-core inductor is normally used to

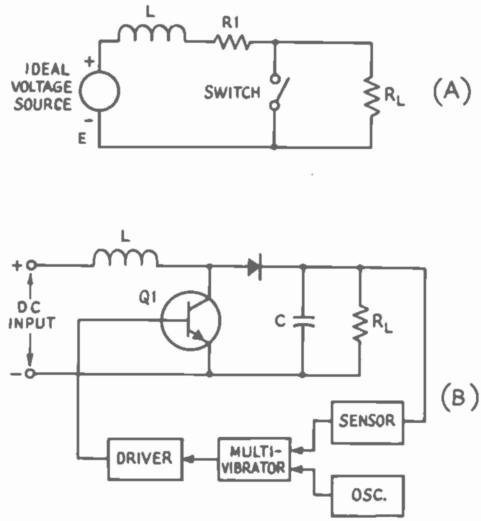


Fig. 5-26 — At A, the fundamental circuit of a flyback switching regulator, and at B, the elements of a practical circuit.

prevent a large inductance change with increased current.

Efficiency of the circuit depends mainly upon the switching and saturation losses of the power transistor. The peak current through the transistor is considerably greater than the input current. The flyback diode must have a fast reverse recovery time and low forward drop. There will be a large current spike through the transistor if the diode is slow.

The complete circuit of a switching regulator is given in Fig. 5-27. This regulator will handle 100 watts of power efficiently, at output voltages as much as 6 volts above the input voltage. The switching rate of the regulator is 9 kHz, and it operates with an input of 22 to 28 volts. Regula-

EXCEPT AS INDICATED, DECIMAL VALUES OF CAPACITANCE ARE IN MICROFARADS (μF); OTHERS ARE IN PICOFARADS (pF OR μμF); RESISTANCES ARE IN OHMS; k = 1 000, M = 1 000 000.

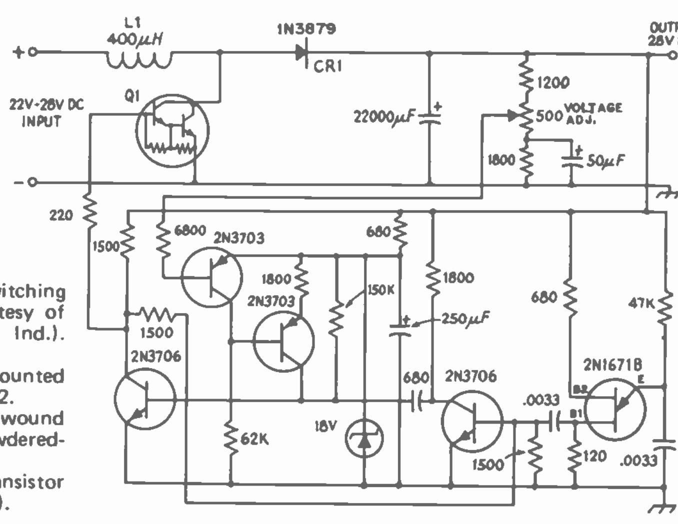


Fig. 5-27 — A 100-W 28-V switching regulator (circuit design courtesy of Delco Electronics, Kokomo, Ind.). All resistors are 1/2 W.

CR1 — Motorola rectifier mounted on Delco heat sink 7281352.

L1 — 124 turns No. 18 wire wound on Arnold BO79024-3 powdered-iron core.

Q1 — Darlington power transistor (Delco DTS 1020 or equiv.).

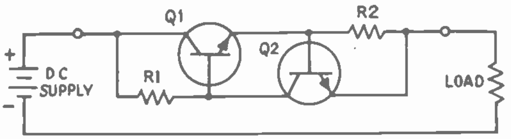


Fig. 5-28 — Two-terminal current limiter. See text for discussion of component values and types.

tion and ripple are less than 1 percent at full output. The switching device, Q1, is a commercially available Darlington transistor.

The efficiency of the circuit drops off at low power levels. This is because the losses of the circuit are not proportional to the output power. Maximum efficiency occurs at about 80 watts because the duty cycle of the transistor is an optimum for the chosen value of the inductor. Whenever the input voltage increases above 28 volts, the output voltage tracks the input. The difference between the two voltages is the drop in the flyback diode.

Output voltage variations resulting from changes in ambient temperature are caused by two major factors; positive temperature coefficient of the Zener diode, and the negative temperature coefficient of the emitter-base junctions of the transistors. One way to compensate partially for temperature is to connect diodes that have negative temperature coefficients in series with the Zener diode.

Two-Terminal Current Limiter

The simple circuit of Fig. 5-28 performs the current limiting function of fuses or circuit breakers. The circuit uses only two transistors and two resistors. The necessary supply voltage for operation is obtained from the power source being protected, with the load functioning as the return to the power source. Q1 is a series element which allows current, up to a desired maximum, to flow to the load. R1 provides a suitable bias for Q1 to permit such current to flow. R2 is a sensing resistor interposed between the series transistor and the load, and provides bias for Q2. Normally this bias is low enough to prevent Q2 from conducting. Q2 controls the bias applied to Q1. When excess current flows through R2 as a result of a circuit malfunction or a short across the load, the voltage drop across R2 rises, biasing Q2 into conduction. When Q2 turns on, it reduces the bias on Q1 and limits the amount of current flow. The maximum amount of current flow can be varied by changing the value of R2. If an adjustable limiting level is desired, R2 may be a variable resistor. The limiting level is an inverse function of the resistance value.

0-25 VOLT ADJUSTABLE POWER SUPPLY

For most amateur work the voltages needed from transistorized power supplies fall into two general ranges; 5 volts for digital circuits and 12 volts. On occasion there is the need for other values and this power supply is capable of providing voltages from 1.3 to 25.4. The heart of the supply is the National LM317K adjustable voltage regulator. In the following configuration the supply is capable of 3 amperes output current throughout its voltage range.

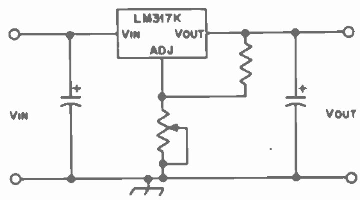


Fig. 1 — Schematic diagram of the basic regulator.

Circuit Description

The LM317K, available in three case styles, is a completely self-contained adjustable regulator. The basic, adjustable, regulated supply requires no more than a dc source, the regulator and two resistors (Fig. 1). In the basic configuration the regulator is capable of supplying up to 1.5 A output current. To provide higher-output currents, an external pass transistor has been added (Fig. 2). C1 and C2 (Fig. 2) are the usual capacitors to improve transient response and reduce the noise on the output voltage. C1 may be eliminated if the regulator is located physically close to the filter capacitor. The 5600-ohm resistor was added in parallel with the potentiometer to limit the output voltage to approximately 25. The ripple on the output voltage is decreased by the addition of C3 to the adjustment lead. If the regulator is going to be used over 25 volts, a diode (CR5) should be connected for protection if the input or output is shorted.

Depending upon the builder's preferences, meters may be added to the power supply. A shunt resistor was added to the meter reading the output current. The method for determining the value of the shunt resistor is discussed in the Test Equipment and Measurements section of this book.

Construction

The transformer secondary voltage should be in the range of 20 to 24 V ac when drawing slightly more than the desired load current. In addition to

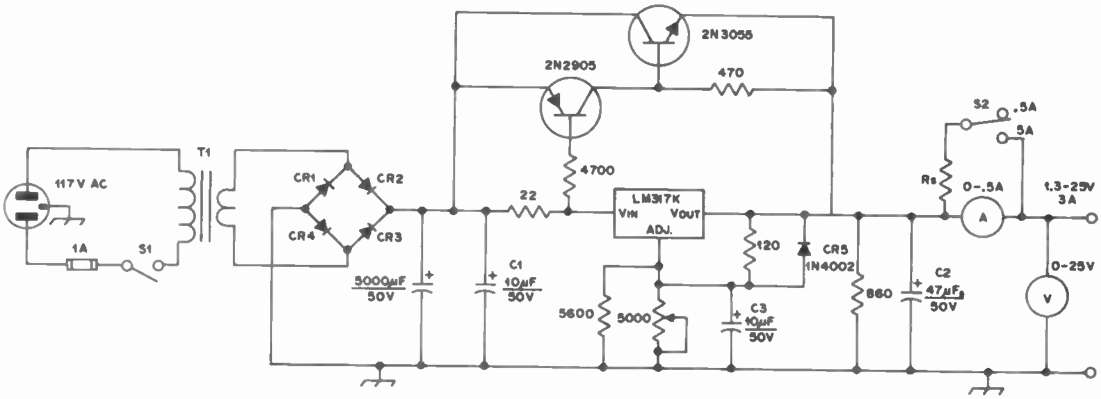


Fig. 2 — Schematic diagram of the complete power supply. All resistors are 1/2 watt composition. C1 through C3 are solid tantalum capacitors. T1 and CR1 through CR4 are discussed in the text. R1 is a panel-mounted linear-taper potentiometer. S1 and S2 are spst toggle switches. The meters are Calectro DI-916 and DI-923. If the pass transistors are not be used, the 22-ohm resistor between C1 and the regulator input to be omitted.

the load current, the transformer must supply a small amount of power to the regulator. The rectifier can be four individual diodes or an encapsulated bridge module. If individual diodes are used their rating should be at least 50 PRV and 1.5 amperes (3 A or more if pass transistors are used). An encapsulated bridge must have the same minimum ratings. A Radio Shack 276-1146 encapsulated bridge will be more than adequate.

The 2N3055 and the regulator will both require a heat sink which can be homemade. A suitable commercially made heat sink would be the Radio Shack 276-1360. Due to the different transformer and heat sink sizes the cabinet can be any type that the builder desires.

The template (Fig. 3) is the etching pattern for a pc board that will hold all the components external to the regulator. This pc board could be etched and then mounted above the regulator. If it is not desired to use a pc board, all the components can be mounted on terminal strips.

One point that should be noted is the proper wire size for the interconnecting wiring. The load regulation is a function of the resistance of the wiring. A value as small as 0.05 ohm can decrease the regulation by as much as a factor of eleven.

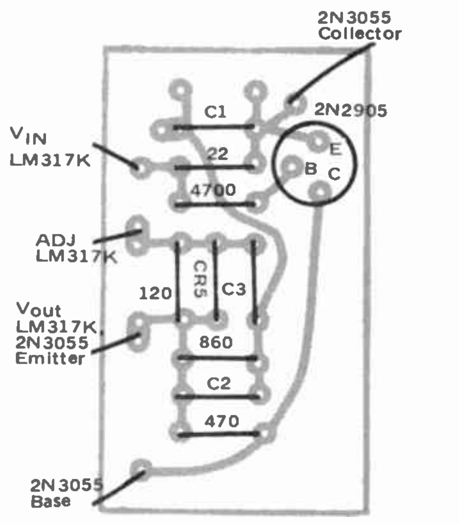


Fig. 3 — Template for pc board. View from the foil side.

Adjustment and Operation

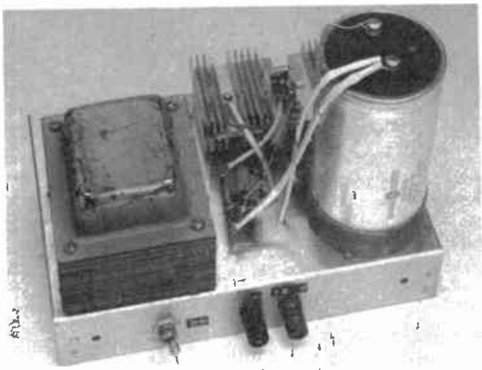
The value of the 5600-ohm resistor may be changed to suit the builder. Lowering its value will decrease the maximum output voltage available. No other adjustment is necessary for operation of this power supply.

A POWER SUPPLY WITH A REBUILT TRANSFORMER

The circuit for the supply is shown in Fig. 1. The power transformer T1 provides approximately 18-volts ac which is rectified via U1 and then regulated at 12 volts. The regulator is a National Semiconductor LM340K-12. A pnp pass transistor is used to increase the current capabilities of the supply. With this supply we have used up to 10 amperes to power an fm transceiver. All of the components were obtained as surplus, with the exception of the power transformer.

Rewinding Power Transformers

The newcomer to ham radio may feel that rewinding a power transformer is an impossible task. However, let us reassure you, it is a very simple process and the only major expenditure is



time. We found that the old tube-type TV sets that are now common at garage and tag sales are almost tailor-made junk boxes for the enterprising amateur. For the type of power supply we describe here, practically any TV transformer has the necessary power capabilities. Twelve volts at 10 amperes equals 120 watts, and even with a 100-percent excess rating (240 watts) such transformers are common in TV sets.

Taking the Transformer Apart

The first step in the rebuilding process is to remove the transformer from the TV chassis. Then, label the windings. The primary or input winding will be connected to the ac line, probably through a switch on the front of the chassis and a fuse or fuse holder on the rear. The 5-volt winding will be connected to the filament terminals (2 and 8) on the rectifier socket, which is usually a 5U4G. Two of the leads from the high-voltage winding will be connected to the plate terminals (4 and 6) on the rectifier-tube socket. The center-tap lead of the high-voltage winding probably will be grounded to the chassis. There will probably be two 6.3-volt windings. The leads from one of these will go into the shielded compartment on top of the TV chassis and be connected to a tube socket in the compartment. The other 6.3-volt winding supplies all the other tube heaters in the set. Tag all leads before removing the transformer.

After identifying the windings, remove the four nuts and bolts that hold the transformer together and also take off the metal covers, assuming the unit has them. Don't worry about the transformer falling apart when you remove the bolts; it won't. Look the unit over carefully and try to determine which layers of windings are which. In most cases the winding nearest the core will be the primary. Usually the order will be something like this: first, the primary; next, the high voltage; then, the 5- and 6.3-volt low-current filament windings; and last, the heavier-current 6.3-volt winding.

Examine the lamination arrangement. Note that the laminations are probably inserted in groups. On one side of the stack there may be three I units and below that three E units, alternating through the entire stack. Note how the top and bottom of the stack are assembled so that you'll be able to put it back in this same order when you complete the winding job.

Getting the laminations apart is not a difficult job, but it should be done carefully. Insert a thin knife blade between the end piece and the rest of the core to break the varnish seal, so the end piece will be loose. Using a block of wood butted against the edge of the piece, drive it out of the core with light taps of a hammer. Alternate between the two ends so the piece will come out straight. Continue by breaking the next group of laminations free with the knife blade, then carefully driving them out. After a few groups have been removed the hammer won't be needed, as the broken-loose laminations can be pulled out by hand. Be careful not to bend the laminations when removing them. If the edges get nicked in hammering, file them smooth before reassembling the core after the new windings are finished.

Once the laminations are finished, remove the

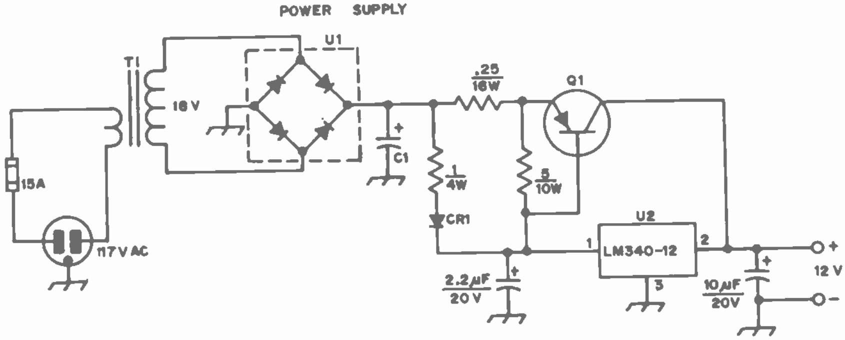


Fig. 1 — Circuit diagram of the power supply. Circuit designations not listed below are for text references.
C1 — 2000 µF (or more), 50-V electrolytic.
CR1 — Silicon diode, 3A 35 PRV.
Q1 — Motorola HEP233, 237 or similar.
U1 — 100 V PRV, 25 A.
U2 — National Semiconductor LM340K-12 regulator.

high-voltage winding by pulling out the wire. If you are lucky you can start it just by pulling on one of the high-voltage leads. However, it is more than likely that the end of the winding will break off, because the wire size will be rather small. If it breaks you'll have to dig in with a knife or probe to get at the wire. Once you get it started the layers come out rather easily. When you get most of the high-voltage winding out you'll see that you can separate the primary winding section from the outer windings. Be careful not to disturb the insulation around the primary winding. Incidentally, in the unit we took apart, and this will probably hold true for most TV transformers, the wire size on the primary was No. 18 enameled.

After you've cleared away the high-voltage winding, remove the 5-volt rectifier-filament winding and *most* carefully count the number of turns. There will probably be approximately 10 turns, but count them to make sure. The number of turns on this winding will tell you how many turns you need for each volt you expect to get with the new windings you will put on. For example, if there are 10 turns on the 5-volt winding, the transformer is wound on the basis of two turns per volt. It doesn't make any difference whether the windings are near the center of the core or at the side; the turns per volt will be the same.

Putting on the New Winding

For this supply, the current rating is 10 amperes, so a wire size that will carry this current is required. No. 12 solid, enamel-covered wire handles our requirements. The transformer used in this supply required two turns per volt, and 18-volts ac was needed. This works out to a total of 36 turns of No. 12. To calculate how much wire you need, take a scrap length of wire or string, make a full turn around the core containing the primary winding, then measure how long the piece of wire or string is. Multiply this by 38, then add about three feet for lead lengths and slop.

Clamp one end of the wire in a vise and making sure there are no kinks, start winding the wire over the section that had the primary winding. Start as close to the edge as possible and keep the wire taut as you wind on the turns. The reason for starting close to the edge is that as you put layers on, each layer has to be progressively narrower, otherwise the end turns may slip off. After the first layer is wound, hold the ends in place with Scotch tape. Ordinary household waxed paper can be used between the layers. A single layer or sheet of paper is adequate insulation between layers. Wrap a sheet tightly around the first layer of winding and fasten the end of the paper with small pieces of Scotch tape. Try to keep the starting point for the next layer as close to the outside turns of the previous layer as possible and always wind in the same direction.

Be sure to bring all leads and taps out on the same side of the core so the transformer covers will go back in place without interference.

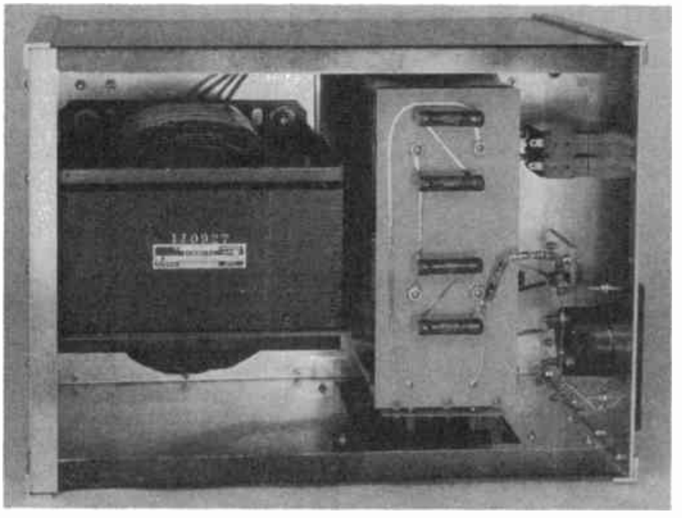
Once all of the turns are on, cover the windings with a couple of layers of electrician's tape (enough to secure the windings). The transformer can now be put back together. If there is too much open area between the top and bottom of the windings and the iron core, make up some smooth wooden wedges and gently drive them between the windings and the iron core. This will help prevent transformer hum or rattle. In our transformer we slipped some tubing insulation over the leads, where they came through the transformer housing, just to prevent chafing of the wire enamel covering.

The rest of the power supply follows conventional practice and additional details can be found in *QST* for November, 1976 (pp. 29-31). Other components are usually obtainable through dealers (such as those listed in *QST* or in the construction chapter of this handbook). Don't overlook the flea markets at hamfests and other amateur activities. This is perhaps the best place for finding a variety of components at low cost.

A "SANITARY" HIGH-VOLTAGE SUPPLY

Power supplies aren't usually noted for adding decor to the shack. Most hams would rather hide them so that non-ham visitors won't ask, "What's that ugly looking thing?" However, an attempt was made to improve the appearance of this model along with the function of providing high voltage for general amplifier purposes. Not all the additions are frivolous. For instance, the use of "rug runners" instead of the usual sharp corners on the bottom of the unit. prevents gouging an easily damaged surface such as a bench or floor.

The diode bridge rectifier is mounted on a separate pc board that can be removed easily. Accidental contact is prevented by a Plexiglas sheet which also permits viewing of the circuit board while it is still in the power supply. Although a sheet-metal cutter and bender were used to fabricate the sides, a "cut-and-file" method could result in a similar job if the builder were willing to spend the time. Either that, or angle brackets (such as those on the front of the unit) could be used inside of the top and bottom covers in order to form an overlap surface for the covers.



Interior view of the supply.

Circuit Details

The power supply employs a full-wave bridge rectifier and is capable of 1-A output at 3400 V dc. Primary-circuit and surge considerations are simplified by the use of 234 V ac instead of 117 V. While the addition of a 234-V line might seem like an unjustified inconvenience, experience has proven this approach to be the most acceptable method. However, additional surge protection is afforded by the use of the 10-Ω, 10-watt resistors in the secondary of T1.

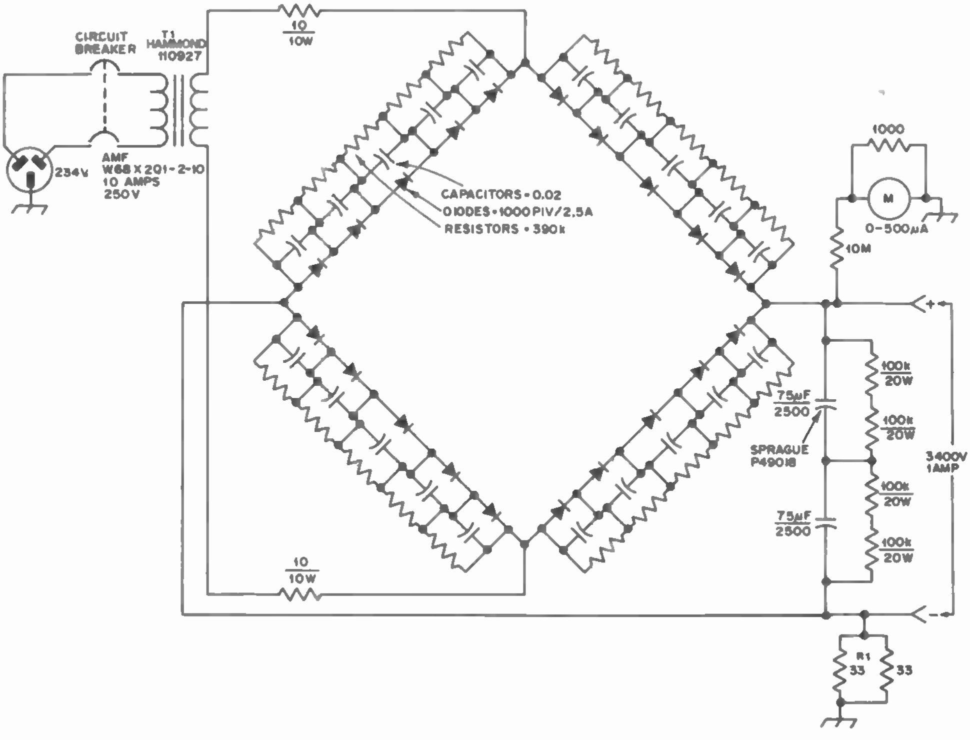


Fig. 1 — Schematic diagram of the power supply showing suggested component values. The paralleled 33-Ω resistors (R1) are for metering purposes using the B— lead. They can be omitted and the minus lead tied to ground if a different system is to be used.

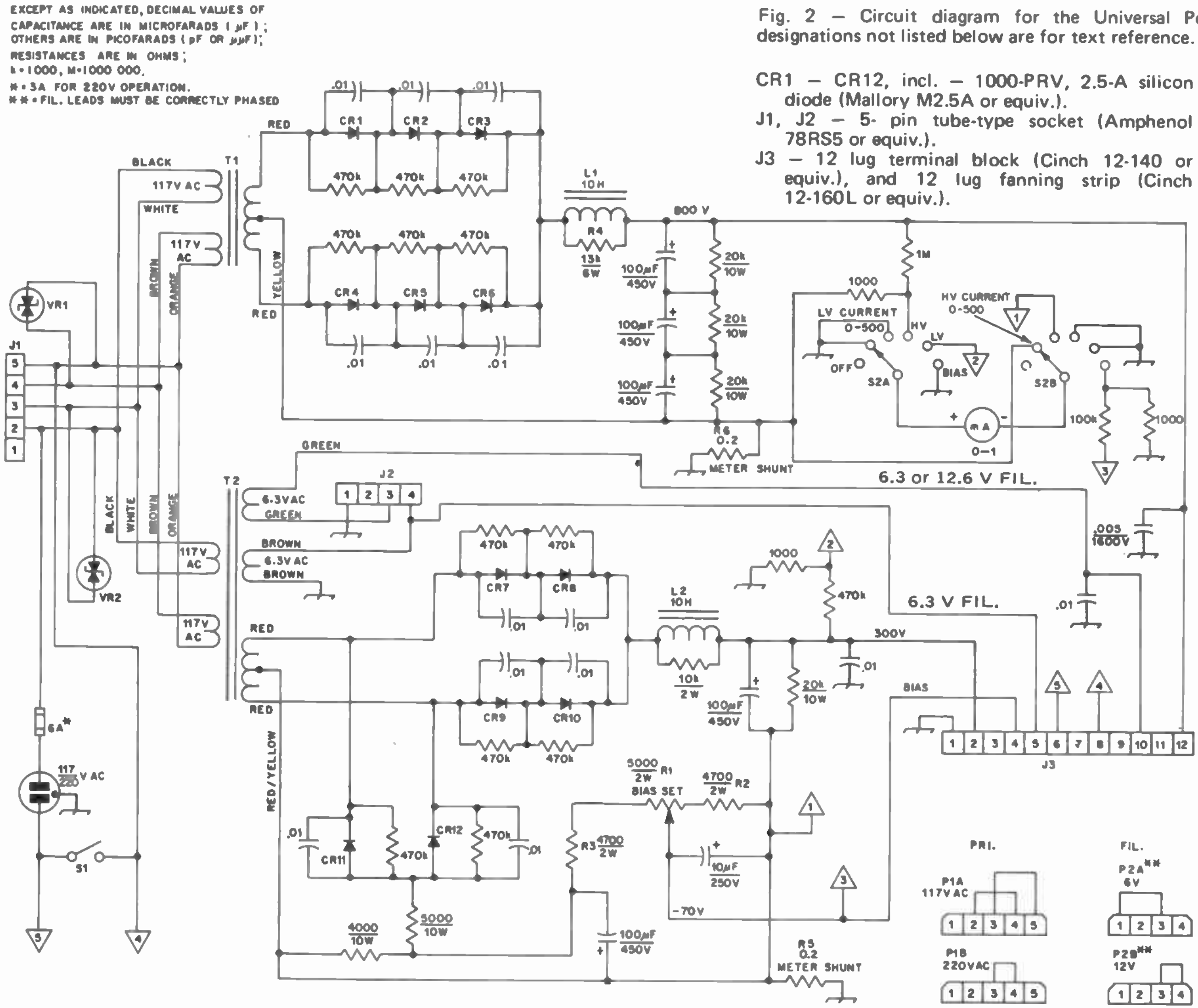


Fig. 2 — Circuit diagram for the Universal Power Supply. Component designations not listed below are for text reference.

CR1 — CR12, incl. — 1000-PRV, 2.5-A silicon diode (Mallory M2.5A or equiv.).
J1, J2 — 5- pin tube-type socket (Amphenol 78RS5 or equiv.).
J3 — 12 lug terminal block (Cinch 12-140 or equiv.), and 12 lug fanning strip (Cinch 12-160L or equiv.).
L1 — 10 H, 200 mA (Hammond 193J).
L2 — 10 H, 300 mA (Hammond 193M).
P1, P2 — 5-pin plugs to mate J1 and J2, 4 req'd (Amphenol 86-PM5 or equiv.).
R1 — 5-watt linear-taper control.
R2, R3 — For text reference.
R4 — Three 39,000-ohm 2-watt resistors connected in parallel.
R5, R6 — See text.
S1 — Spst toggle rated at 6 A or greater.
S2 — 2-pole 6-position rotary, nonshorting (Centralab 1411 or equiv.).
T1 — Dual primary, 117 or 220 V ac; secondary 890 volts each side of center tap at 300 mA (Hammond type 101059).
T2 — Dual primary, 117 or 220 V ac; secondary 350 volts each side of center tap at 175 mA, 6.3 volts ac at 6 A, 6.3 volts ac at 5 A (Hammond special 273BX).
VR1, VR2 — Thyrector assembly (G E 6RS20SP4B4).

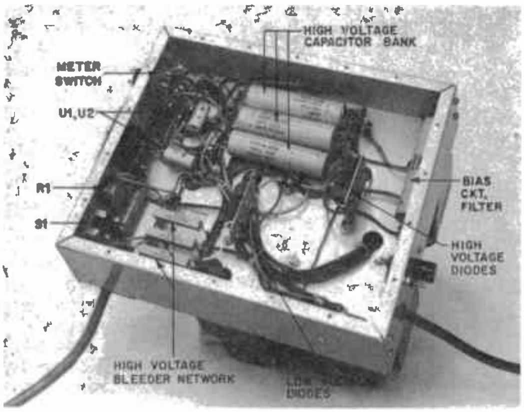


Fig. 3 — Bottom view of the Universal Power Supply.

from the panel of the station transmitting gear. But what happens if an instance arises where a particular voltage (or combination of voltages) is needed for an experimental project? Can that "black box" in the corner be pressed readily into service? And what about the amateur who buys two power supplies for his station because his mobile transceiver cannot be plugged directly into his home-station transmitter power supply? This supply is designed to fill all these needs.

Many of today's commercially available ac supplies are not equipped for 220-volt operation. If the station includes a two-kilowatt amplifier, a separate 220-volt line should be available in the shack. Blinking house lights are not always a result of running a high-powered amplifier. It could be caused by the intermittent 400- or 500-watt load presented by an exciter power supply to the 117-volt source. Connecting the exciter supply to a 220-volt outlet (providing a dual-primary transformer is used) can be helpful in this regard.

Circuit Details

The supply is shown in Figs. 1 through 3. Primary power may be applied to the supply in two ways. First, terminals 6 and 8 of J3 may be shorted together; this is normally the function of the station transmitting equipment on-off switch (see Fig. 2). On the other hand, S1 may be actuated when the supply is used independently. Transient voltages on the ac line are eliminated by Thyrector assemblies VR1 and VR2.

Full-wave rectification is employed in the secondary circuit of each power transformer to develop the three dc operating voltages. Choke-input filtering provides adequate regulation of both the 300- and 800-volt outputs. Both L1 and L2 are shunted with suitable resistors to reduce the possibility of diode damage when primary power to the supply is removed.

The bias voltage is adjustable and may be set to any value between −40 and −80. Should a range between −80 and −130 volts be required, R1 may be interchanged with R3. Likewise, if a range from 0 to −40 volts is needed, R1 may be swapped with R2.

Metering

A six-position switch and a 0-1-mA meter allows monitoring of high and low voltages, the current for each of these, and the bias voltage. The sixth position permits the meter to be disabled. The meter shunts for both current positions of S2 are homemade and provide a full-scale reading of 500 mA on each range. The proper resistance for the shunts is determined by dividing the meter internal resistance (approximately 100 ohms in this case) by 500, and is equal to 0.2 ohm. No. 30 enameled copper wire has a resistance of 105 ohms per 1000 feet, or 0.105 ohm per foot. Extending the division another step, one inch of wire has a resistance of .008 ohm. Approximately 23 inches of wire provided the correct value for the shunts. Each 23-inch length of wire is wound on a 100,000-ohm, two-watt composition resistor which serves as a form.

Construction

The supply is built on a 10 × 8 × 3-inch aluminum chassis. The spot welds at the four corners are reinforced with No. 6 hardware since the transformers are quite heavy. The total weight of the completed supply is slightly more than 40 pounds. Several one-inch-diameter holes are cut in the chassis bottom plate to allow adequate air circulation.

All of the power-supply output voltages are present on a 12-connection terminal block. The end of the cable used to interconnect the supply to the station transceiver is equipped with a 12-lug fanning strip, providing a convenient means to disconnect it.

One special wiring precaution is necessary; the bleeder resistors for both the high and low-voltage circuits should be mounted in the clear to allow plenty of air circulation around them. Perforated aluminum stock is placed over a 1 × 3-inch cut in the chassis which is directly above the mounting position for the 800-volt bleeder network.

Operation

Two jumper plugs are mounted "back-to-back", making the change from 117-volt operation to 220 volts a simple matter of reversing P1. P2 performs an identical function to select 6 or 12 volts for the filament line.

The cost for this project should be under $100, even if all of the parts are purchased new. The price of the two power transformers and two filter chokes comprises approximately 60 percent of the total cost.[1]

[1] A package including the two power transformers and the two filter chokes is available from Hammond Manufacturing Company, Inc., 1051 Clinton Street, Buffalo, NY 14240, for approximately $60. In Canada, the address is Hammond Mfg. Co., Ltd., 394 Edinburgh Rd., North Guelph, Ontario. Catalog available.

A 3000-VOLT POWER SUPPLY

This high-voltage power supply may be used with linear amplifiers that are capable of operating at maximum legal input power levels. It was designed for use with a one-kilowatt 3-500Z amplifier, but with minor modifications to the control circuitry to suit individual circumstances it can be used with amplifiers having a pair of 3-500Z tubes, a single 3-1000Z, 4-1000A, or any tube or tubes calling for 2500 to 3000 volts at up to 700 mA. Examples of such amplifiers may be found in Chapter 6.

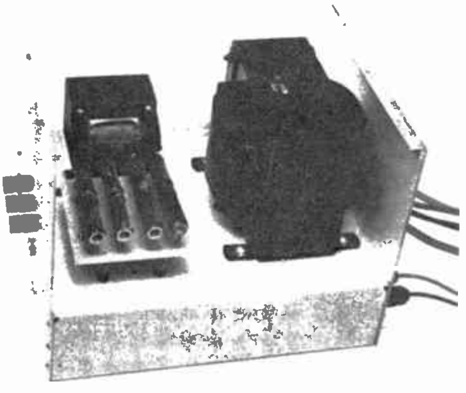


Fig. 1 — Top chassis view of the 3000-volt power supply as constructed by WA1JZC. The circuit board in the foreground holds the bleeder resistors, which are spaced apart and supported a short distance above the board for proper cooling. The large transformer is for the high-voltage supply, and the small transformer provides filament power for the amplifier.

The Circuit

A voltage-doubler circuit connected to the secondary of T1 provides approximately 3000 volts dc. See Fig. 3. The primary of T1 can be operated from either a 117-volt line or a 220-volt source; the latter voltage is preferred. VR1 and VR2 are suppressors included to prevent transients from damaging the high-voltage capacitor bank or the rectifier diodes. Since T1 has two 117-volt primary windings, a suppressor is connected across each. The windings and suppressors are connected in parallel for 117-volt operation, and they are series connected for a 220-volt line.

A relay (K1) is necessary to switch the high-current inrush when the supply is activated. Ordinary toggle switches cannot be used to activate the power supply directly. Surge protection is accomplished by placing R1 in series with one lead of the ac line. K2B shorts out this resistor a few seconds after the main power switch (S1, located on the amplifier front panel) is actuated. A separate line cord for the power supply allows this section to be operated on 220 volts while permitting other circuits in the amplifier to operate on 117 volts. The 120 volts needed to energize the coil of K2 are taken from a half-wave rectified dc supply located on the amplifier chassis. Note that the B-minus terminal is held a few volts above ground by the 15-ohm, 2-watt resistor, for metering purposes in the companion amplifier.

Construction

The power supply is built on a standard 10 × 12 × 3-inch aluminum chassis. Construction is straightforward, as can be seen from Figs. 1 and 2. The front and rear panels are made from 9 × 10-inch pieces of 1/16-inch thick aluminum, and the bottom plate and the U-shaped top cover are made out of perforated aluminum stock.

The primary and control-circuit components, as well as the rectifier board and capacitor bank, are mounted underneath the chassis. Reasonable care must be taken to prevent any part of the primary or control wiring from coming into contact with the high-voltage components. Each of the 100-μF capacitors in the capacitor bank is shunted by a 25,000-ohm, 20-watt wirewound resistor. These resistors equalize the voltage drops across the series-connected capacitors, and also serve as the bleeder resistance. Since these resistors get quite hot during normal operation, they are mounted away from the electrolytic capacitors on a separate circuit board above the chassis, to allow for adequate ventilation. The other large heat-generating components are the power and filament

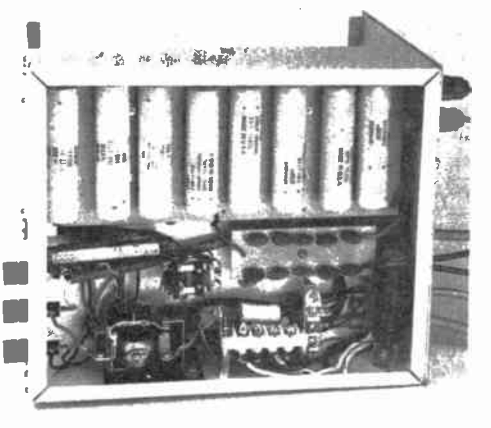


Fig. 2 — The primary and control-circuit components are grouped at the bottom, with the high-voltage capacitor bank and rectifier board occupying the upper portion of this bottom chassis view of the power supply. R1 is visible in the lower right-hand corner.

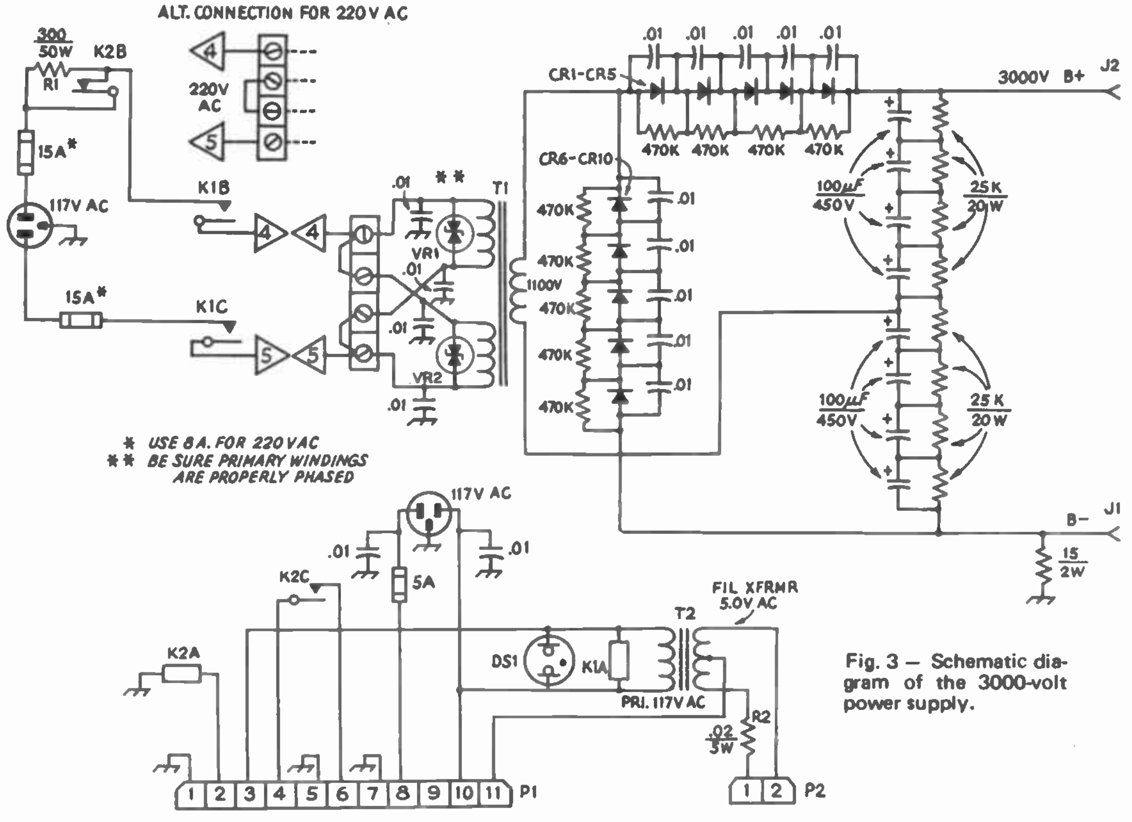


Fig. 3 — Schematic diagram of the 3000-volt power supply.

CR1 — CR10, inc. — 1000-PRV, 2.5-A (Mallory M2.5A or equiv.).
DS1 — 117-volt ac neon pilot lamp assembly.
J1, J2 — High-voltage chassis connector (Millen 37001). K1 — Power relay, dpdt, 117-volt coil (Potter and Brumfeld PR-11AY or equiv.).
K2 — Dpdt 10 A contacts, 120-V dc coil (Potter and Brumfeld KA11DG or equiv.).
P1 — Cable-mounted 11-pin power connector.
P2 — Cable-mounted 2-pin power connector.
R2 — 8 feet No. 14 enam. wire wound on 3-inch long, 3/4-inch dia Plexiglas rod.
T1 — Dual 117-volt primary, 1100-V secondary, 600 VA (Berkshire 6181 or equiv.).
T2 — 117-volt primary; secondary 5.0 volts at 15 A (Stancor P6433 or equiv.).
VR1, VR2 — Transient-voltage suppressor, 120-volt (General Electric 6RS20SP4B4 or equiv.).

transformers (T1 and T2), which are also mounted above chassis.

A small etched circuit board supports CR1 through CR10 and their associated equalizing resistors and transient-suppressing disk capacitors. In actual operation, the filament voltage measured at the amplifier tube socket exceeded the maximum voltage recommended by the tube manufacturer slightly, so R2 was included to reduce the voltage to a suitable value. To avoid excessive voltage drop in the cable connecting T2 with the amplifier, it is recommended that the cable be made of No. 10 wire or larger (in many cases, R2 will not be necessary).

NICKEL-CADMIUM BATTERY CHARGER

Any advantage that a NiCad (nickel-cadmium) battery may have over other types can be lost through improper charging. This information concerning NiCad charging techniques was contributed by WAØUZO. NiCads can even be ruined on the *first* recharging cycle. If connected to a constant-voltage source, initial current may be quite high. Normally, no damage would result unless the battery voltage is low (fully discharged). Using a constant current for battery charging is permissible at the start of the charging cycle, however, as the battery reaches full charge, the voltage may rise to an excessive value.

The correct solution is a combination of the two methods. Any circuit used for charging NiCads should limit both the current and voltage, such as the one described here.

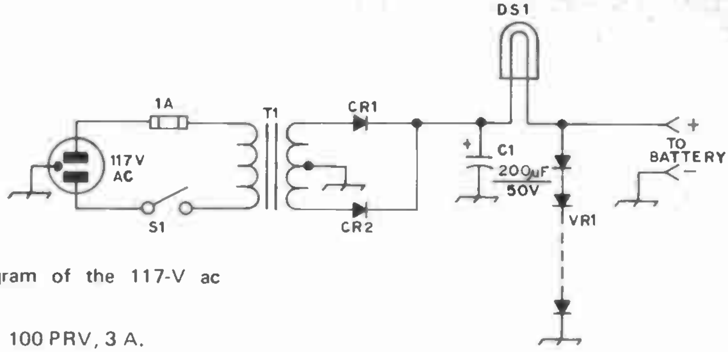


Fig. 1 – Schematic diagram of the 117-V ac charger.
C1 – Electrolytic.
CR1, CR2 – Silicon diodes, 100 PRV, 3 A.
DS1 – See text.
T1 – Primary 117 V ac, secondary 25.6 V at 500 mA. Calectro D1-752 (or equiv.).
VR1 – See text.

Some other precautions which should be observed while charging NiCads are:

1) Battery temperature should be between 40° and 80°F. It should never exceed 100°F.

2) Two or more batteries with the same voltage rating may be charged in parallel, but be sure that the charger has sufficient current capability.

3) Check the manufacturer's data sheet for the maximum allowable charging rate. A typical figure would be ten percent of the ampere-hour rating (a 10-ampere-hour battery would require a current of 1A).

4) Do not attempt to charge two batteries in series with a constant current unless the batteries are of the same type and capacity, and are in the same state of charge (voltage on one may be excessive).

5) To determine the approximate charging time, divide the ampere-hour rating by the charging current used, and multiply the resulting time by 1.25.

Suitable Charging Circuits

Figs. 1 and 2 show two versions of the same basic charging circuit. The circuit shown in Fig. 1 is used with 117 V ac, and the one in Fig. 2 can be used with the car battery. The latter circuit could be connected to the cigarette lighter, and is suitable for battery packs of up to 14 volts.

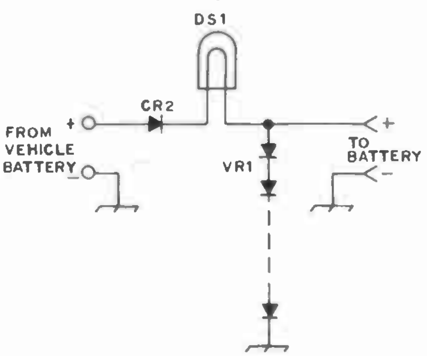


Fig. 2 – Schematic diagram of NiCad battery charger suitable for mobile use. See text for explanation of DS1 and VR1. CR2 protects the components in the event of accidental reversal of input leads. See Fig. 1 for CR2.

The dial lamp (DS1) is used to limit the current. One with a rating of 100 to 150 mA should work fine with most batteries. The voltage rating should be approximately that of the charging source (for example, two 12-V bulbs in series may be necessary if a 26-V supply is used).

The voltage regulator shown in Fig. 3 is based on the fact that a forward-biased diode will not conduct until approximately 0.75 V dc is applied. By adding a suitable number of diodes in series as shown, a voltage regulator for the maximum battery voltage can be built easily. The circuit shown in Fig. 3 can be used in either Fig. 1 or 2, for VR1. It will draw little current until the

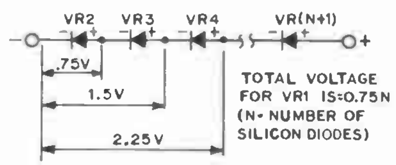


Fig. 3 – Schematic diagram of the voltage regulator (VR1, Figs. 1 and 2).

battery voltage reaches a permissible value during charge. Once the voltage reaches a preset level, the diodes start to conduct and limit any further increases.

Initial Testing

After the circuit is wired and checked, apply power (without a battery connected for charging). The bulb should light to less than full brilliance. Measure the voltage across the regulator. It should be 3 to 8 percent above the rated voltage of the batteries to be charged. Adding or removing some diodes in VR1 may be necessary. Connect the discharged batteries and measure the charging current (either a built-in meter could be used, or a temporary one could be connected in series with the battery). The current should be typically 100 mA with partially discharged batteries. The current will decrease as the charging time increases, and a value of 5 mA indicates a fully charged condition. No damage will result if the batteries are left on charge continuously.

HF Transmitting

Regardless of the transmission mode – code, a-m, fm, single sideband, radioteletype, amateur TV – vacuum tubes and semiconductors are common elements in all transmitters. They are used as oscillators, amplifiers, frequency multipliers and frequency converters. These four building blocks, plus suitable power supplies, are basically all that is required to make any of the popular transmission systems.

The simplest code transmitter is a keyed oscillator working directly into the antenna; a more elaborate (and practical) code transmitter, the type popular with many beginners, will include one or more frequency-multiplication stages and one or more power-amplifier stages. Any code transmitter will obviously require a means for keying it. The bare skeleton is shown in Figs. 6-2A and B. The rf generating and amplifying sections of a double-sideband phone transmitter (a-m or fm) are similar to those of a code transmitter.

The overall design depends primarily upon the bands in which operation is desired and the power output. A simple oscillator with satisfactory frequency stability may be used as a transmitter at the lower frequencies, but the power output obtainable is small. As a general rule, the output of the oscillator is fed into one or more amplifiers to bring the power fed to the antenna up to the desired level.

An amplifier whose output frequency is the same as the input frequency is called a straight amplifier. A buffer amplifier is the term sometimes applied to an amplifier stage to indicate that its primary purpose is one of isolation, rather than power gain.

Because it becomes increasingly difficult to maintain oscillator frequency stability as the frequency is increased, it is most usual practice in working at the higher frequencies to operate the oscillator at a low frequency and follow it with one or more frequency multipliers as required to arrive at the desired output frequency. A frequency multiplier is an amplifier that delivers output at a multiple of the exciting frequency. A doubler is a multiplier that gives output at twice the exciting frequency; a tripler multiplies the exciting frequency by three, etc. From the viewpoint of any particular stage in a transmitter, the preceding stage is its driver.

As a general rule, frequency multipliers should not be used to feed the antenna system directly, but should feed a straight amplifier which, in turn, feeds the antenna system.

Good frequency stability is most easily obtained through the use of a crystal-controlled oscillator, although a different crystal is needed for each frequency desired (or multiples of that frequency). A self-controlled oscillator or VFO (variable-frequency oscillator) may be tuned to any frequency with a dial in the manner of a receiver, but requires great care in design and construction if its stability is to compare with that of a crystal oscillator.

Many transmitters use tubes, but for low-power hf and channelized vhf fm transmitters, transistors are dominant. New solid-state devices are being developed which allow dc inputs of 100 watts or more with a low-level of IM distortion products. As the cost of these transistors is reduced it can be assumed that at some point in the future tubes will be used only for high-power amplification.

The best stage or stages to key in a code transmitter is a matter which is discussed in a later chapter. The oscillator/multiplier/amplifier type of transmitter (Fig. 6-2B) has long been popular. However, the excellent frequency stability and the advantages of grid-block keying (which are explained in the Code Transmission chapter) have

Fig. 6-1 – An amateur's transmitter is his on-the-air voice. He is judged by the quality of that "voice," whatever the mode that he chooses to operate.

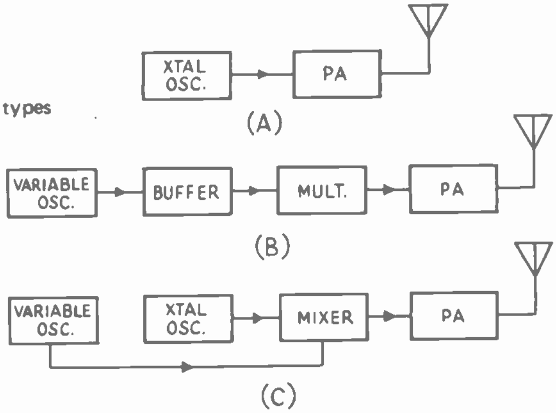


Fig. 6-2 — Block diagrams of the three basic types of transmitters.

made the heterodyne exciter of Fig. 6-2C increasingly popular, in spite of the slightly more complex circuitry required.

An fm transmitter can only be modulated in or following the oscillator stage. An a-m phone transmitter can only be modulated in the output stage, unless the modulated stage is followed by a linear amplifier. However, following an amplitude-modulated stage by a linear amplifier is an inefficient process, convenient as an expedient, but not recommended for best efficiency.

Following the generation of a single-sideband phone signal, its frequency can be changed only by frequency conversion (not multiplication), in exactly the same manner that signals in a receiver are heterodyned to a different frequency. Complete details of ssb transmitter design and construction are given in Chapter 13.

CRYSTAL OSCILLATORS

The frequency of a crystal-controlled oscillator is held constant to a high degree of accuracy by the use of a quartz crystal. The frequency depends almost entirely on the dimensions of the crystal (essentially its thickness); other circuit values have comparatively negligible effect. However, the power obtainable is limited by the heat the crystal will stand without fracturing. The amount of heating is dependent upon the rf cyrstal current which, in turn, is a function of the amount of feedback required to provide proper excitation. Crystal heating short of the danger point results in frequency drift to an extent depending upon the way the crystal is cut. Excitation should always be adjusted to the minimum necessary for proper operation.

The most stable type of crystal oscillator is that which provides only a small voltage output (lightly loaded), and which operates the crystal at a low drive level. Such oscillators are widely used in receivers and heterodyne transmitters. The oscillator/multiplier/amplifier type of transmitter usually requires some power from the oscillator stage. For either type of crystal oscillator, the active element may be a tube or a transistor.

Oscillator Circuits

The simplest crystal-oscillator circuit is shown in Fig. 6-3A. Feedback in this circuit is provided by the gate-source and drain-source capacitance. The circuit shown at B is the equivalent of the tuned-grid, tuned-plate circuit discussed in the chapter on vacuum-tube principles, using the crystal to replace the tuned grid circuit. Although JFETs are shown in the sample circuits at A and B, MOSFETs or triodes may also be employed, using the connections shown in 6-3C through F.

For applications where some power is required from the crystal oscillator, the circuits shown in

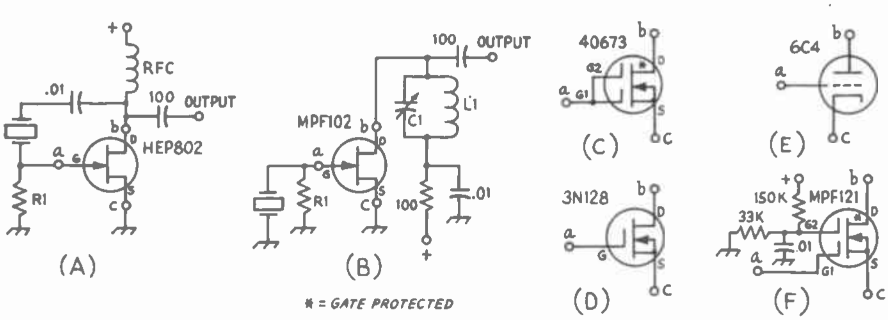


Fig. 6-3 — Simple crystal oscillator circuits. (A) Pierce, (B) FET, (C-F) other devices that can also be used in the circuits of A and B with appropriate changes in supply voltage.

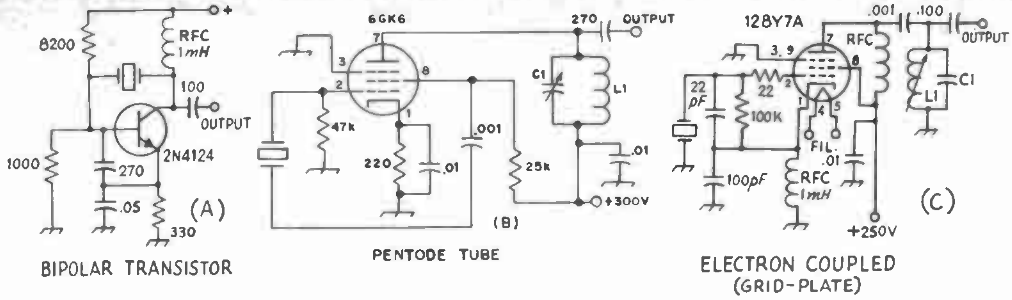


Fig. 6-4 – Crystal-oscillator circuits that are designed to deliver power. L1/C1 resonate at the crystal frequency, or a multiple thereof if the second, third, or fourth harmonic is the desired output frequency.

Fig. 6-4 may be employed. At A, a bipolar transistor is used, while the tube circuits (B, C) are somewhat more complicated. They combine the functions of oscillator and amplifier or frequency multiplier in a single tube. In these circuits, the screen of a tetrode or pentode is used as the plate in a triode oscillator. Power output is taken from a separate tuned tank circuit in the actual plate circuit. Although the oscillator itself is not entirely independent of adjustments made in the plate tank circuit when the latter is tuned near the fundamental frequency of the crystal, the effects can be satisfactorily minimized by proper choice of the oscillator tube.

The oscillators of Fig. 6-4B and 6-4C are a modification of the grid-plate circuit of Fig. 6-3B. In Fig. 6-4C the ground point has been moved from the cathode to the plate of the oscillator (in other words, to the screen of the tube). Excitation is adjusted by proper proportioning of 22- and 100-pF feedback capacitors.

When some types of tubes are used in the circuits of Fig. 6-4B, oscillation will stop when the output plate circuit is tuned to the crystal frequency, and it is necessary to operate with the plate tank circuit critically detuned for maximum output with stability. However, when the 6GK6, 12BY7A, 5763, or the lower-power 6AH6 is used with proper adjustment of excitation, it is possible to tune to the crystal frequency without stopping oscillation. These tubes also operate with less crystal current than most other types for a given power output, and less frequency change occurs when the plate circuit is tuned through the cyrstal frequency (less than 25 Hertz at 3.5 MHz).

Crystal current may be estimated by observing relative brilliance of a 60-mA dial lamp connected in series with the crystal. Current should be held to the minimum for satisfactory output by careful adjustment of excitation. With the operating voltages shown, satisfactory output should be obtained with crystal currents of 40 mA or less.

In these tube circuits, output may be obtained at multiples of the crystal frequency by tuning the plate tank circuit to the desired harmonic, the output dropping off, of course, at the higher harmonics. Especially for harmonic operation, a low-*C* plate tank circuit is desirable.

Practical Considerations

The operation of a crystal oscillator is often hampered because vhf parasitic oscillations also occur in the circuit. An effective way of killing parasitics is the use of a low-value composition resistor or ferrite bead, as shown in Fig. 6-5. The parasitic stopper can be located on the gate (grid or base) lead, and it should be placed as close as possible to the transistor. The circuit at A may be used for low-power applications. If a crystal above 1 MHz is to be used it may be advisable to include a trimmer capacitor across the crystal to allow the crystal frequency to be set exactly.

It is often desirable in fm and ssb gear to use several crystals, switch-selected in a single oscilla-

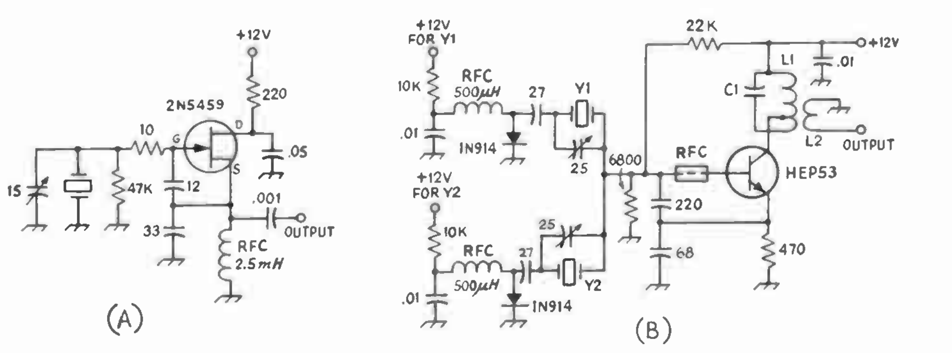


Fig. 6-5 – Two practical crystal-oscillator designs. (A) For low-power output applications such as a conversion oscillator or BFO, (B) an example of diode switching of crystals. The rf choke on the base lead of the transistor is a ferrite bead which prevents vhf parasitic oscillation.

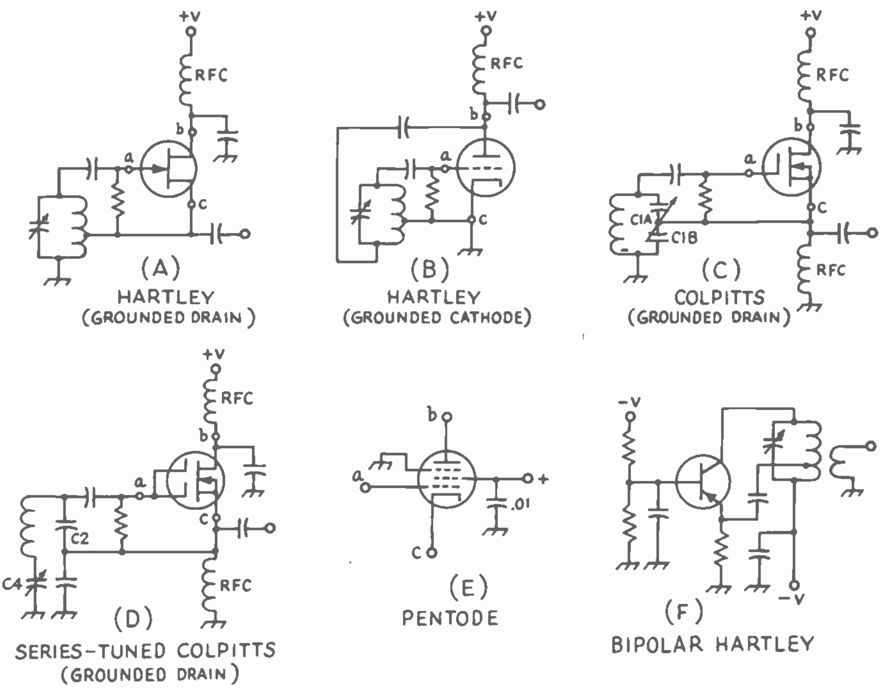


Fig. 6-6 — VFO circuits. The devices shown in Fig. 6-2C through F may also be employed as the active component.

tor. If manual switching is used, the leads to the switch may introduce sufficient additional capacitance to upset the operation of the circuit. Therefore, the use of diode switching, such as shown in Fig. 6-5B, is now popular. Any high-speed switching diode may be employed. The use of diode switching for low-level tank circuits, especially in receivers, has gained wide acceptance. A special diode known as the PIN has been developed for this purpose. In any diode-switching circuit it is important to insure that the switching bias is many times larger than the peak rf voltage present.

VARIABLE-FREQUENCY OSCILLATORS

The frequency of a VFO depends entirely on the values of inductance and capacitance in the circuit. Therefore, it is necessary to take careful steps to minimize changes in these values not under the control of the operator. As examples, even the minute changes of dimensions with temperature, particularly those of the coil, may result in a slow but noticeable change in frequency called drift. The effective input capacitance of the oscillator tube, which must be connected across the circuit, changes with variations in electrode voltages. This, in turn, causes a change in the frequency of the oscillator. To make use of the power from the oscillator, a load, usually in the form of an amplifier, must be coupled to the oscillator, and variations in the load may reflect on the frequency. Very slight mechanical movement of the components may result in a shift in frequency, and vibration can cause modulation.

In the past different techniques have been used to design the VFOs for transmitters and receivers. However, today the same circuits may be used for either application. In receivers the VFO is usually called an HFO.

VFO Circuits

Fig. 6-6 shows the most commonly used circuits. They are all designed to minimize the effects mentioned above. The oscillating circuits in Figs. 6-6A and B are the Hartley type; those in C and D are Colpitts circuits. (See chapter on vacuum-tube principles.) In the circuits of A, B and C, all of the above-mentioned effects, except changes in inductance, are minimized by the use of a high-*Q* tank circuit obtained through the use of large tank capacitances. Any uncontrolled changes in capacitance thus become a very small percentage of the total circuit capacitance.

In the series-tuned Colpitts circuit of Fig. 6-6D (sometimes called the Clapp circuit), a high-*Q* circuit is obtained in a different manner. The tube is tapped across only a small portion of the oscillating tank circuit, resulting in very loose coupling between tube and circuit. The taps are provided by a series of three capacitors across the coil. In addition, the tube capacitances are shunted by large capacitors, so the effects of the tube – changes in electrode voltages and loading – are still

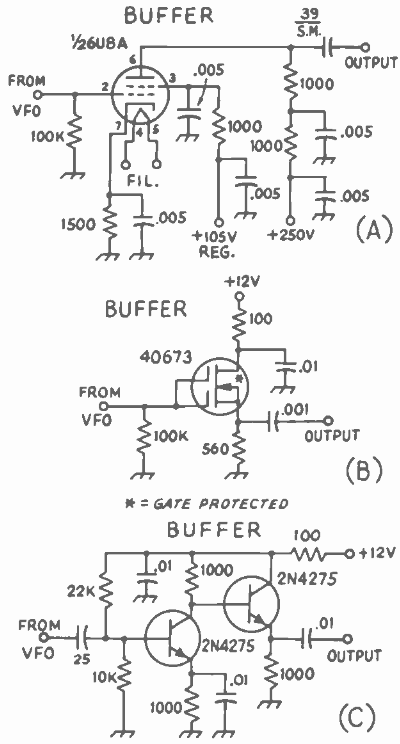


Fig. 6-7 — Isolating stages to be used between a VFO and the following amplifier or mixer stage.

further reduced. In contrast to the preceding circuits, the resulting tank circuit has a high L/C ratio and therefore the tank current is much lower than in the circuits using high-C tanks. As a result, it will usually be found that, other things being equal, drift will be less with the low-C circuit.

For best stability, the ratio of C2 to C4 should be as high as possible without stopping oscillation. The permissible ratio will be higher the higher the Q of the coil and the mutual conductance of the tube. If the circuit does not oscillate over the desired range, a coil of higher Q must be used or the capacitance of C2 and C3 reduced.

The pentode tube of 6-6E or any of the active devices shown in Fig. 6-3 may be used in either the Hartley or Colpitts circuits. Good results can be obtained with both tubes and transistors, so the choice of the active device is often a matter of personal preference.

Load Isolation

In spite of the precautions already discussed, the tuning of later stages in the transmitter may cause a noticeable change in frequency. This effect can be reduced considerably by designing a pentode oscillator for half the desired frequency and doubling frequency in the output circuit.

It is desirable, although not a strict necessity if detuning is recognized and taken into account, to approach as closely as possible the condition where the adjustment of tuning controls in the transmitter, beyond the VFO frequency control, will have negligible effect on the frequency. This can be done by adding isolating stage or stages whose tuning is fixed between the oscillator and the first tunable amplifier stage in the transmitter.

Fig. 6-7A shows such an arrangement that gives good isolation. A pentode tube is operated with a low-impedance resistive load, and regulated screen voltage. At B a simple follower circuit is used. The disadvantage of this circuit is that the level of the output will be quite low, usually less than one volt. Bipolar transistors are used in a direct-coupled follower arrangement in Fig. 6-7C, providing a higher level of output (above 3 V) than was possible with the design shown at B. The ability of a buffer stage to isolate the VFO from the load can be tested simply. Use a receiver to monitor the VFO, and listen as the buffer output is first left open and then shorted. A good buffer will hold the frequency change to less than 100 Hz. Often the frequency change may be in the order of several kHz when this test is made, an indication that the buffer is not doing its job.

Chirp, Pulling and Drift

Any oscillator will change frequency with an extreme change in plate screen voltages, and the use of stabilized sources for both is good practice. But steady source voltages cannot alter the fact of the extreme voltage changes that take place when an oscillator is keyed or heavily amplitude-modulated. Consequently some chirp or fm is the inescapable result of oscillator keying or heavy amplitude modulation.

A keyed or amplitude-modulated amplifier presents a variable load to the driving stage. If the driving stage is an oscillator, the keyed or modulated stage (the variable load) may "pull" the oscillator frequency during keying or modulation. This may cause a "chirp" on cw or incidental fm on a-m phone. In either case the cure is to provide one or more "buffer" or isolating stages between the oscillator stage and the varying load. If this is not done, the keying or modulation may be little better than when the oscillator itself is keyed or modulated.

Frequency drift is minimized by limiting the temperature excursions of the frequency-determining components to a minimum. This calls for good ventilation and a minimum of heat-generating components.

Variable capacitors should have ceramic insulation, good bearing contacts and should preferably be of the double bearing type. Fixed capacitors should have zero-temperature coefficients. The tube socket should have ceramic insulation.

Temperature Compensation

If, despite the observance of good oscillator construction practice, the warm-up drift of an oscillator is too high, it is caused by high-temperature operation of the oscillator. If the ventilation cannot be improved (to reduce the ultimate temperature), the frequency drift of the oscillator can be reduced by the addition of a "temperature-coefficient capacitor." These are available in negative and positive coefficients, in contrast to the zero-coefficient "NPO" types.

Oscillator Coils and Vibration

The Q of inductors either in the VFO circuitry proper or in a succeeding stage should be as high as possible. The coil should be well spaced from shields and other conducting surfaces in order to reduce undesirable eddy-current effects. A heavy-gauge solid wire is recommended for interconnecting leads and lengths should be kept short.

While heating generally produces long-term drift, this is usually not as serious as sudden frequency changes caused by vibration. All components should be securely fastened to the VFO chassis and circuit board with the entire assembly shock mounted on rubber grommet cushions. Care should also be taken concerning the method of coupling the VFO capacitor to the VFO dial. Vibration of the panel on which the dial is mounted should not be transferred to the capacitor shaft.

Filtering

The output of oscillators, mixers and similar stages usually has harmonic and other spurious-frequency energy along with some desired signal. Depending upon the application, such components may result in undesirable effects and require filtering. The circuits shown in Fig. 6-8 can be employed where either a high-pass or low-pass filter characteristic is sufficient. (In some instances, a band-pass filter may be required.)

These filters are based on a Chebyshev design and component values are given in Table 1. The filters are "normalized" to a frequency of 1 MHz and an input and output impedance of 52 ohms. In order to translate the designs to other frequencies, all that is necessary is to divide the component values by the new frequency in MHz. (The 1-MHz value represents a "cutoff" frequency. That is, the attenuation increases rapidly above this frequency in the low-pass case or below f_c in the high-pass application. This effect should not be confused with the variations in attenuation in the passband.) For instance, if it was desired to eliminate harmonics from a VFO at frequencies above 5 MHz, the inductance and capacitance values would be divided by 5.0.

Other impedance levels can also be used by multiplying the inductors by the ratio $Z_o/52$ and the capacitors by $52/Z_o$ where Z_o is the new impedance. This factor should be applied in addition to the ones for frequency translation. Examination of the component values in Table 1 indicates changing the impedance level to a somewhat higher one is advisable if practical. This would avoid having very small inductance values and high capacitance values which might be unhandy in constructing a filter.

The choice of filter model depends mostly on the power level involved. In low-level stages, the effects of variation in insertion loss in the passband usually can be neglected in amateur applications. Consequently, the filter with a 3-dB ripple can be used. However, VSWR is closely related to insertion loss and the 0.1-dB filter should be selected in power-amplifier stages where VSWR could cause harmful effects. The 0.1-dB filter should have a maximum VSWR of approximately 1.4:1. Unfortunately, the 0.1-dB design doesn't have the rapid rolloff that the 3-dB model possesses as can be seen by the tabulated data in Fig. 6-8C. While other filters are possible with different ripple factors, the ones shown in Fig. 6-8 should cover a wide variety of applications.

TABLE 1

Component values for filters shown in Fig. 6-8. Inductances are in microhenries and capacitances are in picofarads.

Fig. 6-8A (LP)	L_1	C_2	L_3	C_4	L_5
Fig. 6-8B (HP)	C_1	L_2	C_3	L_4	C_5
0.1-dB LP	9.49	4197	16.4	4197	9.49
3-dB LP	28.8	2332	37.6	2332	28.6
0.1-dB HP	2669	6.04	1550	6.04	2669
3-dB HP	879	10.9	675	10.9	879

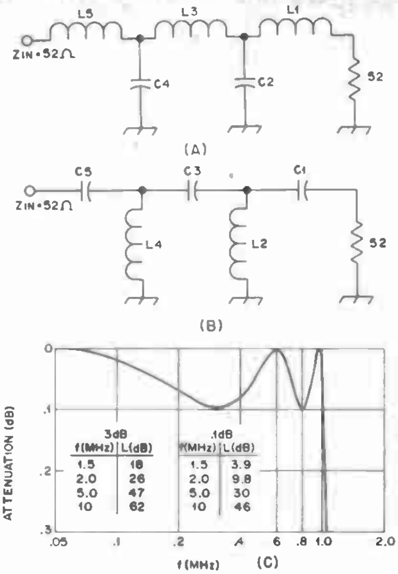


Fig. 6-8 — Low-pass and high-pass Chebyshev filters normalized to a frequency of 1 MHz and input/output impedance terminations of 52 ohms. Component values for 0.1- and 3-dB ripple-factor filters are given in Table 1. Passband frequency response for the 0.1-dB low-pass model is shown in Fig. 6-8C. Curve is similar in shape for the 3-dB model except for scale factor. Response for the high-pass designs can be determined by substituting 1/*f* (MHz) for *f* in Fig. 6-8C.

A PRACTICAL VFO CIRCUIT

The circuit shown in Fig. 6-9 is for a solid-state VFO covering 3.5 to 4 MHz. A number of measures have been taken to prevent harmonic and spurious outputs that so often plague transistor designs. Examination of Fig. 6-9 will show that a diode, CR2 is connected between the signal gate

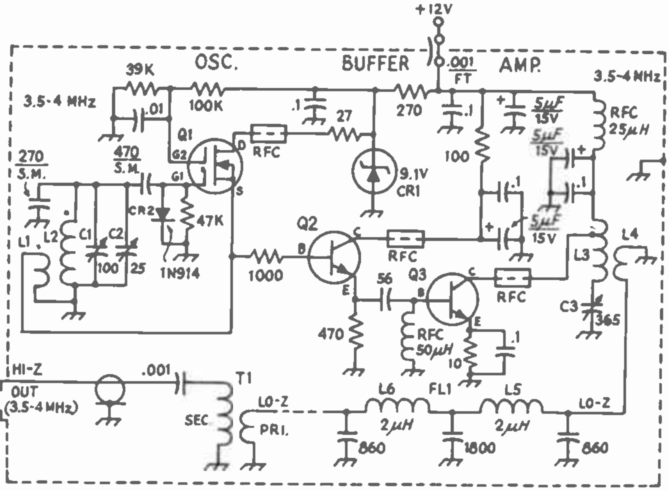


Fig. 6-9 — A typical VFO design showing extensive use of buffering and filtering to achieve a highly stable output with low spurious-frequency content.

of Q1 and ground. This diode should be designed for high-speed switching – a 1N914 is suitable – and should be connected with its anode toward gate 1. It clamps on the positive-going half of the cycle to prevent Q1 from reaching high peak transconductance, the time period when the output from the oscillator is rich in harmonic energy. This technique should be applied to any JFET or MOSFET oscillator, but does not work with bipolar-transistor oscillators. CR2 does not impair the performance of the VFO. Additional harmonics can be generated at Q2 and Q3, so attention must be given to that part of the circuit as well. Note that the collector of Q3 is tapped well down on L3. The tap provides an impedance match for the circuit, but still represents a high impedance at the harmonic frequencies, if not located too near the cold end of L3, thus contributing to a cleaner output signal. However, even though these precautions are taken, it is not uncommon to find that the second and third harmonics from a transistor output stage are only down some 10 to 15 decibels in level from the fundamental signal. By taking the VFO output at low impedance, L4, a low-pass, double-section filter can be used to diminish the harmonic to a level that is some 30 decibels or more below that of the desired output signal. FL1 is designed for 3.5 to 4-MHz use, and assures a clean output signal from the VFO.

VFO Output Level and Impedance

One of the things that perplexes many first-time users of transistorized VFOs is the matter of sufficient signal output to properly excite a transmitter input stage, or to supply adequate injection voltage to a receiver or transmitter mixer. The rms output of a solid-state VFO is limited by its low-impedance output port. In the circuits of Fig. 6-9 the output would usually be taken across the emitter resistor of Q2, the buffer. Typically, the rms output voltage at that point in the circuit will be on the order of 0.5 to 2 volts. Tube mixers can require up to several volts of oscillator signal in order to function properly. Most solid-state transmitters need from 3 to 10 volts of drive on the base of the first power stage, and a reasonable amount of driving power is needed to satisfy this requirement. Driving power is generally required by the grid of the first stage of a tube transmitter. The VFO should, therefore, be capable of supplying from 0.5 to 1 watt of power output. The Class-C amplifier, Q3, provides the needed power output. Should the driven stage present a low-impedance to the VFO, output can be taken directly from the side of FL1 opposite Q3. If, however, the driven stage of the transmitter or receiver has a high input impedance, some method must be used to provide the required impedance transformation, low to high. A broad-band toroidal step-up transformer, T1, is used for this purpose in Fig. 6-9. The secondary of the transformer is resonant somewhere in the operating range of the VFO, and takes advantage of the stray circuit capacitance, normally around 10 pF, to establish resonance. The impedance-transformation ratio is set by adjusting the number of turns on the primary winding. Alternatively, T1 can be replaced by a tuned circuit of conventional design. It can be equipped with a fixed-value capacitor and a slug-tuned inductor, or a fixed-value inductor can be used with a variable capacitor to permit peaking the output at the operating frequency. The use of a tuned circuit will assure somewhat better efficiency than will the broadband transformer, T1. Thus, it can be seen that the circuit must be tailored to the need.

Checking VFO Stability

A VFO should be checked thoroughly before it is placed in regular operation on the air. Since succeeding amplifier stages may affect the signal characteristics, final tests should be made with the complete transmitter in operation. Almost any VFO will show signals of good quality and stability when it is running free and not connected to a load. A well-isolated monitor is a necessity. Perhaps the most convenient, as well as one of the most satisfactory, well-shielded monitoring arrangements is a receiver combined with a harmonic from a frequency standard. (See the Measurements chapter for suitable circuits.) The receiver BFO is turned off and the VFO signal is tuned to beat with the signal from the crystal oscillator instead. In this way any receiver instability caused by overloading the input circuits, which may result in "pulling" of the hf oscillator in the receiver, or by a change in line voltage to the receiver when the transmitter is keyed, will not affect the reliability of the check. Most crystals have a sufficiently low temperature coefficient to give a check on drift as well as on chirp and signal quality if they are not overloaded.

Harmonics of the crystal may be used to beat with the transmitter signal when monitoring at the higher frequencies. Since any chirp at the lower frequencies will be magnified at the higher frequencies, accurate checking can best be done by monitoring at a harmonic.

The distance between the crystal oscillator and receiver should be adjusted to give a good beat between the crystal oscillator and the transmitter signal. When using harmonics of the crystal oscillator, it may be necessary to attach a piece of wire to the oscillator as an antenna to give sufficient signal in the receiver. Checks may show that the stability is sufficiently good to permit oscillator keying at the lower frequencies, where break-in operation is of greater value, but that chirp becomes objectionable at the higher frequencies. If further improvement does not seem possible, it would be logical in this case to use oscillator keying at the lower frequencies and amplifier keying at the higher frequencies.

Premixing

It is difficult to build a variable-frequency oscillator for operation above 10 MHz with drift of only a few Hz. A scheme called premixing, shown in Fig. 6-10A, may be used to obtain VFO output in the 10- to 50-MHz range. The output of a highly stable VFO is mixed with energy from a crystal-controlled oscillator. The frequencies of the two oscillators are chosen so that spurious outputs generated during the mixing process do not fall within the desired output range. A bandpass filter at the mixer output attenuates any out-of-band spurious energy. The charts given in Chapter 8 can be used to choose oscillator combinations which will have a minimum of spurious outputs. Also, Chapter 8 contains a discussion of mixer-circuit design.

PLL

Receivers and transmitters of advanced design are now using phase-locked loops (PLLs) to generate highly stable local oscillator energy up into the microwave region. The PLL has the advantage that no mixing stage is used in conjunction with the output oscillator, so the output energy is quite "clean." The Galaxy R-530, the Collins 651S-1, and the National HRO-600 currently use PLL high-frequency oscillator systems.

The basic diagram of a PLL is shown in Fig. 6-10B. Output from a voltage-controlled oscillator

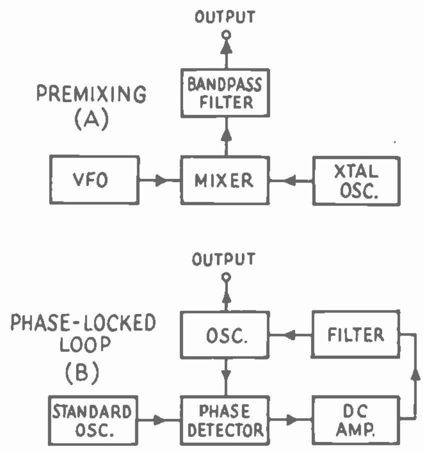


Fig. 6-10 – Block diagrams of the (A) premixing and (B) phase-lock-loop schemes.

(VCO) and a frequency standard are fed to a phase detector which produces an output voltage equal to the difference in frequency between the two signals. The *error* voltage is amplified, filtered, and applied to the VCO. The error voltage changes the frequency of the VCO until it is locked to the standard. The bandwidth of the error-voltage filter determines the frequency range over which the system will remain in phase lock.

Three types of phase-locked loops are now in use. The simplest type uses harmonics of a crystal standard to phase-lock an HFO, providing the injection for the first mixer in a double-conversion receiver. A typical circuit is given in Fig. 6-11. Complete construction details on this PLL were given in *QST* for January, 1972. A second type of phase-locked loop uses a stable mf VFO as the standard which stabilizes the frequency of an hf or vhf VCO. This approach is used in the receiver described by Fischer in *QST*, March, 1970.

The other PLL system also uses a crystal-controlled standard, but with programmable frequency dividers included so that the VCO output is always locked to a crystal reference. The frequency is changed by modifying the instructions to the dividers; steps of 100 Hz are usually employed for hf receivers while 10-kHz increments are popular in vhf gear. The use of a PLL for fm demodulation is covered separately in Chapter 14.

VFO DIALS

One of the tasks facing an amateur builder is the difficulty of finding a suitable dial and drive assembly for a VFO. A dial should provide a sufficiently slow rate of tuning – 10- to 25-kHz per knob revolution is considered optimum – without backlash. Planetary drives are popular because of their low cost; however, they often develop objectional backlash after a short period of use. Several types of two-speed drives are available. They are well suited to homemade amateur

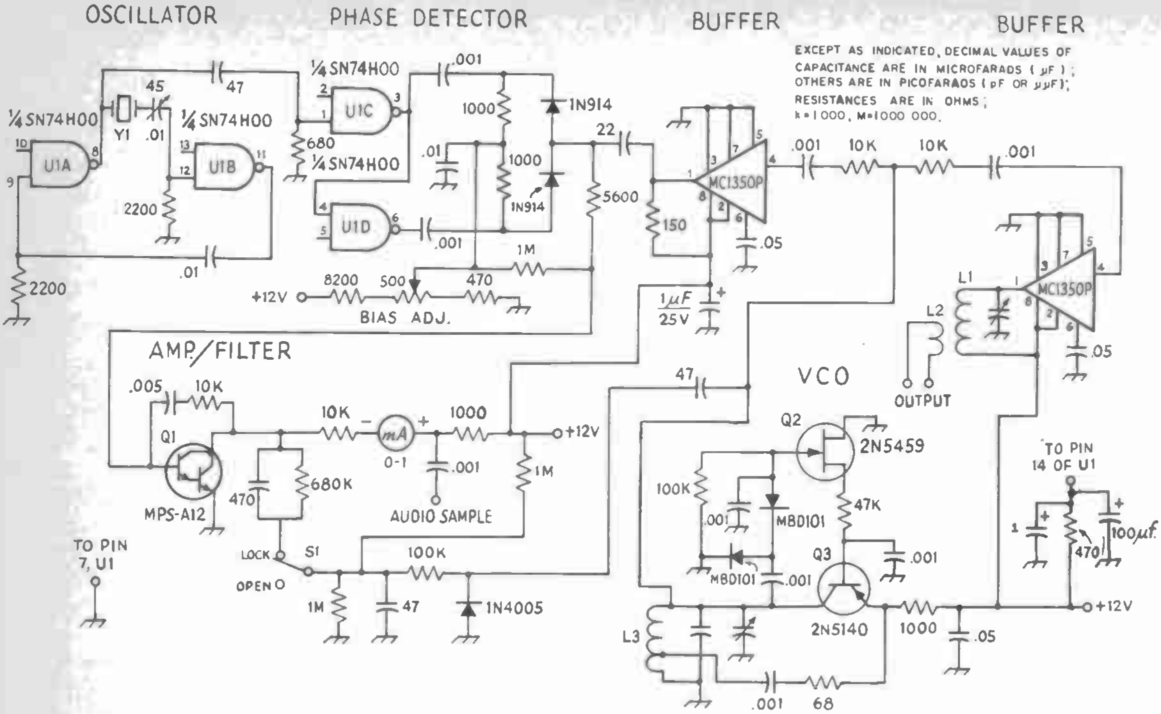


Fig. 6-11 — A practical phase-locked oscillator intended for application as the crystal-controlled HFO in a transmitter or receiver. The crystal frequency should be chosen so that the harmonic content of the standard is sufficient at the desired output frequency. A 200-kHz crystal is good to 40 MHz, 500-kHz crystal to 60 MHz, and 1-MHz crystal to 80 MHz. L1 and L3 are chosen to resonate at the desired output frequency.

equipment. Several of the construction projects described elsewhere in this book employ this type of dial. The Eddystone 898 precision dial has long been a favorite with amateurs, although the need to elevate the VFO far above the chassis introduces some mechanical-stability problems. If a permeability tuned oscillator (PTO) is used, one of the many types of turn counters made for vacuum variable capacitors or rotary inductors may be employed.

Linear Readout

If linear-frequency readout is desired on the dial, the variable capacitor must be only a small portion of the total capacitance in the oscillator tank. Capacitors tend to be very nonlinear near the ends of rotation. A gear drive providing a 1.5:1 reduction should be employed so that only the center of the capacitor range is used. Then, as a final adjustment, the plates of the capacitor must be filed until linear readout is achieved. In a PTO, the pitch of the oscillator coil winding may be varied so that linear frequency change results from the travel of the tuning slug. Such a VFO was described in *QST* for July, 1964. A different approach was employed by Lee (*QST*, November, 1970), using a variable-capacitance diode (Varicap) as the VFO tuning element. A meter which reads the voltage applied to the Varicap was calibrated to indicate the VFO frequency.

Fig. 6-12 — A 5-digit readout using light-emitting diodes.

Electronic Dials

An electronic dial consists of a simplified frequency counter which reads either the VFO or operating frequency of a transmitter or receiver. The advantage of an electronic dial is the excellent accuracy (to one Hertz, if desired) and the fact that VFO tuning does not have to be linear. The readout section of the dial may use neon-glow tubes called Nixies (a trade name of the Burroughs Corp.), or a seven-segment display using incandescent lamps, filament wires in a vacuum tube, or LEDs (light-emitting diodes). A typical LED display is shown in Fig. 6-12. The use of MSI and LSI circuits, some containing as many as 200 transistors on a single chip, reduces the size required for an electronic dial to a few square inches of circuit-board space.

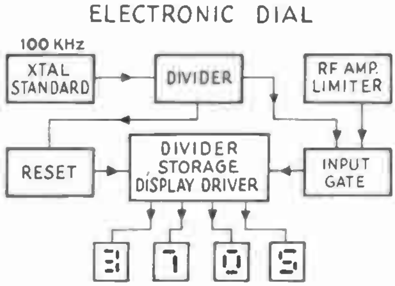


Fig. 6-13 – Block diagram of a frequency counter.

A typical counter circuit is given in Fig. 6-13. The accuracy of the counter is determined by a crystal standard which is often referred to as a clock. The output from a 100-kHz calibration oscillator, the type often used in receivers and transceivers, may be employed if accuracy of 100 Hz is sufficient. For readout down to 1 Hz, a 1- to 10-MHz AT-cut crystal should be chosen, because this type of high-accuracy crystal exhibits the best temperature stability. The clock output energy is divided in decade-counter ICs to provide the pulse which opens the input gate of the counter for a preset time. The number of rf cycles which pass through the gate while it is open are counted and stored. Storage is used so that the readout does not blink. At the end of each counting cycle the information that has been stored activates the display LEDs, which present the numbers counted until another count cycle is complete. A complete electronic dial arranged to be combined with an existing transmitter or receiver was described in *QST* for October 1970. Also, Macleish *et al* reported an adapter which allows a commercially made frequency counter to be mated with ham gear so that the counter performs as an electronic dial (*QST*, May, 1971).

FREQUENCY MULTIPLIERS

Single-Tube Multiplier

Output at a multiple of the frequency at which it is being driven may be obtained from an amplifier stage if the output circuit is tuned to a harmonic of the exciting frequency instead of to the fundamental. Thus, when the frequency at the grid is 3.5 MHz, output at 7 MHz, 10.5 MHz, 14 MHz, etc., may be obtained by tuning the plate tank circuit to one of these frequencies. The circuit otherwise remains the same as that for a straight amplifier, although some of the values and operating conditions may require change for maximum multiplier efficiency.

A practical limit to efficiency and output within normal tube ratings is reached when the multiplier is operated at maximum permissible plate voltage and maximum permissible grid current. The plate current should be reduced as necessary to limit the dissipation to the rated value by increasing the bias and decreasing the loading.

Multiplications of four or five sometimes are used to reach the bands above 28 MHz from a lower-frequency crystal, but in the majority of lower-frequency transmitters, multiplication in a single stage is limited to a factor of two or three. Screen-grid tubes make the best multipliers because their high power-sensitivity makes them easier to drive properly than triodes.

Since the input and output circuits are not tuned close to the same frequency, neutralization usually will not be required. Instances may be encountered with tubes of high transconductance, however, when a doubler will oscillate in t.g.t.p. fashion.

Frequency multipliers using tubes are operated Class C, with the bias and drive levels adjusted for plate-current conduction of less than 180 degrees.

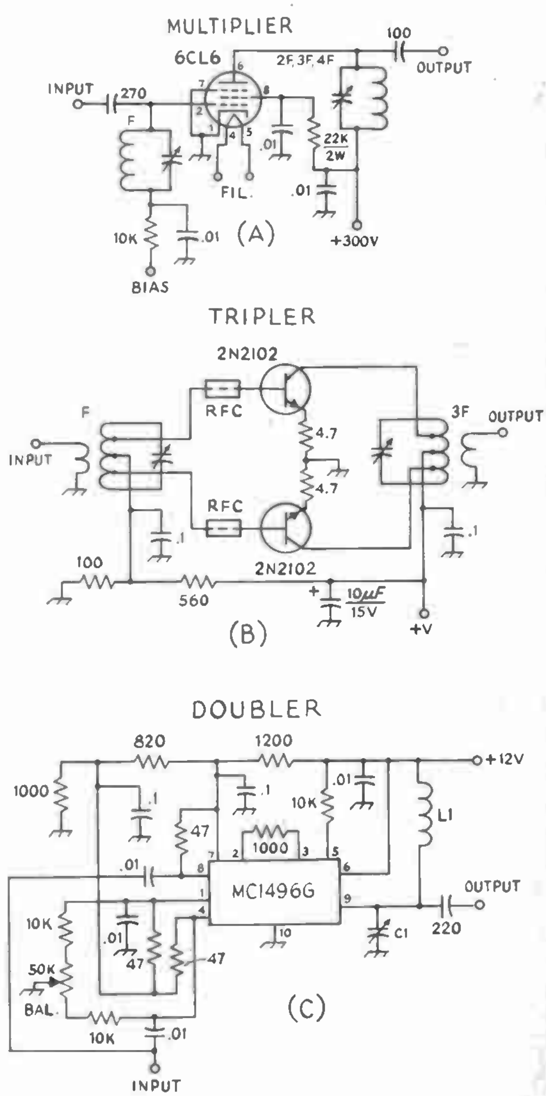


Fig. 6-14 – Frequency-multiplier circuits.

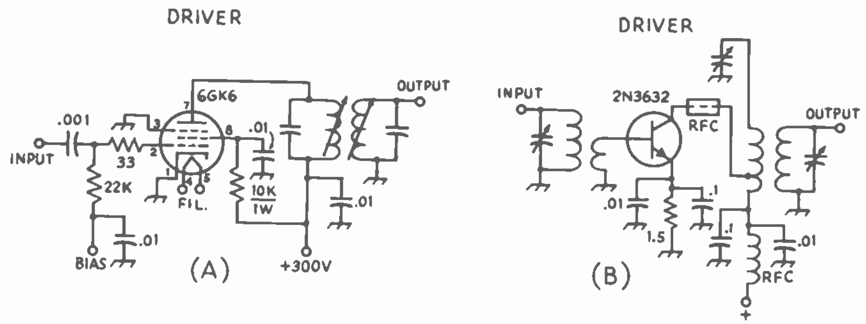


Fig. 6-15 — Driver stages using (A) a pentode tube and (B) a bipolar power transistor.

For maximum efficiency, a doubler requires a plate-conduction angle of about 110 degrees, while a tripler needs 100 degrees, a quadrupler 80 degrees, and a quintupler 65 degrees. For higher orders of multiplication increased bias and more drive are needed.

A typical circuit using a 6CL6 pentode tube is shown in Fig. 6-14A. The input circuit is tuned to the driving frequency while the output tank is set for the desired harmonic. If such a multiplier were to be operated directly into an antenna, additional selectivity would be necessary to prevent the radiation of harmonic energy (other than the desired frequency).

Push-Push Multipliers

A two-tube circuit which works well at even harmonics, but not at the fundamental or odd harmonics, is known as the push-push circuit. The grids are connected in push-pull while the plates are connected in parallel. The efficiency of a doubler using this circuit approaches that of a straight amplifier.

This arrangement has an advantage in some applications. If the heater of one tube is turned off, its grid-plate capacitance, being the same as that of the remaining tube, serves to neutralize the circuit. Thus provision is made for either straight amplification at the fundamental with a single tube, or doubling frequency with two tubes.

Push-Pull Multiplier

A single- or parallel-tube multiplier will deliver output at either even or odd multiples of the exciting frequency. A push-pull stage does not work as a doubler or quadrupler but it will work as a tripler.

Transistor Multipliers

A transistor develops harmonic energy with good efficiency, often causing harmonic-output problems in straight-through amplifiers. Two harmonic-generating modes are present, parametric multiplication and multiplication caused by the nonlinear characteristic presented by the base-collector junction. Transistors may be used in single-ended, push-pull, or push-push circuits. A typical push-pull tripler is shown in Fig. 6-14B. A small amount of forward bias has been added to the bases of the 2N2102s to reduce the amount of drive required. If a high level of drive is available, the bias circuit may be omitted.

A number of integrated circuits can be employed as frequency multipliers. The circuit at C uses a Motorola MC1496G (or the Signetics S5596, or Fairchild μA796) as a doubler. The input signal is balanced out in the IC, so only the desired second harmonic of the input frequency appears at the output. With suitable bypass capacitors this doubler can be used from audio to vhf.

DRIVERS

Pentode tubes are usually chosen for the driver stages of tube transmitters because they provide high amplification, often without requiring neutralization. Many of the receiving-type pentodes and smaller TV sweep tubes may be employed. The 6CL6, 6GK6, 12BY7A, 6BA6, 6AU6, and 6DC6 are often chosen. In cw and fm service the driver stage is operated Class C, while for ssb operation the Class-A mode is preferred to keep distortion to a minimum (third-order products at least 50 dB down). In ssb exciters alc voltage is often applied to a driver stage, in which case a semiremote-cutoff tube is desirable. Sharp-cutoff types are not acceptable because of a rapid increase in distortion as alc voltage drives the grid increasingly negative.

A typical tube driver stage is shown in Fig. 6-15 at A. The output load is a parallel-resonant circuit. Often a bandpass network is used so that the stage does not have to be tuned by a panel control. Also, coupling with a bandpass transformer provides a higher order of attenuation of harmonic and spurious signals. At Fig. 6-15B, a 2N3632 medium-power transistor serves as a Class-C driver. Note that this circuit is not suitable for ssb service.

Broadband Driver

Transistor circuits often require complex interstage coupling networks, because of the low input and output impedance characteristics of bipolar devices. Designing a solid-state multiband hf transmitter often requires some very complex band-switch arrangements. To eliminate this problem, the current trend is to use a broadband multistage driver that covers 3.5 to 30 MHz, for example, without switching or tuning adjustments. A typical circuit, similar to that used in Signal/One's CX-7 transceiver, is shown in Fig.

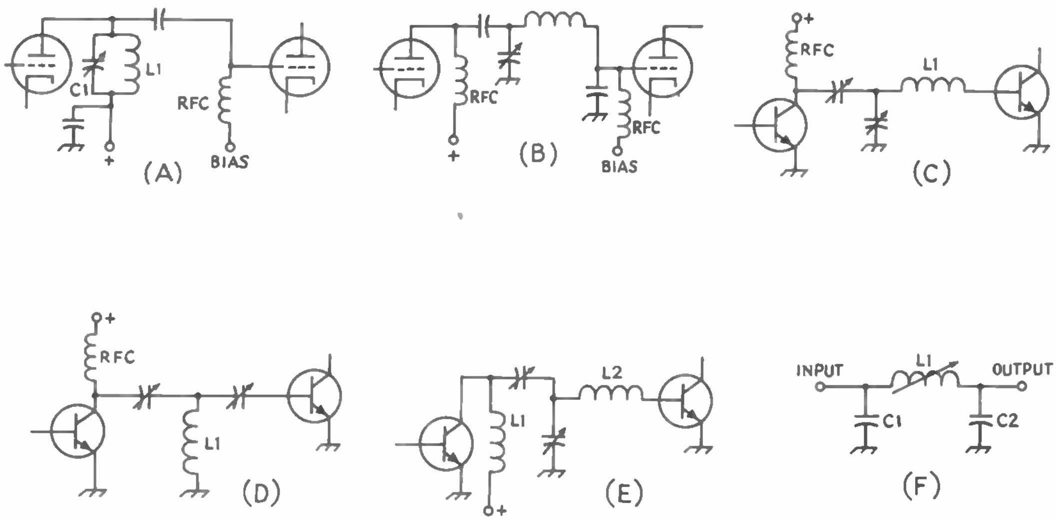


Fig. 6-16 — Interstage coupling networks for (A, B) tubes, (C-E) transistor stages, and (F) a grounded-grid amplifier.

6-17. Only a few millivolts of ssb or cw drive will provide sufficient output to drive a 4CX250B operating Class AB_1. Interstage coupling is accomplished with broadband toroidal transformers. Feedback is added from the collector to the emitter of each bipolar-transistor stage to improve linearity. Output impedance of the broadband driver is approximately 390 ohms.

Interstage Coupling

To achieve the maximum transfer of power between the driver and the succeeding amplifier stage, the output impedance of the driver must be matched to the input impedance of the following amplifier. Some form of rf coupling or impedance-matching network is needed. The capacitive system of Fig. 6-16A is the simplest of all coupling systems. In this circuit, the plate tank circuit of the driver, C1L1, serves also as the grid tank of the amplifier. Although it is used more frequently than any other system, it is less flexible and has certain limitations that must be taken into consideration.

The two stages cannot be separated physically any appreciable distance without involving loss in transferred power, radiation from the coupling lead and the danger of feedback from this lead. Since both the output capacitance of the driver tube and the input capacitance of the amplifier are across the single circuit, it is sometimes difficult to obtain a tank circuit with a sufficiently low Q to provide an efficient circuit at the higher frequencies. The coupling can be varied by altering the capacitance of the coupling capacitor, C2. The driver load impedance is the sum of the amplifier grid resistance and the reactance of the coupling capacitor in series, the coupling capacitor serving simply as a series reactor. The driver load resistance increases with a decrease in the capacitance of the coupling capacitor.

When the amplifier grid impedance is lower than the optimum load resistance for the driver, a transforming action is possible by tapping the grid down on the tank coil, but this is not recommended because it invariably causes an increase in vhf harmonics and sometimes sets up a parasitic circuit.

So far as coupling is concerned, the Q of the circuit is of little significance. However, the other considerations discussed earlier in connection with tank-circuit Q should be observed.

Pi-Network Interstage Coupling

A pi-section tank circuit, as shown in Fig. 6-16B, may be used as a coupling device between screen-grid amplifier stages. The circuit can also be considered a coupling arrangement with the grid of the amplifier tapped down on the circuit by means of a capacitive divider. In contrast to the tapped-coil method mentioned previously, this system will be very effective in reducing vhf harmonics, because the output capacitor provides a direct capacitive shunt for harmonics across the amplifier grid circuit.

To be most effective in reducing vhf harmonics, the output capacitor should be a mica capacitor connected directly across the tube-socket terminals. Tapping down on the circuit in this manner also helps to stabilize the amplifier. Since the coupling to the grid is comparatively loose under any condition, it may be found that it is impossible to utilize the full power capability of the driver stage. If sufficient excitation cannot be obtained, it may be necessary to raise the plate voltage of the

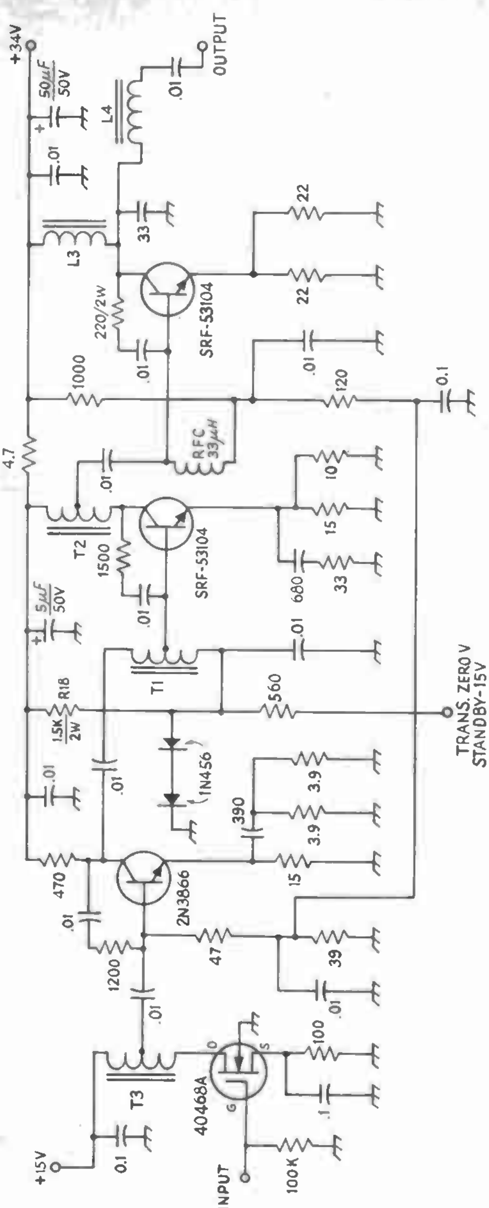


Fig. 6-17 — A solid-state broadband driver for 3 to 30 MHz. The design of transformers T1, T2 and T3 is covered later in the chapter.

driver, if this is permissible. Otherwise a larger driver tube may be required. As shown in Fig. 6-16B, parallel driver plate feed and amplifier grid feed are necessary.

Coupling Transistor Stages

In stages using bipolar power transistors, the input circuit must provide a match between the driver collector and the PA base. The latter exhibits a very low impedance. The input impedance of an rf power transistor is between several tenths of an ohm and several ohms. Generally, the higher the power rating of the device, the lower the input impedance. The base connection also has a reactive component which is capacitive at low frequencies and inductive at higher frequencies. At some frequency, usually between 50 and 150 MHz, the base lead will be self-resonant. The input impedance will vary with drive level, which makes a cut-and-try adjustment of the interstage network necessary.

An interstage network must provide the proper impedance transformation while tuning out reactance in the transistors. The reactive components of the base and collectors of power transistors are of such magnitude that they must be included in any network calculations. Fig. 6-16 shows several networks capable of interstage matching in a multistage transistor amplifier. At C, a T network is pictured. The value of the inductor is chosen so that its reactance is much greater than the capacitive reactance of the second transistor's base circuit. The capacitive divider provides the impedance match between the collector and the base.

The circuit of 6-16D is also basically a T network in which both the inductor and second capacitor are chosen to have reactance of a greater magnitude than the base-emitter capacitance of the second transistor. The circuits of C and D require that the collector of the driver transistor be shunt fed through a high-impedance rf choke. Fig. 6-16E shows a coupling network that eliminates the need for a choke. Here the collector of the driver transistor is parallel-tuned and the base-emitter junction of the following stage is series-tuned.

The remaining circuit, Fig. 6-16F, shows the pi-section network that is often used to match the 50-ohm output of an exciter to a grounded-grid power amplifier. A Q of 1 or 2 is chosen so that the circuit will be broad enough to operate across an amateur band without retuning. The network is designed for a 50-ohm input impedance and to match an output load of 30 to 150 ohms (the impedance range of the cathode of typical grounded-grid stages). Typical LC values are given in the construction projects presented later in this chapter.

RF POWER AMPLIFIER CIRCUITRY

Tube Operating Conditions

In addition to proper tank and output-coupling circuits, an rf amplifier must be provided with suitable operating voltages and an rf driving or excitation voltage. All rf amplifier tubes require a voltage to operate the filament or heater (ac is usually permissible), and a positive dc voltage between the plate and filament or cathode (plate voltage). Most tubes also require a negative dc voltage (biasing voltage) between control grid (grid No. 1) and filament or cathode. Screen-grid tubes require in addition a positive voltage (screen voltage or grid No. 2 voltage) between screen and filament or cathode.

Biasing and plate voltages may be fed to the tube either in series with or in parallel with the associated rf tank circuit as discussed in the chapter on electrical laws and circuits.

It is important to remember that true plate, screen or biasing voltage is the voltage between the particular electrode and filament or cathode. Only when the cathode is directly grounded to the chassis may the electrode-to-chassis voltage be taken as the true voltage. The required rf driving voltage is applied between grid and cathode.

Power Input and Plate Dissipation

Plate power input is the dc power input to the plate circuit (dc plate voltage X dc plate current). Screen power input likewise is the dc screen voltage X the dc screen current.

Plate dissipation is the difference between the rf power delivered by the tube to its loaded plate tank circuit and the dc plate power input. The screen, on the other hand, does not deliver any output power, and therefore its dissipation is the same as the screen power input.

TRANSMITTING-TUBE RATINGS

Tube manufacturers specify the maximum values that should be applied to the tubes they produce. They also publish sets of typical operating values that should result in good efficiency and normal tube life.

Maximum values for all of the most popular transmitting tubes will be found in the tables of transmitting tubes in the last chapter. Also included are as many sets of typical operating values as space permits. However, it is recommended that the amateur secure a transmitting-tube manual from the manufacturer of the tube or tubes he plans to use.

CCS and ICAS Ratings

The same transmitting tube may have different ratings depending upon the manner in which the tube is to be operated, and the service in which it is to be used. These different ratings are based primarily upon the heat that the tube can safely dissipate. Some types of operation, such as with grid or screen modulation, are less efficient than others, meaning that the tube must dissipate more heat. Other types of operation, such as cw or single-sideband phone are intermittent in nature, resulting in less average heating than in other modes where there is a continuous power input to the tube during transmissions. There are also different ratings for tubes used in transmitters that are in almost constant use (CCS – Continuous Commercial Service), and for tubes that are to be used in transmitters that average only a few hours of daily operation (ICAS – Intermittent Commercial and Amateur Service). The latter are the ratings used by amateurs who wish to obtain maximum output with reasonable tube life.

Maximum Ratings

Maximum ratings, where they differ from the values given under typical operating values, are not normally of significance to the amateur except in special applications. No single maximum value should be used unless all other ratings can simultaneously be held within the maximum values. As an example, a tube may have a maximum plate-voltage rating of 2000, a maximum plate-current rating of 300 mA, and a maximum plate-power-input rating of 400 watts. Therefore, if the maximum plate voltage of 2000 is used, the plate current should be limited to 200 mA (instead of 300 mA) to stay within the maximum power-input rating of 400 watts.

SOURCES OF TUBE ELECTRODE VOLTAGES

Filament or Heater Voltage

The heater voltage for the indirectly heated cathode-type tubes found in low-power classifications may vary 10 percent above or below rating without seriously reducing the life of the tube. But the voltage of the higher-power filament-type tubes should be held closely between the rated voltage as a minimum and 5 percent above rating as a maximum. Make sure that the plate power drawn from the power line does not cause a drop in filament voltage below the proper value when plate power is applied.

Thoriated-type filaments lose emission when the tube is overloaded appreciably. If the overload

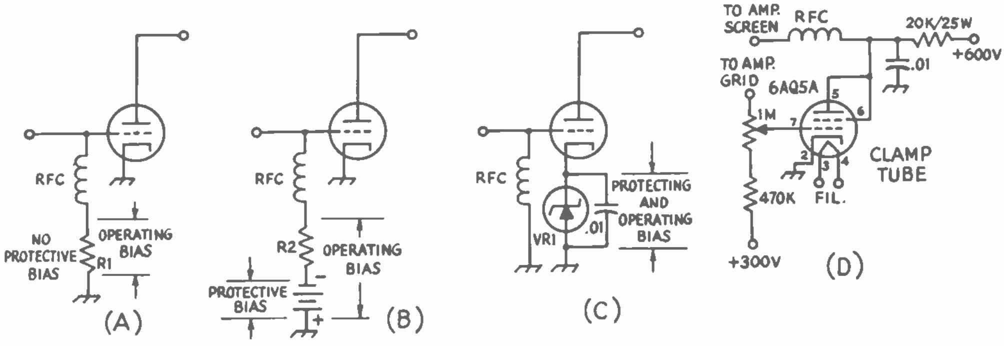


Fig. 6-18 – (A-C) Various systems for obtaining protective and operating bias. (D) Screen clamper circuit for protecting power tetrodes.

has not been too prolonged, emission sometimes may be restored by operating the filament at rated voltage with all other voltages removed for a period of 10 minutes, or at 20 percent above rated voltage for a few minutes.

Plate Voltage

Dc plate voltage for the operation of rf amplifiers is most often obtained from a transformer-rectifier-filter system (see power-supply chapter) designed to deliver the required plate voltage at the required current. However, batteries or other dc-generating devices are sometimes used in certain types of operation (see portable-mobile chapter).

Bias and Tube Protection

Several methods of obtaining bias are shown in Fig. 6-18. At A, bias is obtained by the voltage drop across a resistor in the grid dc return circuit when rectified grid current flows. The proper value of resistance may be determined by dividing the required biasing voltage by the dc grid current at which the tube will be operated. Then, so long as the rf driving voltage is adjusted so that the dc grid current is the recommended value, the biasing voltage will be the proper value. The tube is biased only when excitation is applied, since the voltage drop across the resistor depends upon grid-current flow. When excitation is removed, the bias falls to zero. At zero bias most tubes draw power far in excess of the plate-dissipation rating. So it is advisable to make provision for protecting the tube when excitation fails by accident, or by intent as it does when a preceding stage in a cw transmitter is keyed.

If the maximum cw ratings shown in the tube tables are to be used, the input should be cut to zero when the key is open. Aside from this, it is not necessary that plate current be cut off completely but only to the point where the rated dissipation is not exceeded. In this case plate-modulated phone ratings should be used for cw operation, however.

With triodes this protection can be supplied by obtaining all bias from a source of fixed voltage, as shown in Fig. 6-18B. It is preferable, however, to use only sufficient fixed bias to protect the tube and obtain the balance needed for operating bias from a grid leak. The grid-leak resistance is calculated as above, except that the fixed voltage is subtracted first.

Fixed bias may be obtained from dry batteries or from a power pack (see power-supply chapter). If dry batteries are used, they should be checked periodically, since even though they may show normal voltage, they eventually develop a high internal resistance.

In Fig. 16-8D, bias is obtained from the voltage drop across a Zener diode in the cathode (or filament center-tap) lead. Protective bias is obtained by the voltage drop across VR1 as a result of plate (and screen) current flow. Since plate current must flow to obtain a voltage drop across the resistor, it is obvious that cutoff protective bias cannot be obtained.

The voltage of the cathode biasing Zener diode VR1 should be chosen for the value which will give the correct operating bias voltage with rated grid, plate and screen currents flowing with the amplifier loaded to rated input. When excitation is removed, the input to most types of tubes will fall to a value that will prevent damage to the tube, at least for the period of time required to remove plate voltage. A disadvantage of this biasing system is that the cathode rf connection to ground depends upon a bypass capacitor.

Screen Voltage

For cw and fm operation, and under certain conditions of phone operation (see amplitude-modulation chapter), the screen may be operated from a power supply of the same type used for plate supply, except that voltage and current ratings should be appropriate for screen requirements. The screen may also be operated through a series resistor or voltage-divider from a source of higher voltage, such as the plate-voltage supply, thus making a separate supply for the screen unnecessary. Certain precautions are necessary, depending upon the method used.

It should be kept in mind that screen current varies widely with both excitation and loading. If the screen is operated from a fixed-voltage source, the tube should never be operated without plate voltage and load, otherwise the screen may be damaged within a short time. Supplying the screen through a series dropping resistor from a higher-voltage source, such as the plate supply, affords a measure of protection, since the resistor causes the screen voltage to drop as the current increases, thereby limiting the power drawn by the screen. However, with a resistor, the screen voltage may vary considerably with excitation, making it necessary to check the voltage at the screen terminal under actual operating conditions to make sure that the screen voltage is normal. Reducing excitation will cause the screen current to drop, increasing the voltage; increasing excitation will have the opposite effect. These changes are in addition to those caused by changes in bias and plate loading, so if a screen-grid tube is operated from a series resistor or a voltage divider, its voltage should be checked as one of the final adjustments after excitation and loading have been set.

An approximate value for the screen-voltage dropping resistor may be obtained by dividing the voltage *drop* required from the supply voltage (difference between the supply voltage and rated screen voltage) by the rated screen current in decimal parts of an ampere. Some further adjustment may be necessary, as mentioned above, so an adjustable resistor with a total resistance above that calculated should be provided.

Protecting Screen-Grid Tubes

Considerably less grid bias is required to cut off an amplifier that has a fixed-voltage screen supply than one that derives the screen voltage through a high value of dropping resistor. When a "stiff" screen voltage supply is used, the necessary grid cutoff voltage may be determined from an inspection of the tube curves or by experiment.

When the screen is supplied from a series dropping resistor, the tube can be protected by the use of a clamper tube, as shown in Fig. 6-18D. The grid-leak bias of the amplifier tube with excitation is supplied also to the grid of the clamper tube. This is usually sufficient to cut off the clamper tube. However, when excitation is removed, the clamper-tube bias falls to zero and it draws enough current through the screen dropping resistor usually to limit the input to the amplifier to a safe value. If complete screen-voltage cutoff is desired, a Zener diode may be inserted in the screen lead. The regulator diode voltage rating should be high enough so that it will cease conducting when excitation is removed.

Feeding Excitation to the Grid

The required rf driving voltage is supplied by an oscillator generating a voltage at the desired frequency, either directly or through intermediate amplifiers, mixers, or frequency multipliers.

As explained in the chapter on vacuum-tube fundamentals, the grid of an amplifier operating under Class C conditions must have an exciting voltage whose peak value exceeds the negative biasing voltage over a portion of the excitation cycle. During this portion of the cycle, current will flow in the grid-cathode circuit as it does in a diode circuit when the plate of the diode is positive in respect to the cathode. This requires that the rf driver supply power. The power required to develop the required peak driving voltage across the grid-cathode impedance of the amplifier is the rf driving power.

The tube tables give approximate figures for the grid driving power required for each tube under various operating conditions. These figures, however, do not include circuit losses. In general, the driver stage for any Class C amplifier should be capable of supplying at least three times the driving power shown for typical operating conditions at frequencies up to 30 MHz and from three to ten times at higher frequencies.

Since the dc grid current relative to the biasing voltage is related to the peak driving voltage, the dc grid current is commonly used as a convenient indicator of driving conditions. A driver adjustment that results in rated dc grid current when the dc bias is at its rated value, indicates proper excitation to the amplifier when it is fully loaded.

In coupling the grid input circuit of an amplifier to the output circuit of a driving stage the objective is to load the driver plate circuit so that the desired amplifier grid excitation is obtained without exceeding the plate-input ratings of the driver tube.

Driving Impedance

The grid-current flow that results when the grid is driven positive in respect to the cathode over a portion of the excitation cycle represents an average resistance across which the exciting voltage must be developed by the driver. In other words, this is the load resistance into which the driver plate circuit must be coupled. The approximate grid input resistance is given by:

$$\textit{Input impedance}\ (\text{ohms}) = \frac{\textit{driving power (watts)}}{\textit{dc grid current}\ (\text{mA})^2} \times 620{,}000$$

For normal operation, the driving power and grid current may be taken from the tube tables. Since the grid input resistance is a matter of a few thousand ohms, an impedance step-up is necessary if the grid is to be fed from a low-impedance transmission line

TRANSISTOR RATINGS

Transistor ratings are similar in some respects to the maximum limits given for tubes. However, solid-state devices are generally not so forgiving of overload; they can quickly be ruined if a voltage or current parameter of the device is exceeded. All semiconductors undergo irreversible changes if their temperature is allowed to go above a critical limit.

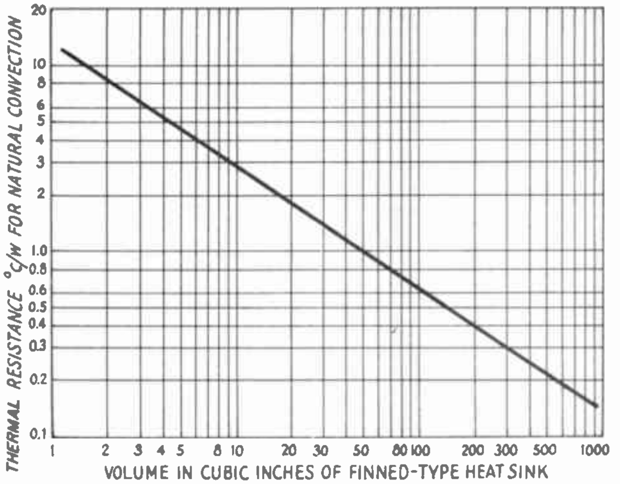
Voltage Rating

In general, the higher the collector-emitter voltage rating of a transistor the less the chance of damage when used as an rf power amplifier. A mismatched load, or the loss of the load entirely, causes high voltages to appear between the collector and emitter of the transistor. If the maximum rating is exceeded, the transistor may break down and pass reverse current. Transistor manufacturers are now including a resistance in series with the emitter lead of each of the many junctions that make up the power transistor as break-down protection. This technique is called ballasting or balanced emitters. Another way to protect a power transistor is to include a Zener diode from collector to emitter. The break-down voltage rating of the diode should be above the peak rf voltage to be developed in the circuit, but below the maximum rating of the power device.

Current and Heat

The current that a power device can stand is related to its ability to dissipate heat. A transistor is physically small, so high-power models must use effective heat radiators, called heat sinks, to insure that the operating temperature is kept to a moderate value even when large currents are flowing through the device.

Cooling considerations for practical solid-state amplifiers are outlined below. Manufacturer's specification sheets describe a safe operating area for an individual power transistor. Also, transistors are rated in terms of power output, rather than input, so it should be remembered that a device specified to deliver 80 watts of output power will probably be running 160 watts or more input. Transistor amplifiers pass an appreciable amount of driver power to the output, as do grounded-grid tube stages, and this fact must also be taken into account by the circuit designer.



(A)

Device Case	*5W*	*10W*	*25W*	*50W*	*100W*
TO-5	17.2	7.2	1.2	.71	.35
TO-44	1.2	9.2	.44	n/a	n/a

(B)

Fig. 6-19 – (A) Graph to determine the thermal resistance of a heat sink of a given size. The heat sink volume may be computed by multiplying cross-sectional area by height. (B) Approximate thermal resistance needed for proper cooling of two types of transistor cases when operated at the proper levels given.

COOLING

Tubes

Vacuum tubes must be operated within the temperature range specified by the manufacturer if long tube life is to be achieved. Tubes with glass envelopes rated at up to 25 watts of plate dissipation may be run without forced-air cooling, if a moderate amount of cooling by convection can be arranged. If a cane-metal enclosure is used, and a ring of 1/4-inch diameter holes are placed around the tube socket, normal air flow can be relied upon to remove excess heat at room temperatures.

For tubes with greater plate dissipation, or those operated with plate currents in excess of the manufacturer's ratings (often the case with TV sweep tubes) forced air cooling with a fan or blower is needed. Fans, especially those designed for cooling hi-fi cabinets, are preferred because they operate quietly. However, all fans lose their ability to move air when excessive back pressure exists. For applications where a stream of air must be directed through a tube socket, a blower is usually required. Blowers vary in their ability to work against back pressure, so this specification should be checked when selecting a particular model. Some air will always leak around the socket and through other holes in a chassis, so the blower chosen should have a capacity which is 30 to 50 percent beyond that called for by the tube manufacturer.

An efficient blower is required when using the external-anode tubes, such as the 4X150A. Such tubes represent a trade-off which allows high-power operation with a physically small device at the expense of increased complexity in the cooling system. Other types of external-anode tubes are now being produced for conductive cooling. An electrical insulator which is also an excellent thermal conductor, such as AlSiMag, couples the tube to a heat sink. Requirements for the heat dissipator are calculated in the same way as for power transistors, as outlined below. Similar tubes are made with special anode structures for water or vapor cooling, allowing high-power operation without producing an objectionable noise level from the cooling system.

Transistor Cooling

Bipolar power transistors usually have the collector connected directly to the case of the device, as the collector must dissipate most of the heat generated when the transistor is in operation. However, even the larger case designs cannot conduct heat away fast enough to keep the operating temperature of the device functioning within the safe area, the maximum temperature that a device can stand without damage. Safe area is usually specified in a device data sheet, often in graphical form. Germanium power transistors may be operated at up to 100 degrees C while the silicon types may be run at up to 200 degrees C. Leakage currents in germanium devices can be very high at elevated temperatures; thus, for power applications silicon transistors are preferred.

A thermal sink, properly chosen, will remove heat at a rate which keeps the transistor junction temperature in the safe area. For low-power applications a simple clip-on heat sink will suffice, while for 100-watts of input power a massive cast-aluminum finned radiator will be necessary. In general, the case temperature of a power transistor must be kept below the point at which it will produce a burn when touched.

Heat-Sink Design

Simple heat sinks, made as described in the Construction Practices chapter, can be made more effective (by 25 percent or more) by applying a coat of flat-black paint. Finned radiators are most effective when placed where maximum air flow can be achieved – outside a case with the fins placed vertically. The size of a finned heat sink required to give a desired thermal resistance, a measure of the ability to dissipate heat, is shown in Fig. 6-19A. Fig. 6-19B is a simplified chart of the thermal resistance needed in a heat sink for transistors in TO-5 and TO-44 cases. These figures

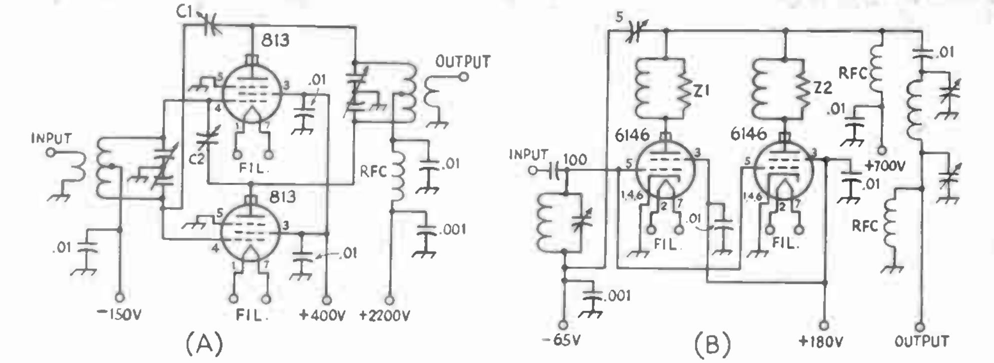


Fig. 6-20 – Typical (A) push-pull and (B) parallel amplifier circuits.

are based on several assumptions, so they can be considered a *worst-case* situation. Smaller heat sinks may be usable.

The thermal design of solid-state circuits has been covered in *QST* for April, 1972. The surface contact between the transistor case and the heat sink is extremely important. To keep the sink from being "hot" with dc, a mica insulator is usually employed between the transistor case and the heat dissipator. Newer types of transistors have a case mounting bolt insulated from the collector so that it may be connected directly to the heat sink. Whatever the arrangement, the use of a conductive compound such as silicone grease (Corning PC-4) is recommended between the transistor and the sink. For high-power designs, it may be desirable to add a small cooling fan, providing a stream of air across the heat sink, to keep the size of the heat dissipator within reasonable limits. Even a light air flow greatly increases the radiator's ability to dispose of excess heat.

OUTPUT POWER FROM TRANSMITTERS

CW or FM: In a cw or fm transmitter, any class of amplifier can be used as an output or intermediate amplifier. (For reasonable efficiency, a frequency multiplier *must* be operated Class C.) Class-C operation of the amplifier gives the highest efficiency (65 to 75 percent), but it is likely to be accompanied by appreciable harmonics and consequent TVI possibilities. If the excitation is keyed in a cw transmitter, Class-C operation of subsequent amplifiers will, under certain conditions, introduce key clicks not present on the keyed excitation (see chapter on Code Transmission). The peak envelope power (PEP) input or output of any cw (or fm) transmitter is the "key-down" input or output.

A-M: In an amplitude-modulated phone transmitter, plate modulation of a Class-C output amplifier results in the highest output for a given input to the output stage. The efficiency is the same as for cw or fm with the same amplifier, from 65 to 75 percent. (In most cases the manufacturer rates the *maximum allowable input* on plate-modulated phone at about 2/3 that of cw or fm.) A plate-modulated stage running 100 watts input will deliver a carrier output of from 65 to 75 watts, depending upon the tube, frequency and circuit factor. The PEP output of any a-m signal is four times the carrier output power, or 260 to 300 watts for the 100-watt input example.

Grid- (control or screen) modulated output amplifiers in a-m operation run at a carrier efficiency of 30 to 35 percent, and a grid-modulated stage with 100 watts input has a carrier output of 30 to 35 watts. (The PEP output, four times the carrier output, is 120 to 140 watts.)

Running the legal input limit in the United States, a plate-modulated output stage can deliver a carrier output of 650 to 750 watts, while a screen- or control-grid-modulated output amplifier can deliver only a carrier of 300 to 350 watts.

SSB: Only *linear* amplifiers can be used to amplify ssb signals without distortion, and this limits the choice of output amplifier operation to Classes A, AB_1, AB_2, and B. The efficiency of operation of these amplifiers runs from about 20 to 65 percent. In all but Class-A operation the indicated (by plate-current meter) input will vary with the signal, and it is not possible to talk about relative inputs and outputs as readily as it is with other modes. Therefore linear amplifiers are rated by PEP (input or output) at a given distortion level, which indicates not only how much ssb signal they will deliver but also how effective they will be in amplifying an a-m signal.

LINEAR AMPLIFIERS FOR A-M: In considering the practicality of adding a linear output amplifier to an existing a-m transmitter, it is necessary to know the carrier output of the a-m transmitter and the PEP output rating of the linear amplifier. Since the PEP output of an a-m signal is four times the carrier output, it is obvious that a linear with a PEP output rating of only four times the carrier output of the a-m transmitter is no amplifier at all. If the linear amplifier has a PEP output rating of 8 times the a-m transmitter carrier output, the output power will be doubled and a 3-dB improvement will be obtained. In most cases a 3-dB change is *just discernible* by the receiving operator.

By comparison, a linear amplifier with a PEP output rating of four times an existing ssb, cw or fm transmitter will *quadruple* the output, a 6-dB improvement, It should be noted that the linear amplifier must be rated for the mode (ssb, cw or fm) with which it is to be used.

GROUNDED-GRID AMPLIFIERS: The preceding discussion applies to vacuum-tube amplifiers connected in a grounded-cathode or grounded-grid circuit. However, there are a few points that apply only to grounded-grid amplifiers.

A tube operated in a given class (AB_1, B, C) will require more driving power as a grounded-grid amplifier than as a grounded-cathode amplifier. This is not because the grid losses run higher in the grounded-grid configuration but because some of the driving power is coupled directly through the tube and appears in the plate load circuit. Provided enough driving power is available, this increased requirement is of no concern in cw or linear operation. In a-m operation, however, the fed-through power prevents the grounded-grid amplifier from being fully modulated (100 percent).

AMPLIFIER CIRCUITS

Parallel and Push-Pull Amplifiers

The circuits for parallel-tube amplifiers are the same as for a single tube, similar terminals of the tubes being connected together. The grid impedance of two tubes in parallel is half that of a single tube. This means that twice the grid tank capacitance shown in Fig. 6-20B should be used for the same Q.

The plate load resistance is halved so that the plate-tank capacitance for a single tube (Fig. 6-24) also should be doubled. The total grid current will be doubled, so to maintain the same grid bias, the grid-leak resistance should be half that used for a single tube. The required driving power is doubled. The capacitance of a neutralizing capacitor should be doubled and the value of the screen dropping resistor should be cut in half.

In treating parasitic oscillation, it is often necessary to use a choke in each plate lead, rather than one in the common lead to avoid building in a push-pull type of vhf circuit, a factor in obtaining efficient operation at higher frequencies.

Two or more transistors are often operated in parallel to achieve high output power, because several medium-power devices often cost less than a single high-power type. When parallel operation is used, precautions must be taken to insure that equal drive is applied to each transistor. Otherwise, one transistor may "hog" most of the drive and exceed its safe ratings.

A basic push-pull circuit is shown in Fig. 6-20A. Amplifiers using this circuit are cumbersome to bandswitch and consequently are not very popular below 30 MHz. However, since the push-pull configuration places tube input and output capacitances in series, the circuit is often used at 50 MHz and higher.

In the circuit shown at A two 813s are used. Cross neutralization is employed, with C1 connected from the plate of one tube to the grid of the second, while C2 is attached in the reverse order.

GROUNDED-GRID AMPLIFIERS

Fig. 6-21A shows the input circuit of a grounded-grid triode amplifier. In configuration it is similar to the conventional grounded-cathode circuit except that the grid, instead of the cathode, is at ground potential. An amplifier of this type is characterized by a comparatively low input impedance and a relatively high driver power requirement. The additional driver power is not consumed in the amplifier but is "fed through" to the plate circuit where it combines with the normal plate output power. The total rf power output is the sum of the driver and amplifier output powers less the power normally required to drive the tube in a grounded-cathode circuit.

Positive feedback is from plate to cathode through the plate-cathode capacitance of the tube. Since the grounded-grid is interposed between the plate and cathode, this capacitance is small, and neutralization usually is not necessary.

In the grounded-grid circuit the cathode must be isolated for rf from ground. This presents a practical difficulty especially in the case of a filament-type tube whose filament current is large. In plate-modulated phone operation the driver power fed through to the output is not modulated.

The chief application for grounded-grid amplifiers in amateur work below 30 MHz is in the case where the available driving power far exceeds the power that can be used in driving a conventional grounded-cathode amplifier.

Screen-grid tubes are also used sometimes in grounded-grid amplifiers. In some cases, the screen

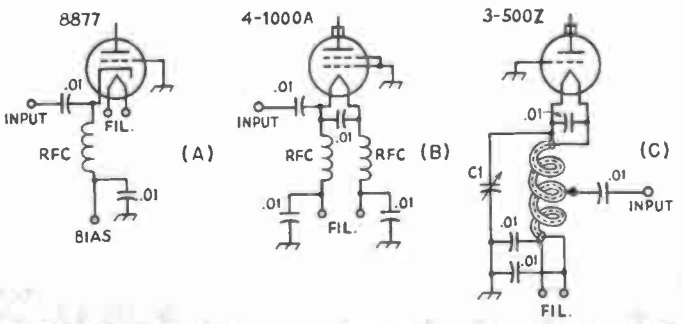


Fig. 6-21 — Input circuits for triode or triode-connected power tubes operated grounded grid.

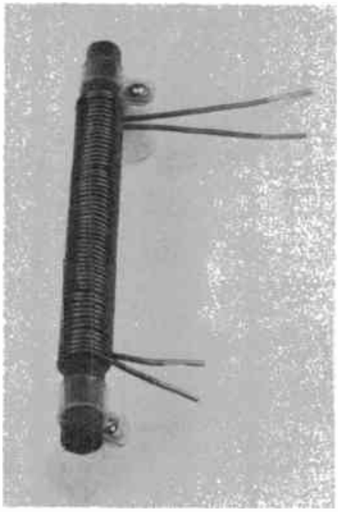


Fig. 6-22 – A 30-A filament choke for a grounded-grid power amplifier consisting of 28 turns of No. 10 enam. wire on a 1/2-inch diameter ferrite rod 7 inches long.

is simply connected in parallel with the grid, as in Fig. 6-21B and the tube operates as a high-μ triode. In other cases, the screen is bypassed to ground and operated at the usual dc potential, as shown at C. Since the screen is still in parallel with the grid for rf, operation is very much like that of a triode except that the positive voltage on the screen reduces driver-power requirements.

In indirectly-heated cathode tubes, the low heater-to-cathode capacitance will often provide enough isolation to keep rf out of the heater transformer and the ac lines. If not, the heater voltage must be applied through rf chokes.

In a directly-heated cathode tube, the filament must be maintained above rf ground. This can be done by using a pair of filament chokes or by using the input tank circuit, as shown in Fig. 6-21C. In the former method, a double solenoid (often wound on a ferrite core) is generally used, although separate chokes can be used. When the tank circuit is used, the tank inductor is wound from two (insulated) conductors in parallel or from an insulated conductor inside a tubing outer conductor. A typical filament choke is shown in Fig. 6-22.

The input impedance of a grounded-grid power stage is usually between 30 and 150 ohms. For circuits similar to those shown in Figs. 6-21A and B some form of input tuning network is needed. A high-*C*, low-*Q* parallel-resonant or pi-section network will suffice. The input network provides benefit other than impedance matching – a reduction in the IM distortion produced by the stage when amplifying an ssb signal. A typical input circuit is shown in Fig. 6-16F. When an amplifier is built for single-band operation, a tank circuit similar to that shown in Fig. 6-21C may be employed. Proper input matching is achieved by tapping the input down on the coil.

TRANSISTOR CIRCUITS

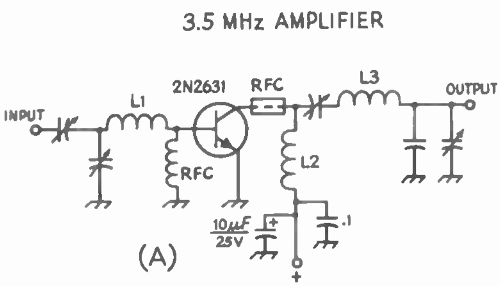
A transistor amplifier requires some means for impedance matching at the input and output of the stage. For conventional narrow-band amplifier designs, impedance matching is achieved with tuned networks (pi, L or T sections or combinations thereof). To simplify band-switching requirements, broadband amplifiers with four octaves or more of bandwidth are desirable. Wide bandwidths are achieved by using a special form of transmission-line transformer for interstage and output coupling that is described later in this chapter.

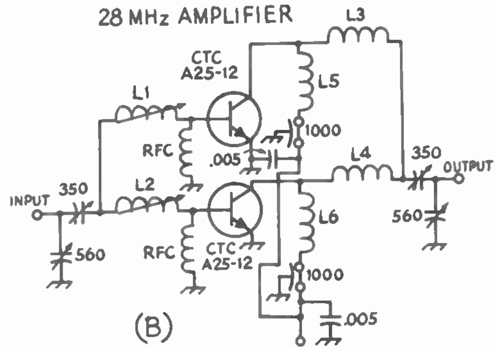
Most solid-state Class-C amplifiers are operated with both the base and emitter leads connected to dc ground. Thus, the transistor is practically off when no driving signal is present. The distortion of the drive signal by such an amplifier is appreciable. However, with cw, fm, or collector-modulated a-m, the harmonics produced are removed from the desired frequency by at least a factor of 2. Thus, harmonic energy can be reduced or eliminated by using appropriate filters.

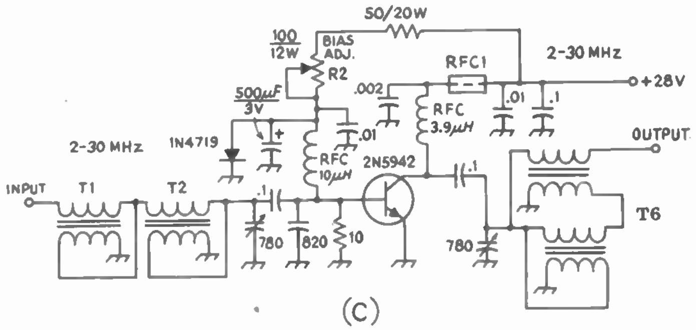
Fig. 6-23A shows a basic Class-C transistor amplifier. The base input is held at dc ground through a radio-frequency choke. A second choke, consisting of two ferrite beads (collector lead), eliminates a tendency to vhf parasitic oscillation. At B, parallel-connected transistors are operated Class C. Adjustment of L1 and L2 provide equal levels of drive. The devices chosen for this circuit are designed for 30- to 50-MHz operation. Below 14 MHz some form of degenerative feedback will be needed to prevent self oscillation, as the gain of the transistors is quite high at lower frequencies.

For ssb operation transistors must be forward biased at the base. The lowest distortion results with Class-A operation, but, efficiency is poor. The best trade off between low distortion and high efficiency is Class-B operation, even though operation in this region introduces some severe requirements for the bias circuit. Whenever a transistor is forward biased, thermal runaway can be a problem. Also, ssb drive varies in amplitude causing large variations in the transistor base current. For best linearity, the dc base-bias voltage should remain constant as the rf drive level is varied. This situation is in conflict with the conditions needed to prevent thermal runaway. Exotic schemes have been designed to provide the proper base bias for Class-B ssb amplification. However, a simple diode circuit such as shown in Figs. 6-23C and D can provide the required dc stability with protection against thermal damage. The ballasted type of transistors are preferred for these circuits. Typical choices for Class-B ssb service are the 2N5941, 2N2942, 2N3375, 2N5070, 2N5071, and the 2N5093. The design of suitable broadband transformers for the circuits of Fig. 6-29 is covered later in this chapter.

The circuits at 6-23C and D are similar except for the choice of the active device. Both designs were developed by K7QWR. The base-bias circuit maintains a steady voltage while supplying current that varies by a factor of 100 to 1 with drive. The gain versus frequency of both circuits follows the







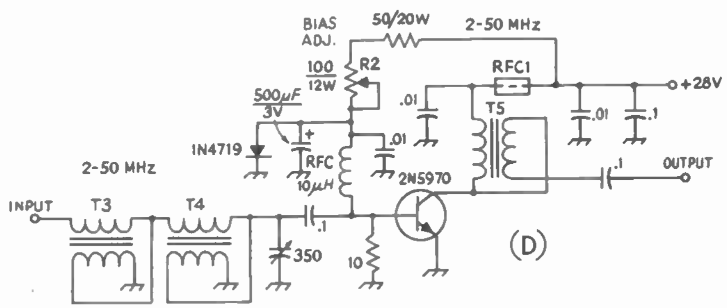


Fig. 6-23 – Some typical transistor power-amplifier circuits. At C, R1 is adjusted for a collector current of 40 mA with no drive, while R2 at D is set for 20 mA collector current with no input. Broadband transformers used consist of the following:

T1, T3, T5 – 6 turns of 2 twisted pairs of No. 26 enam. wire on a Stackpole 57-9322 No. 11 toroid core, connected for 4:1. (See table 6-A.)

T2,T4 – 4 turns of 4 twisted pairs of No. 26 enam. wire on a Stackpole 57-9322 No. 11 toroid core, connected for 4:1.

T6 – 10 turns of 3 twisted pairs of No. 28 enam. wire on two Stackpole 57-9074 No. 11 toroid cores, connected for 9:1

power-output curves of the transistors used, changing from 25 dB at 2 MHz to 13 dB at 30 MHz. IMD is typically 30 dB or more down with either circuit.

RF POWER-AMPLIFIER TANKS AND COUPLING

TANK *Q*

Rf power amplifiers used in amateur transmitters are operated under Class-C or -AB conditions (see chapter on tube fundamentals). The main objective, of course, is to deliver as much fundamental power as possible into a load, *R* without exceeding the tube ratings. The load resistance *R* may be in the form of a transmission line to an antenna, or the input circuit of another amplifier. A further objective is to minimize the harmonic energy (always generated by an amplifier) fed into the load circuit. In attaining these objectives, the *Q* of the tank circuit is of importance. When a load is coupled inductively, the *Q* of the tank circuit will have an effect on the coefficient of coupling necessary for proper loading of the amplifier. In respect to all of these factors, a tank *Q* of 10 to 20 is usually considered optimum. A much lower *Q* will result in less efficient operation of the amplifier tube, greater harmonic output, and greater difficulty in coupling inductively to a load. A much higher *Q* will result in higher tank current with increased loss in the tank coil. Efficiency of a tank circuit is determined by the ratio of loaded *Q* to unloaded *Q* by the relationship:

$$Eff. = 100\left(1 - \frac{Q_L}{Q_U}\right)$$

where Q_L is the loaded *Q* and Q_U is the unloaded *Q*.

The *Q* is determined (see chapter on electrical laws and circuits) by the *L/C* ratio and the load resistance at which the tube is operated. The tube load resistance is related, in approximation, to the ratio of the dc plate voltage to dc plate current at which the tube is operated and can be computed from:

Class-A Tube:

$$R_L = \frac{\text{Plate Volts}}{1.3 \times \text{Plate Current}}$$

Class-B Tube:

$$R_L = \frac{\text{Plate Volts}}{1.57 \times \text{Plate Current}}$$

Class-C Tube:

$$R_L = \frac{\text{Plate Volts}}{2 \times \text{Plate Current}}$$

Transistor:

$$R_L = \frac{(\text{Collector Volts})^2}{2 \times \text{Power Output (Watts)}}$$

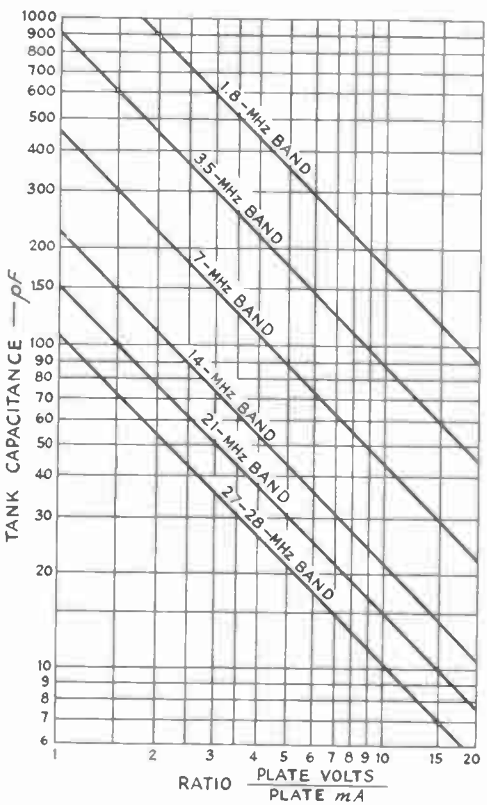


Fig. 6-24 — Chart showing plate tank capacitance required for a Q of 10. Divide the tube plate voltage by the plate current in milliamperes. Select the vertical line corresponding to the answer obtained. Follow this vertical line to the diagonal line for the band in question, and thence horizontally to the left to read the capacitance. For a given ratio of plate voltage/plate current, doubling the capacitance shown doubles the Q. When a split-stator capacitor is used in a balanced circuit, the capacitance of each section may be one half the value given by the chart.

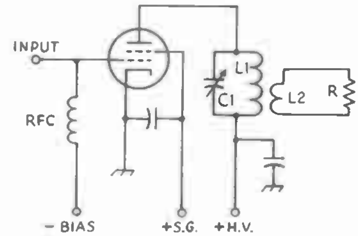


Fig. 6-25 — Inductive-link output coupling circuits.
C1 — Plate tank capacitor — see text and Fig. 6-24 for capacitance.
L1 — To resonate at operating frequency with C1. See *LC* chart and inductance formula in electrical-laws chapter, or use ARRL *Lightning Calculator.*
L2 — Reactance equal to line impedance. See reactance chart and inductance formula in electrical-laws chapter, or use ARRL *Lightning Calculator.*
R — Representing load.

Parallel-Resonant Tank

The amount of C that will give a Q of 10 for various ratios is shown in Fig. 6-24. For a given plate-voltage/plate-current ratio, the Q will vary directly as the tank capacitance, twice the capacitance doubles the Q, etc. For the same Q, the capacitance of *each section* of a split-stator capacitor in a balanced circuit should be half the value shown.

These values of capacitance include the output capacitance of the amplifier tube, the input capacitance of a following amplifier tube if it is coupled capacitively, and all other stray capacitances. At the higher plate-voltage/plate-current ratios, the chart may show values of capacitance, for the higher frequencies, smaller than those attainable in practice. In such a case, a tank Q higher than 10 is unavoidable.

INDUCTIVE-LINK COUPLING

Coupling to Flat Coaxial Lines

When the load R in Fig. 6-25 is located for convenience at some distance from the amplifier, or when maximum harmonic reduction is desired, it is advisable to feed the power to the load through a low-impedance coaxial cable. The shielded construction of the cable prevents radiation and makes it possible to install the line in any convenient manner without danger of unwanted coupling to other circuits.

If the line is more than a small fraction of a wavelength long, the load resistance at its output end should be adjusted, by a matching circuit if necessary, to match the impedance of the cable. This reduces losses in the cable and makes the coupling adjustments at the transmitter independent of the cable length. Matching circuits for use between the cable and another transmission line are discussed in the chapter on transmission lines, while the matching adjustments when the load is the grid circuit of a following amplifier are described elsewhere in this chapter.

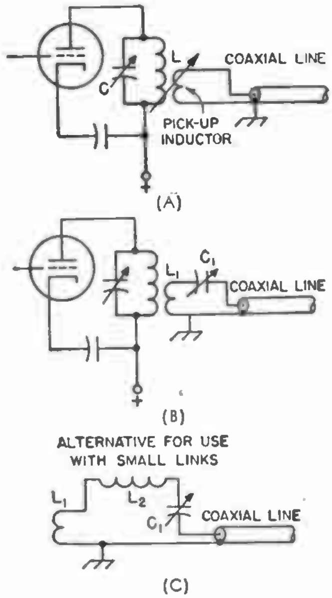


Fig. 6-26 — With flat transmission lines, power transfer is obtained with looser coupling if the line input is tuned to resonance. C1 and L1 should resonate at the operating frequency. See table for maximum usable value of C1. If circuit does not resonate with maximum C1 or less, inductance of L1 must be increased or added in series at L2.

Table 6-A

Capacitance in pF Required for Coupling to Flat Coaxial Lines with Tuned Coupling Circuit[1]

Frequency Band MHz	*Characteristic Impedance of Line* 52 *ohms*	75 *ohms*
3.5	450	300
7	230	150
14	115	75
21	80	50
28	60	40

[1] Capacitance values are maximum usable.
Note: Inductance in circuit must be adjusted to resonate at operating frequency.

Assuming that the cable is properly terminated, proper loading of the amplifier will be assured, using the circuit of Fig. 6-26A, if

1) The plate tank circuit has reasonably higher value of Q. A value of 10 is usually sufficient.

2) The inductance of the pickup or link coil is close to the optimum value for the frequency and type of line used. The optimum coil is one whose self-inductance is such that its reactance at the operating frequency is equal to the characteristic impedance, Z_0, of the line.

3) It is possible to make the coupling between the tank and pickup coils very tight.

The second in this list is often hard to meet. Few manufactured link coils have adequate inductance even for coupling to a 50-ohm line at low frequencies.

If the line is operating with a low SWR, the system shown in Fig. 6-26A will require tight coupling between the two coils. Since the secondary (pickup coil) circuit is not resonant, the leakage reactance of the pickup coil will cause some detuning of the amplifier tank circuit. This detuning effect increases with increasing coupling, but is usually not serious. However, the amplifier tuning must be adjusted to resonance, as indicated by the plate-current dip, each time the coupling is changed.

Tuned Coupling

The design difficulties of using "untuned" pickup coils, mentioned above, can be avoided by using a coupling circuit tuned to the operating frequency. This contributes additional selectivity as well, and hence aids in the suppression of spurious radiations.

If the line is flat the input impedance will be essentially resistive and equal to the Z_0 of the line. With coaxial cable, a circuit of reasonable Q can be obtained with practicable values of inductance and capacitance connected in series with the line's input terminals. Suitable circuits are given in Fig. 6-26 at B and C. The Q of the coupling circuit often may be as low as 2, without running into difficulty in getting adequate coupling to a tank circuit of proper design. Larger values of Q can be used and will result in increased ease of coupling, but as the Q is increased the frequency range over which the circuit will operate without readjustment becomes smaller. It is usually good practice, therefore, to use a coupling-circuit Q just low enough to permit operation, over as much of a band as is normally used for a particular type of communication, without requiring retuning.

Capacitance values for a Q of 2 and line impedances of 52 and 75 ohms are given in the accompanying table. These are the *maximum* values that should be used. The inductance in the circuit should be adjusted to give resonance at the

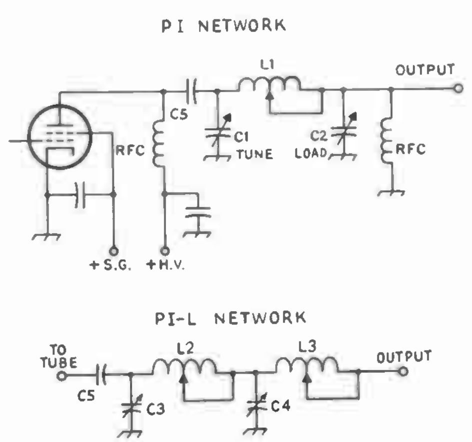


Fig. 6-27 — Pi and pi-L output-coupling networks.

operating frequency. If the link coil used for a particular band does not have enough inductance to resonate, the additional inductance may be connected in series as shown in Fig. 6-26C.

Characteristics

In practice, the amount of inductance in the circuit should be chosen so that, with somewhat loose coupling between L1 and the amplifier tank coil, the amplifier plate current will increase when the variable capacitor, C1, is tuned through the value of capacitance given by the table. The coupling between the two coils should then be increased until the amplifier loads normally, without changing the setting of C1. If the transmission line is flat over the entire frequency band under consideration, it should not be necessary to readjust C1 when changing frequency, if the values given in the table are used. However, it is unlikely that the line actually will be flat over such a range, so some readjustment of C1 may be needed to compensate for changes in the input impedance of the line. If the input impedance variations are not large, C1 may be used as a loading control, no changes in the coupling between L1 and the tank coil being necessary.

The degree of coupling between L1 and the amplifier tank coil will depend on the coupling-circuit Q. With a Q of 2, the coupling should be tight – comparable with the coupling that is typical of "fixed-link" manufactured coils. With a swinging link it may be necessary to increase the Q of the coupling circuit in order to get sufficient power transfer. This can be done by increasing the L/C ratio.

PI AND PI-L OUTPUT TANKS

A pi-section and pi-L tank circuit may also be used in coupling to an antenna or transmission line, as shown in Fig. 6-27. The optimum values of capacitance and inductance are dependent upon values of amplifier power input and output load resistance.

Values for L and C may be taken directly from the charts of Fig. 6-28 if the output load resistance is the usual 52 ohms. It should be borne in mind that these values apply only where the output load is resistive, i.e., where the antenna and line have been matched. Fig. 6-28 and 6-28A were provided by W6FFC.

Output-Capacitor Ratings

The voltage rating of the output capacitor will depend upon the SWR. If the load is resistive, receiving-type air capacitors should be adequate for amplifier input powers up to 2 kW PEP when

	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

Fig. 6-28 – Chart to determine the values of L and C needed for a pi (A) and pi-L (B) network to match a range of input impedances to a 50-ohm load.

R1 Ohms	F MHz	C1 pF	L1 μH	C2 pF	R2 Ohms	Q Qual.
50	3.5	2600	0.94	4153	10	2.9
50	7.0	1179	0.49	1678	10	2.6
50	14.0	579	0.25	801	10	2.5
50	21.0	384	0.16	528	10	2.5
50	29.7	266	0.12	351	10	2.5
50	3.5	2098	1.27	2811	20	2.3
50	7.0	952	0.67	1220	20	2.1
50	14.0	467	0.34	593	20	2.1
50	21.0	310	0.23	393	20	2.0
50	29.7	214	0.16	268	20	2.0
50	3.5	2098	1.43	2533	30	2.3
50	7.0	952	0.76	1131	30	2.1
50	14.0	467	0.38	553	30	2.1
50	21.0	310	0.26	367	30	2.0
50	29.7	214	0.18	253	30	2.0
50	3.5	2098	1.55	2290	40	2.3
50	7.0	952	0.83	1033	40	2.1
50	14.0	467	0.42	506	40	2.1
50	21.0	310	0.28	336	40	2.0
50	29.7	214	0.20	232	40	2.0
50	3.5	2098	1.66	2098	50	2.3
50	7.0	952	0.88	952	50	2.1
50	14.0	467	0.45	467	50	2.1
50	21.0	310	0.30	310	50	2.0
50	29.7	214	0.21	214	50	2.0
50	3.5	2098	1.66	2098	50	2.3
50	7.0	952	0.88	952	50	2.1
50	14.0	467	0.45	467	50	2.1
50	21.0	310	0.30	310	50	2.0
50	29.7	214	0.21	214	50	2.0
75	3.5	1399	2.21	1630	50	2.3
75	7.0	634	1.17	731	50	2.1
75	14.0	311	0.59	358	50	2.1
75	21.0	207	0.40	238	50	2.0
75	29.7	143	0.28	164	50	2.0
100	3.5	1049	2.72	1337	50	2.3
100	7.0	476	1.43	591	50	2.1
100	14.0	234	0.72	288	50	2.1
100	21.0	155	0.48	191	50	2.0
100	29.7	107	0.35	131	50	2.0

R1 Ohms	F MHz	C1 pF	L1 μH	C2 pF	R2 Ohms	Q Qual.
125	3.5	839	3.19	1124	50	2.3
125	7.0	381	1.67	488	50	2.1
125	14.0	187	0.84	237	50	2.1
125	21.0	124	0.56	157	50	2.0
125	29.7	86	0.40	107	50	2.0
150	3.5	699	3.62	957	50	2.3
150	7.0	317	1.89	405	50	2.1
150	14.0	156	0.95	196	50	2.1
150	21.0	103	0.64	129	50	2.0
150	29.7	71	0.45	88	50	2.0
175	3.5	599	4.03	816	50	2.3
175	7.0	272	2.09	333	50	2.1
175	14.0	133	1.05	159	50	2.1
175	21.0	89	0.70	105	50	2.0
175	29.7	61	0.50	70	50	2.0
200	3.5	569	4.26	822	50	2.5
200	7.0	258	2.22	334	50	2.3
200	14.0	127	1.12	160	50	2.2
200	21.0	84	0.74	105	50	2.2
200	29.7	58	0.53	70	50	2.2
225	3.5	543	4.48	827	50	2.7
225	7.0	246	2.34	335	50	2.4
225	14.0	121	1.18	160	50	2.4
225	21.0	80	0.79	106	50	2.4
225	29.7	55	0.56	70	50	2.3
250	3.5	520	4.68	831	50	2.9
250	7.0	236	2.45	336	50	2.6
250	14.0	116	1.23	160	50	2.5
250	21.0	77	0.82	106	50	2.5
250	29.7	53	0.59	70	50	2.5
275	3.5	499	4.86	834	50	3.0
275	7.0	227	2.56	336	50	2.7
275	14.0	111	1.29	160	50	2.7
275	21.0	74	0.86	106	50	2.7
275	29.7	51	0.61	70	50	2.6
300	3.5	481	5.04	836	50	3.2
300	7.0	218	2.66	337	50	2.9
300	14.0	107	1.34	160	50	2.8
300	21.0	71	0.89	106	50	2.8
300	29.7	49	0.64	70	50	2.8

Fig. 6-28A — The following data is for a pi network with a Q of 2 at the top of each band. The Q shown is that for the same inductor at the bottom of the band. The capacitors are shown for the bottom of the band to indicate the maximum capacitance needed. If the transformation ratio exceeds 70 percent of maximum, the Q has been automatically recalculated in order to retain the characteristics of a pi network and that new value shown. Do not forget which end of the network represents 50 ohms!

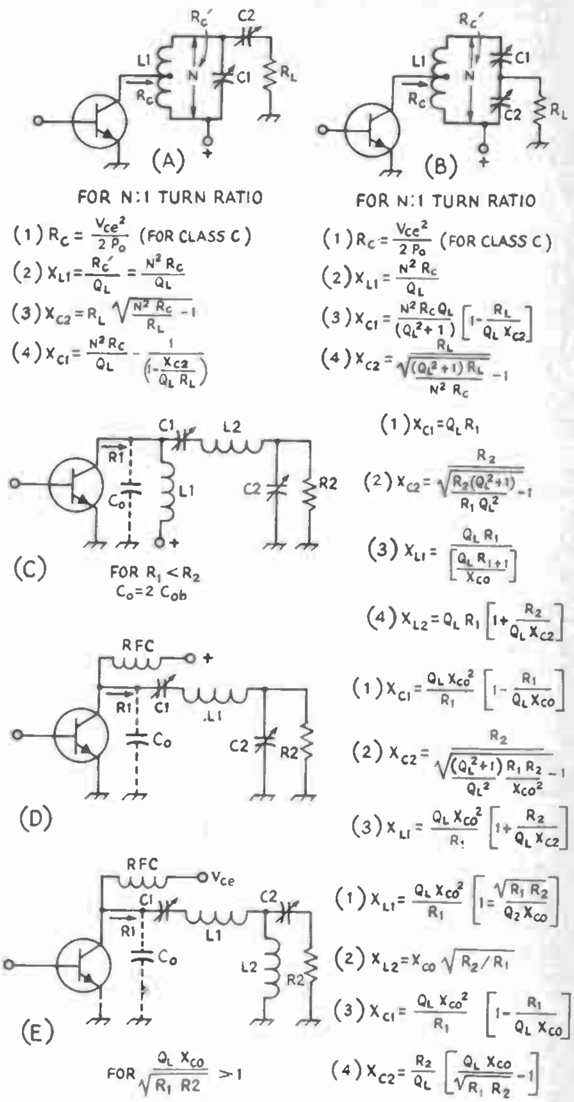
feeding 52- 75-ohm loads. In obtaining the larger capacitances required for the lower frequencies, it is common practice to switch one or more fixed capacitors in parallel with the variable air capacitor. While the voltage rating of a mica or ceramic capacitor may not be exceeded in a particluar case, capacitors of these types are limited in current-carrying capacity. Postage-stamp silver-mica capacitors should be adequate for amplifier inputs over the range from about 70 watts at 28 MHz to 400 watts at 14 MHz and lower. The larger mica capacitors (CM-45 case) having voltage ratings of 1200 and 2500 volts are usually satisfactory for inputs varying from about 350 watts at 28 MHz to 1 kW at 14 MHz and lower. Because of these current limitations, particularly at the higher frequencies, it is advisable to use as large an air capacitor as practicable, using the micas only at the lower frequencies. Broadcast-receiver replacement-type capacitors can be obtained reasonably. Their voltage insulation should be adequate for inputs of 1000 watts or more.

TRANSISTOR OUTPUT CIRCUITS

Since rf power transistors have a low output impedance (on the order of 5 ohms or less), the problem of coupling the transistor to the usual 50-ohm load is the reverse of the problem with a vacuum-tube amplifier. The 50-ohm load must be transformed to a low resistance.

Figs 6-29A and B show two types of parallel-tuned circuits used to couple the load to the collector circuit. The collector is tapped down on the inductor in both cases. C1 provides tuning

Fig. 6-29 — Typical transistor output-matching networks.



for the collector and C2 adjusts the coupling to the load to achieve the proper impedance transformation. The use of the tapped connection to the inductor helps to maintain the loaded Q of the circuit while minimizing variations in tuning with changes in the junction capacitance of the transistor.

Circuits of Figs. 6-29C through E are not dependent upon coupling coefficient of a tapped coil for load-impedance transformation, making them more suitable for use at hf than either A or B. The collector-emitter capacitance (C_o) of the transistor is a major factor in the calculations used to design these circuits. Unfortunately C_o is not constant, so cut-and-try adjustments are usually necessary to optimize a particular circuit.

Early tests of transistor rf power amplifiers should be made with low voltage, a dummy load and no drive. Some form of output indicator should be included. When it has been established that no instability exists, the drive can be applied in increments and adjustment made for maximum output. The amplifier should never be operated at high voltage and no load.

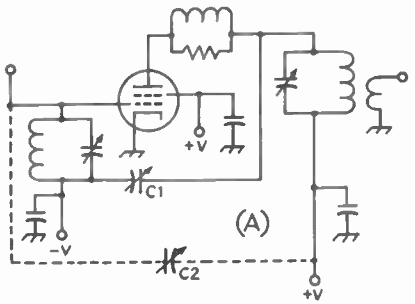
BROADBAND COUPLING

The techniques of broadband-transformer construction use transmission-line elements. A transformer consists of a short transmission line (one-eighth wavelength or less) made from a twisted-wire pair, coaxial or strip line, wound on a high-permeability toroid core to improve the low-frequency characteristics. At vhf the core may be omitted. Only discrete impedance transformations are possible; typical ratios are 9/4:1, 4:1, 9:1, 16:1, and 25:1. The higher ratios are difficult to achieve in practice, so several 4:1 transformers are employed for a large transformation ratio as shown in Fig. 6-23. Hybrid transformers, providing the 180-degree phase shift for input and output matching to push-pull stages, may also be made using broadband techniques.

Large toroid cores are not required for moderate power levels. A one-half inch diameter core is sufficient for operation at 100 watts at the low impedance levels found in transistor circuits. Because the current is high it is important to keep the resistance of the conductors low. Multiconductor leads (3 or 4 strands of No. 26 enam., twisted) or the flat enam. strip used for transformer windings) are suitable. Some typical designs are shown in Table 6-II.

STABILIZING AMPLIFIERS

A straight amplifier operates with its input and output circuits tuned to the same frequency. Therefore, unless the coupling between these two circuits is brought to the necessary minimum, the amplifier will oscillate as a tuned-plate tuned-grid circuit. Care should be used in arranging components and wiring of the two circuits so that there will be negligible opportunity for coupling external to the tube or transistor itself. Complete shielding between input and output circuits usually is required. All rf leads should be kept as short as possible and particular attention should be paid to the rf return paths from input and output tank circuits to emitter or cathode. In general, the best arrangement using a tube is one in which the cathode connection to ground, and the plate tank circuit are on the same side of the chassis or other shielding. The "hot" lead from the input tank (or driver plate tank) should be brought to the socket through a hole in the shielding. Then when the grid tank capacitor or bypass is grounded, a return path through the hole to cathode will be encouraged, since transmission-line characteristics are simulated.



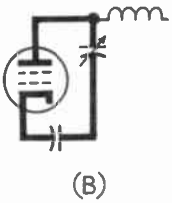


Fig. 6-30 – (A) A neutralizing scheme may use either C1 or C2 to cancel the effect of grid-to-plate capacitance in the tube (B) Vhf parasitic circuit shown with heavy lines.

A check on external coupling between input and output circuits can be made with a sensitive indicating device, such as the wavemeter shown in the Measurements chapter. The amplifying device is removed. With the driver stage running and tuned to resonance, the indicator should be coupled to the output tank coil and the output tank capacitor tuned for any indication of rf feedthrough. Experiment with shielding and rearrangement of parts will show whether the isolation can be improved. For additional information on transistor circuits see Chapter 4.

Screen-Grid Tube Neutralizing Circuits

The plate-grid capacitance of screen-grid tubes is reduced to a fraction of a picofarad by the interposed grounded screen. Nevertheless, the power sensitivity of these tubes is so great that only a very small amount of feedback is necessary to start oscillation. To assure a stable amplifier, it is usually necessary to load the grid circuit, or to use a neutralizing circuit.

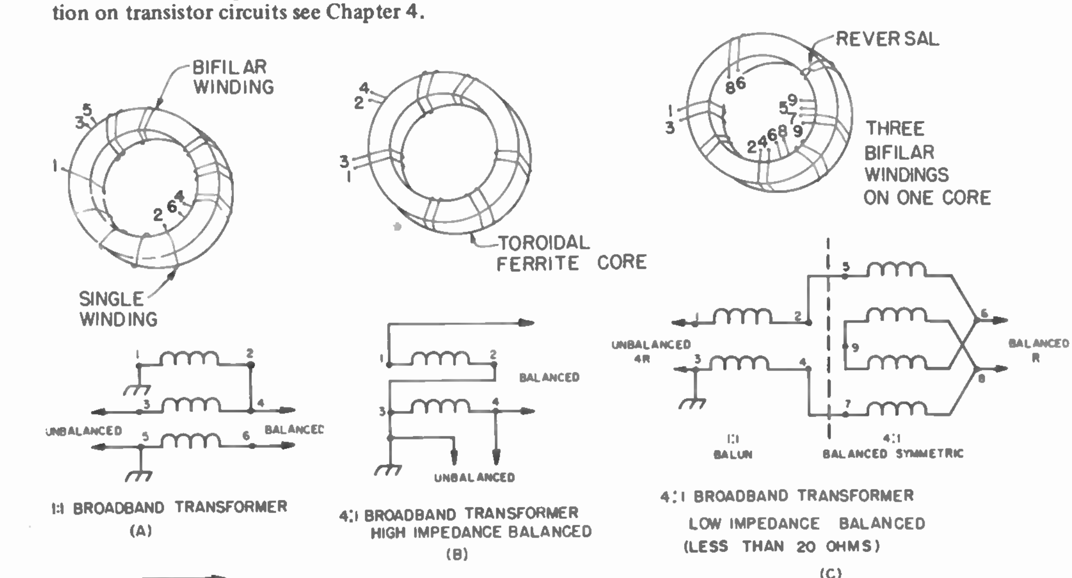
The capacitive neutralizing system for screen-grid tubes is shown in Fig. 6-30A. C1 is the neutralizing capacitor. The capacitance should be chosen so that at some adjustment of C1,

$$\frac{C1}{C3} = \frac{\textit{Tube grid-plate capacitance (or } C_{GP}\textit{)}}{\textit{Tube input capacitance (or } C_{IN}\textit{)}}$$

The grid-cathode capacitance must include all strays directly across the tube capacitance, including the capacitance of the tuning-capacitor stator to ground. This may amount to 5 to 20 pF. In the case of capacitance coupling, the output capacitance of the driver tube must be added to the grid-cathode capacitance of the amplifier in arriving at the value of C1.

Neutralizing a Screen-Grid Amplifier Stage

There are two general procedures available for indicating neutralization in a screen-grid amplifier stage. If the screen-grid tube is operated with or without grid current, a sensitive output indicator



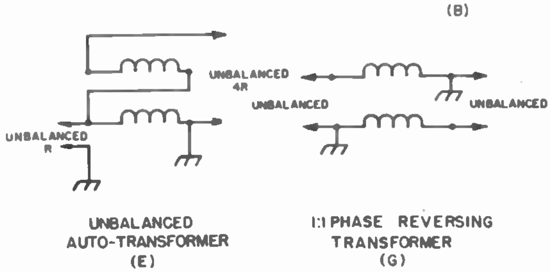


Table 6-11 – Basic broadband balun transformers. Bifilar windings are six to ten turns, depending on the ferrite-core permeability. A suitable ferrite material is $Q1$ with a permeability of 125. Very small size cores (1/4-to 3/4-inch OD) may be used for receiving and low-power applications. For full-power applications a 2-1/2-inch OD $Q1$ core with 1/2-inch cross section wound with No. 14 Formex copper wire, seven turns per winding, is recommended.

can be used. If the screen-grid tube is operated with grid current, the grid-current reading can be used as an indication of neutralization. When the output indicator is used, both screen and plate voltages must be removed from the tubes, but the dc circuits from the plate and screen to cathode must be completed. If the grid-current reading is used, the plate voltage may remain on but the screen voltage must be zero, with the dc circuit completed between screen and cathode.

The immediate objective of the neutralizing process is reducing to a minimum the rf driver voltage fed from the input of the amplifier to its output circuit through the grid-plate capacitance of the tube. This is done by adjusting carefully, bit by bit, the neutralizing capacitor or link coils until an rf indicator in the output circuit reads minimum, or the reaction of the unloaded plate-circuit tuning on the grid-current value is minimized.

The wavementer shown in the Measurements chapter makes a sensitive neutralizing indicator. The wavemeter coil should be coupled to the output tank coil at the low-potential or "ground" point. Care should be taken to make sure that the coupling is loose enough at all times to prevent buring out the meter or the rectifier. The plate tank capacitor should be readjusted for maximum reading after each change in neutralizing.

When the grid-current meter is used as a neutralizing indicator, the screen should be grounded for rf and dc, as mentioned above. There will be a change in grid current as the unloaded plate tank circuit is tuned through resonance. The neutralizing capacitor (or inductor) should be adjusted until this deflection is brought to a minimum. As a final adjustment, screen voltage should be returned and the neutralizing adjustment continued to the point where minimum plate current, maximum grid current and maximum screen current occur simultaneously. An increase in grid current when the plate tank circuit is tuned slightly on the high-frequency side of resonance indicates that the neutralizing capacitance is too small. If the increase is on the low-frequency side, the neutralizing capacitance is too large. When neutralization is complete, there should be a slight decrease in grid current on either side of resonance.

Grid Loading

The use of a neutralizing circuit may often be avoided by loading the grid circuit if the driving stage has some power capability to spare. Loading by tapping the grid down on the grid tank coil (or the plate tank coil of the driver in the case of capacitive coupling), or by a resistor from grid to cathode is effective in stabilizing an amplifier.

VHF Parasitic Oscillation

Parasitic oscillation in the vhf range will take place in almost every rf power amplifier. To test for vhf parasitic oscillation, the grid tank coil (or driver tank coil in the case of capacitive coupling) should be short-circuited with a clip lead. This is to prevent any possible t.g.t.p. oscillation at the operating frequency which might lead to confusion in identifying the parasitic. Any fixed bias should be replaced with a grid leak of 10,000 to 20,000 ohms. All load on the output of the amplifier should be disconnected. Plate and screen voltages should be reduced to the point where the rated dissipation is not exceeded. If a Variac is not available, voltage may be reduced by a 117-volt lamp in series with the primary of the plate transformer.

With power applied only to the amplifier under test, a search should be made by adjusting the input capacitor to several settings, including minimum and maximum, and turning the plate capacitor through its range for each of the grid-capacitor settings. Any grid current, or any dip or flicker in plate current at any point, indicates oscillation. This can be confirmed by an indicating absorption wavemeter tuned to the frequency of the parasitic and held close to the plate lead of the tube.

The heavy lines of Fig. 6-30B show the usual parasitic tank circuit, which resonates, in most cases, between 100 and 200 MHz. For each type of tetrode, there is a region, usually below the parasitic frequency, in which the tube will be self-neutralized. By adding the right amount of inductance to the parasitic circuit, its resonant frequency can be brought down to the frequency at which the tube is self-neutralized. However, the resonant frequency should not be brought down so low that it falls close to TV Channel 6 (88 MHz). From the consideration of TVI, the circuit may be loaded down to a frequency not lower than 100 MHz. If the self-neutralizing frequency is below 100 MHz, the circuit should be loaded down to somewhere between 100 and 120 MHz with inductance. Then the parasitic can be suppressed by loading with resistance. A coil of 4 or 5 turns, 1/4 inch in diameter, is a good starting size. With the tank capacitor turned to maximum capacitance, the circuit should be checked with a GDO to make sure the resonance is above 100 MHz. Then, with the shortest possible leads, a noninductive 100-ohm 1-watt resistor should be connected across the entire coil. The amplifier should be tuned up to its highest-frequency band and operated at low voltage. The tap should be moved a little at a time to find the minimum number of turns required to suppress the parasitic. Then voltage should be increased until the resistor begins to feel warm after several minutes of operation, and the power input noted. This input should be compared with the normal input and the *power* rating of the resistor increased by this proportion; i.e., if the power is half normal, the wattage rating should be doubled. This increase is best made by connecting 1-watt *carbon* resistors in parallel to give a resultant of about 100 ohms. Or, one of the Globar surge-protection resistors may be used. As power input is increased, the parasitic may start up again, so power should be applied only momentarily until it is made certain that the parasitic is still suppressed. If the parasitic starts up again when voltage is raised, the tap must be moved to include more turns. So long as the parasitic is suppressed, the resistors will heat up only from the operating-frequency current. In grounded-grid

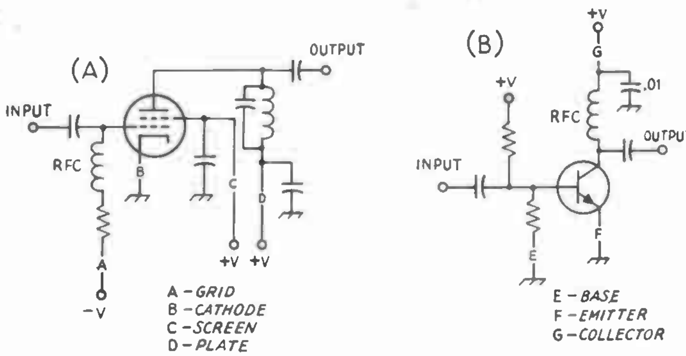


Fig. 6-31 – Metering circuits for (A) tubes and (B) transistors. To measure current, connect a meter at the point shown *in series with the lead.* For voltage measurements, connect the meter from the point indicated *to the common or ground connection.*

circuits it is useful to locate the parasitic suppressor in the cathode lead, as the rf power level is less than at the plate terminal.

Since the resistor can be placed across only that portion of the parasitic circuit represented by L_p, the latter should form as large a portion of the circuit as possible. Therefore, the tank and bypass capacitors should have the lowest possible inductance and the leads shown in heavy lines should be as short as possible and of the heaviest pratical conductor. This will permit L_p to be of maximum size without tuning the circuit below the 100-MHz limit.

Another arrangement that has been used successfully in transistor and low-level tube stages is to place one or more ferrite beads over the input or output leads, as close as possible to the amplifying device. The beads have sufficient low-*Q* inductance at vhf to discourage any tendency toward parasitic oscillation.

Low-Frequency Parasitic Oscillation

The screening of most transmitting screen-grid tubes is sufficient to prevent low-frequency parasitic oscillation caused by resonant circuits set up by rf chokes in grid and plate circuits. When rf chokes are used in both grid and plate circuits of a triode amplifier, the split-stator tank capacitors combine with the rf chokes to form a low-frequency parasitic circuit, unless the amplifier circuit is arranged to prevent it. Often, a resistor is substituted for the grid rf choke, which will produce the desired result. This resistance should be at least 100 ohms. If any grid-leak resistance is used for biasing, it should be substituted for the 100-ohm resistor.

Transistor LF Parasitics

Using transistors with shunt feed often means low-frequency parasitic trouble. A word about this problem is in order as it usually doesn't occur in vacuum-tube circuits and is often a rough problem for the newcomer to solid state. These parasitics manifest themselves as a wide spectrum of white noise (hash) around and below the operating frequency. They can often be heard on a broadcast receiver several feet away from a transmitter under test. The desired signal may sound clean, so it is necessary to check far below the operating frequency. Two transistor characteristics combine to cause this trouble. First, transistors have higher gain at lower frequencies than they do at hf. Second, interelement capacitances vary over a wide range of changes in voltage, the result being varactor action that causes spurious outputs. The best way to avoid the problem is to use a minimum of inductance in the collector circuit. Large chokes are unsatisfactory. Series feed is a good answer as no choke is needed. Bypass capacitors should be the minimum value required. Decoupling on power leads between stages should have at least two capacitors, one effective at the operating frequency and a second large capacitor that is good at low frequencies.

METERING

Fig. 6-31 shows how a voltmeter and milliammeter should be connected to read various voltages and currents. Voltmeters are seldom installed permanently, since their principal use is in preliminary checking. Also, milliammeters are not normally installed permanently in all of the positions shown. Those most often used are the ones reading grid current and plate current, or grid current and cathode current, or collector current.

Milliammeters come in various current ranges. Current values to be expected can be taken from the tube tables and the meter ranges selected accordingly. To take care of normal overloads and pointer swing, a meter having a current range of about twice the normal current to be expected should be selected.

Grid-current meters connected as shown in Fig. 6-31 and meters connected in the cathode circuit need no special precautions in mounting on the transmitter panel so far as safety is concerned. However, milliammeters having metal zero-adjusting screws on the face of the meter should be recessed behind the panel so that accidental contact with the adjusting screw is not possible, if the meter is connected in any of the other positions shown in Fig. 6-31. The meter can be mounted on a small subpanel attached to the front panel with long screws and spacers. The meter opening should be covered with glass or celluloid. Illuminated meters make reading easier. Reference should also be made to the TVI chapter of this *Handbook* in regard to wiring and shielding of meters to suppress TVI.

COMPONENT RATINGS

Output Tank Capacitor Voltage

In selecting a tank capacitor with a spacing between plates sufficient to prevent voltage breakdown, the peak rf voltage across a tank circuit under load, but without modulation, may be taken conservatively as equal to the dc plate or collector voltage. If the dc supply voltage also appears across the tank capacitor, this must be added to the peak rf voltage, making the total peak voltage twice the dc supply voltage. If the amplifier is to be plate-modulated, this last value must be doubled to make it four times the dc plate voltage, because both dc and rf voltages double with 100-percent amplitude modulation. At the higher voltages, it is desirable to choose a tank circuit in which the dc and modulation voltages do not appear across the tank capacitor, to permit the use of a smaller capacitor with less plate spacing.

Capacitor manufacturers usually rate their products in terms of the peak voltage between plates. Typical plate spacings are shown in the following table, 6-III.

Output tank capacitors should be mounted as close to the tube as temperature considerations will permit, to make possible the shortest capacitive path from plate to cathode. Especially at the higher frequencies where minimum circuit capacitance becomes important, the capacitor should be mounted with its stator plates well spaced from the chassis or other shielding. In circuits where the rotor must be insulated from ground, the capacitor should be mounted on ceramic insulators of size commensurate with the plate voltage involved and – most important of all, from the viewpoint of safety to the operator – a well-insulated coupling should be used between the capacitor shaft and the dial. *The section of the shaft attached to the dial should be well grounded.* This can be done conveniently through the use of panel shaft-bearing units.

Table 6-III

Typical Tank-Capacitor Plate Spacings

Spacing (In.)	*Peak Voltage*	*Spacing (In.)*	*Peak Voltage*	*Spacing (In.)*	*Peak Voltage*
0.015	1000	0.07	3000	0.175	7000
0.02	1200	0.08	3500	0.25	9000
0.03	1500	0.125	4500	0.35	11000
0.05	2000	0.15	6000	0.5	13000

Tank Coils

Tank coils should be mounted at least their diameter away from shielding to prevent a marked loss in Q. Except perhaps at 28 MHz it is not important that the coil be mounted quite close to the tank capacitor. Leads up to 6 or 8 inches are permissible. It is more important to keep the tank capacitor as well as other components out of the immediate field of the coil. For this reason, it is preferable to mount the coil so that its axis is parallel to the capacitor shaft, either alongside the capacitor or above it.

Wire Sizes for Transmitting Coils for Tube Transmitters

Power Input (Watts)	*Band (MHz)*	*Wire Size*
1000	28–21	6
	14–7	8
	3.5–1.8	10
500	28–21	8
	14–7	12
	3.5–1.8	14
150	28–21	12
	14–7	14
	3.5–1.8	18
75	28–21	14
	14–7	18
	3.5–1.8	22
25 or less*	28–21	18
	14–7	24
	3.5–1.8	28

* Wire size limited principally by consideration of Q.

There are many factors that must be taken into consideration in determining the size of wire that should be used in winding a tank coil. The considerations of form factor and wire size that will produce a coil of minimum loss are often of less importance in practice than the coil size that will fit into available space or that will handle the required power without excessive heating. This is particularly true in the case of screen-grid tubes where the relatively small driving power required can be easily obtained even if the losses in the driver are quite high. It may be considered preferable to take the power loss if the physical size of the exciter can be kept down by making the coils small.

Transistor output circuits operate at relatively low impedances because the current is quite high. Coils should be made of heavy wire or strap, with connections made for the lowest possible resistance. At vhf stripline techniques are often employed, as the small inductance values required for a lumped inductance become difficult to fabricate.

RF Chokes

The characteristics of any rf choke will vary with frequency, from characteristics resembling those of a parallel-resonant circuit, of high impedance, to those of a series-resonant circuit, where the impedance is lowest. In between these extremes, the choke will show varying amounts of inductive or capacitive reactance.

In series-feed circuits, these characteristics are of relatively small importance because the rf voltage across the choke is negligible. In a parallel-feed circuit, however, the choke is shunted across the tank circuit, and is subject to the full tank rf voltage. If the choke does not present a sufficiently high impedance, enough power will be abosrbed by the choke to cause it to burn out.

To avoid this, the choke must have a sufficiently high reactance to be effective at the lowest frequency, and yet have no series resonances near the higher-frequency bands.

THE K1ZJH SOLID-STATE TRANSCEIVER

Construction details for an all-band solid-state transceiver are presented here for the more adventurous builders. Beginners are not encouraged to attempt duplicating the circuits of this advanced project. Printed-circuit templates or layouts are *not* available. The transceiver has many features making it suited to fit a variety of amateur needs. Five-hundred kHz coverage of the 80-, 40-, 20- and 15-meter bands is provided, as well as 500 kHz of the 10-meter band (28.5 to 29.0 MHz). Cw (peak ssb) power output is 20 watts on 80, 40 and 20 meters, dropping to a maximum of 9 watts on 10 meters. Cw operation was included, and the ssb VOX circuit is combined with the cw sidetone generator to allow semi-break-in operation. Wherever possible, FET and IC circuitry was employed.

In keeping with the state-of-the art, an electronic dial, or frequency counter, accurately counts and displays the operating frequency of the transceiver. Because the counter derives its final count as a product of the oscillators and the BFO, it will be as accurate as its time base allows. A more detailed description of counter operation will be given later.

Because portability was desired, the complete transceiver, including speaker and power supply, was assembled in an LMB CO-1 enclosure. Although the author's unit (K1ZJH) was built for operation at 117 volts, the 14.5-volt circuitry makes battery operation possible.

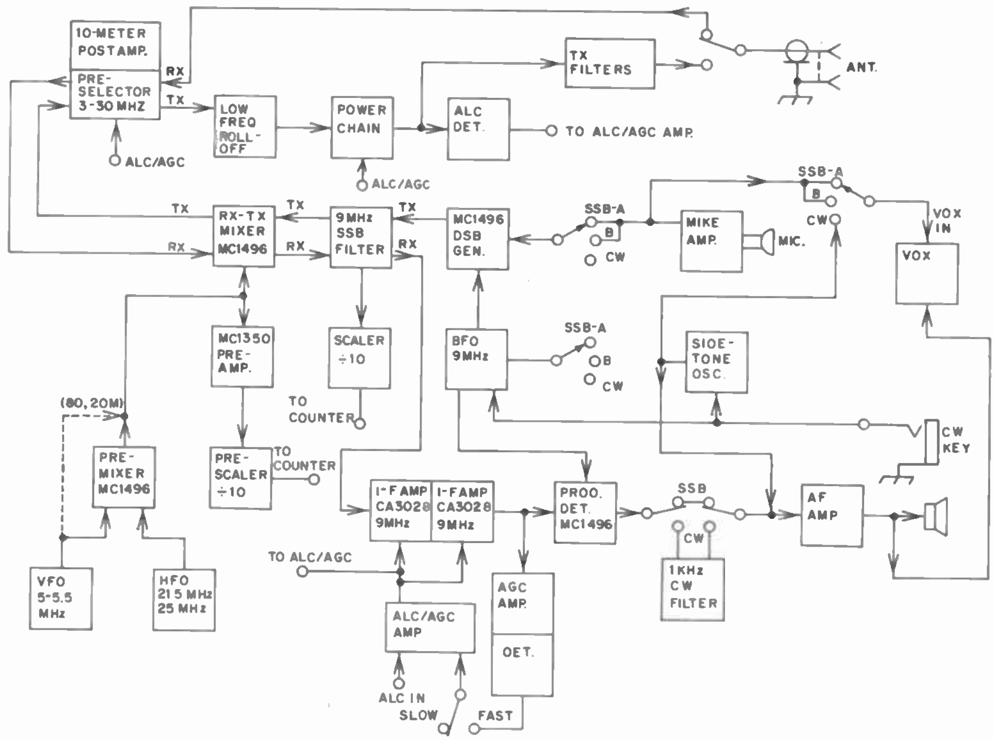
Design Notes

Without a doubt one of the larger problems plaguing the designer of a multiband transceiver is the generation of the local oscillator signal. One approach is to switch bands with the VFO, hoping it will remain stable over wide ranges of operating parameters. Other alternatives are a multiconversion arrangement involving the use of a combination of i-f ranges, or using a premixing scheme which directly synthesizes the mixer injection by combining the VFO with other oscillators. Single conversion and one VFO range (the premixing method) is used here.

The use of the popular 5- to 5.5-MHz VFO and 9-MHz i-f allows 80- and 20-meter coverage to be produced without premixing. Forty-meter operation is accomplished by premixing the 5-MHz VFO output with the 21.5-MHz heterodyne oscillator to generate a mixer injection signal at 16.5 to 16.0 MHz. When combined with 9 MHz, 40-meter operation is obtained. The remaining bands, 10 and 15 meters, are produced with a 25-MHz HFO. On 15 meters, the VFO is mixed with the 25-MHz HFO to produce a 30.0 to 30.5-MHz premixer output. When mixed with the 9-MHz i-f, the result is the 21.0 to 21.5-MHz amateur band. For 10 meters, the VFO is mixed with the 25-MHz HFO, and the 20.0 to 19.5-MHz product is selected from the premixer output. Combined with the 9-MHz i-f, coverage of the 29 to 28.5-MHz portion of the 10-meter band is provided. This premixing scheme produces a few minor operational quirks. Sideband inversion or VFO tuning inversion occurs on some bands because of down conversion in one of the mixer stages. For this reason, the sideband positions on the mode switch are marked A and B. The variable up or down premixing conversion does not make it practical to offset the VFO to compensate for changes in frequency when sidebands are changed. This would result in an error of a few kHz when using a mechanical dial arrangement, but the counter quickly displays the new frequency. The counter is programmed to correct for VFO tuning inversion on those bands where it occurs.

The VFO and HFO

An MPF-102 JFET Colpitts oscillator is the heart of the VFO. The gate of the MPF-102 is diode clamped to minimize harmonic generation on positive voltage peaks. Two transistor buffer stages follow to reduce the effects of loading and to provide an amplified low-impedance VFO output. Receiver offset tuning which is provided by a diode across the main-tuning capacitor allows up to a 3-kHz offset during receive. Voltage for the MPF-102 is regulated. The Zener will dissipate heat and should not be located in close proximity to frequency determining components. Main-tuning capacitor, C1, is a rather costly item and was used because one was made available for this project; it is one of the better capacitors available for VFO service. The large value of fixed-input capacitance associated with the Colpitts oscillator swamped C1 and it was not possible to obtain full 500-kHz VFO range with the *LC* ratio. L2 serves to expand the frequency coverage available with the 35-pF capacitor to a little over 500 kHz. The VFO tuning range should be adjusted to give about 20 kHz of overlap on band edges. The slug of L1 will largely determine total VFO spread, and trimmer capacitor C_a (45 pF across C1) is used to set the VFO frequency. Both adjustments will interact.



Block diagram for the K1ZJH transceiver. The counter circuit is not included.

The VFO circuit board and associated VFO components were mounted on a 1/8-inch thick aluminum plate for mechanical rigidity. The gear drive mechanism (not available commercially) has a gear ratio of 16:1 providing about 33 kHz per dial revolution. This is about the fastest tuning rate desirable for comfortable tuning of ssb signals.

The internal speaker is located within the VFO enclosure (just below the VOX amplifier). This is a compromise necessitated because of tight cabinet space. Fortunately the VFO is not microphonic, and when possible, a larger and better sounding external speaker is used. The photographs reveal the counter board which is located directly beneath the VFO. During operation, the components on this board get warm and radiate considerable heat. However, after about a 15-minute warm-up period, thermal equilibrium is reached, and VFO drift stabilizes to less than 100 Hz per hour.

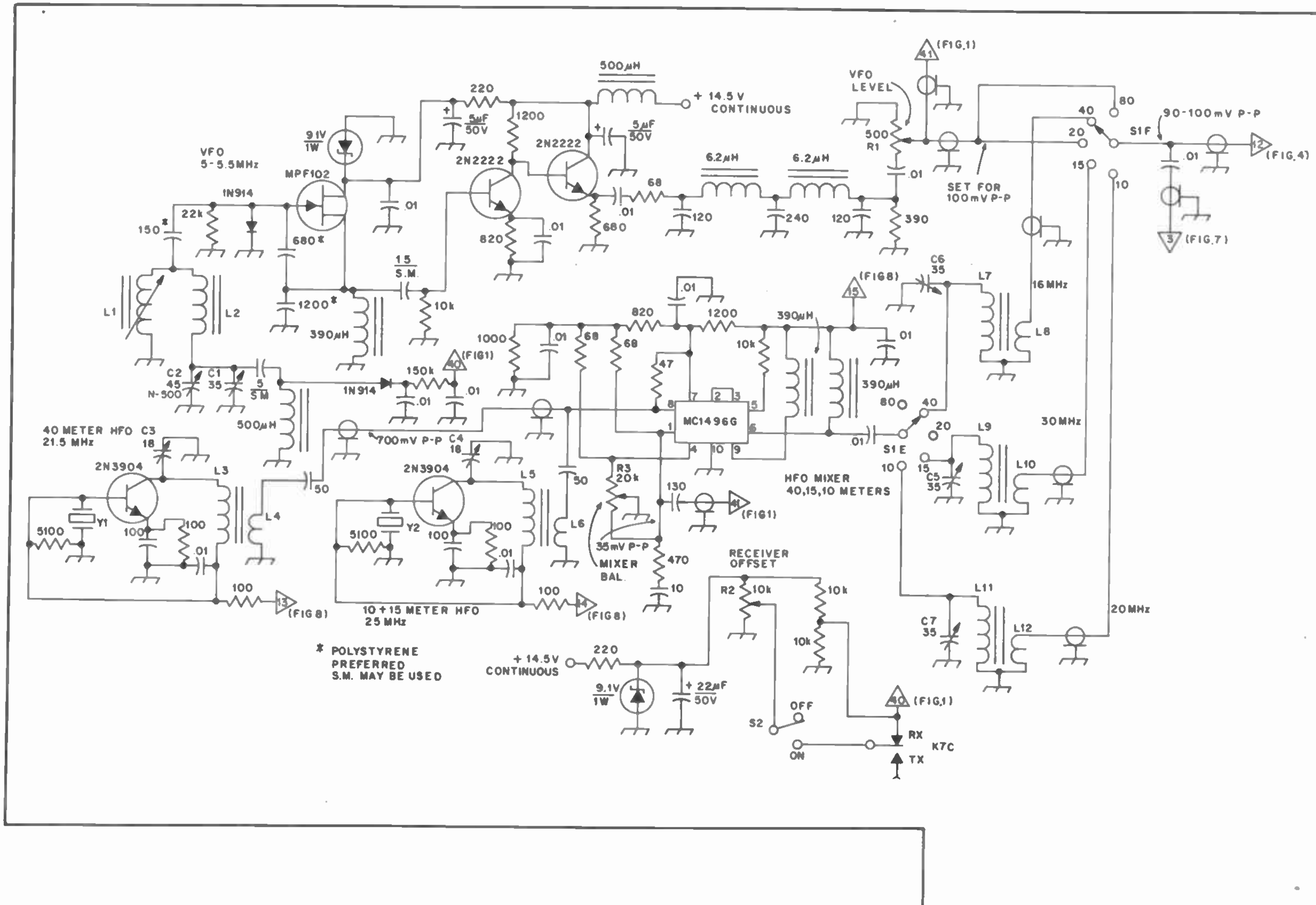
Polystyrene capacitors should be used where indicated. These have a better tolerance to temperature variations (either from external sources or internal rf heating) than their silver mica counterparts. Mica capacitors can become leaky with age because of dielectric deterioration, a bane to stable VFO operation. Drifting can be reduced by selecting the "N" coefficient of trimmer C_a. An N-500 capacitor was optimum compensation for this unit.

Any harmonics generated in the VFO buffer stages are reduced in a half-wave filter. The VFO output level is adjustable by a 500-ohm composition trimmer, and should be set for 100 millivolts peak to peak to supply proper injection levels for the MC1496 mixers. On 20 and 80 meters, the VFO directly feeds the MC1496 balanced mixer. On 10, 15 and 40 meters, the VFO output is premixed with one of the HFO oscillators in the MC1496 premixer to produce the appropriate injection range for the mixer.

The HFO oscillators are used only on 40, 15 and 10 meters. Only one HFO oscillator is active at any one time; the operating voltage for the appropriate HFO is supplied through band-switch section nine. On 40 meters, the 21.5-MHz oscillator is active, while on 15 and 10 meters the 25-MHz oscillator is used. Except for the operating frequency, both oscillator circuits are identical. The HFO circuit board is mounted on the rear VFO partition. The crystals used in the HFO are inexpensive, general purpose (third overtone) types supplied by International Crystals.

The HFO Premixer and Preselector

The doubly-balanced design gives adequate isolation of the input signals and only one tuned



VFO 5-5.5MHz
MPF102
1N914
9.1V 1W
2N2222
2N2222
500μH
+14.5V CONTINUOUS
6.2μH
6.2μH
VFO LEVEL
500 R1
SET FOR 100mV P-P
90-100mV P-P
S1F
(FIG.1)
(FIG.4)
(FIG.7)
(FIG.8)
L1
L2
C2 45 N-500
C1 35
390μH
15 S.M.
150k
1N914
40 METER HFO 21.5 MHz
C3 18
2N3904
Y1
5100
L3
L4
700mV P-P
C4 18
2N3904
Y2
L5
L6
10+15 METER HFO 25 MHz
MC1496G
R3 20k
35mV P-P MIXER BAL.
HFO MIXER 40,15,10 METERS
S1E
C6 35
L7
L8
16 MHz
30MHz
L9
L10
C5 35
L11
L12
C7 35
20MHz
RECEIVER OFFSET
R2 10k
S2
OFF
ON
RX
TX
K7C
* POLYSTYRENE PREFERRED S.M. MAY BE USED
+14.5V CONTINUOUS
9.1V 1W
22μF 50V

Fig. 1 — Circuit diagram for the transceiver VFO, HFO, and mixer.
C1 — 35-pF variable (Millen 28035MKBB or equiv.).
L1 — 24 turns No. 26 enam. on 1/2-inch ceramic form tuned with a red slug.
L2 — 9 turns No. 20 enam. on T-50-2 core.
L3, L5 — 18 turns No. 24 enam. on T-50-2 toroid core.
L4, L6 — 2 turns No. 24 enam. over L3 and L5 respectively.
L7 — 19 turns No. 24 enam. on T-50-2 toroid core.
L8 — 2 turns No. 24 enam. over L7.
L9 — 9 turns No.24 enam. on T-50-2 toroid core.
L10 — 2 turns No. 24 enam. over L9.
L11 — 14 turns No. 24 enam. on T-50-2 toroid core.
L12 — 2 turns No. 24 enam. over L11.
VR1 — silicon, 1000 PRV, 1 A.
Y1 — 21.5 MHz third-overtone general-purpose.
Y2 — 25.0 MHz third-overtone general-purpose.

stage is needed for each premixer output. The necessary 500-kHz pass band is easily obtained. Voltage for the premixer is supplied through band-switch section nine only on 40, 15 and 10 meters. The premixer is mounted in the VFO enclosure on the partition shield parallel to the bandswitch. HFO injection measured on pin 8 of the MC1496 premixer should be to 700 millivolts p-p. VFO injection supplied to pin one should be in the area of 35 millivolts p-p; if not, the 130-pF coupling capacitor can be changed accordingly. Injection levels to both MC1496 mixer stages are critical. When using higher injection levels spurious, raspy buzzsaw sounding signals will be heard randomly as the receiver is tuned across the bands. Motorola mentions likelihood of self oscillation when the devices are driven from a high-impedance source. Low-impedance drive sources are used in the transceiver. Careful regulation of the maximum mixer injection levels seems to be the best safeguard for stable MC1496 mixer operation.

Band-switch section five selects the correct premixer *LC* output network. These are mounted on the premixer pc board, and leads between the bandswitch and premixer should be short and direct. The low-impedance outputs are brought to band-switch section six in a braided shield. Section six selects the proper injection signal for the receiver-transmitter mixer and the counter input shaping circuit.

The preselector circuitry is located near the front of the transceiver, adjacent to the VFO compartment. It is shielded on three sides by the front panel, a VFO compartment partition, and the rear partition separating the preselector and receiver-transmitter mixer compartments. The preselector circuit is fairly conventional using two MPF-102 FETs in cascade configuration. Gate bias is under alc/agc control. Resistive loading of the 80 and 40 input links reduces excessive and unneeded gain on these two bands. Band-switch sections one, two, three and four are dedicated to preselector coil switching. Section two should be shielded from section three to isolate the input and output stages of the preselector. The shield is grounded directly to the adjacent shield partition separating the VFO and preselector compartments. Shielding fabricated from thin copper flashing on the preselector pc board further isolates the input and output circuitry. The center shield of the preselector tuning capacitor is grounded with a short section of braid to a pc board shield directly beneath it. The higher frequency coils were placed closest to the band switch permitting short lead lengths.

Both input and output impedances of the preselector are approximately 50-70 ohms. Relays K1 and K2 switch the preselector between the receiver and transmitter portions of the transceiver. The author acquired a handful of these miniature relays, and they are used in several areas of the transceiver to switch rf. The use of relay switching may seem archaic in a modern transceiver but these little relays are quiet, offer a good degree of switching isolation, readily switch low-impedance rf with minimum loss, and are available surplus at reasonable prices. On receive, the preselector input comes from the antenna relay through a rf attenuator gain control. This method of gain control is not only effective, but does not affect the total dynamic agc range of the receiver. On 80 and 40 meters the gain and noise figure of the preselector is of little practical value and the rf-gain control will likely be run with almost full attenuation on these bands to allow for good, strong-signal handling characteristics in the receiver. The preselector output in receive goes to the receiver-transmitter mixer stage for conversion to the 9-MHz receiver i-f. In transmit, the preselector input is switched to the receiver-transmitter mixer output. The preselector output, through an attenuator rf-drive-level control, goes to the transmitter power amplifier chain during transmit.

10-Meter Post Amplifier and Receiver-Transmitter Mixer

It was found that both the preselector and transmitter power chain suffered from a lack of pep on ten meters. To improve receiver and transmitter performance on 10 meters, a 28-MHz post amplifier was incorporated. The post amplifier, a MFE-121 dual-gate MOSFET, like the preselector, is under alc/agc control. Operating voltage is supplied to the post amplifier through band-switch section nine on ten meters. Powering the amplifier energizes relay K2 placing the amplifier in line with the low-impedance output of the preselector. The three tuned stages of the post amplifier offer a large degree of selectivity and require careful stagger tuning to obtain uniform gain across the 10-meter band. Stability of the amplifier was achieved by stagger tuning and by resistive loading of the input stage. Placement and shielding of the amplifier stage are critical.

One mixer does the necessary conversion functions for both the receiver and transmitter sections of the transceiver. Another Motorola MC1496 IC is

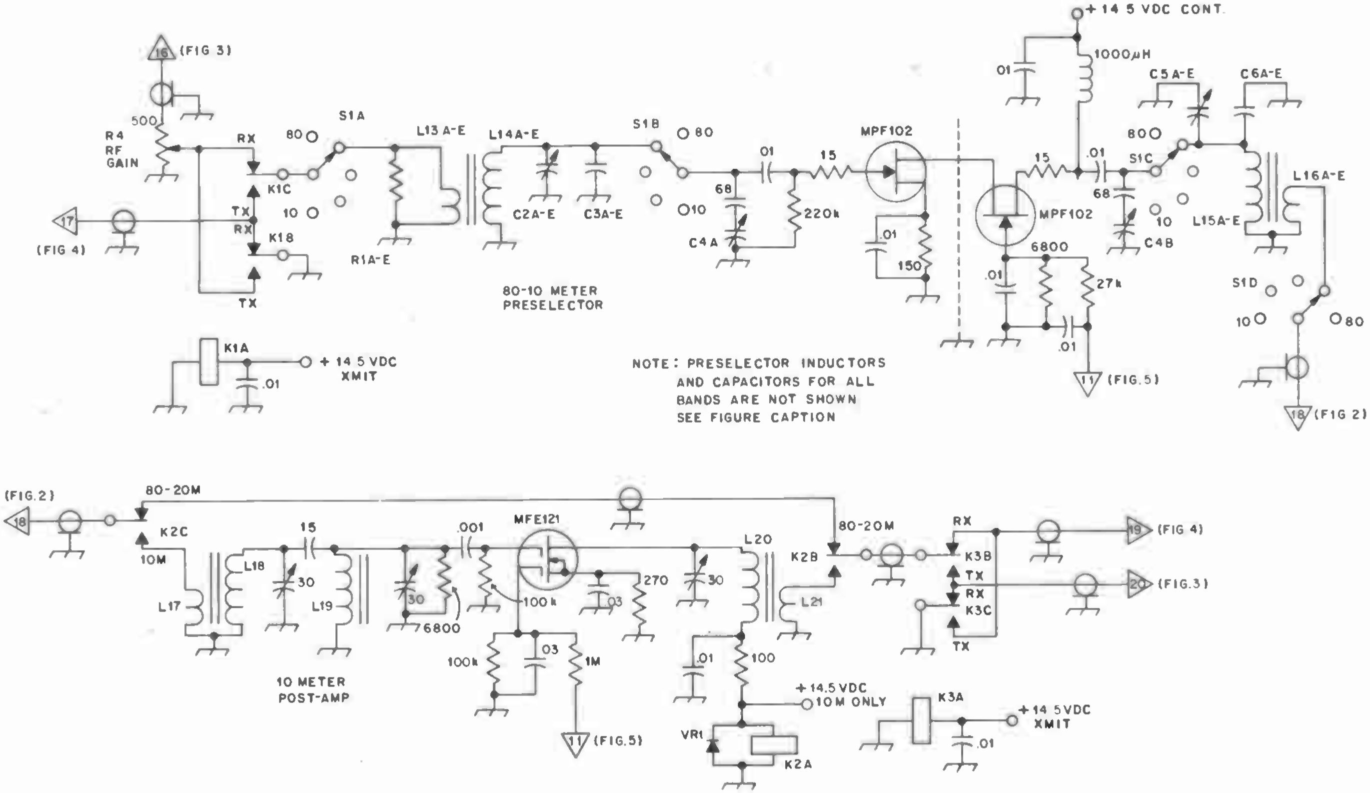

+14.5 VDC CONT.
1000μH
(FIG 3)
R4
RF
GAIN
500
S1A
RX
80
K1C
TX
10
K1B
(FIG 4)
L13 A-E
L14A-E
R1A-E
C2A-E
C3A-E
S1B
68
C4A
220k
MPF102
15
.01
150
C5A-E
C6A-E
S1C
68
C4B
L15A-E
L16A-E
6800
27k
(FIG.5)
S1D
(FIG 2)
80-10 METER
PRESELECTOR
K1A
+ 14.5 VDC
XMIT
NOTE: PRESELECTOR INDUCTORS
AND CAPACITORS FOR ALL
BANDS ARE NOT SHOWN
SEE FIGURE CAPTION
(FIG.2)
80-20M
K2C
10M
L17
L18
30
L19
6800
.001
MFE121
100k
.03
270
1M
L20
L21
K2B
100
+14.5 VDC
10M ONLY
VR1
K2A
10 METER
POST-AMP
RX
K3B
TX
K3C
(FIG.3)
K3A
+14.5VDC
XMIT

Fig. 2 — Circuit diagram for the transceiver preselector and 10-meter post amplifier. Preselector inductors and capacitors are assigned a suffix A for 80 meters, B for 40 meters, C for 20 meters, D for 15 meters, and E for 10 meters. R1A is 27 ohms and R1B is 47 ohms. R1C, R1D, and R1E are not used. Toroid cores are from Amidon Assoc.

C2A-E, C5A-E — 45-pF trimmer.
C3A, C6A — not used.
C3B, C6B — 120 pF.
C3C, C6C — 62 pF.
C3D, C3E, C6D, C6E — not used.
C4 — 100-pF per section dual (Millen 26100RM or equiv.).
K1, K2, K3 — 12 V dc dpdt.
L13A — 4 turns, No. 24 enam.
L13B — 3 turns, No. 24 enam.
L13C — 1 turn, No. 24 enam.
L13D — 1 turn, No. 24 enam.
L13E — 1 turn, No. 24 enam.
L14A, L15A — 70 turns, No. 26 enam. on a T-50-2 toroid core.
L14B, L15B — 23 turns, No. 24 enam. on a T-50-2 core.
L14C, L15C — 12 turns, No. 24 enam. on a T-50-2 core.
L14D, L15D — 10 turns, No. 22 enam. on a T-50-2 core.
L14E, L15E — 10 turns, No. 22 enam. on a T-50-2 core.
L16A — 33 turns, No. 24 enam.
L16B — 3 turns, No. 24 enam.
L16C — 3 turns, No. 24 enam.
L16D — 1 turn, No. 24 enam.
L16E — 1 turn, No. 24 enam.
L17 — 2 turns, No. 26 enam. wound over L18.
L18 — 16 turns, No. 24 enam. wound on a T-23-2 core.
L19 — 15 turns, No. 24 enam. wound on a T-37-2 core.
L20 — 19 turns, No. 24 enam. wound on a T-37-2 core.
L21 — 3 turns, No. 26 enam. wound on L20.
VR1 — 100 PRV, 1 A.

used in the receiver-transmitter mixer, and the circuit is quite similar to the premixer stage. Like the preselector, miniature hermetic relays are used to switch the mixer between receive and transmit functions. Injection levels to the receiver-transmitter mixer are critical. Ninety millivolts p-p should be measured at pin 4 of the MC1496 mixer.

In receive, the mixer output is matched to the 560-ohm crystal-filter impedance with a pi-network. During transmit, individual fixed-tuned pi networks for each of the five bands are selected by band-switch sections seven and eight to couple and match the receiver-transmitter mixer output to the preselector. The pi networks also offer selectivity and rejection to out-of-band signals.

Because individual wiring techniques and parts tolerances vary, the values for coils L24A-E and capacitors C12A-E are approximate. Ceramic or mica trimmers may be substituted for the fixed values specified for C12A-E. Note that these capacitors and toroidal coils L24A-E were mounted on, and supported by, band-switch sections seven and eight.

I-F Crystal Filter

The Essel M9A filter is used with good results in the transceiver. Because the filter directly affects the receiver selectivity, transmitter bandwidth and carrier suppression, a good quality filter is mandatory. KVG, Swan and others offer ssb filters that would be suitable for use in this transceiver. Diode switching selects the proper filter paths for transmitter or receiver operation. Filter ports Rxa and Rxb are associated with the receiver; likewise, ports Txa and Txb are used with the transmitter circuitry. Modern eight-pole filters are capable of more than 100 dB of stop-band attenuation and if they are to be used to full advantage, care is required to maintain good isolation between filter ports.

The I-F Stages, Product Detector and AGC/ALC

The i-f stages, product detector and agc/alc circuits are contained on one pc-board assembly. The i-f stages are under alc/agc control; agc voltage, applied to pin seven of the CA3028A i-f amplifiers varies, from two volts for minimum gain to 12 volts during periods of no agc action. The two i-f stages can produce over 50 dB of gain and, without adequate shielding and power supply decoupling between stages, instability can develop. Toroidal cores in the tuned i-f stages reduce the chance for unwanted interstage coupling. The turns ratio of input transformer L25-L26 performs impedance matching between the filter and i-f stages. Capacitive dividers perform impedance transformations between the i-f stages and to the product detector. The output of the last i-f stage is heavily loaded to prevent overdriving the product detector and to help stabilize the i-f stages. The MC1496 product detector has a dynamic range of 90 dB and a 12-dB conversion gain. Output filtering is simplified as the double balanced design reduces the level of the i-f signal or BFO energy appearing in the output. Two hundred millivolts p-p of BFO energy are

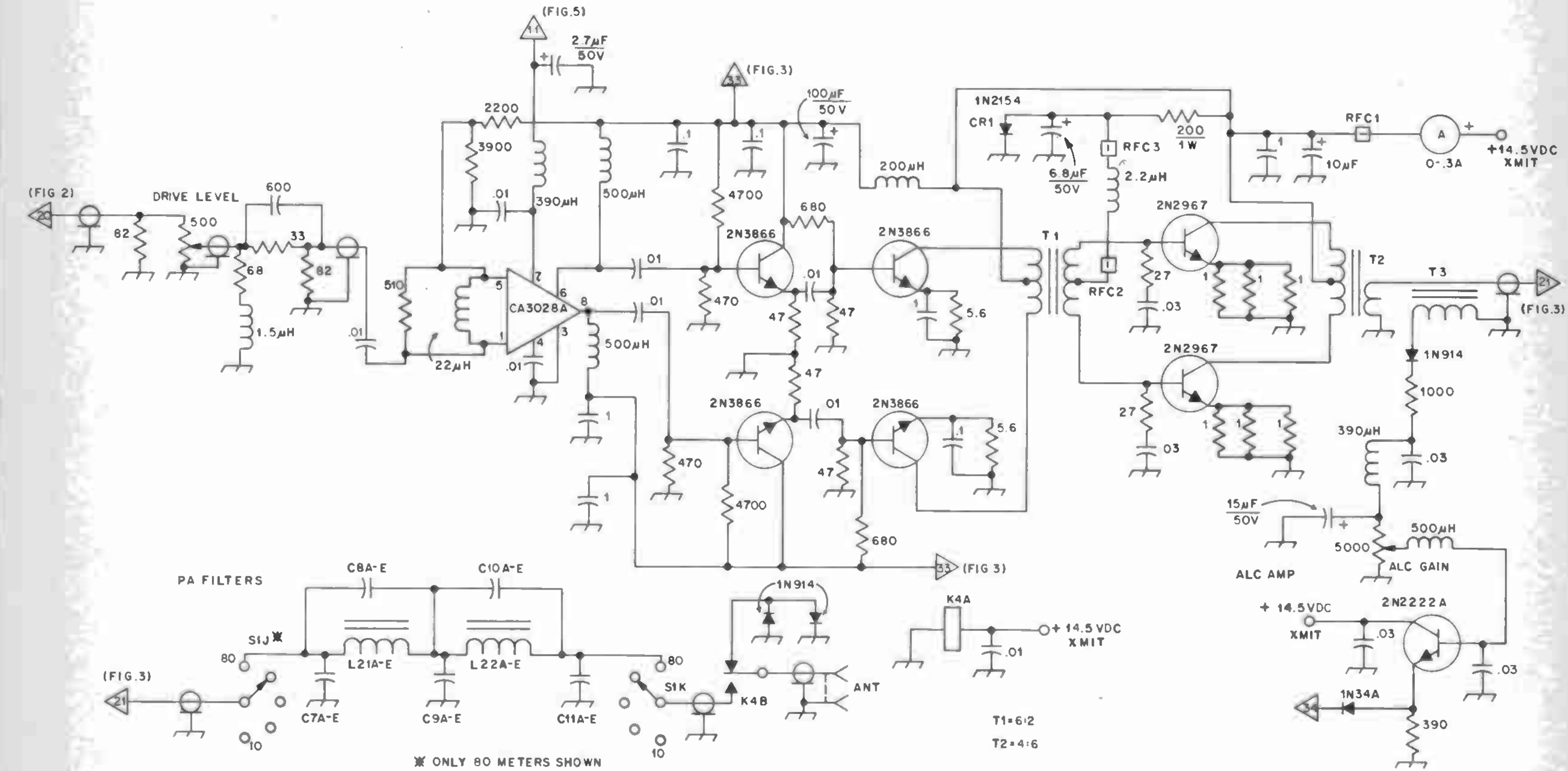

(FIG.5)
2.7µF 50V
(FIG.3)
100µF 50V
1N2154
CR1
6.8µF 50V
200 1W
RFC1
RFC3
2.2µH
A
0-.3A
+14.5VDC XMIT
10µF
2200
3900
.01
390µH
500µH
4700
680
2N3866
200µH
2N2967
T1
T2
T3
(FIG 2)
DRIVE LEVEL
600
500
82
33
68
1.5µH
510
22µH
CA3028A
470
47
5.6
RFC2
27
.03
(FIG.3)
1N914
1000
390µH
15µF 50V
5000
500µH
ALC GAIN
ALC AMP
2N2222A
+ 14.5VDC XMIT
1N34A
390
(FIG 3)
PA FILTERS
C8A-E
C10A-E
S1J
L21A-E
L22A-E
C7A-E
C9A-E
C11A-E
80
10
S1K
K4B
ANT
K4A
+ 14.5 VDC XMIT
T1=6:2
T2=4:6
ONLY 80 METERS SHOWN

Fig. 3 — Circuit diagram for the transmitter power chain. See Table I for capacitor values.
CR1 — 1N2154
K4 — 12 V dc, spdt.
L21A — 18 turns, No. 16 enam. on a T-80-2 core. Inductance is a 1.9 μH.
L21B — 15 turns, No. 16 enam. on a T-60-6 core. Inductance is 1.08 μH.
L21C — 10 turns, No. 16 enam. on a T-60-6 core. Inductance is 0.57 μH.
L21D — 9 turns, No. 16 enam. on a T-80-6 core. Inductance is 0.41 μH.
L21E — 8 turns, No. 16 enam. on a T-80-6 core. Inductance is 0.33 μH.
L22A — 16 turns, No. 16 enam. wound on a T-80-2 core. Inductance is 1.46 μH. L22B — 13 turns, No. 16 enam. on a T-80-2 core. Inductance is 0.76 μH.
L22C — 9 turns, No. 16 enam. on a T-80-6 core. Inductance is 0.41 μH.
L22D — 8 turns, No. 16 enam. on a T-80-6 core. Inductance is 0.27 μH.
L22E — 7 turns, No. 16 enam. on a T-80-6 core. Inductance is 0.19 μH.
RFC1 — Six beads from Amidon.
RFC2, RFC3 — One bead from Amidon.
T1 — Primary 6 turns ct; secondary 2 turns ct. See text.
T2 — Primary 4 turns ct; secondary 6 turns. See text.
T3 — 26 turns, No. 30 enam. on 0.3 in. Q1 type core.

supplied to the product detector. Part of the i-f output is sampled and amplified by a CA3028 rf amplifier. The amplified i-f signal is detected in a diode voltage doubler, and the dc signal is fed to a cascaded dc amplifier using a 2N3903 and a 2N3905 transistor. The 2N3905 controls the S meter and supplies the alc/agc control voltage. The agc/alc control bus is a common line, allowed by bilateral stages common to both receiver and transmitter functions.

Two agc ranges select fast or slow agc characteristics. The one-mA S meter, a small surplus movement, indicates relative received signal strengths and transmitter alc levels. In transmit, voltage is not supplied to the i-f stages, product detector or CA3028 agc amplifier. The 2N3903 and 2N3905 dc amplifiers remain active, and alc voltage developed in the PA is fed back to the dc amplifiers for alc control voltage generation.

Audio Amplifier

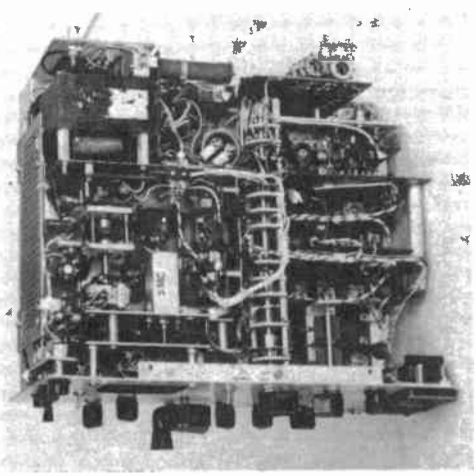
A large variety of IC amplifiers is available that could be used in the audio stage of the transceiver. A Sprague 2277 was used as it was readily available at a nearby Radio Shack store. The RS-2277 is actually two two-watt rms amplifiers in one package and was intended for small stereo systems. It requires a minimum of external components and provides good gain. Only one section of the 2277 was used. It delivers slightly less than two watts with the 14.5-volt supply. Large ground foil areas left on the pc board will provide a satisfactory heatsink for the IC. The amplifier runs continuously to allow sidetone monitoring for cw operation. The 1500-μF capacitor from pin 14 of the 2277 is needed for decoupling and stable operation. The audio output will drive most low-impedance loads. Cw selectivity is obtained through a simple bandpass audio filter built from 44-mH telephone loading coils, and the 1000-Hz filter is adequate for casual cw operation.

A sharper filter skirt may be obtained by removing the two 2700-ohm resistors across the 44-mH coils (at the expense of a noticeable increase in ringing). During cw operation, the sidetone oscillator is active on keydown and is fed to the audio amplifier. The amplifier assembly is small and is located parallel to the front panel behind the audio-gain control.

The SSB Generator and Associated Circuits

The two sideband crystals are those supplied by the filter manufacturer and should be set to the frequencies recommended. Rather than using a third crystal for cw generation, it is possible to pull the 9.0017-MHz crystal within the filter passband with a diode-switched trimmer. For cw generation the BFO oscillator transistor emitter is keyed. When receiving in the cw mode, relay contacts K7B energize the BFO for product detector and counter operation. During sideband operation, contacts of mode switch S1B prevent keying the BFO. While in cw operation, voltage is supplied to the twin-tee sidetone oscillator by the contacts of S1A of the

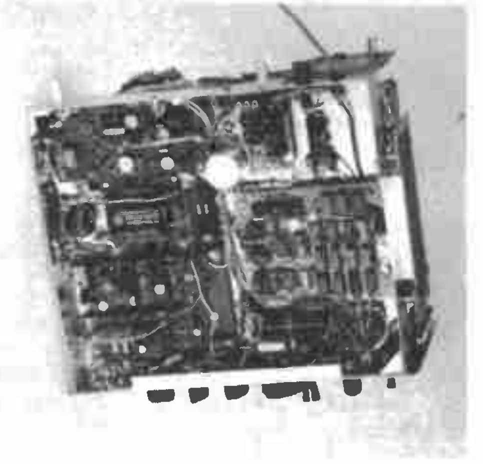
Top view of the transceiver.



mode switch. The emitter of the sidetone oscillator transistor is also keyed during cw operation. Isolation diode CR4 prevents relay contacts of K7B from activating the sidetone generator in receive which could result in false tripping of the semi-break-in circuit. Because the BFO signal is a part of the display frequency, during key-up periods in transmit the display readouts are blanked.

Mode switch contacts S1A supply voltage to the mic amplifiers in either sideband position. An RCA CA3018A transistor array in the VOX circuit reduces the number of active components to a few devices. In the sideband positions, mode switch contacts of S1C feeds audio to the VOX circuit for normal VOX operation. In cw, contacts of S1C feed sidetone audio to the VOX input allowing semi-break-in operation, if desired. The VOX gain, delay and anti trip controls are front-panel mounted for accessibility.

The VOX circuit energizes relay K7. This relay does voltage switching and primarily controls receive or transmit operation. When VOX operation is not desired, the mic PTT contacts will manually key K7, or a front-panel mounted switch allows the transceiver to be locked in transmit. The VOX assembly is mounted to the speaker support bracket within the VFO compartment. Another Motorola MC1496 IC is used for the balanced modulator. Audio is fed to pin four of the IC, BFO injection to pin eight. Best carrier rejection at 9 MHz results with 80 millivolts p-p of BFO injection. The BFO level may be varied by changing the value of the 82-pF coupling capacitor. A multiturn trimpot is used for the carrier balance control to allow for accurate and stable nulling of carrier. Because no rf is present on the carrier balance control, it may be mounted away from the sideband generator. During cw operation the carrier balance is deliberately upset by placing a potential on pin one of the MC1496, permitting the BFO energy to pass through the modulator stage. The balanced modulator is not powered during receive conditions which prevents spurious 9-MHz signals from reaching the filter and the receiver i-f stages. Besides the modulator, BFO



Bottom view of the transceiver.

injection is supplied to the product detector and to the BFO counter-prescaler circuit. Excessive radiation of the BFO signal, through careless shielding or bypassing, could reach the i-f stages causing agc action or i-f blocking to occur.

Transmitter Power Chain

The development of a high-gain broad-band power amplifier able to span 30 MHz can be difficult. Fortunately, two previously published articles laid the basic ground work for the transceiver's power chain.[1] The circuitry was intended for 20-meter service and to obtain adequate performance at 10 meters required several circuit revisions. This included replacing the 2N2102 stages with 2N3866 transistors. The CA3028 predriver, like the preselector and 10-meter postamplifier, is under alc control. Drive to the predriver stage is manually regulated by the drive-level control. A low-frequency rolloff *RC* network after the drive-level control helps to equalize for the decrease in transistor gain at the higher frequencies. The push-pull output of the predriver drives two class AB 2N3866 emitter follower stages. These, in turn, drive two push-pull AB 2N3866 transistors in common-emitter configuration. Heatsink fins are needed on the 2N3866 transistors.

The final amplifier is a 12-volt version of one described by Lowe.[2] Construction of T1 and T2 is somewhat unconventional. The original amplifier described by Lowe was designed for 28-volt operation allowing the use of inexpensive surplus transistors. Since 14.5-volt circuitry is used for the rf stages, this involved changing the turns ratios of T1 and T2 and necessitated using 12-volt transistors in place of the 2N3632s. Degenerative feedback in the PA circuit improves linearity with a small sacrifice in output power. Excessive rf drive causes the CA3028 predriver to saturate before danger-

TABLE I

Capacitor values in pF.

C7A – 560	C10A – 300
C7B – 270	C10B – 160
C7C – 150	C10C – 75
C7D – 100	C10D – 51
C7E – 68	C10E – 39
C8A – 100	C11A – 390
C8B – 51	C11B – 200
C8C – 24	C11C – 100
C8D – 15	C11D – 68
C8E – 10	C11E – 51
C9A – 910	C12A – 62
C9B – 500	C12B – 43
C9C – 250	C12C – 10
C9D – 160	C12D – 22
C9E – 120	C12E – None used

[1] Hill, "Single Band SSB Transceiver," *Ham Radio*, Nov., 1973.

[2] Lowe, "A 15-Watt Output Solid-State Linear Amplifier for 3.5 to 30 MHz," *QST*, Dec., 1971.

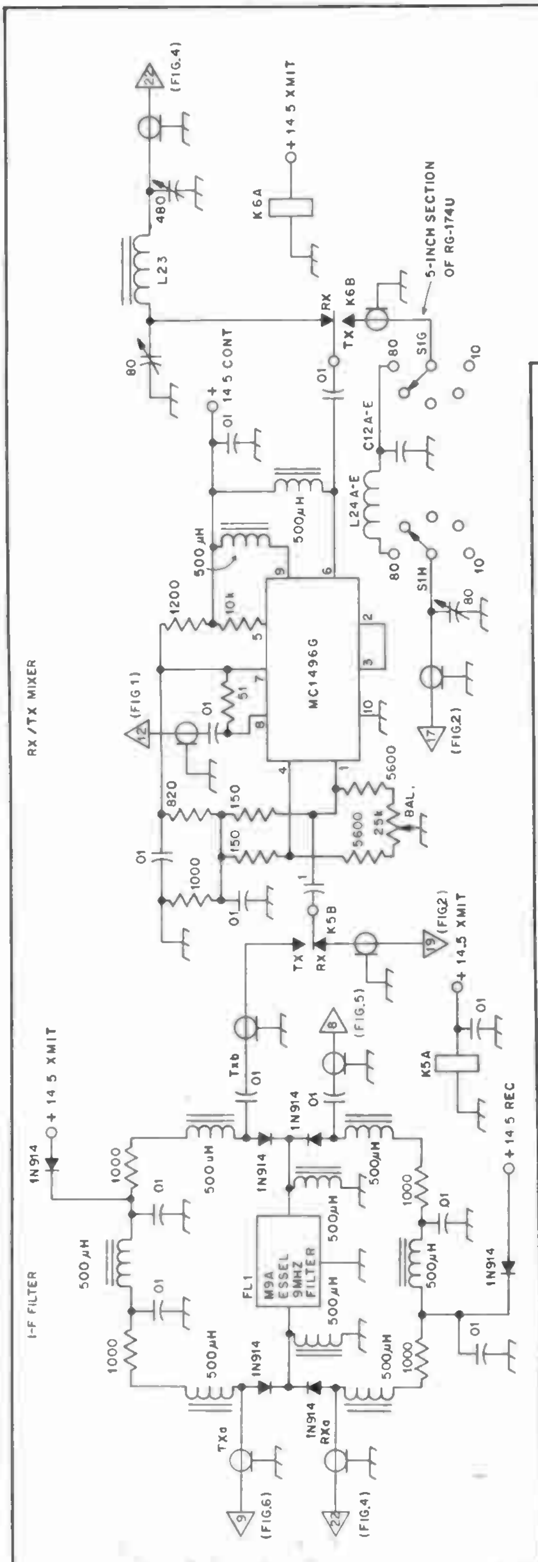


Fig. 4 — Circuit diagram for the i-f filter switching circuit and for the receiver-transmitter mixer section. See Table I for capacitor values.
K5, K6 — 12 V dc, spdt.
L23 — 40 turns. No. 28 enam. wound on a T-50-2 core.
L24A — 62 turns, No. 26 enam. wound on a T-50-2 core.
L24B — 35 turns, No. 26 enam. wound on a T-50-2 core.
L24C — 22 turns. No. 22 enam. wound on a T-50-2 core.
L24D — 12 turns, No. 22 enam. wound on a T-50-2 core.
L24E — 4 turns, No. 22 enam. wound on a T-50-2 core.

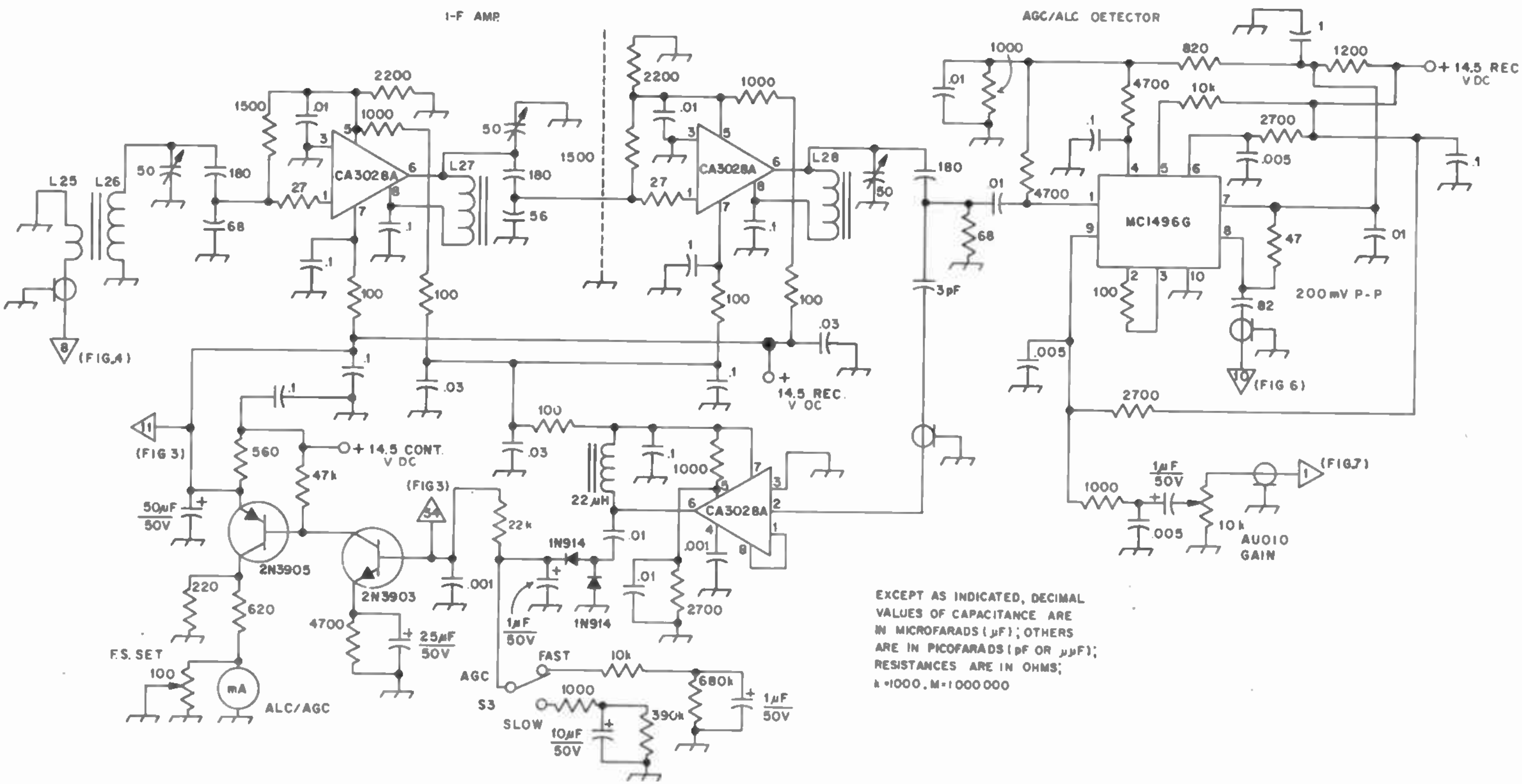
ously high base currents are reached in the final amplifier.

T3 samples PA rf output. The secondary of T3 is detected to produce a proportional alc dc output. Alc action is regulated by a front-panel alc control. The 2N2222 alc amplifier increases the dc signal to a level usable for the cascaded dc alc/agc amplifier. The 1N34 diode prevents the 390-ohm 2N2222 emitter resistor from loading the agc voltage developed in the voltage-doubler detector. Because of the broad-band nature of the amplifier, undesirable energy could be readily coupled to the antenna. Five elliptical low-pass filters, one for each individual band, are used to filter the transmitter output. Band-switch sections 10 and 11 select the appropriate filter for the band in use.[3]

The power chain was built on double-clad epoxy board. A 12-volt power amplifier develops large rf currents, and stable efficient operation demands low-impedance ground paths. The top foil was left on the board to form a ground plane. Leads above ground going to lower foil runs are isolated by reaming away copper around the lead holes. Leads going to lower ground foil runs should also be soldered to the upper ground plane foil for good bonding and to reduce ground loops. The board was mounted in a section of 1/8-inch aluminum channel that serves as a shield and a heatsink for the final transistors. The PA transistors were mounted directly on the aluminum channel. The use of metal spacers and screws to mount the board to the aluminum channel assured adequate rf grounding.

Spectral analysis of the two-tone ssb output signal shows third-order IMD products to be 27 dB down at full power. With a slight reduction in drive, better than 30 dB are obtainable. Broadband white noise, harmonics and other undesirable spurious signals are well within acceptable minimum levels. Because excessive drive levels cause saturation of the CA3028 predriver, it is good practice not to use excessive drive on cw to keep harmonic generation at a minimum. Total transmitter current drain is monitored with a three-ampere meter movement. Proper drive levels for ssb operation must be initially determined using an

[3] Schubert, "Lowpass Filters for Solid State Linears," *Ham Radio*, March, 1974.



EXCEPT AS INDICATED, DECIMAL VALUES OF CAPACITANCE ARE IN MICROFARADS (μF) ; OTHERS ARE IN PICOFARADS (pF OR μμF); RESISTANCES ARE IN OHMS; k = 1000, M = 1 000 000

Fig. 5 — Circuit diagram for the i-f amplifier, the agc-alc system, and the product detector.
L25 — 15 turns of No. 24 enam.
L26, L27, L28 — 25 turns, No. 24 enam. wound on a T-50-2 core.

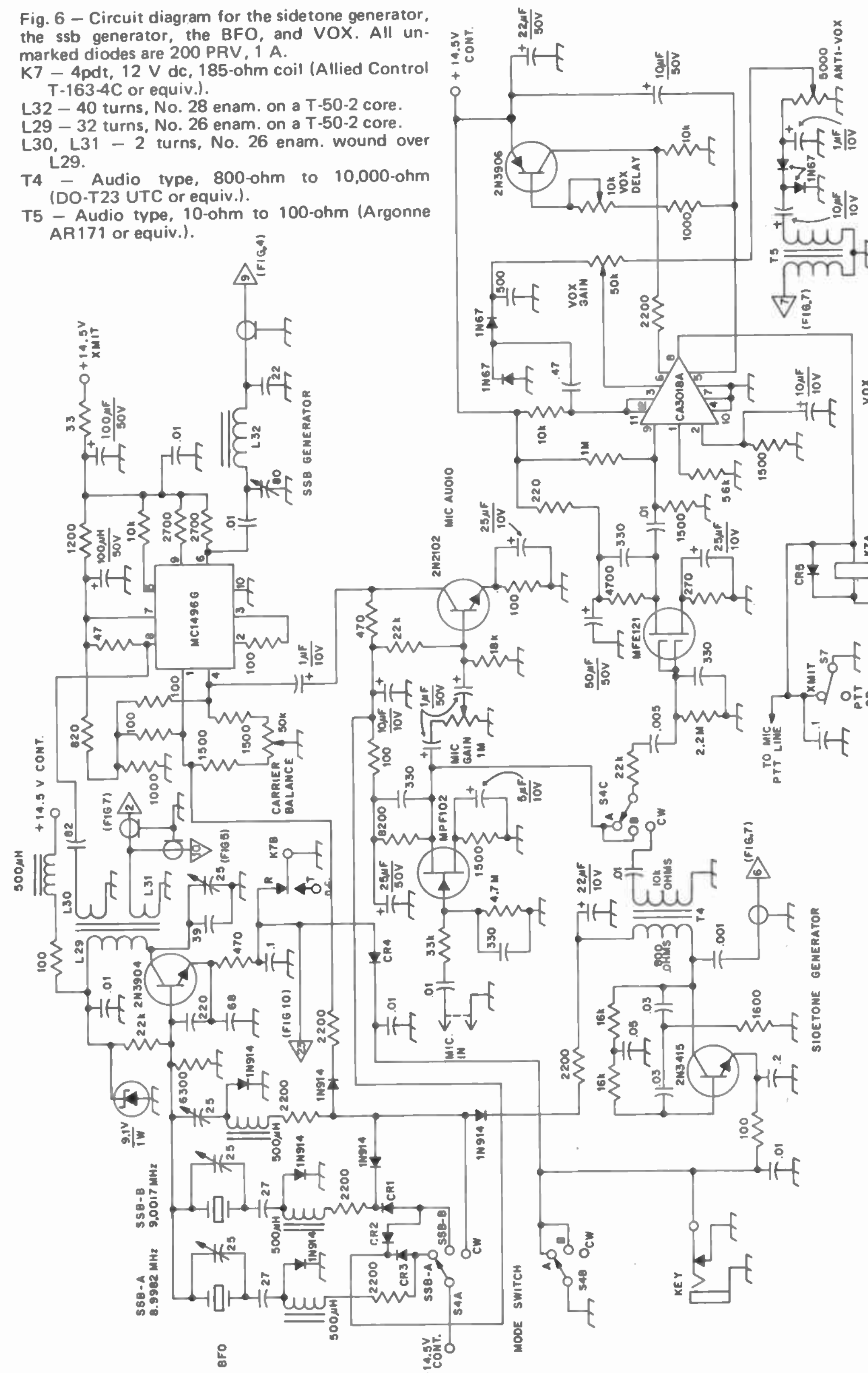
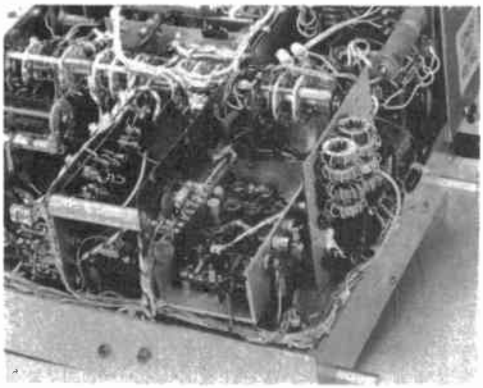


Fig. 6 – Circuit diagram for the sidetone generator, the ssb generator, the BFO, and VOX. All unmarked diodes are 200 PRV, 1 A.
K7 – 4pdt, 12 V dc, 185-ohm coil (Allied Control T-163-4C or equiv.).
L32 – 40 turns, No. 28 enam. on a T-50-2 core.
L29 – 32 turns, No. 26 enam. on a T-50-2 core.
L30, L31 – 2 turns, No. 26 enam. wound over L29.
T4 – Audio type, 800-ohm to 10,000-ohm (DO-T23 UTC or equiv.).
T5 – Audio type, 10-ohm to 100-ohm (Argonne AR171 or equiv.).



Rear view of the transceiver.

oscilloscope. Once correct peak current is determined, the alc control should be set to show signs of alc action on voice peaks. The peak meter reading depends on the characteristics of the meter used and upon the voice characteristics of the operator.

Band-Switch Notes

The 11 section ceramic band switch used in the transceiver is assembled from several switches removed from scrapped electronic assemblies. Various switch manufacturers sell the necessary wafer and drive assemblies to produce almost any custom switch required. If available, a fiber drive shaft will reduce unwanted coupling between switch sections. Shields are needed for isolation between switch wafer sections 2 and 3, 4 and 5, 6 and 7, 8 and 9, and between 9 and 10. These shields are grounded to nearby shield partitions. The metal switch hardware at the end of the assembly should be grounded. In the author's unit, the vertical elliptical-filter pc board was supported in part by the rear of the switch assembly, and ground foil remaining on the board provided a low-impedance ground path to the chasis. The transmitter power runs straight through from the mixer output representing a large value of in-line gain produced by the preselector and power chain circuitry. It doesn't take much unwanted coupling to produce feedback paths. Careful shielding, bonding, and elimination of ground loops are important for stable operation.

Power Supply

The power supply provides regulated 5-volts dc and 14.5-volts dc for the transceiver. The five-volt supply is for the TTL logic and LED displays in the frequency counter; the 14.5-volt supply powers the rf and audio sections. A National LM 309 or Fairchild 7805 IC is the heart of the 5-volt supply. The IC has internal current limiting and automatic over temperature cutoff for self protection. The output is electronically filtered. A series resistor reduces heatsink requirements for the 5-volt regulator.

The 14.5-volt supply centers around a Fairchild 7815 regulator. This IC has the same features of the 7805 except it furnishes 15 volts. Because the 7815 alone will not meet the current demands of the transceiver, two MJE-3055 pass transistors are used to amplify the current capability of the 14.5-volt supply. Both pass transistors and IC regulators require a heatsink. The dropping resistor for the 5-volt regulator should be mounted so that it will dissipate heat without adversely affecting nearby circuits. An SCR crowbar circuit blows the power supply fuse if regulator or pass transistor failure allows more than 18 volts to appear on the 14-volt line. Large transients will also activate the SCR.

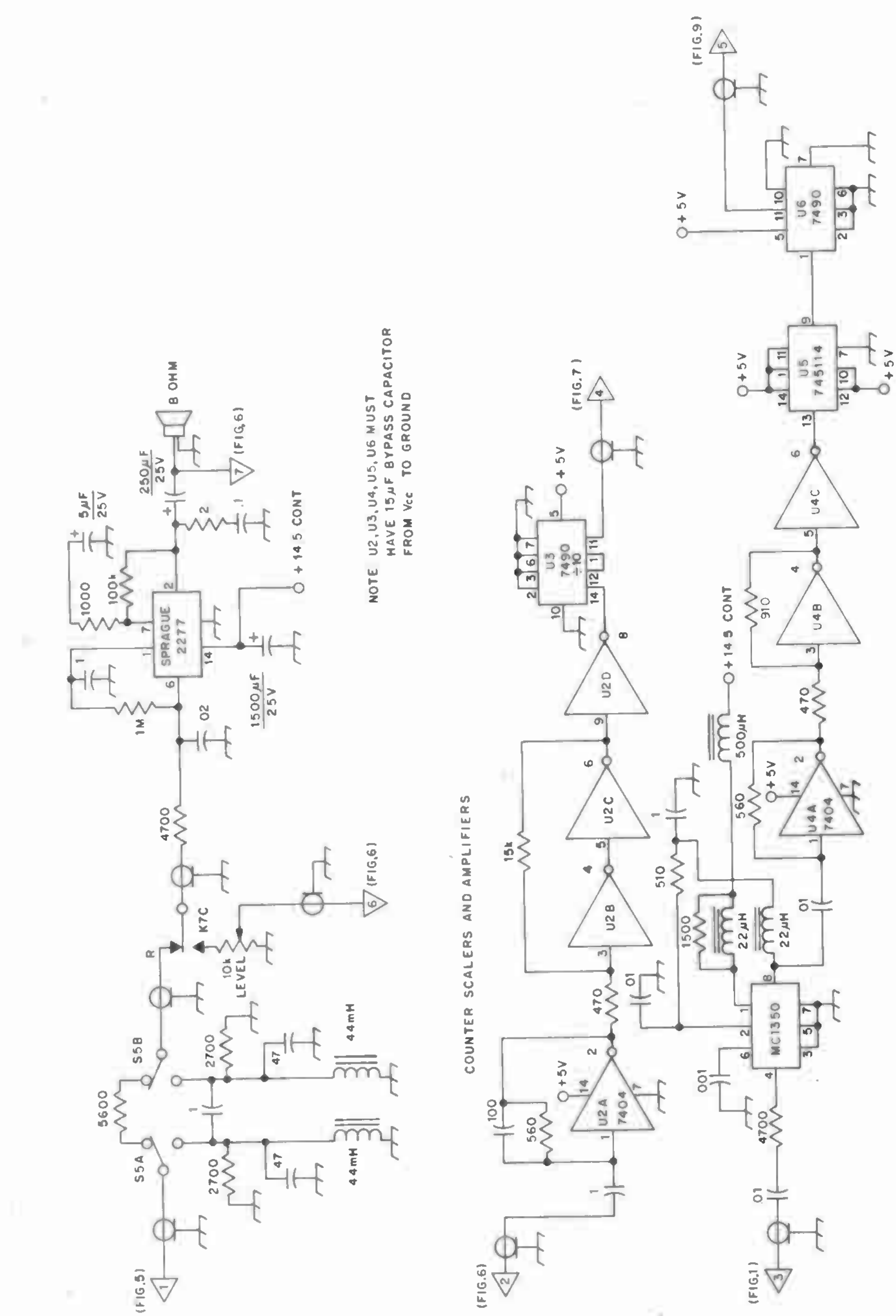
If battery operation is contemplated, the 5-volt regulator should be operated directly from the 14.4-volt source without the series dropping resistor. Operation directly from an 18- to 28-volt supply may be accomplished, provided the regulators and pass transistors are used for voltage reduction.

The Frequency Display

Two separate amplifiers and scaler stages are used to amplify and shape the rf voltage levels to be TTL compatible. Since the 74192 up-down counters will only operate reliably to 25 MHz in this application, each counter input is prescaled by a factor of ten. The highest scaled input to the counter will be about 3 MHz on 15 meters from the 30-MHz premixer output. The BFO amplifier uses a 7404 hex inverter biased for amplifier service, and amplifies and shapes the 9-MHz BFO signal to a 5-volt logic level. A 7490 decade counter divides the BFO signal down to 900 kHz. These amplifiers and scalers should be well shielded and bypassed as their harmonics radiate well into the hf region. A MC1350 IC serves to amplify, as applicable, either the premixer or VFO output. Another 7404 hex inverter is used to further amplify and shape the rf signal to a usable logic level. It was found that a 7490 would not operate reliably in the presence of an rf field with the maximum 30-MHz premixer output. The solution was to use a high speed 74S114 as a divide-by-two prescaler followed by a 7490 divide-by-five stage.

The basic counter circuit evolved from an article appearing in the 1974, May-June issue of the *Air Force MARS Newsletter*. As originally

Fig. 7 — Circuit diagram for the cw filter, the counter scalers and amplifiers, and the audio amplifier stage for the receiver.



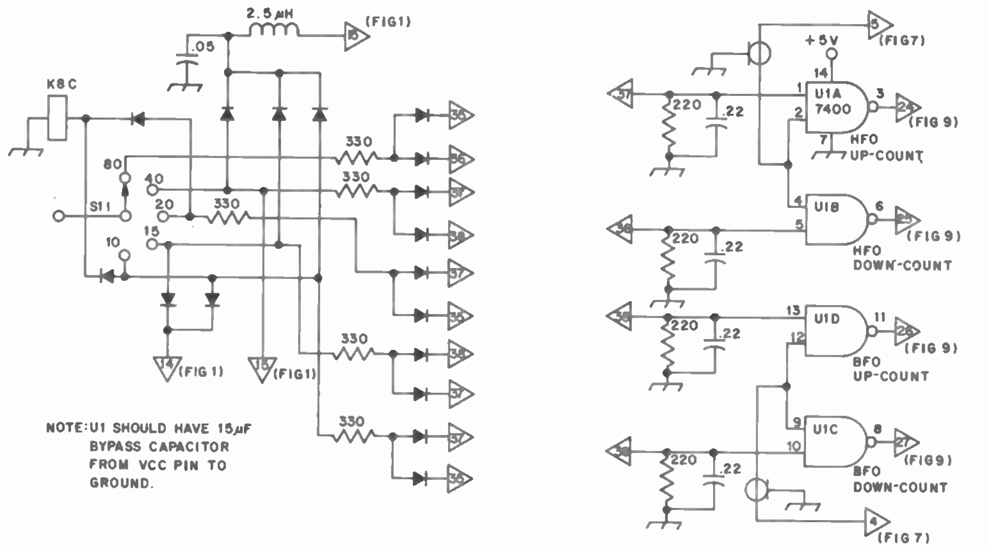


Fig. 8 — Circuit diagram for the diode matrix to select the preselector output for the up-down converters. K8 is a 12-V dc dpdt unit. All diodes are power type, 200 PRV, 1 A.

intended, the counter was designed to count up the high-frequency oscillator of a Swan transceiver, and then change over and count down to 5.5-MHz carrier generating oscillator; the difference of these two (the operating frequency of the transceiver) would then be transferred, latched and displayed. With the premixing arrangement used in this transceiver, it is necessary to count up or down either the premixer or BFO signals. On 20 and 10 meters, a double up count of the BFO and premixer signals is needed to reproduce the transceiver operating frequency.

Fig. 8 illustrates the use of U1, a 7400 gate. This gate, in conjunction with a diode matrix from band-switch section nine, serves to select the appropriate prescaler output for the up- or down-counter inputs. Relay K8 is energized on 20 or 10 meters allowing two sequential up counts of the scaler outputs to occur in lieu of the normal up-down count sequence.

The three 74164 registers form the gating circuit for the counter. With initially all of the registers at zero, the first ten clock pulses enable the up-counter gate; the down counter is disabled. The eleventh clock pulse disables the up-counter input, and enables the down-counter input until the 20th clock pulse, when both inputs are disabled. On 10 or 20 meters, the eleventh clock pulse enables the second up-counter input and disables the first up-counter input (U18A-B).

On the 22nd clock pulse, the 7475 latches are set with the count present at the 74192 outputs. The 22nd pulse also clears both the 74192 counters and 74164 gating registers. Because it is possible for the 74192 counters to be reset before the 7475s latch, causing a loss of the count, inverters U13E and U13F provide a small time delay before clearing the 74192 counters.

The counter displays four digits. With the 100-Hz clock rate selected, the hundreds of kHz, tens of kHz, kHz, and hundreds of Hz are displayed. A complete counter cycle takes 200 ms, which provides rapid updating of changes in operating frequency. A selectable 10-Hz clock rate allows displays to within 10 Hz to be displayed, with a loss of the 100-kHz position and a 2.2-second counter update period. The MHz decimal point to the left of the MSD is illuminated to indicate when the higher clock rate is selected.

The counter was assembled on a small square of Vectorbord. This allowed numerous circuit changes to be made in the construction of the counter. The counter circuitry, like the prescalers, generates large amounts of rf hash and should be well bypassed and shielded from the receiver circuitry. The use of sockets is recommended. Ideally the counter should be constructed on double-sided pc board with an upper ground plane. The readouts were mounted in a Monsanto MHO4C circularly polarized bezel. In operation the last digit varies up to ±1 count, possibly caused by the random loss of one or two counts depending upon the phase relationship between the gating clock pulses and the counter input signals. Placing the scalers after the counter gating may have eliminated this condition, but it is only a minor annoyance.

A double-pole center-off switch is used to select the 10-Hz or 100-Hz clock rate. The center-off position removes the 5-V dc supply to the counter circuitry, allowing considerable power savings when the counter is not needed. The counter adds a touch of class to the transceiver as well as operating flexibility and convenience. On the minus side, it must be admitted that a substantial portion of the transceiver power requirements and cost are represented in the counter section. Some amateurs may wish to forego the counter in favor of substituting a mechanical dial arrangement.

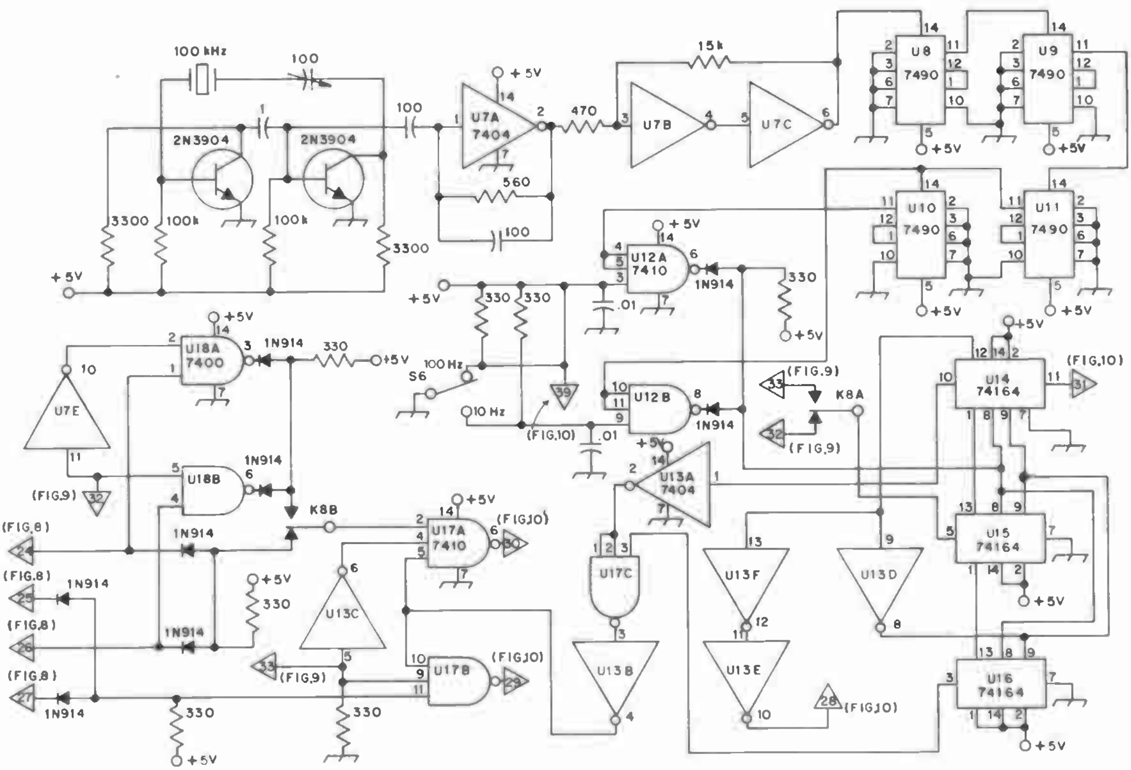


Fig. 9 — Circuit diagram for the counter circuits.

Transceiver Alignment

It is assumed that the builder is experienced in aligning ssb equipment, and space permits only a cursory alignment procedure to be outlined. Access to a wideband calibrated oscilloscope is recommended. The VFO range can be set with either a frequency counter or general-coverage receiver. Disabling the BFO prescaler and selecting 80 or 20 meters will allow the internal counter to display the VFO frequency.

The BFO tank should be adjusted so the oscillator will start reliably when power is applied. The trimmer for the BFO crystals should be set using a counter. If one is not available, disable the premixer prescaler and use the internal counter. On 40 meters, the 21.5-MHz HFO should be adjusted so that it starts reliably. The 25-MHz oscillator should be adjusted in a similar manner on either 10 or 15 meters. Set both of the MC1496 mixer balancing pots to mid position. On the appropriate bands, adjust the premixer tanks for maximum output at mid band. Use an rf voltmeter or suitable scope on the low-impedance premixer output for these measurements. With the scope still on the premixer output, carefully adjust the balance pot until input injection feedthrough is minimal. Adjust the 9-MHz i-f receiver stages using a signal generator. Adjust for maximum S-meter deflection. The 9-MHz receiver-transmitter mixer output can be tuned in a similar fashion. The third harmonic of the VFO will beat with the 25-MHz oscillator to produce a strong 9-MHz signal on 10 and 15 meters. Tune the receiver to 28.666 MHz. Carefully null the signal using the receiver transmitter mixer balance pot. The preselector coils may be peaked on receive using a signal generator, or by on-the-air signals. Verify that the MC1496 mixer injection levels (including balanced modulator and BFO) are correct. In cw transmit, peak the 9-MHz balanced modulator output. If transmit drive levels are low, repeak the preselector coils for maximum transmitter drive. In ssb transmit, adjust the carrier-balance pot for minimal carrier feedthrough. Peak the 80 pF receiver-transmitter mixer pi-network loading trimmer for maximum 10-meter cw output.

The mic gain control is fixed. First disable the transmitter power chain. Then, using a suitable scope, monitor the low-impedance receiver-transmitter mixer output. Using the station mic and speaking in a slightly louder than normal voice, adjust the mic-gain control to a point just below where flat topping begins in the receiver-transmitter mixer. This insures adequate ssb drive levels on 10 and 15 meters. Changing microphones will require this procedure to be repeated.

Preselector tuning during transmit should be smooth; if the preselector tuning peaks abnormally sharp on some bands, it is a probable indication of

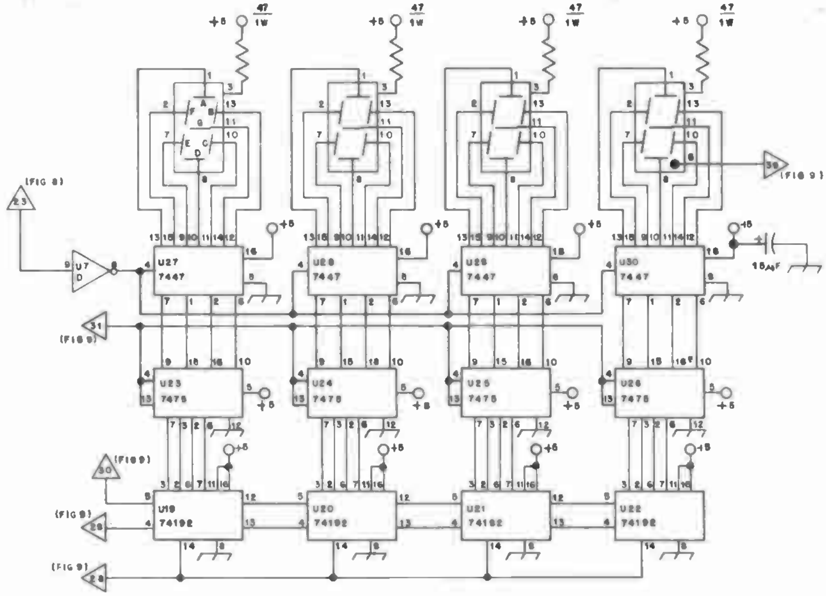


Fig. 10 — Circuit diagram for the display unit.

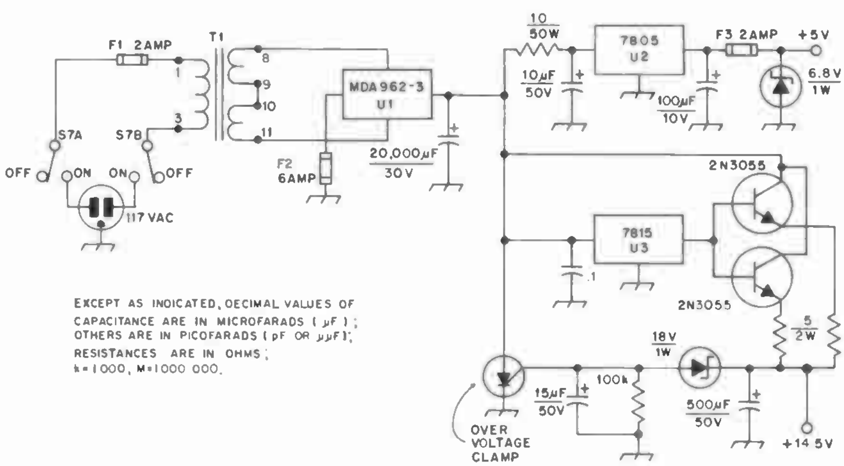


Fig. 11 — Circuit diagram for the transceiver power supply. U2 and U3 are manufactured by Fairchild. VR1 is a General Electric type C35B. U1 is from Motorola. T1 is a Stancor RT-204.

regeneration. Likewise, the drive-level control should exercise a smooth linear control over output power. In ssb transmit, with the rf-drive level at maximum, it should be possible to tune the preselector throughout its range with no indication of spurious oscillations.

The receiver performs as well as most on the amateur market today. The 20-watt output power places the transmitter above the QRP class, although some discretion is needed when rubbing shoulders with high-power stations on a crowded band. Reports have praised the audio quality, and many stations refused to believe the transmitter was running only 20 watts.

SIMPLE TRANSMITTING FILTERS

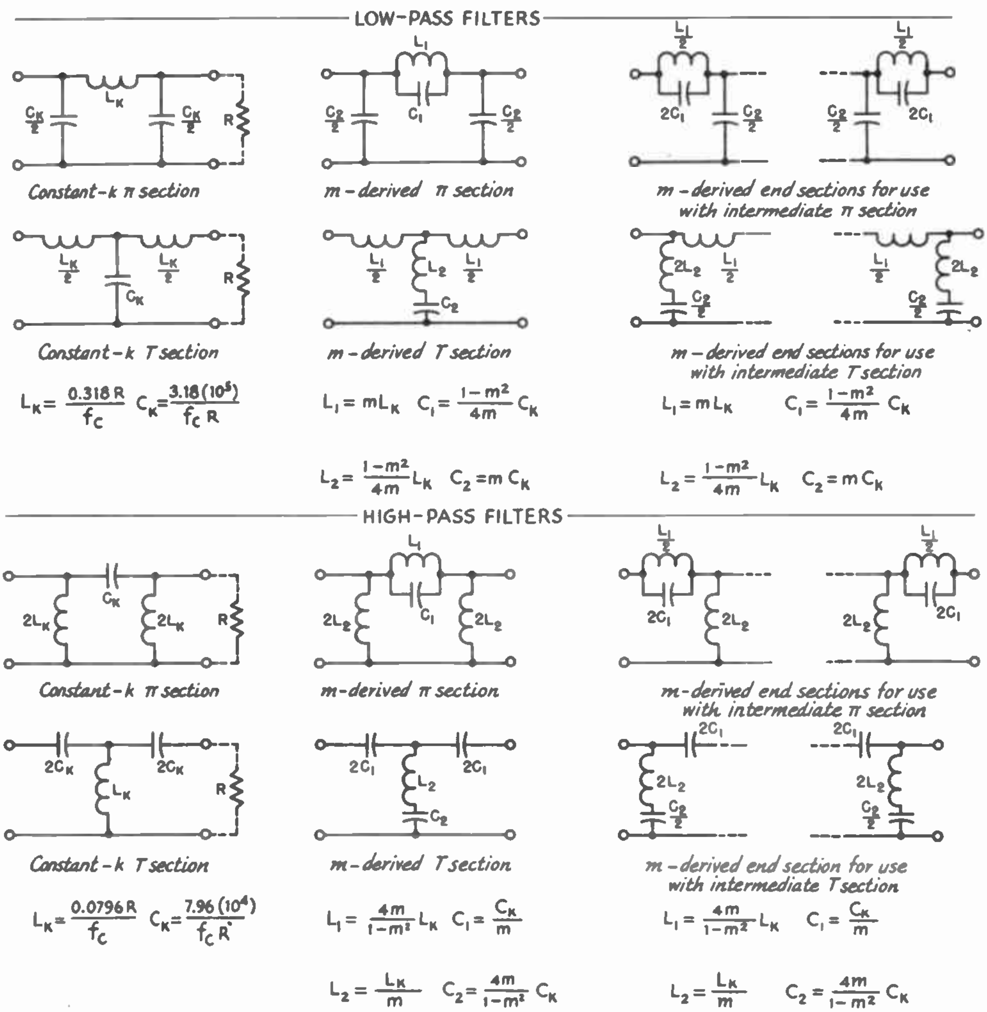
While the filters in this chart represent somewhat obsolete designs, the circuits are likely to be encountered because of their popularity. The terms *wave filters* or *image-parameter* filter are often applied to this type of network. A basic limitation on image-parameter filters is that a terminating impedance that varies in a complicated manner is required if the exact filter response is to be realized. Consequently, such designs are only approximate in nature and should not be used if close tolerence on attenuation is required.

The units for L, C, R, and f are microhenries, picofarads, ohms and megahertz, respectively. If the so-called *m-derived* filter section is to be used, the value of m in the chart can be found from

$$m = \sqrt{1 - \left(\frac{f_c}{f_\infty}\right)^2}$$ for the low-pass filter and

$$m = \sqrt{1 - \left(\frac{f_\infty}{f_c}\right)^2}$$ for the high-pass filter.

where f_c is the cutoff frequency and f_∞ is a frequency of high attenuation.



A 160-METER AMPLIFIER

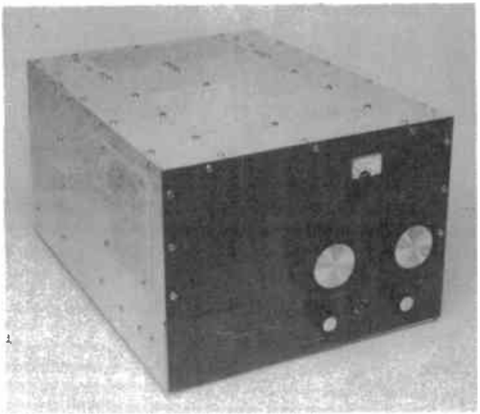


Fig. 1 — Front view of the 160-meter amplifier. Note the use of perforated aluminum stock to permit ventilation of both the rf and power supply compartments. The large front-panel knob on the right controls C3, while the adjacent knob to the left controls C2. The power switch, S1, is controlled by the smaller knob located beneath C3. Both S1 and S2, the meter switch, are mounted below the chassis, and DS1 is mounted between the two switches.

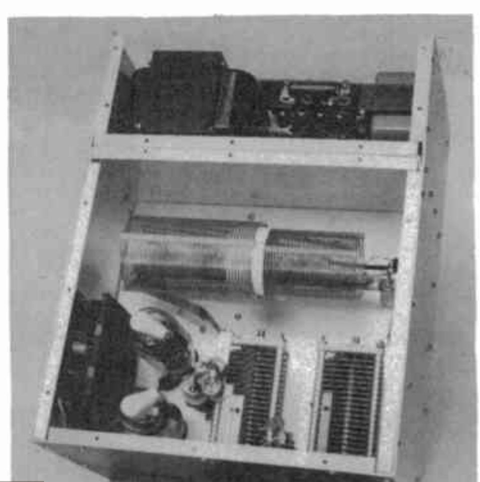
Anyone who has operated in the 160-meter band lately can attest to the fact that interest in the "top band" is on the upswing. With only a handful of manufacturers producing gear for 160, this band is somewhat of a "homebrewers' haven." Most operation takes place during the evening hours, because the high level of daytime ionospheric absorption makes communication (other than strictly local) all but impossible for low powered stations. Summertime static makes things even more difficult. At present, amateurs occupy this band on a shared basis with various radionavigation services, with maximum input power limitations imposed to prevent harmful interference from occuring. These restrictions are greatest between sundown and sunrise, when the potential for interference is at maximum. However, during the daylight hours, amateurs in 29 states are permitted to use up to 1000 watts power input, while in the other 21, the maximum is 500 watts, in selected segments of the band.[1] The amplifier described below is for use with 160-meter exciters in the 50- to 100-watt output class, for ssb and cw operation.

Circuit Data

A pair of 572B/T160L triodes are used in a cathode-driven, grounded-grid configuration (see Fig. 3). A small amount of operating bias is provided by the 3.9-volt, 10-watt Zener diode in series with the cathode return lead, and the tubes are completely cut off during nontransmitting periods by opening that lead with K1A to reduce unnecessary power consumption and heat generation. The other contacts on K1 perform all necessary antenna switching functions for transceive or separate transmitter/receiver operation. Drive power from the exciter is fed to the directly heated cathodes through a parallel combination of three .01 μF disk capacitors, and a resonant cathode tank circuit helps minimize the amount of drive required. The filament choke, RFC2, isolates the driving signal from the filament transformer. A B&W FC-15A choke was used here. A single power switch, S1, applies 117 V ac to the primaries of both the power and filament transformers simultaneously, as the 572B's require no significant warmup time. S1 also activates the cooling fan, B1, and the front-panel pilot light assembly, DS1. The self-contained high-voltage power supply uses a straightforward voltage doubler circuit. No-load voltage is approximately 3100 V dc, dropping to 2600 V dc under one kilowatt key-down conditions. R2 limits the initial surge current to the filter capacitor bank to prevent exceeding the current handling capability of the rectifier string when the supply is first turned on.

A single 0-1 mA meter is used to monitor either plate voltage or cathode current. To measure plate voltage, a multiplier consisting of five series-connected 1-megohm 1-watt resistors with one end tied to the B plus line is switched in series with the meter to provide a full-scale reading of 5000 volts. A 1000-ohm one-watt resistor between the bottom of the meter multiplier and ground prevents the full B plus voltage from appearing across the meter switch, S2, when it is in the other position. To measure cathode current, the meter is placed in

Fig. 2 — Top view of the amplifier. The rf components occupy the foreground, while the heat-generating power-supply components are visible behind the compartment shield at the rear.



[1] A chart of U.S. and Canadian 160-meter sub-allocations is available from ARRL Headquarters; send a stamped, self-addressed envelope and request form S-15A.

Fig. 3 — Circuit diagram for the 160-meter amplifier. Fixed-value capacitors are ceramic disk unless otherwise indicated. Polarized capacitors are electrolytic. All resistors are 1/2-watt composition unless noted otherwise.

B1 — 117-volt axial fan (Rotron Whisper Fan or equiv.).
C1 — Parallel combination of one 5000, 2000, and 1000-pF silver-mica capacitors.
C2, C3 — 250-pF air variable, .075-inch spacing (E. F. Johnson 154-9 or equiv.).
C4, C5, C6 — Transmitting capacitor, 1000-pF "doorknob" (Centralab 858S or equiv.).
CR1-CR12, incl. — 1000 PRV, 2.5-A silicon (Motorola HEP 170 or equiv.).
DS1 — 117-volt ac neon pilot lamp assembly.
J1 — Phone jack.
J2, J3 — Coax chassis-mounting connector, type SO-239.
K1 — 3pdt, 10-A contacts, 110-V dc coil (Potter and Brumfield KUP14D15 or equiv.).

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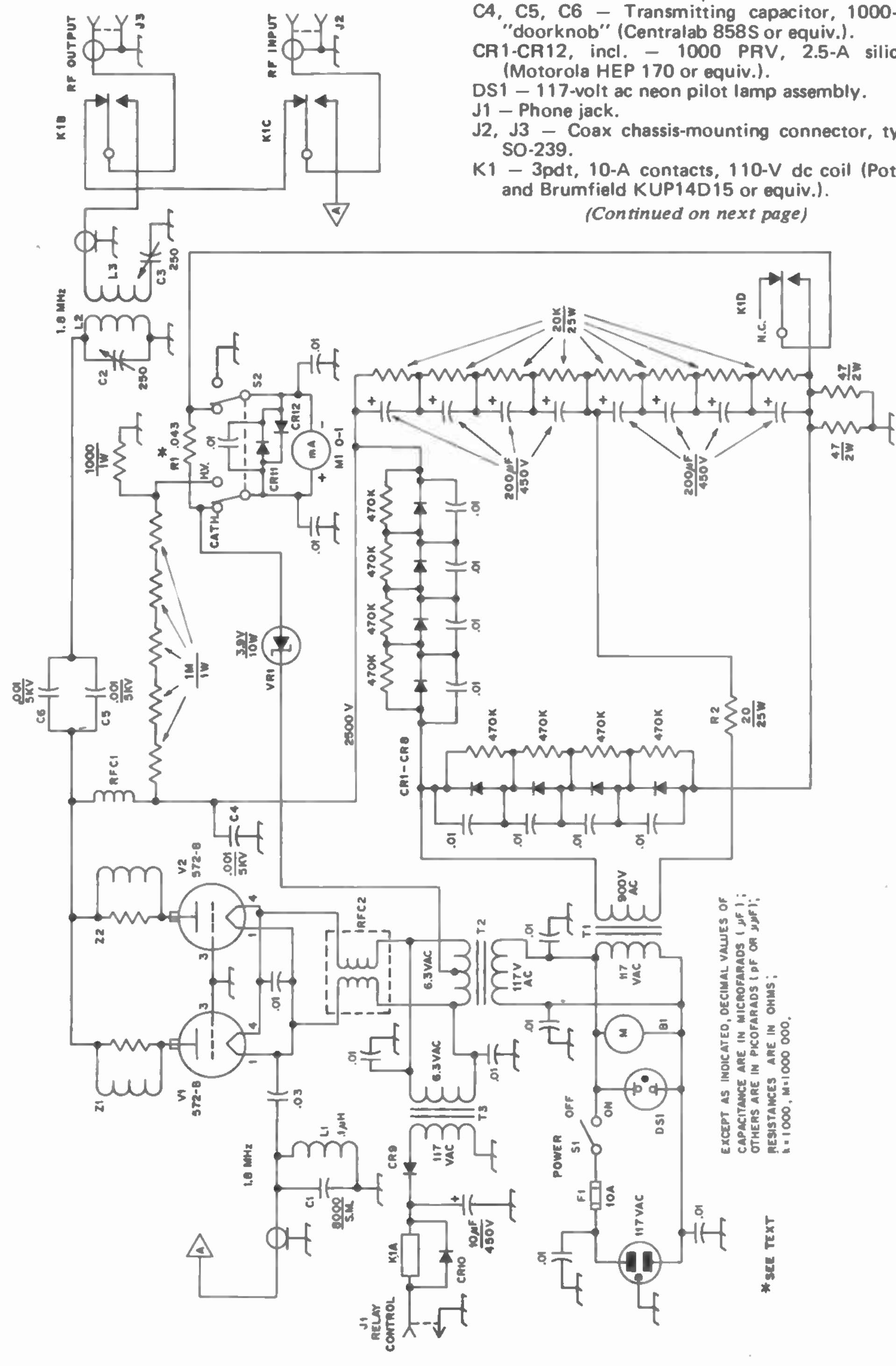


L1 – 1.0 μH
L2, L3 – See text.
M1 – 1-mA dc (Simpson model 2121 or equiv.).
RFC1 – 1.0 mH, 500 mA (E. F. Johnson 102-752 or equiv.).
S1 – Spst rotary switch.
S2 – Dpdt rotary switch.
T1 – 117-volt primary; secondary 625-0-625 volts ac (ct not used) at 450 mA (Hammond No. 720).
T2 – 117-volt primary; secondary 6.3 V ct at 10 A (Stancor P-6464 or equiv.).
T3 – 117-volt primary; secondary 6.3 V ac.
VR1 – Zener, 3.9-V, 10-watt (Motorola HEP Z3500 or equiv.).

parallel with shunt R1, which remains in series with the cathode return lead at all times. To obtain a full-scale reading of one ampere, a shunt resistance of .043 ohms was used with the Simpson model 2121 meter, as it has an internal resistance of 43 ohms (see Chapter 17).

As this amplifier is designed for monoband operation, the mechanical and electrical complexities and compromises involved in the bandswitching of an output network are not a factor here. Tuned-link coupling is used in the output circuit. The grid of each 572B is tied directly to chassis ground, using short leads, to avoid problems with instability. Parasitic suppressors Z1 and Z2 also contribute to stability. Neutralization is not necessary.

B&W Miniductor stock is used at L2 and L3. L2 is made from 43 turns of B&W 3034 (No. 14 wire, 8 tpi, 3-inch dia.) and L3 is made from 39 turns of B&W 3030 (No. 14 wire, 8 tpi, 2-1/2-inch dia.). The coils are supported on a 10-inch strip of bakelite which is mounted on three 1-1/2-inch steatite insulating cones. L2 is epoxied into place on the side of the bakelite strip nearest the tubes. L3 will be partially inserted into the cold end of L2, and is epoxied into place after initial adjustments have been made. L3 must be able to slide freely inside L2 without making electrical contact. The first 10 turns of L3 may be covered with a layer of Scotch No. 27 glass insulating tape. Leads from L3 are made with teflon-insulated flexible stranded wire to allow the coil a degree of freedom of movement during initial adjustment. Rf output from L3 is connected to K1B through a short length of RG-58/U coaxial cable.

Meter shunt R1 is made by winding 12-1/2 inches of No. 26 enam. wire around a 1-megohm 2-watt resistor. If the meter used has an internal resistance other than 43 ohms, the appropriate shunt resistance value may be wound by referring to the copper wire resistance table in Chapter 18.

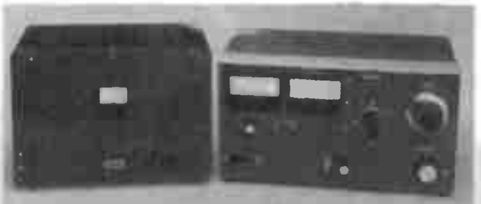
Parasitic suppressors Z1 and Z2 are each made with 3-1/2 turns of No. 14 enam. wire wound around the parallel combination of three 82-ohm, 1-watt composition resistors, mounted right at each plate cap.

Operation

The power supply should be tested before rf drive is applied to the amplifier. For initial tests, it is desirable to control the power transformer primary voltage with a Powerstat, while leaving the filament transformer primary and fan connected directly to the 117 V ac line. *Remember at all times that lethal voltages exist both above and below chassis.* Do not make any internal adjustments with the power on, or even with the power off until the bleeders have fully discharged the filter capacitors (at least 40 seconds with this particular amplifier). It is good practice to clip a lead from the B-plus terminal to ground after the capacitors have discharged, whenever working inside the amplifier (remember to remove it before applying power!). The tuned-input circuit (L1-C1), should be checked with a grid-dip meter for resonance at the frequency segment of interest. K1 must be closed during transmit; this may be effected by shorting the wire from J1 to ground with a relay inside the 160-meter exciter, or with an external switch. Starting with a plate voltage of about 1500 volts, drive is applied through J2 and C2 is adjusted for maximum rf output as indicated on an external rf wattmeter or relative output indicator. C3 is then adjusted for maximum output. The plate voltage may now be advanced to its normal level. The link may be moved in or out (with power off) and C2 and C3 again adjusted until the highest efficiency is obtained. At that point the link, L3, may be epoxied in place. In the amplifier described here, the optimum position for L3 was when eight of its turns were inside L2. This may be used as a starting point for the adjustment. Normal tune-up procedure involves only the adjustment of C2 and C3 for maximum output, within the maximum legal power limits, of course. During normal operation the 572B anodes may glow with a dull red color. The tubes draw about 50 mA resting current, when K1 is closed and no drive is applied.

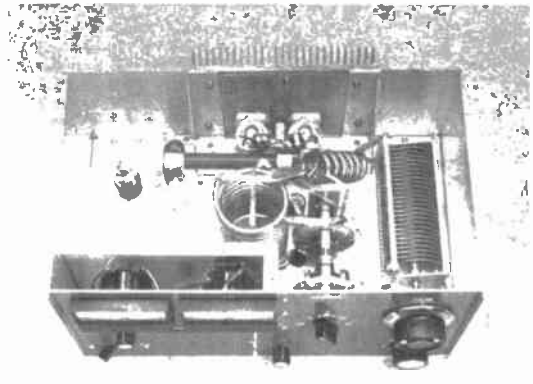
A CONDUCTION-COOLED TWO-KILOWATT AMPLIFIER

One of the major concerns when dealing with high power amplifiers is heat and how to reduce it. The usual method has been to use a large fan or blower, but this solution is generally noisy. By using the principles of heat transfer, a noiseless amplifier can be made with the use of an adequate heat sink and conduction-cooled tubes.



The amplifier shown in the photographs and schematically in Fig. 1 uses a pair of recently designed 8873 conduction-cooled triode tubes. The circuit configuration is grounded grid and uses no tuned-input tank components. When properly adjusted, the amplifier is capable of IMD characteristics which are better than can be achieved by a

A 2-kW Amplifier



Top view of the 80-through 10-meter conduction-cooled amplifier. The chassis is 17 × 12 × 3 inches (43.2 × 30.5 × 7.6 cm) and is totally enclosed in a shield. A separate partition was fabricated to prevent rf leakage through the meter holes in the front panel. An old National Radio Company vernier dial is used in conjunction with the plate tuning capacitor to provide ease of adjustment (especially on 10 meters). The position of the dial for each band is marked on the dial skirt with a black pen and india ink.

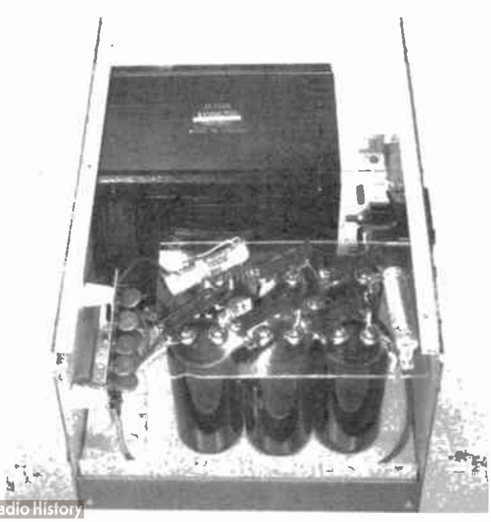
typical exciter, therefore the added complexity of band switching a tuned-input circuit was deemed unnecessary.

Construction

Building an amplifier such as this is often an exercise in adapting readily available components to a published circuit. For this reason, a blow-by-blow description of this phase of the project will not be given. An effort was made, however, to use parts which are available generally, and should the builder desire, this model could be copied verbatim.

The most difficult constructional problem is that of aligning the tube sockets correctly. It is imperative that the sockets be aligned so that when the tubes are mounted in place, the flat surfaces of the anodes fit smoothly and snugly against the thermal-link heat-transfer material. Any misalignment here could destroy the tubes (or tube) the first time full power is applied. The mounting holes for the tube sockets are enlarged to allow final positioning after the tubes are "socked" in place with the clamping hardware. Pressure must be applied to the anodes so that they are always snug against the thermal link. The hardware used to perform this function must be nonconducting material capable of withstanding as much as 250°C. The pressure bracket used here was fabricated from several Millen jack-bar strips (metal clips removed) mounted in back-to-back fashion. The entire assembly is held in place by means of a long piece of No. 10 threaded brass rod which passes through a small hole in the center of the heat sink. An attempt to give meaningful comments about how tight the tubes should be pressured to the copper and aluminum sink will not be given. Suffice it to say that the tubes should fit flat and snugly against the thermal hardware. The heat sink was purchased from Thermaloy and is connected to a 1/4-inch thick piece of ordinary copper plate. The total cost for the copper and the aluminum sink is somewhat more than the price of a good centrifugal blower ($30) but the savings offered by not having to purchase special tube sockets and glass chimneys overcomes the cost differential.

The power supply is built on a separate chassis because the plate transformer is bulky and cumbersome. A special transformer was designed for this amplifier by Hammond Transformer Co. Ltd., of Guelph, Ont. Canada. The transformer contains two windings, one is for the plate supply to be used in a voltage-doubler circuit and the other is for the tube filaments. The power supply produces 2200 volts under a load of 500mA, and is rated for 2000 watts. The Hammond part number is given in Fig. 1. All of the interconnections for power-supply control and the operating voltages needed by the amplifier are carried by a seven-conductor cable. This excludes the B plus, however, which is connected between the units by means of a piece of test-probe wire (5-kV rating) with Millen high-voltage connectors mounted at both ends. The seven-conductor cable is made from several pieces of two-conductor household wire (No. 10) available at most hardware stores. Since the main power switch is mounted on the front panel of the amplifier, the power supply may be placed in some remote position, out of the way from the operator (not a bad idea!). A high-voltage meter was included with the power supply so that it could be used with other amplifiers. It serves no purpose with this system. The main amplifier deck has provisions for monitoring the plate voltage.



Top view of the power supply built by WA1JZC showing the technique for mounting the filter-capacitor bank. The diodes are mounted on a printed-circuit board which is fastened to the rear of the cabinet with cone insulators and suitable hardware.

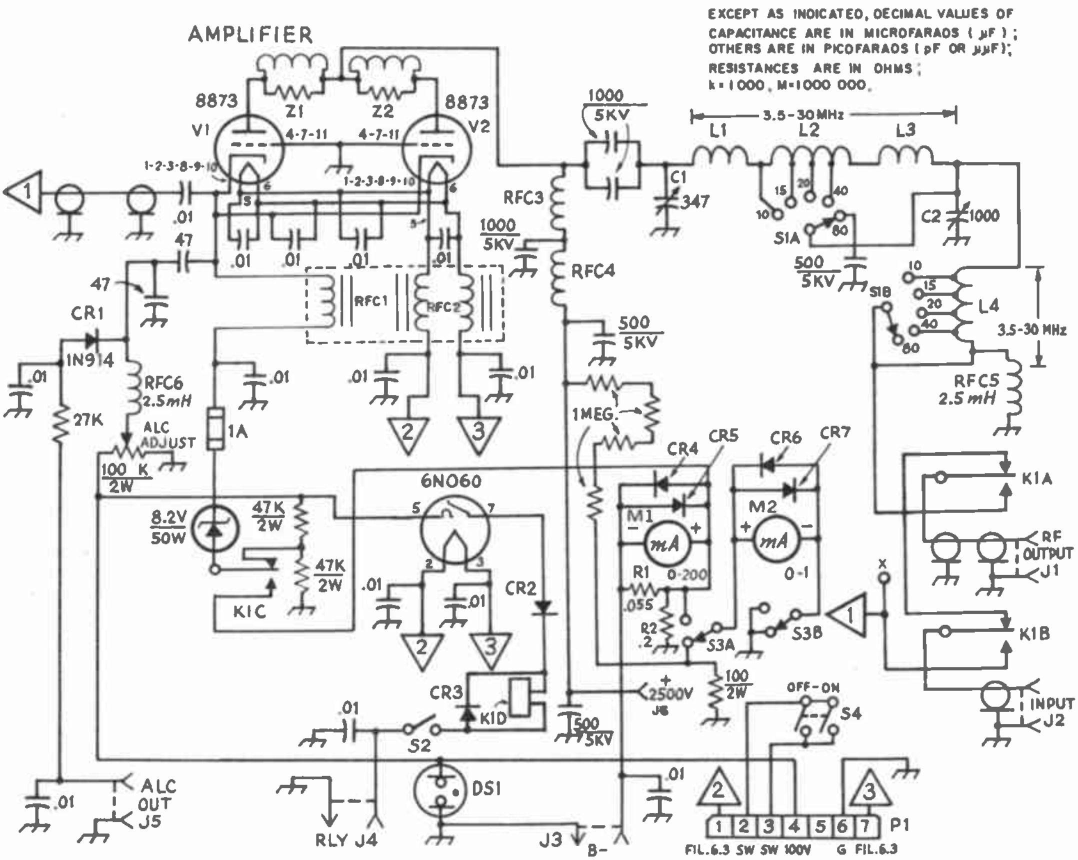


Fig. 1 — Circuit diagram for the 8873 conduction-cooled amplifier. Component designations not listed below are for text reference. RFC1 and RFC2 are wound on the same ferrite rod in the same direction; three wires are wound together (Amidon MU-125 kit). Tube sockets for V1 and V2 are E.F. Johnson 124-0311-100. The thermal links are available from Eimac with the tubes. The heat sink is part number 2559-080-A000 from Astrodyne Inc., 353 Middlesex Ave., Wilmington, MA 01887, and costs approximately $20.

C1 — Transmitting air variable, 347 pF (E.F. Johnson 154-0010-001).
C2 — Transmitting air variable, 1000 pF (E.F. Johnson 154-30).
CR2-CR7, incl. — 1000 PRV, 2.5 A (Motorola HEP170).
J1 — SO-239 chassis mounted coaxial connector.
J3, J4, J5 — Phono jack, panel mount.
J6 — High-voltage connection (Millen 37001).
K1 — Enclosed, three-pole relay, 110-volt dc coil (Potter and Brumfield KUP14D15).
L1 — 4-3/4 turns of 1/4-inch copper tubing, 1-3/4-inch inside diameter, 2-1/4 inches long.
L2 — 12-1/2 turns, 1/4-inch copper tubing, 2-3/4-inch inside diameter, tap at one turn from connection point with L1, 2-1/2 inches for 20 meters, 7-3/4 turns for 40 meters.
L3 — 11-1/2 turns, 2-inch diameter, 6 tpi (Barker and Williamson 3025).
L4 — 10 turns, 2-inch diameter, 6 tpi, with taps at 3 turns for 10 meters, 3-1/2 turns for 15 meters, 4-3/4 turns for 20 meters, 6-3/4 turns for 40 meters; all taps made from junction of L3 (Barker and Williamson 3025).
M1 — 200 mA full scale, 0.5-ohm internal resistance (Simpson Electric Designer Series Model 523).
M2 — 1 mA full scale, 43 ohms internal resistance (Simpson Electric, same series as M1).
R1 — Meter shunt, .05555 ohms constructed from 3.375 feet of No. 22 enam. wire wound over the body of any 2-watt resistor higher than 100 ohms in value.
R2 — Meter shunt, 0.2 ohms made from five 1-ohm, 1-watt resistors connected in parallel.
RFC1, RFC5, RFC6 — 2.5 mH (Millen 34300-2500).
RFC3 — Rf choke (Barker and Williamson Model 800 with 10 turns removed from the bottom end).
RFC4 — 22 µH (Millen 34300).
S1 — High-voltage band-selector style, double pole, six position (James Millen 51001 style).
Z1, Z2 — 2 turns 3/8-inch-wide copper strap wound over three 100-ohm, 2-watt resistors connected in parallel.

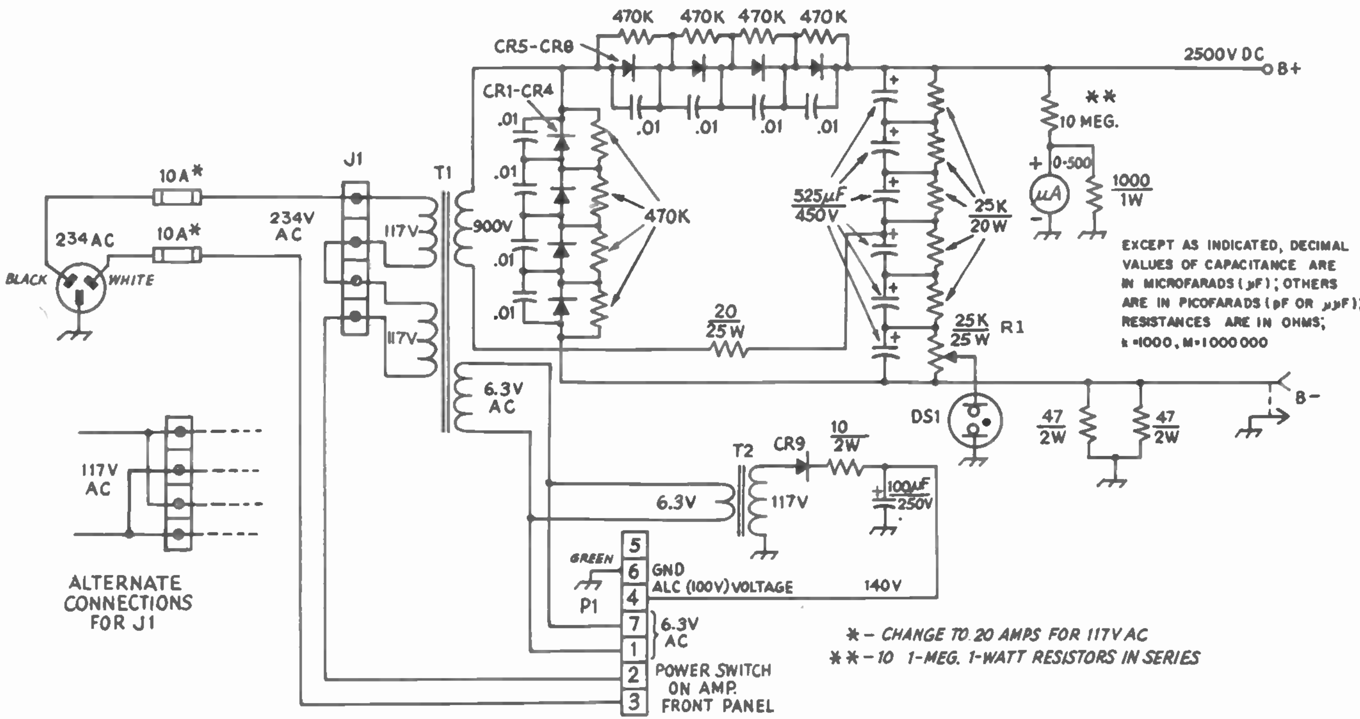


Fig. 2 – Circuit diagram for the power supply. The power transformer is available from Hammond; type no. 101165. CR1 through CR9 are 2.5 A, 1000 PRV; see Fig. 1 for suitable part number. T2 is Stancor part number P-8190 and is rated for 6.3 volts at 1.2 amperes. DS1 is a 117-volt neon pilot lamp assembly. The tap at R1 should be set for 5000 ohms to the B minus lead. Adjustments to this tap cannot be made while voltage is applied to the power supply. *If the pilot lamp does not glow properly, remove the ac cord, allow suitable time for the high-voltage to bleed to zero, and apply a screwdriver between the B-plus line and ground before making any adjustments!*

A conventional household light switch may be used for S4. If the switch is to be mounted horizontally, be sure to use a contactor device and not a mercury type (which operates in a vertical position only). A double-pole switch was used with both poles connected in parallel. The rating is 220 V at 10 A per section.

The RF Deck

The two sections of the pi-L network are isolated from each other by placing one of them under the chassis. Although not shown in the photograph, a shield was added to prevent rf energy from entering the control section underneath the chassis. The shield divides the chassis between the tube sockets and the inductors. The loading capacitor is mounted directly beneath the plate-tuning capacitor. This scheme provides an excellent mechanical arrangement as well as a neat front-panel layout.

The 8873s require a 60-second warmup time, and accordingly, a one-minute time-delay circuit is included in the design. The amplifier IN/OUT switch is independent of the main power switch and the time delay. Once the delay circuit "times out," the amplifier may be placed in or out of the line to the antenna, whenever desired. A safety problem exists here: there is no large blower running, and there are no brightly illuminated tubes to warn the operator that the amplifier is turned on. Except for the pilot lamp on the front panel, one might be fooled into believing the amplifier is turned off! And if the pilot lamp should burn out, there is *absolutely* no way to tell if the power is turned on (with the resultant high voltage at the anodes of the 8873s). *Beware!*

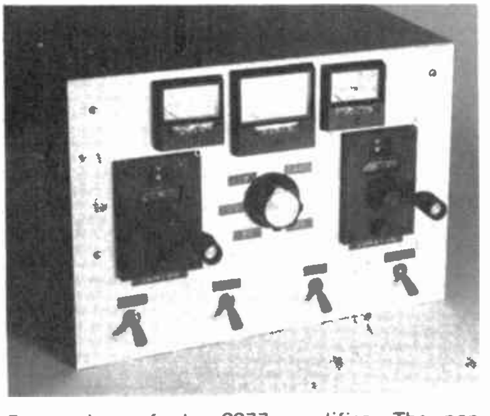
Operation

Tuning a pi-L-output circuit is somewhat different than tuning a conventional pi-network because the grid current should be monitored closely. Grid current depends on two items, drive power and amplifier loading. The procedure found to be most effective is to tune for maximum power output with the loading sufficiently heavy to keep the grid current below the maximum level while adjusting the drive power for the proper amount of plate current. The plate current for cw operation should be 450 mA and approximately 900 mA under single-tone tuning conditions for ssb. This presents a problem since it is not legal to operate under single-tone tuning conditions for ssb. Sixty watts of drive power will provide full input levels. For use with high-power exciters, see *QST* for October, 1973.

A TWO-KILOWATT AMPLIFIER USING THE EIMAC 8877 TRIODE

One of the easier projects for the amateur to undertake is the construction of an amplifier for use on the hf bands. Generally speaking, the mechanical aspects of the construction are more difficult to handle than the electrical ones. And, as with any construction project, acquiring the parts can be difficult. The two-kilowatt amplifier shown here is designed for dependable service at the maximum legal power input allowed in the United States. The component ratings are generous and the construction is heavy duty. Since power handling capability is typically determined by physical size, most of the components used here are large and accordingly, a split arrangement has been employed allowing the placement of the power supply on a separate chassis from the amplifier compartment.

Another feature sets this amplifier apart from most others described in the literature; the air is exhausted from the top of the tube socket instead of the conventional pressurized chassis air-flow system.



Front view of the 8877 amplifier. The non-sequential numbering of the band switch is discussed in the text. A switch is provided to allow the selection of proper bias for the mode in use at the time.

The Circuit

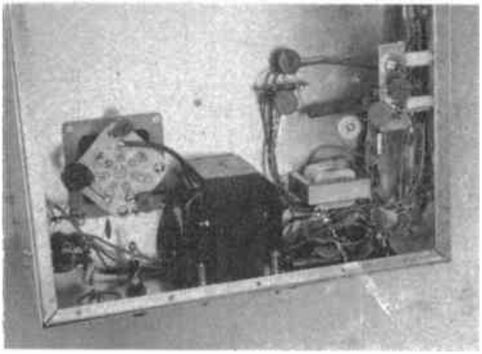
The triode, a 3CX1500/8877, is connected in a grounded-grid configuration which provides about the most simple layout possible. The output tank circuit is a pi-network with vacuum-variable capacitors used for both input and output tuning. A 2.5-mH rf choke is connected between the output end of the tank and ground to prevent B plus from appearing at the antenna terminals should C1 develop a short.

A passive, untuned, capacitor-coupled circuit is used to apply rf-drive energy to the 8877 cathode. Since a moderate amount of bias is permitted, L1 is incorporated to provide isolation from rf to the bias-developing Zener diodes. The highest recommended bias voltage for use on ssb is 8.2, but for cw operation, where IMD is not important, 22 volts is developed which nearly places the tube at cutoff (zero no-signal plate current). A 10-kΩ resistor is included in series with the Zener-diode circuit to assure complete cutoff of the 8877 during receiving periods. It is switched out of the circuit during transmit periods by a set of contacts on K1.

Antenna transfer along with bias switching is accomplished with two relays. Sequencing can be an important factor since it is very undesirable to provide drive to the 8877, remove its bias, all before the output circuit relay has closed and stopped "bouncing." This is accomplished by the use of a vacuum relay shown as K2 in Fig. 1. K2 is many times faster in operation than K1 and accordingly, the antenna is placed on the amplifier output circuit well ahead of drive arriving at the cathode of the 8877. Voltage to operate both relays is developed by T2. Since the relays are connected in a fashion to allow straight-through operation of an exciter or transceiver to the antenna in the de-energized position, interruption of the voltage from T2 during transmit periods with it is undesirable; to have the amplifier "on line" and developing power, is all that need be done. S4 serves that function.

Metering of three operating conditions of the amplifier is accomplished with three meters instead of one or two. The purpose is first, to eliminate a switch for selection, and second, to provide continuous indication of the important parameters of the 8877. Of course, operator error is reduced since it is impossible to assume a meter is measuring one thing while a switch is selected for another. Both plate and grid current meters are direct-reading instruments – no shunts are needed. A string of resistors is used at R1 to multiply the scale of the 500-μA meter to indicate zero through 5000 volts. R1 is constructed of 10 resistors, one-watt in size and one-megohm in value. The purpose is to keep the applied voltage across each resistor below 600. A 1000-ohm resistor is included at the meter end of R1 to keep the voltage low should the meter winding become an open circuit.

Another feature of this particular amplifier is the use of a motorized Powerstat for control of the high-voltage circuit. The ability to select the operating plate voltage from the front panel of the amplifier is a feature desired by the builder of the project and need not be duplicated. If voltage control is not used, the power transformer used in the high-voltage power supply should be selected to provide about 3000 volts. This is a suitable compromise for efficient cw and ssb operation at the maximum power input levels. In actual operation, the amplifier shown in the photographs is



Bottom view of the amplifier chassis. L1 is shown near the tube socket. It is wound on a T-50-2 Amidon toroid core. The Zener diodes are mounted on a plate and secured to the side chassis wall.

used with 3400 volts during ssb operation and 2500 volts for cw conditions.

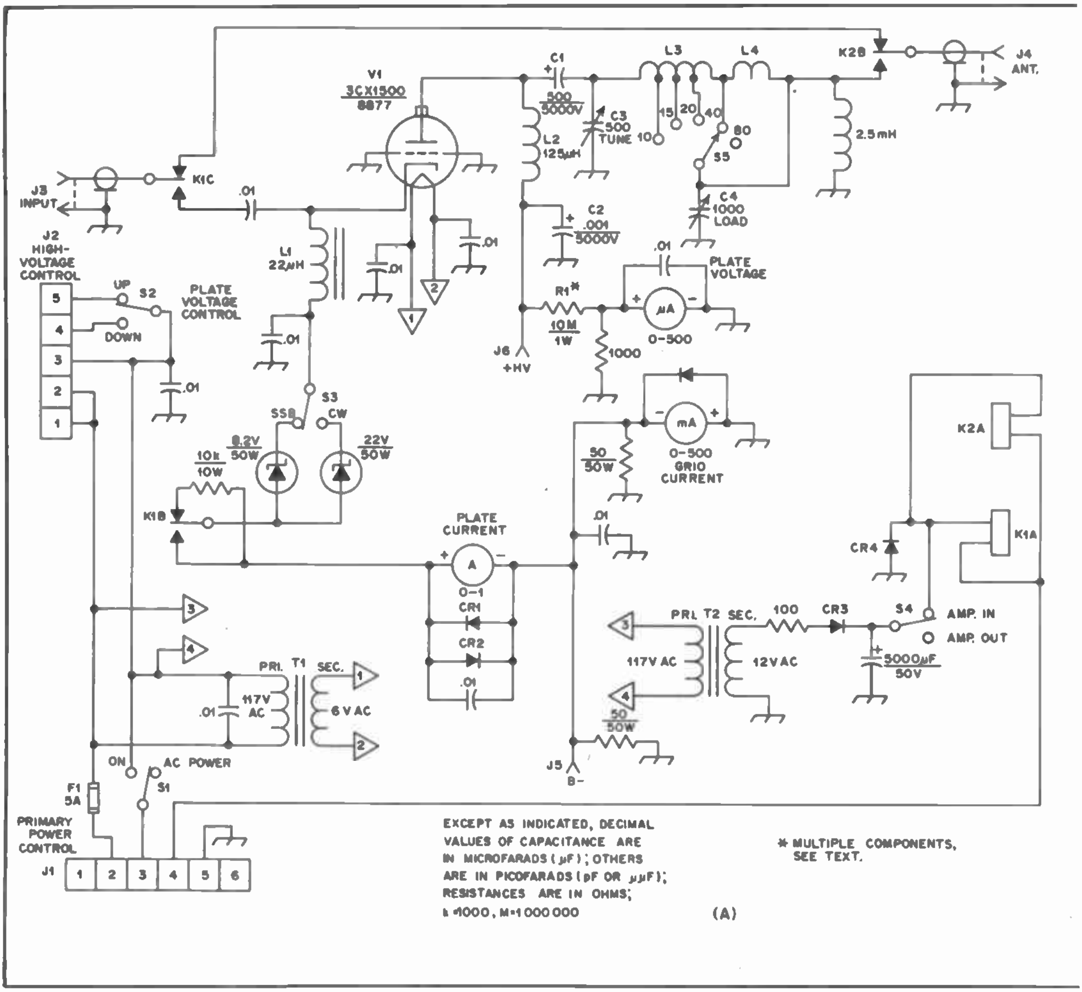
Several 50-ohm, 10-watt resistors have been placed at various points in the B-minus circuit, both in the amplifier-chassis compartment and on the power supply chassis. This prevents the B-minus lead from creeping above "almost ground" potential should a defect develop in the grid-current metering circuit. Also, included in the B-plus lead is a 10-ohm resistor which will help prevent component damage should a direct short take place in the amplifier compartment.

Mechanical Construction

The split-chassis configuration offers several advantages. First, it allows the amplifier compartment to be somewhat more compact because the power supply can be located elsewhere. This consumes less space on the operating desk. It also divides the weight into parts; the heavier section may be placed on the floor. The power supply may be equipped with wheels to give it mobility. The only disadvantage with having a two chassis system comes when portability is desired.

The power supply has been assembled on an aluminum plate which is 1/4-inch thick. Casters are provided because the plate transformer itself weighs about 80 pounds. The capacitor bank for filtering has also been mounted on the aluminum plate. A circuit board is used to interconnect the capacitors and is supported above the plate with ceramic pillar insulators. A screened covering is provided to keep unwanted objects from contacting the high-voltage system. The power supply relay (T3 primary connection) and the Powerstat have been assembled separately and may be interconnected to the power supply chassis plate via an inconnecting cable. Mounting both the plate transformer and the Powerstat on the same chassis would render it unmovable!

The power supply bleeder-resistor network

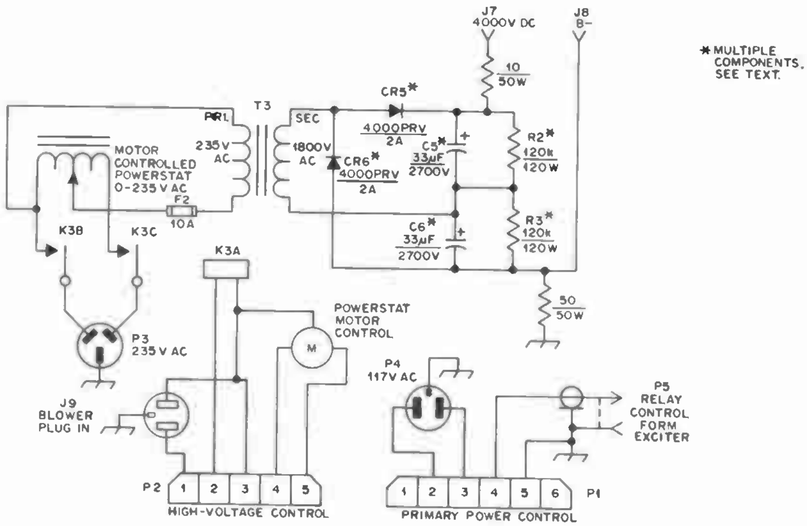


must be placed in a position to allow air to flow past it and rise through the top screen cover. R2 and R3 each consist of six resistors (12 total) rated at 20-kΩ and 20 watts each. A similar bank of 12 capacitors connected in series constitutes the filter network. Each filter capacitor in the circuit has one of the bleeder resistors connected directly across it in order to assure equal voltage division. Each capacitor is rated at 200 µF and 450 volts.

CR5 and CR6 consist of series diodes similar in hookup to the capacitor and resistor network described above. Each individual diode is rated at 1000 volts; the four series connected equal 4000 PRV. The current rating for each diode is two amperes.

No provisions have been made to operate this plate supply from a 117-volt ac source. Accordingly, if one wishes to have such capability, a suitable transformer must be substituted for T3 shown in Fig. 1. It should be pointed out, however, that the plate supply is the only portion to operate with a 234-volt line. The amplifier filament circuit and voltage-source circuits for the relays along with the blower all operate from 117 V ac.

The amplifier portion of this system is constructed on an aluminum chassis which is 14 × 17 × 4 inches. The 14-inch dimension was chosen as the front-panel side to conserve space on the operating table. A bottom cover for the chassis is cut from a large section of aluminum perforated stock while the area above the chassis top is completely sealed and made airtight. The amplifier top cover is solid stock but has a four-inch flange (stovepipe material) mounted directly above the 8877 tube. Hot air is exhausted via this port using an external "blower" which has been set up to draw air rather than force it. The procedure is simple; just connect the four-inch hose coming from the pipe flange to the blower intake port. Place the blower exhaust outlet in a position so that it will not be restricted. The cold air is drawn in under the amplifier chassis, passes through the 8877 and socket, then out the stovepipe to the blower. The air from the blower is heated and



(B)

Fig. 1 — Circuit diagram for the 8877 two-kilowatt amplifier. Component designations not listed below are for text reference only.

C1 — 500-pF transmitting capacitor, 5000 volts (Centralab 858 series).
C2 — .001-µF transmitting capacitor, 5000 volts (Centralab 858 series).
C3 — Vacuum variable, 500 pF maximum, 7500 volts.
C4 — Vacuum variable, 1000 pF maximum, 5000 volts.
C5, C6 — Six 200-µF units (See text.).
K1 — Dpdt, 5-A contacts. Coil voltage is 12 volts dc.
K2 — Vacuum relay, spst. (Torr Electronics TFI or equiv.)
K3 — Power relay, 10-A contacts, 117-V ac coil.
L1 — 25 turns wound on an Amidon T-50-2 toroid core.
L2 — 125 µH, 2 A (Hammond Mfg.).
L3 — Strap-wound inductor, 12 µH total tapped at 2-1/2 turns for 10 meters, 3 turns for 15 meters, 5 turns for 20 meters and 14 turns for 40 meters. (E. F. Johnson 232-626).
L4 — 5 turns of 1/4-inch copper tubing wound the same diameter as L3.
S1, S3, S4 — spst, 3 A (Radio Shack).
S2 — spdt, 3 A with spring return to center off position (Radio Shack).
T1 — primary 117 V ac, secondary 5.0 V ac, 10 A. (Hammond Transformer).
T2 — primary 117 V ac, secondary 12 V ac at 3 A.
T3 — plate transformer, 234-volt ac primary, 1770-volt (Hammond Transformer 105677).

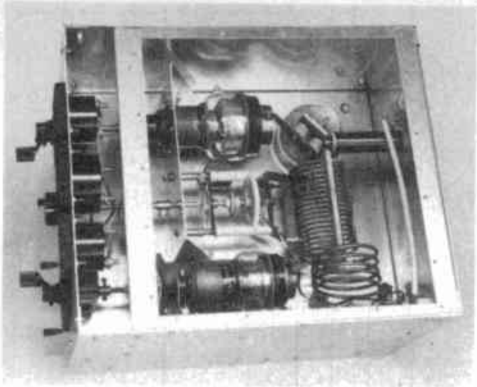
should not be directed at anything which might run normally warm (power supply components). In fact, the warm air may either be directed out of the radio shack in the event the heat is undesirable or the heat may be applied to one's feet in the winter season if the system is being used in an unheated basement atmosphere. Of course, one of the key features of a solid shield enclosure for the rf compartment is the reduction of unwanted radiation of fundamental or spurious energy which could cause TVI.

Care must be given to the mechanical installation of the high-voltage connectors and the cable used to transport 3600 volts of dc from the power supply to the anode of the 8877. Millen high-voltage connectors were used throughout. One problem developed during the testing phase of this project. A steel screw and nut were used to mount one of the connectors and apparently the Bakelite material cracked during installation. A discharge path developed across the crack creating loud noises and popping fuses. Nylon or Teflon hardware is recommended for mounting the Millen connectors if voltages above 3000 are anticipated.

Operation

Since this project is one which should not be undertaken by an inexperienced builder, some of the basic steps of pretest will not be discussed in detail here. Suffice it to say that ordinary primary voltage checks and switching should be confirmed as being in correct working order before placing primary power to the high-voltage supply. The 3000-volt circuit *must* be treated with respect – it can seriously injure or even kill a person coming in contact with it! Operation of the motor-controlled Powerstat can be determined by operating the system on the 117-volt primary, leaving the plug to the 234-volt line disconnected.

A word of caution: *The air flow system must always be used when any power is applied to the 8877 – even filament.* And rf drive power should



Top view of the 8877 amplifier. The three meters are separated from the rf compartment with an aluminum shield.

never reach the 8877 *unless* that tube has plate voltage applied. Of course, if one applies plate voltage and drive, he should be prepared to dissipate the power output from the amplifier into a dummy load of suitable rating.

One particular disadvantage of having the amplifier completely enclosed in a solid shield is the inability of the operator to visually spot any arc or component failure. During the initial testing of this amplifier, occasionally an arc would occur. While the arc was audible, the operator had to inspect the inner compartment very carefully to determine the cause of the malady. In fact, the arc had to be "encouraged" to a point where damage was easily identified!

Actual operation of the amplifier is quite simple. A feature which simplifies tuneup is the use of turn-counting dials for both the plate tuning and plate loading. Once the proper tuning has been established, one can log the numbers and return to them anytime. It is quite easy to touch up the dial settings for proper operating conditions once the approximate settings have been determined.

The coil-tap positions shown in the caption for Fig. 1 are given for proper operation at 3000 volts. Slightly better efficiency is possible by increasing the plate voltage to 3600 for two-kilowatt PEP ssb operation. The same is true of lowering the voltage to 2500 during cw conditions. The actual plate current to which the amplifier is driven should be determined in conjunction with the full-load plate voltage.

For cw tuneup and operation, the amplifier should be adjusted for maximum output power (usually determined using an externally mounted rf wattmeter) while maintaining proper grid current under conditions of one kilowatt input as determined by the combination of plate current and voltage. The proper settings will have been found when the plate meters indicate one-kilowatt input, the grid current shows 40 mA, and maximum output power occurs in conjunction with a "dip" in plate current and a "peak" in grid current all at the same time as the plate tuning control is adjusted. A condition of high grid current is usually a result of insufficient loading or too much drive power. If a low grid current condition exists and loading control decrease doesn't correct it, more drive power is indicated.

Tuneup for single sideband at the two-kilowatt level can be done only during dummy load conditions because it requires key-down conditions in excess of the legal-limit power restrictions. There is no way to tune this amplifier into an antenna at reduced power input and then drive it up to the two-kilowatt rated input point. The procedure for adjustment is identical to the one described above for cw operation. The one exception is plate-power input as indicated by the meters should be two kilowatts. Then, when the proper settings have been determined, the ssb drive signal is adjusted so that peak readings of plate current show about one half of that shown for key-down operation.

The lack of a tuned-input circuit solves several problems normally encountered when constructing an amplifier. The main advantage is that there need not be two band-switch decks with long leads (or even worse, two band switches!) nor space given to the inductors and capacitors. The driving impedance of the 8877 is very nearly 50 ohms and requires very little power to drive it to full power input. The tuned-circuit characteristics would reduce the drive requirement even further and should be considered by anyone wishing to use a 20-watt driver. For those amateurs using modern-day exciters in the 100-watt output class, some reduction in exciter gain control may be necessary. Exciters with more than about 150 watts of output power available should not be used without due consideration being given to an attenuator. The measured power required for this particular amplifier and tube shown was 50 watts to achieve a kilowatt on cw and about 70 watts for two kilowatts (this was in conjunction with a plate voltage change between modes) for ssb service. Slightly more drive power was required on 10 meters.

Problems

Some of the difficulties which come with making an amplifier of this category operate correctly are worthy of mention. First, as discussed earlier, some components which seemed adequate for 2500-volt systems failed when 3600 volts of dc was applied. The Millen connectors were one example. While the failure was not the fault of the component, care must be given to the maintenance of insulation integrity. Rf voltages developed in the tank circuit of this unit are substantial. Several problems with the band-switch contacts were encountered. Even though the plate output inductor has unused turns shorted out by the band switch, the shorted section of the coil can (and most likely will) have extremely high voltages present on it. In order to solve an arcing problem one of the band-switch tap positions had to be swapped with another. The band-switch markings on the front panel are not sequential. A solution to the problem would no doubt come if one were to use a continuously shorting switch.

Another difficulty which plagued this constructor was the propensity for the amplifier to show large amounts of negative grid current on 80 meters along with erratic plate current and poor efficiency. The calculated value for L2 is 95 μH. The original choke used at this point measured 87 μH even though an inductance of 95 was called for. After many hours of tank-circuit troubleshooting, it was decided to change the plate choke. The new one had an inductance of 125 μH and the 80-meter problems disappeared.

Collecting the Parts

Most of the component parts used in the project were purchased from surplus dealers or were donated by the manufacturer. Special thanks to James Millen Inc., Hammond Transformer Corporation, Simpson Electric, and Eimac for keeping the total cost of the parts to well within the allowable limits of any ARRL Lab project.

AN AMPLIFIER FOR QRP TRANSCEIVERS

The circuit of Fig. 1 shows a 15-watt output plug-in amplifier suitable for use with the HW-7 transceiver, or with any 2-watt class rig designed for hf-band work. The nucleus of the circuit is Motorola's new MRF-line transistor, the 449A. Unit cost is $13 for the part, and it will deliver up to 30 watts of output to 30 MHz as a Class C amplifier. Operating voltage is 13.0 dc. Rf drive requirements for full output are under 1 watt. Another member of this transistor family is the MRF450A, which will provide 50 watts of output with 2 watts of drive. It costs $16.50 according to a quote from an East Coast supplier.

Table 1 lists *L* and *C* values for 15 watts of output. The network is based on a loaded *Q* of 4. The X_L and X_C values given in Fig. 1 can be used to obtain inductance and capacitance values for frequencies other than 7, 14, and 21 MHz.

A 50-ohm, 3-dB attenuator pad is used at the amplifier input to assure less than 1 watt of drive without the need to modify the output stage or driver of the HW-7. The base shunting resistors of Q1 consume additional drive power while aiding amplifier stability. Another advantage of the attenuator is that it provides a resistive termination for the HW-7 across its operating range – 7 to 21 MHz. Without the attenuator the complex input impedance of Q1 would be reflected through T1, and the reactance seen by the HW-7 could be troublesome. A complex X_L and X_C condition exists at the input of a power transistor, and reactance amounts vary with operating frequency.

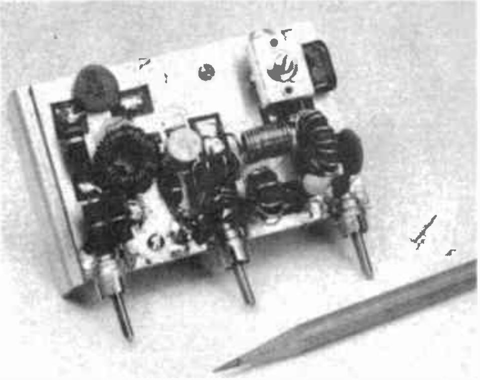
Circuit Notes

A *conventional* broadband toroidal transformer is used at T1. Equal performance was noted when comparing this transformer to *transmission-line* transformers. The latter consisted of two 4:1 transformers in cascade, effecting the desired 16:1 transformation ratio. A conventional transformer requires but one toroid core, and it is easy to build. For that reason it is specified here.

RFC1 serves as a low-*Q* collector choke, RFC2 is used as a decoupling choke to prevent rf energy from entering the HW-7 via the 13-volt line.

Performance

Motorola rates the MRF449A at 50-percent efficiency. Our lab findings bring that figure closer to 60 percent at 15-watts output. The output waveform from the circuit of Fig. 1 is exceptionally "sanitary." No distortion could be seen on the sine wave, as viewed on a 50-MHz scope, while delivering 15 watts into a 50-ohm dummy load.



Shown here is the assembled amplifier as originally described in *QST* for December, 1975.

TABLE I

Band	*L1*	*L2*	*C1*	*C2*
7 MHz	0.6 μH, 13 T No. 22 enam.,5/16" ID, no core	1.1 μH, 14 T No. 22 enam. on T-68-2 toroid core	450-pF mica trimmer	820-pF silver mica
14 MHz	0.3 μH, 8 T No. 22 enam., 5/16" ID, no core.	0.55 μH, 9 T No. 22 enam.. on T-68-6 toroid core	450-pF mica trimmer	220-pF silver mica
21 MHz	0.19 μH, 5 T No. 20 enam., 5/16" ID, no core	0.39 μH, 6 T No. 22 enam., on T-68-6 toroid core	450-pF mica trimmer	None

L1 coils are airwound. L2 coils are on Amidon toroid cores.

Harmonic energy was at least 40 dB below carrier level.

Short test periods were established with the amplifier output port shorted and open (30 seconds maximum), and no damage to Q1 resulted. It is stressed, however, that the amplifier should *always* have a 50-ohm termination during operation to assure proper performance and transistor longevity. Gain will be 8.7 dB at 15-watts output (HW-7 2-watt reference). It can be seen that a significant improvement in signal readability will result from using the amplifier when band conditions are poor.

Final Comments

Three phono plugs are soldered to the amplifier pc board. Mating phono jacks (J1, J2, and J3) are located on the rear panel of the HW-7. The coaxial cable between the HW-7 PA and the antenna relay is opened to permit insertion of the amplifier. The HW-7 PA output is routed to P1 of Fig. 1, and the amplifier output is fed into the HW-7, then to the antenna relay, through P3. To reinstate the HW-7, simply jumper J1 and J3 with a short length of 50-ohm coaxial cable. P2 and J2 permit the operator to obtain 13.0 operating volts for the amplifier from inside the HW-7, if desired. A 3-A

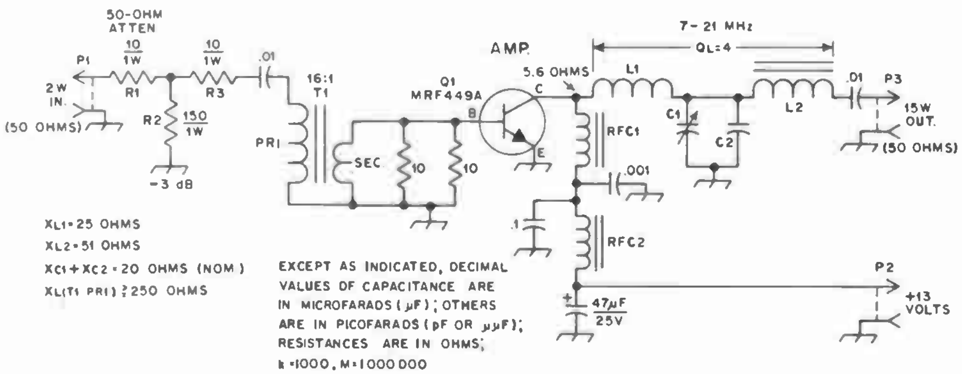


Fig. 1 — Schematic diagram of the 15-watt amplifier. Fixed-value capacitors are disk ceramic unless otherwise noted. Resistors are 1/2-watt composition unless specified differently. The 47-μF capacitor can be electrolytic or tantalum.

C1 — 450-pF Mica compression trimmer (Arco-ElMenco 466 or equivalent).

C2 — See Table I.

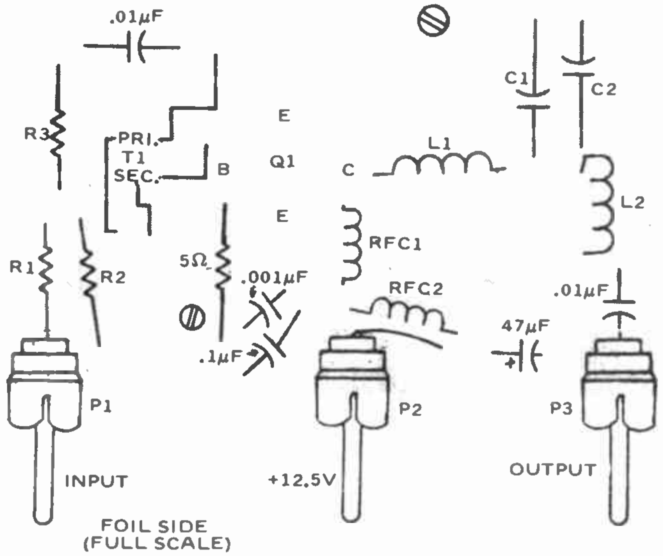
L1, L2 — See Table I.

P1-P3, incl. — Phono plugs soldered to edge of pc board.

Q1 — Motorola MRF449A strip-line stud transistor.

RFC1, RFC2 — 7 turns No. 20 enam. wire on 0.5-inch OD toroid ferrite core with 125 permeability (Amidon Assoc. FT-50-61 core or equiv.), 3μH.

T1 — Primary, 32 turns No. 24 enam. on Amidon T-68-2 core (7μH). Secondary, 8 turns No. 24 over primary winding.



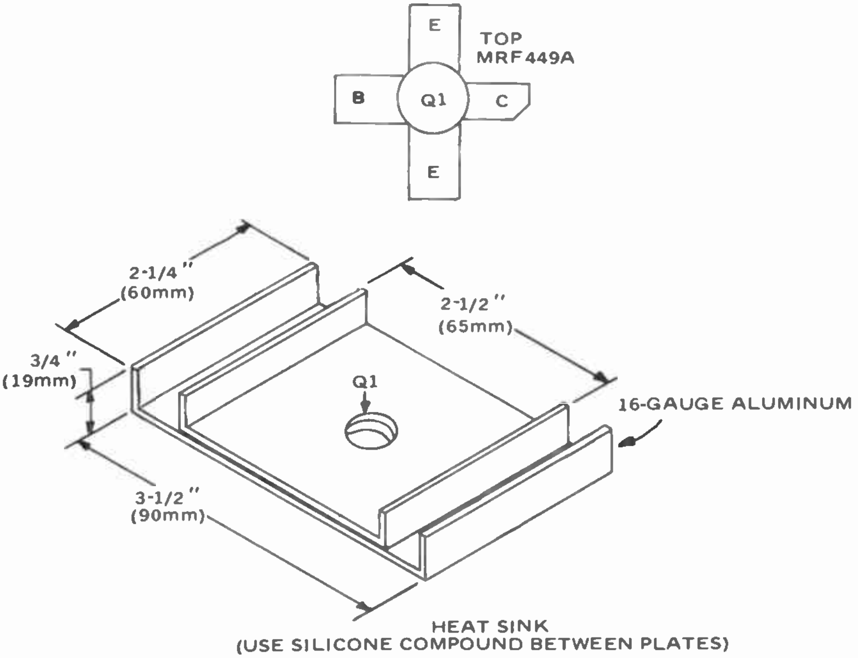


Fig. 2 — Scale layout of the pc board. Heat-sink dimensions are given in English and metric. The heat sink is held to the double-clad pc board by means of the transistor stud and two 4-40 screws and nuts. A substantially larger heat sink should be used if the amplifier is revised for 30 watts of output.

regulated supply is recommended for the *overall* system when running 15 watts of output.

Details of the pc-board layout and heat sink are given in Fig. 2. The transistor body must make firm contact with the heat sink. Silicone heat-sink compound should be used between Q1 and the heat sink. Tighten the transistor stud nut *with care* lest the stud be broken off.

Tune up in the center of the cw band by adjusting C1 to provide maximum rf output into a 50-ohm load. Amplifier bandwidth will be sufficient for all of each cw band.

If a band-switched version of the circuit is desired, build the amplifier in a separate box and use a two-pole, three-position, ceramic-insulated wafer switch to select the T networks. The leads from the switch to the networks and Q1 must be kept short to preserve the network characteristics. Switched-lead inductances will become part of the network, so they must be kept to minimum lengths.

Detailed network design information can be obtained from the *QST* beginner's series, "Learning to Work with Semiconductors," April-October, inclusive, 1975. Those wishing to operate the amplifier at 30-watts output can redesign the T network to match a 2.8-ohm collector impedance, using the equations in the beginner's series.

VHF and UHF Transmitting

Before planning operation on the frequencies above 50 MHz, we should understand the FCC rules, as they apply to the bands we are interested in. The necessary information is included in the allocations table in the first chapter of this *Handbook* and in *The Radio Amateur's License Manual*, but some points will bear emphasis here.

Standards governing signal quality in the 50-MHz band are the same as for all lower amateur frequencies. Frequency stability, modulation, keying characteristics, and freedom from spurious products must be consistent with good engineering practice. Simultaneous amplitude and frequency modulation is prohibited. These standards are not *imposed by law* on amateur frequencies from 144 MHz up. This is not to say that we should not strive for excellence on the higher bands, as well as on 50 MHz, but it is important to remember that we may be cited by FCC for failing to meet the required standards in 50-MHz work.

A sideband signal having excessive bandwidth, an a-m signal whose frequency jumps when modulation is applied, an fm signal that is also amplitude-modulated, a cw signal with excessive keying chirp or objectionable key clicks – any of these is undesirable on any band, but they are all *illegal* on 50 MHz. Any of them could earn the operator an FCC citation in 50-MHz work. And misinterpretation of these points in an FCC examination could cost the would-be amateur his first ticket.

The frequencies above 50 MHz were once a world apart from the rest of amateur radio, in equipment required, in modes of operation and in results obtained. Today these worlds blend increasingly. Thus, if the reader does not find what he needs in these pages to solve a transmitter problem, it will be covered in the hf transmitting chapter. This chapter deals mainly with aspects of transmitter design and operation that call for different techniques in equipment for 50 MHz and up.

DESIGNING FOR SSB AND CW

Almost universal use of ssb for voice work in the hf range has had a major impact on equipment design for the vhf and even uhf bands. Many amateurs have a considerable investment in hf sideband gear. This equipment provides accurate frequency calibration and good mechanical and electrical stability. It is effective in cw as well as ssb communication. These qualities being attractive to the vhf operator, it is natural for him to look for ways to use his hf gear on frequencies above 50 MHz.

Thus increasing use is being made of vhf accessory devices, both ready made and home-built. This started years ago with the vhf converter, for receiving. Rather similar conversion equipment for transmitting has been widely used since ssb began taking over the hf bands. Today the hf trend is to one-package stations, called transceivers. The obvious move for many vhf men is a companion box to perform both transmitting and receiving conversion functions. Known as transverters, these are offered by several transceiver manufacturers. They are also relatively simple to build, and are thus likely projects for the home-builder of vhf gear.

Transverter vs. Separate Units

It does not necessarily follow that what is popular in hf work is ideal for vhf use. Our bands are wide, and piling-up in a narrow segment of a band, which the transceiver encourages, is less than ideal use of a major asset of the vhf bands – spectrum space. Separate ssb exciters and receivers, with separate vhf conversion units for transmitting and receiving, tend to suit our purposes better than the transceiver-transverter combination, at least in home-station service.

Future of Other Modes

It should not be assumed that ssb will monopolize voice work in the world above 50 MHz in the way that it has the amateur voice frequencies below 29 MHz. Sideband is unquestionably far superior to other voice modes for weak-signal DX work, but where there is plenty of room, as there is in all vhf and higher bands, both amplitude and frequency modulation have merit. A low-powered a-m transmitter is a fine construction project for a vhf beginner, and fm has been gaining in popularity rapidly in recent years. A reprint of a very popular 4-part *QST* series describing a complete two-band vhf station for the beginner is available from ARRL for 50 cents.

The decline in use of amplitude modulation has been mainly in high-powered stations. The heavy-iron modulator seems destined to become a thing of the past, but this should not rule out use of a-m. Many ssb transceivers are capable of producing high-quality a-m, and one linear amplifier stage can build as little as 2 watts a-m output up to 200 watts or so, with excellent voice quality, if the equipment is adjusted with care. It should be remembered that the transmitting converter (or heterodyne unit as it is often called) is not a sideband device only. It will serve equally well with a-m, fm or cw drive.

THE OSCILLATOR-MULTIPLIER APPROACH

Where modes other than ssb are used, most vhf transmitters have an oscillator, usually in the hf range, one or more frequency multiplier stages, and at least one amplifier stage. The basics of this type of transmitter are well covered in the preceding chapter, so only those aspects of design that are of special concern in vhf applications will be discussed here.

Oscillators

Because any instability in the oscillator is multiplied along with the frequency itself, special attention must be paid to both mechanical and electrical factors in the oscillator of a vhf transmitter. The power source must be pure dc, of unvarying voltage. The oscillator should run at low input, to avoid drift due to heating. Except where fm is wanted, care should be taken to isolate the oscillator from the modulated stage or stages.

Crystal oscillators in vhf transmitters may use either *fundamental* or *overtone* crystals. The fundamental type is normally supplied for frequencies up to 18 MHz. For higher frequencies the overtone type is preferred in most applications, though fundamental crystals for up to about 30 MHz can be obtained on order. The fundamental crystal oscillates on the frequency marked on its holder. The marked frequency of the overtone type is approximately an odd multiple of its fundamental frequency, usually the third multiple for frequencies between 12 and 54 MHz, the fifth for roughly 54 to 75 MHz, and the seventh or ninth for frequencies up to about 150 MHz. Crystals are seldom used for direct frequency control above about 75 MHz in amateur work, though crystals for 144-MHz oscillation can be made.

Most fundamental crystals can be made to oscillate on at least the third overtone, and often higher, with suitable circuits to provide feedback at the desired overtone frequency. Conversely, an overtone crystal is likely to oscillate on its fundamental frequency, unless the tuned circuit is properly designed. An overtone crystal circuit should be adjusted so that there is no oscillation at or near one-third of the frequency marked on the holder, nor should there be energy detectable on the *even* multiples of the fundamental frequency.

It should be noted that the overtone is not necessarily an exact multiple of the fundamental. An 8000-kHz fundamental frequency does not guarantee overtone oscillation on 24.000 MHz, though it may work out that way in some circuits, with some crystals. Overtone crystals can also be made to oscillate on other overtones than the intended one. A third-overtone 24-MHz crystal can be used for its fifth overtone, about 40 MHz, or its seventh, about 56 MHz, by use of a suitable tuned circuit and careful adjustment of the feedback.

Variable-frequency oscillators are in great demand for vhf-transmitter frequency control, but except where heterodyning to a higher frequency is used, as opposed to frequency multiplication, the VFO is generally unsatisfactory. Small instabilities, hardly noticeable in hf work, are multiplied to unacceptable proportions in the oscillator-multiplier type of transmitter. The fact that many such unstable VFO rigs are on the air, particularly on 6 meters, does not make them desirable, or even *legal*. Only careful attention to all the fine points of VFO design and use can result in satisfactory stability in vhf transmitters.

Frequency Multipliers

Frequency multiplication is treated in Chapter 6. The principal factor to keep in mind in multipliers for the vhf bands is the probability that frequencies other than the desired harmonics will be present in the output. These can be sources of TVI in vhf transmitters. Examples are the 9th harmonic of 6 MHz and the 7th harmonic of 8 MHz, both falling in TV Channel 2. The 10th harmonic of 8-MHz oscillators falling in Channel 6 is a similar problem. These unwanted multiples can be held down by the use of the highest practical degree of selectivity in interstage coupling circuits in the vhf transmitter, and by proper shielding and interstage impedance matching. This last is particularly important in transistor frequency multipliers and amplifiers. More on avoiding TVI will be found later in this chapter, and in the chapter on interference problems.

The varactor multiplier (see Chapter 4) is much used for developing power in the 420-MHz band. Requiring no power supply, it uses only driving power from a previous stage, yet quite high orders of efficiency are possible. Two examples are shown later in this chapter. A 220-MHz exciter tuned down to 216 MHz makes a good driver for a 432-MHz varactor doubler. More commonly used is a tripler such as the one described in this chapter, using 144-MHz drive. The output of a varactor multiplier tends to have appreciable amounts of power at other frequencies than the desired, so use of a strip-line or coaxial filter is recommended, whether the multiplier drives an amplifier or works into the antenna directly.

AMPLIFIER DESIGN AND OPERATION

Amplifiers in vhf transmitters all once ran Class C, or as near thereto as available drive levels would permit. This was mainly for high-efficiency cw, and quality high-level amplitude modulation. Class C is now used mostly for cw or fm, and in either of these modes the drive level is completely uncritical, except as it affects the operating efficiency. The influence of ssb techniques is seen clearly in current amplifier trends. Today Class AB_1 is popular and most amplifiers are set up for linear amplification, for ssb and – to a lesser extent – a-m. The latter is often used in connection with small amplitude-modulated vhf transmitters, having their own built-in audio equipment. Where a-m output is available from the ssb exciter, it is also useful with the Class AB_1 linear amplifier, for only a watt or two of driver output is required.

There is no essential circuit difference between the AB_1 linear amplifier and the Class-C amplifier;

only the operating conditions are changed for different classes of service. Though the plate efficiency of the AB_1 linear amplifier is low in a-m service, this type of operation makes switching modes a very simple matter. Moving toward the high efficiency of Class C from AB_1, for cw or fm service, is accomplished by merely raising the drive from the low AB_1 level. In AB_1 service the efficiency is typically 30 to 35 percent. No grid current is ever drawn. As the grid drive is increased, and grid current starts to flow, the efficiency rises rapidly. In a well-designed amplifier it may reach 60 percent, with only a small amount of grid current flowing. Unless the drive is run well into the Class C region, the operating conditions in the amplifier can be left unchanged, other than the small increasing of the drive, to improve the efficiency available for cw or fm. No switching or major adjustments of any kind are required for near-optimum operation on ssb, a-m, fm or cw, if the amplifier is designed primarily for AB_1 service. If high-level a-m were to be used, there would have to be major operating-conditions changes, and very much higher available driving power.

Tank-Circuit Design

Except in compact low-powered transmitters, conventional coil-and-capacitor circuitry is seldom used in transmitter amplifiers for 144 MHz and higher frequencies. U-shaped loops of sheet metal or copper tubing, or even copper-laminated circuit board, generally give higher Q and circuit efficiency at 144 and 220 MHz. At 420 MHz and higher, coaxial tank circuits are effective. Resonant cavities are used in some applications above 1000 MHz. Examples of all types of circuits are seen later in this chapter. Coil and capacitor circuits are common in 50-MHz amplifiers, and in low-powered, mobile and portable equipment for 144 and even 220 MHz.

Stabilization

Most vhf amplifiers, other than the grounded-grid variety, require neutralization if they are to be satisfactorily stable. This is particularly true of AB_1 amplifiers, which are characterized by very high power sensitivity. Conventional neutralization is discussed in Chapter 6. An example is shown in Fig. 7-1A.

A tetrode tube has some frequency where it is inherently neutralized. This is likely to be in the lower part of the vhf region, for tubes designed for hf service. Neutralization of the opposite sense may be required in such amplifiers, as in the example shown in Fig. 7-1B.

Conventional screen bypassing methods may be ineffective in the vhf range. Series-tuning the screen to ground, as in 7-1C, may be useful in this situation. A critical combination of fixed capacitance and lead length may accomplish the same result. Neutralization of transistorized amplifiers is not generally practical, at least where bipolar transistors are used.

Parasitic oscillation can occur in vhf amplifiers, and, as with hf circuits, the oscillation is usually at a frequency considerably higher than the operating frequency, and it cannot be neutralized out. Usually it is damped out by methods illustrated in Fig. 7-2. Circuits A and B are commonly used in 6-meter transmitters. Circuit A may absorb sufficient fundamental energy to burn up in all but low-power transmitters. A better approach is to use

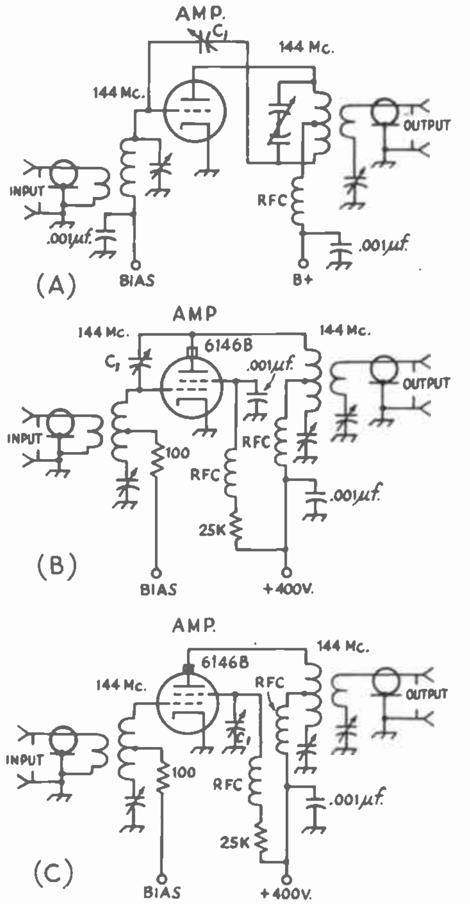


Fig. 7-1 — Representative circuits for neutralizing vhf single-ended amplifiers. The same techniques are applicable to stages that operate in push-pull. At A, C1 is connected in the manner that is common to most vhf or uhf amplifiers. The circuits at B and C are required when the tube is operated above its natural self-neutralizing frequency. At B, C1 is connected between the grid and plate of the amplifier. Ordinarily, a short length of stiff wire can be soldered to the grid pin of the tube socket, then routed through the chassis and placed adjacent to the tube envelope, and parallel to the anode element. Neutralization is effected by varying the placement of the wire with respect to the anode of the tube, thus providing variable capacitance at C1. The circuit at C is a variation of the one shown at B. It too is useful when a tube is operated above its self-neutralizing frequency. In this instance, C1 provides a low-Z screen-to-ground path at the operating frequency. RFC in all circuits shown are vhf types and should be selected for the operating frequency of the amplifier.

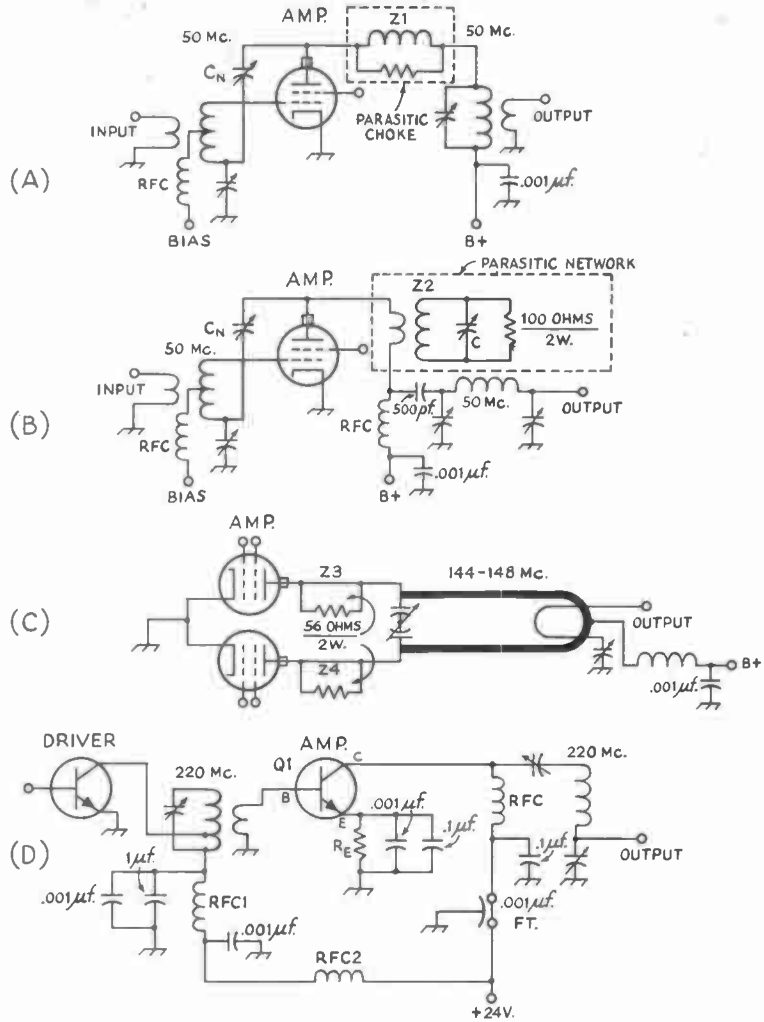


Fig. 7-2 — Representative circuits for vhf parasitic suppression are shown at A, B, and C. At A, Z1 (for 6-meter operation) would typically consist of 3 or 4 turns of No. 14 wire wound on a 100-ohm 2-watt non-inductive resistor. Z1 overheats in all but very low power circuits. The circuit at B, also for 6-meter use, is more practical where heating is concerned. Z2 is tuned to resonance at the parasitic frequency by C. Each winding of Z2 consists of two or more turns of No. 14 wire — determined experimentally — wound over the body of a 100-ohm 2-watt (or larger) noninductive resistor. At C, an illustration of uhf parasitic suppression as applied to a 2-meter amplifier. Noninductive 56-ohm 2-watt resistors are bridged across a short length of the connecting lead between the tube anode and the main element of the tank inductor, thus forming Z3 and Z4.

The circuit at D illustrates how bypassing for both the operating frequency and lower frequencies is accomplished. Low-frequency oscillation is discouraged by the addition of the 0.1-μF disk ceramic capacitors. RFC1 and RFC2 are part of the decoupling network used to isolate the two stages. This technique is not required in vacuum-tube circuits.

the selective circuit illustrated at B. The circuit is coupled to the plate tank circuit and tuned to the parasitic frequency. Since a minimum amount of the fundamental energy will be absorbed by the trap, heating should no longer be a problem.

At 144 MHz and higher, it is difficult to construct a parasitic choke that will not be resonant at or near the operating frequency. Should uhf parasitics occur, an effective cure can often be realized by shunting a 56-ohm 2-watt resistor across a small section of the plate end of the tuned circuit as shown in Fig. 7-2, at C. The resistor should be attached as near the plate connector as practical. Such a trap can often be constructed by bridging the resistor across a portion of the flexible strap-connector that is used in some transmitters to join the anode fitting to the plate-tank inductor.

Instability in solid-state vhf and uhf amplifiers can often be traced to oscillations in the lf and hf regions. Because the gain of the transistors is very high at the lower frequencies, instability is almost certain to occur unless proper bypassing and decoupling of stages is carried out. Low-frequency oscillation can usually be cured by selecting a bypass-capacitor value that is effective at the frequency of oscillation and connecting it in parallel with the vhf bypass capacitor in the same part of the circuit. It is not unusual, for example, to employ a 0.1-μF disk ceramic in parallel with a .001-μF disk capacitor in such circuits as the emitter, base, or collector return. The actual values used will depend upon the frequencies involved. This technique is shown in Fig. 7-2D. For more on transmitter stabilization, see Chapter 6.

TIPS ON AB_1 LINEAR AMPLIFIERS

As its name implies, the function of a linear is to amplify an amplitude-modulated signal in a manner so that the result is an exact reproduction of the driving signal. (Remember, ssb is a form of amplitude modulation.) The nature of the a-m signal with carrier is such that linear amplification of it is inherently an inefficient process, in terms of power input to power output, which is the conventional way of looking at amplifier efficiency. But when all factors are considered, particularly the very small exciter power required and elimination of the cumbersome and expensive high-level plate-modulation equipment, "efficiency" takes on a different meaning. Viewed in this way, the Class-AB_1 a-m linear has only two disadvantages: it is incapable of providing as much power output (within the amateur power limit of 1 kW) as the high-level-modulated amplifier, and it requires considerable skill and care in adjustment.

The maximum plate efficiency possible with an AB_1 a-m linear is about 35 percent. The power output in watts that is possible with a given amplifier tube is roughly half its rated plate dissipation. If the first factor is exceeded the result is poor quality and splatter. If the second is ignored, the tube life is shortened markedly.

There being no carrier to worry about in ssb operation, the linear amplifier can run considerably higher efficiency in amplifying ssb signals, and the popularity of ssb has brought the advantages of the linear amplifier for all classes of service into focus. The difference between a-m with carrier and ssb without carrier, in the adjustment of a linear, is mainly a matter of the drive level. Drive can never be run up to the point where the stage begins to draw grid current, but it can run close with ssb, whereas it must be held well below the grid-current level when the carrier is present.

With a-m drive the plate and screen currents must remain steady during modulation. (The screen current may be negative in some amplifiers, so observation of it is simpler if the screen-current meter is the zero-center type.) The plate, screen and grid meters are the best simple indicator of safe AB_1 operation, but they do not show whether or not you are getting all you can out of the amplifier. The signal can be monitored in the station receiver, if the signal in the receiver can be held below the point at which the receiver is overloaded. Cutting the voltage from a converter amplifier stage is a good way to do this. But the only way to know for sure is to use an oscilloscope.

One that can be used conveniently is the Heath Monitor Scope, any version. Some modification of the connections to this instrument may be needed, to prevent excessive rf pickup and resultant pattern distortion, when using it for vhf work. Normally a coupling loop within the scope, connected between two coaxial fittings on the rear of the instrument, is used. The line from the transmitter to the antenna or dummy load runs through these two fittings. For vhf service, a coaxial T fitting is connected to one of these terminals, and the line is run through it, only. With full power it may even be necessary to remove the center pin from the T fitting, to reduce the input to the scope still further, particularly in 144-MHz service.

Really effective adjustment of the linear amplifier, whether with ssb or a-m drive, involves many factors. The amplifier must be loaded as heavily as possible. Its plate and grid circuits must be tuned carefully for maximum amplifier output. (Detuning the grid circuit is *not* the way to cut down drive.) If the power level is changed, all operating conditions must be checked carefully again. Constant metering of the grid, screen and plate currents is very helpful. One meter, switched to the various circuits, is definitely not recommended. A relative-power indicator in the antenna line is a necessity.

All this makes it appear that adjustment of a linear is a very complex and difficult process, but with experience it becomes almost second nature, even with all the points that must be kept in mind. It boils down to keeping the amplifier adjusted for maximum power output, and the drive level low enough so that there is no distortion, but high enough so that maximum efficiency is obtained. Practice doing this with the amplifier running into a dummy load, and the process will soon become almost automatic. Your amateur neighbors (and perhaps TV viewers nearby, as well) will appreciate your cooperation!

About Driver Stages

If the amplifier is capable of reproducing the driving signal exactly, it follows that the driver quality must be above reproach. This is quite readily assured, in view of the low driving power required with the AB_1 linear. Only about two watts exciter power is needed to drive a grounded-cathode AB_1 linear of good design, so it is possible to build excellent quality and modulation charac-

Fig. 7-3 – The 6-meter transverter, with shield cover in place. Large knobs are for amplifier tuning and loading. Small knob, lower right is for a meter sensitivity control. The meter switch is just above it.

teristics into the a-m driver or ssb exciter. If this is done, and the amplifier is operated properly, the result can be a signal that will bring appreciative and complimentary reports from stations worked, on both a-m and ssb.

VHF TVI CAUSES AND CURES

The principal causes of TVI from vhf transmitters are as follows:

1) Adjacent-channel interference in Channels 2 and 3 from 50 MHz.

2) Fourth harmonic of 50 MHz in Channels 11, 12 or 13, depending on the operating frequency.

3) Radiation of unused harmonics of the oscillator or multiplier stages. Examples are 9th harmonic of 6 MHz, and 7th harmonic of 8 MHz in Channel 2; 10th harmonic of 8 MHz in Channel 6; 7th harmonic of 25-MHz stages in Channel 7; 4th harmonic of 48-MHz stages in Channel 9 or 10; and many other combinations. This may include i-f pickup, as in the cases of 24-MHz interference in receivers having 21-MHz i-f systems, and 48-MHz trouble in 45-MHz i-fs.

4) Fundamental blocking effects, including modulation bars, usually found only in the lower channels, from 50-MHz equipment.

5) Image interference in Channel 2 from 144 MHz, in receivers having a 45-MHz i-f.

6) Sound interference (picture clear in some cases) resulting from rf pickup by the audio circuits of the TV receiver.

There are other possibilities, but nearly all can be corrected completely, and the rest can be substantially reduced.

Items 1, 4 and 5 are receiver faults, and nothing can be done at the transmitter to reduce them, except to lower the power or increase separation between the transmitting and TV antenna systems. Item 6 is also a receiver fault, but it can be alleviated at the transmitter by using fm or cw instead of a-m phone.

Treatment of the various harmonic troubles, Items 2 and 3, follows the standard methods detailed elsewhere in this *Handbook*. It is suggested that the prospective builder of new vhf equipment familiarize himself with TVI prevention techniques, and incorporate them in new construction projects.

Use as high a starting frequency as possible, to reduce the number of harmonics that might cause trouble. Select crystal frequencies that do not have harmonics in TV channels in use locally. Example: The 10th harmonic of 8-MHz crystals used for operation in the low part of the 50-MHz band falls in Channel 6, but 6-MHz crystals for the same band have no harmonic in that channel.

If TVI is a serious problem, use the lowest transmitter power that will do the job at hand. Keep the power in the multiplier and driver stages at the lowest practical level, and use link coupling in preference to capacitive coupling. Plan for complete shielding and filtering of the rf sections of the transmitter, should these steps become necessary.

Use coaxial line to feed the antenna system, and locate the radiating portion of the antenna as far as possible from TV receivers and their antenna systems.

50-MHZ TRANSVERTER

With the increase in use of ssb on the vhf bands, there is much interest in adapting hf ssb gear to use on higher frequencies. The transverter of Fig. 7-3 will provide transceiver-style operation on 50 MHz, when used with a low-powered 28-MHz transceiver. The output of the transmitter portion is about 40 watts, adequate for much interesting work. It can be used to drive an amplifier such as the grounded-grid 3-500Z unit described later in this chapter. The receiving converter combines simplicity, adequate gain and noise figure, and freedom from overloading problems.

Circuit Details

The receiving front end uses a grounded-gate JFET rf amplifier, Q1 in Fig. 7-5, followed by a dual-gate MOSFET mixer, Q2. Its 22-MHz injection voltage is taken from the oscillator and buffer stages that also supply injection for transmitter mixing. The difference frequency is 28 MHz, so the transceiver dial reading bears a direct 28-50 relationship to the 50-MHz signal being received. For more detail on the converter construction and adjustment, see Fig. 9-9 and associated text. The transverter uses the grounded-gate rf amplifier circuit, while the converter referred to above has a grounded source, but they are quite similar otherwise.

The triode portion of a 6LN8, V1A, is a

22-MHz crystal oscillator. The pentode, V1B, is a buffer, for isolation of the oscillator, and increased stability. Injection voltage for the receiving mixer is taken from the buffer output circuit, L8, through a two-turn link, L9, and small-diameter coax, to gate 2 of the mixer, through a 10-pF blocking capacitor.

The grid circuit of the 6EJ7 transmitting mixer, V2, is tuned to 22 MHz and is inductively coupled to the buffer plate circuit. The 28-MHz input is applied to the grid circuit through a link around L11, and small-diameter coax. The mixer output, L12, is tuned to the sum frequency, 50 MHz, and coupled to a 6GK6 amplifier, V3, by a bandpass circuit, L12 and L13. The 6GK6 is bandpass-coupled to the grid of a 6146 output stage, V4. This amplifier employs a pi-network output stage.

The 6146 plate dissipation is held down during the receiving periods by fixed bias that is switched in by relay K1. The mixer and driver tubes have their screen voltage removed during receiving, by the same relay, which also switches the antenna and 28-MHz input circuits for transmitting and receiving. The relay is energized by grounding pin 7 of P1 through an external switch, or by the VOX relay in the transceiver.

Construction

A 7 X 9 X 2-inch aluminum chassis is used for the transverter, with a front panel 6 inches high, made of sheet aluminum. The top and sides are enclosed by a one-piece cover of perforated aluminum. The output-stage tuning control, C5, is on the upper left of the panel, 2 inches above the chassis. The loading control, C6, is immediately below, under the chassis. The meter, upper right, monitors either 6146 plate current or relative output, as selected by the switch, S1, immediately below it. A sensitivity control for calibrating the output-metering circuit completes the front-panel controls.

The output connector, J2, is centered on the rear apron of the chassis, which also has the input jack, J1, the 8-pin connector, P1, and the bias-adjusting control mounted on it.

The meter is a 1-mA movement, with multiplier resistors to give a full-scale reading on a current of 200 mA. The front cover snaps off easily, to allow calibration marks to be put on as desired.

An enclosure of perforated aluminum, 3 1/4 inches high, 4 inches wide and 4 3/4 inches long shields the 6146 and its plate circuit. There is also an L-shaped shield around the 6146 socket, under the chassis.

The receiving converter is built on a 2 1/2 X 4 1/4-inch etched board, and mounted vertically in a three-sided shield of sheet aluminum. Before mounting the converter shield, be sure to check for clearance with the terminals on the meter. Remember, the meter has full plate voltage on it when the switch is set to read plate current, even when the transverter is in the receiving mode.

Testing of the transverter was done with the General-Purpose Supply for Transceivers, described in the power supply chapter. Separate provision must be made for 12 volts dc for the receiving converter.

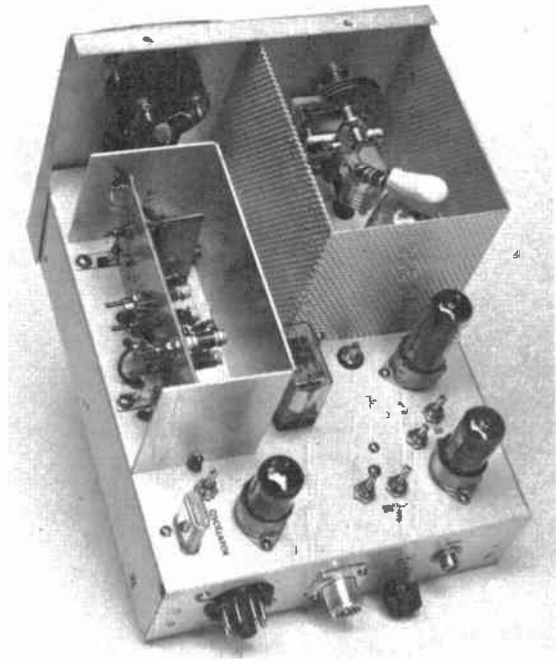


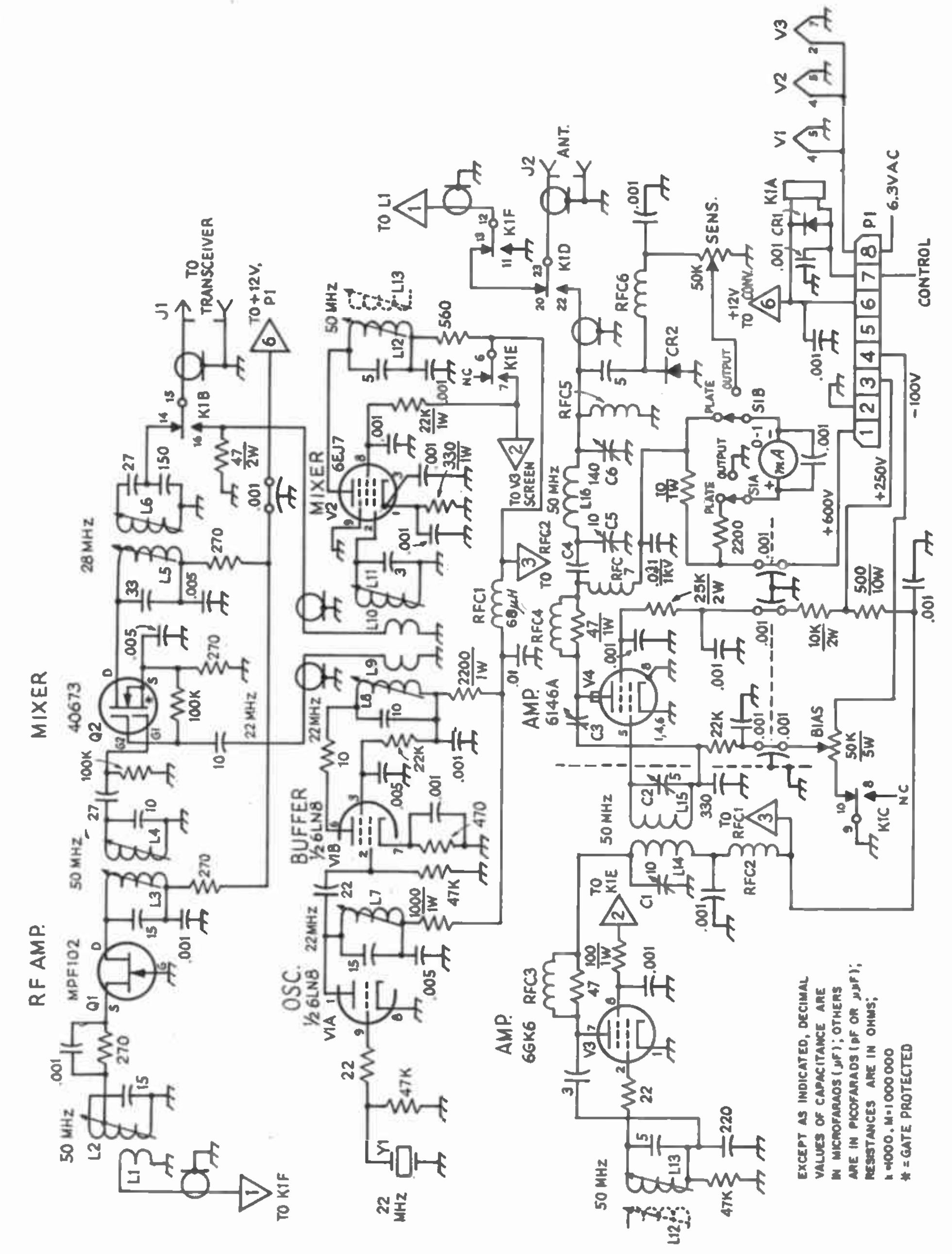
Fig. 7-4 – Top view of the transverter. The receiving converter is inside the shield at the left. The 22-MHz crystal oscillator and buffer are in the left rear portion of the chassis. In the right corner is the transmitting mixer. Above it is the first amplifier. The 6146 output amplifier is in the shielded compartment at the left front.

Injection voltage, signal input and i-f output connections to the converter are made with small-diameter coax. These and the 12-volt wiring are brought up through small holes in the chassis, under the converter. As seen in Fig. 9-11, the input JFET, Q1, is on the left. The mixer is near the center. The 28-MHz output coils, L5 and L6, are just to the right of Q2.

Note that there are two sets of relay contacts, K1D and K1F, in series in the receiver line. This guarantees high isolation of the receiver input, to protect the rf amplifier transistor. Another protective device is the diode, CR1, across the coil of the relay. If there are other relays external to this unit that use the same 12-volt supply, it is advisable to put diodes across their coils also. Spikes of several volts can be induced with making and breaking of the coil circuits.

Adjustment

A dip meter is very useful in the preliminary tuning. Be sure that L7 and L8 are tuned to 22 MHz and L12 and L13 are tuned to 50 MHz. The driver and output circuits should also be tuned to 50 MHz. Check to be sure that slug-tuned coils really tune *through* the desired frequency. Quite



often troubles are eventually traced to coils where the circuit is only approaching resonance as the core centers in the winding. Such a circuit will appear to work, but drive will be low, and spurious outputs will tend to be high. This is a common trouble in overtone oscillators, with slug-tuned coils.

Once the circuits have been set approximately, apply heater and plate voltage to the oscillator, and tune L7 for best oscillation, as checked with a wavemeter or a receiver tuned to 22 MHz. Connect a 28-MHz receiver to the input, J1, and apply dc to the converter. It should be possible to hear a strong local station or test signal immediately. Peak all coils for best reception, then stagger-tune L5 and L6 for good response across the first 500 kHz of the band.

Before applying plate voltage to the 6146, it is advisable to protect the tube during tuneup by inserting a 1500- or 2000-ohm 25-watt resistor in series with the plate supply. Connect a 50-ohm load to the output jack, and energize K1. Adjust

Fig. 7-5 – Schematic diagram and part information for the 50-MHz transverter.
C1 – 10-pF subminiature variable (Hammarlund MAC-10).
C2 – 5-pF subminiature variable (Hammarlund MAC-5).
C3 – 2 1/2-inch length No. 14 wire, parallel to and 1/4 inch away from tube envelope. Cover with insulating sleeve.
C4 – 500-pF 3000-volt disk ceramic.
C5 – 10-pF variable (Johnson 149-3, with one stator and one rotor plate removed).
C6 – 140-pF variable (Millen 22140).
CR1 – 1N128 diode.
CR2 – 1N83A diode.
J1 – Phono jack.
J2 – Coaxial jack, SO-239.
K1 – 6-pole double-throw relay, 12-volt dc coil.
L1 – 2 turns small insulated wire over ground end of L2.
L2, L3, L4 – 10 turns No. 24 enamel closewound on J. W. Miller 4500-4 iron-slug form.
L5, L6 – 12 turns No. 24 enamel on J. W. Miller 4500-2 iron-slug form.
L7, L8, L11 – Iron-slug coils adjusted for 4.1, 5.5 and 5.5 μH, respectively (Miller 4405).
L9, L10 – 2 turns small insulated wire over ground ends of L8 and L11.
L12, L13 – 1-μH iron-slug coil J. W. Miller 4403, 3 turns removed.
L14 – 7 turns No. 20, 1/2-inch dia, 1/2 inch long (B & W 3003).
L15 – Like L14, but 6 turns.
L16 – 6 turns No. 20, 5/8-inch dia, 3/4 inch long (B & W 3006).
P1 – 8-pin power connector.
RFC1 – 68-μH rf choke (Millen 34300).
RFC2 – 8.2-μH rf choke (Millen J-300).
RFC3 – 5 turns No. 22 on 47-ohm 1/2-watt resistor.
RFC4 – 4 turns No. 15 on 47-ohm 1-watt resistor.
RFC5, RFC6, RFC7 – 8.2-μH rf choke (Millen 34300).
S1 – Dpdt toggle.
Y1 – 22-MHz overtone crystal (International Crystal Co., Type EX).

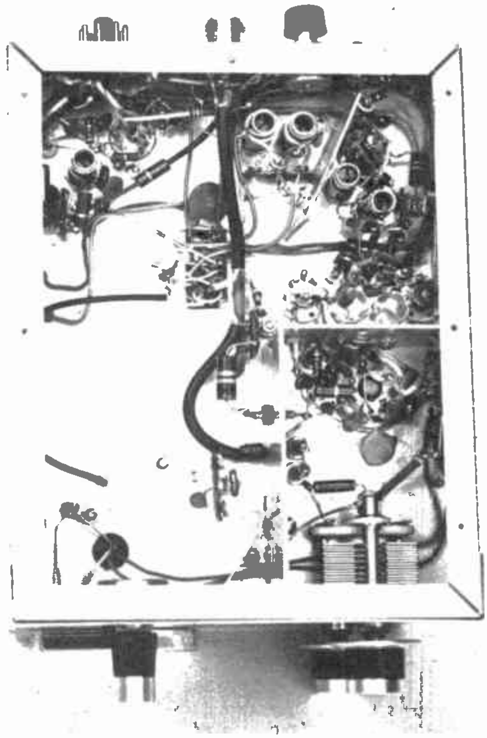


Fig. 7-6 – Bottom of the transverter, with the 6146 socket inside the shield compartment at the right. Three sets of inductively-coupled circuits are visible in the upper-right corner. The first two, near the top of the picture, are on 22 MHz. Next to the right and down, are the mixer plate and first-amplifier grid circuits. The self-supporting 6GK6 plate and 6146 grid coils are just outside the amplifier shield compartment. The large variable capacitor is the loading control.

the bias control for 25 to 30 mA plate current. Apply a small amount of 28-MHz drive. A fraction of a watt, enough to produce a dim glow in a No. 47 pilot lamp load, will do. Some output should be indicated on the meter, with the sensitivity control fully clockwise. Adjust the amplifier tuning and loading for maximum output, and readjust all of the 50-MHz circuits likewise.

After the circuits have been peaked up, adjust the bandpass circuits by applying first a 28.1-MHz input and then a 28.4-MHz input, and peaking alternate coils until good operation is obtained over the range of 50.0 to 50.5 MHz. Most ssb operation currently is close to 50.1 MHz, so uniform response across a 500-kHz range is not too important, if only this mode is used. If the 10-meter transceiver is capable of a-m operation, and you want to use this mode, coverage up to 50.5 with uniform output may be more desirable. Adjust the position of the neutralizing wire, C3, for minimum rf in L16, with drive on, but no screen or plate voltage on the 6146.

Now apply full plate voltage. With no drive, set the bias adjustment for a 6146 plate current of 25 to 30 mA. With the dummy load connected, experiment with the amount of drive needed to reach maximum plate current. Preferably, use a scope to check for flat-topping as the drive is increased. An output of 40 watts, cw, should be obtainable. The quality of the ssb signal is determined first by the equipment generating it, but it can be ruined by improper operation. Over driving the mixer or the 6146, and improper loading of the amplifier will cause distortion and splatter. Continuous monitoring with a scope is the best preventive measure.

Because of the frequencies mixed, and the bandpass coupling between stages, the output of the transverter is reasonably clean. Still, use of an antenna coupler or filter between the transverter and antenna is good insurance. The same treatment of the transverter output is desirable when driving a linear amplifier.

Fig. 7-7 — Panel view of the 2-meter transverter. This version is patterned after a transmitting converter design by K9UIF. The on-off switches for ac and dc sections of the power supply are mounted on the front panel of the unit as are the pilot lamps and plate meter for the PA stage. The tuning controls for the various stages are accessible from the top of the chassis.

A 2-METER TRANSVERTER

This transverter is designed to be used with any 14- or 28-MHz ssb exciter capable of delivering approximately 20 watts peak output. It is stable both in terms of frequency and general operating conditions. It can provide up to 20 watts PEP output at 144 MHz – sufficient, say, for driving a pair of 4CX250 tubes in Class C for cw operation, or the same pair of tubes can be operated AB_1 to provide 1200 watts PEP input with this unit as a driver. The output signal is clean and TVI should not be experienced except where receiver faults are involved.

It is not recommended that beginners attempt this project since vhf ssb circuits require special care in their construction and operation, sometimes a requirement that is a bit beyond the inexperienced builder.

How It Operates

Starting with V1A, the oscillator, Fig. 7-8, a 43.333-MHz or overtone crystal is used at Y1 to provide the local-oscillator signal for the exciter. Output from V1A is amplified by V1B to a suitable level for driving the tripler, V2. 130-MHz or 116-MHz energy is fed to the grids of V3, a 6360 mixer, by means of a bandpass tuned circuit, L3,C1, and L4,C2. The selectivity of this circuit is high, thus reducing unwanted spurious energy at the mixer grids.

Output from the exciter is supplied through an attenuator pad at J1 and is injected to the mixer, V3, at its cathode circuit, across a 270-ohm resistor. The attenuator pad can be eliminated if a very low-power exciter is to be used. The values shown in Fig. 7-8 were chosen for operation with a Central Electronics 20A exciter operating at full input, or nearly so. The amount of driving power needed at the cathode of V3 is approximately 4 or 5 watts PEP.

B1 – Small 15-volt battery.
C1 – 20-pF miniature variable (E. F. Johnson 160-110 suitable).
C2, C3, C5 – 10-pF per section miniature butterfly (E. J. Johnson 167-21 suitable).
C4 – 5-pF per section miniature butterfly (E. F. Johnson 160-205 suitable).
C6 – 20-pF miniature variable (same as C1).
I1, I2 – 117-Vac neon panel lamp assembly.
J1-J3, incl. – SO-239-style coax connector.
J3 – Closed-circuit phone jack.
L1 – 15 turns No. 28 enam. wire, close-wound, on 1/4-inch dia slug-tuned form (Millen 69058 form suitable).
L2 – 12 turns No. 28 enam. wire, close-wound, on same type form as L1.
L3 – 5 turns No. 18 wire space-wound to 7/8-inch length, 1/2-inch dia, center-tapped.
L4 – 3 turns No. 18 wire, 1/2-inch dia, 3/8-inch long, center-tapped.
L5 – 5 turns No. 18 wire, 1/2-inch dia, 5/8-inch long, center-tapped.
L6 – 3 turns No. 18 wire, 1/2-inch dia, 5/8-inch long, center-tapped.
L7 – 4 turns No. 18 wire, 1/2-inch dia, 1/2-inch long, center-tapped.
L8 – 1-turn link of insulated hookup wire, 1/2-inch dia, inserted in center of L7.
L9 – 2 turns of insulated hookup wire over L3.
M1 – 0 to 200-mA dc meter.
P1 – 11-pin chassis-mount male plug (Amphenol 86PM11).
R1 – 50,000-ohm linear-taper, 5-watt control.
RFC1-RFC3, incl. – 2.7-μH rf choke (Millen 34300-2.7).
S1, S2 – Spst rocker-type switch (Carling TIGK60).
Y1 – 43.333-MHz third-overtone crystal for 14-MHz input. If a 28-MHz transceiver will be used, a 38.667-MHz crystal is required.

After the 130-MHz and 14-MHz signals are mixed at V3, the *sum* frequency of 144-MHz is coupled to the grids of V4, the PA stage, by means of another bandpass tuned circuit – further reducing spurious output from the exciter. PA stage V4 operates in the AB_1 mode. Its idling plate current is approximately 25 mA. The plate current rises to approximately 100 mA at full input.

If cw operation is desired, the grid-block keying circuit in the mixer stage (J3) can be included. If ssb operation is all that is contemplated, the minus 100-volt bias line can be eliminated along with J3, R1, and the shaping network at J3. In that case the 15,000-ohm grid resistor from the center tap of L4 would be grounded to the chassis.

The receiving section uses a low-noise uhf MOSFET as the rf amplifier and a second dual-gate MOSFET as the mixer. See Fig. 7-10. The gate-1 and drain connections of the rf amplifier are tapped down on the tuned circuits so that unconditional stability is achieved without neutralization. Oscillator energy is sampled with a two-turn link wound over L3. A short length of RG-58A/U carries the injection energy to Q2. The converter is built in a 5 X 2 1/4 X 2 1/4-inch box constructed from four pieces of double-sided circuit board that have been soldered on all abutting edges. The unit is mounted on the transverter front panel.

Fig. 7-8 — Schematic diagram of the transmitting converter portion of the transverter. Fixed-value capacitors are disk ceramic unless noted differently. The polarized capacitor is electrolytic. Fixed-value resistors are 1/2-watt carbon unless otherwise noted.

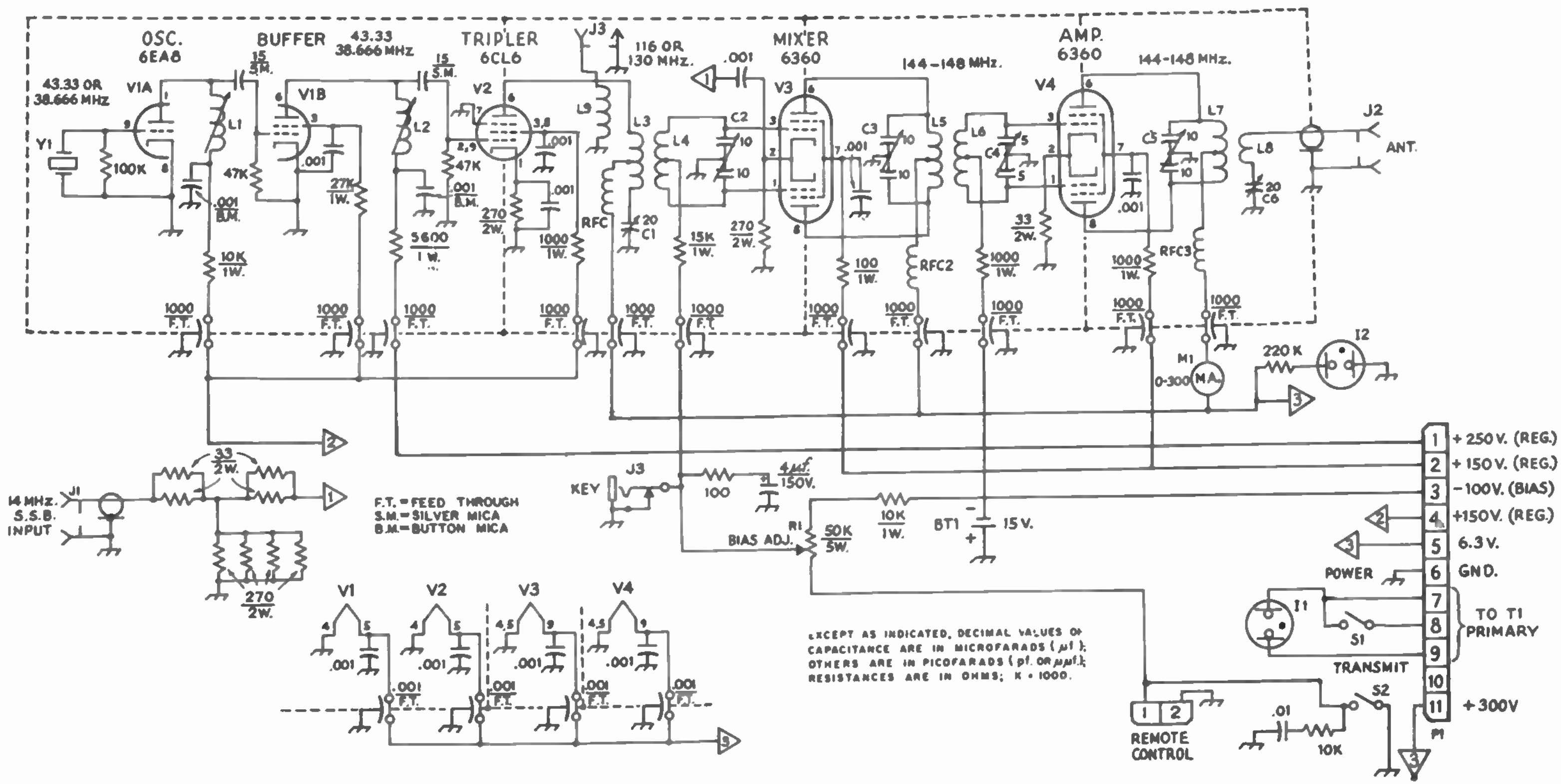


Fig. 7-9 — Inside view of the converter. Shields are used between the rf amplifier input and output circuits, and between the latter and the mixer input circuit. The cable entering the bottom side of the enclosure carries the oscillator injection energy. Output to the associated receiver or transceiver is taken through the jack to the left.

Construction Notes

The photographs show the construction techniques that should be followed for duplicating this equipment. The more seasoned builder should have no difficulty changing the prescribed layout to fit his particular needs, but the shielding and bypassing methods used here should be adhered to even if changes are made.

An 8 X 12 X 3-inch aluminum chassis is used for this equipment. An internal chassis, 5 inches wide, 3 inches deep, and 12 inches long, is made from flashing copper and installed along one edge of the main chassis. This method makes it possible to solder directly to the chassis for making positive ground connections rather than rely on mechanical joints. Shield partitions are made of copper and are soldered in place as indicated on the schematic diagram and in the photo. An aluminum bottom plate is used to enclose the underside of the chassis for confining the rf.

Feedthrough capacitors are used to bring power leads into the copper compartment. Though this adds somewhat to the overall cost of the project, it provides excellent bypassing and decoupling, thus reducing unwanted interstage coupling. It also contributes to TVI reduction. Most surplus houses stock feedthrough capacitors, and offer them at reasonable cost.

Tune-Up

An antenna-changeover relay and a set of normally-open relay contacts, both operated by the exciter, must be provided. The remote control leads, from P2, should be connected to the relay contacts. With power applied to the converter, L12 should be set for maximum noise input to the transceiver. Then, using a signal generator or off-the-air weak signal, peak L9, L10 and L11 for best signal-to-noise ratio.

The transmitter section can be powered by the circuit of Fig. 7-12, or the builder can design a supply of his own choice. Regulated voltages are

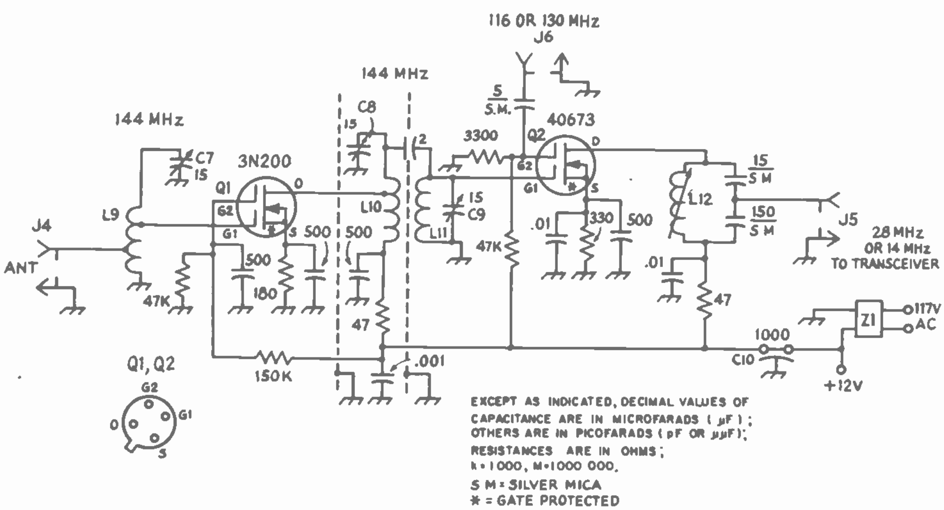
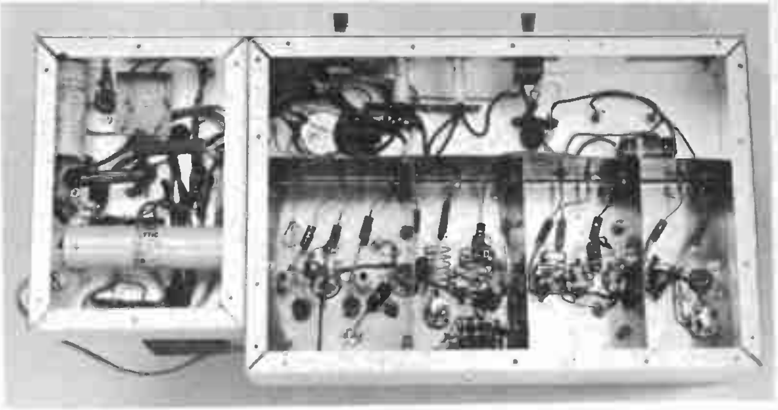


Fig. 7-10 — Diagram of the converter section. Resistors are 1/4-watt composition and capacitors are disk ceramic, except as noted otherwise.

C7-C9, incl. — Air variable, pc mount (Johnson 189-505-5).
C10 — Feedthrough type.
L9 — 4 1/2 turns, No. 18 tinned wire, 1/4-inch ID. Tap at 1 1/2 turns up from the ground end for the antenna connection, and at 3 turns for the Q1 gate.
L10 — 4 1/2 turns, No. 18 tinned wire, 1/4-inch ID. Tap at 3 turns up from the cold end for the Q1 drain connection.
L11 — 5 turns No. 18 tinned wire, 1/4-inch ID.
L12 — 1.99-2.42-µH slug-tuned coil, pc mount, for 28-MHz output (J. W. Miller 46A226CPC); or, for 14-MHz output, 7.3-8.9-µH (J. W. Miller 46A826CPC).
J4-J6, incl. — Phono type.
Q1, Q2 — RCA dual-gate MOSFET.
Z1 — 12-V miniature power supply, transistor radio type.

Fig. 7-11 – Looking into the bottom of the chassis, the rf section is enclosed in a shield compartment made from flashing copper. Additional divider sections isolate the input and output tuned circuits of the last three stages of the exciter. Feedthrough capacitors are mounted on one wall of the copper compartment to provide decoupling of the power leads.



recommended for best operation.

With a dummy load connected to J2, apply operating voltage. Couple a wavemeter to L1 and tune the oscillator plate for maximum output. Then, detune the slug of L1 slightly (toward minimum inductance) to assure reliable oscillator starting. Couple the wavemeter to L2 and tune for peak output. With the wavemeter applied to L4, adjust C1 and C2 for maximum indicated output.

The next step is to connect the transceiver to J1 and supply just enough drive to cause a rise in PA plate current of a few milliamperes. Tune C3 and C4 for maximum indicated plate current at M1, then adjust C5 and C6 for maximum power output to the dummy load. C1, C2, C3 and C4 should be readjusted at this point for maximum plate current of the PA stage. Use only enough drive to bring the PA plate current up to 100 mA at maximum dc input power.

A closed-circuit keying jack is used at J3 so that the mixer stage is not biased to cutoff during voice operation. Inserting the key permits full bias to be applied, thus cutting off V3. R1 should be adjusted for complete cutoff of V3 when the key is open.

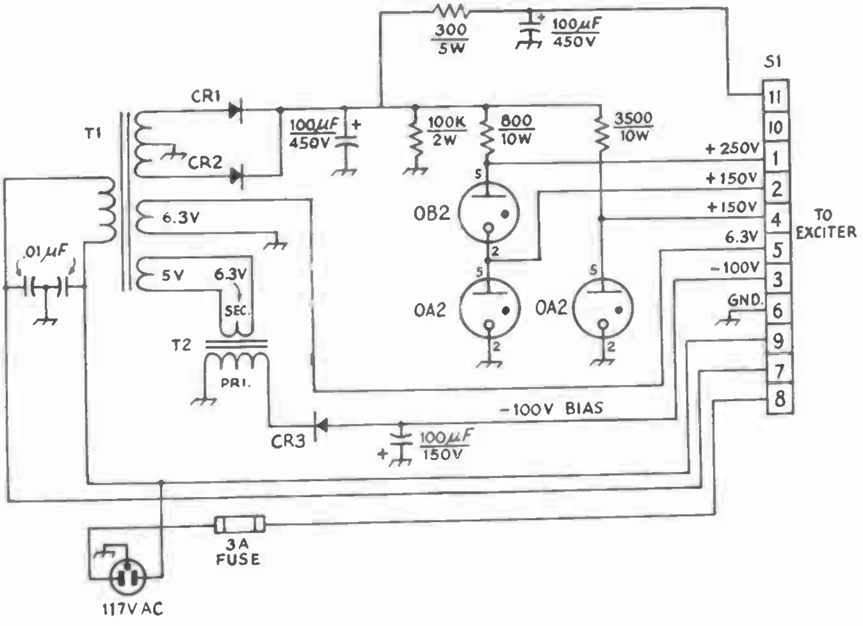


Fig. 7-12 – Schematic of the power supply section. On-off switches for the ac and dc circuits are mounted in the rf deck along with the pilot lamps. Polarized capacitors are electrolytic, others are disk ceramic. CR1 and CR2 are 1000-volt, 1-ampere silicon diodes. CR3 is a 200-PRV 600-mA silicon diode. T1 is a power transformer with a 540-volt ct secondary at 120 mA. Filament windings are 5 volts at 3 A, and 6.3 volts at 3.5 A. T2 is a 6.3-volt, 1-ampere filament transformer connected back to back with the 5-volt winding of T1. S1 is an 11-pin socket (female). A 10,000-ohm resistor and a .01-µF disk capacitor are connected in series between the center tap of T1's secondary and ground for transient suppression when S2 is switched to on. The suppressor is mounted at S2, in the rf deck.

A LINEAR TRANSMITTING CONVERTER

Linear transmitting converters offer several advantages over frequency multiplication schemes for the 70-cm band. Any mode which the low-frequency exciter is capable of generating can be translated to the higher frequency. Frequency stability is similar to that of the hf unit, and all the hf operating features are retained when using the transmitting converter. Single sideband and cw operation are prevalent on Mode B operation of the OSCAR series, and a linear transmitting converter permits this option.

A traditional approach (using tubes) is employed here. This unit is a modified version of one appearing in *QST* for November, 1973, by W2AIH. Detailed construction notes are described in that article. The output is 35 watts from a modified

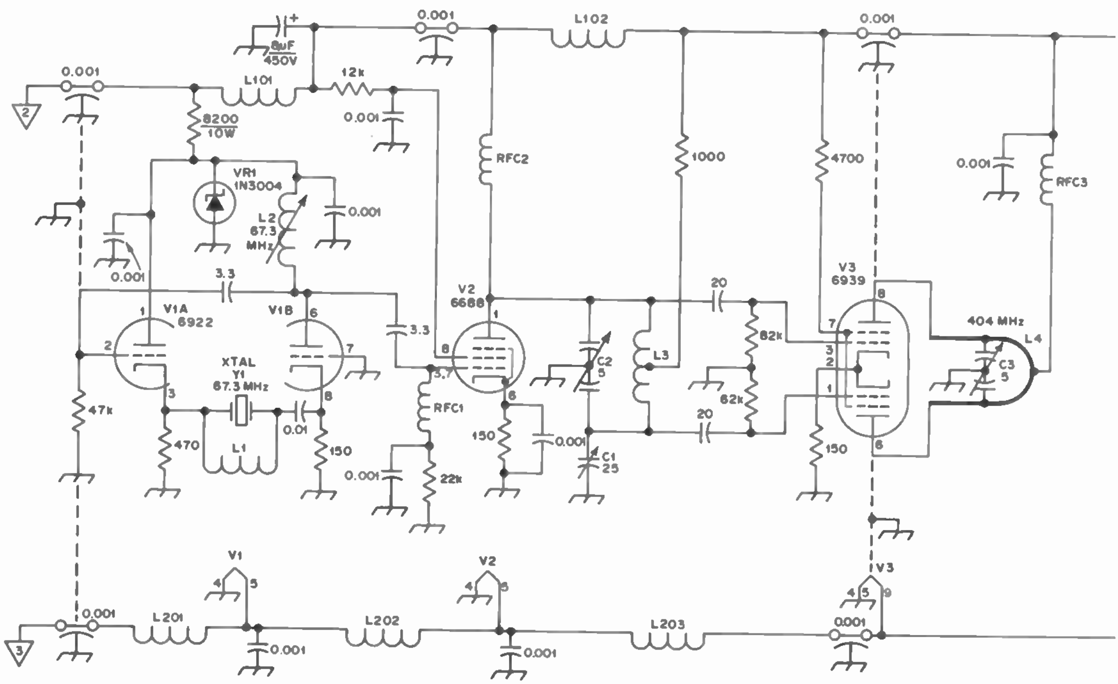
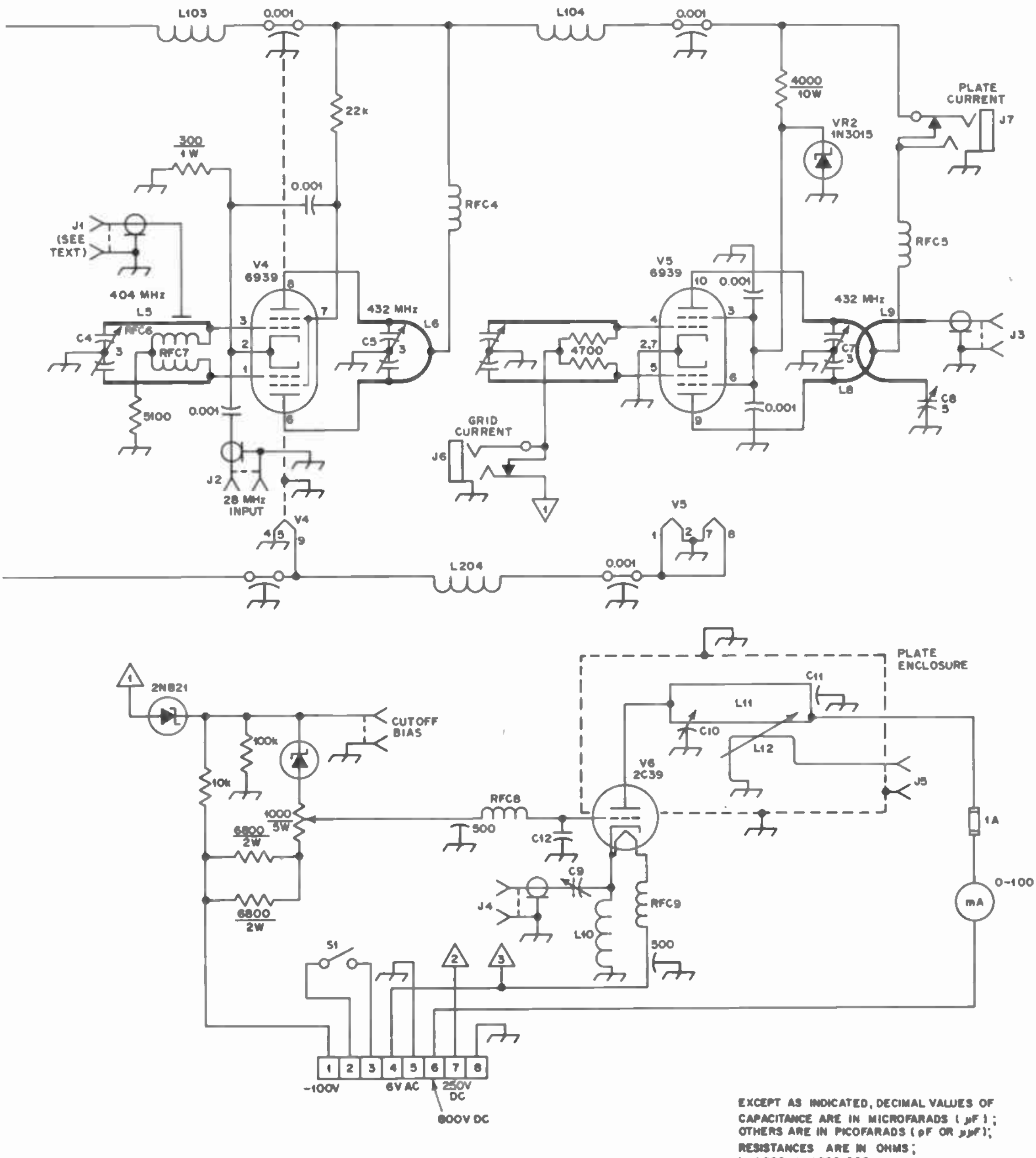


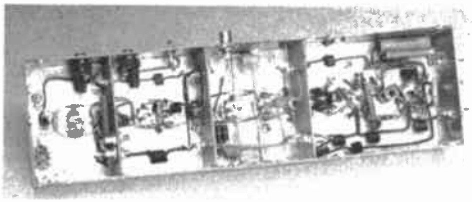
Fig. 1 — Schematic diagram of the heterodyne exciter for 432 MHz.

C1 — 5-25-pF ceramic trimmer.
C2, C3 — 1.8-5-pF air variable. (E. F. Johnson 160-205.)
C4, C5, C6, C7 — 1.5-3-pF air variable. (E. F. Johnson 160-203.)
C8 — 1.5-5-pF air variable. (E. F. Johnson 160-102.)
C9 — Trimmer capacitor, 15 pF.
C10 — See Fig. 2 and text.
C11, C12 — These components are built into T44 cavity.
J1-J5, inclusive — Coaxial connector, type BNC.
J6, J7 — Closed-circuit jack. Insulate from chassis.
L1 — 21 turns No. 26 enam. close-wound on 3/16-inch dia. plastic rod.
L2 — 7 turns No. 20 enam. close-wound on 3/8-inch dia. slug-tuned form.
L3 — 5 turns No. 16 enam. 1/2-inch dia., center-tapped, turns spaced one wire diameter.
L4, L6 — Hairpin loop, 1-7/8-inch long, 7/8-inch wide, No. 14 tinned.
L5, L7 — Each two pieces No. 14 tinned, 3 inches long. Formed as shown in Fig. 3 and the photographs.
L8 — Hairpin loop 1-7/8-inch long, 1 inch wide, No. 12 enam., with plate connectors. See Fig. 4.
L9 — Hairpin loop 1-inch long, 1-inch wide, No. 12 enam., spaced 1/8-inch above L8.
L10, L11, L12 — Part of T44 cavity.
L101-L104, inclusive — 4 turns No. 20 insulated hookup wire, 1/4-inch diameter, close-wound.
L201-L205, inclusive — 4 turns No. 18 insulated hookup wire, 1/4-inch diameter, close-wound.
RFC1 — 22 turns No. 26 enam. on 3/16 inch diameter plastic rod. Spaced one wire diameter.
RFC2-RFC5, inclusive — 5 turns No. 22 insulated hookup wire, 3/16-inch diameter.
RFC6, RFC7 — 5 turns No. 26 enam., 1/8-inch diameter, close-wound.
RFC8, RFC9 — Part of T44 cavity.
VR1 — 90-volt, 10-watt Zener diode, 1N3004 or equiv.
VR2 — 200-volt, 10-watt Zener diode, 1N3015 or equiv. For direct chassis mounting of the diode, use the reverse-polarity version of VR1 and VR2, i.e., 1N3004R and 1N3015R, respectively.
Y1 — Overtone crystal, 67.333 MHz.

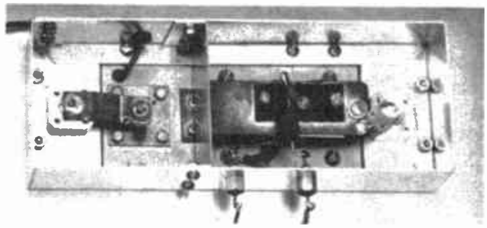
surplus Motorola T-44 transmitting module. OSCAR Mode B requires less than 10 watts erp for effective access of the satellite. Feed-line loss in a typical installation will drop the 35-watt output to the level where an omnidirectional, unity-gain antenna will be sufficient for use with OSCAR. For other applications, the most popular kilowatt amplifier for 432 MHz, a design by K2RIW (*QST* for April and May, 1972), requires in excess of 20 watts of drive for maximum output. The output of this transmitting converter handily provides for both situations.

Injection to the mixer is provided by multiplication of 67.33-MHz energy from the oscillator. The 6922 dual triode is used as a Butler oscillator. The plate circuit of V1B is capacitively coupled to the input of the 6688 pentode doubler, V2. The following stage uses a 6939 as a tripler for an output frequency of 404 MHz. Inductive coupling is used between the tripler and the 6939 mixer,





Bottom view showing stages for V1 through V5.



Bottom view of T44 cavity.

V4, and between the output of the mixer and the 6939 amplifier stages (L4, L5/L6 and L7, respectively). A 6939 driver-amplifier stage runs Class AB1 for good suppression of unwanted mixing products. The final amplifier is a cathode-driven 2C39 triode. Input power to the cathode is coupled through a variable capacitor to provide an impedance match to the exciter. The plate and cathode hardware are modified components from a T44 amplifier. A flapper type of capacitor is used to tune the plate circuit to resonance. The mica dielectric in the bypass capacitors is replaced with 5-mil Teflon sheet. This is all the modification required for the 2C39 PA.

Tune-up is simply a matter of applying voltage to the unit and adjusting the tuned circuits for maximum indication on a wavemeter. Coupling between L3, and L4/L5 and L6 should be as loose as possible while still providing adequate drive for the 6939s. Final adjustment is made by applying approximately 0.5 watt of drive and adjusting for maximum power output as indicated by a wattmeter in the feed line. For normal use, approximately 0.5 watt of 28-MHz energy is adequate for full output with minimum distortion. More drive will cause distortion in the mixer and will generate spurious signals within and out of the ham band.

T44 Cavity Modification

It was felt that the inductive tuning arrangement was somewhat inefficient so this was replaced with a capacitive type instead. A "flapper" capacitor of spring brass is attached to the newly fabricated top cover. The top cover is bent from aluminum sheet stock (.035 inch). Tuning is accomplished by means of a screw which bears against the spring-brass capacitor plate. Some adjustment of the line that forms L11 may also be required to bring the plate tank near resonance. Final adjustment is accomplished by means of C10.

Optional LO Output

In some applications, rf energy at the LO frequency (404 MHz) is required. This is accomplished by mounting a BNC connector (with the pin sawed off) near L5. Refer to the bottom view of the driver stages. The connector can be seen just above C4 in the third compartment from the left.

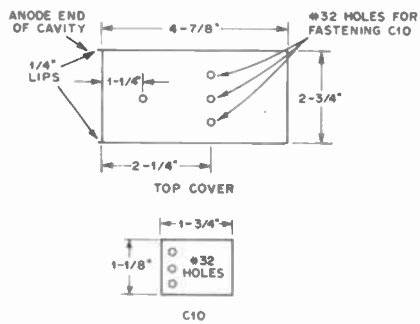
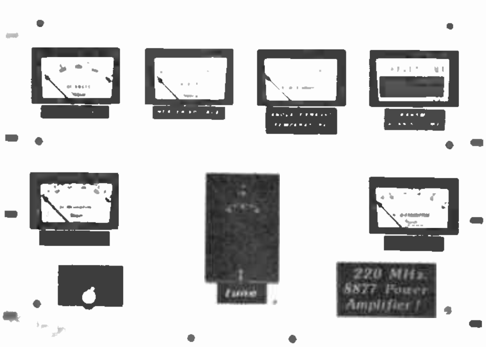


Fig. 2 — Layout details for modified top cover and brass capacitor flap.

A 220-MHz HIGH-POWER AMPLIFIER

Circuits for 220-MHz power amplifiers have long been designed around the external-anode tetrode. While these tubes offer high gain, instability problems have caused many builders considerable consternation over the years. Multiple-tube amplifiers are often necessary to obtain the high power levels many moonbouncers and weak-signal specialists require. Push-pull amplifiers have been tried with moderate success, and recently parallel-tube designs have found favor.[1]

Modern computer-aided tube designs have brought forth high-μ triodes such as the



[1]Knadle, "A Strip-Line Kilowatt Amplifier for 432 MHz," *QST* for April and May, 1972.

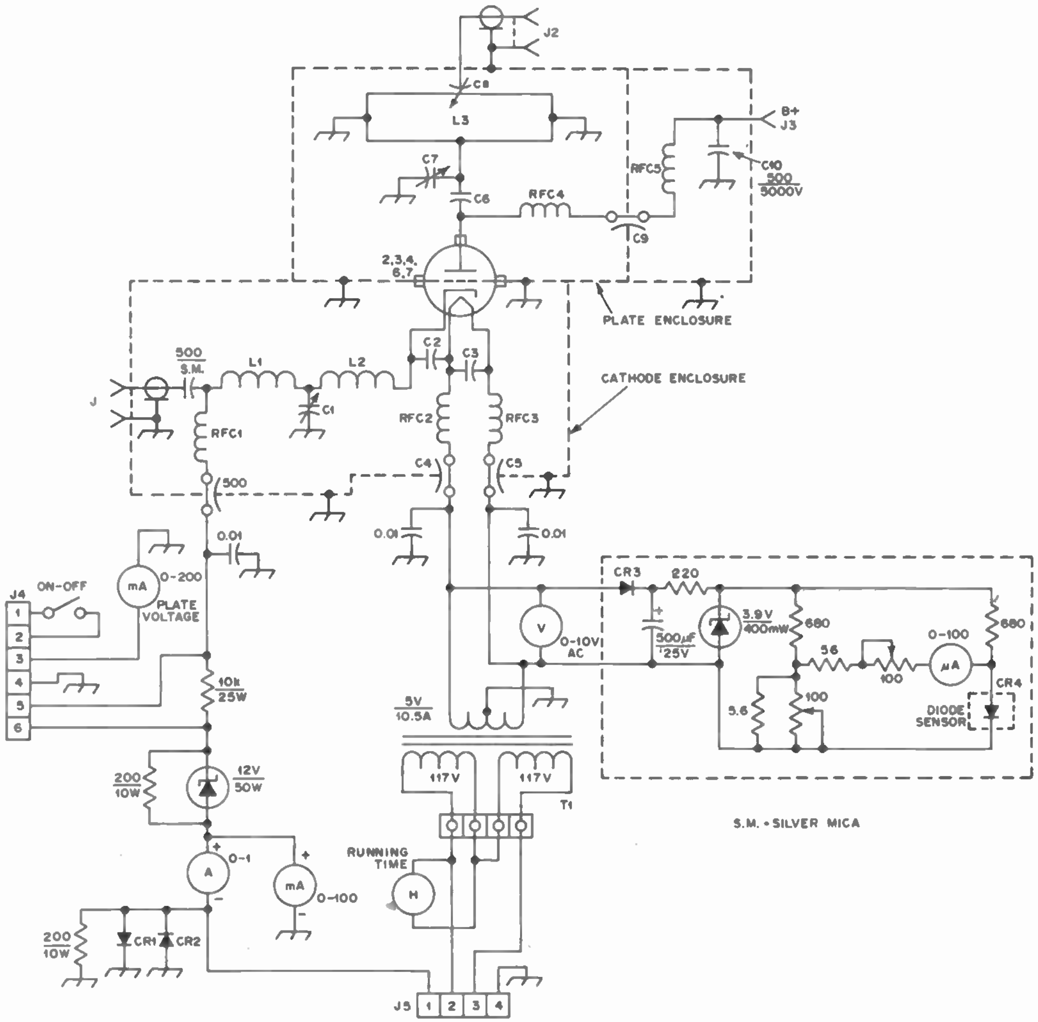


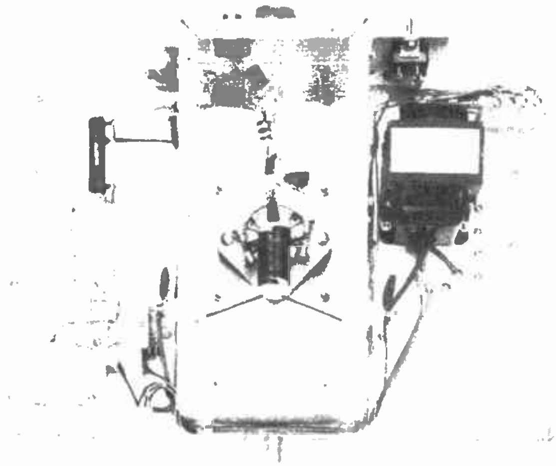
Fig. 1 — Schematic diagram of the amplifier. Unless otherwise specified, all capacitors are disk ceramic and resistors are 1/2-watt carbon composition.

C1 — Air variable, 15 pF.
C2, C3 — Button mica, 500-pF, 500-V rating.
C4-C9, inclusive — Teflon capacitor (use 10-mil Teflon sheet) see Fig. 2 for dimensions.
C10 — Doorknob capacitor, 500 pF, 5-kV rating.
CR1-CR4, inclusive — HEP170 or equiv.
J1, J2 — Coaxial receptacle, type N.
J3 — High-voltage connector (Millen).
L1 — 3 turns No. 14, 1/4 inch ID, 3/4 inch long.
L2 — 1/4 inch wide, 2-3/8 inch long copper flashing strap.
L3 — Plate inductor (see Fig. 2)
RFC1 — 8 turns No. 18 enam. 1/2 inch dia. 3/4 inch long.
RFC2, RFC3 — 10 turns No. 18 enam. bifilar wound on 3/4 inch Teflon rod close wound.
RFC4 — 5 turns No. 16 enam. 1/2 inch ID, 3/4 inch long.
RFC5 — 12 turns No. 18 enam. wound on 1-MΩ 2-watt composition resistor.
T1 — Filament transformer, 5.0 V at 10.5 A.

3CX1500A7/8877, a 1500-watt dissipation external-anode triode with maximum ratings good through 250 MHz. The ceramic insulation allows a heavy flow of rf current through the tube, with no loss of stability in a properly designed circuit. Low heater requirements (5 V at 10.5 A) add to the appeal of the 8877. This amplifier employs the 3CX1500A7/8877 in a cathode-driven circuit. The grid is grounded directly to the chassis, adding to the stability. The amplifier is unconditionally stable — more so than some amplifiers built for the hf region.

Circuit Details

The input circuit consists of a T network. Medium values of *Q* were chosen to provide high efficiency. Both the cathode and the heater are operated at the same rf potential; the heater is held above rf ground by the impedance of the filament choke. The plate tank is a pair of quarter-

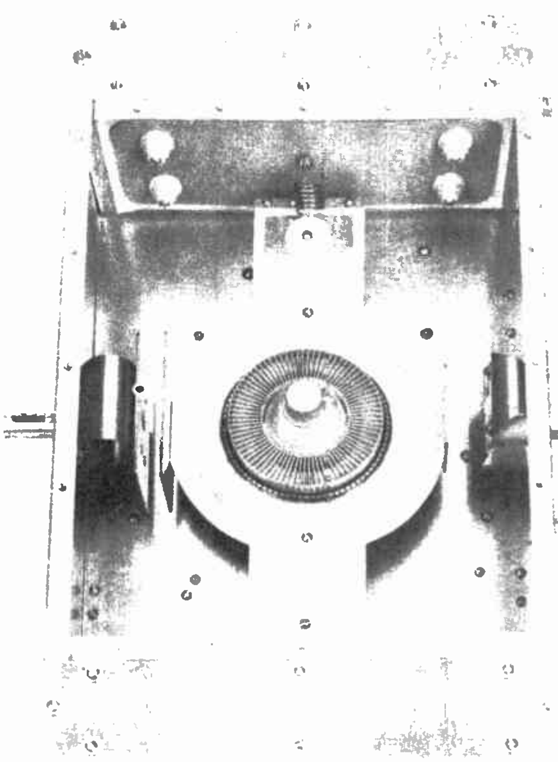


Bottom view of the amplifier. RFC2 and RFC3 can be seen above tube socket (bifilar winding). Copper strap is L2 shown connected to C1. Small coil is L1 and larger coil is RFC1. The grid of the tube should be grounded to the chassis with finger stock similar to that used in the plate line. Component mounted on the heat sink at left is the Zener diode used for biasing purposes.

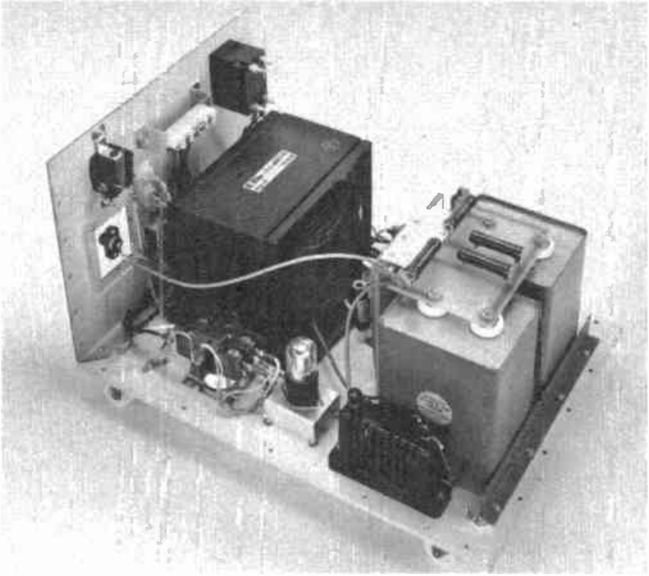
wavelength striplines placed symmetrically about the tube.[2] This arrangement permits a more uniform flow of current through the anode, preventing "hot spots" on the anode conducting surface. Additionally, tube output capacitance is effectively halved, as one half the tube capacitance (13 pF) is used to load each stripline. Striplines act as low-pass circuit elements even with the high unloaded-Q conditions found at 220 MHz. Linear inductors also offer control of odd-mode harmonics. No spurious responses could be found in this amplifier up through the 900-MHz region.

A stripline impedance can be varied by changing its width and relation to its ground planes.

[2] Barber, Rinaudo, Orr and Sutherland, "Modern Circuit Design for VHF Transmitters," *CQ* for November and December, 1965.



Interior view of the amplifier. Contact of the tube with "hot side" of C6 is accomplished with suitable finger stock (available from the tube manufacturer). This conductor, in conjunction with a similar one separated by the Teflon insulator, forms the L3/C6 combination. The entire assembly is sandwiched together by means of four insulated bushings (approximately 3/4 inch [19 mm] diameter). Placement of bushings is not critical. RFC4 can be seen at the right connected to C9. C8 is seen at the bottom center of the photo and has a nominal spacing of 1 inch (25 mm) to similar plate soldered to L3. Tuning capacitor, C7, can be seen in upper center of photo (see Fig. 2 for dimensions). Drive mechanism can be of builder's choice or see reference 3.



Interior view of the power supply.

Physical dimensions of the tube limit the position of the stripline above one ground plane. In order to utilize commercially available chassis, the stripline was placed 1-1/4 inch (32 mm) above one side of an inverted 4-inch (102 mm) high chassis. This means that approximately 75 percent of the rf current flows through the chassis, but only 25 percent flows through the top shield cover. The small percentage flowing through the top reduces the effect of any mechanical anomalies associated with a removable cover.

For quarter-wavelength lines, the ratio of line impedance to reactance should be between 1.5 and 2.0 for the best bandwidth. Taking stray capacitance into account, expected tuning capacitance and tube output capacitance gives a value of 55 Ω for X_C. Values of line impedance versus line length for resonance at 222 MHz were computed on a programmable calculator for impedances between 30 and 100 ohms. These were plotted on a graph. Final dimensions were determined using this system, choosing dimensions that fell into the middle of the graph, thus allowing for any unpredicted effects.

The plate blocking capacitor consists of a sandwich of brass plate and the stripline, with Teflon sheet as the dielectric. This forms a very low-loss, high-voltage capacitor. The plate bypass capacitor is built along the same principles. A piece of circuit board was once again sandwiched with Teflon sheet to the side wall of the chassis. This technique is used effectively throughout as an inexpensive bypass or feedthrough capacitor at vhf.

Amplifier output is coupled through a capacitive probe. Transformation of the load impedance to the tube resonant-load impedance is achieved by means of a series reactance (the loading capacitor). The tuning capacitor is solidly grounded by means of a flexible strap of negligible inductance. Me-

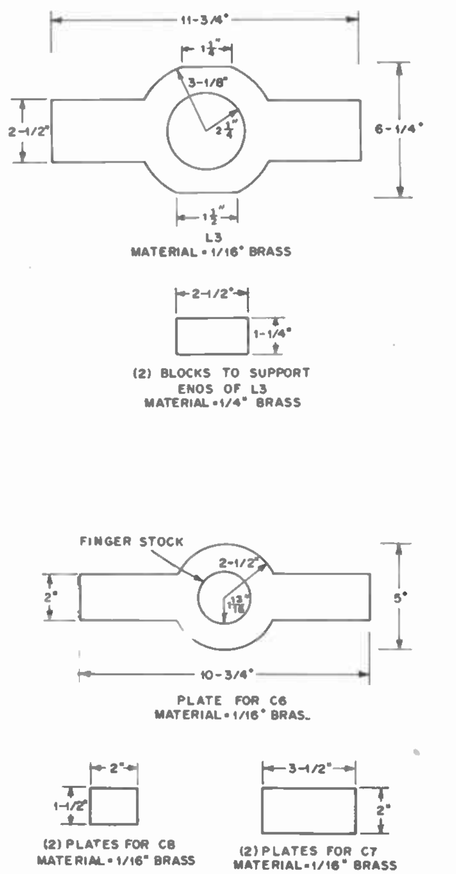


Fig. 2 — Construction details of plate line and associated components.

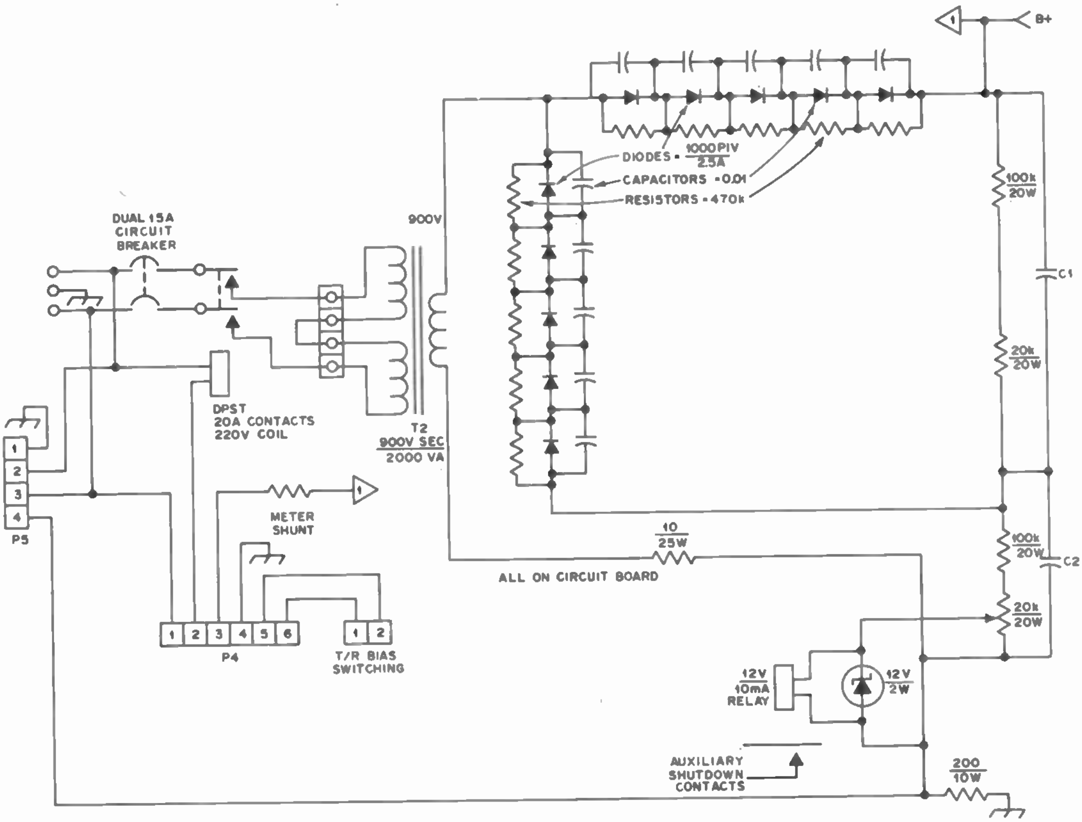


Fig. 3 — Schematic diagram of a suitable power supply for the amplifier. Unless otherwise specified, all capacitors are disk ceramic (1000 V) and resistors are 1/2-watt carbon composition.

C1, C2 — Filter capacitor, 74 μF at 2500 V (Sprague Electric).
Meter shunt used should be one supplied with movement for 4-kV full-scale.

chanical details were described by Sutherland.[3]

A rather elaborate metering system is employed. Although all of the meters provide useful data, only the plate and grid meters are necessary for proper amplifier use. At a repeater site where key-down service is the rule rather than the exception, measurement of heater usage and voltage provide data requisite to tube replacement. The anode exhaust-temperature metering circuit takes advantage of a thermal property of semiconductors. As the temperature changes the forward resistance of a diode changes in a nearly linear manner. The diode sensor is made a part of a bridge circuit, allowing calibrated operation. Calibration parts may be determined by packing the diode in ice for the low point (0°C) and immersing it in boiling water for the high point (100°C). The amount of heat dissipated by the tube is inversely proportional to the efficiency for a given power input. Low heat dissipation yields longer tube life.

High-power amplifiers require considerable attention to cooling. The plate compartment is pressurized by air from an external blower, and holes in the chassis allow a portion of this air to pass through the grid and cathode structure. Most of the air flows through the anode, a hand-made Teflon chimney, then out the top cover. Aluminum screening is tightly bonded around these two openings. No radiated rf could be detected around the chassis except within one inch of the anode exhaust hole.

To commence operation, the input should be adjusted for minimum VSWR with no voltages applied. The covers should be in place whenever voltage is present. Drive should *never* be applied without plate voltage and a load connected if the filament is energized. Cooling air must always be supplied whenever the filament is turned on.

After a 60-second warmup small amounts of drive may be applied. The plate circuit is then tuned for maximum output indication. The drive level is then increased. Tuning and loading follow the normal procedure for any cathode-driven amplifier: Adjustments are made for maximum output and efficiency. When the desired plate output power has been achieved, the input circuit should be adjusted for minimum input VSWR.

[3]Sutherland, "High-Performance 144 MHz Power Amplifier," *Ham Radio* for August, 1971.

GROUNDED-GRID 50-MHZ AMPLIFIER

Increasing use of 50-MHz transceivers and transmitters having outputs of 25 watts or more has created a demand for amplifiers to be used with such equipment as the driver. The grounded-grid amplifier of Fig. 7-27 is designed for this use. With 30 watts or more of driving power it will deliver 600 watts cw output. As a Class-B linear, single-tone conditions, its rated PEP output is 750 watts.

Circuit

The Eimac 3-500Z triode is designed for grounded-grid service. As may be seen from Fig. 7-30, driving power is applied to the filament circuit, which must be kept above rf ground by means of high-current bifilar rf chokes, RFC1 and RFC2. These are a central feature of the bottom view, Fig. 7-29. The input impedance is low, so the input circuit, L1,C1, tunes broadly, and the 50-ohm line from the exciter is tapped well up on L1. The plate circuit is merely a coil of copper tubing, L2, inductively tuned by means of a "shorted turn" of copper strip, rotated inside its cold end. See Fig. 7-28. Tuning is smooth and the rotating loop avoids many problems commonly encountered in tuning high-powered amplifiers by conventional methods. Plate voltage is shunt fed to the tube, to prevent the high dc voltage from accidentally appearing on the output coupling loop or on the antenna line.

Most of the lower part of the schematic diagram has to do with control and metering, and is largely self-explanatory. The exciter voice-control relay shorts out R1, allowing grid current to flow, and making the amplifier operative, if the filament and primary-control switches, S1 and S2, have been closed. Feeding ac voltage to the plate-supply relay through J4, J5 and P1 makes application of plate voltage without the filament and blower being on impossible.

Construction

The amplifier chassis is aluminum, 10 X 12 X 3 inches in size, with the tube socket centered 3 1/8 inches from the front edge. The sheet-aluminum panel is 10 inches high. The decorative edging is "cove molding," used by cabinet makers for counter tops. Sides and back are also sheet aluminum. Where they need not be removable, parts are fastened together by pop-riveting. Tools and rivets for this work can be found in most hardware stores. Perforated aluminum (cane metal) is used for the top, and for covering the panel viewing hole.

Stretch the wire for the bifilar rf chokes, before winding. Then, with the wires side by side, under tension, wind them on a form of wood or metal. This is left in until the choke ends are soldered in position. Then remove the form and coat the windings with coil cement, to help maintain turn alignment.

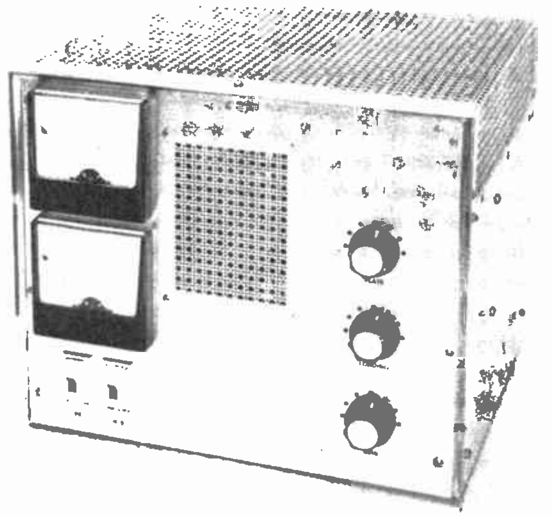


Fig. 7-27 — Table-top 50-MHz amplifier of grounded-grid design, only 10 X 12 inches in size. Grid and plate current are monitored simultaneously. Knobs at the right are for input tuning, bottom, amplifier loading, center, and plate tuning, top.

Connections to the grid terminals (on opposite sides of the socket) are made with short 1/4-inch copper straps soldered to the pins and bolted to the chassis with No. 6 screws, nuts and lock-washers. Be sure that a clean, tight rf ground results.

In Fig. 7-28 it will be seen that the hot end of L2 is supported on the top of the two blocking capacitors, C3 and C4, which in turn, are mounted on the Teflon rod that serves as the form for RFC3. The ground end of L2 is supported on a vertical post made of 3/8-inch copper tubing, 1 3/8 inches high. The end of the coil can be fitted with a heavy copper lug, or pounded flat. A hole is drilled in the flat portion and a 2-inch brass bolt runs through it and the post and chassis. Be sure that there is a permanent solid rf ground at this point.

The shunt-feed rf choke is effectively across the tuned circuit, so it must be a good one. Hand-winding as described below is strongly recommended, as no ready-made choke is likely to be as good. Teflon is slippery, so a light thread cut in the form will help keep the winding in place. If this cannot be done, prepare and wind two wires, as for the filament chokes. Feed the wire ends through one hole in the form, and wind a bifilar coil. Pull the other ends through the finish hole, bending one

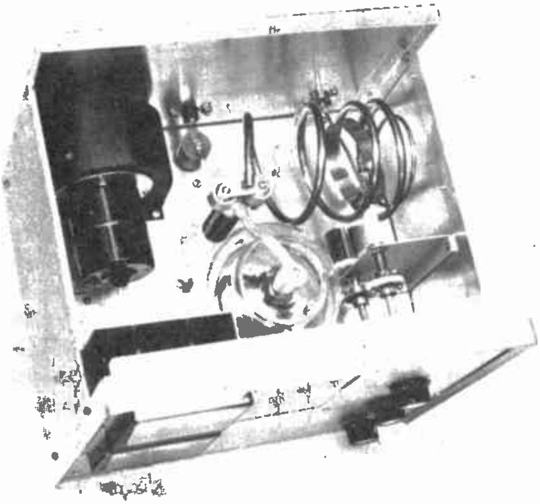


Fig. 7-28 – Interior view of the 50-MHz amplifier shows the shorted-turn tuning system, plate coil and output coupling, upper right. The tuning and loading controls are mounted on a bracket to the right of the 3-500Z tube and chimney. Meter shielding is partially visible in the left front corner.

back tightly at the hole edge. Remove the other winding, which should leave a tight evenly-spaced coil that makes an excellent vhf choke.

The blocking capacitors, C3 and C4, are mounted between brass plates, one of which is fastened to the top of the rf choke form with a sheet-metal screw. The other plate is connected to the hot end of L2 by means of a wrap-around clip of flashing copper. The lead to the tube plate cap is made with braid removed from a scrap of coax. A strip of flashing copper about 1/4 inch wide is also good for this. Use a good heat-dissipating connector such as the Eimac HR6.

The shorted-turn tuning ring is centered between the first two turns of L2. The ring is attached to a ceramic pillar, and that to a 1/4-inch shaft, the end of which is tapped for 8-32 thread. This shaft runs through a bearing mounted in a bracket 4 inches high and 2 3/4 inches wide, fastened to the chassis and the side of the enclosure. The output loading capacitor, C6, is also mounted on this bracket. It is one inch above the chassis, and the tuning-ring shaft is 3 1/4 inches above the chassis. The input tuning capacitor, C1, is mounted under the chassis, with equal spacing between the three, for symmetrical appearance.

The output coupling loop, L3, is just inside the cold end of L2. It can be adjusted for optimum coupling by "leaning" it slightly into or out of L2. Be sure that it clears the shorted turn throughout movement of the latter.

The coaxial output jack, J3, is on the rear wall of the enclosure. A small bracket of aluminum grounds it to the chassis, independent of the bonding between the chassis and the enclosure. Plate voltage enters through a Millen 37001 high-voltage connector, J2, on the rear wall, and is bypassed immediately inside the compartment with a TV "doorknob" high-voltage capacitor, C5.

The blower assembly in the left rear corner of the chassis draws air in through a hole in the back of the compartment, and forces it down into the enclosed chassis. The only air path is then back up through the socket and chimney (Eimac parts SK-410 and SK-406 recommended) and out through the top of the enclosure. The data sheet for the 3-500Z specifies an air flow of at least 13 cubic feet per minute, when the tube is operated at 500 watts plate dissipation. The ac leads for the blower motor come into the enclosure on feedthrough capacitors.

The meters are enclosed in a shield fastened to the front and side panels. Meter terminals are bypassed for rf inside the shield, and leads come through the chassis on feedthrough capacitors. The rocker-type switches just below the meters have built-in illumination. The high-voltage switch is not meant to control the plate supply directly, but rather through a relay, as in the 3000-volt supply shown in Chapter 5. The plate meter is in the negative lead, so be sure that your supply is compatible with this arrangement. Do not use this system where a potential difference exists between the amplifier and power supply chassis. All power leads are made with shielded wire (Belden 8862) and all exposed points are bypassed to ground.

Adjustment and Use

Do not apply drive to the 3-500Z without the plate voltage being on. Also, it is recommended that initial testing be done with low drive, and with a plate voltage of 1500 or less. With a 50-ohm load

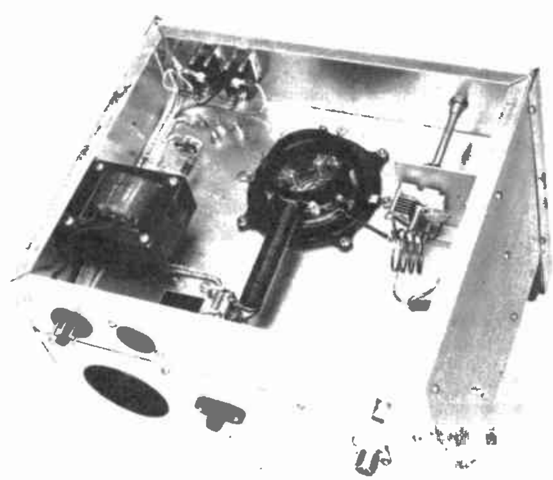


Fig. 7-29 – With the bottom cover removed, a look into the chassis from the rear shows the input circuit, L1,C1, right, the bifilar filament chokes, foreground, filament transformer and control switches. Opening in the rear wall is for air intake.

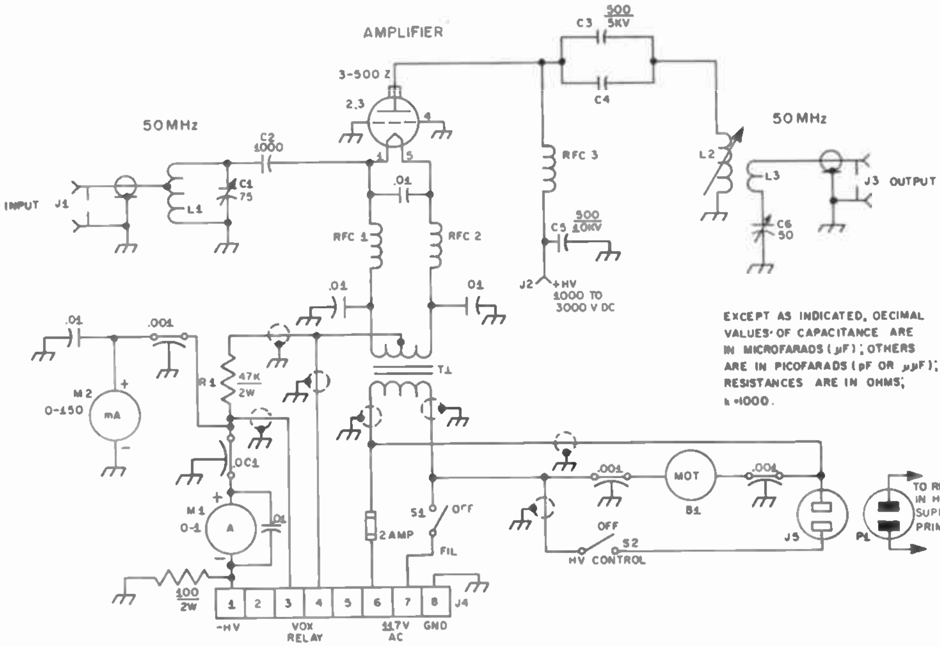


Fig. 7-30 — Schematic diagram and parts information for the 50-MHz grounded-grid amplifier.

B1 — Blower, 15 ft^3/min or more.
C1 — 75-pF variable (Johnson 167-4).
C2 — 1000-pF dipped mica.
C3, C4 — 500-pF 5-kV transmitting ceramic (Centralab 858S-500).
C5 — 500-pF, 10-kV or more, TV "Doorknob."
C6 — 50-pF variable (Johnson 167-3).
J1 — BNC coaxial receptacle.
J2 — High-voltage connector (Millen 37001).
J3 — Type N coaxial receptacle.
J4 — 8-pin male power connector, chassis-mounting.
J5 — AC receptacle, chassis-mounting.
L1 — 4 turns No. 12 enam, 1 inch long, 1inch dia. Tap 2 1/2 turns from ground end.
L2 — 3 1/2 turns 1/4-inch copper tubing, 3 1/2-inch dia, 5 1/4 inches long. Diameter is finished dimension, not that of form used for winding. See text and photo for turn spacing. Tuning ring is closed loop of 1/2-inch copper strip, 2 5/8-inch dia.
L3 — 1 turn, 3-inch dia, and leads, made from one piece of 1/8-inch copper tubing or No. 8 wire.
M1 — DC meter, 0-1 ampere (Simpson Wide-Vue, Model 1327).
M2 — 0-300 mA, like M1.
P1 — AC plug, on cable to power supply.
R1 — 47,000-ohm 2-watt resistor.
RFC1, RFC2 — 21 turns each, No. 12 enam, 1/2-inch dia, bifilar.
RFC3 — 30 turns No. 20 enam, spaced wire dia, on 3/4-inch Teflon rod, 3 3/4 inches long. Drill end holes 1/2 and 2 3/4 inches from top.
S1, S2 — Spst, rocker-type, neon-lighted (Carling LT1L, with snap-in bracket).
T1 — Filament transformer, 5 V, 15 A (Stancor P6433; check any electrical equivalent for fit under 3-inch chassis).

connected to J3, apply 1000 to 1500 volts through J2, and turn on the driver. Adjust the tuning ring inside L2 for a dip in plate current. Tune C1 for maximum grid current. Tune C6 and adjust the position of L3 with respect to L2 for maximum output. If the amplifier seems to be running properly, connect an SWR bridge between the driver and J1, and check reflected power. It should be close to zero. If otherwise, adjust the tap position on L1.

Tuning range of the plate circuit can be checked with a grid-dip meter, with the power off the amplifier. The range is affected by turn spacing overall, and at the cold end. The closer the first two turns are together the greater the effect of the tuning ring. No other tuning device is used, so some experimentation with diameter and length of L2 may be needed if you want other than the 49.8 to 52.7 MHz obtained with the graduated turn spacing visible in the interior view. The highest frequency is reached with the ring in a vertical plane. Dimensions that affect tuning range are as follows: Grounded support for L2 – 1 1/8 inches from right side of chassis, and 3 1/4 inches from rear. RFC3 mounting position – 4 inches from rear and 5 1/2 inches from left. Shorted turn approximately centered between turns 1 and 2 of L2. The start of L3 bends from the stator of C6 to near the start of L2. The end toward J2 passes between the first two turns of L2, clearing the tuning ring in any position of the latter.

Once the amplifier seems to work normally at moderate plate voltages, apply higher, up to the maximum of 3000. Plate current, with no drive, should be about 160 mA. It can be lowered by inserting 0.1 to 0.4 ohm in series with R1 and the filament center-tap. A Zener diode, 2 to 9 volts, 10 watts, could do this job, as well.

Keep the amplifier tuned for maximum output. Do not decouple to reduce output; cut down drive and/or plate voltage instead. Adjustment for linear operation requires a scope. Maximum output, minimum plate current and maximum grid current should all occur at the same setting of the plate tuning. If they do not, the output loading is over coupled, or there is regeneration in the amplifier. The plate-current dip at resonance is noticeable and smooth, but not of great magnitude.

Typical operating conditions given by the manufacturer, and in the tube-data section of the *Handbook*, are guides to good practice. The amplifier works well with as little as 1000 volts on the tube plate, so varying the ac voltage to the plate-supply transformer is a convenient way to control power level. It is seldom necessary to run the maximum legal power in vhf communication, so some provision for this voltage control is recommended. With just one high-voltage supply needed and no critical tuning adjustments, power variations from 100 to 600 watts output are quickly and easily made. This amplifier was built by Tom McMullen, W1SL, and first described in *QST* for November, 1970.

A 2-KW PEP AMPLIFIER FOR 144 MHz

Large external-anode triodes, in a cathode-driven configuration, offer outstanding reliability, stability and ease in obtaining high power at 144 MHz. The selection is somewhat limited and they are not inexpensive. Data on the recently introduced 3CX1500A7/8877, a high-mu, external-anode power triode, appeared very promising. A reasonable heater requirement (5V at 10 A) and an inexpensive socket and chimney combination made the tube even more attractive.

The techniques employed in the design and construction of the cathode-driven 3CX1500A7/8877 amplifier described here have removed many of the mechanical impositions of other designs. Those interested in obtaining complete constructional details should refer to the two part article appearing in December, 1973, and January, 1974 *QST*.

The plate tank operates with a loaded Q on the order of 40 at 2-kW PEP and 80 at 1 kW. Typical loaded Q values of 10 to 15 are used in hf amplifiers. In comparison, we are dealing with a relatively high loaded Q, so losses in the strip-line tank-circuit components must be kept very low. To this end, small diameter Teflon rods are used as mechanical drive for the tuning capacitor and for physical support as well as mechanical drive for the output-coupling capacitor. The tuning vane or flapper capacitor is solidly grounded, through a wide flexible strap of negligible inductance, directly to the chassis in close proximity to the grid-return point. A flexible-strap arrangement, similar to that of the tuning capacitor, is used to connect the output coupling capacitor to the center pin of a type N coaxial connector mounted in the chassis base. Ceramic (or Teflon) pillars, used to support the air strip line, are located under the middle set of plate-line dc isolation bushings. This places these pillars well out of the intense rf field associated with the tube, or high-impedance end of the line. In operation, plate tuning and loading is quite smooth and stable, so a high-loaded Q is apparently not bothersome in this respect.

In this amplifier, output coupling is accomplished by the capacitive probe method. As pointed out by Knadle[1] "Major advantages of capacitive probe coupling are loading linearity and elimination of moving contact surfaces."

Capacitive-probe coupling is a form of "reactive transformation matching" whereby the feed-line (load) impedance is transformed to the tube resonant-load impedance (R_O) of 1800 ohms (at the 2-kW level) by means of a series reactance (a capacitor in this case). At the 1-kW level, R_O is approximately twice that at the 2-kW PEP level. Therefore, the series coupling capacitor should be variable and of sufficient range to cover both power levels. Formulas to calculate the transformation values have been presented in *QST*.[2]

The electro-mechanical method of probe coupling used in this amplifier is easy to assemble and provides good electrical performance. Also, it has no moving-contact surfaces and enables placement of the output coupling, or loading, control on the front panel of the amplifier for ease in adjustment.

The grid- and cathode-metering circuits employed are conventional for cathode-driven amplifiers. The multimeter, a basic 0-1 mA movement, is switched to appropriate monitoring points.

An rf-output monitor is a virtual necessity in

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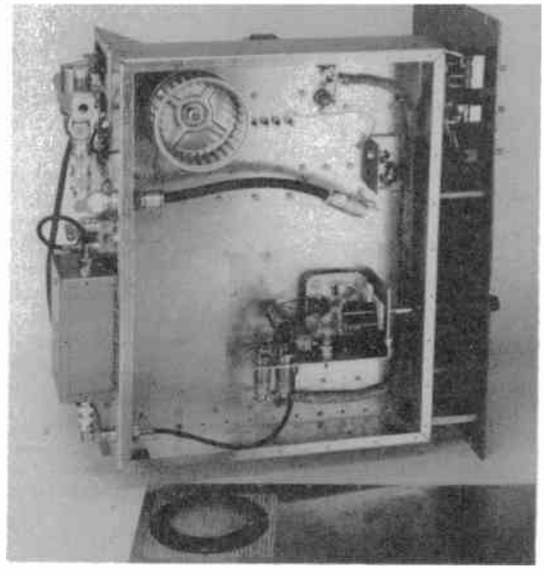
[1] Knadle, "A Strip-Line Kilowatt Amplifier for 432 MHz," *QST*, in two parts: Part I, April, 1972, p. 49; Part II, May, 1972, p. 59.

[2] Belcher, "Rf Matching Techniques, Design and Example," *QST*, October, 1972.

2-kW PEP Amplifier for 144 MHz



The placement of input-circuit components and supporting bracket may be seen in this bottom view. When the bottom cover is in place, the screened air inlet allows the blower to pull air in, pressurizing the entire under-chassis area. The Minibox on the rear apron is a housing for the input reflectometer circuit.

vhf amplifiers to assure maximum power transfer to the load while tuning. Most capacitive-probe output coupling schemes presented to date do not lend themselves to built-in relative-output monitoring circuits. In this amplifier, one of these built-in circuits is achieved quite handily. The circuit consists of a 10:1 resistive voltage divider, diode rectifier, filter and adjustable indicating instrument. Two 7500-ohm, 2-watt carbon resistors are located in the plate compartment connected between the type N rf-output connector and a BNC connector. A small wire was soldered to the center pin of the BNC connector, inside a Minibox, with the 1500-ohm, 1-watt composition resistor and the rectifier diode joined at this point. Relative output voltage is fed, via feedthrough capacitors, to the level-setting potentiometer and multimeter switch.

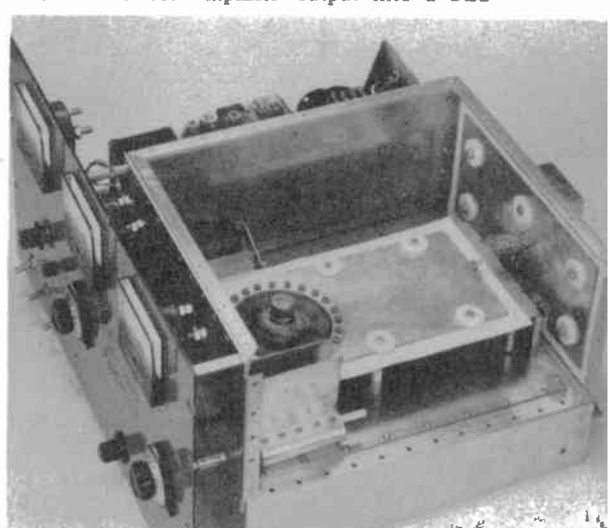
A calibrated string of 2-watt composition resistors, totaling 5 megohms, was installed to facilitate "on-the-spot" determination of power input, and to attest to the presence or absence of high voltage in the plate tank circuit. A full-scale range of 5000 volts is obtained with the 0-1 mA meter. If desired, the builder may use ten 500-K-Ω, 2-watt, 1-percent resistors for the string and reasonable accuracy will be obtained. Of course this monitor feature may be eliminated if other means are used to measure and monitor plate voltage.

The amplifier is unconditionally stable, with no parasitics. To verify this, a zero bias check for stability was made. This involved shorting out the Zener diode in the cathode return lead, reducing bias to essentially zero volts. Plate voltage was applied, allowing the tube to dissipate about 885 watts. The input and output circuits were then tuned through their ranges with no loads attached. There was no sign of output on the relative output meter and no change in the plate and grid currents. As with most cathode-driven amplifiers, there is a slight interaction between grid and plate currents during normal tune-up under rf-applied conditions. This should not be misconstrued as amplifier instability.

Tolerances of the Zener diode used in the cathode return line will result in values of bias voltage and idling plate currents other than those listed in Table I. The 1N3311, a 20-percent tolerance unit, is rated at 12 volts nominal but actually operates at 10 volts in this amplifier (within the 20-percent tolerance).

All testing and actual operation of this amplifier was conducted with a Raytrack high-voltage power supply used in conjunction with the author's 6-meter amplifier. The power supply control and output cable harness was moved from one amplifier to the other, depending on the desired frequency of operation.

Drive requirements were measured for plate power-input levels of 1000 and 1600 watts with a Bird Model 43 Thru Line Wattmeter and a plug of known accuracy. Output power was measured simultaneously with drive requirements at the 1000 and 1600 watt plate power input levels. A second Bird model 43 with a 1000-watt plug was used to measure amplifier output into a Bird



The tube and plate line is in place, with the top and side of the compartment removed for clarity. The plate-tuning vane is at bottom center. A bracket is attached to the side panel to support the rear of the Teflon rod supporting the tuning vane. The coil at the opposite end of the plate line is RFC1, connected between the high-voltage-bypass plate and the top section of the plate-line sandwich. Items outside the tube enclosure include the filament transformer, blower motor, relays, and a power supply to operate a VOX-controlled relay system.

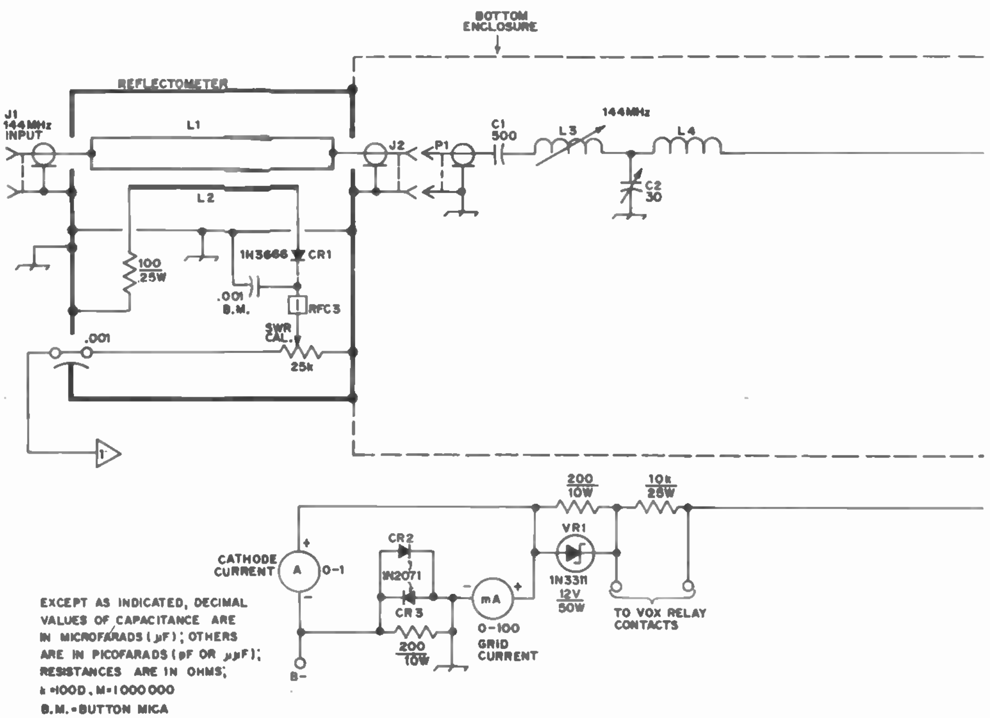
1000-watt Termaline load. A 2500-watt plug would be necessary to determine output power at the 2-kW input level, so I stopped at the 1000-watt output point and worked backwards to calculate apparent stage gain and efficiency.

Efficiency measurements also were made employing the "tube air-stream heat-differential" method. Several runs were made at 885 watts static dc and normal rf input. Apparent efficiencies of 62 to 67 percent were noted. These values were about 5-percent higher than the actual power output values given in Table I. Both efficiency measurement schemes serve to confirm that the amplifier is operating at the upper limit of the theoretical 50-60-percent efficiency range for typical Class AB2 amplifiers.

To commence routine operation, the variable capacitor in the input circuit should be set at the point where lowest input VSWR was obtained during the "cold-tube" initial tube-up. The ability of the plate tank to resonate at 144-145 MHz with the top cover in place should be verified with a grid-dip meter, via a one-turn link attached to the rf output connector. Top and bottom covers are then secured. *As with all cathode driven amplifiers, excitation should never be applied when the tube*

Fig. 1 — Schematic diagram of the amplifier. Included is information for the input reflectometer used as an aid to tuning the cathode circuit for low SWR. C7, C8, and C9 are fabricated as described in the text and Fig. 2.

B1 — Blower. Fasco 59752-IN or Dayton 2C610. Wheel diameter is 3-13/16 inches.
C2 — 5- to 30-pF air variable. Hammarlund HF-30-X or equiv.
C3, C4, C5, C6 — 0.1 μF, 600-V, 20-A feedthrough capacitor. Sprague 8OP3 or equiv.
J1, J2, J6 — Coaxial chassis-mount connectors, type BNC.
J3 — Coaxial connector, type N.
J4 — Coaxial panel jack, UG-22B/U (Amphenol 82-62 or equiv.).
J5 — HV connector (James Millen 37001 or equiv.).
L1 — Double-sided pc board, 1-1/4 × 4-7/16 inches.
L2 — 4-1/4 inches of No. 18 wire. L1 and L2 are part of the input reflectometer circuit described in the text under the heading of "Support Electronics."
L3 — 6 turns No. 18 enam., 5/8-in. long on 3/8-in. dia form (white slug).
L4 — 3 turns No. 14 enam., 5/8-in. long × 9/16-in. ID. Lead length to L3 is 5/8-in. Lead length to cathode bus is 3/4-in.
L5 — Air-dielectric strip line. See text and Fig. 2.
P1 — Coaxial cable connector, type BNC.
P2 — Coaxial cable connector, type N.
R1 — Meter range multiplier. Ten 500-K Ω, 2-watt composition resistors in series.
RFC1 — 7 turns No. 16 tinned, 1/2-in. ID × 1-in. long.
RFC2 — 18 turns No. 18 enam., close wound on 1-megohm, 2-watt composition resistor.
RFC3, RFC4 — Each 2 ferrite beads on component leads.
RFC5, RFC6 — 10 turns No. 12 enam. bifilar wound, 5/8-in. dia.
S1 — Single-pole, three position rotary switch, non-shorting contacts.
T1 — 5-V, 10-A secondary, center tap not used. (Stancor P-6135 or equiv.).
VR1 — 12-V, 50-watt Zener diode.

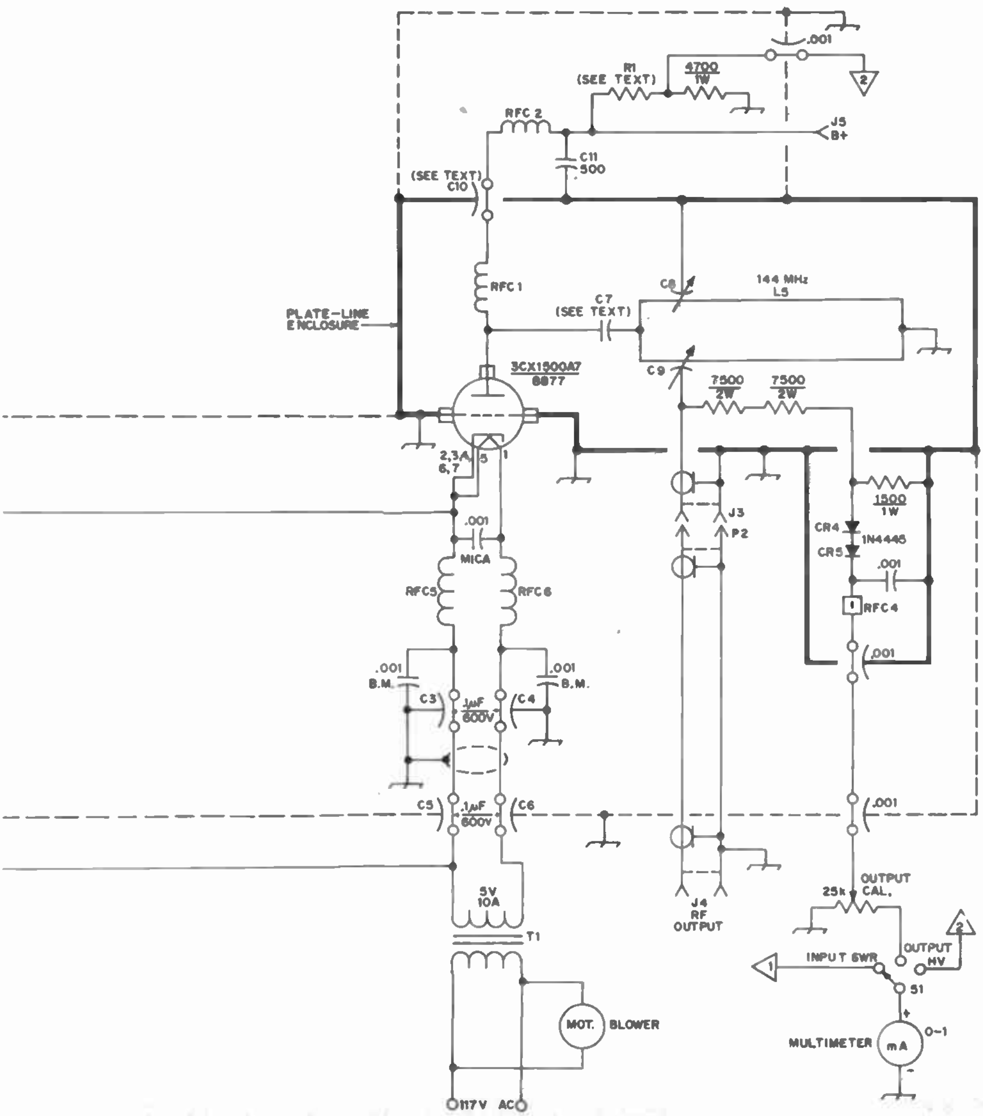


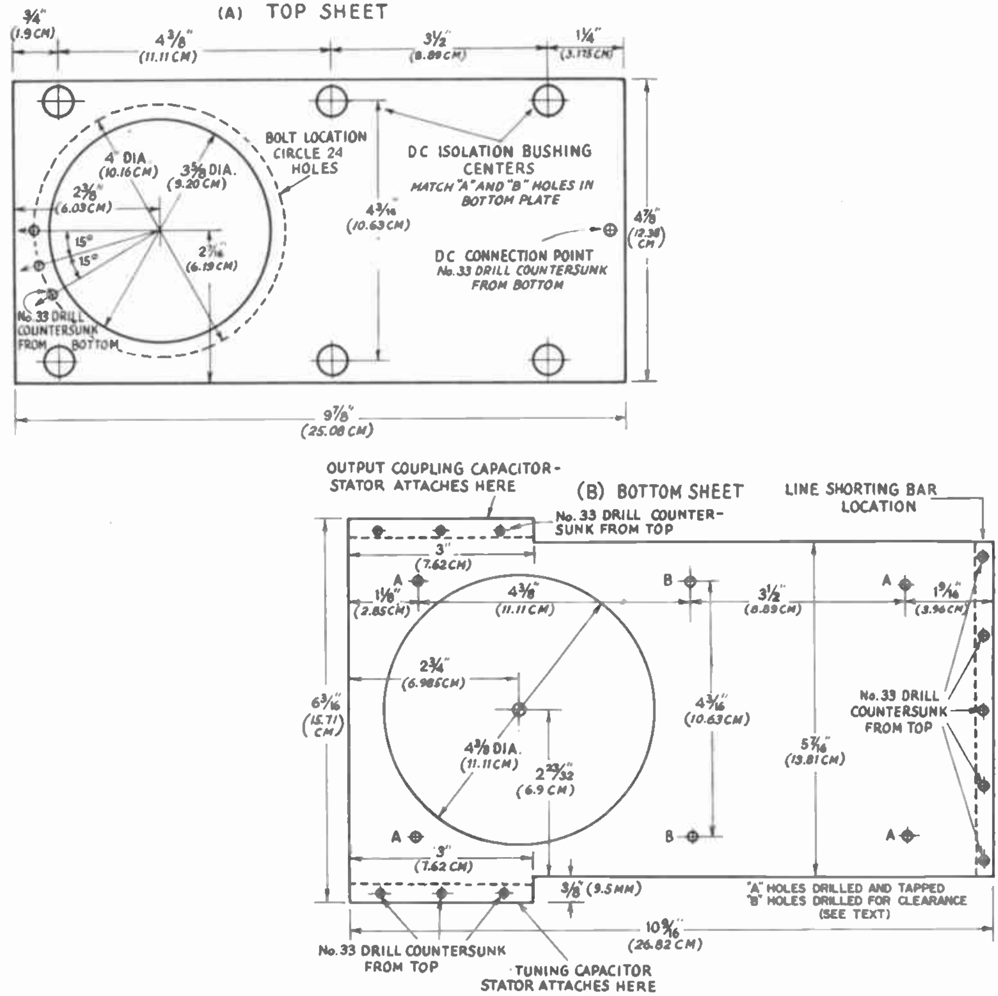
heater is activated and plate voltage is removed. Next, turn on the tube heater and blower simultaneously, allowing 90 seconds for warm-up. Plate potential between 2400-3000 volts then may be applied and its presence verified on the multimeter. The power supply should be able to deliver 800 mA or so. With the VOX relay actuated, resting current should be indicated on the cathode meter. A small amount of drive is applied and the plate tank circuit tuned for an indication of maximum relative power output. The cathode circuit can now be resonated, tuning for minimum reflected power on the reflectometer, and not for maximum drive power transfer. Tuning and loading of the plate-tank circuit follows the standard sequence for any cathode driven amplifier. Resonance is accompanied by a moderate dip in plate/cathode current, a rise in grid current and a considerable increase in relative power output. Plate-current dip is not absolutely coincident with maximum power output but it is very close. Tuning and output-loading adjustments should be for maximum efficiency and output as indicated on the output meter. Final adjustment for lowest VSWR at amplifier input should be done when the deisred plate input-power level has been reached.

Table I

Performance Data

Power input, watts	1000	1600
Plate voltage	2600	2450
Plate current (single tone)	385 mA	660 mA
Plate current (idling)	50 mA	50 mA
Grid bias	−10 V	−10 V
Grid current (single tone)	35 mA	54 mA
Drive power, watts	18	41
Efficiency (apparent)	59.5%	61.8%
Power gain (apparent)	15.2 dB	13.9 dB
Power output, watts	595	1000





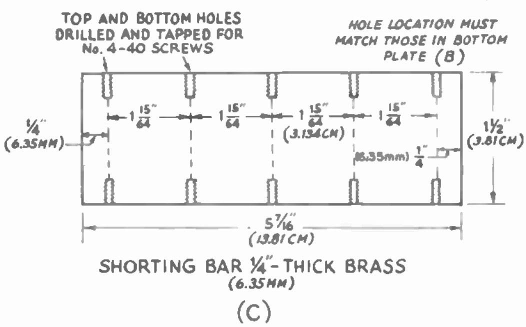


Fig. 2 — Dimensions and layout information for the plate line. The two brass plates and a Teflon sheet form a sandwich with the plate nearest the chassis being at dc ground potential. The top plate carries high voltage and is connected to the tube anode. The shorting bar (C) is permanently attached between the bottom plate and the chassis, replacing the sliding short that is visible in the photographs.

A TRIPLER AMPLIFIER FOR 432 MHZ

Equipment for 432 MHz varies in style, size, complexity, and ancestry. Some stations use converted uhf fm transmitters that once saw duty in taxicabs or the like. Others have been able to build up-converters using tubes such as the 6939.[1] Others have pressed their 144-MHz equipment into service by employing an active frequency tripler.[2,3]

[1]Moretti, "A Heterodyne Exciter for 432 MHz," *QST*, November 1973, (also see Feedback, *QST*, March, 1974, page 83).

[2]Knadle, "High Efficiency Parallel Kilowatt for 432 MHz," *QST*, April, 1972.

[3]Knadle, "Dual-Band Stripline Amplifier-Tripler for 144 and 432 MHz," *Ham Radio*, February, 1970.

The design criteria for a desirable amplifier were simple – a table-top conduction-cooled (quiet) unit that would deliver 100-watts output at a drive level of less than 10 watts. The table-top configuration would be more attractive to many station owners than would the old reliable rack-and-panel system of days gone by. The conduction-cooling requirement was to get away from the blower/air-hose/insulated-box problems that follow the usual external-anode design. At the 100-watt output level, some transmission-line loss could be tolerated and still allow the use of a modest antenna for satellite access.

Amplifier Circuitry

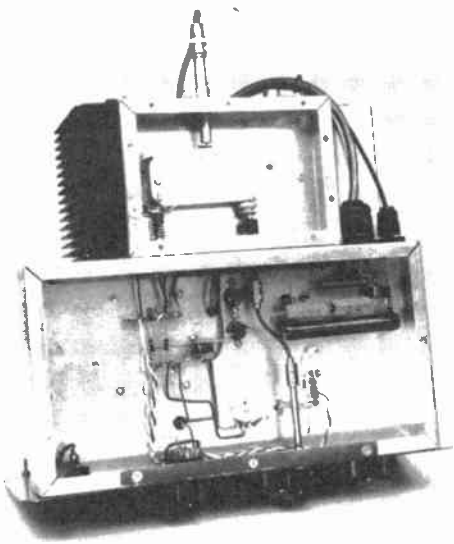
The amplifier draws heavily upon previous designs that utilized the air-cooled, external-anode tubes, as shown in Fig. 1 and in the photographs. A half-wave grid line is fabricated from double-sided pc-board material. The input-coupling method departs slightly from previous examples, but only in the mechanics of adjustment. The plate line is similar to published information, with slight variations in the method of tuning.

Input coupling to the amplifier is by means of a capacitive probe to the grid line. A small tab of copper is soldered to the grid line and forms one side of the capacitor. A disk on the center conductor of a coaxial section is the movable portion of the coupling. This coaxial section is fabricated from pieces of brass tubing that will slide together, telescope fashion. A BNC chassis-mount fitting with the threads filed down is soldered into the inner, movable piece of tubing to allow ease of connection from the exciter. A piece of copper wire and a couple of Teflon disks extend the center conductor for attachment of the capacitor plate inside the grid compartment. Once adjusted, the sliding portion is held in place by means of a small compression clamp.

The plate line is the familiar half-wavelength variety, with capacitive tuning provided by movable vanes or "flappers." In earlier versions using this tuning scheme, the flappers were moved by means of string that was allowed to wind or unwind around a shaft, providing front-panel control. After a few instances of loss of control, caused by the nylon fishing line melting or becoming untied, the writers decided that there had to be a better way. Accordingly, the cam-on-a-rod method was tried and found satisfactory. Both plate-tuning and output-coupling flappers are adjusted in this manner (Fig. 2).

Cooling

Several tests were performed to check the effectiveness of the thermal-link/heat-sink cooling system. With the aid of Temprobes,[4] it was determined that the tube would stay within maximum temperature ratings while dissipating 100 to 200 watts of dc. A liberal coating of thermal-conducting grease was used to aid heat transfer. More on this subject later.



A look at the bottom of the amplifier reveals the grid compartment (top center) and the ac and dc connection cables from the power supply. A grid line is tuned by means of a butterfly type of capacitor mounted on phenolic so that the total capacitance is reduced. A small disk on the end of a coaxial section provides capacitive input coupling to the grid circuit. The flexible coupling shown here has since been replaced by two universal-joint type of connectors to remove some annoying backlash in the tuning control. A high-wattage dropping resistor, part of the screen supply circuitry, is shown at the right.

[4]Temprobes° Test Kit, by Tempil°, Hamilton Blvd., South Plainfield, NJ 07080.

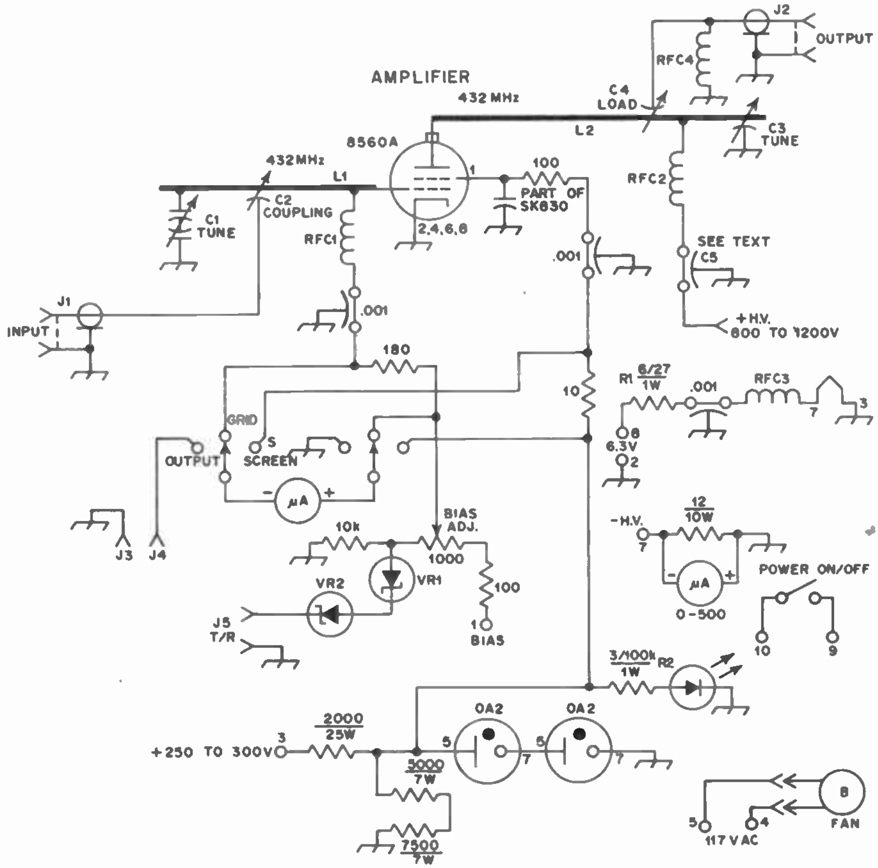


Fig. 1 — Schematic diagram of the 432-MHz amplifier.

C1 — 1.8- to 5.1-pF air variable, E. F. Johnson 160-0205-001. Mount on phenolic bracket.

C2 — 1/2-inch dia. disk on center conductor of coaxial extension. See text and photograph.

C3, C4 — Spring-brass flapper type tuning capacitors. See text and Fig. 2.

C5 — 2-1/2 X 4-inch pc board, single-sided, with .01-inch thick Teflon sheet for insulation to chassis. Copper-foil side mounted toward the chassis wall.

CR1 — 1/4-inch dia. LED.

J1 — BNC chassis-mount connector with threads filed to fit inside brass sleeve.

J2 — Type "N" coaxial connector.

J3, J4 — Tip jacks or binding posts.

J5 — Phono type connector. External relay contacts should be wired to short J5 for "carrier-on" condition.

J6 — High-voltage connector, James Millen 37001.

L1 — 1-3/4 X 4-inch double-sided pc board, spaced 7/8-inch from chassis.

L2 — 3-1/2 X 6-1/4-inch double-sided pc board or aluminum strip. Length from tip of line to tube center is 7-1/8 inches. See Fig. 2.

Heat Sink — Astrodyne No. 3216-0500-A0000, 5 X 5 inches. Can be painted flat black or anodized for better dissipation.

R1 — 27 ohm, 1-W resistor, 6 in parallel.

R2 — 100-kΩ 1-W resistor, 3 in parallel.

Early tests with only dc applied, and later ones with the full dc and rf voltages present, confirmed that at the 100-watt output level no forced-air cooling was required. At higher output levels of 175 to 200 watts, the temperatures on the anode and heat sink were still below the maximum allowed by the manufacturer, but high enough that it was felt prudent to add a quiet "whisper" fan for safety. Operational tests proved that the added background noise was not distracting to the operator.

Construction

There are several configurations possible for the package, and the constructor should feel free to mold them to fit his idea of how things should be assembled. An LMB cabinet (CO-1) was selected for an enclosure because it matches many of the "gray boxes" found in a lot of shacks. Rather than mount the heat sink through an unsightly hole in the rear panel of the cabinet, it was decided to mount the amplifier parallel to the front panel.

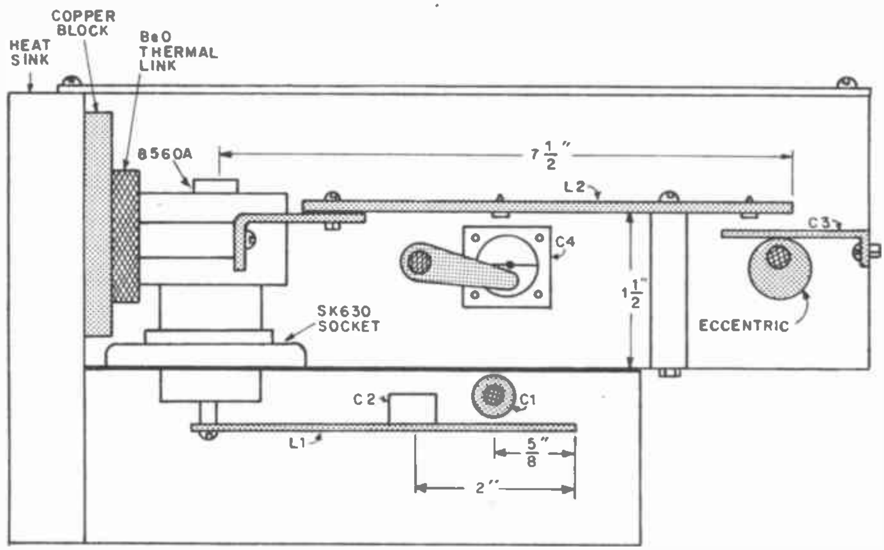


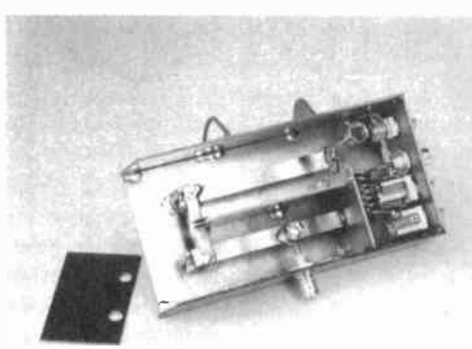
Fig. 2 — Cutaway drawing from the side of the grid and plate compartments. The plate line may be made of two pieces, as shown here, or of one single piece of aluminum strip. C4 is shown from the end-on view. The arm that moves C4 and the eccentric that moves C3 are fastened to their insulated shafts by epoxy cement. Small Teflon buttons prevent accidental shorts between the capacitors and the plate line.

This places the heat sink inside, but there is adequate ventilation through the box to allow proper cooling. This mounting scheme also permits a fan to be mounted inside, so that there are no awkward protuberances to worry about behind the cabinet. A standard size chassis is used to fill the gap between the panel and the amplifier proper, and incidentally to provide a mounting space for peripheral electronics. As long as the parts placement within the amplifier grid and plate compartments is not changed from the design given here, it will not matter what is done externally.

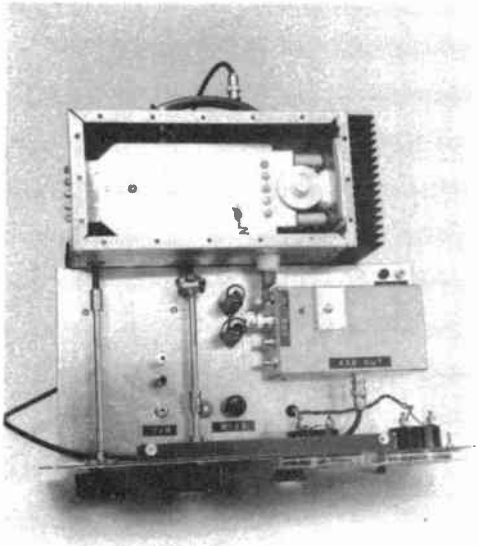
The grid compartment is a 5 X 7 X 2-inch aluminum chassis with captive nuts in the bottom lip to permit securing the bottom plate. For the plate compartment a 5 X 10 X 3-inch aluminum chassis was modified to provide better mounting surfaces for the heat sink and to allow the plate-tuning flapper to be mounted on the end wall of the compartment. One end of the chassis was removed and pieces of aluminum angle stock were fastened around the open end. These pieces were drilled to accept No. 8-32 screws that thread into tapped holes in the heat sink. Tapped holes in the top surface of the heat sink and captive nuts in the top lips of the chassis permit a perforated top plate to be fastened securely for minimum rf leakage. Total dimensions are given in Fig. 2.

Tube Placement

An Eimac SK-630 socket and SK-1920 thermal link are used in mounting the tube and conducting the heat away from the anode. The thermal link is made of toxic beryllium oxide (BeO). The manufacturer's caution against abrasion, fractures, or disposal should be heeded. Parts placement in the anode-block area is critical if efficient heat transfer and minumum strain on the tube are to be obtained. The tube socket must have sufficient clearance in its mounting hole that some lateral movement toward or away from the heat sink is allowed. The socket is secured to the chassis with



The varactor tripler is assembled in a box made from double-sided pc board.



The amplifier chassis is mounted parallel to the front panel. A varactor-diode tripler is mounted on the subchassis, at the right. This view of the amplifier shows the ceramic insulators that provide pressure to hold the tube anode against the thermal link and the heat-sink assembly. A half-wavelength plate line occupies most of the length of the chassis, with a flapper type of tuning capacitor mounted on the left wall. The two VR tubes, center, are regulators for the screen voltage. Insulated shafts extend into the plate compartment, under the plate line, where they rotate eccentric disks to provide tuning control. Two tip jacks at the extreme right allow a cooling fan to be connected, if needed for higher power operation.

the usual toe clamps supplied. Because of the rim formed on the socket by the integral screen-bypass capacitor, a spacer is needed between the thermal link and the heat sink. A piece of copper, 1/4-inch thick and about 2-3/4 X 4-1/2-inches, serves as the spacer, as well as providing excellent heat transfer to the inner face of the heat sink. This copper spacer and the BeO thermal link are both held in place between the tube anode and the heat sink by the pressure applied by the ceramic pillars. The anode end of the plate line is bent up to form a surface that will permit screws to thread into the insulators. In the early version of the amplifier this shaped and bent piece of aluminum was only long enough to provide some mounting surface to which the plate line (double-sided pc board in this instance) was fastened by means of five No. 6-32 screws and nuts with lock washers. The photograph shows this particular scheme in the top view. A later version had the pc board replaced with an aluminum strip of the same size. A still later test was made with the anode-clamp/plate line all constructed from one piece of aluminum. No difference in plate-circuit performance could be noticed, which was the reason for the tests of different materials.

A moderate coating of thermal-conducting grease should be applied between the copper plate, the heat sink, the thermal link, and the anode block. Don't overdo it, however. In one test a glob of the material found its way down to the screen ring, and the combination of rf and dc voltages between the screen and plate caused the material to break down.

The Tripler

The tripler is responsible for the "Tr" part of the name. The frequency tripler, using a varactor diode, is essentially a duplicate of the one described in other ARRL publications.[5] A slight change was made to permit easier adjustment; a 1000-ohm resistor was added in series with the normal bias resistor across the diode. This permits the diode current to be monitored during the tune-up procedure. A rough approximation of correct adjustment can be obtained by tuning the input circuitry for maximum voltage across the 1000-ohm resistor, and then adjusting the idler circuit and the output circuits for a dip in this reading. These adjustments should be made with the varactor output connected to a suitable 50-ohm load; reactive loads will cause the readings to be erratic and confusing. Final adjustments should be made with the aid of SWR meters and a sensitive wavemeter or other spectral-output indicating system. Once the tripler is adjusted for proper operation into a dummy load, *don't touch it.* Further adjustments should be done at the tube grid-input circuit.

Because the tripler construction and the peripheral-electronics chassis layout were not carefully coordinated, there is a distressing lack of space to adjust the tripler input circuits while in place (as can be seen in the photograph). However, if the builder will move the location of the voltage regulator tubes an inch or two to the left, there should be no problem. The tripler is fastened to the chassis by means of spade lugs extending from the vertical members of the tripler box.

Power Supply

Most of the earlier testing of this unit was performed while using the Heath HP-23A to supply

[5] *Radio Amateur's Handbook*, ARRL, 52nd Edition, Chapter 7.

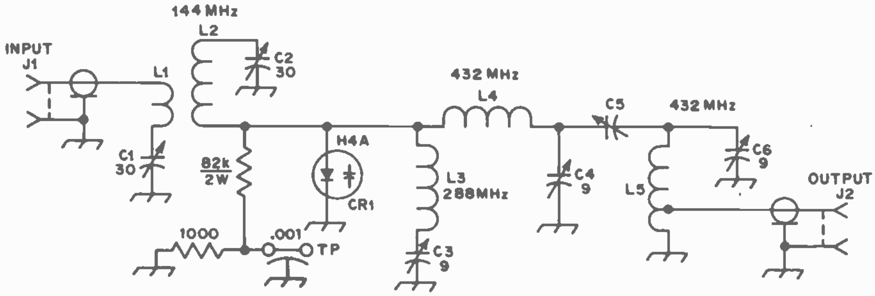


Fig. 3 — Schematic diagram of the varactor tripler.

all voltages. The amplifier can be operated at the 80- to 100-watt output level without unduly taxing the capability of this supply. Accordingly, the wiring and plug connections were made up with this feature in mind. When a larger supply was constructed for tests at the 200-watt level, connections were made compatible with those on the Heath supply as far as practicable. When using the HP-23A, provision must be made to drop the filament potential to the nominal 6.0 V required by the 8560A heater. A voltage-dropping resistor for this purpose is located under the support chassis. Heater voltage should be measured *at the tube socket*, not at the power supply. The newer power supply, HP-23B, can be used if the series resistance added is sufficient to drop the potential from 12 to 6 V as needed by the tube.

Adjustment and Operation

Initial testing should be performed while operating the amplifier at reduced plate and screen voltages, if possible. Output coupling should be at maximum, and the input-coupling probe should be near maximum. Again, do not adjust the tripler circuits to make up for misadjustment of the amplifier. Drive power should be adjusted by increasing or decreasing the 144-MHz excitation to the tripler. An output power indicator should be used as an aid in adjustment of the amplifier.

TABLE I

Operating Conditions

144-MHz drive power watts	*432-MHz drive power watts*	*432-MHz output power watts*
4	2	30
8	4	50
10	5	80
15	7	100
18	9	140

Ep – 1000 V.
Ip – 60 mA, zero signal.
Ip – 300 mA, single tone (cw), 140 W output.

Provisions were made in the wiring to the multimeter switch to display a sample of rf energy, such as might be obtained from a directional coupler.[6] The input-probe spacing and the grid-line tuning should be adjusted for maximum drive to the tube; this should be concurrent with minimum SWR as seen by the tripler. Move the coupling probe in small increments – the proper position will tend to be somewhat difficult to find. Output coupling and plate tuning should be adjusted for maximum output. The reason for starting with maximum coupling is that with minimum coupling and reactive loads, the amplifier could be unstable. Loading should be decreased until there is a smooth, but not sharp, dip in plate current. A reading in the vicinity of 250 to 300 mA at resonance is about right, at a plate potential of 800 V. As with most tubes in this family, maximum output is seldom achieved at minimum plate current. Use the output power as an indication of proper operation, but be sure that the screen is not abused – *small* amounts of *negative* screen current are no cause for alarm. In all cases, do not exceed the power dissipation rating of the tube element concerned.

It is not practical to operate this tube in this configuration at more than 1200-V plate potential. Tests were made at 1500 V, with disastrous results. At that dc level, with the added rf voltage, the stress across the BeO thermal link caused it to become very "unhappy." This caused it to produce frying sounds, which made the authors unhappy. The condition also caused a reduction of plate-circuit efficiency and much unwanted heating of nearby metal parts. Investigation of the phenomenon showed that the high *Q* of the circuit caused the fault. Rather than do a complete redesign of the plate circuit, and because the initial goal was a 100-watt unit, the decision was made to leave well enough alone and recommend a 1200-volt limit. This unit was originally described by *QST* for January, 1976.

[6] McMullen, "The Line Sampler," *QST*, April 1972. Also in *FM and Repeaters for the Radio Amateur*, Chapter 10, and *The Radio Amateur's VHF Manual*, Chapter 14.

A LOW-DRIVE 6-METER PA

Recently, there have been some excellent articles on 6-meter amplifiers in the 1- and 2-kW PEP levels. Usually grounded-grid design is used and the amplifiers require exciters in the 100-watt class. The new popular solid-state 6-meter transceivers that develop approximately 10-watts PEP fall short of the necessary drive for grounded-grid design. This amplifier can easily be driven to 500-watts PEP input by transmitters in this power class.

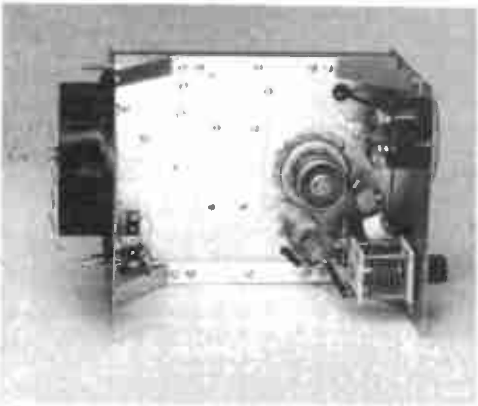
The Circuit

A single 4CX250B is used in a conventional grounded-cathode arrangement (Fig. 1). The tuned grid circuit and the pi-network in the output is a standard design that works well on 6 meters. Driving power is fed into the tuned grid circuit through a 50-ohm T pad. Selection of the correct T-pad value will compensate for driving power of a watt or so up to 25 watts. R5 provides very heavy swamping and assures that the amplifier is completely stable. If R5 resistor is omitted for drive of less than 1 watt, the amplifier will have to be neutralized. Another advantage of the T pad and resistive input is a more constant load to the driving stage.

In the plate circuit, heavy copper-strap conductors are used to provide low inductance leads. The output capacitance of the 4CX250B (4.4 pF) plus strays and the plate tuning capacitance should be 10 to 12 pF for a reasonable circuit Q. The amplifier plate circuit should resonate at 50 MHz with the tuning capacitor (C1) as near minimum value as possible.

Designed and built by Dick Stevens, W1QWJ, this amplifier fills the need for a low-drive model usable with 10-watt exciters.

When drive is provided by a transceiver, a dpdt relay (K1) places the amplifier in the line in the transmit condition and connects the antenna to the transceiver in the receive condition. Cutoff bias is applied to the amplifier in the receive condition and is reduced to the operating value while transmitting by grounding one end of the bias potentiometer, R4. A double set of VOX send/receive contacts is required to perform these two functions (K2).



Top view of the amplifier (note paralleled ceramic capacitors for C4).

Construction

An LMB CO-7 cabinet is used as the basic amplifier housing. It is necessary to add 4 small brackets to stiffen the front and back panels. Two pieces of 1/2 × 1/2-inch Reynolds aluminum angle stock are added to the sides of the built-in chassis to provide additional strength and provide an air

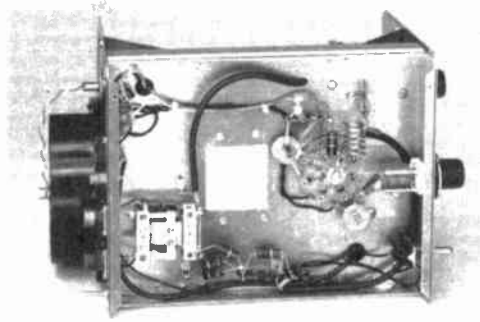
TABLE I

Pad Values for Input Attenuator

Attn (dB)	*R1*	*R2*	*R3 (ohms)*
0	none	–	–
6	18	18	68
10	27	27	39
20	43	43	11

seal between the bottom and top of the chassis. All the perforated holes above the top of the chassis *must* be covered with masking tape to make the top portion of the cabinet airtight. Directly beneath the 4CX250B tube socket, a large hole is punched in the bottom of the cabinet for an air entrance. The photograph of the amplifier shows that the 4CX250B does not have a chimney. It was later found that the chimney *must* be used to provide adequate cooling. A 4-inch diameter hole is cut in the back panel of the cabinet and a 5-inch Roton Whisper fan is mounted over the hole to exhaust air from the cabinet.

Air flow is through the bottom of the cabinet, through the socket of the 4CX250B, through the chimney into the anode and out of the cabinet through the exhaust fan. Very little blower noise is generated using this method of cooling as compared to the conventional squirrel-cage blower fan. The amplifier construction is quite simple as can be seen from the photographs and can be duplicated easily.



Bottom view of the amplifier.

Results

A suitable power supply is shown in Fig. 2. With 2000 volts on the anode and a plate current of 250 mA, the power output as measured by a Bird Thruline wattmeter into a Bird dummy load was 325 watts. This agrees closely with the tube specification sheets.

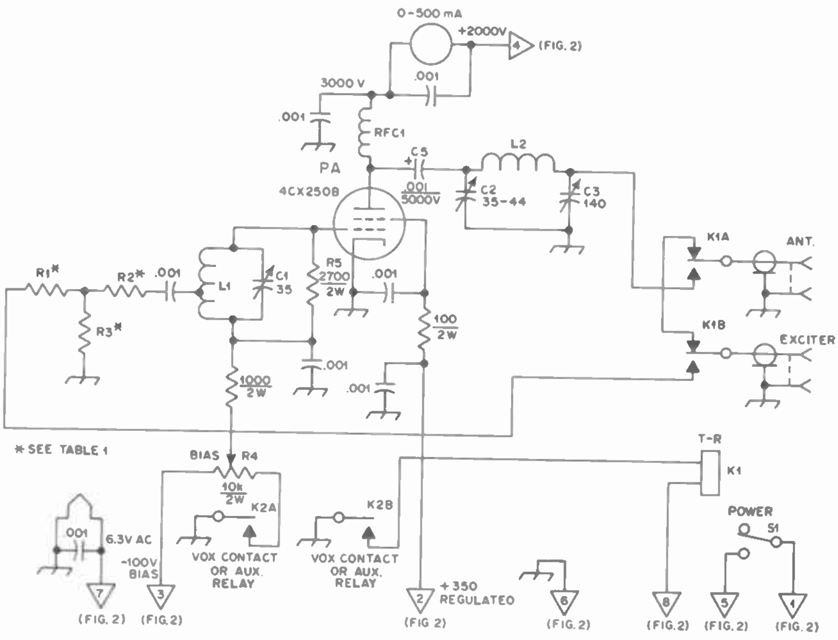


Fig. 1 — Schematic diagram of the amplifier. Unless otherwise specified, capacitors are disk ceramic and resistors are carbon composition.

C1 — 50 pF, receive spacing.
C2 — 25 pF, 3 kV (surplus cap. in unit).
C3 — 140 pF, receive spacing.
C4 — Cer. cap. 2 paralleled 500 pF, 5 kV.
K1 — Dpdt relay, 12-V coil (can have dc/power type contacts but rf design preferable).
K2 — Dpdt relay. Either T-R contacts in exciter or aux. relay if only spst option available.
L1 — 6 turns No. 14 solid wire, 1/2-inch dia, 1-1/4 inch long. Tap 1-1/2 turns from gnd end.
L2 — 5 turns No. 10 solid wire, 1-3/8 dia, 2 inch long (see text).
RFC1 — 35 turns No. 22 enam. wire on 5/8-inch dia cer. ins.

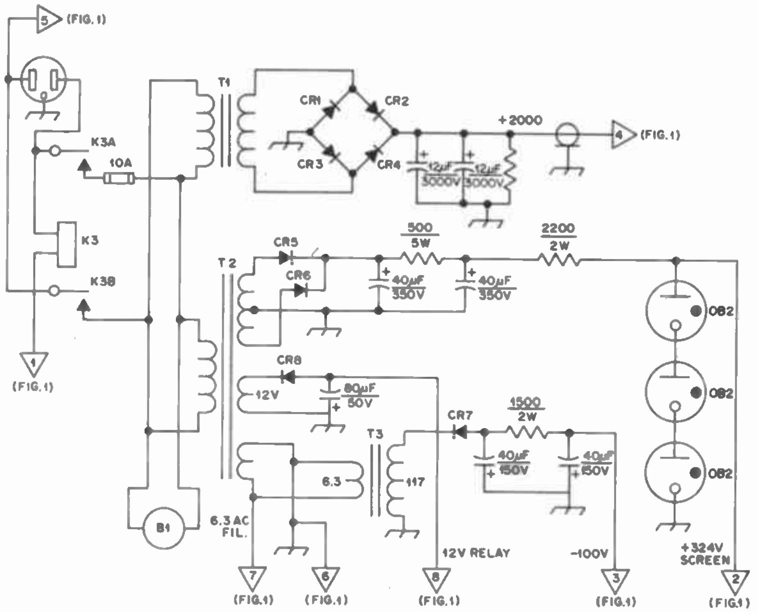


Fig. 2 — Power supply for the amplifier.
B1 — Blower motor (see text).
CR1-CR4, incl. — Each leg consists of 2, series silicon diodes (1 A, 1000 PRV).
CR5, CR6, CR7 — Silicon diode, 1 A, 1000 PRV.
CR8 — Silicon diode, 3 A, 100 PRV.
K3 — Power relay, dpdt 10 A, contacts 117-V ac coil.
T1 — Power transformer, 1400 V ac, 500 mA.
T2 — Power transformer, 500 V at 100 mA sec. 12-V, 1-A sec. and 6.3-V, 3-A sec.
T3 — Filament transformer, 6.3 V, 1A.

Receiving Systems

The performance of a communications receiver can be measured by its ability to pick up weak signals and separate them from the noise and interference while at the same time holding them steady at the same dial settings. The difference between a good receiver and a poor one can be the difference between copying a weak signal well, or perhaps not copying it at all.

Whether the receiver is of home-made or commercial origin, its performance can range from excellent to extremely poor, and high cost or circuit complexity cannot assure proper results. Some of the simplest of receivers can provide excellent results if careful attention is given to their design and proper use. Conversely, the most expensive of receivers can provide poor results if not operated in a competent manner. Therefore, the operator's success at sorting the weak signals out of the noise and interference is dependent upon the correct use of a properly designed, correctly operated receiver.

Communications receivers are rated by their sensitivity (ability to pick up weak signals), their selectivity (the ability to distinguish between signals that are extremely close together in terms of frequency), and by their stability. The latter trait assures that once a stable signal is tuned in it will remain tuned without periodic retuning of the receiver controls (especially the main tuning and BFO controls).

A well-designed modern receiver must be able to receive all of the popular modes of emission if it is to be truly versatile. It should be capable of handling cw, ssb, a-m, fm, and RTTY signals.

The type of detection to be used will depend on the job the receiver is called upon to do. Simple receivers consisting of a single stage of detection (regenerative detector) followed by a one- or two-stage audio amplifier are often adequate for portable and emergency use over short distances. This type of receiver can be quite compact and light weight and can provide many hours of operation from a dry-battery pack if transistorized circuitry is used. Similarly, superregenerative detectors can be used in the same way, but are suitable for copying only a-m and wide-band fm signals. Superheterodyne receivers are the most popular and are capable of better performance than the foregoing types. *Heterodyne* detectors are used for ssb and cw reception in the latter. If a regenerative detector is made to oscillate and provide a steady signal, it is known as an autodyne detector. A beat-frequency oscillator, or BFO, is used to generate a steady signal in the superheterodyne receiver. This signal is applied to the detector stage to permit the reception of ssb and cw signals.

Communications receivers should have a slow tuning rate and a smooth-operating tuning-dial mechanism if any reasonable degree of selectivity is used. Without these features cw and ssb signals are extremely hard to tune in. In fact, one might easily tune past a weak signal without knowing it was there if a fast tuning rate were used.

Fig. 8-1 – The success of amateur on-the-air operation is, in a large part, determined by a receiver. A good receiver, mated with a good pair of ears, is an unbeatable combination.

RECEIVER CHARACTERISTICS

Sensitivity

In commercial circles "sensitivity" is defined as the signal at the input of the receiver required to give a signal-plus-noise output some stated ratio (generally 10 dB) above the noise output of the receiver. This is a useful sensitivity measure for the amateur, since it indicates how well a weak signal will be heard. However, it is not an absolute method, because the bandwidth of the receiver plays a large part in the result.

The random motion of the molecules in the antenna and receiver circuits generates small

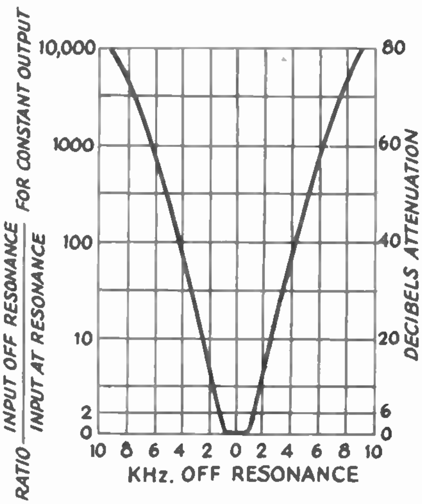


Fig. 8-2 – Typical selectivity curve of a modern superheterodyne receiver. Relative response is plotted against deviations above and below the resonance frequency. The scale at the left is in terms of voltage ratios, the corresponding decibel steps are shown at the right.

voltages called thermal-agitation noise. Thermal-agitation noise is independent of frequency and is proportional to the (absolute) temperature, the resistive component of the impedance across which the thermal agitation is produced, and the bandwidth. Noise is generated in vacuum tubes and semiconductors by random irregularities in the current flow within them; it is convenient to express this shot-effect noise as an equivalent resistance in the grid circuit of a noise-free tube. This equivalent noise resistance is the resistance (at room temperature) that placed in the grid circuit of a noise-free tube will produce plate-circuit noise equal to that of the actual tube. The equivalent noise resistance of a vacuum tube increases with frequency.

An ideal receiver would generate no noise in its tubes or semiconductors and circuits, and the minimum detectable signal would be limited only by the thermal noise in the antenna. In a practical receiver, the limit is determined by how well the amplified antenna noise overrides the other noise of the input stage. (It is assumed that the first stage of any good receiver will be the determining factor; the noise contributions of subsequent stages should be insignificant by comparison.) At frequencies below 20 or 30 MHz the site noise (atmospheric and man-made noise) is generally the limiting factor.

The degree to which a practical receiver approaches the quiet ideal receiver of the same bandwidth is given by the noise figure of the receiver. Noise figure is defined as the ratio of the signal to noise power ratio of the ideal receiver to the signal-to-noise power ratio of the actual receiver output. Since the noise figure is a ratio, it is usually given in decibels; it runs around 5 to 10 dB for a good communications receiver below 30 MHz. Although noise figures of 2 to 4 dB can be obtained, they are of little or no use below 30 MHz except in extremely quiet locations or when a very small antenna is used. The noise figure of a receiver is not modified by changes in bandwidth.

Selectivity

Selectivity is the ability of a receiver to discriminate against signals of frequencies differing from that of the desired signal. The overall selectivity will depend upon the selectivity and the number of the individual tuned circuits.

The selectivity of a receiver is shown graphically by drawing a curve that gives the ratio of signal strength required at various frequencies off resonance to the signal strength at resonance, to give constant output. A resonance curve of this type is shown in Fig. 8-2. The bandwidth is the width of the resonance curve (in Hz or kHz) of a receiver at a specified ratio; in the typical curve of Fig. 8-2 the bandwidths for response ratios of 2 and 1000 (described as "-6 dB" and "-60 dB") are 2.4 and 12.2 kHz respectively.

The bandwidth at 6 dB down must be sufficient to pass the signal and its sidebands if faithful reproduction of the signal is desired. However, in the crowded amateur bands, it is generally advisable to sacrifice fidelity for intelligibility. The ability to reject adjacent-channel signals depends upon the skirt selectivity of the receiver, which is determined by the bandwidth at high attenuation. In a receiver with excellent skirt selectivity, the ratio of the 6-dB bandwidth to the 60-dB bandwidth will be about 0.2 for code and 0.3 for phone. The minimum usable bandwidth at 6 dB down is approximately 150 Hz for code reception and approximately 2000 Hz for phone.

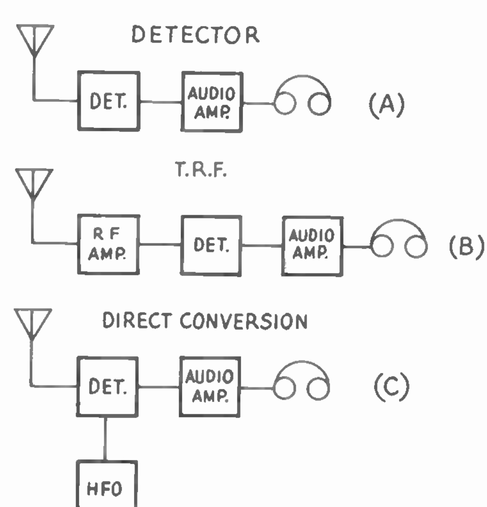


Fig. 8-3 – Block diagrams of three simple receivers.

Stability

The stability of a receiver is its ability to "stay put" on a signal under varying conditions of gain-control setting, temperature, supply-voltage changes and mechanical shock. The term "unstable" is also applied to a receiver that breaks into oscillation or a regenerative condition with some settings of its controls that are not specifically intended to control such a condition.

SIMPLE RECEIVERS

The simplest receiver design consists of a detector followed by an audio amplifier, as shown in Fig. 8-3A. Obviously, the sensitivity of the detector determines how well this receiver will work. Various schemes have been developed to increase detector sensitivity, including the regenerative and superregenerative detectors described later in this chapter. Another way to increase receiver sensitivity is to add one or more rf-amplifier stages before the detector. This approach is called the tuned-radio-frequency, or TRF receiver, Fig. 8-3B.

Another design which has become popular for use in battery-powered equipment is the direct-conversion receiver, Fig. 8-3C. Here, a detector is employed along with a variable-frequency oscillator which is tuned just slightly off the frequency of the incoming signal to produce a beat note. A narrow-bandwidth audio filter located between the detector and the aduio amplifier provides selectivity. However, the lack of automatic gain control limits the range over which the receiver can handle strong signals unless a manual rf-gain control is employed. FETs and ICs can be used as detectors to provide up to 90 dB of dynamic range – typically 3 μV to 100 mV of input signal.

DETECTION AND DETECTORS

Detection (demodulation) is the process of extracting the signal information from a modulated carrier wave. When dealing with an a-m signal, detection involves only the rectification of the rf signal. During fm reception, the incoming signal must be converted to an a-m signal for detection. See Chapter 14.

Detector sensitivity is the ratio of desired detector output to the input. Detector linearity is a measure of the ability of the detector to reproduce the exact form of the modulation on the incoming signal. The resistance or impedance of the detector is the resistance or impedance it presents to the circuits it is connected to. The input resistance is important in receiver design, since if it is relatively low it means that the detector will consume power, and this power must be furnished by the preceding stage. The signal-handling capability means the ability to accept signals of a specified amplitude without overloading or distortion.

Diode Detectors

The simplest detector for a-m is the diode. A germanium or silicon crystal is an imperfect form of diode (a small current can usually pass in the reverse direction), but the principle of detection in a semiconductor diode is similar to that in a vacuum-tube diode.

Circuits for both half-wave and full-wave diodes are given in Fig. 8-4. The simplified half-wave circuit at Fig. 8-4A includes the rf tuned circuit, L2C1, a coupling coil, L1, from which the rf energy is fed to L2C1, and the diode, CR1, with its load resistance, R1, and bypass capacitor, C2.

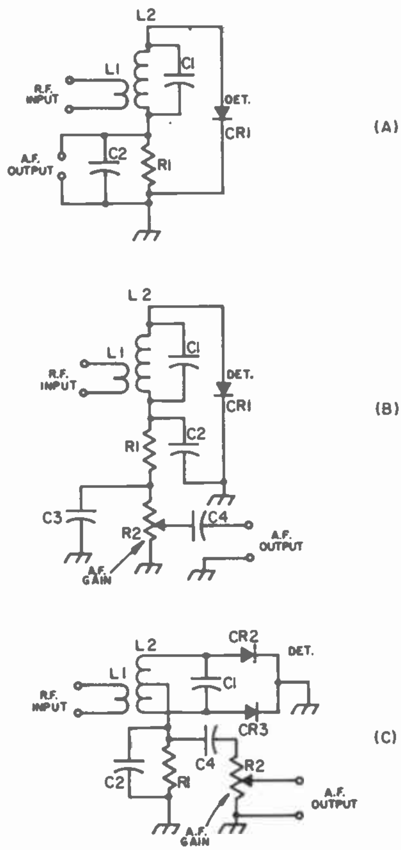


Fig. 8-4 – Simplified and practical diode detector circuits. A, the elementary half-wave diode detector; B, a practical circuit, with rf filtering and audio output coupling; C, full-wave diode detector, with output coupling indicated. The circuit, L2C1, is tuned to the signal frequency; typical values for C2 and R1 in A and C are 250 pF and 250,000 ohms, respectively; in B, C2 and C3 are 100 pF each; R1, 50,000 ohms; and R2, 250,000 ohms. C4 is 0.1 μF and R3 may be 0.5 to 1 megohm.

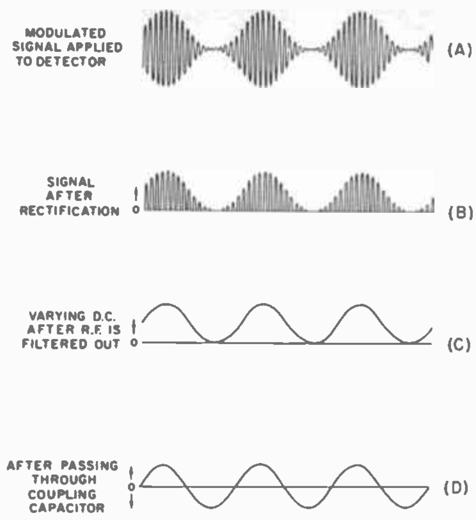


Fig. 8-5 — Diagrams showing the detection process.

The progress of the signal through the detector or rectifier is shown in Fig. 8-5. A typical modulated signal as it exists in the tuned circuit is shown at A. When this signal is applied to the rectifier, current will flow only during the part of the rf cycle when the anode is positive with respect to cathode, so that the output of the rectifier consists of half-cycles of rf. These current pulses flow in the load circuit comprised of R1 and C2, the resistance of R1 and the capacitance of C2 being so proportioned that C2 charges to the peak value of the rectified voltage on each pulse and retains enough charge between pulses so that the voltage across R1 is smoothed out, as shown in C. C2 thus acts as a filter for the radio-frequency component of the output of the rectifier, leaving a dc component that varies in the same way as the modulation on the original signal. When this varying dc voltage is applied to a following amplifier through a coupling capacitor (C4 in Fig. 8-4), only the *variations* in voltage are transferred, so that the final output signal is ac, as shown in D.

In the circuit at 8-4B, R1 and C2 have been divided for the purpose of providing a more effective filter for rf. It is important to prevent the appearance of any rf voltage in the output of the detector, because it may cause overloading of a succeeding amplifier stage. The audio-frequency variations can be transferred to another circuit through a coupling capacitor, C4. R2 is usually a "potentiometer" so that the audio volume can be adjusted to a desired level.

Coupling from the potentiometer (volume control) through a capacitor also avoids any flow of dc through the moving contact of control. The flow of dc through a high-resistance volume control often tends to make the control noisy (scratchy) after a short while.

The full-wave diode circuit at 8-4C differs in operation from the half-wave circuit only in that both halves of the rf cycle are utilized. The full-wave circuit has the advantage that rf filtering is easier than in the half-wave circuit. As a result, less attenuation of the higher audio frequencies will be obtained for any given degree of rf filtering.

The reactance of C2 must be small compared to the resistance of R1 at the radio frequency being rectified, but at audio frequencies must be relatively large compared to R1. If the capacitance of C2 is too large, response at the higher audio frequencies will be lowered.

Compared with most other detectors, the gain of the diode is low, normally running around 0.8 in audio work. Since the diode consumes power, the *Q* of the tuned circuit is reduced, bringing about a reduction in selectivity. The loading effect of the diode is close to one half the load resistance. The detector linearity is good, and the signal-handling capability is high.

Plate Detectors

The plate detector is arranged so that rectification of the rf signal takes place in the plate circuit of the tube or the collector of an FET. Sufficient negative bias is applied to the grid to bring the plate current nearly to the cutoff point, so that application of a signal to the grid circuit causes an increase in average plate current. The average plate current follows the changes in the signal in a fashion similar to the rectified current in a diode detector.

In general, transformer coupling from the plate circuit of a plate detector is not satisfactory, because the plate impedance of any tube is very high when the bias is near the plate-current cutoff point. The same is true of a JFET or MOSFET. Impedance coupling may be used in place of the

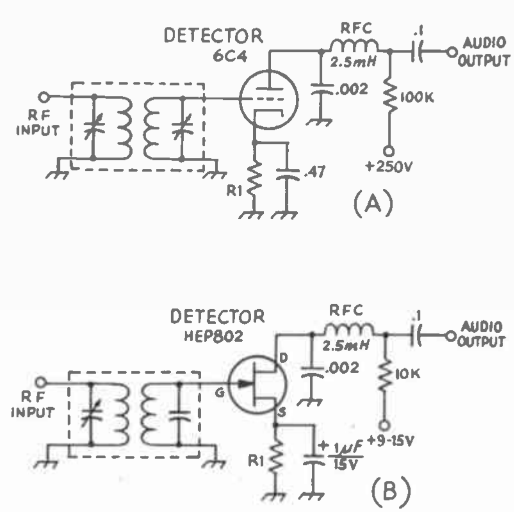


Fig. 8-6 — Circuits for plate detection. A, triode; B, FET. The input circuit, L2C1, is tuned to the signal frequency. Typical values for R1 are 22,000 to 150,000 ohms for the circuit at A, and 4700 to 22,000 ohms for B.

resistance coupling shown in Fig. 8-6. Usually 100 henrys or more of inductance are required.

The plate detector is more sensitive than the diode because there is some amplifying action in the tube or transistor. It will handle large signals, but is not so tolerant in this respect as the diode. Linearity, with the self-biased circuits shown, is good. Up to the overload point the detector takes no power from the tuned circuit, and so does not affect its Q and selectivity.

Infinite-Impedance Detector

The circuit of Fig. 8-7 combines the high signal-handling capabilities of the diode detector with low distortion and, like the plate detector, does not load the tuned circuit it connects to. The circuit resembles that of the plate detector, except that the load resistance, 27kΩ, is connected between source and ground and thus is common to both gate and drain circuits, giving negative feedback for the audio frequencies. The source resistor is bypassed for rf but not for audio, while the drain circuit is bypassed to ground for both audio and radio frequencies. An rf filter can be connected between the cathode and the output coupling capacitor to eliminate any rf that might otherwise appear in the output.

The drain current is very low at no signal, increasing with signal as in the case of the plate detector. The voltage drop across the source resistor consequently increases with signal. Because of this and the large initial drop across this resistor, the gate usually cannot be driven positive by the signal.

HETERODYNE AND PRODUCT DETECTORS

Any of the foregoing a-m detectors becomes a heterodyne detector when a local-oscillator (BFO) is added to it. The BFO signal amplitude should be 5 to 20 times greater than that of the strongest incoming cw or ssb signal if distortion is to be minimized. These heterodyne detectors are frequently used in receivers that are intended for a-m as well as cw and ssb reception. A single detector can thus be used for all three modes, and elaborate switching techniques are not required. To receive a-m it is merely necessary to disable the BFO circuit.

The name product detector has been given to heterodyne detectors in which special attention has been paid to minimizing distortion and intermodulation (IM) products. Product detectors have been thought of by some as a type of detector whose output signal vanishes when the BFO signal is removed. Although some product detectors function that way, such operation is not a criterion. A *product* is something that results from the combination of two or more things, hence any heterodyne detector can rightfully be regarded as a product detector. The two input signals (i-f and BFO) are fed into what is essentially a mixer stage. The difference in frequency (after filtering out and removing the i-f and BFO signals from the mixer output) is fed to the audio amplifier stages and increased to speaker or headphone level. Although product detectors are intended primarily for use with cw and ssb signals, a-m signals can be copied satisfactorily on receivers which have good i-f selectivity. The a-m signal is tuned in as though it were an ssb signal. When properly tuned, the heterodyne from the a-m carrier is not audible.

A triode product-detector circuit is given in Fig. 8-8A. The i-f signal is fed to the grid of the tube, while the BFO energy is supplied to the cathode. The two signals mix to produce audio-frequency output from the plate circuit of the tube. The BFO voltage should be about 2 V rms and the signal should not exceed 0.3 V rms for linear detection. The degree of plate filtering required will depend on the frequency of operation. The values shown in Fig. 8-8A are sufficient for 450-kHz operation. At low frequencies more elaborate filtering is needed. A similar circuit using a JFET is shown at B.

In the circuit of Fig. 8-8C, two germanium diodes are used, though a 6AL5 tube could be substituted. The high back resistance of the diodes is used as a dc return; if a 6AL5 is used the diodes must be shunted by 1-megohm resistors. The BFO signal should be at least 10 or 20 times the amplitude of the incoming signal.

At Fig. 8-8D a two-diode circuit, plus one transistor, provides both a-m and product detection. This circuit is used in the Drake SPR-4 receiver. Balanced output is required from the BFO. The a-m detector is forward biased to prevent the self-squelching effect common to single-diode detectors (caused by signals of low level not exceeding the forward voltage drop of the diode). The IC detector given in Fig. 8-8E has several advantages. First, the BFO injection only needs to be equal to the input signal, because of the additional amplification of the BFO energy which takes place within the IC. Also, output filtering is quite simple, as the double-balanced design reduces the level of i-f signal and BFO voltage appearing in the output circuit. Motorola's MC1496G has a dynamic range of 90 dB and a conversion gain of about 12 dB, making it a good choice for use in a direct-conversion receiver.

A multipurpose IC i-f amplifier/detector/agc system, the National Semiconductor LM373, is shown in Fig. 8-8F. A choice of a-m, ssb, cw, and fm detection is available, as well as a 60-dB-range agc system and i-f amplification of 70 dB. Recovered audio is typically 120 mV. L1C1 tune to the i-f frequency.

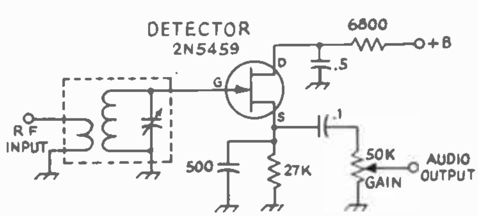


Fig. 8-7 — The infinite-impedance detector. The input circuit, L2C1, is tuned to the signal frequency.

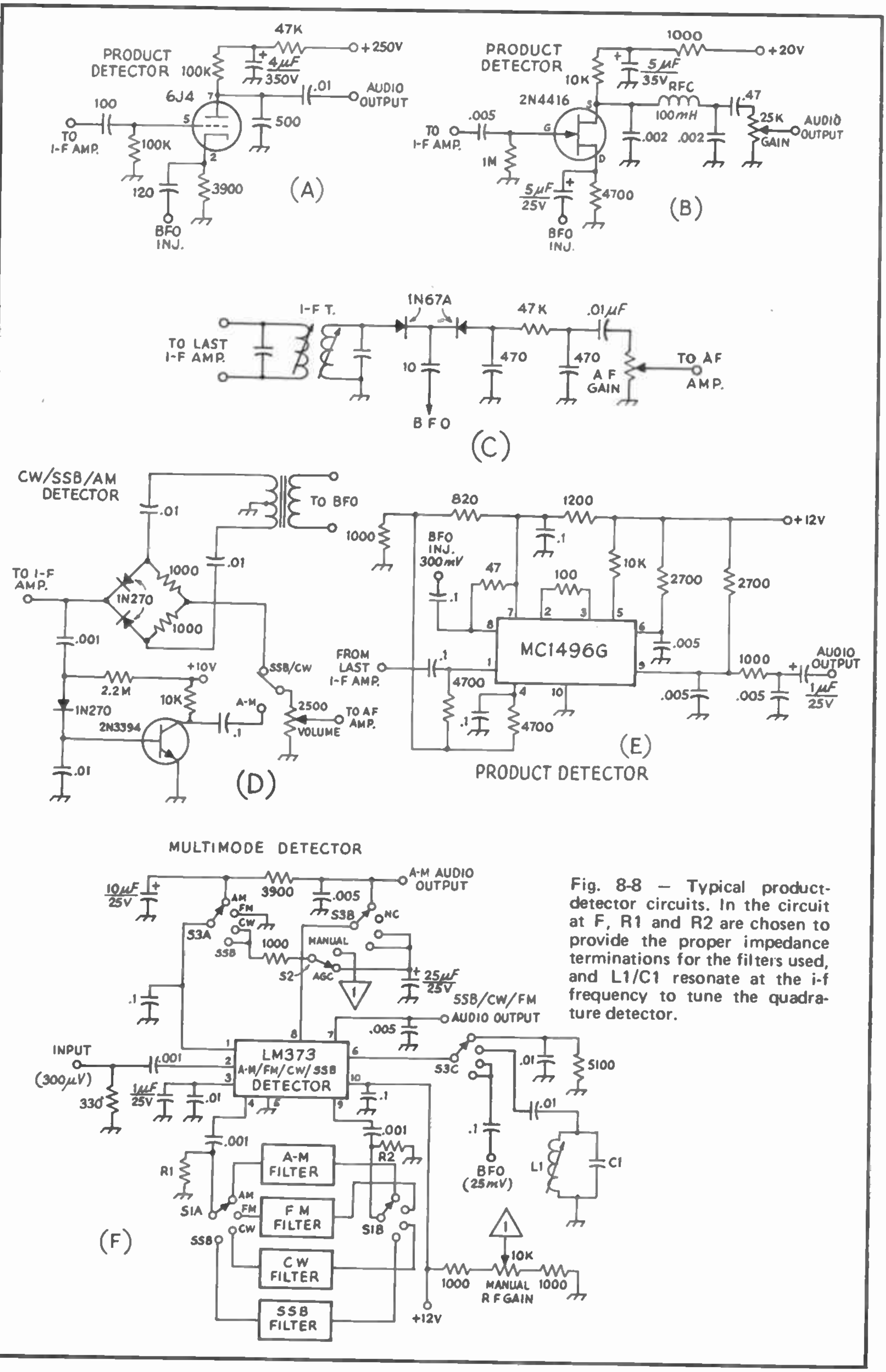


Fig. 8-8 — Typical product-detector circuits. In the circuit at F, R1 and R2 are chosen to provide the proper impedance terminations for the filters used, and L1/C1 resonate at the i-f frequency to tune the quadrature detector.

REGENERATIVE DETECTORS

By providing controllable rf feedback (regeneration) in a triode, pentode, or transistorized-detector circuit, the incoming signal can be amplified many times, thereby greatly increasing the sensitivity of the detector. Regeneration also increases the effective *Q* of the circuit and thus the selectivity. The grid-leak type of detector is most suitable for the purpose.

The grid-leak detector is a combination diode rectifier and audio-frequency amplifier. In the circuit of Fig. 8-9A, the grid corresponds to the diode plate and the rectifying action is exactly the same as in a diode. The dc voltage from rectified-current flow through the grid leak, R1, biases the grid negatively, and the audio-frequency variations in voltage across R1 are amplified through the tube as in a normal af amplifier. In the plate circuit, R2 is the plate-load resistance and C3 and RFC a filter to eliminate rf in the output circuit.

A grid-leak detector has considerably greater sensitivity than a diode. The sensitivity is further increased by using a screen-grid tube instead of a triode. The operation is equivalent to that of the triode circuit. The screen bypass capacitor should have low reactance for both radio and audio frequencies.

The circuit in Fig. 8-9B is regenerative, the feedback being obtained by feeding some signal from the drain circuit back to the gate by inductive coupling. The amount of regeneration must be controllable, because maximum regenerative amplification is secured at the critical point where the circuit is just about to oscillate. The ciritical point in turn depends upon circuit conditions, which may vary with the frequency to which the detector is tuned. An oscillating detector can be detuned slightly from an incoming cw signal to give *autodyne* reception. The circuit of Fig. 8-9B uses a control which varies the supply voltage to control regeneration. If L2 and L3 are wound end to end in the same direction, the drain connection is to the outside of the "tickler" coil, L3, when the gate connection is to the outside end of L2.

Although the regenerative detector is more sensitive than any other type, its many disadvantages commend it for use only in the simplest receivers. The linearity is rather poor, and the signal-handling capability is limited. The signal-handling capability can be improved by reducing R1 to 0.1 megohm, but the sensitivity will be decreased. The degree of antenna coupling is often critical.

A bipolar transistor is used in a regenerative detector hookup at C. The emitter is returned to dc ground through a 1000-ohm resistor and a 50,000-ohm regeneration control. The 1000-ohm resistor keeps the emitter above ground at rf to permit feedback between the emitter and collector. A 5-pF capacitor (more capacitance might be required) provides the feedback path. C1 and L2 comprise the tuned circuit, and the detected signal is taken from the collector return through T1. A transistor with medium or high beta works best in circuits of this type and should have a frequency rating which is well above the desired operating frequency. The same is true of the frequency rating of any FET used in the circuit at B.

Superregenerative detectors are somewhat more sensitive than straight regenerative detectors and can employ either tubes or transistors. An in-depth discussion of superregenerative detectors is given in Chapter 9.

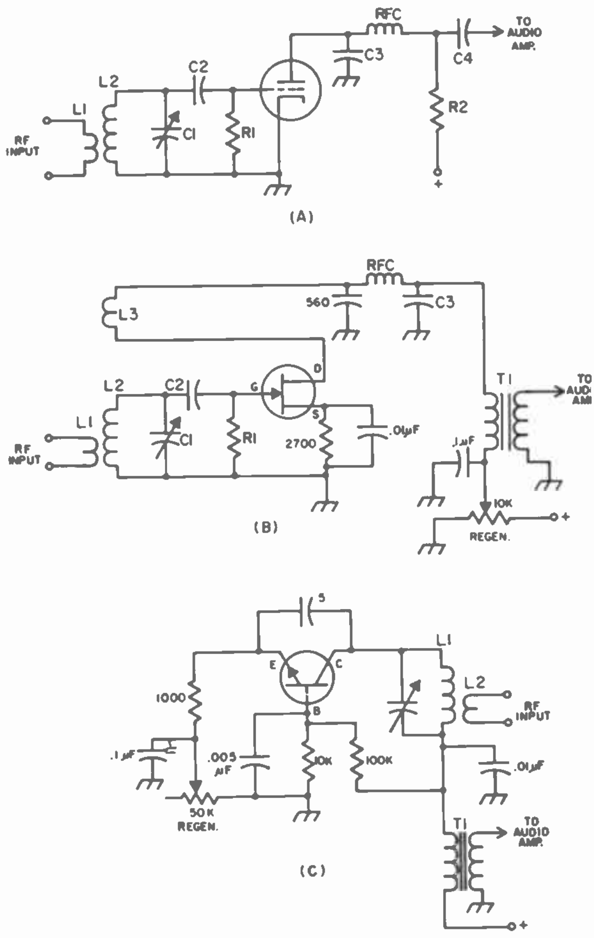


Fig. 8-9 – (A) Triode grid-leak detector combines diode detection with triode amplification. Although shown here with resistive plate load, R2, an audio choke coil or transformer could be used.

(B) Feeding some signal from the drain circuit back to the gate makes the circuit regenerative. When feedback is sufficient, the circuit will oscillate. The regeneration is adjusted by a 10,000-ohm control which varies the drain voltage.

(C) An npn bipolar transistor can be used as a regenerative detector too. Feedback occurs between collector and emitter through the 5-pF capacitor. A 50,000-ohm control in the emitter return sets the regeneration. Pnp transistors can also be used in this circuit, but the battery polarity must be reversed.

Tuning

For cw reception, the regeneration control is advanced until the detector breaks into a "hiss" which indicates that the detector is oscillating. Further advancing of the regeneration control will result in a slight decrease in the hiss.

Code signals can be tuned in and will give a tone with each signal depending on the setting of the tuning control. A low-pitched beat note cannot be obtained from a strong signal because the detector "pulls in" or "blocks."

The point just after the detector starts oscillating is the most sensitive condition for code reception. Further advancing the regeneration control makes the receiver less prone to blocking, but also less sensitive to weak signals.

If the detector is in the oscillating condition and an a-m phone signal is tuned in, a steady audible beat-note will result. While it is possible to listen to phone if the receiver can be tuned to exact zero beat, it is more satisfactory to reduce the regeneration to the point just before the receiver goes into oscillation. This is also the most sensitive operating point.

TUNING METHODS

Tuning

The resonant frequency of a circuit can be shifted by changing either the inductance or the capacitance in the circuit. Panel control of inductance (permeability-tuned oscillator, or PTO) is used to tune a few commercial receivers, but most receivers depend upon panel-mounted variable capacitors for tuning.

Tuning Rate

For ease in tuning a signal, it is desirable that the receiver have a tuning rate in keeping with the type of signal being received and also with the selectivity of the receiver. A tuning rate of 500 kHz per knob revolution is normally satisfactory for a broadcast receiver, but 100 kHz per revolution is almost too fast for easy ssb reception – around 25 to 50 kHz being more desirable.

Band Changing

The same coil and tuning capacitor cannot be used for, say, 3.5 to 14 MHz because of the impracticable maximum-to-minimum capcitance ratio required. It is necessary, therefore, to provide a means for changing the circuit constants for various frequency bands. As a matter of convenience the same tuning capacitor usually is retained, but new coils are inserted in the circuit for each band.

One method of changing inductances is to use a switch having an appropriate number of contacts, which connects the desired coil and disconnects the others. The unused coils are sometimes short-circuited by the switch, to avoid undesirable self-resonances.

Another method is to use coils wound on forms that can be plugged into suitable sockets. These plug-in coils are advantageous when space is at a premium, and they are also very useful when considerable experimental work is involved.

Bandspreading

The tuning range of a given coil and variable capacitor will depend upon the inductance of the coil and the change in tuning capacitance. To cover a wide frequency range and still retain a suitable tuning rate over a relatively narrow frequency range requires the use of bandspreading. Mechanical bandspreading utilizes some mechanical means to reduce the tuning rate; a typical example is the two-speed planetary drive to be found in some receivers. Electrical bandspreading is obtained by using a suitable circuit configuration. Several of these methods are shown in Fig. 8-10.

In A, a small bandspread capacitor, C1 (15- to 25-pF maximum), is used in parallel with capacitor C2, which is usually large enough (100 to 140 pF) to cover a 2-to-1 frequency range. The setting of C2 will determine the minimum capacitance of the circuit, and the maximum capacitance for bandspread tuning will be the maximum capacitance of C1 plus the setting of C2. The inductance of the coil can be adjusted so that the maximum-minimum ratio will give adequate bandspread. It is almost impossible, because of the nonharmonic relation of the various band limits, to get full bandspread on all bands with the same pair of capacitors. C2 is variously called the bandsetting or main-tuning capacitor. It must be reset each time the band is changed.

If the capacitance change of a tuning capacitor is known, the total fixed shunt capacitance (Fig. 8-10A) for covering a band of frequencies can be found from Fig. 8-11.

Example: What fixed shunt capacitance will allow a capacitor with a range of 5 to 30 pF to tune 3.45 to 4:05 MHz?

$$(4.05 - 3.45) \div 4.05 = 0.148$$

From Fig. 8-11, the capacitance ratio is 0.38, and hence the minimum capacitance is $(30 - 5) \div 0.38 = 66$ pF. The 5-pF minimum of the tuning capacitor, the tube capacitance and any stray capacitance must be included in the 66 pF.

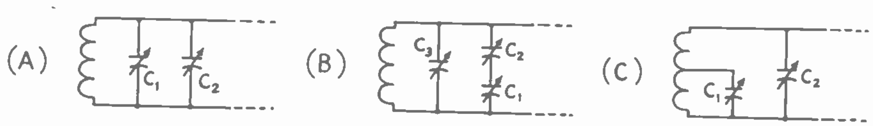


Fig. 8-10 – Essentials of the three basic bandspread tuning systems.

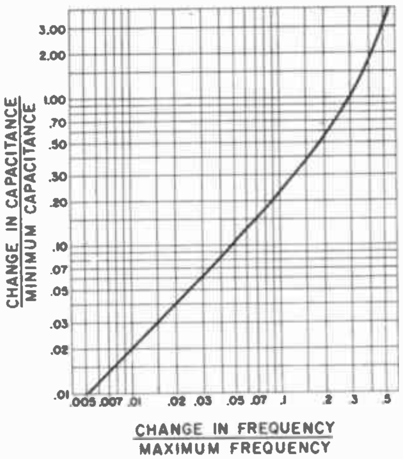


Fig. 8-11 — Minimum circuit capacitance required in the circuit of Fig. 8-10A as a function of the capacitance change and the frequency change. Note that *maximum* frequency and *minimum* capacitance are used.

The method shown at Fig. 8-10B makes use of capacitors in series. The tuning capacitor, C1, may have a maximum capacitance of 100 pF or more. The minimum capacitance is determined principally by the setting of C3, which usually has low capacitance, and the maximum capacitance by the setting of C2, which is in the order of 25 to 50 pF. This method is capable of close adjustment to practically any desired degree of bandspread. Either C2 or C3 must be adjusted for each band or separate preadjusted capacitors must be switched in.

The circuit at Fig. 8-10C also gives complete spread on each band. C1, the bandspread capacitor, may have any convenient value; 50 pF is satisfactory. C2 may be used for continuous frequency coverage ("general coverage") and as a bandsetting capacitor. The effective maximum-minimum capacitance ratio depends on C2 and the point at which C1 is tapped on the coil. The nearer the tap to the bottom of the coil, the greater the bandspread, and vice versa. For a given coil and tap, the bandspread will be greater if C2 is set at higher capacitance. C2 may be connected permanently across the individual inductor and preset, if desired. This requires a separate capacitor for each band, but eliminates the necessity for resetting C2 each time.

Ganged Tuning

The tuning capacitors of the several rf circuits may be coupled together mechanically and operated by a single control. However, this operating convenience involves more complicated construction, both electrically and mechanically. It becomes necessary to make the various circuits track – that is, tune to the same frequency for a given setting of the tuning control.

True tracking can be obtained only when the inductance, tuning capacitors, and circuit inductances and minimum and maximum capacitances are identical in all "ganged" stages. A small trimmer or padding capacitor may be connected across the coil, so that various minimum capacitances can be compensated. The use of the trimmer necessarily increases the minimum circuit capacitance but is a necessity for satisfactory tracking. Midget capacitors having maximum capacitances of 15 to 30 pF are commonly used.

The same methods are applied to bandspread circuits that must be tracked. The inductance can be trimmed by using a coil form with an adjustable brass (or copper) core. This core material will reduce the inductance of the coil, raising the resonant frequency of the circuit. Powdered-iron or ferrite core material can also be used, but will lower the resonant frequency of the tuned circuit because it increases the inductance of the coil. Ferrite and powdered-iron cores will raise the Q of the coil provided the core material is suitable for the frequency being used. Core material is now available for frequencies well into the vhf region.

The Superheterodyne

In a superheterodyne receiver, the frequency of the incoming signal is heterodyned to a new radio frequency, the intermediate frequency (abbreviated "i-f"), then amplified, and finally detected. The frequency is changed by modulating the output of a tunable oscillator (the high-frequency, or local oscillator) by the incoming signal in a mixer or converter stage to produce a side frequency equal to the intermediate frequency. The other side frequency is rejected by selective circuits. The audio-frequency signal is obtained at the detector. Code signals are made audible by heterodyne reception at the detector stage; this oscillator is called the "beat-frequency oscillator" or BFO. Block diagrams of typical single- and double-conversion receivers are shown in Fig. 8-12.

As a numerical example, assume that an intermediate frequency of 455 kHz is chosen and that the incoming signal is at 7000 kHz. Then the high-frequency oscillator frequency may be set to 7455 kHz in order that one side frequency (7455 minus 7000) will be at 455 kHz. The high-frequency oscillator could also be set to 6545 kHz and give the same difference frequency. To produce an audible code signal at the detector of, say, 1000 Hz, the heterodyning oscillator would be set to either 454 or 456 kHz.

The frequency-conversion process permits rf amplification at a relatively low frequency, the i-f. High selectivity and gain can be obtained at this frequency, and this selectivity and gain are constant. The separate oscillators can be designed for good stability and, since they are working at frequencies considerably removed from the signal frequencies, they are not normally "pulled" by the incoming signal.

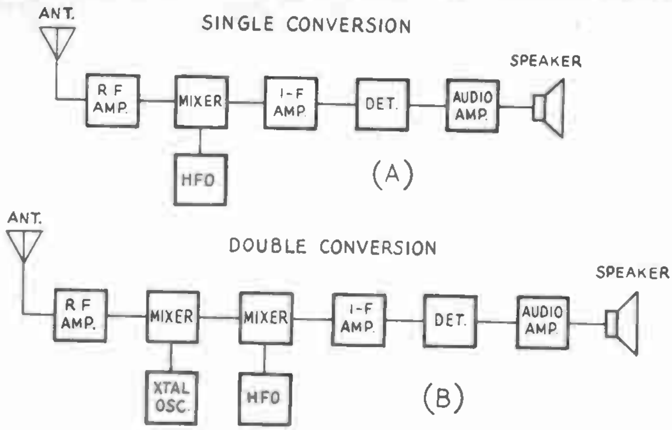


Fig. 8-12 — Block diagrams of a (A) single- and (B) double-conversion superheterodyne receiver.

Images

Each hf oscillator frequency will cause i-f response at two signal frequencies, one higher and one lower than the oscillator frequency. If the oscillator is set to 7455 kHz to tune to a 7000-kHz signal, for example, the receiver can respond also to a signal on 7910 kHz, which likewise gives a 455-kHz beat. The undesired signal is called the image. It can cause unnecessary interference if it isn't eliminated.

The radio-frequency circuits of the receiver (those used before the signal is heterodyned to the i-f) normally are tuned to the desired signal, so that the selectivity of the circuits reduces or eliminates the response to the image signal. The ratio of the receiver voltage output from the desired signal to that from the image is called the signal-to-image ratio, or image ratio.

The image ratio depends upon the selectivity of the rf tuned circuits preceding the mixer tube. Also, the higher the intermediate frequency, the higher the image ratio, since raising the i-f increases the frequency separation between the signal and the image and places the latter further away from the resonance peak of the signal-frequency input circuits.

The Double-Conversion Superheterodyne

At high and very-high frequencies it is difficult to secure an adequate image ratio when the intermediate frequency is of the order of 455 kHz. To reduce image response the signal frequently is converted first to a rather high (1500, 5000, or even 10,000 kHz) intermediate frequency, and then – sometimes after further amplification – converted to a lower i-f where higher adjacent-channel selectivity can be obtained. Such a receiver is called a double-conversion superheterodyne (Fig. 8-12B).

Other Spurious Responses

In addition to images, other signals to which the receiver is not tuned may be heard. Harmonics of the high-frequency oscillator may beat with signals far removed from the desired frequency to produce output at the intermediate frequency; such spurious responses can be reduced by adequate selectivity *before* the mixer stage, and by using sufficient shielding to prevent signal pickup by any means other than the antenna. When a strong signal is received, the harmonics generated by rectification in the detector may, by stray coupling, be introduced into the rf or mixer circuit and converted to the intermediate frequency, to go through the receiver in the same way as an ordinary signal. These "birdies" appear as a heterodyne beat on the desired signal, and are principally bothersome when the frequency of the incoming signal is not greatly different from the intermediate frequency. The cure is proper circuit isolation and shielding.

Harmonics of the beat oscillator also may be converted in similar fashion and amplified through the receiver; these responses can be reduced by shielding the beat oscillator and by careful mechanical design.

MIXER PRODUCTS

Additional spurious products are generated during the mixing process, and these products are the most troublesome of all, as it is difficult indeed to eliminate them unless the frequencies chosen for the mixing scheme are changed. The tables and chart given in Fig. 8-13 will aid in the choice of spurious-free frequency combinations, and they can be used to determine how receiver "birdies" are being generated. Only mixer products that fall close to the desired frequency are considered, as they are the ones that normally cause trouble. The horizontal axis of the chart is marked off in steps from 3 to 20, and the vertical axes from 0 to 14. These numbers can be taken to mean either kilohertz or megahertz, depending on the frequency range used. Both axes must use the same reference; one cannot be in kHz and the other in MHz.

Spurious Response Chart

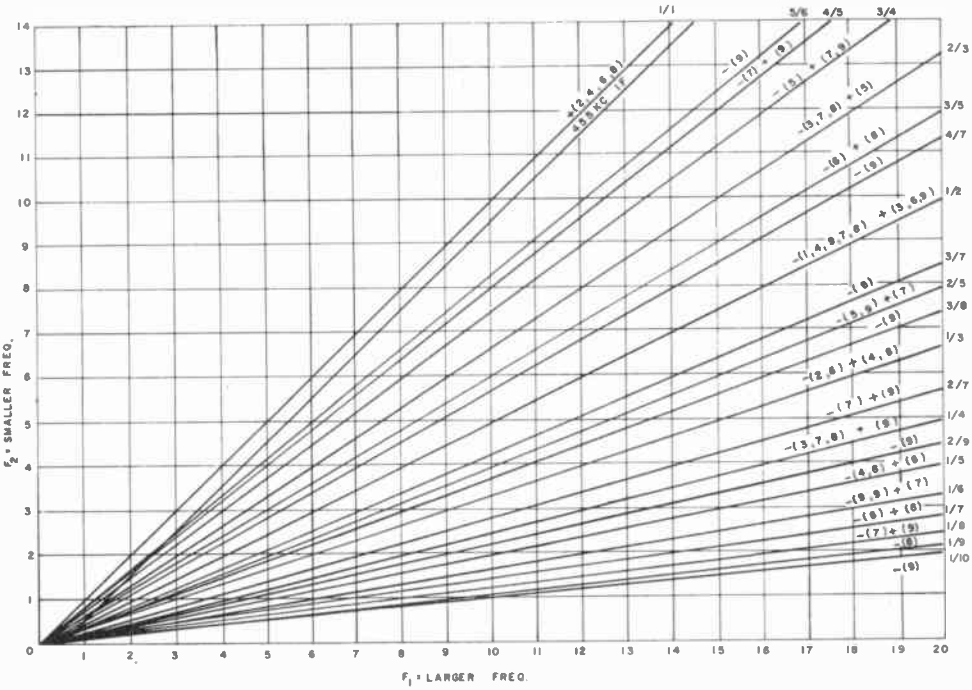


TABLE 1

$F_2 \sim F_1$

ORDER	1	2	3	4	5	6	7	8	9
1/1		•20 •02		•13 •31		•24 •42		•35 •53	
1/2	10		•12 •30	31	32	•33 •51	52	53	•54 •72
1/3		20		•22 •40		42 51		•53 •71	
1/4			30		•32 •50		52	71	
1/5				40		•42 •60		62	
1/6					50		•52 •70		72
1/7						60		•62 •80	
1/8							70		•72 •90
1/9								80	
1/10									90
2/3			21		•23 •41		43	53	

TABLE 2

$F_2 \sim F_1$

ORDER	1	2	3	4	5	6	7	8	9
2/5				41			•43 •61		63
2/7							61		•63 •81
2/9									81
3/4					32		•34 •52		54
3/5						42		•44 •62	
3/7								62	
3/8									72
4/5							43		•45 •63
4/7									63
5/6									54

• INDICATES SUM MIXING
OTHER — DIFF MIXING

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Fig. 8-13 — Chart to aid in the calculation of spurious frequencies generated during the mixing process.

To demonstrate the use of the chart, suppose an amateur wanted to mix a 6- to 6.5-MHz VFO output with a 10-MHz ssb signal to obtain output in the 80-meter band (the same problem as with a receiver that tunes 3.5 to 4 MHz, using a 6- to 6.5-MHz VFO to heterodyne to a 10-MHz i-f). Thus, F1 is 10 MHz and F2 is 6 to 6.5 MHz. Examination of the chart shows the intersection of these frequencies to be near the lines marked 2/3 and 3/5. In the case of the transmitter, difference (subtractive) mixing is to be used. The order of the products that will be close to the desired mixer output frequency is given on each line in parentheses. A plus sign in front of the parentheses indicates the product order in a sum (additive) mix, and a minus sign the order of a difference mix. For this example, the chart indicates the 3rd-, 7th-, and 8th-order products in a 2/3 relationship are going to be near the 80-meter band, plus the 6th-order product of the 3/5 relationship.

The exact frequencies of these products can be found with the help of the two small tables in Fig. 8-13. The product orders from 1 to 9 are given for all the product lines on the chart. The first digit of each group in a box is the harmonic of the lower frequency, F2, and the second digit is the harmonic of the larger frequency, F1. The dot indicates sum mixing and no dot indicates products in a difference mix. In the example, the chart shows that the 2/3 relationship will yield a 3rd-order product 2F2-F1, a 7th-order product 4F2-3F1, and an 8th-order product 5F2-3F1.

(Continued on next page)

(2 X 6) — 10 = 2
(2 X 6.5) — 10 = 3 (3rd order)

(4 X 6) — (3 X 10) =-6
(4 X 6.5) — (3 X 10) =-4 (7th order)

(5 X 6) — (3 X 10) = 0
(5 X 6.5) — (3 X 10) = 2.5 (8th order)

The 3/4 relationship produces a 6th-order product 4F2-2F1.

(4 X 6) — (2 X 10) = 4
(4 X 6.5) — (2 X 10) = 6

Thus, the ranges of spurious signals near the desired output band are 2 to 3 MHz, 6 to 4 MHz, 0 to 2.5 MHz, and 4 to 6 MHz. The negative sign indicates that the 7th-order product moves in the opposite direction to the normal output frequency, as the VFO is tuned. In this example proper mixer operation and sufficient selectivity following the mixer should keep the unwanted products sufficiently down in level without the use of filters or traps. Even-order products can be reduced by employing a balanced or doubly balanced mixer circuit, such as shown in Fig. 8-16.

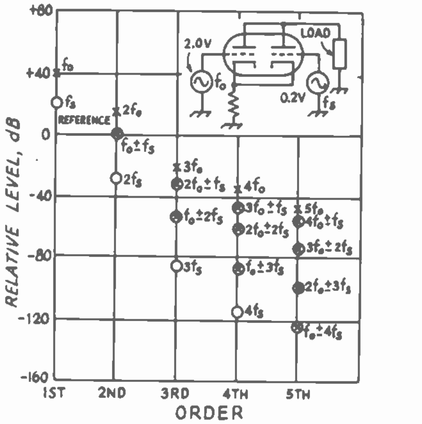


Fig. 8-14 — Chart showing the relative levels of spurious signals generated by a 12AU7A mixer.

The level of spurious products to be found in the output of a 12AU7 have been calculated by V. W. Bolie, using the assumption that the oscillator injection voltage will be 10 times (20 dB) greater than the input signal. This information is given in Fig. 8-14 for 1st- to 5th-order products. It is evident from the chart that multiples of the oscillator voltage produce the strongest of the undesired products. Thus, it follows that using a balanced-mixer design which reduces the level of oscillator signal in the output circuit will decrease the strength of the unwanted products.

MIXERS

A circuit tuned to the output frequency is placed in the plate circuit of the mixer, to offer a high impedance load for the output current that is developed. The signal- and oscillator-frequency voltages appearing in the plate circuit are rejected by the selectivity of this circuit. The output tuned circuit should have low impedance for these frequencies, a condition easily met if neither is close to the output frequency.

The conversion efficiency of the mixer is the ratio of output voltage from the plate circuit to rf signal voltage applied to the grid. High conversion efficiency is desirable. The device used as a mixer also should be low noise if a good signal-to-noise ratio is wanted, particularly if the mixer is the first active device in the receiver.

A change in oscillator frequency caused by tuning of the mixer grid circuit is called pulling. Pulling should be minimized, because the stability of the whole receiver or transmitter depends critically upon the stability of the hf oscillator. Pulling decreases with separation of the signal and hf-oscillator frequencies, being less with higher output frequencies. Another type of pulling is caused by lack of regulation in the power supply. Strong signals cause the voltage to change, which in turn shifts the oscillator frequency.

Circuits

If the mixer and high-frequency oscillator are separate tubes or transistors, the converter portion is called a "mixer." If the two are combined in one tube envelope (as is often done for reasons of economy or efficiency), the stage is called a "converter." In either case the function is the same.

Typical mixer circuits are shown in Figs. 8-15 and 8-16. The variations are chiefly in the way in which the oscillator voltage is introduced. In 8-15A, a pentode functions as a plate detector at the output frequency; the oscillator voltage is capacitance-coupled to the grid of the tube through C2. Inductive coupling may be used instead. The conversion gain and input selectivity generally are good, so long as the sum of the two voltages (signal and oscillator) impressed on the mixer grid does not exceed the grid bias. It is desirable to make the oscillator voltage as high as possible without exceeding this limitation. The oscillator power required is negligible. The circuit is a sensitive one and makes a good mixer, particularly with high-transconductance tubes like the 6CY5, 6EJ7 or 6U8A (pentode section). Triode tubes can be used as mixers in grid-injection circuits, but they are commonly used at 50 MHz and higher, where mixer noise may become a significant factor. The triode mixer has the lowest inherent noise, the pentode is next, and the multigrid converter tubes are the noisiest.

In the circuit of Fig. 8-15A the oscillator voltage could be introduced at the cathode rather than at the control grid. If this were done, C3 would have to be removed, and output from the oscillator would be coupled to the cathode of the mixer through a .001-μF capacitor. C2 would also be discarded. Generally, the same rules apply as when the tube uses grid injection.

It is difficult to avoid "pulling" in a triode or pentode mixer, and a pentagrid mixer tube provides much better isolation. A typical circuit is shown in Fig. 8-15B, and tubes like the 6BA7 or 6BE6 are commonly used. The oscillator voltage is introduced through an "injection" grid. Measurement of the rectified current flowing in R2 is used as a check for proper oscillator-voltage amplitude. Tuning of the signal-grid circuit can have little effect on the oscillator frequency because the injection grid is isolated from the signal grid by a screen grid that is at rf ground potential. The pentagrid mixer is much noisier than a triode or pentode mixer, but its isolating characteristics make it a very useful device.

Penagrid tubes like the 6BE6 or 6BA7 are somtimes used as "converters" performing the dual function of mixer and oscillator. The usual circuit resembles Fig. 8-15B except that the No. 1 grid connects to the top of a grounded parallel-tuned circuit by means of a larger grid-blocking capacitor, and the cathode (without R1 and C3) connects to a tap near the grounded end of the coil. This forms a Hartley oscillator circuit. Correct location of the cathode tap is indicated by the grid current; raising the tap increases the grid current because the strength of oscillation is increased.

The effectiveness of converter tubes of the type just described becomes less as the signal frequency is increased. Some oscillator voltage will be coupled to the signal grid through "space-charge" coupling, an effect that increases with frequency. If there is relatively little frequency difference between oscillator and signal, as for example a 14- or 28-MHz signal and an i-f of 455 kHz, this voltage can become considerable because the selectivity of the signal circuit will be unable to reject it. If the signal grid is not returned directly to ground, but instead is returned through a resistor or part of an agc system, considerable bias can be developed which will cut down the gain. For this reason, and to reduce image response, the i-f following the first converter of a receiver should be not less than 5 or 10 percent of the signal frequency.

Diodes, FETs, ICs, and bipolar transistors can be used as mixers. Examples are given in Figs. 8-15 and 8-16. A single-diode mixer is not shown here since its application is usually limited to circuits operating in the uhf region and higher. A discussion of diode mixers, plus a typical circuit, is given in Chapter 9.

Oscillator injection can be fed to the base or emitter elements of bipolar-transistor mixers, Fig. 8-15C. If emitter injection is used, the usual emitter bypass capacitor must be removed. Because the dynamic characteristics of bipolar transistors prevent them from handling high signal levels, FETs are usually preferred in mixer circuits, although they do not provide the high conversion gain available with bipolar mixers. FETs (Fig. 8-15D and E) have greater immunity to cross-modulation and overload than bipolar transistors, and offer nearly square-law performance. The circuit at D uses a junction FET, N-channel type, with oscillator injection being supplied to the source. The value of the source resistor should be adjusted to provide a bias of approximately 0.8 volts. This value offers a good compromise between conversion gain and good intermodulation-distortion characteristics. At this bias level a local-oscillator injection of approximately 1.5 volts is desirable for good conversion gain. The lower the oscillator-injection level, the lower the gain. High injection levels improve the mixers immunity to cross-modulation.

A dual-gate MOSFET is used as a mixer at E. Gate 2 is used for injecting the local-oscillator signal while gate 1 is supplied with signal voltage.

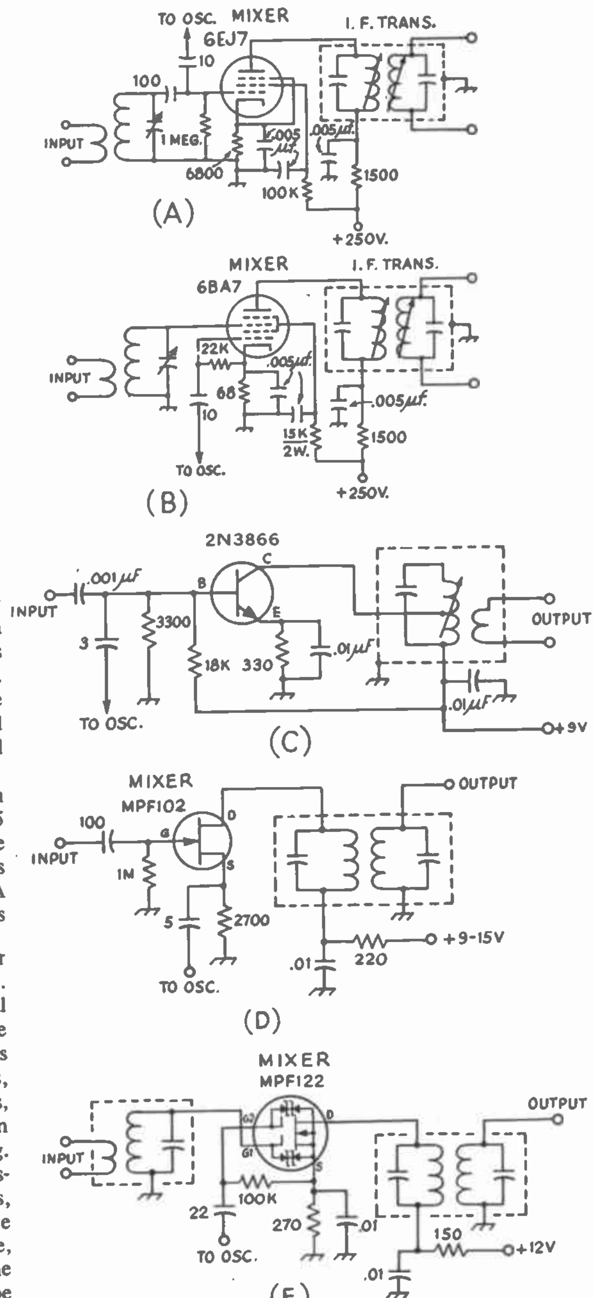


Fig. 8-15 — Typical single-ended mixer circuits.

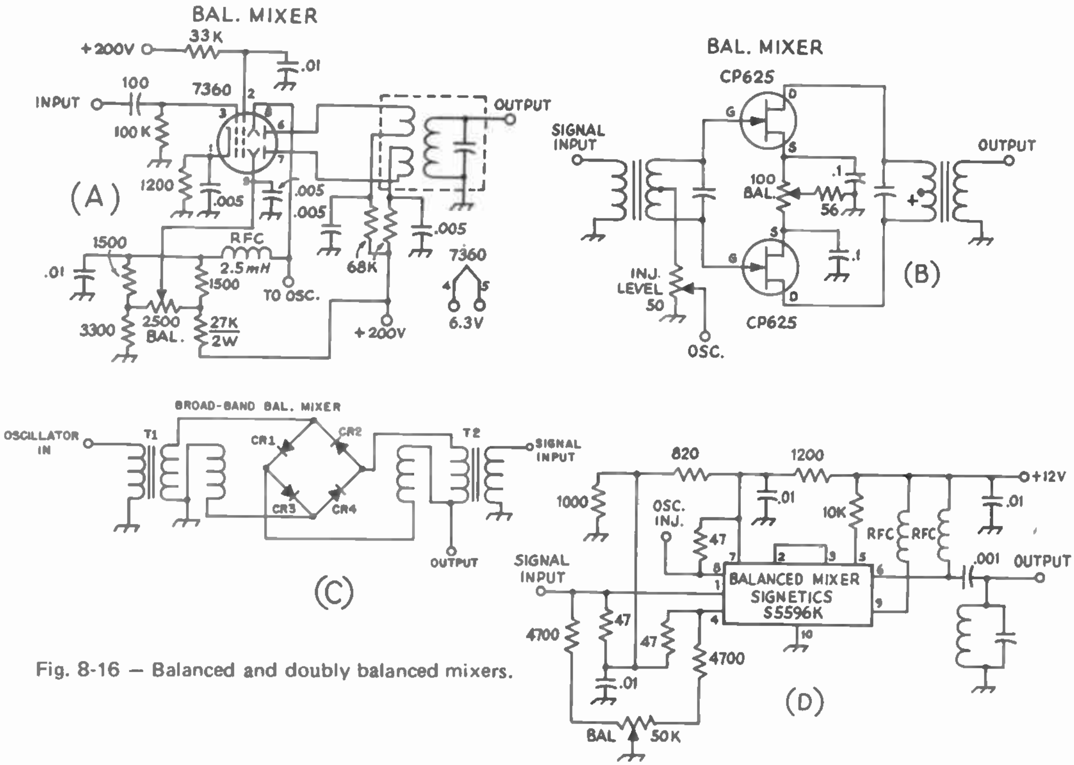


Fig. 8-16 — Balanced and doubly balanced mixers.

This type of mixer has excellent immunity to cross-modulation and overload. It offers better isolation between the oscillator and input stages than is possible with a JFET mixer. The mixers at D and E have high-Z input terminals, while the circuit at C has a relatively low-Z input impedance. The latter requires tapping the base down on the input tuned circuit for a suitable impedance match.

BALANCED MIXERS

The level of input and spurious signals contained in the output of a mixer may be decreased by using a balanced or doubly balanced circuit. The balanced mixer reduces leakthrough and even-order harmonics of one input (usually the local oscillator) while the doubly balanced designs lower the level of spurious signals caused by both the signal and oscillator inputs. One type of balanced mixer uses a 7360 beam-deflection tube, connected as shown in Fig. 8-16A. The signal is introduced at the No. 1 grid, to modulate the electron stream running from cathode to plates. The beam is deflected from one plate to the other and back again by the BFO voltage applied to one of the deflection plates. (If oscillator radiation is a problem, push-pull deflection by both deflection plates should be used.) At B, two CP625 FETs are used; these devices have a large dynamic range, about 130 dB, making them an excellent choice for either a transmitting or receiving mixer. Dc balance is set with a control in the source leads. The oscillator energy is introduced at the center tap of the input transformer.

In the circuit of Fig. 8-16C, hot-carrier diodes are employed as a broad-band balanced mixer. With careful winding of the toroid-core input and output transformers, the inherent balance of the mixer will provide 40- to 50-dB attenuation of the oscillator signal. The transformers, T1 and T2, having trifilar windings – using No. 32 enamel wire, 12 turns on a 1/2-inch core will provide operation on any frequency between 500 kHz and 100 MHz. Using Q3 cores the upper-frequency range can be extended to 300 MHz. CR1 to CR4, inc, comprise a matched quad of Hewlett-Packard HPA 5082-2805 diodes. Conversion loss in the mixer will be 6 to 8 dB.

Special doubly balanced mixer ICs are now available which can simplify circuit construction, as special balanced transformers are not required. Also, the ICs produce high conversion gain. A typical circuit using the Signetics S5596K is shown in Fig. 8-16D. The upper frequency limit of this device is approximately 130 MHz.

THE HIGH-FREQUENCY OSCILLATOR

Stability of the receiver is dependent chiefly upon the stability of the tunable hf oscillator, and particular care should be given this part of the receiver. The frequency of oscillation should be insensitive to mechanical shock and changes in voltage and loading. Thermal effects (slow change in frequency because of tube, transistor, or circuit heating) should be minimized. See Chapter 6 for sample circuits and construction details.

THE INTERMEDIATE-FREQUENCY AMPLIFIER

One major advantage of the superhet is that high gain and selectivity can be obtained by using a good i-f amplifier. This can be a one-stage affair in simple receivers, or two or three stages in the more elaborate sets.

Choice of Frequency

The selection of an intermediate frequency is a compromise between conflicting factors. The lower the i-f, the higher the selectivity and gain, but a low i-f brings the image nearer the desired signal and hence decreases the image ratio. A low i-f also increases pulling of the oscillator frequency. On the other hand, a high i-f is beneficial to both image ratio and pulling, but the gain is lowered and selectivity is harder to obtain by simple means.

An i-f of the order of 455 kHz gives good selectivity and is satisfactory from the standpoint of image ratio and oscillator pulling at frequencies up to 7 MHz. The image ratio is poor at 14 MHz when the mixer is connected to the antenna, but adequate when there is a tuned rf amplifier between antenna and mixer. At 28 MHz and on the very high frequencies, the image ratio is very poor unless several rf stages are used. Above 14 MHz, pulling is likely to be bad without very loose coupling between mixer and oscillator. Tuned-circuit shielding also helps.

With an i-f of about 1600 kHz, satisfactory image ratios can be secured on 14, 21 and 28 MHz with one rf stage of good design. For frequencies of 28 MHz and higher, a common solution is to use double conversion, choosing one high i-f for image reduction (5 and 10 MHz are frequently used) and a lower one for gain and selectivity.

In choosing an i-f it is wise to avoid frequencies on which there is considerable activity by the various radio services, since such signals may be picked up directly by the i-f wiring. Shifting the i-f or better shielding are the solutions to this interference problem.

Fidelity; Sideband Cutting

Amplitude modulation of a carrier generates sideband frequencies numerically equal to the carrier frequency plus and minus the modulation frequencies present. If the receiver is to give a faithful reproduction of modulation that contains, for instance, audio frequencies up to 5000 Hz, it must at least be capable of amplifying equally all frequencies contained in a band extending from 5000 Hz above or below the carrier frequency. In a superheterodyne, where all carrier frequencies are changed to the fixed intermediate frequency, the i-f amplification must be uniform over a band 5-kHz wide, when the carrier is set at one edge. If the carrier is set in the center, at 10-kHz band is required. The signal-frequency circuits usually do not have enough overall selectivity to affect materially the "adjacent-channel" selectivity, so that only the i-f-amplifier selectivity need be considered.

If the selectivity is too great to permit uniform amplification over the band of frequencies occupied by the modulated signal, some of the sidebands are "cut." While sideband cutting reduces fidelity, it is frequently preferable to sacrifice naturalness of reproduction in favor of communications effectiveness.

The selectivity of an i-f-amplifier, and hence the tendency to cut sidebands increases with the number of tuned circuits and also is greater the lower the intermediate frequency. From the standpoint of communication, sideband cutting is never serious with two-stage amplifiers at frequencies as low as 455 kHz. A two-stage i-f-amplifier at 85 or 100 kHz will be sharp enough to cut some of the higher frequency sidebands, if good transformers are used. However, the cutting is not at all serious, and the gain in selectivity is worthwhile in crowded amateur bands as an aid to QRM reduction.

Circuits

I-f amplifiers usually consist of one or more stages. The more stages employed, the greater the selectivity and overall gain of the system. In double-conversion receivers there is usually one stage at the first i-f, and sometimes as many as three or four stages at the second, or last, i-f. Most single-conversion receivers use no more than three stages of i-f amplification.

A typical vacuum-tube i-f stage is shown in Fig. 8-17 at A. The second or third stages would simply be duplicates of the stage shown. Remote cutoff pentodes are almost always used for i-f amplifiers, and such tubes are operated as Class-A amplifiers. For maximum selectivity, double-tuned transformers are used for interstage coupling, though single-tuned inductors and capacitive coupling can be used, but at a marked reduction in selectivity.

Agc voltage can be used to reduce the gain of the stage, or stages, by applying it to the terminal marked AGC. The agc voltage should be negative. Manual control of the gain can be effected by lifting the 100-ohm cathode resistor from ground and inserting a potentiometer between it and ground. A 10,000-ohm control can be used for this purpose. A small amount of B-plus voltage can be fed through a dropping resistor (about 56,000 ohms from a 250-volt bus) to the junction of the gain control and the 100-ohm cathode resistor to provide an increase in tube bias in turn reducing the mutual conduction of the tube for gain reduction.

An integrated-circuit i-f amplifier is shown at B. A positive-polarity agc voltage is required for this circuit to control the stage gain. If manual gain control provisions are desired, a potentiometer can be used to vary the plus voltage to the agc terminal of the IC. The control would be connected between the 9-volt bus and ground, its movable contact wired to the agc terminal of the IC.

A dual-gate MOSFET i-f amplifier is shown at B. Application of negative voltage to gate 2 of the

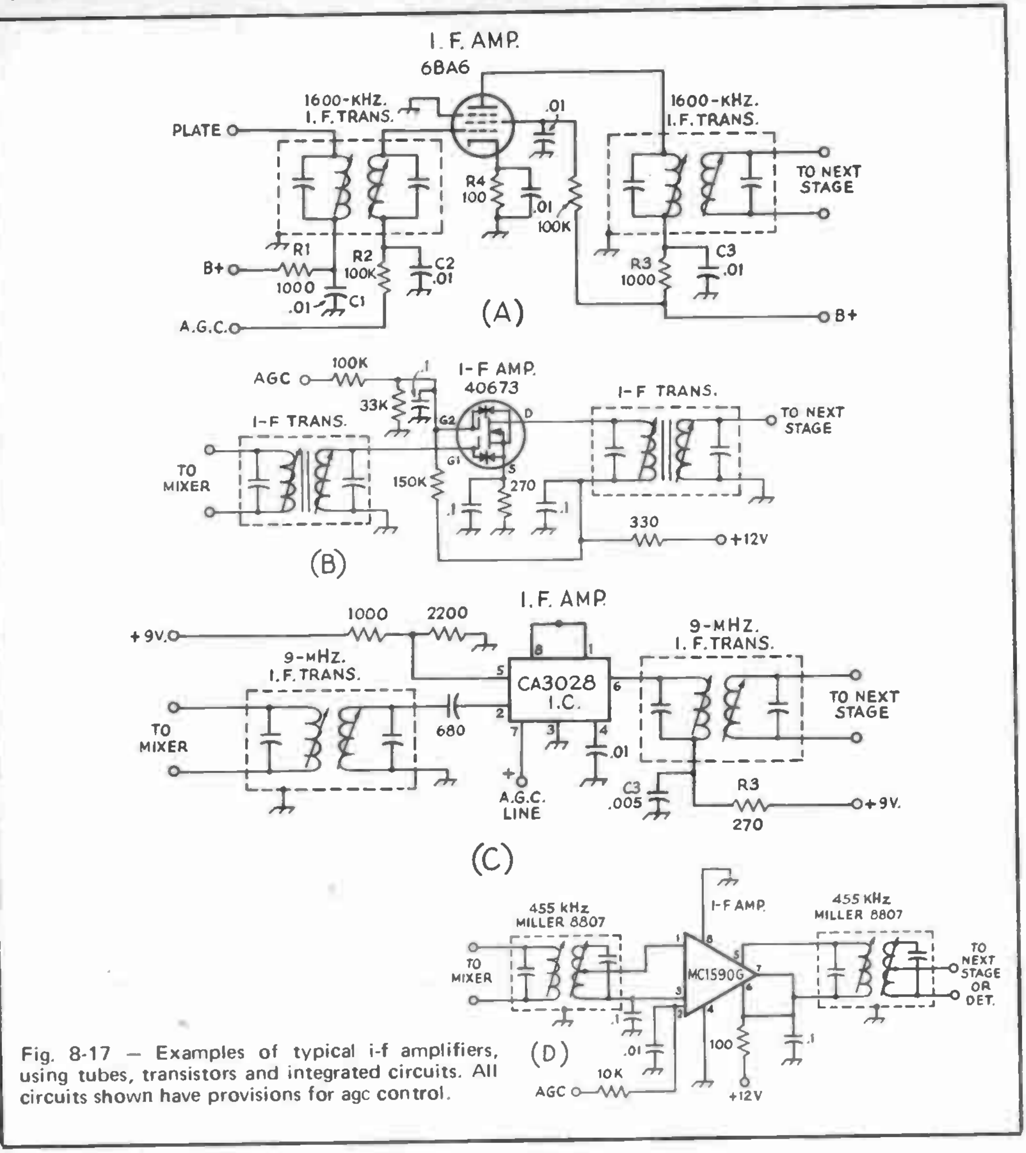


Fig. 8-17 — Examples of typical i-f amplifiers, using tubes, transistors and integrated circuits. All circuits shown have provisions for agc control.

device reduces the gain of the stage. To realize maximum gain when no agc voltage is present, it is necessary to apply approximately 3 volts of positive dc to gate 2. Neutralization is usually not required with a MOSFET in i-f amplifiers operating up to 20 MHz. Should instability occur, however, gate 1 and the drain may be tapped down on the i-f transformer windings.

High-gain linear ICs have been developed specifically for use as receiver i-f amplifiers. A typical circuit which uses the Motorola MC1590G is shown at D; 70 dB of gain may be achieved using this device. Agc characteristics of the IC are excellent. A 4-volt change at the agc terminal produces 60-dB change in the gain of the stage. Agc action starts at 5 volts, so a positive agc system with a fixed dc level must be employed.

Tubes for I-f Amplifiers

Variable-μ (remote cutoff) pentodes are almost invariably used in i-f amplifier stages, since grid-bias gain control is practically always applied to the i-f amplifier. Tubes with high plate resistance will have least effect on the selectivity of the amplifier, and those with high mutual conductance will give greatest gain. The choice of i-f tubes normally has no effect on the signal-to-noise ratio, since this is determined by the preceding mixer and rf amplifier.

The 6BA6, 6BJ6 and 6BZ6 are recommended for i-f work because they have desirable remote cutoff characteristics.

When two or more stages are used the high gain may tend to cause troublesome instability and oscillation, so that good shielding, bypassing, and careful circuit arrangement to prevent stray coupling between input and output circuits are necessary.

When vacuum tubes are used, the plate and grid leads should be well separated. When transistors are used, the base and collector circuits should be well isolated. With tubes it is advisable to mount the screen-bypass capacitor directly on the bottom of the socket, crosswise between the plate and grid pins, to provide additional shielding. As a further precaution against capacitive coupling, the grid and plate leads should be "dressed" close to the chassis.

I-f Transformers

The tuned circuits of i-f amplifiers are built up as transformer units consisting of a metal shield container in which the coils and tuning capacitors are mounted. Both air-core and powered-iron-core universal-wound coils are used, the latter having somewhat higher Qs and hence greater selectivity and gain. In universal windings the coil is wound in layers with each turn traversing the length of the coil, back and forth, rather than being wound perpendicular to the axis as in ordinary single-layer coils. In a straight multilayer winding, a fairly large capacitance can exist between layers. Universal winding, with its "criss-crossed" turns, tends to reduce distributed-capacitance effects.

For tuning, air-dielectric tuning capacitors are preferable to mica compression types because their capacitance is practically unaffected by changes in temperature and humidity. Iron-core transformers may be tuned by varying the inductance (permeability tuning), in which case stability comparable to that of variable air-capacitor tuning can be obtained by use of high-stability fixed mica or ceramic capacitors. Such stability is of great importance, since a circuit whose frequency "drifts" with time eventually will be tuned to a different frequency than the other circuits, thereby reducing the gain and selectivity of the amplifier.

The normal interstage i-f transformer is loosely coupled, to give good selectivity consistent with adequate gain. A so-called diode transformer is similar, but the coupling is tighter, to give sufficient transfer when working into the finite load presented by a diode detector. Using a diode transformer in place of an interstage transformer would result in loss of selectivity; using an interstage transformer to couple to the diode would result in loss of gain.

Besides the conventional i-f transformers just mentioned, special units to give desired selectivity characteristics have been used. For higher-than-ordinary adjacent-channel selectivity, triple-tuned transformers, with a third tuned circuit inserted between the input and output windings, have been made. The energy is transferred from the input to the output windings via this tertiary winding, thus adding its selectivity to the over-all selectivity of the transformer.

Selectivity

The overall selectivity of the i-f amplifier will depend on the frequency and the number of stages. The following figures are indicative of the bandwidths to be expected with good-quality circuits in amplifiers so constructed as to keep regeneration at a minimum:

Tuned Ckts.	*Freq.*	*Circuit Q*	*Bandwidth, kHz −6 dB*	*−20 dB*	*−60 dB*
4	50 kHz	60	0.5	0.95	2.16
4	455 kHz	75	3.6	6.9	16
6	1600 kHz	90	8.2	15	34

THE BEAT OSCILLATOR AND DETECTOR

The detector in a superheterodyne receiver functions the same way as do the simple detectors described earlier in this chapter (Fig. 8-4), but usually operates at a higher input level because of the amplification ahead of it. The detectors of Fig. 8-4 are satisfactory for the reception of a-m signals. When copying cw and ssb signals it becomes necessary to supply a beat-oscillator (BFO) signal to the detector stage as described in the earlier section on product detectors. Suitable circuits for variable-frequency and crystal-controlled BFOs are given in Chapter 6.

AUTOMATIC GAIN CONTROL

Automatic regulation of the gain of the receiver in inverse proportion to the signal strength is an operating convenience in phone reception, since it tends to keep the output level of the receiver constant regardless of input-signal strength. The average rectified dc voltage, developed by the received signal across a resistance in a detector circuit, is used to vary the bias on the rf and i-f amplifier stages. Since this voltage is proportional to the average amplitude of the signal, the gain is reduced as the signal strength becomes greater. The control will be more complete and the output more constant as the number of stages to which the agc bias is applied is increased. Control of at least two stages is advisable.

Carrier-Derived Circuits

A basic diode-detector/agc-rectifier circuit is given at Fig. 8-18A. Here a single germanium diode serves both as a detector and an agc rectifier, producing a negative-polarity agc voltage. Audio is taken from the return end of the i-f transformer secondary and is filtered by means of a 47,000-ohm resistor and two 470-pF capacitors.

At B, CR1 (also a germanium diode) functions as a detector while CR2 (germanium) operates as an agc rectifier. CR2 furnishes a negative agc voltage to the controlled stages of the receiver. Though solid-stage rectifiers are shown at A and B, vacuum-tube diodes can be used in these circuits. A 6AL5 tube is commonly used in circuits calling for two diodes (B), but a 1-megohm resistor should be shunted across the right-hand diode if a tube is used.

The circuit at C shows a typical hookup for agc feed to the controlled stages. S1 can be used to disable the agc when this is desired. For tube and FET circuits the value of R1 and R2 can be 100,000 ohms, and R3 can be 470,000 ohms. If bipolar transistors are used for the rf and i-f stages being controlled, R1 and R2 will usually be

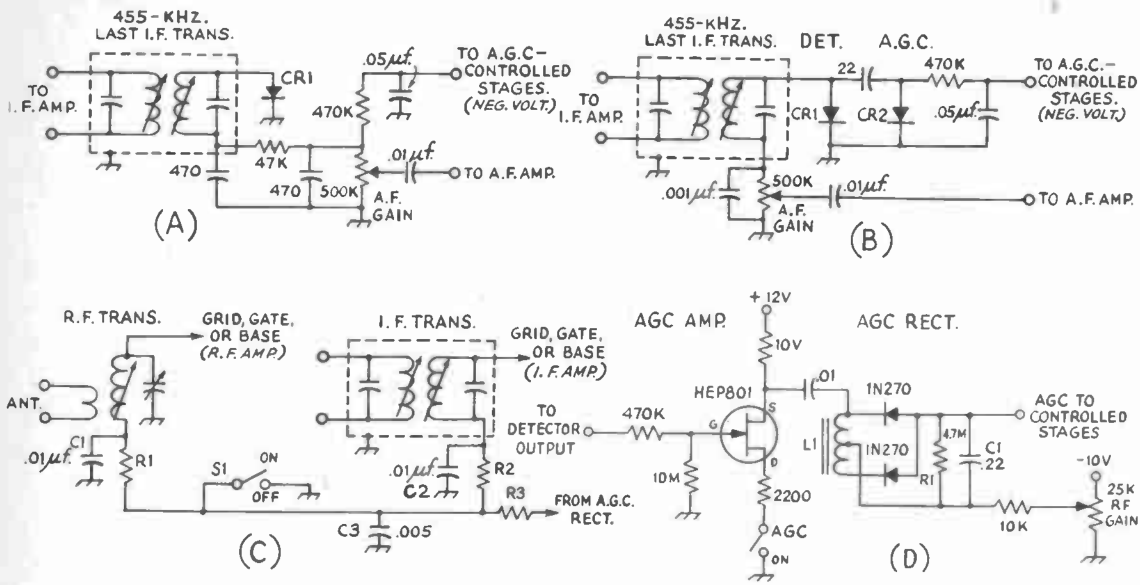


Fig. 8-18 – Methods for obtaining rectified voltage. At A the detector furnishes agc voltage. B shows separate diodes being used for the detector and agc circuits. C illustrates how negative agc voltage is fed to the rf and i-f stages of a typical receiver. D shows an audio-derived agc scheme. S1 is used to disable the agc when desired. R1, R2 and R3 in combination with C1, C2, and C3, are used for rf decoupling. Their values are dependent upon the device being used – tube or transistor. CR1 and CR2 at A and B are germanium diodes.

between 1000 and 10,000 ohms, depending upon the bias network required for the transistors used. R3 will also be determined by the bias value required in the circuit.

Agc Time Constant

The time constant of the resistor-capacitor combinations in the agc circuit is an important part of the system. It must be long enough so that the modulation on the signal is completely filtered from the dc output, leaving only an average dc component which follows the relatively slow carrier variations with fading. Audio-frequency variations in the agc voltage applied to the amplifier grids would reduce the percentage of modulation on the incoming signal. But the time constant must not be too long or the agc will be unable to follow rapid fading. The capacitance and resistance values indicated in 8-18A will give a time constant that is satisfactory for average reception.

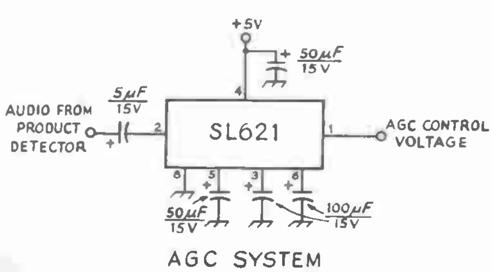


Fig. 8-19 – An IC agc system.

Cw and Ssb

Agc can be used for cw and ssb reception but the circuit is usually more complicated. The agc voltage must be derived from a rectifier that is isolated from the beat-frequency oscillator (otherwise the rectified BFO voltage will reduce the receiver gain even with no signal coming through). This is done by using a separate agc channel connected to an i-f amplifier stage ahead of the second detector (and BFO) or by rectifying the audio output of the detector. If the selectivity ahead of the agc rectifier isn't good, strong adjacent-channel signals may develop agc voltages that will reduce the receiver gain. When clear channels are available, however, cw and ssb agc will hold the receiver output constant over a wide range of signal inputs. Agc systems designed to work on these signals should have fast-attack and slow-decay characteristics to work satisfactorily, and often a selection of time constants is made available.

Audio-Derived Agc

Agc potential for use in a cw/ssb receiver may also be obtained by sampling the audio output of the detector and rectifying this signal. A typical circuit is shown in Fig. 8-18D. The JFET stage amplifies the audio signal; the output of the HEP801 is coupled to the secondary of an audio transformer, L1. The time constant of the agc line is established by R1C1. Manual gain control can be accomplished by adding a variable negative voltage to the common lead of the audio rectifier.

An improved audio-derived agc circuit is shown in Fig. 8-19, using the Plessey Microelectronics SL-621 integrated circuit. This design provides the fast-attack, slow-decay time constant required for ssb reception. High-level pulse signals that might "hang up" the agc system are sampled by the IC input circuit, activating a trigger which provides a fast-discharge path for the time-constant capacitor. Thus, noise bursts will not produce a change in the level of agc output voltage.

NOISE REDUCTION

Types of Noise

In addition to tube and circuit noise, much of the noise interference experienced in reception of high-frequency signals is caused by domestic or industrial electrical equipment and by automobile ignition systems. The interference is of two types in its effects. The first is the "hiss" type, consisting of overlapping pulses similar in nature to the receiver noise. It is largely reduced by high selectivity in the receiver, especially for code reception. The second is the "pistol-shot" or "machine-gun" type, consisting of separated impulses of high amplitude. The "hiss" type of interference usually is caused by commutator sparking in dc and series-wound ac motors, while the "shot" type results from separated spark discharges (ac power leaks, switch and key clicks, ignition sparks, and the like).

The only known approach to reducing tube and circuit noise is through the choice of low-noise front-end active components and through more overall selectivity.

Impulse Noise

Impulse noise, because of the short duration of the pulses compared with the time between them, must have high amplitude to contain much average energy. Hence, noise of this type strong enough to cause much interference generally has an instantaneous amplitude much higher than that of the signal being received. The general principle of devices intended to reduce such noise is to allow the desired signal to pass through the receiver unaffected, but to make the receiver inoperative for amplitudes greater than that of the signal. The greater the amplitude of the pulse compared with its time of duration, the more successful the noise reduction.

Another approach is to "silence" (render inoperative) the receiver during the short duration time of any individual pulse. The listener will not hear the "hole" because of its short duration, and very effective noise reduction is obtained. Such devices are called "blankers" rather than "limiters."

In passing through selective receiver circuits, the time duration of the impulses is increased, because of the Q of the circuits. Thus, the more selectivity ahead of the noise-reducing device, the more difficult it becomes to secure good pulse-type noise suppression. See Fig. 8-22.

Audio Limiting

A considerable degree of noise reduction in code reception can be accomplished by amplitude-limiting arrangements applied to the audio-output circuit of a receiver. Such limiters also maintain the signal output nearly constant during fading. These output-limiter systems are simple, and they are readily adaptable to most receivers without any modification of the receiver itself. However, they cannot prevent noise peaks from overloading previous stages.

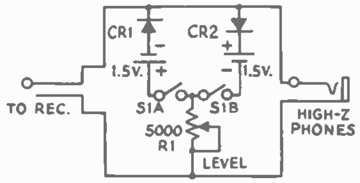


Fig. 8-20 — Circuit of a simple audio limiter/clipper. It can be plugged into the headphone jack of the receiver. R1 sets the bias on the diodes, CR1 and CR2, for the desired limiting level. S1 opens the battery leads when the circuit is not being used. The diodes can be 1N34As or similar.

NOISE-LIMITER CIRCUITS

Pulse-type noise can be eliminated to an extent which makes the reception of even the weakest of signals possible. The noise pulses can be clipped, or limited in amplitude, at either an rf or af point in the receiver circuit. Both methods are used by receiver manufacturers; both are effective.

A simple audio noise limiter is shown at Fig. 8-20. It can be plugged into the headphone jack of the receiver and a pair of headphones connected to the output of the limiter. CR1 and CR2 are wired to clip both the positive and negative peaks of the audio signal, thus removing the high spikes of pulse noise. The diodes are back-biased by 1.5-volt batteries to permit R1 to serve as a clipping-level control. This circuit also limits the amount of audio reaching the headphones. When tuning across the band, strong signals will not be ear-shattering and will appear to be the same strength as the weaker ones. S1 is open when the circuit is not in use to prevent battery drain. CR1 and CR2 can be germanium or silicon diodes, but 1N34As are generally used. This circuit is usable only with high-impedance headphones.

The usual practice in communications receivers is to use low-level limiting, Fig. 8-21. The limiting can be carried out at rf or af points in the receiver, as shown. Limiting at rf does not cause poor audio quality as is sometimes experienced when using series or shunt af limiters. The latter limits the normal af signal peaks as well as the noise pulses,

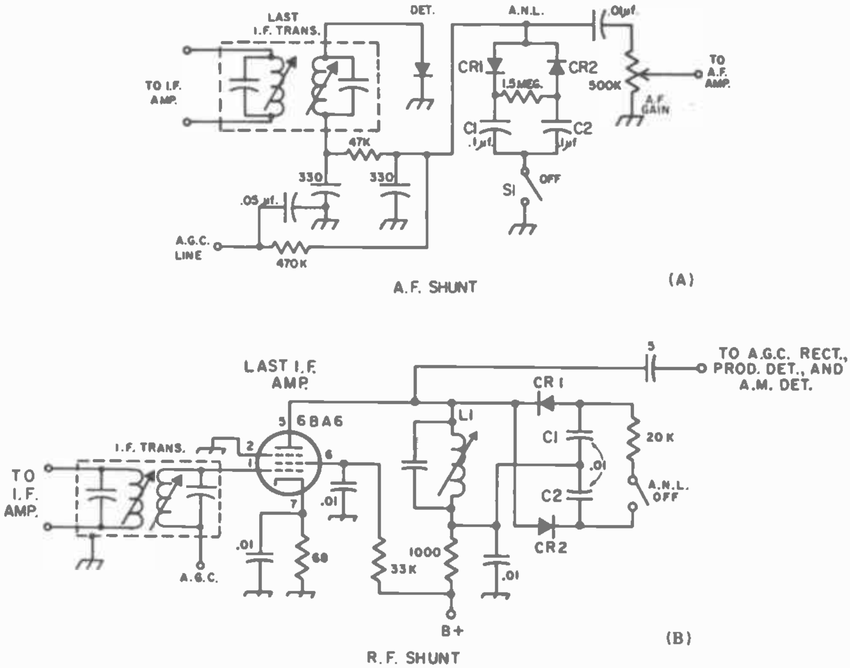
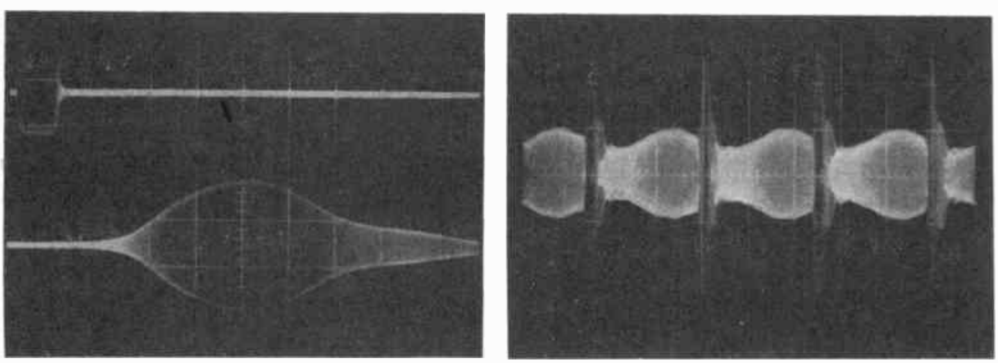


Fig. 8-21 — Typical rf and af anl circuits. A shows the circuit of a self-adjusting af noise limiter. CR1 and CR2 are self-biased silicon diodes which limit both the positive and negative audio and noise-pulse peaks. S1 turns the limiter on or off; B shows an rf limiter of the same type as A, but this circuit clips the positive and negative rf peaks and is connected to the last i-f stage. This circuit does not degrade the audio quality of the signal as does the circuit of A.

(A)  (B)

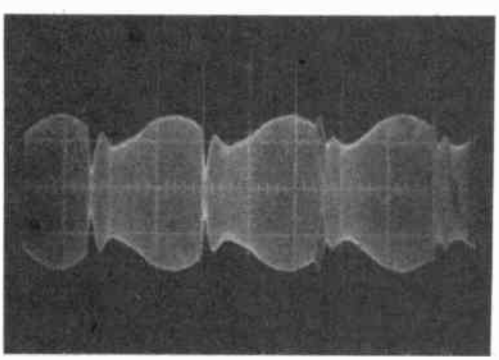
(C) 

Fig. 8-22 — The delay and lengthening of a noise pulse when passed through a 2-kHz wide amplifier with good skirt selectivity (4 kHz at −60 dB). (B) a 3.75-MHz carrier modulated 30 percent, interfered with by noise pulses. The noise pulses were originally 1000 times the amplitude of the signal; they have been reduced (and lengthened) by overload in the i-f. The i-f bandwidth is 5 kHz. Sweep speed = 1 millisecond/cm. (C) Same as B but with a noise blanker on.

giving an unpleasant audio quality to strong signals.

In a series-limiting circuit, a normally conducting element (or elements) is connected in the circuit in series and operated in such a manner that it becomes nonconductive above a given signal level. In a shunt limiting circuit, a nonconducting element is connected in shunt across the circuit and operated so that it becomes conductive above a given signal level, thus short-circuiting the signal and preventing its being transmitted to the remainder of the amplifier. The usual conducting element will be a forward-biased diode, and the usual nonconducting element will be a back-biased diode. In many applications the value of bias is set manually by the operator; usually the clipping level will be set at about 1 to 10 volts.

The af shunt limiter at A, and the rf shunt limiter at B operate in the same manner. A pair of self-biased diodes are connected across the af line at A, and across an rf inductor at B. When a steady cw signal is present the diodes barely conduct, but when a noise pulse rides in on the incoming signal, it is heavily clipped because capacitors C1 and C2 tend to hold the diode bias constant for the duration of the noise pulse. For this reason the diodes conduct heavily in the presence of noise and maintain a fairly constant signal output level. Considerable clipping of cw signal peaks occurs with this type of limiter, but no apparent deterioration of the signal quality results. L1 at C is tuned to the i-f of the receiver. An i-f transformer with a conventional secondary winding could be used in place of L1, the clipper circuit being connected to the secondary winding; the plate of the 6BA6 would connect to the primary winding in the usual fashion.

I-F NOISE SILENCER

The i-f noise silencer circuit shown in Fig. 8-23 is designed to be used ahead of the high-selectivity section of the receiver. Noise pulses are amplified and rectified, and the resulting negative-going dc pulses are used to cut off an amplifier stage during the pulse. A manual "threshold" control is set by the operator to a level that only permits rectification of the noise pulses that rise above the peak amplitude of the desired signal. The clamp transistor, Q3, short circuits the positive-going pulse "overshoots." Running the 40673 controlled i-f amplifier at zero gate 2 voltage allows the direct application of agc voltage. See July 1971 *QST* for additional details.

SIGNAL-STRENGTH AND TUNING INDICATORS

It is convenient to have some means by which to obtain *relative* readings of signal strength on a communications receiver. The actual meter readings in terms of S units, or decibels above S9, are of little consequence as far as a meaningful report to a distant station is concerned. Few signal-strength meters are accurate in terms of decibels, especially across their entire indicating range. Some manufacturers once established a standard in which a certain number of microvolts were equal to S9 on the meter face. Such calibration is difficult to maintain when a number of different receiver circuits are to be used. At best, a meter can be calibrated for one receiver – the one in which it will be used. Therefore, most S meters are good only as relative indicating instruments for comparing the strength of signals at a given time, on a given amateur band. They are also useful for "on-the-nose-tuning" adjustments with selective receivers. If available, a signal generator with an accurate output attenuator can be used to calibrate an S meter in terms of microvolts, but a different calibration chart will probably be required for each band because of probable differences in receiver sensitivity from band to band. It is helpful to establish a 50-μV reading at midscale on the meter so that the very strong signals will crowd the high end of the meter scale. The weaker signals will then be spread over the lower half of the scale and will not be compressed at the low end. Midscale on the meter can be called S9. If S units are desired across the scale, below S9, a marker can be established at every 6 dB point.

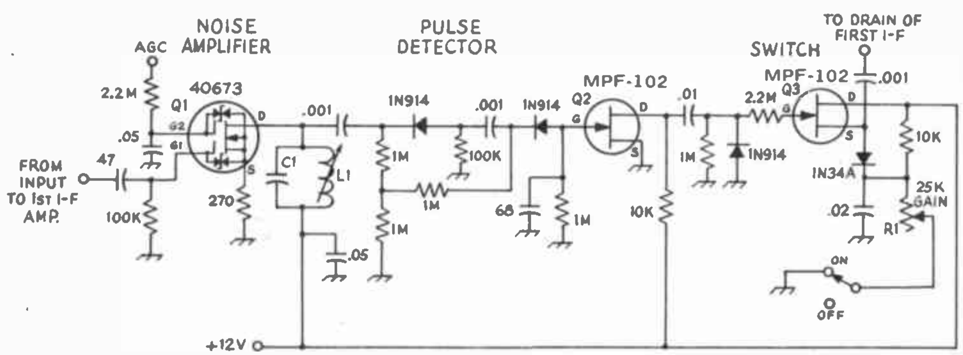
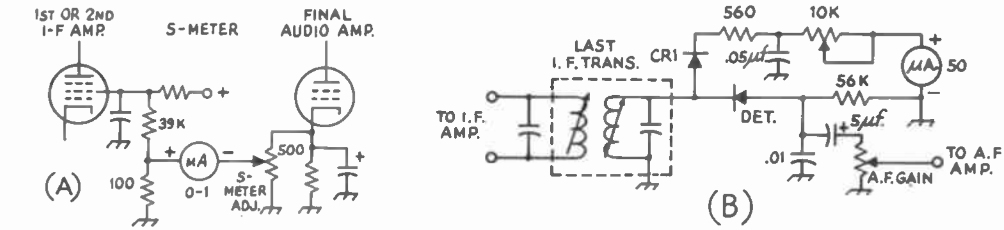


Fig. 8-23 – Diagram of the noise blanker. L1 and C1 are chosen to resonate at the desired i-f.



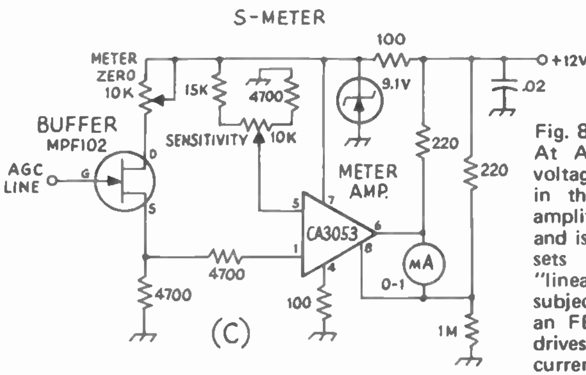


Fig. 8-24 – Practical examples of S-meter circuits. At A, a meter measures the change in screen voltage of an i-f amplifier stage, caused by changes in the agc voltage applied to the grid of the amplifier. At B, the i-f signal is rectified by CR1 and is fed to a 50-µA meter. A 10,000-ohm control sets the sensitivity and also functions as a "linearizing" resistor to make the meter less subject to the square-law response of CR1. At C, an FET samples the voltage on the agc line and drives an IC amplifier which provides the required current swing to operate a 1-mA meter.

S-METER CIRCUITS

A very simple meter indicator is shown at Fig. 8-24B. Rectified i-f is obtained by connecting CR1 to the take-off point for the detector. The dc is filtered by means of a 560-ohm resistor and a .05-µF capacitor. A 10,000-ohm control sets the meter at zero reading in the absence of a signal and also serves as a "linearizing" resistor to help compensate for the nonlinear output from CR1. The meter is a 50-µA unit, therefore consuming but a small amount of current from the output of the i-f.

Another simple approach is to meter the change in screen voltage of an i-f amplifier stage. The swing in screen potential is caused by changes in the agc voltage applied to the stage. A reference voltage is obtained from the cathode of the audio-output stage. A 1-mA meter is suitable for the circuit shown in Fig. 8-24A. At C, a more complex design is employed which can operate directly from the agc line of a transistorized receiver. The sensitivity of the metering circuit is adjusted by changing the gain of the IC meter amplifier. An FET buffer is employed to insure that loading of the agc line will be negligible.

IMPROVING RECEIVER SELECTIVITY

INTERMEDIATE-FREQUENCY AMPLIFIERS

One of the big advantages of the superheterodyne receiver is the improved selectivity that is possible. This selectivity is obtained in the i-f amplifier, where the lower frequency allows more selectivity per stage than at the higher signal frequency. For normal a-m (double-sideband) reception, the limit to useful selectivity in the i-f amplifier is the point where too many of the high-frequency sidebands are lost. The limit to selectivity for a single-sideband signal, or a double-sideband a-m signal treated as an ssb signal, is about 2000 Hz, but reception is much more normal if the bandwidth is opened up to 2300 or 2500 Hz. The correct bandwidth for fm or pm reception is determined by the deviation of the received signal; sideband cutting of these signals results in distortion. The limit to useful selectivity in code work is around 150 or 200 Hz for hand-key speeds, but this much selectivity requires excellent stability in both transmitter and receiver, and a slow receiver tuning rate for ease of operation.

Single-Signal Effect

In heterodyne cw (or ssb) reception with a superheterodyne receiver, the beat oscillator is set to give a suitable audio-frequency beat note when the incoming signal is converted to the intermediate frequency. For example, the beat oscillator may be set to 454 kHz (the i-f being 455 kHz) to give a 1000-Hz beat note. Now, if an interfering signal appears at 453 kHz or if the receiver is tuned to heterodyne the incoming signal to 453 kHz, it will also be heterodyned by the beat oscillator to produce a 1000-Hz beat. Hence every signal can be tuned in at two places that will give a 1000-Hz beat (or any other low audio frequency). The audio-frequency image effect can be reduced if the i-f selectivity is such that the incoming signal, when heterodyned to 453 kHz, is attenuated to a very low level.

When this is done, tuning through a given signal will show a strong response at the desired beat note on one side of zero beat only, instead of the two beat notes on either side of zero beat characteristic of less-selective reception, hence the name: single-signal reception.

The necessary selectivity is not obtained with nonregenerative amplifiers using ordinary tuned circuits unless a low i-f, or a large number of circuits, is used.

Regeneration

Regeneration can be used to give a single-signal effect, particularly when the i-f is 455 kHz or lower. The resonance curve of an i-f stage at critical regeneration (just below the oscillating point) is extremely sharp, a bandwidth of 1 kHz at 10 times down and 5 kHz at 100 times down being obtainable in one stage. The audio-frequency image of a given signal thus can be reduced by a factor of nearly 100 for a 1000-Hz beat note (image 2000 Hz from resonance).

Regeneration is easily introduced into an i-f amplifier by providing a small amount of capacity coupling between grid and plate. Bringing a short length of wire, connected to the grid, into the vicinity of the plate lead usually will suffice. The feedback may be controlled by a cathode-resistor gain control. When the i-f is regenerative, it is preferable to operate the tube at reduced gain (high bias) and depend on regeneration to bring up the signal strength. This prevents overloading and increases selectivity.

The higher selectivity with regeneration reduces the over-all response to noise generated in the earlier stages of the receiver, just as does high selectivity produced by other means, and therefore improves the signal-to-noise ratio. However, the regenerative gain varies with signal strength, being less on strong signals.

Crystal Filters; Phasing

A simple means for obtaining high selectivity is by the use of a piezoelectric quartz crystal as a selective filter in the i-f amplifier. Compared to a good tuned circuit, the *Q* of such a crystal is extremely high. The crystal is ground resonant at the i-f and used as a selective coupler between i-f stages. For single-signal reception, the audio-frequency image can be reduced by 50 dB or more. Besides practically eliminating the af image, the high selectivity of the crystal filter provides good discrimination against adjacent signals and also reduces the broadband noise.

BAND-PASS FILTERS

A single high-*Q* circuit (e.g., a quartz crystal or regenerative stage) will give adequate single-signal cw reception under most circumstances. For phone reception, however, either single-sideband or a-m, a band-pass characteristic is more desirable. A band-pass filter is one that passes without unusual attenuation a desired *band* of frequencies and rejects signals outside this band. A good band-pass filter for single-sideband reception might have a bandwidth of 2500 Hz at -6 dB and 4 kHz at -60 dB; a filter for a-m would require twice these bandwidths if both sidebands were to be accommodated, thus assuring suitable fidelity.

The simplest band-pass crystal filter is one using two crystals, as in Fig. 8-25A. The two crystals are separated slightly in frequency. If the frequencies are only a few hundred Hz apart the characteristic is a good one for cw reception. With crystals about 2 kHz apart, a reasonable phone characteristic is obtained. Fig. 8-2 shows a selectivity characteristic of an amplifier with a bandpass (at -6 dB) of 2.4 kHz, which is typical of what can be expected from a two-crystal bandpass filter.

More elaborate crystal filters, using four and six crystals, will give reduced bandwidth at -60 dB without decreasing the bandwidth at -6 dB. The resulting increased "skirt selectivity" gives better rejection of adjacent-channel signals. "Crystal-lattice" filters of this type are available commercially for frequencies up to 40 MHz or so, and they have also been built by amateurs from inexpensive transmitting-type crystals. (See Vester, "Surplus-Crystal High-Frequency Filters," *QST*, January, 1959; Healey, "High-Frequency Crystal Filters for SSB," *QST*, October, 1960.)

Two half-lattice filters of the type shown at Fig. 8-25A can be connected back to back as shown at B. The channel spacing of Y1 and Y2 will depend upon the receiving requirements as discussed in the foregoing text. Ordinarily, for ssb reception (and nonstringent cw reception) a frequency separation of approximately 1.5 kHz is suitable. The overall i-f strip of the receiver is tuned to a frequency which is midway between Y1 and Y2. C1 is tuned to help give the desired shape to the passband. L1 is a bifilar-wound toroidal inductor which tunes to the i-f frequency by means of C1. The values of R1 and R2 are identical and are determined by the filter response desired. Ordinarily the ohmic value is on the order of 600 ohms, but values as high as 5000 ohms are sometimes used. The lower the value of resistance, the broader and flatter will be the response of the filter. Though the circuit at B is shown in a transistorized circuit, it can be used with vacuum tubes or integrated circuits as well. The circuit shows an i-f frequency of 9 MHz, but the filter can be used at any desired frequency below 9 MHz by altering the crystal frequencies and the tuned circuits. Commercial versions of the 9-MHz lattice filter are available at moderate cost.[1] War-surplus FT-241 crystals in the 455-kHz range are inexpensive and lend themselves nicely to this type of circuit.

Mechanical filters can be built at frequencies below 1 MHz. They are made up of three sections; an input transducer, a mechanically resonant filter

[1] Spectrum International, P. O. Box 87, Topsfield, MA 01983. Also, McCoy Electronics Co., Mount Holly Springs, PA.

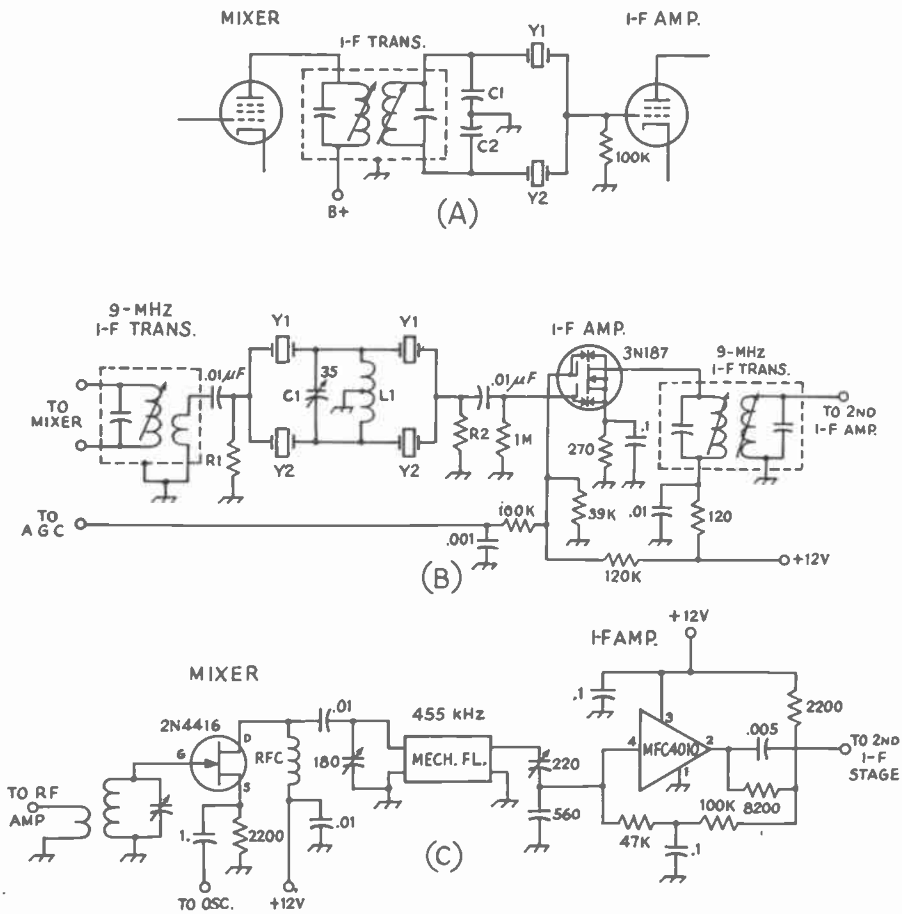


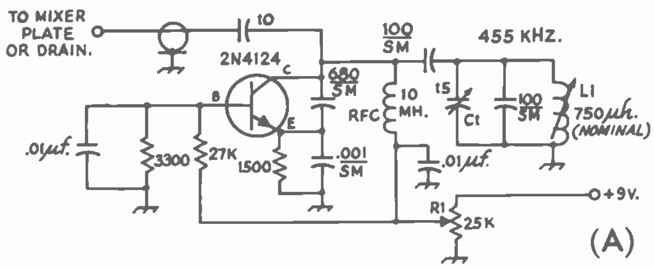
Fig. 8-25 — A half-lattice bandpass filter at A; B shows two half-lattice filters in cascade; C shows a mechanical filter.

section, and an output transducer. The transducers use the principle of magneto-striction to convert the electrical signal to mechanical energy, then back again. The mechanically resonant section consists of carefully machined metal disks supported and coupled by thin rods. Each disk has a resonant frequency dependent upon the material and its dimensions, and the effective Q of a single disk may be in excess of 2000. Consequently, a mechanical filter can be built for either narrow or broad passband with a nearly rectangular curve. Mechanical filters are available commercially and are used in both receivers and single-sideband transmitters. They are moderately priced.

The signal-handling capability of a mechanical filter is limited by the magnetic circuits to from 2 to 15 volts rms, a limitation that is of no practical importance provided it is recognized and provided for. Crystal filters are limited in their signal-handling ability only by the voltage breakdown limits, which normally would not be reached before the preceding amplifier tube was overloaded. A more serious practical consideration in the use of any high-selectivity component is the prevention of coupling "around" the filter, externally, which can only degrade the action of the filter.

The circuit at Fig. 8-25C shows a typical hookup for a mechanical filter. FL1 is a Collins 455-FB-21, which has an ssb band-pass characteristic of 2.1 kHz. It is shown in a typical solid-state receiver circuit, but can be used equally as well in a tube-type application.

Placement of the BFO signal with respect to the passbands of the three circuits at A, B, and C, is the same. Either a crystal-controlled or self-excited oscillator can be used to generate the BFO signal and the usual practice is to place the BFO signal at a frequency that falls at the two points which are approximately 20 dB down on the filter curve, dependent upon which sideband is desired. Typically, with the filter specified at C, the center frequency of FL1 is 455 kHz. To place the BFO at the 20-dB points (down from the center-frequency peak) a signal at 453 and 456 kHz is required.



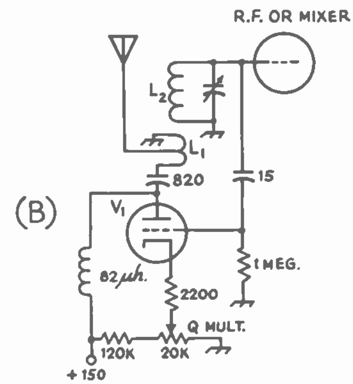


Fig. 8-26 – An i-f Q multiplier for use with a bipolar transistor (A). At B, a tube-type rf Q multiplier which can be used at the first stage of the receiver. The antenna coil is used for feedback to V1, which then introduces "negative resistance" to L2.

Q Multiplier

The "*Q* Multiplier" is a stable regenerative stage that is connected in parallel with one of the i-f stages of a receiver. In one condition it narrows the bandwidth and in the other condition it produces a sharp "null" or rejection notch. A "tuning" adjustment controls the frequency of the peak or null, moving it across the normal passband of the receiver i-f amplifier. The *shape* of the peak or null is always that of a single tuned circuit (Fig. 2-42) but the effective *Q* is adjustable over a wide range. A *Q* Multiplier is most effective at an i-f of 500 kHz or less; at higher frequencies the rejection notch becomes wide enough (measured in Hz) to reject a major portion of a phone signal. Within its useful range, however, the *Q* Multiplier will reject an interfering carrier without degrading the quality of the desired signal.

In the "peak" condition the *Q* Multiplier can be made to oscillate by advancing the "peak" (regeneration) control far enough and in this condition it can be made to serve as a beat-frequency oscillator. However, it cannot be made to serve as a selective element and as a BFO *at the same time.* Some inexpensive receivers may combine either a *Q* Multiplier or some other form of regeneration with the BFO function, and the reader is advised to check carefully any inexpensive receiver he intends to buy that offers a regenerative type of selectivity, in order to make sure that the selectivity is available when the BFO is turned on.

A representative circuit for a transistorized *Q*-multiplier is given in Fig. 8-26A. The constants given are typical for i-f operation at 455 kHz. L1 can be a J. W. Miller 9002 or 9102 slug-tuned inductor. A 25,000-ohm control, R1, permits adjustment of the regeneration. C1 is used to tune the *Q*-multiplier frequency back and forth across the i-f passband for peaking or notching adjustments. With circuits of this type there is usually a need to adjust both R1 and C1 alternately for a peaking or notching effect, because the controls tend to interlock as far as the frequency of oscillation is concerned. A *Q*-multiplier should be solidly built in a shielded enclosure to assure maximum stability.

Q multipliers can be used at the front end of a receiver also, as shown at B in Fig. 8-26. The enhancement of the *Q* at that point in a receiver greatly reduces image problems because the selectivity of the input tuned circuit is increased markedly. The antenna coil, L1, is used as a feedback winding to make V1 regenerative. This in effect adds "negative resistance" to L2, increasing its *Q*. A 20,000-ohm control sets the regeneration of V1, and should be adjusted to a point just under regeneration for best results. Rf *Q* multiplication is not a cure for a poor-quality inductor at L2, however.

T-Notch Filter

At low intermediate frequencies (50 – 100 kHz) the T-notch filter of Fig. 8-27 will provide a sharp tunable null.

The inductor *L* resonates with *C* at the rejection frequency, and when $R = QX_L/4$ the rejection is maximum. (X_L is the coil-reactance and *Q* is the coil *Q*.) In a typical 50-kHz circuit, *C* might be 3900 pF making *L* approximately 2.6 mH. When *R* is greater than the maximum-attenuation value, the circuit still provides some rejection, and in use the inductor is detuned or shorted out when the rejection is not desired.

At higher frequencies, the T-notch filter is not sharp enough with available components to reject only a narrow band of frequencies.

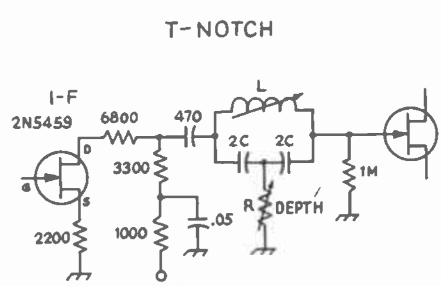
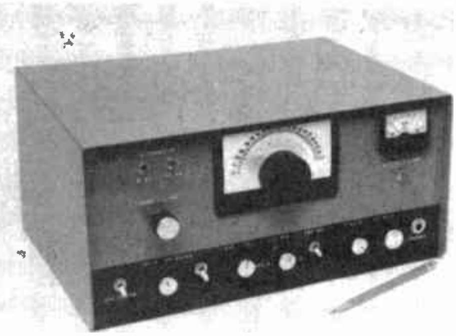


Fig. 8-27 – Typical T-notch (bridged-T) filter, to provide a sharp notch at a low i-f. Adjustment of L changes the frequency of the notch; adjustment of R controls the notch depth.

SOME RECEIVER DESIGN NOTES

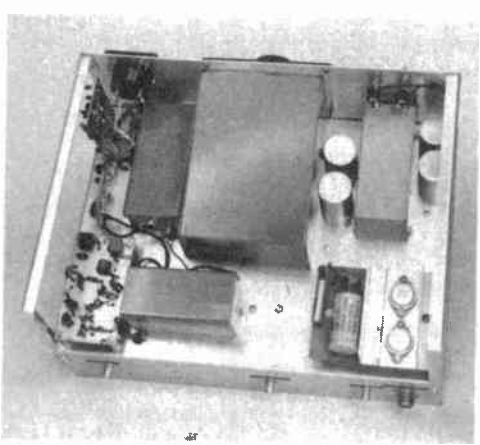
The receiver to be discussed in these notes incorporates some advanced ideas and has been included for its value as a theory article rather than a construction project. Consequently, templates will not be available. For further information, see article by W1CER in June and July *QST* for 1976. Also, a set of converters has been designed by W7ZOI and the article appears in *QST* for June, 1976. The basic receiver covers the 1.8-MHz band.



Front-End Features

Although the circuit treated here is for a one-band receiver (1.8 to 2.0 MHz), the design procedures are applicable to any amateur band in the hf spectrum. Down converters to cover 80 through 10 meters can be employed and they are founded on the same concepts to be discussed here.

Fig. 1 shows the rf amplifier mixer, and post-mixer amplifier. What may seem like excessive elaboration in design is a matter of personal whim, but the features are useful, nevertheless. For example, the two front-end attenuators aren't essential to good performance, but are useful in making accurate measurements (6, 12 or 18 dB) of signal levels during on-the-air experiments with other stations (antennas, amplifiers and such). Also, FL2, a fixed-tuned 1.8- to 2-MHz bandpass filter, need not be included if the operator is willing to repeak the three-pole tracking filter (FL1) when tuning about in the band. The fixed-tuned filter is useful when the down converters are in use.



Top-chassis view of the receiver. The *R-C* active filter and audio preamplifier are built on the pc board at the upper left. To the right is the BFO module in a shield box. The agc circuit is seen at the lower left, and to its right is the i-f strip in a shield enclosure. The large shield box at the upper center contains the VFO. To its right is the tunable front-end filter. The three-section variable capacitor is inside the rectangular shield box. The audio-amplifier module is seen at the lower right. The small board (mounted vertically) at the left center contains the product detector. Homemade end brackets add mechanical stability between the panel and chassis, and serve as a support for the receiver top cover.

The benefits obtained from a highly selective tunable filter like FL1 are seen when strong signals are elsewhere in (or near) the 160-meter band. Insertion loss was set at 5 dB in order to narrow the filter response. In this example the high-*Q* slug-tuned inductors are isolated in aluminum shields, and the three-section variable capacitor which tunes them is enclosed in a shield made from pc-board sections. Bottom coupling is accomplished with small toroidal coils.

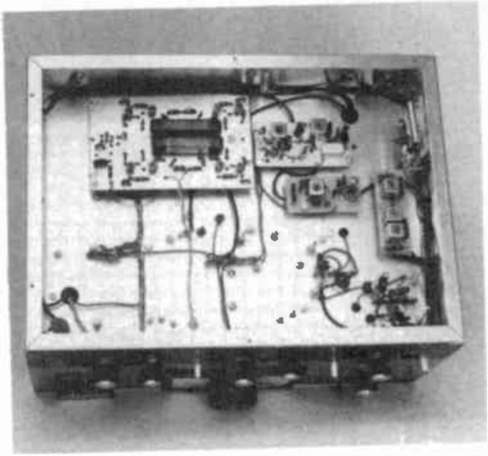
Rf amplifier Q1 was added to compensate for the filter loss. It is mismatched intentionally by means of L10 and L11 to restrict the gain to 6-dB maximum. Some additional mismatching is seen at L12, and the mixer is overcoupled to the FET-tuned output tank to broaden the response (1.8 to 2 MHz). The design tradeoffs do not impair performance. The common-gate rf stage has good dynamic range and IMD characteristics.

The doubly balanced diode-ring mixer (U1) was chosen for its excellent reputation in handling high signal levels, having superb port-to-port signal isolation, and because of its good IMD performance. The module used in this design is a commercial one which contains two broadband transformers and four hot-carrier diodes with matched characteristics. The amateur can build his own mixer assembly in the interest of reduced expense. At the frequencies involved in this example, it should not be difficult to obtain performance equal to that of a commercial mixer.

A diplexer is included at the mixer output (L13 and the related .002 capacitors). The addition was worthwhile, as it provided an improvement in the noise floor and IMD characteristics of the receiver. The diplexer works in combination with matching network L14, a low-pass L-type circuit. (The diplexer is a high-pass network which permits the 56-ohm terminating resistor to be seen by the mixer without degrading the 455-kHz i-f. The low-pass portion of the diplexer helps reject all frequencies above 455 kHz so that the post-mixer amplifier receives only the desired information.) The high-pass section of the diplexer starts rolling off at 1.2 MHz. A reactance of 66 ohms was chosen to permit use of standard-value capacitors in the low-Q network.

A pair of source-coupled JFETs is used in the post-mixer i-f preammplifier. The 10,000-ohm gate resistor of Q2 sets the transformation ratio of the L network at 200:1 (50 ohms to 10 K-Ω). An L network is used to couple the preamplifier to a diode-switched pair of Collins mechanical filters which have a characteristic impedance of 2000 ohms. The terminations are built into the filters.

Gain distribution to the mixer is held to near unity in the interest of good IMD performance. The preamplifier gain is approximately 25 dB. The choice was made to compensate for the relatively high insertion loss of the mechanical filters – 10 dB. Without the high gain of Q2 and Q3, there would be a deterioration in noise figure.



Considerable space remains beneath the chassis for the addition of accessory circuits or a set of down converters. At the lower right are the adjustment screws for the tunable filter, plus the bottom-coupling toroids. At the right center is the fixed-tuned front-end filter. To the left is the rf-amplifier module. A 100-kHz MFJ Enterprises calibrator is seen at the far upper right. Immediately to its left is the mixer/amplifier assembly. The large board at the upper center contains the i-f filters and post-filter amplifier. Most of the amplifier components have been tacked beneath the pc board because of design changes which occurred during development.

Local Oscillator

A low noise floor and good stability are essential traits of the local oscillator in a quality receiver. The requirements are met by the circuit of Fig. 2. Within the capabilities of the ARRL lab measuring procedures, it was determined that VFO noise was at least 90 dB below fundamental output. Furthermore, stability at 25° C ambient temperature was such that no drift could be measured from a cold start to a period three hours later. Mechanical stability is excellent: Several sharp blows to the VFO shield box caused no discernible shift in a cw beat note while the 400-Hz i-f filter was actuated. VFO amplifier Q14 is designed to provide the recommended +7-dBm mixer injection. Furthermore, the output pi tank of Q14 is of 50-ohms characteristic impedance. Though not of special significance in this application, the measured harmonic output across 50 ohms is −36 dB at the second order and −47 dB at the third order.

Filter Module

In the interest of minimizing leakage between the filter input to output ports, diode switching (Fig. 3) was used. The advantage of this method is that only dc switching is required, thereby avoiding the occasion for unwanted rf coupling across the contacts and wafers of a mechanical switch. 1N914 diodes are used to select FL3 (400-Hz bandwidth) or FL4 (2.5-kHz bandwidth). Reverse bias is applied to the nonconducting diodes. This lessens the possibility of leakage through the switching diodes. Because the Collins filters have a characteristic impedance of 2000 ohms, the output coupling capacitors from each are 120 pF rather than low-reactance .01-μF units, as used at the filter inputs. Without the smaller value of capacitance, the filters would see the low base impedance of Q4, the post-filter i-f amplifier. The result would be one of double termination in this case, leading to a loss in signal level. Additionally, the 120-pF capacitors help to divorce the input capacitance of the amplifier stage. The added capacitance would have to be subtracted from the 350- and 510-pF resonating capacitors at the output ends of the filters.

The apparent overall receiver gain is greatest during cw reception, owing to the selectivity of cw filter FL3. To keep the S-meter readings constant for a given signal level in the ssb and cw modes, R7 has been included in the filter/amplifier module. In the cw mode, R7 is adjusted to bias Q4 for an S-meter reading equal to that obtained in the ssb mode. Voltage for the biasing is obtained from the diode switching line during cw reception.

Although a 2N2222A is not a low-noise device, the performance characteristics are suitable for this circuit. A slight improvement in noise figure would probably result from the use of an MPF102, 40673 or low-noise bipolar transistor in that part of the circuit.

Performance Notes

The tuning range of the receiver is 200 kHz. This means that for use with converters the builder will have to satisfy himself with the cw or ssb band

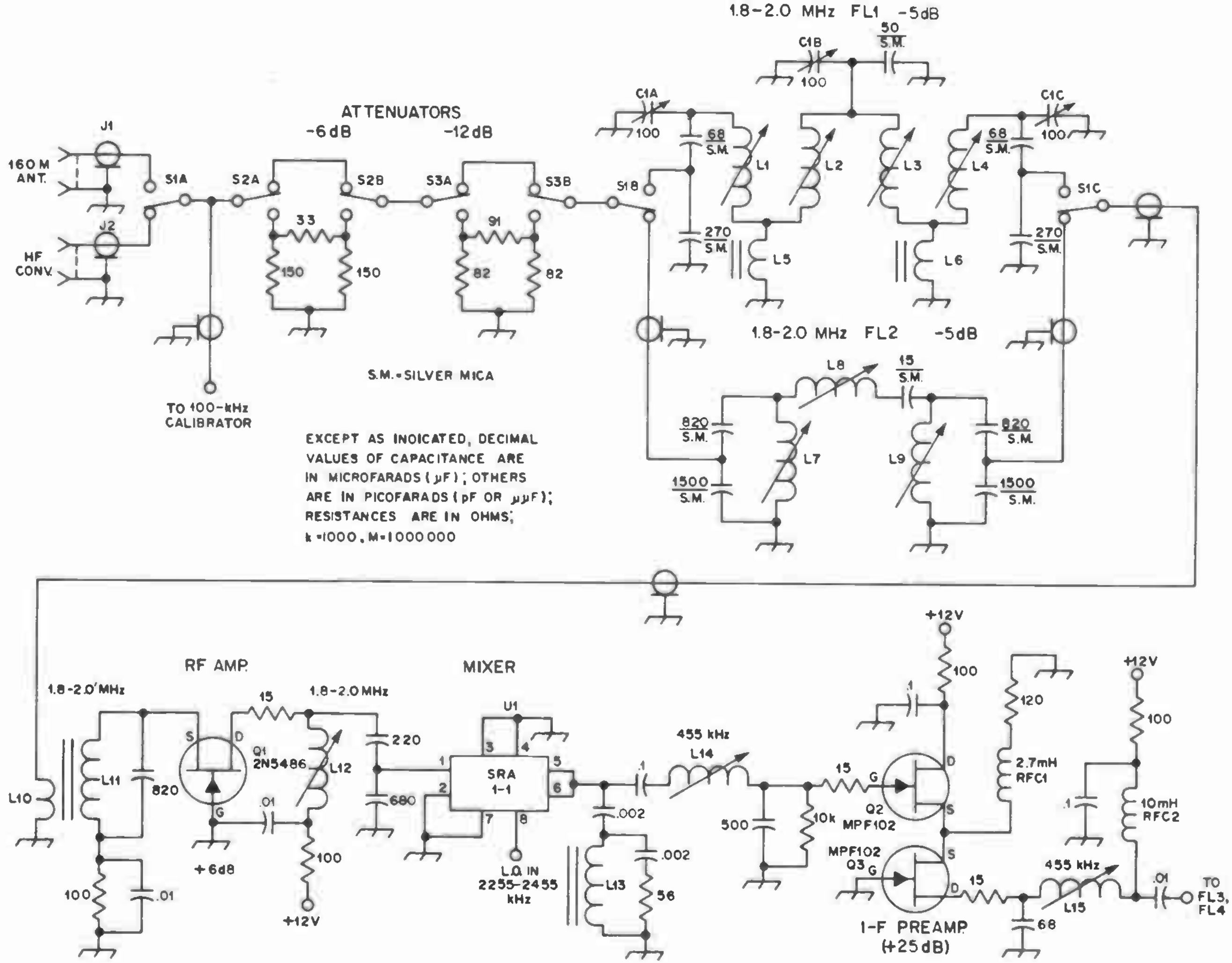

1.8–2.0 MHz FL1 -5dB
ATTENUATORS
-6dB
-12dB
160 M ANT.
HF CONV.
J1
J2
S1A
S2A
S2B
S3A
S3B
S1B
S1C
C1A
C1B
C1C
L1
L2
L3
L4
L5
L6
1.8–2.0 MHz FL2 -5dB
L7
L8
L9
TO 100-kHz CALIBRATOR
S.M. = SILVER MICA
EXCEPT AS INDICATED, DECIMAL VALUES OF CAPACITANCE ARE IN MICROFARADS (μF); OTHERS ARE IN PICOFARADS (pF OR μμF); RESISTANCES ARE IN OHMS; k = 1000, M = 1000 000
RF AMP.
1.8-2.0 MHz
Q1 2N5486
+6dB
+12V
MIXER
U1
SRA 1-1
L.O. IN 2255–2455 kHz
455 kHz
L14
Q2 MPF102
MPF102 Q3
2.7mH RFC1
10mH RFC2
L15
TO FL3, FL4
I-F PREAMP (+25dB)

Fig. 1 — Schematic diagram of the receiver front end. Fixed-value capacitors are disk ceramic unless otherwise noted. Resistors are 1/2-W composition. All slug-tuned inductors are contained in individual shield cans which are grounded.

C1 — Three-section variable, 100 pF per section. Model used here obtained as surplus.
J1 — SO-239.
J2 — Phono jack.
L1, L4 — 38 to 68 µH, Q_u of 175 at 1.8 MHz, slug tuned (J. W. Miller 43A685CBI in Miller S-74 shield can).
L2, L3 — 95 to 187 µH, Q_u of 175 at 1.8 MHz, slug tuned (J. W. Miller 43A154CBI in S-74 shield can).
L5, L6 — 1.45-µH toroid inductor, Q_u of 250 at 1.8 MHz. 15 turns No. 26 enam. wire on Amidon T-50-2 toroid.
L7, L9 — 13-µH slug-tuned inductor (J. W. Miller 9052).
L8 — 380-µH slug tuned inductor (J. W. Miller 9057).
L10 — 16 turns No. 30 enam. wire over L11 winding.
L11 — 45 turns No. 30 enam. wire on Amidon T-50-2 toriod, 8.5 µH.
L12 — 42-µH slug-tuned inductor, Q_u of 50 at 1.8 MHz (J. W. Miller 9054).
L13 — 8.7-µH toroidal inductor, 12 turns No. 26 enam. wire on Amidon FT-37-61 ferrite core.
L14 — 120 to 280 µH, slug-tuned inductor (J. W. Miller 9056).
L15 — 1.3 to 3.0 mH, slug-tuned inductor (J. W. Miller 9059).
Q1, Q2, Q3 — Motorola JFET.
RFC1 — 2.7-mH miniature choke (J. W. Miller 70F273A1).
RFC2 — 10-mH miniature choke (J. W. Miller 70F102A1).
S1 — Three-pole, two-position phenolic wafer switch.
S2, S3 — Two-pole, double-throw miniature toggle.
U1 — Mini-Circuits Labs. SRA-1-1 doubly balanced diode mixer, (2913 Quentin Rd., Brooklyn, NY 11229).

segments. The alternatives are to increase the local oscillator tuning range to 500 kHz, or use a multiplicity of converters to cover the cw and ssb portions of each band.

Some severe lab tests were undertaken with the completed receiver, aimed at learning how "crunchproof" the front end really was. A quarter-wavelength end-fed wire (inverted L) was matched to the receiver 50-ohm input port. The far end of the antenna was situated three feet away from the W1AW end-fed Zepp antenna. A pk-pk voltage of 15 was measured across the 50-ohm receiver input jack by means of a Tektronix model 453 scope while W1AW was operating. Now, that's a lot of rf energy! With that high level of rf voltage present, a 10-µV signal was fed into the receiver and spotted 2 kHz away from the W1AW operating frequency. No evidence of cross modulation could be observed, and desensitization of the receiver could not be discerned by ear. The spread from 1.8 to 2 MHz was tuned, and no IM products were heard.

Dynamic range tests were performed in accordance with the Hayward paper in *QST* for July, 1975. Noise floor was −135 dBm, IMD was 95 dB and 1 dB of blocking occurred at some (undetermined) point greater than 123 dB above the noise floor. The latter measurement is inconclusive because blocking did not become manifest within the output capability of the model 80 generators used in the ARRL lab. The resultant receiver noise figure at 1.8 MHz is 13 dB, which is more than adequate for the high atmospheric noise level on 160 meters.

A receiver i-f system should be capable of providing a specific gain, have an acceptable noise figure and respond satisfactorily to the applied agc. Two of the more serious shortcomings in some designs are poor agc (clicky, pumping or inadequate range) and insufficient i-f gain.

Because of past successes, it was elected to use a pair of RCA CA3028A ICs in the i-f strip. Somewhat greater i-f dynamic range is possible with MC1590G ICs, and they are the choice of many builders. However, the CA3028As, configured as differential amplifiers, will provide approximately 70 dB of gain per pair when operated at 455 kHz. This gives an agc characteristic from maximum gain to full cutoff which is entirely acceptable for most amateur work.

Fig. 4 shows the i-f amplifiers, product detector and Varicap-tuned BFO. Transformer coupling is used between U2 and U3 and also between U3 and the product detector. The 6800-ohm resistors used across the primaries of T2 and T3 were chosen to force an impedance transformation which the transformers can't by themselves provide: Available Miller transformers with a 30,000-ohm primary to 500-ohm secondary characteristic are used. U2 and U3 have 10- and 220-ohm series resistors in the signal lines. These were added to discourage vhf parasitic oscillations.

Agc is applied to pin 7 of each IC. Maximum gain occurs at +9 V, and minimum gain results when the agc voltage drops to its low value — +2 V. The agc is rf-derived, with i-f sampling for the agc

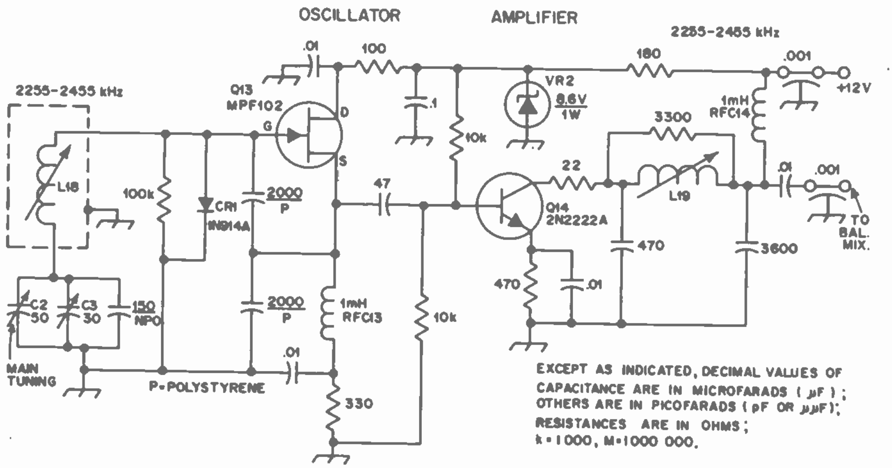


Fig. 2 — Circuit diagram of the local oscillator. Capacitors are disk ceramic unless specified differently. Resistors are 1/2-W composition. Entire assembly is enclosed in a shield box made from pc-board sections.

C2 — Double-bearing variable capacitor, 50 pF.
C3 — Miniature 30-pF air variable.
CR1 — High-speed switching diode, silicon type 1N914A.
L18 — 17- to 41-μH slug-tuned inductor, Q_u of 175 (J. W. Miller 43A335CBI in Miller S-74 shield can).
L19 — 10- to 18.7-μH slug-tuned pc-board inductor (J. W. Miller 23A155 RPC).
RFC13, RFC14 — Miniature 1-mH rf choke (J. W. Miller 70F103A1).
VR2 — 8.6-V, 1-W Zener diode.

amplifier being done at pin 6 of U3 through a 100-pF blocking capacitor.

The 1000-ohm decoupling resistors in the 12-V feed to U2 and U3 drop the operating voltage to +9. This aids stability and reduces i-f system noise. The amplifier strip operates with unconditional stability.

Product Detector

A quad of 1N914A diodes is used in the product detector. Hot-carrier diodes may be preferred by some, and they may lead to slightly better performance than the silicon units chosen. A trifilar broadband toroidal transformer, T4, couples the i-f amplifier to the detector at a 50-ohm impedance level. BFO injection is supplied at 0.7-V rms.

BFO Circuit

In the interest of lowering the cost of this project, a Varicap (CR10 of Fig. 4) is used to control the BFO frequency. Had a conventional system been utilized, three expensive crystals would have been needed to handle upper sideband, lower sideband and cw. The voltage-variable capacitor tuning method shown in Fig. 4 is satisfactory if the operator is willing to change the operating frequency of the BFO when changing receive modes. Adjustment is done by means of front-panel control R1. Maximum drift with this circuit was measured as 5 Hz from a cold start to a time three hours later. A Motorola MV-104 tuning diode is used at CR10.

To vary the BFO frequency from 453 to 457 kHz, the diode is subjected to various amounts of back bias, applied by means of R1. Regulated voltage (VR1) is applied to the oscillator and tuning diode.

Q6 functions as a Class A BFO amplifier/buffer. It contains a pi-network output circuit and has a 50-ohm output characteristic. The main purpose of the amplifier stage is to increase the BFO injection power without loading down the oscillator.

AGC Circuit

Fig. 5 shows the agc amplifier, rectifier, dc source follower and op-amp difference amplifier. An FET is used at Q10 because it exhibits a high-input impedance, and will not, therefore, load down the primary of T3 in Fig. 4. Q1 is direct coupled to a pnp transistor, Q11. Assuming that Rs and R2 are treated as a single resistance, Rs, the Q10/Q11 gain is determined as: Gain (dB) = 20 log Rc ÷ Rs. Control R2 has been included as part of Rs to permit adjustment of the agc loop gain. Each operator may have a preference in this regard. The writer has the agc set so it is fully actuated at a signal-input level of 10 μV. Agc action commences at 0.2 μV (1 dB of gain compression).

Agc disabling is effected by removing the operating voltage from Q10 and Q11 by means of S5. Manual i-f gain control is made possible by

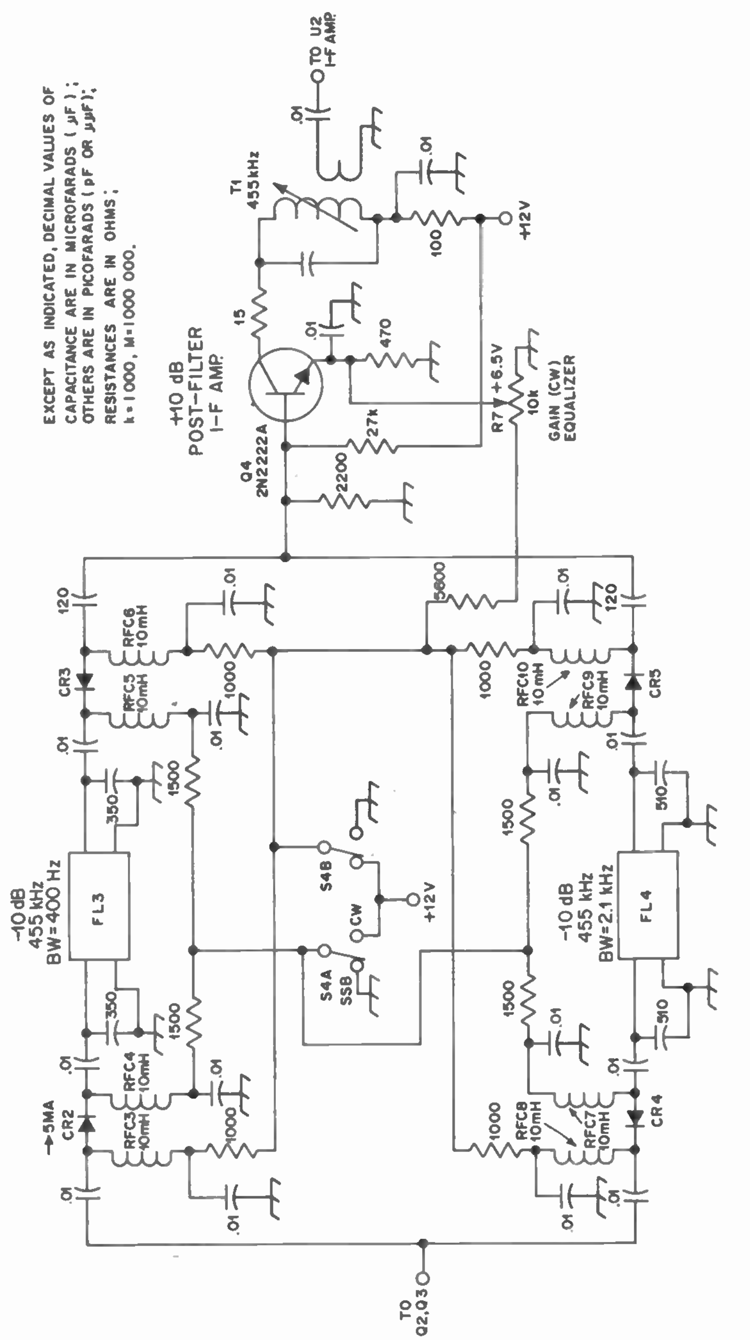


Fig. 3 — Schematic diagram of the filter and i-f post-filter amplifier. Capacitors are disk ceramic. Resistors are 1/2-W composition.

CR2-CR5, incl. — High-speed silicon switching diode, 1N914A.

FL3 — Collins mechanical filter F455FD-04.

FL4 — Collins mechanical filter F455FD-25.

RFC3-RFC10, incl. — 10-mH miniature rf choke (J. W. Miller 70F102A1).

R7 — Pc-board control, 10,000 ohms, linear taper.

S4 — Double-pole, double-throw toggle or wafer.

T1 — Miniature 455-kHz i-f transformer (J. W. Miller 2067, 30,000 to 500 ohms).

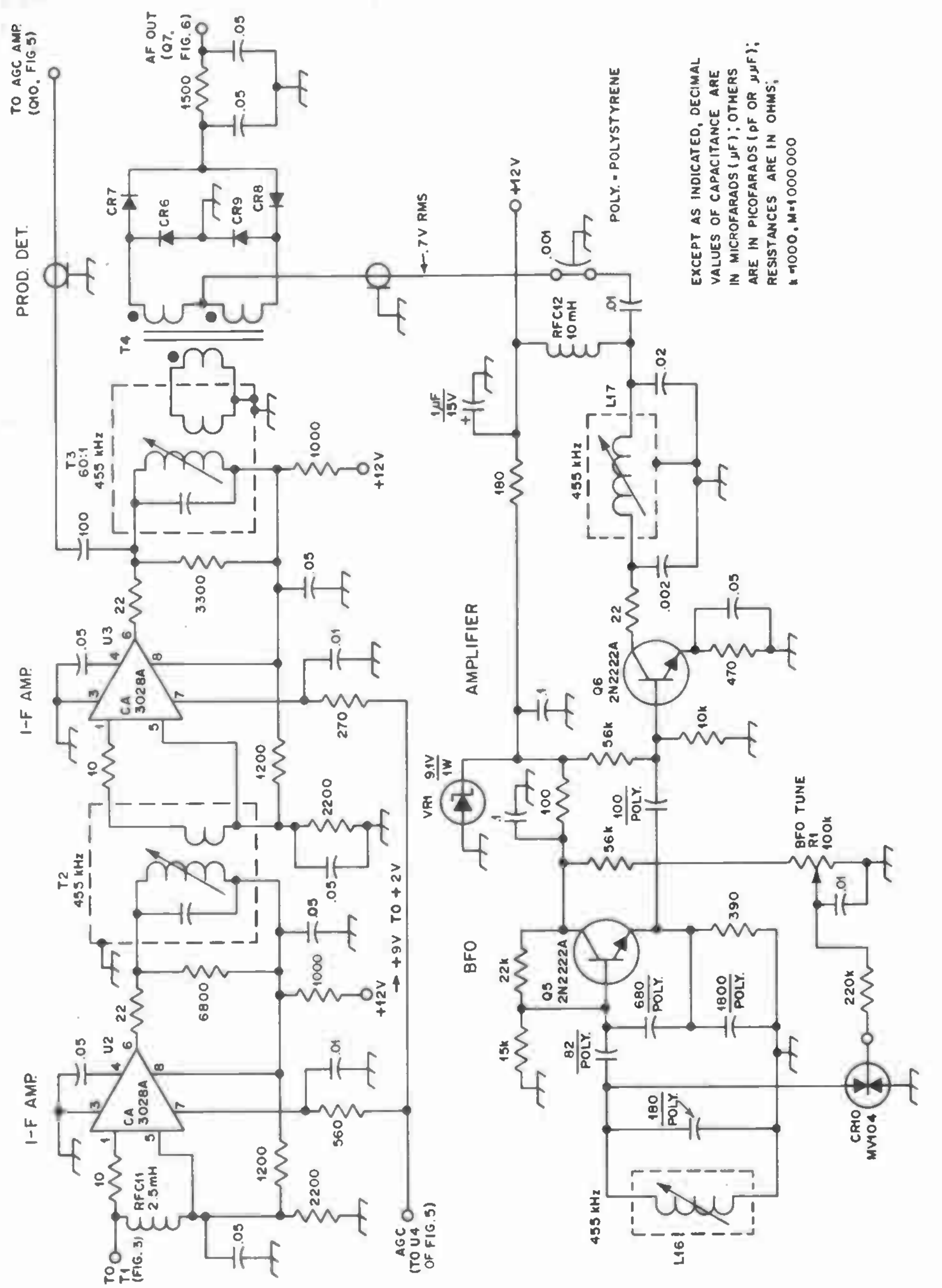


Fig. 4 — Circuit of the i-f amplifier, BFO, and product detector. Capacitors are disk ceramic unless noted differently. Fixed-value resistors are 1/2-W composition. Dashed lines show shield enclosures. The BFO and i-f circuits are installed in separate shield boxes. The *R-C* active filter and af preamplifier are on a common circuit board which is not shielded.

CR6-CR9, incl. — High-speed silicon, 1N914A or equiv.

CR10 — Motorola MV-104 Varicap tuning diode.

L16 — Nominal 640-μH slug-tuned inductor (J. W. Miller 9057).

L17 — Nominal 60-μH slug-tuned inductor (J. W. Miller 9054).

R1 — 100,000-ohm linear-taper composition control (panel mount).

RFC11 — 2.5-mH miniature choke (J. W. Miller 70F253A1).

RFC12 — 10-mH miniature choke (J. W. Miller 70F102A1).

T2, T3 — 455-kHz i-f transformer. See text. (J. W. Miller 2067).

T4 — Trifilar broadband transformer. 15 trifilar turns of No. 26 enam. wire on Amidon T-50-61 toroid core.

U2, U3 — RCA IC.

VR1 — 9.1-V, 1-W Zener diode.

adjusting R3 of Fig. 5. Agc delay is approximately one second. Longer or shorter delay periods can be established by altering the values of the Q14 gate resistor and capacitor. Agc amplifier gain is variable from six to 40 dB by adjusting R2. The arrangement at Q14 and U4 was adapted from a design by W7ZOI. Agc action is smooth, and there is no evidence of clicks on the attack during strong-signal periods. At no time has agc "pumping" been observed.

Audio System

A major failing of many receivers is poor-quality audio. For the most part this malady is manifest as cross-over distortion in the af-output amplifier. Moreover, some receivers have marginal audio-power capability for normal room volume when a loudspeaker is used. Some transformerless single-chip audio ICs (0.25- to 2-W class) exhibit a prohibitive distortion characteristic, and this is especially prominent at low signal levels. The unpleasant effect is one of "fuzziness" when listening to low-level signals. Unfortunately, external access to the biasing circuit of such ICs is not typical, owing to the unitized construction of the chips.

Since "sanitary" audio is an important feature of a quality communications receiver, a circuit containing discrete devices was used. The complimentary-symmetry output transistors and the op-amp driver are configured in a manner similar to that used by Jung in his *Op Amp Cookbook* by Howard Sams. Maximum output capability is 3.5 W into an 8-ohm load. An LM-301A driver was chosen because of its low-noise profile. There has been no aural evidence of distortion at any signal level while using the circuit of Fig. 6. The game played in this situation is one of having considerably more audio power available than is ever needed – a rationale used in hi-fi work.

R-C Active CW Filter

A worthwhile improvement in signal-to-noise ratio can be realized during weak-signal reception by employing an *R-C* active bandpass filter. A two-pole version (FL5) is shown in Fig. 6. A peak frequency of 800 Hz results from the *R* and *C* values given.

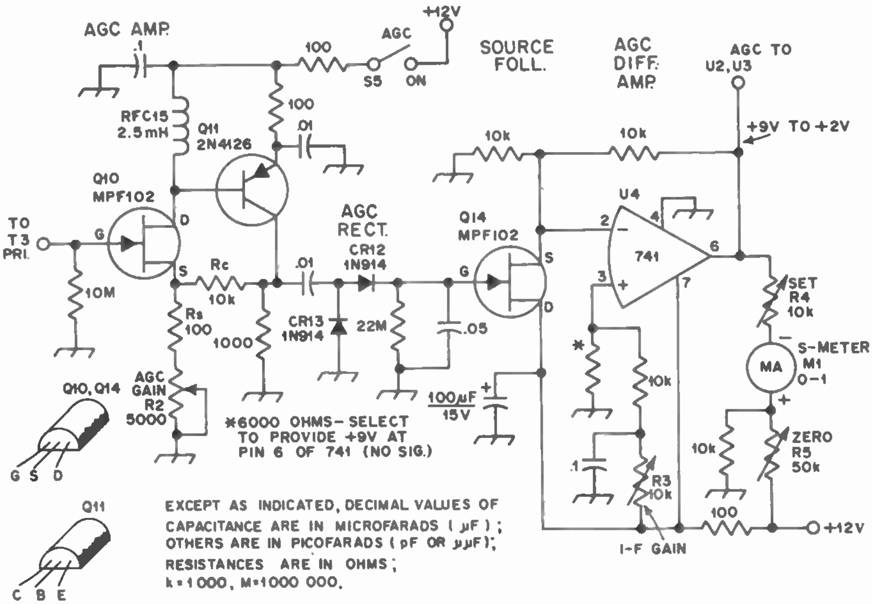


Fig. 5 – Schematic diagram of the agc system. Capacitors are disk ceramic except when polarity is indicated, which signifies electrolytic. Fixed-value resistors are 1/2-W composition. This module is not enclosed in a shield compartment.

CR12, CR13 – High-speed silicon. 1N914A or equiv.
M1 – 0 to 1-mA meter.
Q10, Q11, Q14 – Motorola transistor.
R2, R4, R5 – Linear-taper composition pc-board mount control.
R3 – 10,000-ohm linear-taper control, panel mounted.
RFC15 – 2.5-mH miniature choke (J. W. Miller 70F253A1).
S5 – Single-pole, single-throw toggle.
U4 – Dual-in-line 8-pin 741 op amp.

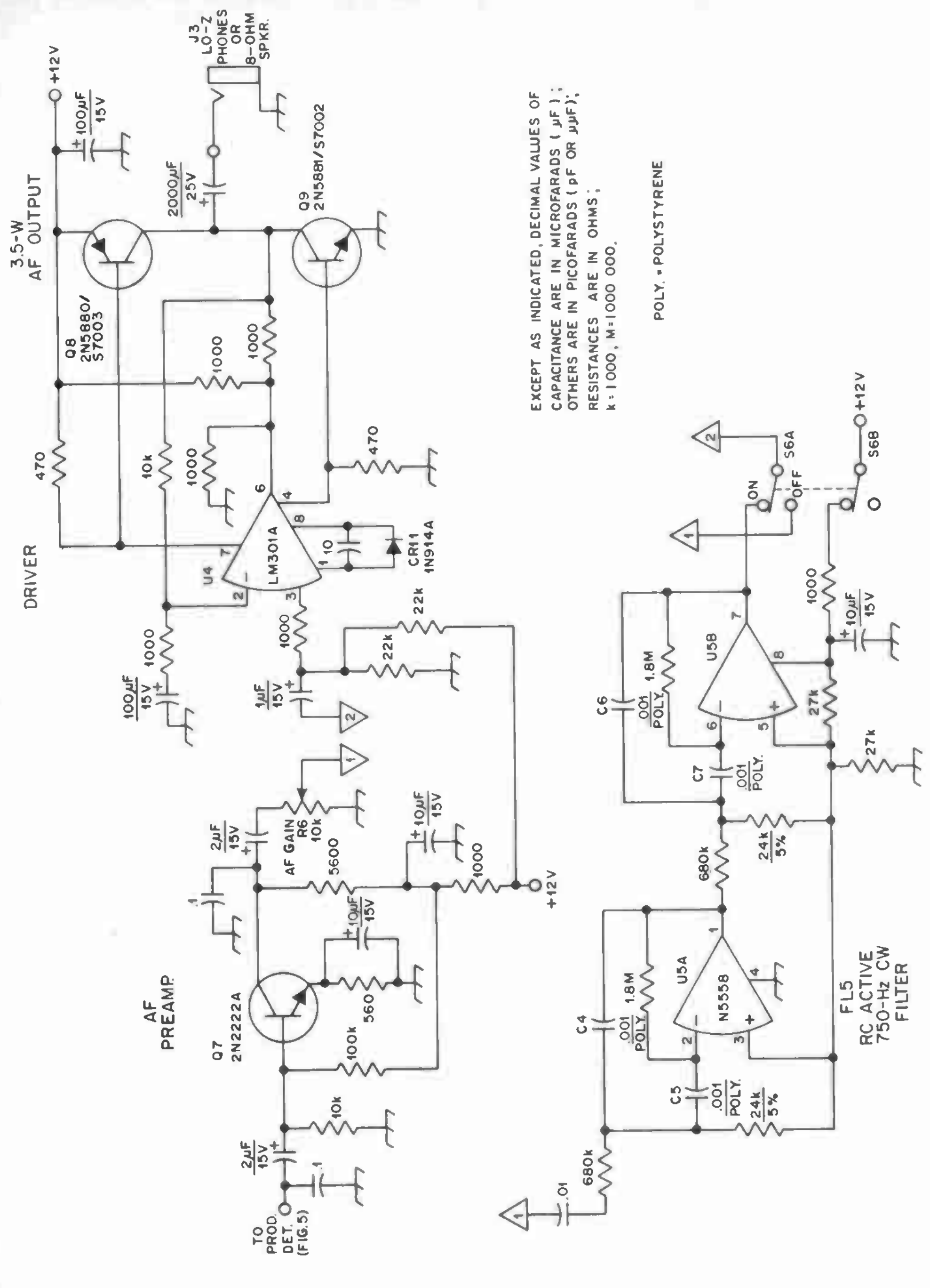


Fig. 6 — Diagram of the audio amplifier and *R-C* active filter. Capacitors are disk ceramic unless otherwise noted. Polarized capacitors are electrolytic or tantalum. Fixed-value resistors are 1/2-W composition. This circuit is not contained in a shield box. Heat sinks are used with Q8 and Q9. C4-C7, incl. — See text.

CR11 — High-speed silicon, 1N914A or equiv.
J3 — Phone jack.
R6 — 10,000-ohm audio-taper composition control, panel mounted.
S6 — Double-throw, double-pole toggle.
U4 — National Semiconductor LM-301A IC.
U5 — Signetics N5558 dual op-amp IC.

The benefits of FL5 are similar to those described by Hayward in his "Competition-Grade CW Receiver" article. He used a second i-f filter (at the i-f strip output) to reduce wide-band noise from the system. The *R-C* active filter serves in a similar manner, but performs the signal "laundering" at audio rather than at rf. The technique has one limitation – monotony in listening to a fixed-frequency beat note, which is dictated by the center frequency of the filter. The *R-C* filter should be designed to have a peak frequency which matches the cw beat-note frequency preferred by the operator. That is, if the BFO is adjusted to provide an 800-Hz cw note, the center frequency of FL5 should also be 800 Hz.

Experience with FL5 in this receiver has proved in many instances that weak DX signals on 160 meters could be elevated above the noise to a Q5 copy level, while without the filter solid copy was impossible. It should be stressed that high-*Q* capacitors be used from C4 to C7, inclusive, to assure a sharp peak response. Polystyrene capacitors satisfy the requirement. To ensure a well-defined (minimum ripple) center frequency, the capacitors should be matched closely in value (5 percent or less). Resistors of 5-percent tolerance should be employed in the circuit, where indicated in Fig. 6.

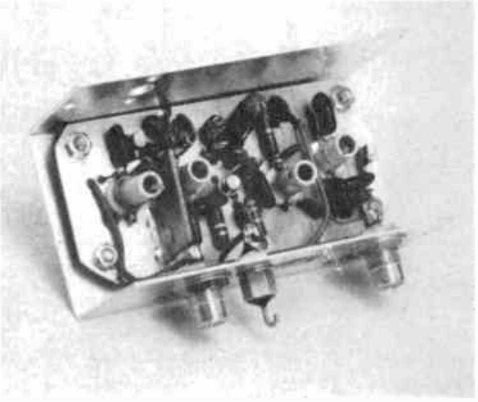
NBS EARS FOR YOUR HAM-BAND RECEIVERS

Radio amateurs are becoming worshipers of the sun, constantly peering at projections of the surface for signs of black spots; "sun spots," they call them.

Why has interest in sun spots suddenly turned many amateurs into solar astronomers? The answer is becoming more obvious each day. The state of the earth's ionosphere and geomagnetic field, and therefore radio propagation for a given time, is directly related to conditions on the sun. Sun-spot activity will soon be on the increase, and geomagnetic activity is constantly changing. As a result, unexpected, often undetected, band openings are occurring. But clouds and other weather phenomena can make it impossible for observers to see the sun, much less any spots that might otherwise be visible. That geomagnetic disturbance business – how can one tell when such an event will occur? Well, one way is to consult the "DXer's Crystal Ball."[1]

If you don't have a Ouija board at your disposal, the National Bureau of Standards (NBS) stations WWV and WWVH offer an alternative source of information on solar and geomagnetic activity. Propagation bulletins are broadcast hourly by these stations, and articles in *QST* and this Handbook have shown several ways in which this information can be put to work in helping amateurs make better use of their air time.

Some amateurs may have a problem using this information source because a large portion of amateur gear manufactured in recent years is for ham-bands-only reception. Some receivers do offer an "extra" band, usually 15 MHz, which is useful sometimes, in some areas of the world, but not in others. An inexpensive solution to the problem for those who want to receive the NBS stations' transmissions, but don't want to spend the money for a general-coverage receiver, is a converter which uses one of the amateur frequencies for an i-f output. Selection of the proper component values allows the potential user to build a converter that will cover the WWV or WWVH frequency most usable at his location.



WWV converter as nested in the chassis. The shield shown in the photograph was found to be unnecessary since stray coupling between the input and output of the rf amplifier proved not to be a problem.

The converter described here, when used with an amateur-bands-only receiver, provides for reception of 10-, 15- or 25-MHz NBS stations WWV or WWVH. The receiver, when tuned to 4, 14 or 21 MHz, serves as the i-f amplifier, detector and audio stages. The low current drain of the conver-

[1] Tilton, "The DXer's Crystal Ball," Parts I through III, *QST*, June, August and September, 1975.

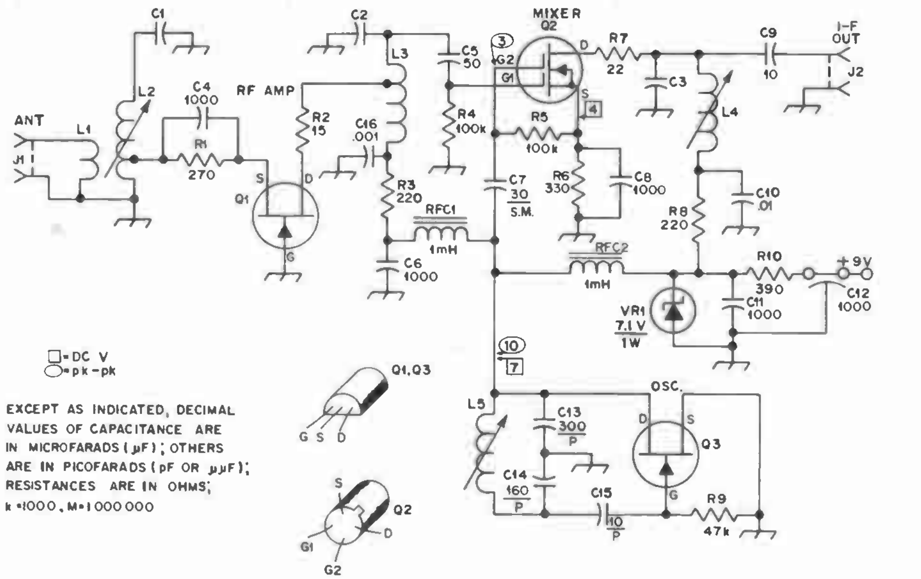


Fig. 1 — Schematic diagram, of the WWV-to-ham bands converter. The oscillator output frequency of 11 MHz was chosen to provide the reception of the three most commonly used WWV frequencies (10, 15 and 25 MHz), without the need to change the oscillator frequency.

ter (15 mA typical) lends itself to operation from a 9-volt transistor-radio battery and to use with QRP equipment.

The Circuit

The schematic diagram of the converter is given in Fig. 1. With the exception of the Miller coil forms, nearly all of the components used can be purchased from Radio Shack or Lafayette Radio Electronic stores. For coverage of the 10-, 15- and 25-MHz WWV frequencies, component values of the three tuned circuits in the rf-amplifier and mixer stages must be selected from Table 1. This approach reduces the complexity of the converter by eliminating band-switching circuitry, but restricts the converter to use on only one NBS frequency at a time.

A common-gate JFET rf amplifier provides 8 dB of gain in this converter and has good IMD and overload immunity. A 40673 MOSFET is used as the mixer in the converter. The output circuit of the mixer uses a low value of coupling capacitor as an alternative to an rf voltage divider or other output-coupling technique. This was done as a parts-saving step and does not seem to degrade the performance of the converter significantly.

VR1 provides adequate regulation of the V+ to the converter pc board. The regulator diode is

TABLE I

	C1 – C2	*C3*	*L1*	*L2 – L3*	*L4*
10 MHz	90 pF	22 pF	2-1/2 turns No. 34 enamel over L2.	Same as L5*	Same as L5*
15 MHz	43 pF	300 pF	1-1/2 turns No. 24 enamel over L2	Same as L5*	5.5 μH (nom.) Miller 46A566CPC
25 MHz	22 pF	48 pF	1-1/2 turns No. 24 enamel over L2	1.8 μH(nom.) Miller 46A186CPC	Same as L5*

*L5 – 2.42-2.96 μH, Miller 46A276CPC

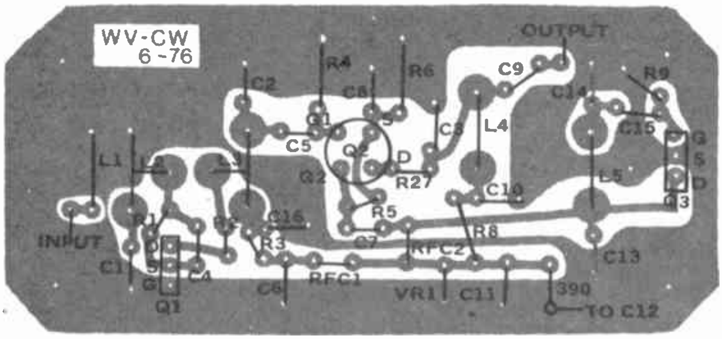


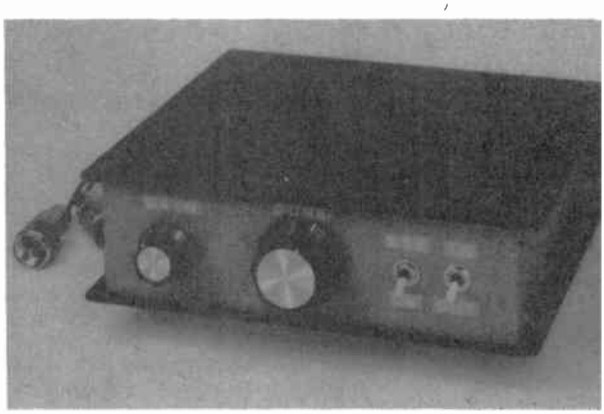
Fig. 2 — Etching pattern and parts-placement guide for the converter circuit board. 1/2-watt resistors were used throughout, but 1/4-watt resistors may also be used if preferred.

placed on the V+ line for the entire circuit of the converter; the converter, therefore, is operating at 7.1 volts. Any voltage from 9 to 18 volts will power the converter.

The converter is housed in an aluminum Minibox; dimensions of the box are 4 X 2-1/8 X 1-5/8 inches. Radio Shack part number 270-239 is suitable. As can be noted from the photograph, the converter pc board was laid out to facilitate 1/2-watt resistors, but 1/4-watt resistors are accept able since power consumption for the converter is very low. Silver-mica or polystyrene capacitors should be used for C7, C13, C14 and C15 because they aid stability in the oscillator circuit. Disc-ceramic capacitors are suitable for use in the remainder of the converter circuit.

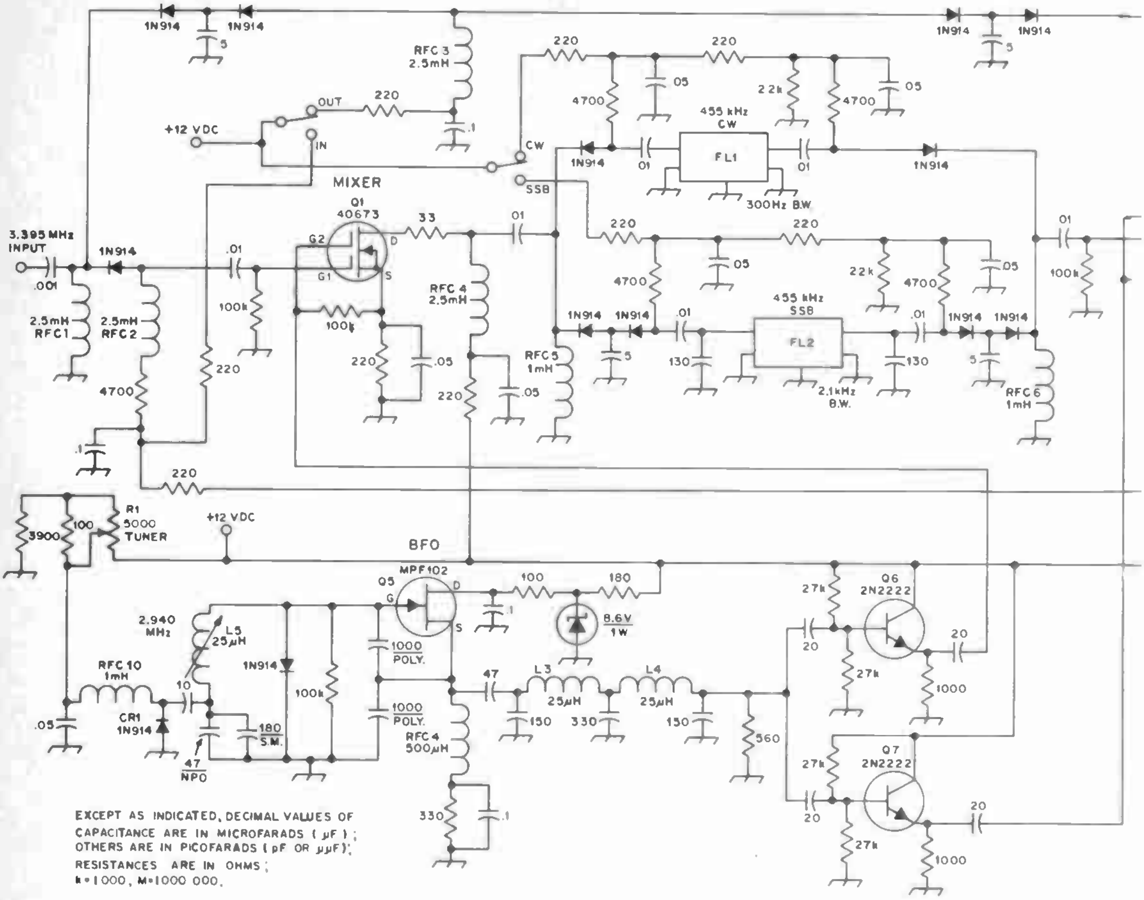
A BAND-PASS TUNER FOR ADJUSTABLE SELECTIVITY

Many modern receivers have fixed-frequency oscillator circuits for injection at the product detector during ssb and cw operation. The frequency of the fixed-frequency oscillator determines the characteristic range of the received signal. For cw operation, it determines the BFO offset frequency and accordingly the pitch of the signals heard. A receiver which has a variable BFO (sometimes called a pitch control) allows the operator to select the pitch at which the cw signals are centered in the filter response. During ssb reception, the control may be used to select a pleasing quality for the incoming signal. A pitch control is undesirable, however, if the BFO in the receiver is to act as a carrier oscillator in the transmitter for transceive operation. In this case, the BFO must be on exactly the same frequency for both receive and transmit periods; indeed many times the same oscillator is used. Transceive operation is not compatible with a pitch control, and unfortunately the operator is required to accept the resultant BFO offset. While it is possible to shift a BFO/carrier oscillator to obtain a desirable receiver response characteristic, the affect to the transmitted signal could be detrimental. These disadvantages can be overcome by incorporating the circuit shown in Fig. 1 into the station receiver or transceiver. It may be used with any apparatus



having an i-f of 3.395 MHz. Other i-f ranges could be used if appropriate changes are made to the Band-Pass Tuner VFO input and output circuits. The experienced builder should not encounter any difficulty.

The purpose of the Band-Pass Tuner is to allow the operator to adjust the range of frequencies



which pass through the filter without changing the frequency of the receiver or the transmit offset frequency. The unit could be built for cw or ssb operation only. Several filters, however, may be included in the circuit and switched to coincide with the desired mode of transmission. When the Band-Pass Tuner is interconnected to a transceiver, one needs to assure that the transmit signal does not pass through the Band-Pass Tuner system. This could be accomplished with either a double-pole relay or a diode switching array as shown in the circuit diagram. Generally speaking, the latter provides more isolation than the mechanical relay and is recommended.

Circuit Functions

The main function of the Band-Pass Tuner is to convert the i-f signal to a different frequency where sharp-skirt filters may be used to increase the selectivity of a receiving system. After the signal is filtered, it is then converted back to the receiver (or transceiver) i-f. The technique used to obtain variable band-pass response is to employ a VFO for the conversion oscillator and use this energy not only to convert the signal from 3.395 MHz down to 455 kHz, but also to convert the signal back to the receiver i-f range. Since the down conversion is equal to the up conversion (the same oscillator is used for both), changing the VFO frequency does not change the frequency of the received signal. The output frequency is always equal to the input frequency; the VFO only changes the position of the signal around the 455 kHz filter system.

A Circuit diagram for the Band-Pass Tuner is given in Fig. 1. Dual-gate MOSFETs are used to accomplish both the down and the up conversion. Since the phone-band mechanical filter has considerable insertion loss (about 10 dB), an RCA 40673 amplifier is included to bring the signal up to the proper level for re-entering the receiver i-f system. The 40673 could be controlled by voltage supplied from the receiver agc bus; however, it was not necessary. Normal agc action of the receiver seemed unaffected by the inclusion of the Band-Pass Tuner.

The VFO must be capable of reaching stability from a cold start in just a few minutes. A VFO which drifts more than a kilohertz during warmup will cause the operator to have to readjust the

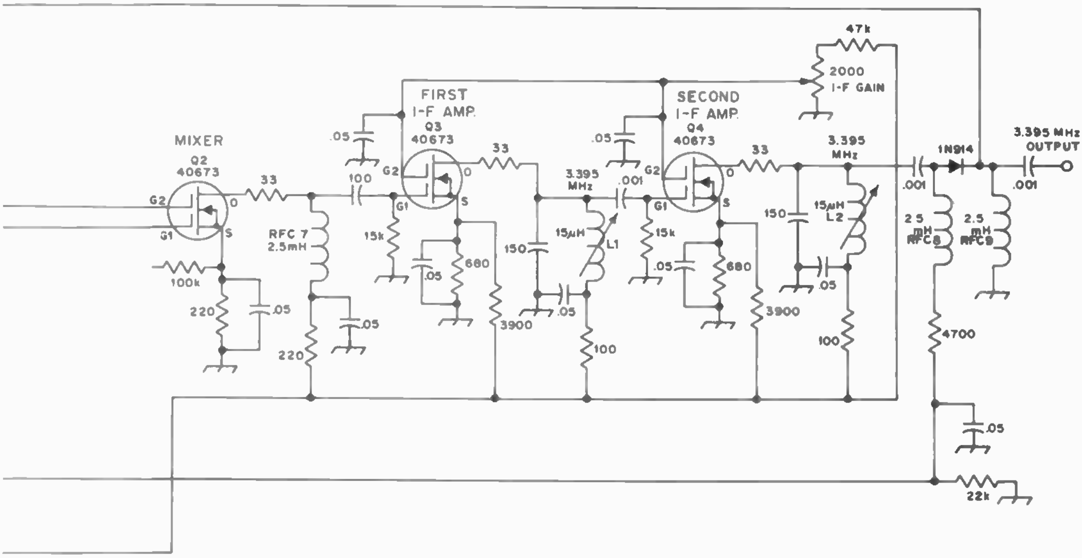


Fig. 1 — Circuit diagram for the Band-Pass Tuner. All resistors are 1/2-watt composition. FL1 and FL2 are Collins types. RFC's are from Millen. All capacitors are disc ceramic. The output of the Pass-Band Tuner should be connected in the receiver at a point before the receiver i-f filter system. The i-f control should be adjusted to give suitable gain when the unit is placed in operation as outlined in the text.

tuning during this period. Since the tuning is accomplished with a varactor diode which has a limited range, drift must be held to within a few hundred Hertz for warmup and within about 25 Hz for normal operation. The voltage source for the VFO must be regulated. CR1 in Fig. 1 serves that function.

Most receivers with a 3.395 MHz second i-f have a first i-f in the range of 8 MHz. Since the third harmonic of the VFO falls in that range, a low-pass filter was incorporated to eliminate any receiver spurious response. The tenth harmonic of the VFO is the only one to fall in an hf amateur band. It was not detectable on the receiver shown in the photographs.

A pair of emitter followers are incorporated in the VFO injection line; each one offers isolation between the VFO and the associated mixer. Additionally, isolation is provided between the input mixer, U1 and the output mixer, U3. Insufficient isolation on this line could cause poor selectivity characteristics (parallel path to the filters) and instability.

Construction

Several methods may be used to enclose the Band-Pass Tuner. If sufficient space is available in the receiver or transceiver, and if the control functions can be made part of the front-panel layout (without drilling holes!), then mounting the circuit boards internally would be desirable. The other option is to fabricate a cabinet or obtain a sufficiently large Minibox to house the circuit boards and controls.

All of the components for the VFO, converters, and amplifiers are included on one board. The filter switching network and the filters are mounted on the same circuit board.

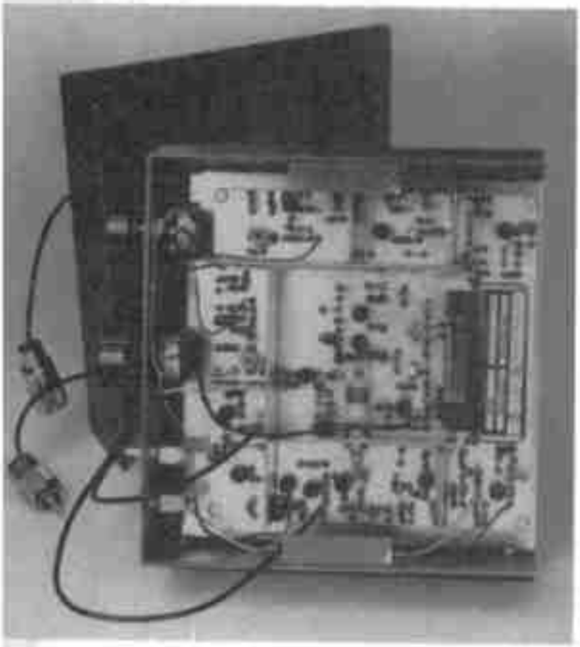
Shielded leads should be used between the input and output points on the circuit board and the station receiver.

Receiver Interconnection and Alignment

The input terminal of the Band-Pass Tuner should be connected to the output of the second mixer in the receiver. The output of the Band-Pass Tuner connects directly to the input of the receiver crystal filter.

Alignment is simple. Before the Band-Pass Tuner is installed in the receiver, tune in the receiver crystal calibrator and log the S-meter reading. Install the Band-Pass Tuner. Set the VFO for 2.940 MHz. A frequency counter is handy for this step, however a general-coverage receiver is suitable. The band-pass tuning control R5, should cause a frequency change of approximately three kilohertz either side of 2.940 MHz.

With the receiver and the Band-Pass Tuner turned on, and the calibrator tuned in, the band-pass control (VFO) should be adjusted to show a peak reading on the S-Meter. Then the gain



Top view of the Band-Pass Tuner

control, R1, should be set so that the receiver S-meter reading under the above conditions is the same as the reading taken before the Band-Pass Tuner was installed. The gain control will no doubt have to be changed during normal operation if more than one filter is used.

Operation

The incorporation of the Band-Pass Tuner does not change the normal operation of the receiver. S1 may be used to select an out condition (Band-Pass Tuner out of the circuit), the ssb filter, or a sharper cw filter. The band-pass tuning control may be adjusted to give the desired filter response in relation to the signal being received.

With most receivers, the change from one sideband to the other at the receiver mode switch will not require a change in the position of the band-pass tuning control. It will, however, change the tuning direction of the response. For instance, if the tuning control is rotated in one direction to favor a higher pitch with usb operation, when used on lower sideband, the same direction of rotation will cause the favored pitch to become lower.

A COMMUNICATIONS RECEIVER WITH DIGITAL FREQUENCY READOUT

Front view of the communications receiver built by WA1JZC. The receiver controls are grouped at the right side of the front panel, with the digital readout occupying the left side.

This solid-state receiver will enable the operator to tune the amateur bands from 160 to 10 meters in 500 kHz segments. An integral part of the unit is a digital frequency counter that may be used to display the received frequency directly to the nearest 100 Hz. The basic receiver consists of a single-conversion tunable i-f which covers a range that includes the 160-meter band. Converters for each additional desired band are placed ahead of the tunable i-f and may be built into the same enclosure. This approach combines the virtues of high performance, moderate complexity, and reasonable cost with plenty of flexibility.

Circuit Overview

The design objectives incorporated in this receiver include optional coverage of all of the amateur bands below 30 MHz, ability to withstand strong signals without cross modulation or overloading, selectable phone or cw bandwidth i-f filters, extensive use of diode switching, and direct display of the frequency of incoming signals. The signal path through the receiver may be traced with the aid of the block diagram, Fig. 1.

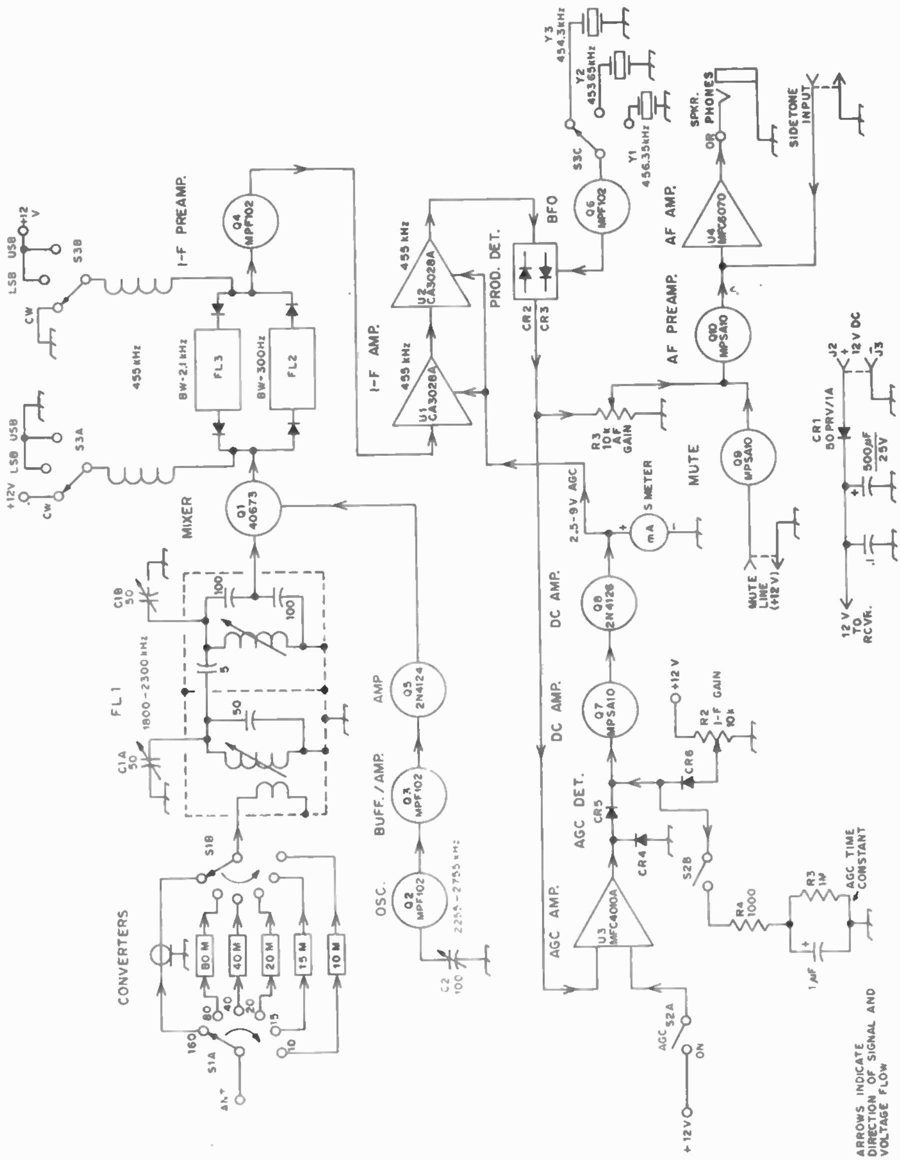
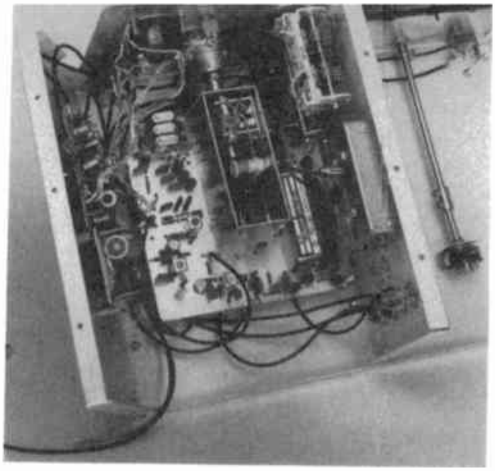


Fig. 1 — Block diagram of the solid-state communications receiver. A polarity-guarding diode, CR1, is shown at the lower right. This diode and the two bypass capacitors associated with it are mounted just inside the receiver cabinet near J2 and J3. These components do not appear in the schematic diagrams of this article.

Band switch S1 connects the rf input, i-f input, and positive supply voltage to the desired converter board. The circuits used in the 40-, 20-, 15-, and 10-meter converters are of similar design. Each one consists of an FET rf amplifier, a crystal-controlled heterodyne oscillator (HFO), and a dual-gate MOSFET mixer stage. No rf amplifier is included in the 80-meter design; an input bandpass network takes its place. The HFO and mixer in the 80-meter circuit are similar to those of the other converters. In each case, the desired output from the mixer is the difference between the incoming signal frequency and the HFO frequency. This difference frequency will fall within the range covered by the tunable i-f, 1800 to 2300 kHz. For 160-meter operation, S1 connects the antenna lead straight through to the input of the tunable i-f, bypassing the converters. No rf amplifier is needed (or used) on 160 meters. In the tunable i-f, a bandpass filter, FL1, is used ahead of a dual-gate MOFSET mixer, Q1. Oscillator injection to Q1, supplied by a three-stage variable frequency oscillator assembly (Q2, Q3, and Q5), may be tuned over a 500-kHz wide range. In this case, the desired mixer output is at 455 kHz. I-f selectivity following the mixer is established by means of a narrow-bandwidth crystal filter, FL2, for cw operation, or by a mechanical filter with a somewhat broader response, FL3, for use with ssb. A diode-switching network permits either filter to be chosen. An i-f preamplifier (Q4) following the filters compensates for filter insertion loss. Two stages of i-f amplifica-



View of the receiver compartment prior to installation of the converter boards.

tion (U1 and U2) after the preamplifier provide ample gain and dynamic range. A passive product detector (CR2, CR3) at the i-f output receives injection from a crystal-controlled BFO. (Q6). The BFO incorporates diode switching of three crystals, one for cw, one for lower sideband reception, and one for upper sideband reception. The audio output from the product detector is split into two independent paths. One path leads to an a-f preamplifier (Q10) followed by an integrated audio power amplifier (U4) which drives a speaker or headphones. The other path runs to an audio-derived agc circuit consisting of U3, CR4 and CR5, Q7, and Q8 plus associated components. The agc is applied to U1 and U2, and additionally provides an S-meter indication on M1.

Tunable I-F Section

A tuning range of 1800 to 2300 kHz was chosen for this portion of the receiver (Fig. 2A). A front-end filter, peaked by means of a front-panel mounted variable capacitor, is used to assure rejection of potentially troublesome out-of-band signals on 160-meters, particularly those of local bc stations. Inexpensive loopstick inductors are used at L2 and L3. It was necessary to remove 125 turns of wire from each so that they would be suitable for tuning from 1800 to 2300 kHz with the split-stator variable capacitor used in the filter. The unloaded Q of each, after modification, is 125 at 1.9 MHz. An active mixer was chosen for the first stage of the tunable i-f section to limit the number of receiver stages required for suitable overall gain. Q1, the dual-gate MOSFET, operates with the signal gate tapped down on FL1 to reduce the possibility of over-loading and cross modulation. The conversion gain is roughly 10 dB, and that more than compensates for the insertion loss of the tunable filter, FL1. It is wise to keep the signal voltage at gate 1 as low as practicable to assure good mixer performance. A small amount of forward bias is applied to both gates of the 40673 mixer to increase the linearity and conversion gain of that stage. A 33-ohm resistor is used in the drain to prevent vhf parasitic oscillations.

Local Oscillator

Perhaps the most important consideration when choosing a vfo circuit is its freedom from drift. The incorporation of narrow-bandwidth filters (FL2, FL3) in this receiver makes local oscillator stability a must, as any change in the LO frequency can be easily detected. The local oscillator depicted in Fig. 2B is unconditionally stable, from both a mechanical and an electrical standpoint. Oscillator drift, measured from a cold start, was less than 75 Hz after 30 minutes of operation. VFOs patterned after this circuit have been used in several different applications, and comparable stability has been achieved in every case.

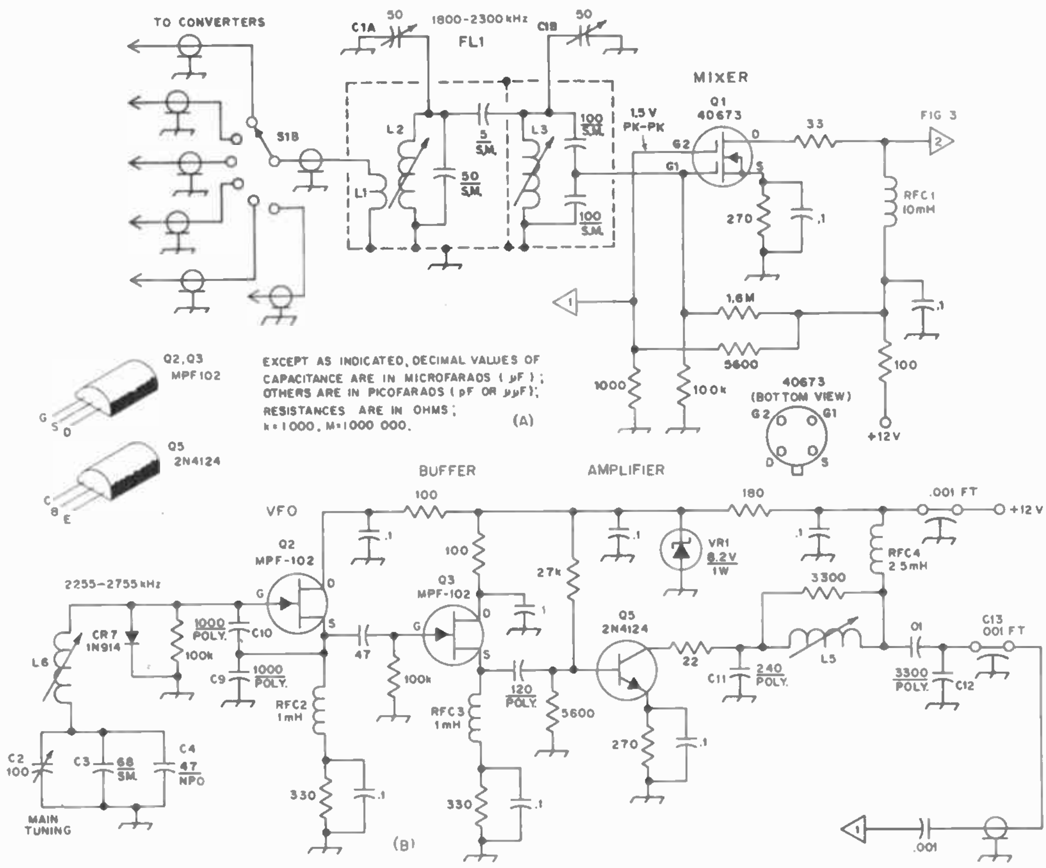
With the circuit constants shown in Fig. 2B, the vfo covers the desired frequency range of 2255 kHz to 2755 kHz, with an additional 15 kHz on either side of that range. VFO linearity across the entire tuning range is not as exacting a requirement when an "electronic dial" is used to display the receiver frequency as it is when some type of mechanical readout is used. Using a 10:1 ratio ball bearing drive to turn the shaft of C2, a tuning rate of 80 kHz per main-tuning-knob revolution is obtained at the low frequency end of the range, and 125 kHz per revolution at the high frequency end of the range. If personal preference dictates a slower tuning rate, dual ratio 36:1 and 6:1 drives are available.

Referring to the circuit diagram, Q2 is used in a series-tuned Colpitts oscillator configuration. Polystyrene capacitors were chosen for their temperature stability at C9 and C10 because these components are part of the frequency-determining network. If the inductor specified for L6 is not available, its replacement must exhibit a similarly high Q (approximately 150 measured at 2.5 MHz) to assure proper oscillation. With the exception of C2, C3, C4, and L6, all VFO components are mounted on a single printed-circuit board. C3 and C4 are soldered directly between the C2 stator terminal and the (grounded) capacitor body to minimize lead inductance effects. Inasmuch as the position of the powdered iron core of L6 is variable, the added inductance of the wire from C2 to L6, and from L6 to the pc board is taken into account during the initial VFO alignment. Nevertheless, from the standpoint of mechanical and thermal stability, it is desirable to keep the connecting leads short and direct, using a stiff gauge of wire.

Q3 is connected as a source follower, and is intended to act as a buffer in order to isolate the oscillator circuit from later stages. The buffer drives a single-transistor amplifier stage, Q5. Output from the VFO is taken off of the Q5 collector through a pi-network tank circuit, composed of C11, L5, C12, and C13. Operating voltages for the three transistorized vfo stages are derived from a Zener regulated 8.2-volt bus which is, in turn, derived from the 12-volt receiver power supply. A variation of plus or minus 15% in the receiver supply voltage results in less than a 5 Hz change in the VFO frequency. Additional details concerning the design of the VFO appear in the November, 1974 issue of *QST*, page 22.

Fig. 2 — Schematic diagram of the 160-meter front-end mixer and local oscillator. The mixer is at A, and the oscillator and buffer are at B. Fixed-value capacitors are disk ceramic unless otherwise indicated. Resistors can be 1/4- or 1/2-watt composition types unless specified differently.

C1 — Dual-section 50-pF variable.
C2 — Single-section 100-pF variable (J.W. Miller 2101 or equiv.).
L1 — 3 turns small-diameter insulated wire wound over ground end of L2.
L2, L3 — Radio Shack No. 270-376 ferrite bc antenna with 125 turns of wire removed (see text).
L5 — Pc-mount slug-tuned coil, 10.0 - 18.7 μH (J.W. Miller 23A155RPC).
L6 — Slug-tuned coil — 3/8-inch diameter ceramic form, red core, 18.8 — 41.0 μH (Miller 42A335CBI).
RFC1 — 10-mH miniature encapsulated inductor (Millen Mfg. Co. J302-10,000).
RFC2, RFC3 — 1-mH miniature encapsulated inductor (Millen J302-1000).
RFC4 — 2.5-mH miniature encapsulated inductor (Millen J302-2500).
S1 — Four-pole, six-position rotary switch.
VR1 — 8.2-V, 1-W Zener diode.



I-F Strip

The output of the mixer, Q1, contains not only the desired difference between the VFO and the (post-conversion) incoming signal frequencies, but the sum of these frequencies and higher order products of these frequencies as well. I-f selectivity is developed at the output of the mixer by means of sharp filters resonant at 455 kHz, which both reject the unwanted mixing products and establish the skirt selectivity of the receiver. A choice of i-f bandwidths, one suitable for cw reception and the other tailored for ssb, is made available by the use of two separate filters, FL2 and FL3. FL2 is a crystal filter with a bandwidth of 300 Hz measured at −6 dB from the peak response. FL3 is a 2.1 kHz bandwidth mechanical filter.

A diode switching arrangement, shown schematically in Fig. 4, is used to choose the appropriate filter. When the mode switch (S3) is set to cw, a dc path is created from the positive supply bus through R8, RFC5, CR11, RFC7, RFC8, RFC10, CR12, RFC11, and R9 to ground. Simultaneously, CR9 and CR10 become reverse biased, so that they look like very high impedances. When diodes CR11 and CR12 are forward biased, they appear as very low impedances, thereby opening an rf path from the drain of Q1, through C14, CR11, C15, FL2, C16, CR12, and C17, to the gate of the

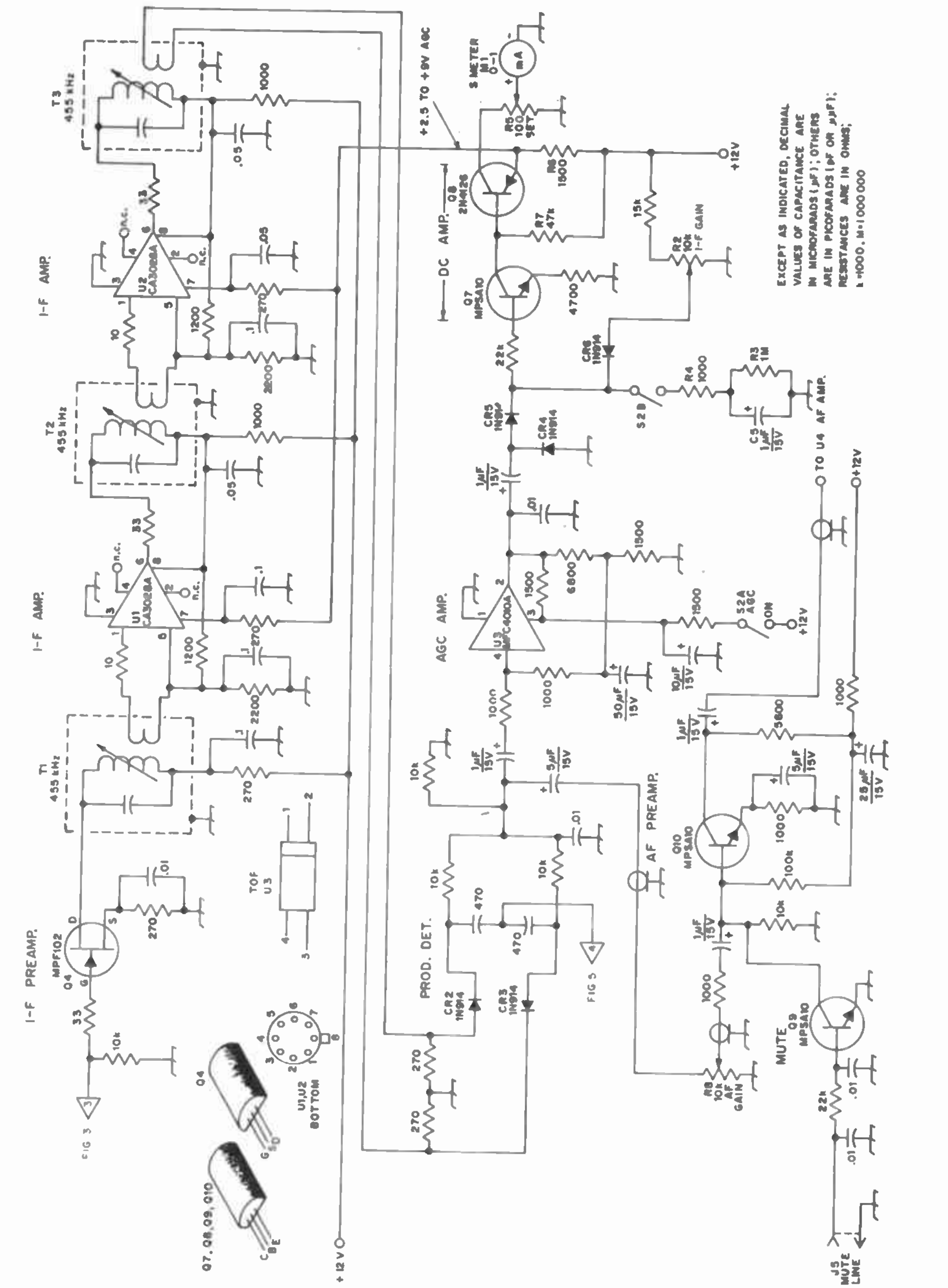


Fig. 3 — Schematic diagram of the i-f, agc, and audio preamplifier circuits. Capacitors are disk ceramic except those with polarity marked, which are electrolytic. Fixed-value resistors can be 1/4- or 1/2-watt composition unless otherwise noted. Numbered components not appearing in the parts list are so numbered for text discussion.

CR2-CR6, incl. — High-speed silicon switching diode, 1N3063, 1N914, or equiv.

J5 — Phono jack, single-hole mount.

M1 — 0 to 1-mA meter (Simpson No. 2121).

R2, R8 — 10,000-ohm linear-taper control.

R5 — 100-ohm pc-board-mount control (Mallory MTC-12L1 or equiv.).

S2 — Two-pole, single-throw toggle. Subminiature type used in this example.

T1-T3, incl.— Single-tuned miniature 455-kHz i-f transformer, 30,000-ohm primary to 500-ohm secondary (Radio Shack No. 273-1383). Use the black core at T1, yellow core at T2, and white core at T3.

U1, U2 — RCA integrated circuit.

U3 — Motorola integrated circuit.

i-f preamp (Q4). When the mode switch is placed in the lower sideband or upper sideband position, CR11 and CR12 become reverse biased, while CR9 and CR10 become forward biased, opening an rf path through FL3, and closing off the path through FL2. With this system, isolation between the filter inputs and outputs as well as between the filters is good, and since the mode switch carries dc only, no special precautions need be taken with the switching lead dress. If insufficient isolation exists between the filters, the characteristics of the narrower filter will be degraded by the wider filter. If this is the case, the use of two series-connected diodes in place of the single diodes on either side of the wider filter (and the possible addition of a small capacitor from the junction of those diodes to ground) should improve the isolation. However, no such degradation was observed when the circuit, as depicted in Fig. 4, was used. The values for C18 and the series combination of C19 and C20 are chosen to resonate with the inductance of the mechanical filter input and output transducers at 455 kHz.

An FET preamplifier stage follows the i-f filter assembly to compensate for the insertion loss of the filters, and establishes the noise figure of the i-f strip. An MPF102 was used at Q4 because of its low noise characteristics. A single-tuned 455-kHz i-f transformer is used to couple the output from Q4 to the input of U1.

Two stages of i-f amplification are provided by U1 and U2 (Fig. 3). RCA CA3028A integrated circuits were chosen for use in the i-f chain because they are inexpensive and easy to work with. In this circuit they are connected as differential amplifiers. Audio-derived agc is applied to terminal 7 of each IC (+2.5 to +9 volts), the constant-current-source bases. The dynamic range of the i-f system is approximately 60 dB.

A passive product detector was chosen over an active one because of its simplicity and good signal-handling capability. A pair of high-speed switching diodes (1N3063) were chosen because of their low cost and easy availibility.

BFO

An MPF102 JFET (Q6) functions in a Pierce crystal controlled BFO (Fig. 5). Separate crystals are used for cw, lower sideband, and upper sideband. Diode switching is used to select the proper crystal, with one section of the mode switch, S3C, performing the function. Again, lead dress to the mode switch is not critical, because only dc is being carried by those leads. The BFO crystal frequency chosen for cw operation is 454.3 kHz, 700 Hz below the center frequency of the i-f. This results in an audio beat note of 700 Hz when a cw signal is peaked in the passband. The lower sideband crystal frequency is 453.650 kHz, and the upper sideband crystal frequency is 456.350 kHz. Note the (suppressed) carrier frequency that the receiver is tuned to changes by 2.7 kHz when the mode switch is changed to the opposite sideband. BFO injection to the product detector is 7 volts peak to peak.

Audio-Derived AGC

Audio output from the product detector is split into two channels, one line feeding the agc strip and the other running to the audio amplifier circuit. An MFC4010A low-cost IC provides 60 dB of gain and serves as the agc amplifier (U3 of Fig. 3). Output from U3 is rectified by means of a voltage doubler consisting of two 1N914 diodes. Because of the high-gain capability of U3 it tends to be unstable at frequencies above the audio range. Addition of the .01-μF bypass capacitor

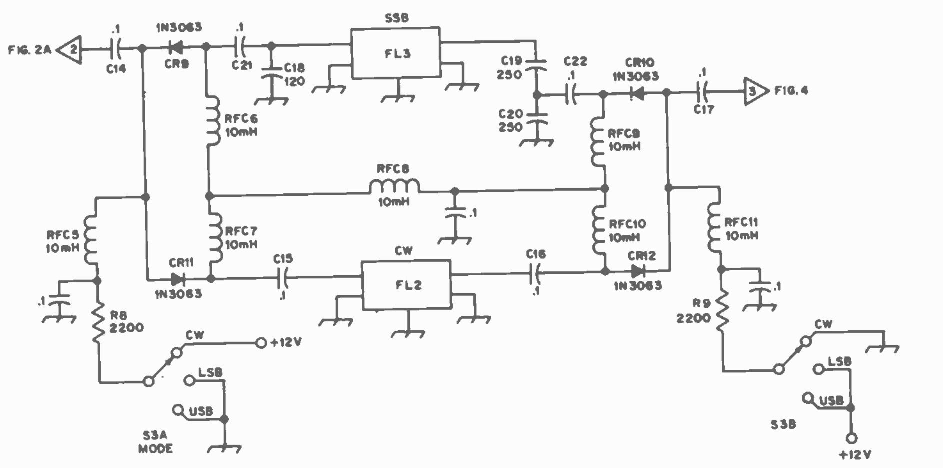


Fig. 4 — Circuit diagram of the i-f filter diode switching network. Numbered components not appearing in the parts list are so numbered for text reference purposes.

FL3 — 2.1-kHz bandwidth mechanical filter, 455-kHz center frequency (Collins F455FA21, Collins Radio Company, 4311 Jamboree Blvd., Newport Beach, CA 92663).

FL2 — 300-Hz bandwidth crystal filter, 455-kHz center frequency (Collins X455KF300, see *QST* Ham-Ads to obtain the names of suppliers).

RFC5 — RFC11 — incl. 10-mH miniature encapsulated inductor (Millen J302-10,000).

S3 — Four pole, three position rotary switch.

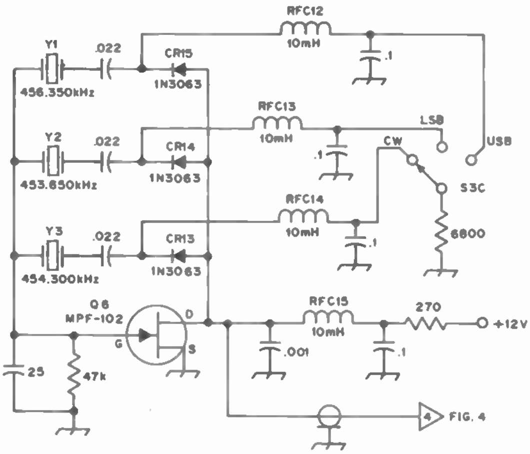


Fig. 5 — Diagram of the receiver beat frequency oscillator showing the use of diode switching of BFO crystals.

CR13—CR15, incl. — High-speed silicon switching diode, 1N3063, 1N914, or equiv.
RFC 12-15, incl. — 10-mH miniature encapsulated inductor (Millen J302- 10,000).
Y1 — 456.350-kHz crystal in HC-6/U holder (International Crystal Mfg. Co. type CS, 10 North Lee, Oklahoma City, OK 73102).
Y2 — 453.650-kHz crystal in HC-6/U holder (International type CS).
Y3 — 454.300-kHz crystal in HC-6/U holder (International type CS).

from terminal 2 to ground cured all signs of unstable operation in this circuit. Stubborn cases may require some additional bypassing at terminal 4 of U3. If so, use only that amount necessary to assure stability.

Rectified audio voltage from CR4 and CR5 is supplied to a two-transistor dc amplifier, Q7 and Q8. Agc voltage is taken from the emitter of Q8. Its amount varies with the incoming signal level, and changes as the current-caused voltage drop across the 1500-ohm emitter resistor, R6, shifts in value. S-meter M1 follows the same excursions in current at Q8.

Manual i-f gain control is possible by means of potentiometer R2. It supplies dc voltage to the base of Q7, thereby causing a voltage drop across R7, which causes Q8 to conduct more heavily. As a result, the voltage drop across R6 increases and reduces the agc voltage to lower the gain of the i-f system. The same action takes place during normal agc action. Diode CR6 acts as a gate to prevent the dc voltage provided by CR4 and CR5 from being disturbed by the presence of R2. Maximum i-f gain occurs when the arm of R2 is closest to ground. R3 and C5 establish the agc time constant. The value of R4 can be tailored to provide the attack-time characteristics one prefers. Slower or faster agc time constants can be obtained by changing the values of R3 and C5. The final values will be a matter of operator preference; no two people seem to agree on which time constant is best.

Audio System

Low-cost components are used in the audio system of Fig. 3 and Fig. 6. The circuit performs well and delivers undistorted af output up to one watt in level. An MPSA10 transistor is employed as an audio preamplifier. Muting is provided for by means of another MPSA10, Q9. A positive-polarity voltage is fed to the base of Q9 from the transmitter changeover system to saturate the muting transistor. When in the saturated mode, Q9 shorts out the base of Q10 to silence the receiver. The audio output circuit, U4 of Fig. 6, was borrowed from MFJ Enterprises and is that used in their 1-watt module, No. 1000. Those wishing to do so may order the assembly direct from MFJ.

Provisions are made for feeding a side-tone signal into terminal 3 of U4. This will permit monitoring one's sending even though the receiver is muted by means of Q9. U4 remains operative at all times.

HF-Band Converters

The same pattern is followed for the individual crystal-controlled converters used from 40 through 10 meters (Fig. 7). The 80-meter design is slightly different and is seen in Fig. 8. Separate converters were incorporated to eliminate the need for complicated band switching, and also to permit

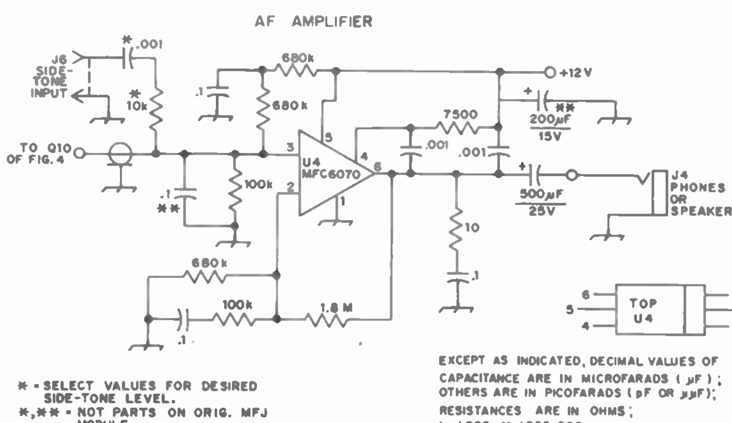
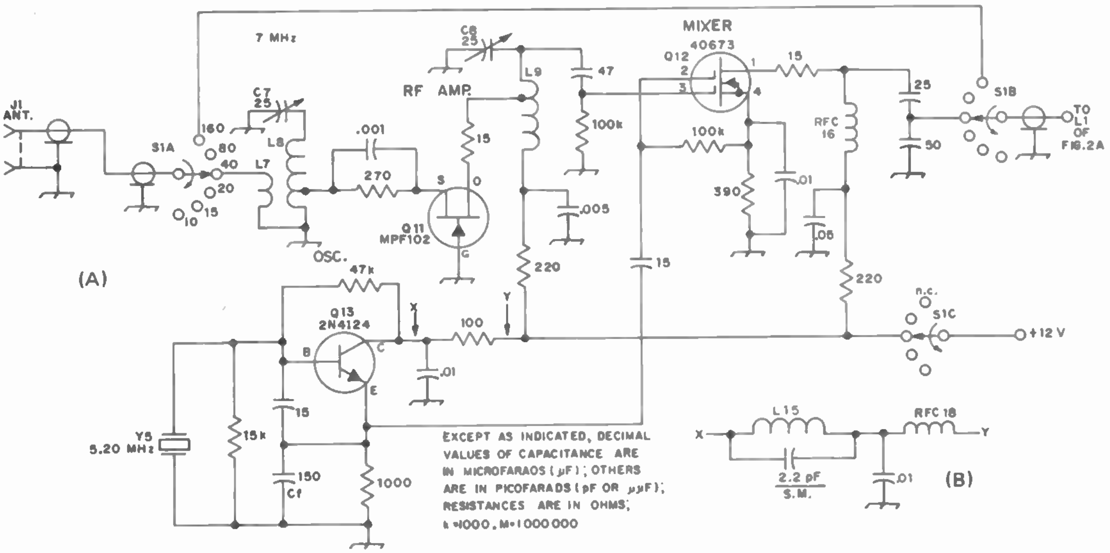


Fig. 6 — Schematic diagram of the MFJ Enterprises 1-watt audio module used in the receiver.

Fig. 7 — Schematic diagram of the 40-meter converter at A, with 10-meter oscillator modification at B. Capacitors are disc ceramic (fixed-value types). Resistors can be 1/4- or 1/2-watt composition types.

C7, C8 — Miniature ceramic or compression trimmer, 5- to 20-pF range.
CF — 39 pF for 20-, 15-, and 10-meter converters.
J1 — Coax connector of builder's choice.
L7 — 2 turns of No. 28 enam. wire over grounded end of L8.
L8 — 40 meters, 50 turns of No. 28 enam. on Amidon T-50-2 toroid core. Tap 8 turns above ground (13 µH, Qu = 180). 20 meters, 44 turns No. 28 enam. on Amidon T-50-6 toroid core. Tap 6 turns above ground (8 µH, Qu = 180). 15 meters, 25 turns No. 28 enam. on Amidon T-50-6 toroid core. Tap 4 turns above ground (4 µH, Qu = 150). 10 meters, 20 turns No. 28 enam. on Amidon T-50-6 toroid core. Tap 3 turns above ground (3.5 µH Qu = 150)
L9 — 40 meters, same as L8, but tap at 25 turns. 20 Meters, same as L8, but tap at 22 turns. 15 meters, same as L8, but tap at 12 turns. 10 meters, same as L8, but tap at 10 turns.
RFC16 — 500-µH miniature encapsulated inductor (Millen J302-500).
Y5 — International Crystal type CS crystal in type FM-1 holder; 40 meters, 5.2 MHz, fundamental mode; 20 meters, 12.2 MHz, fundamental mode; 15 meters, 19.2 MHz, fundamental mode; 10 meters (28.0- to 28.5-MHz coverage) 26.2 MHz, (28.5-to 29.0-MHz coverage) 26.7 MHz, (29.0-to 29.5-MHz coverage) 27.2 MHz, (29.5-to 30-MHz coverage) 27.7 MHz, all third overtone.
L15 — Pc mount slug-tuned coil, 1.5 µH nominal. Miller 46A156CPC or equiv.
RFC18 — 2.2 µH rf choke.



optimization of circuit values for each band of interest. The system used in this receiver calls for switching of only dc and 50-ohm circuitry. Low-impedance switching eliminates problems caused by long switch leads. Switching at high-impedance points, which is the usual technique in multiband receivers, can impair the quality of the tuned circuits and makes isolation of critical circuits more difficult.

A common-gate JFET rf amplifier provides 10 dB of gain in these converters and has good IMD and overload immunity. A 40673 MOSFET is used as the mixer in each converter. Output is taken at the i-f from a broadly resonant circuit formed by a 500-µH rf choke and an rf voltage divider which uses a series capacitor combination (25 and 50 pF). The divider provides a low-impedance pickoff point for the i-f output line to the tunable i-f receiver section.

The 40- through 15-meter converters employ simple Colpitts oscillators. A high-beta transistor is used for the oscillator. It has an f_T of approximately 200 MHz. The circuit for the 10-meter converter oscillator differs slightly from the others in that the HFO uses third overtone rather than fundamental mode crystals, necessitating the insertion of a collector tank circuit tuned to the overtone frequency.

A different design is used in the 80-meter converter, wherein a bandpass filter is used as the input fixed-tuned circuit. This technique was necessary to assure ample bandwidth from 3.5 to 4.0 MHz without the need to have a panel-mounted peaking control. The bandwidth is usable for an 80- and 75-meter frequency spread of 1 MHz.

A Pierce oscillator is used in the 80-meter front-end module to assure plenty of feedback for the 1700-kHz crystal.

Frequency Display Design Approach

The operation of the frequency display may be followed with the aid of the simplified block diagram, Fig. 9. With the conversion scheme utilized in the receiver, the received zero-beat frequency is equal to the sum of the VFO and HFO frequencies minus the BFO frequency, or in the special case of 160-meter operation, the difference between the VFO and BFO frequencies. Accordingly, the display has three identical input

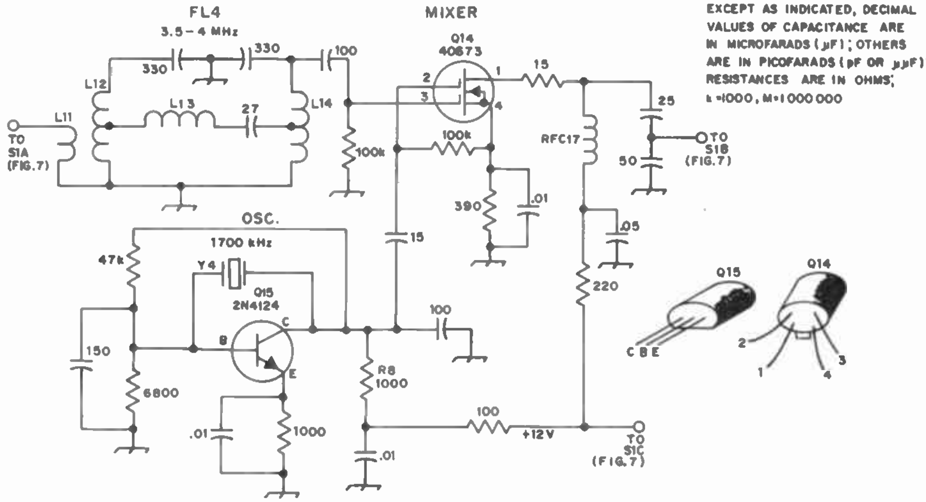


Fig. 8 — Schematic diagram of the 80-meter converter. Capacitors are disc ceramic. Resistors can be 1/4- or 1/2-watt composition types.

L11 — 4 turns No. 28 enam. wire over grounded end of L12.

L12 — 36 turns No. 28 enam. wire on Amidon T-50-2 toroid core (5.5 μH, *Q*u = 175). Tap at 18 turns.

L13 — 68-μH miniature rf choke (*Q*v of 50 or greater). Millen 34300-68 used in this example.

L14 — Same as L12.

RFC17 — 500-μH miniature encapsulated inductor (Millen J302-500).

Y4 — International type CS crystal in F-700 holder, 1.7 MHz.

networks – one for each oscillator – which amplify and square the incoming waveforms. The outputs of these preconditioning networks are sequentially gated into a chain of seven type SN74192 presettable up/down decade counters, U12-U18. The oscillator frequency addition and subtraction functions are performed by this counter chain. The events in a standard count sequence occur in this order: The counter chain is reset to zero and then placed in the count-up mode. The VFO signal is gated in and counted for 100 milliseconds. The counter is then placed in the count-down mode and the BFO signal is gated in and counted for 100 milliseconds (effectively subtracting the BFO frequency from the VFO frequency digitally). During the next 100 millisecond period, the counter is again placed in the count-up mode and the HFO signal is gated in and counted. At this point, the output of the counter chain ICs represents the receiver zero beat fre-

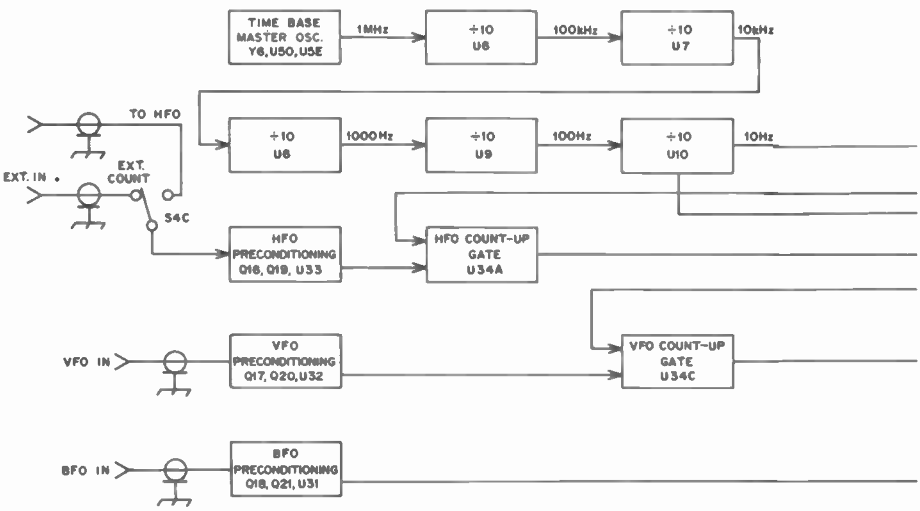
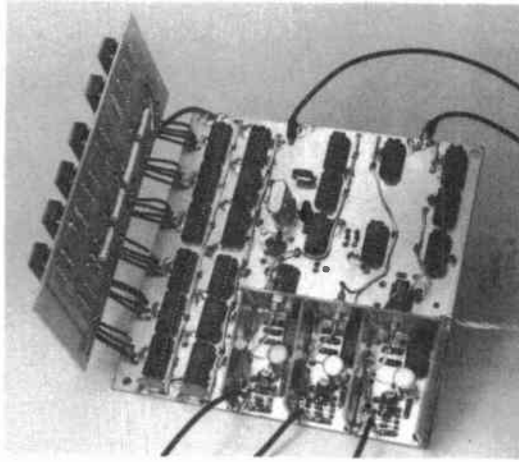


Fig. 9 — Simplified block diagram of the receiver frequency display.

quency. This output is in Binary Coded Decimal (BCD) form. The information from the last six SN74192s in the chain, U13-U18, is stored in six type SN7475 latches (in BCD form) and the entire counter chain is reset to zero during the ensuing 100 millisecond period. The standard count sequence is then repeated. Each of the latches is followed by a type SN7447 IC which decodes the BCD input and provides an output suitable for driving a seven-segment readout. Thus, the display is updated every 400 milliseconds. The output of U12 is not displayed, in order to avoid a distracting last-digit flicker. Accordingly, the 6-digit frequency display reads accurately to the nearest 100-Hz. The timing of the oscillator gating, the latch pulse, and the counter reset pulse is determined by a crystal-controlled oscillator/divider chain consisting of U5D, U5E, and U6-U11.

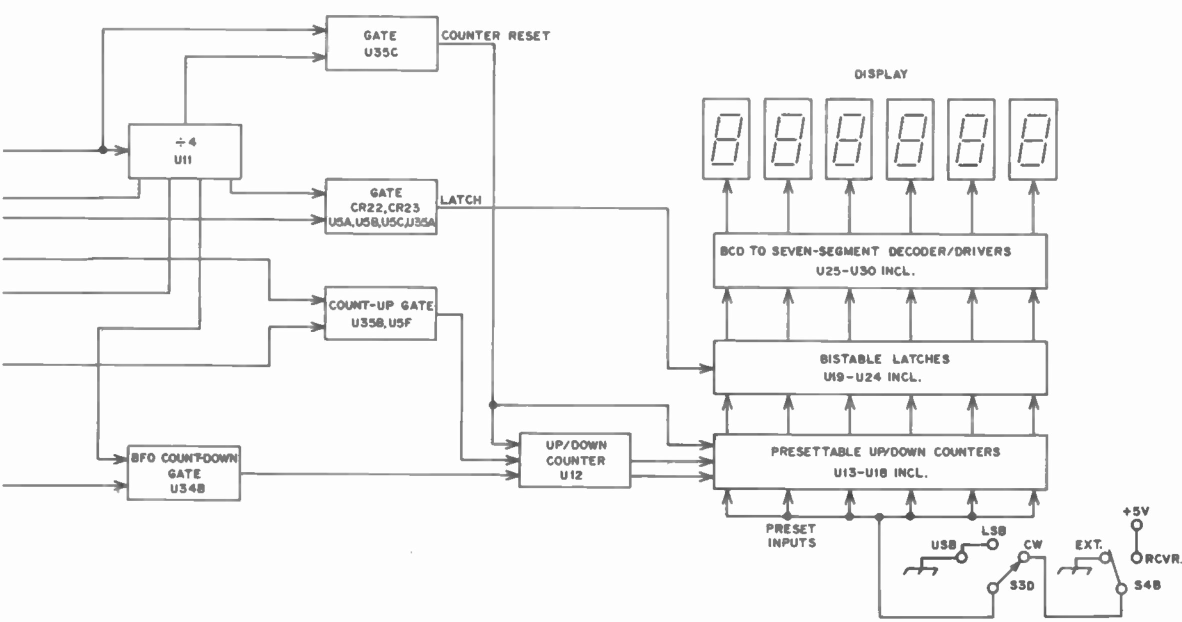
Although this system is satisfactory for ssb reception, it has the drawback that it is necessary for the operator to zero beat an incoming cw signal to read its exact frequency. By taking advantage of the presettable input feature of the SN74192 ICs, the display can be altered during cw reception to read the signal's zero beat frequency while the signal is peaked in the crystal filter passband. Inasmuch as the BFO frequency is 700 Hz below the passband center frequency, in order to obtain the desired readout it is necessary to start counting from "negative 700 Hz" rather than from zero by presetting the counter chain to 999930 every time a count sequence is begun.



Top view of frequency display circuit board. Visible in the foreground are the three shielded input wave-shaping networks.

Frequency Display Operation

Fig. 10 is a complete schematic diagram of the frequency display. The design of each of the input preconditioning networks is identical to that used by Blakeslee (*QST* for June, 1972, pages 31-32). In the HFO shaping network, protection against possible damage to Q16 caused by the application of too great an input voltage is accorded by CR16 and CR17, which conduct if the absolute value of the input signal exceeds approximately 0.6 volts. Q16 and Q19 form a two-stage amplifier that presents a high impedance to the input signal and a low impedance to the succeeding stage. Four sections of a type SN74HO4 high-speed TTL hex inverter are used to convert the incoming sinusoidal HFO signal to a square wave. U33F, operating as an amplifier, drives a Schmitt Trigger composed of U33E and U33D. U33C acts as an output buffer. A type SN7404 can be used in place of the SN74HO4 in the VFO and BFO shaping networks inasmuch as they operate at relatively low frequen-



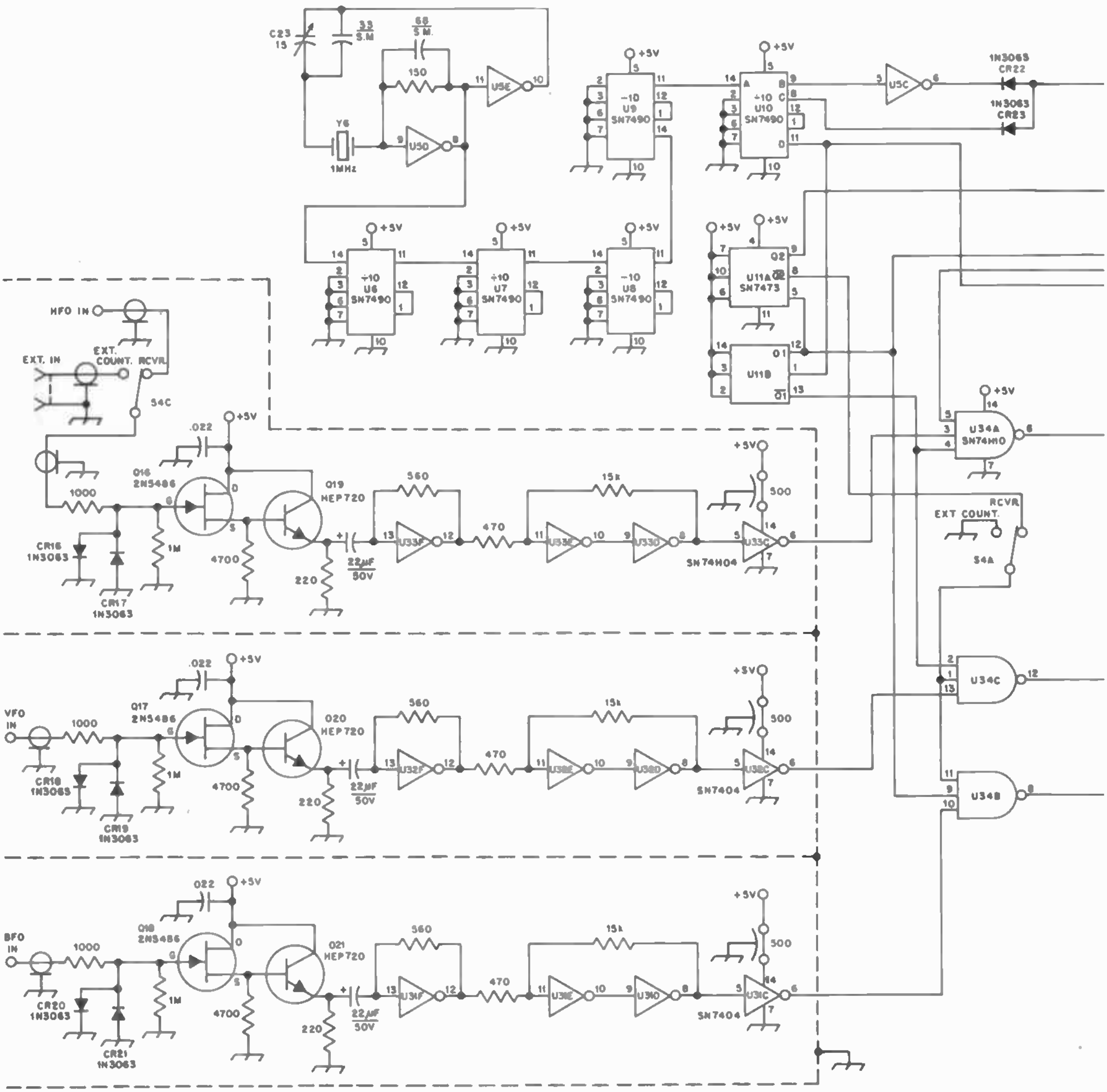
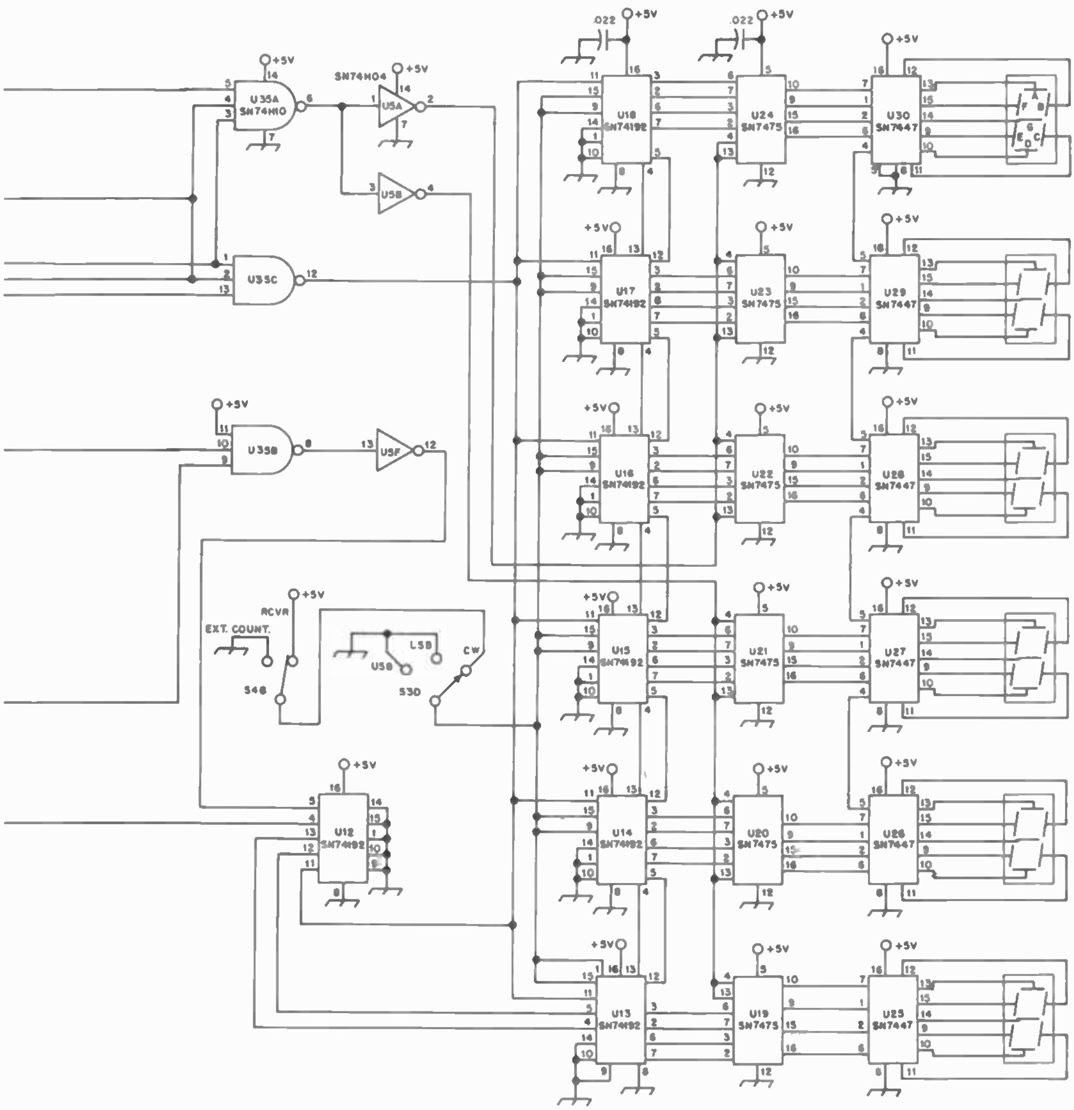


Fig. 10 – Schematic diagram of the receiver frequency display. U5-U30 are Texas Instruments SN7400 series TTL integrated circuits, or equiv.

cies. Each input network is connected to its appropriate oscillator through a short length of RG-174 miniature coaxial cable and a small value coupling capacitor. The smallest value of capacitance that will provide reliable counting should be employed. For example, 27 pF was used to couple from the VFO. Each heterodyne oscillator crystal will require its own coupling capacitor, and one of the receiver band-switch decks may be used to select the proper capacitor for each band. For use of the counter to display an external signal, such as a transmitter's frequency, the HFO input network is switched to receive signals from a front-panel mounted BNC jack, by means of S4C. The typical input circuit sensitivity is 50 to 100 mV.

The three preconditioning networks operate continuously and independently of each other. The shaped oscillator signals are gated sequentially to the input of the counter chain, each for a 100 millisecond interval. Both the duration of the interval and the order in which the receiver oscillators are sampled are governed by a crystal-controlled time base. Two sections of a type SN7404 hex inverter (U5D and U5E), a 1-MHz crystal (Y6), plus a handful of other parts constitute the master oscillator. C23 is a pc-mount air



variable capacitor that permits an incremental adjustment of the oscillator frequency.

The master oscillator is followed by a frequency divider network, U6-U11. U6-U10 are type SN7490 decade counters. Each SN7490 is composed of a divide-by-two section and a divide-by-five section. In this application, the sections are cascaded resulting in divide-by-ten operation. Four outputs are available from each SN7490 connected in this configuration, corresponding to a BCD representation of the number of input pulses previously applied to the IC, from zero to nine and then back to zero again. These outputs are labeled "A" (pin 12), "B" (pin 9), "C" (pin 8), and "D" (pin 11) representing the 2^0, 2^1, 2^2, and 2^3 bits respectively. For every ten pulses applied to the input (pin 14), one pulse appears at the "D" output, giving the effect of dividing the input frequency by ten (see Fig. 11). Five such divide-by-ten stages are cascaded in the time base resulting in division by 100,000 – with an input frequency of 1 MHz from the master oscillator, the frequency of the square wave appearing at the "D" output of U10 is 10 Hz. A slight shift of the master-oscillator frequency away from 1 MHz, due to perhaps temperature or voltage fluctuations, will also show up at U10, but divided by 100,000 – a good reason for starting with a high crystal frequency. It is obvious that with large-amplitude harmonic-rich square waves of several different frequencies present in the time base, the frequency display must be well shielded from the receiver in order to prevent the appearance of strong birdies all across the dial.

The output pulses from U10 are asymmetric – low for 80 milliseconds and high for 20 milliseconds. U11 is a type SN7473 dual flip-flop with both sections connected for divide-by-two operation and cascaded. Each time a negative-going pulse edge from U10 appears at the clock input of the first section of U11 (pin 1), the Q1 output (along with its complement, $\overline{Q1}$) changes state. Since U11 encounters a negative-going pulse edge once every 100 milliseconds, the Q1 output is alternately low for 100 milliseconds and high for 100 milliseconds. The Q1 output is used to establish the duration of the receiver-oscillator sampling interval.

Four distinct 100 millisecond intervals are necessary (VFO count, BFO count, HFO count, latch and reset). A means for distinguishing between these four intervals is provided by dividing the time base frequency (at Q1) by two in the second section of U11. Thus the Q2 output of U11 (along with its complement $\overline{Q2}$) is alternately low for 200 milliseconds and high for 200 milliseconds. The four distinct states occur with Q1 low and Q2 low, Q1 high and Q2 low, Q1 low and Q2 high, and Q1 high and Q2 high. An equivalent description of these states is that $\overline{Q1}$ and $\overline{Q2}$ are high, Q1 and $\overline{Q2}$ are high, $\overline{Q1}$ and Q2 are high, and Q1 and Q2 are high respectively. The sequential gating to the counter of the preconditioned VFO, BFO, and HFO signals is performed by U34, a type SN74H10 triple three-input NAND gate. The output of a three-input positive logic NAND gate is high under all input conditions except when all three inputs are in a high state simultaneously, when the output is forced low. The preconditioned VFO signal is applied continuously to one of the inputs of U34C. The other two inputs are tied to the $\overline{Q1}$ and $\overline{Q2}$ outputs respectively of U11. Whenever either $\overline{Q1}$ or $\overline{Q2}$ or both $\overline{Q1}$ and $\overline{Q2}$ are low, the output of U34C is held high. During the 100 millisecond long interval when both $\overline{Q1}$ and $\overline{Q2}$ are simultaneously high, however, the output of U34C follows the excursions of the VFO input, going low when the square wave VFO input is high and vice versa. Similarly, there are inputs to U34B from Q1, $\overline{Q2}$, and the BFO; to U34A from $\overline{Q1}$, Q2, and the HFO. In each case, when a signal is not being gated by a section of U34, the output of that section is held high.

The ICs used in the counter chain are SN74192 presettable up/down decade counters. Like the SN7490, each SN74192 has four terminals, "A" (pin 3), "B" (pin 2), "C" (pin 6), and "D" (pin 7) for BCD output. Additionally, the IC has "A" input, "B" input, "C" input", and "D input" terminals (pins 15, 1, 10, and 9 respectively) which may be used to initialize or preset the BCD output of the counter to some particular desired nonzero state from which point the count begins. For example, to load the preset inputs with the count of nine (the BCD representation of nine is 1001) the A and D inputs would be tied to +5 volts and the B and C inputs would be grounded. The information at the preset inputs is transferred to the output terminals when a negative-going pulse is applied to the counter reset or load terminal (pin 11), and as long as this terminal is held low, the counter is inhibited from counting pulses applied to either of its clock inputs. Each SN74192 has two clock inputs, one labeled count up (pin 5) and the other labeled count down (pin 4). When pin 4 is held at +5 volts and pulses are applied to pin 5, each incoming pulse increases the BCD output by 1 count. Likewise, when pin 5 is at a high logic level and pulses are applied to pin 4, each incoming pulse decreases the BCD output by 1 count. A single SN74192 can count from 0 to 9. If a greater number of pulses is to be counted, several of the ICs may be cascaded. Borrow and carry output terminals (pins 13 and 12 respectively) are provided for this purpose. To cascade several counters, the borrow output of the first IC is connected to the count down input of the next IC in line and

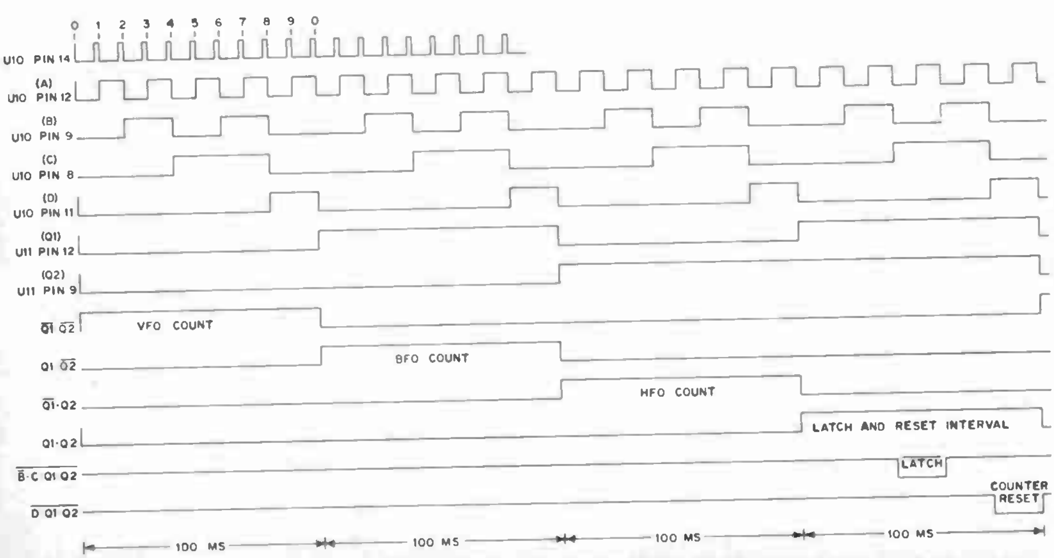


Fig. 11 — Chart showing waveforms associated with a standard count sequence.

the carry output is connected to the count up input of the next IC. In this configuration, if the output of a counter is nine and one pulse is applied to the count up input, the output of the first counter will go to zero and one pulse will be transferred to the count up input of the next counter in line. The case of the count down and borrow terminals is analogous.

In this frequency display, seven counters are cascaded with the result that a maximum of 9,999,999 counts can be registered before the chain resets to zero. If a 10-MHz signal was gated into the counter chain for 100 milliseconds, one million counts would be registered. The output of the last counter in the chain would be 1, while the other counter outputs would be zero. Thus, with a 100 millisecond sampling interval, the last counter indicates the tens of Megahertz, the counter immediately before indicates the units of Megahertz, the counter before that indicates the number of hundreds of kilohertz, and so on up to the first counter which indicates the number of tens of Hertz. Because of the plus-or-minus one count accuracy in the sampling process, which takes place three times in the making of each combined count, there is a tendency for the first counter in the chain to reflect this uncertainty by taking on several different nearby values even though the input frequency is kept constant. For this reason, only the outputs of U13-U18 are displayed on the seven-segment readouts (the display reads to the nearest 100 Hz rather than to the nearest 10 Hz). The maximum rated clock input frequency of the SN74192 is at least 20 MHz, but it is listed in the data books as typically 30 MHz, so it may be necessary to hand pick the first IC in the chain for proper high frequency operation.

The outputs of the VFO and HFO gates (U34C and U34A) are combined by U35B and U5F, and applied to the count up input of U12, while the output of the BFO gate (U34B) is applied directly to the count down input of U12, so that during the first 100 milliseconds of a count sequence the counter chain counts up the VFO frequency, during the second 100 ms period the BFO frequency is counted down, and during the third 100 ms period the HFO frequency is counted up. The result of this activity is that at the end of 300 ms, the outputs of the counter chain represent the receiver zero beat frequency. During the fourth 100 ms period, the outputs of the SN74192s are stored in six SN7475 latches, and then after 80 ms of this period have passed, the counter ICs are reset to zero (or to their preset values) by a 20 ms long reset pulse applied to pin 11. The entire sequence then repeats. The reset pulse is derived from the time base. U35C is one section of a SN74H10 triple three-input NAND gate. The inputs to U35C come from the "D" output of U10, and the Q1 and Q2 outputs of U11. By referring to Fig. 11, it can be seen that the only time that these three inputs are simultaneously high is during this 20 ms interval, forcing the output of U35C and the reset inputs of U12-U18 low.

When the receiver mode switch (S3) is in the cw position, the preset inputs of U13-U18 are set to 999930, while for ssb operation all of the preset inputs are set to 0. S4 chooses either receiver frequency display or external frequency count operation. When set to external, the counter preset inputs are set to 000000 regardless of the receiver mode switch position, the input to the HFO preconditioning network is removed from the HFO and connected to the front panel BNC jack, and U34B and U34C are disabled so that only the external signal source is counted up. The rest of the counter functions are unchanged for external operation.

The action of the SN7475 quad latches (U19-U24) is straightforward. Each SN7475 has four one-bit latches, each with an input and output terminal plus two clock lines (one for each pair of latches). When the clock line is at a high logic level, the logic level applied to each input will appear at its own output, and the output will follow any changes in the input. When the clock line is at a low logic level, the level that was present at each output at the time of the high to low transition of the clock pulse remains at that level regardless of the state of the input. In this application, the BCD output from each of the counter ICs, U13-U18, is applied to the inputs of a SN7475 latch. During the 300 ms long counting period, the clock line is kept at a low level. About midway into the fourth 100 ms period in the standard sequence, the clock line goes positive for 20 ms, transferring the BCD information present at the counter outputs to the latch outputs, and then the clock line goes negative again before the reset pulse occurs. The latching pulse is derived from the time base. Referring to Fig. 11, the pulse labeled $\overline{\text{LATCH}}$ is the output from U35A when the "C" output of U10, the Q1 and Q2 outputs of U11, and the complement of the "B" output of U10 are at a high logic level simultaneously. Note that in order to "synthesize" a four-input NAND gate from a three-input NAND gate, diodes CR22 and CR23 were added in front of one of the SN74H10 inputs. The LATCH pulse is obtained by inverting the $\overline{\text{LATCH}}$ pulse. U5A drives the clock lines of three of the SN7475s, while U5B drives the other three.

Each of the SN7475s drives a type SN7447 BCD to seven segment decoder/driver. The function of these ICs (U25-U30) is to convert the BCD code from the latches into a form suitable for driving seven segment display devices.

Construction Details

The completed receiver was assembled inside a "wrap around" style cabinet (LMB CO-1) measuring 6-1/2 × 14-1/2 × 13-inches (HWD). A homemade aluminum inner enclosure, attached to the front panel, was partitioned with an aluminum shield into two equal-size compartments – one side to house the receiver boards and the other to house the frequency display. The liberal use of perforated aluminum screening in the top and outside walls of the frequency counter compartment allows for adequate ventilation of that subsystem. No ventilation of the receiver section was deemed necessary as no heat-generating components are involved there.

The physical location of the front-panel controls was dictated as much by functional requirements as by operational considerations. The layout depicted in the photograph of the front panel is compact yet uncluttered. As far as possible, each individual control is placed near the circuit being controlled. A "commercial look" is imparted to the unit by the use of Kurz-Kasch series 1657 and 700 knobs. A Digbezel model 930-70 (Nobex Components, 1027 California Drive, Burlingame, CA 94010) filter/bezel assembly measuring approximately 5-1/4 X 1-3/4-inches is used to frame the seven-segment readouts. A red filter was chosen for this application, but filters are available for the bezel in amber, green, and neutral tints as well.

Printed circuit boards are used extensively in the construction of this receiver. The pattern used for interconnecting the twenty-five ICs on the main frequency-display board is sufficiently complex to warrant the use of photosensitive resist or silkscreening techniques in the preparation of the board. While the less complex layout of the receiver boards does not require the use of such methods of pc fabrication, the high quality of the results that may be obtained more than justify the slight additional costs involved. G-10 glass-epoxy board, 1/16-inch thick, with 1-oz. (per square foot) copper is used in all cases.

With the exception of U25 through U30 and the seven-segment displays, all of the frequency counter components are mounted on a 6 × 6-inch double-sided pc board. The pattern of interconnecting copper paths is etched on the underside of the board, while the top side is left as a solid ground plane, broken only at the places where component leads project through the board. If sockets are used, short bus-wire jumpers from the donut pads provided on the underside of the board to the top foil may be used to ground the necessary IC pins. Alternatively, individual component leads may be grounded by directly soldering them to the foil. A No. 65 drill was used for the IC and transistor pins, while the remaining holes were drilled with a No. 60 bit. The removal of copper foil from around the ungrounded leads is best accomplished by drilling those holes first, and then scraping away the top ground foil around each hole with a hand-held large-diameter drill bit (1-4-inch is satisfactory for the purpose). This process may be speeded up by the use of a drill press, although extreme caution must be exercised in order to prevent the drill bit from going completely through the board. The holes for grounded leads may then be drilled.

The liberal use of bypass capacitors in combination with a low-impedance ground return is vital in order to prevent clock pulses, switching transients, and other "garbage" from travelling along the dc supply lines to the sensitive counter input wave-shaping networks, resulting in improper operation. The configuration of the +5-volt dc supply bus resembles a five-toothed comb of jumper wires, laid across the board's upper side. The far end of each of the "teeth" is bypassed with a .022 μF disc ceramic capacitor. Shield compartments made from inch-high strips of double-sided pc board isolate the three input-shaping networks from each other and from the rest of the counter. Signal leads from the three receiver oscillators are carried to the input networks via short lengths of miniature coaxial cable (RG-174 or equiv.), and the dc supply line to each compartment passes through a .001-μF feedthrough capacitor.

Space is at a premium inside the input compartments, requiring the use of miniature vertical-mount electrolytics and 1/4-watt resistors. Low-profile sockets (Texas Instruments type C93) were used with all of the ICs on the counter board. The clock oscillator crystal, Y6, was soldered directly into the circuit with its case grounded to the top foil. A three-section, two-position rotary switch (S4), mounted with an "L" bracket to the chassis bottom to the rear of the counter board, serves as the external count/receiver count switch. Input signals from external sources are routed from the front-panel-mounted BNC jack to S4 via a length of RG-174. Standoff insulators, 1-1/2-inch high, are used to support the counter board at its corners.

The display board containing U25 through U30, as well as the seven-segment readout devices, is mounted vertically behind the front panel in such a manner that only the readouts are visible through the bezel. This board is supported partly by the stiffness of the leads that interface it with the main board, and partly by means of two No. 6 screws and metal spacers affixed to the panel. No circuit board pattern is supplied for the display board because of the wide variety of display devices and BCD decoder/driver ICs available to the builder at modest cost. LED readouts, fluorescent readouts and gas-discharge devices, to name a few, are competitive in price to the incandescent seven-segment displays that were used here, and each type has its own pin-numbering scheme and driver requirements, as well as physical mounting requirements.

The receiver is built from a number of subassemblies, with the principal one incorporating the better part of the tunable i-f. This 6 × 6-inch pc board contains the front-end mixer stage, the i-f filters and amplifiers, the product detector, and the BFO. Additionally, it supports the local oscillator module. Separate and smaller boards are used for the agc and audio circuits and each of the converters. A complete set of templates for the boards used in the receiver and in the frequency display is available from ARRL for $2 and a self-addressed, stamped business size envelope.

Physically small components are employed wherever possible, resulting in compact board layouts. Vertical format low voltage electrolytics such as the Sprague 503D series are to be preferred over axial lead types, which must be mounted upright to fit on the board. Sprague Hypercon low-voltage ceramic disc capacitors are ideally suited due to their miniature size for use wherever a 0.1- or .01-μF value is called for. Millen J302 series encapsulated inductors are used liberally. They are compact and are designed for high-density pc mounting. Either quarter-watt or half-watt resistors can be used, although with the larger size it may be necessary to stand some of the resistors on end. T1-T3 are inexpensive imported 455-kHz i-f transformers that come four to a package. International Crystal type F-605 nylon sockets were used for Y1-Y3, which are not available with wire leads. All of the semiconductor devices used in the receiver are soldered directly into the circuit board, although sockets are available for U1 and U2 if desired.

An outboard power supply capable of providing 12 volts at 150 mA maximum for the receiver and 5.0 volts at 1.5A for the frequency display is required. It is adviseable to use separate receiver and display supplies rather than a common transformer and rectifier feeding two voltage regulators in order to minimize the chances of coupling counter "hash" into the receiver via the +12-volt supply line. A front-panel-mounted miniature toggle switch is used to apply 117-v ac to the remote supply transformer primaries. Proper bypassing of all supply leads entering the cabinet will aid in rejection of extraneous signals. A loudspeaker may be installed in the same cabinet as the power supply for operating convenience.

The local oscillator enclosure is made from four pieces of double-sided circuit board material soldered together at the edges. The assembly measures approximately 2-1/4 × 1-3/4 × 4-1/8-inches (HWD), and is bolted down to the main receiver board by four spade lugs. The main tuning capacitor, C2, is first attached to one of the side walls of the compartment with the two specially threaded machine screws provided with the capacitor. A 3/8-inch diameter hole drilled in the front compartment wall provides adequate clearance for the capacitor shaft. L6 is mounted on the same wall as C2, centered in the enclosure and directly above the VFO printed-circuit board which is affixed by soldering its ground foil along three edges to the enclosure walls. A 10:1 reduction epicyclic drive. Jackson Brothers number 5857 (available from M. Swedgal, 258 Broadway, New York, NY 10007) is used to turn the main-tuning capacitor shaft. This drive unit requires 1-1/8-inch clearance behind the front panel. A one-piece nickel-plated brass coupling is used between the drive and the capacitor shaft. Four 2-inch long No. 6 machine screws are used to support the large receiver board, which must be installed at the height that permits precise alignment between these shafts.

The main receiver board and the agc/audio preamp boards are fabricated from double-sided copper clad material. All of the other receiver boards use single-sided stock. The agc/audio preamp board and the MFJ audio board (available from MFJ Enterprises, P.O. Box 494, Mississippi State, MS 39762) are mounted with No. 6 hardware and half-inch spacers on one of the aluminum enclosure walls. The input bandpass filter, FL1, is enclosed in an alumimum Minibox measuring 1-1/8 × 2-1/8 × 3-1/4-inches HWD (Bud CU-2117A) which is mounted to the opposite enclosure wall. All interconnections between boards are made with RG-174 miniature coaxial cable. Spade lugs support the converter boards from the rear wall of the receiver compartment.

Alignment

The only adjustment that should be necessary with the frequency display is the setting of the time base crystal oscillator frequency to 1 MHz. A clip lead from the antenna input jack of a general-coverage receiver loosely coupled to the oscillator will pick up enough harmonic energy to allow for precise zero beating of the oscillator frequency to WWV at 10 MHz. A nonmetallic alignment tool should be used to turn the shaft of C23, which should provide a wide enough incremental frequency adjustment.

The 160-meter tunable i-f should be adjusted before the converters are checked out. After verifying with an oscilloscope or a general-coverage receiver that the VFO is indeed oscillating, the VFO tuning range may be set properly with the aid of the frequency display (set to external count). With the plates of C2 fully meshed, the core of L6 should be moved in or out until the frequency display reads approximately 2240 kHz. Then note the displayed frequency with C2 rotated to minimum capacitance, which should be close to 2770 kHz. If the upper frequency obtained departs significantly from the above reading, the slug of L6 should be moved to a point that provides approximately equal amounts of overlap on either side of the desired 2255-2755 kHz range.

The BFO should be checked for proper operation. At this point, the display may be connected in the receiver count mode. If all is well it will indicate a received frequency between 1800 and 2300 kHz. With a signal generator set to 1800 kHz connected to the input of FL1 and the plates of C1 fully meshed, a beat note should be heard in the speaker or headphones as the receiver is tuned to the signal generator frequency. L2 and L3 may now be adjusted to peak the signal on the S-meter or in the loudspeaker. If these coils are not carefully tuned the two filter sections will not track well, resulting in an undesirable broad response or perhaps even a double peak. Now, tune T1, T2, and T3 for maximum signal output. Spot checks across the 1800 to 2300 kHz range can be made to assure that the input bandpass filter is optimized and is tunable across the entire frequency spread. Check to see that the mode switch selects both the proper crystals in the BFO and the correct i-f filter. All that remains to be adjusted now is the S-meter control, R5. With the agc on, but with the i-f gain (R2) set at minimum sensitivity, adjust R5 to give full-scale deflection of M1. This procedure will complete the tune-up of the main portion of the receiver.

Checkout of the converters is similarly easy. The trimmers should be adjusted for peak response at the center of each frequency band of interest. A signal generator is useful for this procedure, although on-the-air signals, preferably weak ones, will suffice.

Alignment of the receiver is complete at this point. The frequency display should be operative and accurate on each band.

VHF and UHF Receiving Techniques

Adequate receiving capability is essential in vhf and uhf communication, whether the station is a transceiver or a combination of separate transmitting and receiving units, and regardless of the modulation system used. Transceivers and fm receivers are treated separately in this *Handbook*, but their performance involves basic principles that apply to all receivers for frequencies above 30 MHz. Important attributes are good signal-to-noise ratio (low noise figure), adequate gain, stability, and freedom from overloading and other spurious responses.

Except where a transceiver is used, the vhf station often has a communications receiver for lower bands, with a crystal-controlled converter for the vhf band in question ahead of it. The receiver serves as a tunable i-f system, complete with detector, noise limiter, BFO and audio amplifier. Unless one enjoys work with communications receivers, there may be little point in building this part of the station. Thus our concern here will be mainly with converter design and construction.

Choice of a suitable communications receiver for use with converters should not be made lightly, however. Several degrees of selectivity are desirable: 500 Hz or less for cw, 2 to 3 kHz for ssb, 4 to 8 kHz for a-m phone and 12 to 36 kHz for fm phone are useful. The special requirements of fm phone are discussed in Chapter 14. Good mechanical design and frequency stability are important. Image rejection should be high in the range tuned for the converter output. This may rule out 28 MHz with receivers of the single-conversion type having 455-kHz i-f systems.

Broad-band receiving gear of the surplus variety is a poor investment at any price, unless one is interested only in local work. The superregenerative receiver, though simple to build and economical to use, is inherently lacking in selectivity. With this general information in mind, this section will cover vhf and uhf receiver "front ends" stage by stage.

RF AMPLIFIERS

Signal-to-Noise Ratio: Noise of one kind or another limits the ability of any receiving system to provide readable signals, in the absence of other kinds of interference. The noise problem varies greatly with frequency of reception. In the hf range man-made, galactic and atmospheric noise picked up by the antenna and amplified by all stages of the receiver exceeds noise generated in the receiver itself. Thus the noise figure of the receiver is not of major importance in weak-signal reception, up to at least 30 MHz.

At 50 MHz, external noise still overrides receiver noise in any well-designed system, even in a supposedly "quiet" location. The ratio of external to internal noise then drops rapidly with increasing signal frequency. Above 100 MHz or so external noise other than man-made is seldom a problem in weak-signal reception. Noise characteristics of transistors and tubes thus become very important in receivers for 144 MHz and higher bands, and circuit design and adjustment are more critical than on lower frequencies.

The noise figure of receivers using rf amplifiers is determined mainly by the first stage, so solving the internal-noise problem is fairly simple. Subsequent stages can be designed for selectivity, freedom from overloading, and rejection of spurious signals, when a good rf amplifier is used.

Gain: It might seem that the more gain an rf amplifier has, the better the reception, but this is not necessarily true. The primary function of an rf amplifier in a vhf receiver is to establish the noise figure of the system; that is, to override noise generated in later stages. One good rf stage is usually enough, and two is the usual maximum requirement.

Once the system noise figure is established, any further gain required may be more readily obtained in the intermediate frequency stages, or even in the audio amplifier. Using the minimum rf gain needed to set the overall noise figure of the receiver is helpful in avoiding overloading and spurious responses in later circuits. For more on rf gain requirements, see the following section on mixers.

Stability: Neutralization or unilaterialization (see chapter on semiconductors) may be required in rf amplifiers, except where the grounded-gate circuit or its tube equivalent is used. Amplifier neutralization is accomplished by feeding the energy from the output circuit back into the input, in such amount and phase as to cancel out the effects of device capacitance and other unwanted input-output coupling that might cause oscillation or other regenerative effects. Inductive neutralization is shown in Fig. 9-1B and C. Capacitive arrangements are also usable. Examples of both will be seen later in this chapter.

An rf amplifier may not actually oscillate if operated without neutralization, but noise figure and bandwidth of the amplifier may be better with

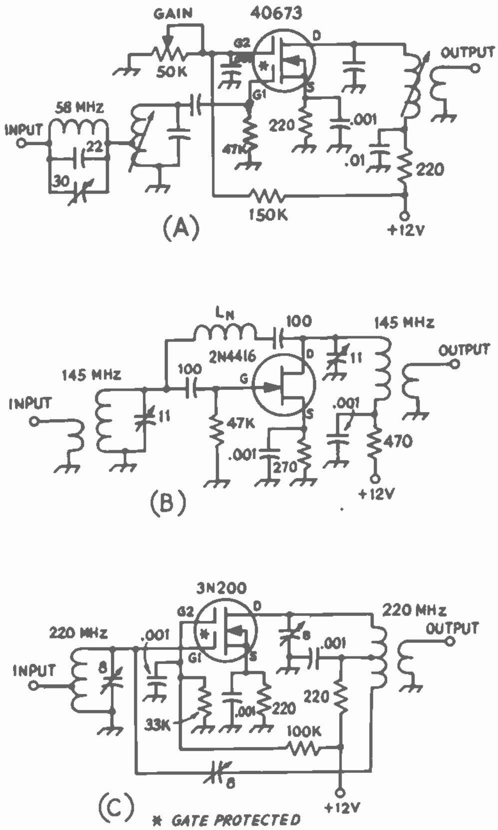


Fig. 9-1 – Typical grounded-source rf amplifiers. The dual-gate MOSFET,A, is useful below 500 MHz. The junction FET,B, and neutralized MOSFET,C, work well on all vhf bands. Except where given, component values depend on frequency.

it. Any neutralization adjustment reacts on the tuned circuits of the stage, so the process is a repetitive cut-and-try one. The objective should be greatest margin of signal over noise, rather than maximum gain without oscillation. A noise generator is a great aid in neutralization, but a weak signal can be used if the job is done with care.

Overloading and Spurious Signals: Except when some bipolar transistors are used, the rf amplifier is not normally a major contributor to overloading problems in vhf receivers, though excessive rf gain can cause the mixer to overload more readily. Overloading is usually a matter of mixer design, with either transistors or tubes. Images and other spurious responses to out-of-band signals can be kept down by the use of double-tuned circuits between the rf and mixer stages, and in the rf amplifier input circuit. In extreme cases, such as operation near to fm or TV stations, coaxial or other high-*Q* input circuits are helpful in rejecting unwanted frequencies.

Using RF Preamplifiers

It is important to design the front-end stages of a vhf receiver for optimum performance, but we often want to improve reception with equipment already built. Thousands of fm receivers formerly in commercial service, now revamped for amateur work in the 50-, 144- and 420-MHz bands, were built before modern low-noise tubes and transistors were available. Though otherwise useful, these receivers have excessively-high noise figure. Many other commercial and home-built vhf converters and receivers are also not as sensitive as they might be.

Though it would be better to replace the rf stages of such equipment with more modern devices, the simpler approach is usually to add an outboard rf amplifier using a low-noise tube or transistor. In the fm example, the quieting level of some receivers can be improved by as much as 10 dB by addition of a simple transistor amplifier. Similar improvement in noise figure of some receivers for other modes is also possible; particularly band-switching communications receivers that have vhf coverage.

Common circuits for rf preamplifier service are shown in Figs. 9-1, 2 and 3. Examples of amplifier construction are given later in this chapter. Circuits shown in the vhf converters described can also be adapted to preamplifier service.

Circuit discussion is cumbersome if we use strictly-correct terms for all tube and transistor amplifiers, so tube terminology will be used here for simplification. The reader is asked to remember

Amplifier Circuitry

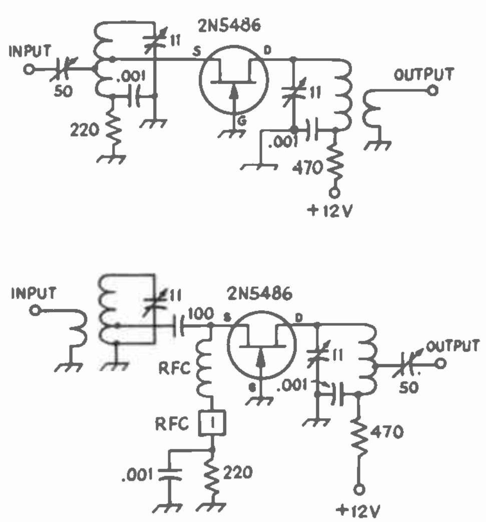


Fig. 9-2 – Grounded-gate FET preamplifier tends to have lower gain and broader frequency response than other amplifiers described.

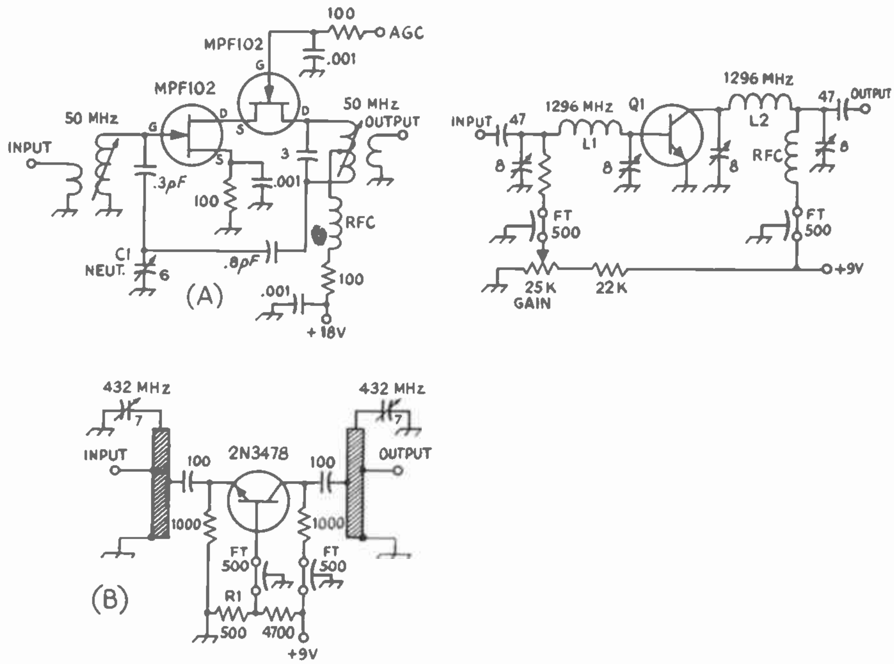


Fig. 9-3 — (A) Cascode amplifier circuit combines grounded-source and grounded-gate stages, for high gain and low noise figure. Though JFETs are shown, the cascode principle is usable with MOSFETs as well. (B and C) Examples of uhf preamplifier construction using bipolar transistors.

that "gate" may also imply "base" for bipolar transistors, or "grid" for tubes. "Source" should be read as "emitter" for the bipolar, and as "cathode" for the FET.

Amplifiers may be the grounded-source type, Fig. 9-1; grounded-gate, 9-2; or a combination of both, 9-3. The dual-gate MOSFET circuit, 9-1A, works well up to 300 MHz, but JFET and bipolar devices are superior for 420 MHz and higher. The gain and noise figure of a dual-gate MOSFET are adequate at 300 MHz, and it is simple and readily adapted to automatic gain control.

Triode tubes and FET transistors usually require neutralization for optimum noise figure with the grounded-cathode circuit. Inductive neutralization is shown in Fig. 9-1B, and the capacitive method shown at C works equally well. Examples will be seen later in this chapter. The 58-MHz trap circuit in Fig. 9-1A is discussed in the following section on mixers.

An alternative to neutralization lies in use of the grounded-gate circuit, Fig. 9-2. Its stage gain is lower and its bandwidth generally greater than with the grounded-cathode circuit. The input impedance is low, and the input circuit is tapped to provide a proper impedance match. A broad-band amplifier may be made with a low-impedance line connected directly to the input element, if selectivity is not required at this point for other reasons. Tubes designed for grounded-grid service include the 417A/5842, 416B, 7768 and the various "lighthouse" types, though almost any triode or triode-connected tetrode can be used. JFETs work well in grounded-gate circuits. In the grounded-grid amplifier, the tube heater becomes effectively a part of the tuned circuit, so some form of high-current rf choke is required. Ferrite-bead chokes work well.

The cascode circuit, Fig. 9-3, combined grounded-source and grounded-gate stages, securing some of the advantages of both. Fig. 9-3B shows a grounded-base bipolar transistor amplifer. The value of R1 should be chosen experimentally to achieve best sensitivity.

Front-End Protection

The first amplifier of a receiver is susceptible to damage or complete burnout through application of excessive voltage to its input element by way of the antenna. This can be the result of lightning discharges (not necessarily in the immediate vicinity), rf leakage from the station transmitter through a faulty send-receive relay or switch, or rf power from a nearby transmitter and antenna system. Bipolar transistors often used in low-noise uhf amplifiers are particularly sensitive to this trouble. The degradation may be gradual, going unnoticed until the receiving sensitivity has become very poor.

No equipment is likely to survive a direct hit from lightning, but casual damage can be prevented by connecting diodes back-to-back across the input circuit. Either germanium or silicon vhf diodes can be used. Both have thresholds of conduction well above any normal signal level, about 0.2 volt for germanium and 0.6 volt for silicon. The diodes used should have fast switching times. Computer diodes such as the 1N914 and hot-carrier types are suitable. A check on weak-signal reception should be made before and after connection of the diodes.

RF SELECTIVITY

The weakest point in any vhf or uhf receiver is the front-end circuit. Solid-state devices with high sensitivity, wide dynamic range and freedom from overload are now available. Thus, the quality of a front-end circuit is usually determined by how the active devices are used and the degree of rf selectivity included. High selectivity at vhf and uhf is not easy to achieve. Many lumped-constant tuned circuits are needed for even a moderate degree of selectivity at the signal frequency. Several tuned circuits before the first active stage (rf amplifier or mixer) will have sufficient loss to limit the sensitivity of the receiver. If lumped-constant circuits are employed, rf amplifiers can be interspaced between the *LC* elements to make up losses. High gain is not needed or desirable, so FETs operated grounded-gate are preferred.

For improved rf selectivity a helical resonator, a device which consists of a shield and a coil may be employed. One end of the coil is attached to the shield, as shown in Fig. 9-4, and the other end is open-circuited, except for a tuning capacitor. Helical resonators are electrically equivalent to a

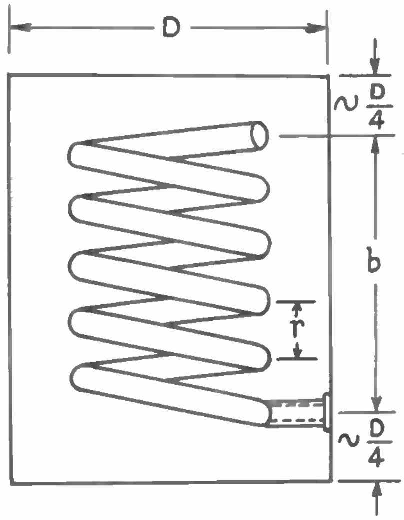


Fig. 9-4 — Outline sketch of resonator.

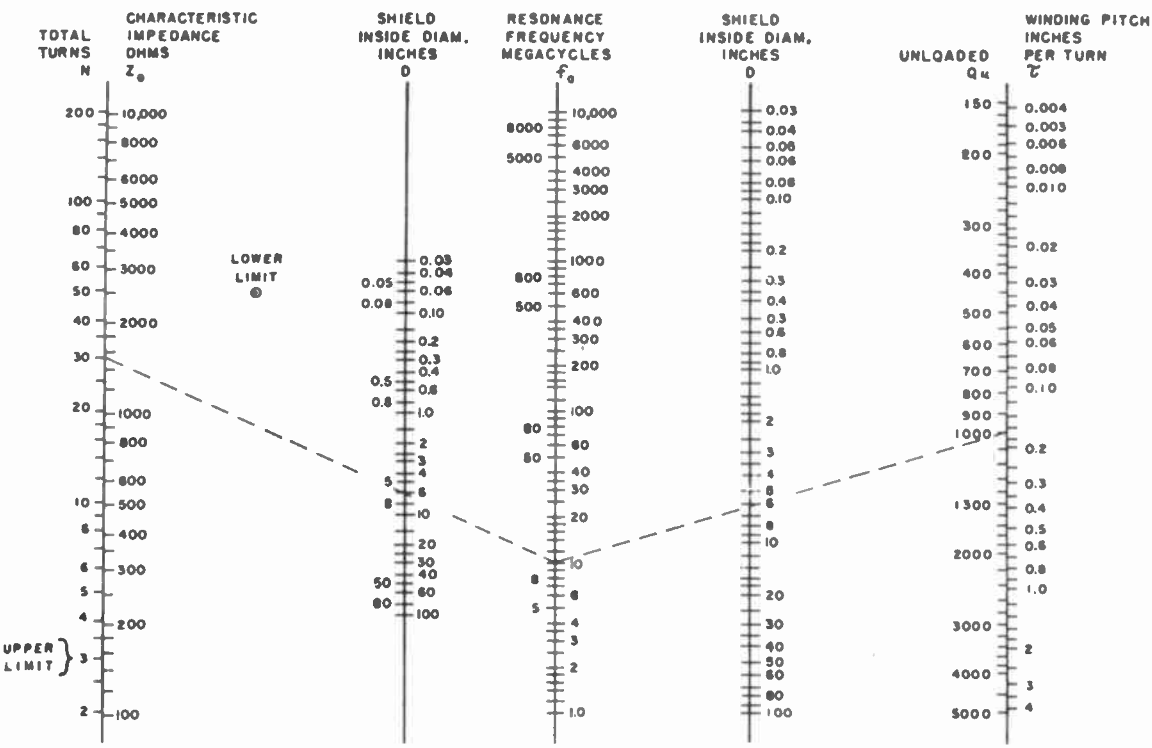


Fig. 9-5 — Design chart for quarter-wave helical resonators.

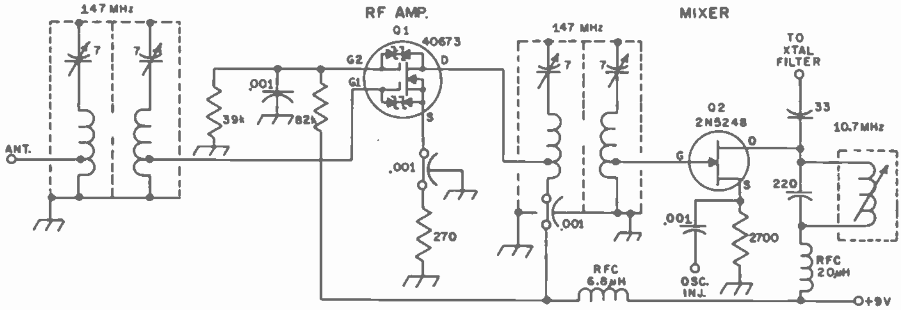


Fig. 9-6 – Schematic diagram of the Johnson 504 front-end circuit.

quarter-wave transmission-line resonator but are physically much smallter. Resonators can be built exhibiting *Q* of 1000 or more at vhf and uhf. Because the *Q* is so high, front-end circuits can be designed using helical resonators which provide a high degree of selectivity without high losses, at least a low and moderate power levels.

The inductance element in a helical resonator should be made as large as possible and capacitance kept to a minimum for best performance. Probe, tap or aperture coupling may be employed. The basic form of a helical resonator is shown in Fig. 9-4. A low-loss air-insulated trimmer or disk plunger may be used to tune the resonator. The capacitor must be much higher *Q* than the resonator to be useable. The usual precautions for fabricating high-Q coils must be observed when building a helical resonator. A protective silver plating is recommended for the coil and shield for units to be used above 100 MHz. The shield should be seamless and all joints should be effectively soldered to keep resistance to a minimum. The coil and shield should be made using heavy stock to assure mechanical stability.

Fig. 9-5 can be used to obtain approximate design information accurate to plus or minus ten percent. Complete design equations for helical resonators are beyond the scope of this text, but they may be found in Macapline and Schildknecht, "Coaxial Resonators with Helical Inner Conductor," *Proceeding of the IRE*,December, 1959.

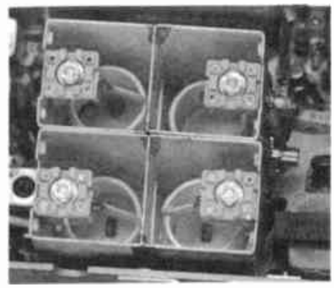


Fig. 9-7 – Close-up view of the helical resonators with the covers removed. The rf amplifier stage is constructed on the outside wall of the upper-right-hand resonator. Details are given in the text.

An application of helical resonators in a 146-MHz front-end circuit is shown in Figs. 9-6 and 9-7. This circuit is used in the Johnson 504 transceiver. The helical resonators consist of 5-3/4 turns of No. 12 wire contained in a rectangular 1 × 1 × 2-inch cavity. Both the coil and enclosure are silver plated. The coil is 5/8 inch inside diameter and 5/8 inch long, tuned with a 7-pF miniature air-variable capacitor. The 50-ohm input tap is at 1/4 turn from the ground end of the coil, an indication of the high impedance achieved. Coupling between individual resonators is through a 1/2 × 1/4 -inch aperture, or "window." Layout details can be seen in Fig. 9-7.

MIXERS

Conversion of the received energy to a lower frequency, so that it can be amplified more efficiently than would be possible at the signal frequency, is a basic principle of the superheterodyne receiver. The stage in which this is done may be called a "converter," or "frequency converter," but we will use the more common term, *mixer*, to avoid confusion with *converter*, as applied to a complete vhf receiving accessory. Mixers perform similar functions in both transmitting and receiving circuits, and mixer theory and practice are treated in considerable detail elsewhere in this *Handbook.*

A receiver for 50 MHz or higher usually has at least two such stages; one in the vhf or uhf converter, and usually two or more in the communications receiver that follows it. We are concerned here with the first mixer.

Diode Mixer: There are many types of mixers, the simplest being merely a diode with the signal and energy on the heterodyning frequency fed into it, somewhat in the manner of the 1296-MHz example, Fig. 9-8A. The mixer output includes both the sum and difference frequencies. Either can be used, but in this application it is the difference, since we are interested in going lower in frequency.

With a good uhf diode in a suitable circuit, a diode mixer can have a fairly low noise figure, and this is almost independent of trequencv, well into the microwave region. The effectiveness of most active mixers falls off rapidly above 400 MHz, so the diode mixer is almost standard practice in amateur microwave communication. All diode mixers have some conversion *loss*. This must be added to the noise figure of the i-f amplifier following, to determine the overall system noise figure. Low-noise design in the first i-f stage is thus mandatory, for good weak-signal reception with a diode mixer having no rf amplifier preceding it. Purity of the heterodyning energy and the level of injection to the mixer are other factors in the performance of diode mixers.

Balanced mixers using hot-carrier diodes are capable of noise figures 1 to 2 dB lower than the best point-contact diodes. Hot-carrier diodes are normally quite uniform, so tedious selection of matched pairs (necessary with other types of diodes) is eliminated. They are also rugged, and superior in the matter of overloading.

The i-f impedance of a balanced hot-carrier diode mixer (Fig. 9-8B) is on the order of 90 ohms, when the oscillator injection is about one milliwatt. Thus the mixer and a transistorized i-f amplifier can be separated physically, and connected by means of 93-ohm coax, without an output transformer.

Conversion loss, around 7 dB, must be added to the noise figure of the i-f system to determine the overall system noise figure. Unless a low-noise preamplifier is used ahead of it, a communications receiver may have a noise figure of about 10 dB, resulting in an overall noise figure of 17 dB or worse for a vhf system with any diode mixer. A good i-f preamplifier could bring the receiver noise figure down to 2 dB or even less, but the system noise figure would still be about 9 dB; too high for good reception.

An amplifier at the signal frequency is thus seen to be required, regardless of mixer design, for optimum reception above 50 MHz. The rf gain, to override noise in the rest of the receiver, should be greater than the sum of noise figures of the mixer and the i-f system. Since the noise figure of the better rf amplifiers will be around 3 dB, the gain should be at least 20 dB for the first example in the previous paragraph, and 12 dB for the second.

Tube and Transistor Mixers: Any mixer is prone to overloading and spurious responses, so a prime design objective should be to minimize these problems. FET mixers have become standard practice at vhf. JFETs are slightly better than MOSFETs, although the junction types require

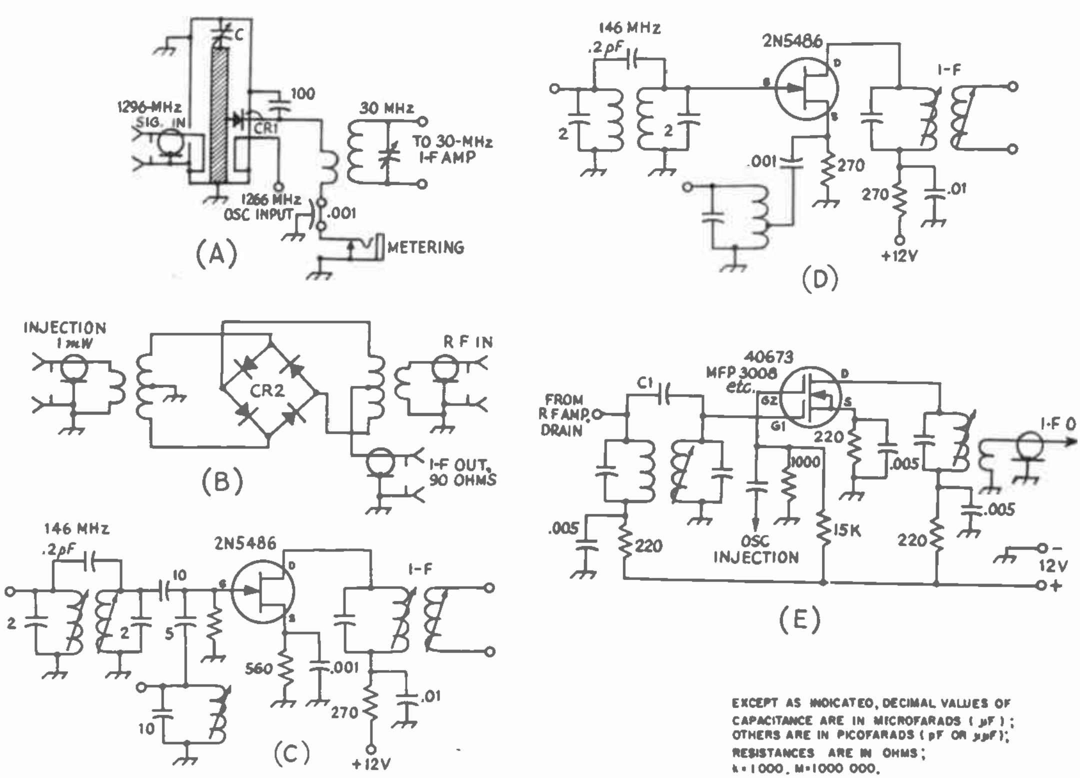


Fig. 9-8 — Vhf and uhf mixer circuits. A diode mixer for 1296 MHz, with a coaxial circuit for the signal frequency, is shown in A. CR1 is a uhf diode, such as the 1N21 series. A balanced mixer, as in B, gives improved rejection of the signal and injection frequencies. If hot-carrier diodes are used for CR2, sorting for matched characteristics is eliminated. Gate and source injection of a JFET mixer are shown at C and D, respectively.

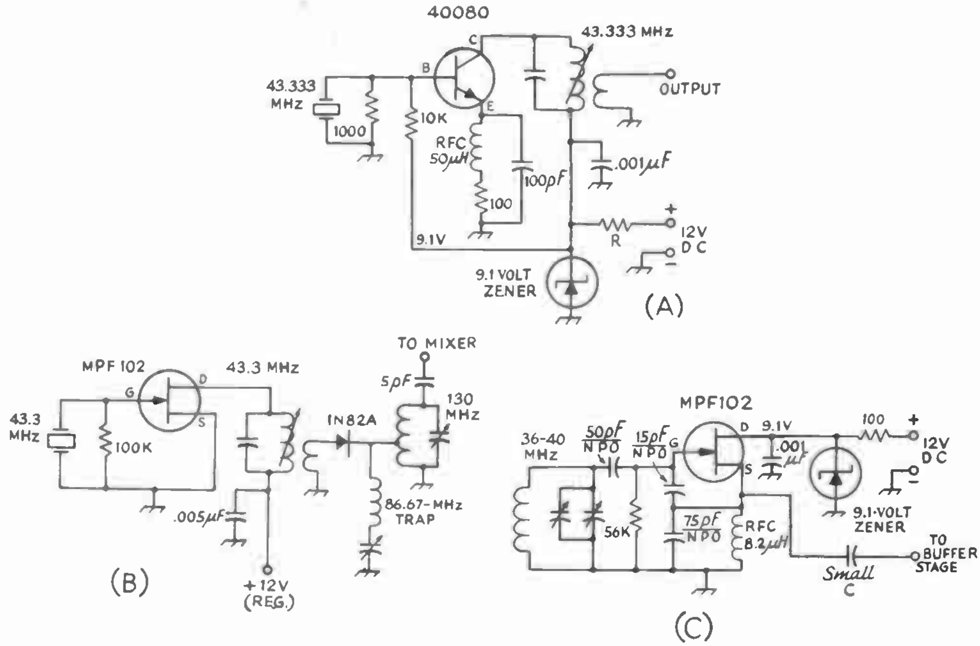


Fig. 9-9 — A simple overtone crystal oscillator for vhf converters, (A) has Zener voltage regulation. An FET overtone oscillator and diode multiplier, (B) supply injection for a 144-MHz converter with a 14-MHz i-f. Series trap absorbs unwanted second harmonic at 86 MHz. A triode oscillator would use essentially the same circuit. A tunable oscillator, as shown at C, would be suitable for a simple 50-MHz receiver with a broad i-f system.

more power from the injection source. When the local-oscillator frequency is far removed from the input frequency, the scheme of Fig. 9-8C can be used. The diagram at 9-8D is needed if the oscillator frequency is within 20 percent of the signal frequency.

The injection level from the oscillator affects mixer performance. Until it affects the mixer adversely in other ways, raising the injection level raises the mixer conversion gain. A simple check is made by observing the effect on signal-to-noise ratio as the injection is varied. At preferred injection levels, the gain will vary but the signal-to-noise ratio will not change. The injection should then be set for conversion gain a few decibels above that at which lower injection causes a drop in signal-to-noise ratio.

Double-tuned circuits in the mixer and the rf amplifier, as shown in several of the schematic diagrams in this chapter, help to keep down mixer response to signals outside the intended tuning range.

The insulated-gate FET is superior to other transistors for mixer service in the matter of overloading. An example is given in Fig. 9-8E. An objection to the MOSFET, the ease with which it can be damaged in handling, has been taken care of by building-in protective diodes in devices such as the MPF122, 40673, and 3N187. Units so designed require no special care in handling, and they work as well as their more fragile predeccessors. Insulated-gate MOSFETs have resistance to overloading which, while superior to most tubes, is not as good as the best JFETs.

Pentode or tetrode tubes make simple and effective mixers, up to 150 MHz or so. Triodes work well at any frequency, and are preferred in the high vhf range. Diode mixers are common in the 420-MHz band and higher.

INJECTION STAGES

Oscillator and multiplier stages that supply heterodyning energy to the mixer should be as stable and free of unwanted frequencies as possible. Stability is no great problem in crystal-controlled converters, if the oscillator is run at low input and its supply voltage is regulated. Simple Zener regulation, as in Fig. 9-9A, is adequate for a transistorized overtone oscillator. A higher order of regulation is needed for tunable oscillators. See Chapter 5 for suitable regulated power supplies.

Unwanted frequencies generated in the injection stages can beat with signals that are outside the intended tuning range. In a typical example, Fig. 9-9B, an FET overtone oscillator on 43.333 MHz feeds a diode tripler to 130 MHz. This frequency beats with signals between 144 and 148 MHz, to give desired responses at 14 to 18 MHz. The multiplier stage also has some output at twice the crystal frequency, 86.666 MHz. If allowed to reach the mixer, this can beat with fm broadcast signals in the 100-MHz region that leak through the rf circuits of the converter. There are many such annoying possibilities, as any vhf enthusiast living near high-powered fm and TV stations has found out.

Spurious frequencies can be kept down by using the highest practical oscillator frequency, no multiplier in a 50-MHz converter, and as few as possible for higher bands. Some unwanted harmonics are unavoidable, so circuit precautions are often needed to prevent both these harmonics and the unwanted signals from reaching the mixer. Selective coaxial or helical-resonator circuits are practical aids in uhf receivers. Trap circuits of various kinds may be needed to "suck out" energy on troublesome frequencies.

The series trap in Fig. 9-5B reduces the level of the 86-MHz second harmonic of the crystal frequency. A 58-MHz parallel-tuned trap, Fig. 9-1A, prevents the entry of Channel 2 TV signals that could otherwise beat with the second harmonic of a 36-MHz oscillator in a 50-MHz converter that works into a 14-MHz i-f (36 × 2 - 14=58).

Unwanted frequencies also increase the noise output of the mixer. This degrades performance in a receiver having no rf amplifier, and makes the job of an amplifier, if used, more difficult.

Frequency multipliers in vhf receivers generally follow transmitting practice, except for their low power level. The simple diode multiplier of Fig. 9-9B will often suffice. Its parallel-tuned 130-MHz circuit emphasizes the desired third harmonic, while the series circuit suppresses the unwanted second harmonic. The trap is tuned by listening to a spurious fm broadcast signal and tuning the series capacitor for minimum interference. The tripler circuit should be peaked for maximum response to a 2-meter signal. Do not detune this circuit to lower injection level. This should be controlled by the voltage on the oscillator, the coupling between the oscillator and multiplier, or by the coupling to the mixer from the 130-MHz circuit.

Tunable Oscillators

Any tunable vhf receiver must employ a variable oscillator. At this point the intermediate frequency is fixed, and the oscillator tunes a range higher or lower than the signal frequency by the amount of the i-f. In the interest of stability, it is usually lower. In Fig. 9-9C a simple JFET oscillator tunes 36 to 40 MHz, for reception of the 50-MHz band with a fixed 14-MHz i-f. Its stability should be adequate for a-m or fm reception with a relatively broad i-f, but it is unlikely to meet the requirements for ssb or cw reception, even for 50 MHz, and certainly not for higher bands.

Practically all vhf reception with high selectivity uses double-conversion schemes, with the tunable oscillator serving the second conversion. Such hf oscillators are treated in Chapter 6. They should run at the lowest practical input level, to minimize drift caused by heating. The supply should be well-regulated pure dc. Mechanically-rugged components and construction are mandatory. The circuits should be shielded from the rest of the receiver, and coupling to the mixer should be as light as practical. Drift cycling due to heating can be minimized if the oscillator is kept running continuously.

THE SUPERREGENERATIVE RECEIVER

Though the newcomer may not be too familiar with the superregenerative detector, the simple "rushbox" was widely used in early vhf work. Nothing of comparable simplicity has been found to equal its weak-signal reception, inherent noise-limiting and agc action, and freedom from overloading and spurious responses. But like all simple devices the superregenerator has limitations. It has little selectivity. It makes a high and unpleasant hissing noise, and it radiates a broad interfering signal around its receiving frequency.

Adding an rf amplifier will improve selectivity and reduce detector radiation. High-Q tuned circuits aid selectivity and improve stability. Use of superregeneration at 14 to 18, 26 to 30 MHz, or some similar hf range, in the tunable element of a simple superheterodyne receiver, works fairly well as a simple tuner for vhf converters. None of these steps corrects the basic weaknesses entirely, so the superregenerator is used today mainly where simplicity, low cost and battery economy are major considerations. Cw and narrow-band fm signals cannot be received using a superregenerative receiver.

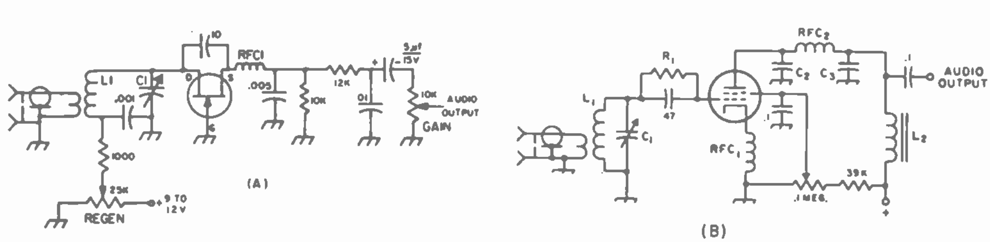


Fig. 9-10 — Circuits of typical superregenerative detectors using a field-effect transistor, A, and a tetrode tube, B. Regeneration is controlled by varying the drain voltage on the detector in the transistor circuit, and the screen voltage in the tetrode or pentode. Values of L1 and C1 should be adjusted for the frequency involved, as should the size of the rf choke, RFC1.

C2, C3 — .001-µF disk ceramic. Try different values up to .005 for desired audio quality.
R1 — 2 to 10 megohms.
L2 — Small audio or filter choke; not critical.
RFC1 — Single-layer rf choke, to suit frequency.
RFC2 — 85-mH rf choke.

Typical superregenerative detector circuits are shown in Fig. 9-10. High-transconductance FET• and high-beta vhf transistors are favored. The power source should be well-filtered and of low impedance. Fresh or well-charged batteries are ideal. Regeneration is controlled by varying the gain of the stage.

SERIES-RESONANT BYPASSING

Inexpensive disk-ceramic and "dog-bone" types of capacitors are relatively ineffective for bypassing above approximately 100 MHz. This is because of their considerable lead inductance, even when they are connected as close to the elements to be bypassed as possible. Actually this lead inductance can be used to advantage by selecting lead lengths that make the capacitor series-resonant at the frequency to be bypassed.

This approach is recommended by WA2KYF, who supplied the information in Table 9-I, showing capacitor and lead-length combinations for effective bypassing of rf energy at frequencies commonly encountered in vhf work. The values are not particularly critical, as a series-resonant circuit is broad by nature. The impedance of a series-resonant bypass is very close to zero ohms at the frequency of resonance, and it will be lower than most conventional capacitors for a considerable range of frequency either side of resonance.

A high-capacitance short-lead combination is preferable to a lower value with longer leads, because the former will be less likely to allow unwanted coupling to other circuits. For example, a 100-pF capacitor with 1/4-inch leads is a better bet than a 25-pF with 1-inch leads, for bypassing at 144 MHz. The series-resonant bypass is worth a try in any circuit where instability is troublesome, and conventional bypassing has been shown to be ineffective.

TABLE 9-I

Values of capacitance in pF required for resonance of frequencies commonly encountered in amateur-band vhf work, for leads of 1/4, 1/2 and 1 inch in length.

Frequency MHz	*1/4-Inch Leads*	*1/2-Inch Leads*	*1-Inch Leads*
48–50	800	400	200
72	390	180	91
96	220	100	56
144	100	47	25
220	39	20	10

MOSFET PREAMPLIFIERS FOR 10, 6, AND 2 METERS

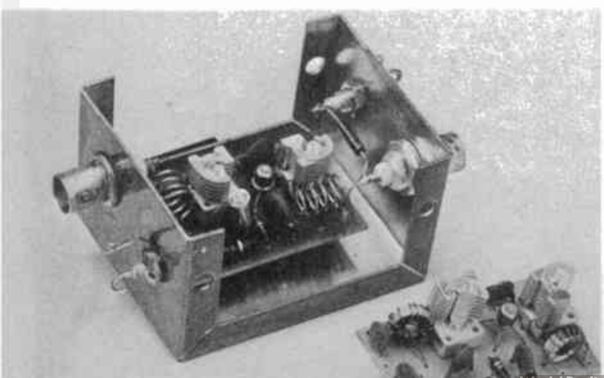
Where an hf or vhf receiver lacks gain, or has a poor noise figure, an external preamplifier can improve its ability to detect weak signals. This preamplifier uses an RCA 40673 dual-gate MOSFET. Designs for using this device as a mixer or as a preamplifier abound and many of them are excellent.

When it comes to simplicity, small size, good performance, low cost, and flexibility, a design by Gerald C. Jenkins, W4CAH, certainly qualifies.

Where the preamplifier really shines is in pepping up the performance of some of the older ten-meter receivers that many have pressed into service. A six-meter version is also very useful for any of the modes of communication available on that band.

The voltage dropping resistor, R4, and the Zener diode, VR1, may be of the value necessary to obtain 9 to 12 V dc for operation of the unit. By increasing the resistance and dissipation rating of R4 and VR1, the preamplifier may be operated from the 150- to 200-V supply found in many tube-type receivers.

The layout of the board is so simple that it is hardly worth the effort of making a negative for the photo-etch process. A Kepro resist-marking pen was used with success on several boards. Another approach – and one that is highly recommended –



Two versions of the preamplifier. The one in the box is for 2-meter use. Toroids are used in the six-meter version (right) and in the ten-meter preamplifier (not shown). Input is at the right on both units. The extra rf choke and feedthrough capacitor on the right end of the Minibox are for decoupling a crystal-current metering circuit that is part of a 2304-MHz mixer.

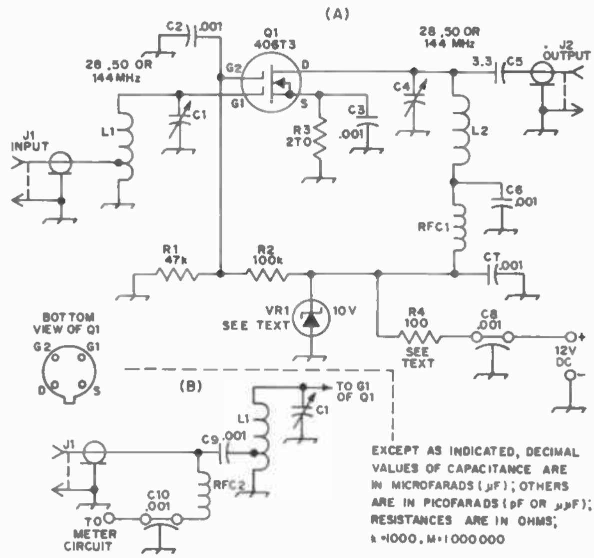


Fig. 1 — Schematic diagram for the preamplifier. Part designations not listed below are for pc board placement purposes. Alternative input circuit for use with microwave diode mixer is shown at B.
C1, C4 — See Table I.
C2, C3, C5, C6, C7, C9 — Disk ceramic.
C8 — .001 feedthrough capacitor.
J1, J2 — Coaxial connectors. Phono-type, BNC or SO-239 acceptable.
L1, L2 — See Table I.
R4 — 3 turns No. 28 enam. on ferrite bead. A 220-ohm, 1/2-watt resistor may be substituted.
RFC2 — 33 μH, iron-core inductor. Millen J300-33 or J. W. Miller 70F335A1.

is to cover the copper with masking tape, transfer the pattern with carbon paper, then cut away the tape to expose the part to be etched. On small, simple boards the masking-tape method is hard to beat.

The pc board may be mounted in almost any small enclosure. Construction is not tricky or difficult. It should take only a few minutes to complete the unit after the board is prepared. The board is fastened in the enclosure by means of one metal standoff post and a No. 4 screw and nut. Input and output connectors are not critical; phono-type jacks may be used in the interest of low cost.

Adjustment is so easy that it almost needs no description. After connecting the amplifier to a receiver, simply tune the input (C1) and the output (C4) for maximum indication on a weak signal. One possible area of concern might be that the toroids used in the ten- and six-meter versions are not always uniform in permeability, as purchased from various suppliers. However, it is an easy matter to add capacitance or remove a turn as required to make the circuits resonate at the correct frequency.

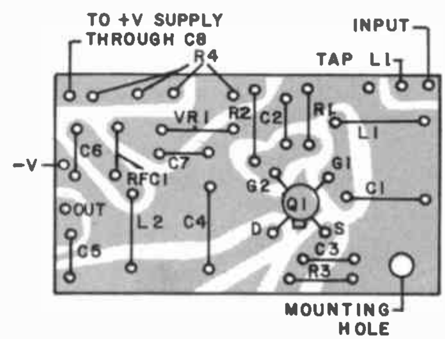


Fig. 2 — Full-scale layout and parts placement guide for the pc board. Foil side shown.

Table I

	28 MHz	*50 MHz*	*144 MHz*
L1	17 turns No. 28 enam. on Amidon T-50-6 core. Tap at 6 turns from ground end	12 turns No. 26 enam. on Amidon T-37-10 core. Tap at 5 turns from ground end.	5 turns No. 20 tinned 1/2-inch ID × 1/2-inch long. Tap at 2 turns from ground end.
L2	Same as L1, without tap.	Same as L1, without tap.	4 turns No. 20 tinned like L1, without tap.
C1, C4.	15 to 60-pF ceramic trimmer. Erie 538-002F.	1.8- to 16.7-pF air variable. E. F. Johnson 189-506-005.	1.5- to 11.6-pF air variable. E. F. Johnson 189-504-005.

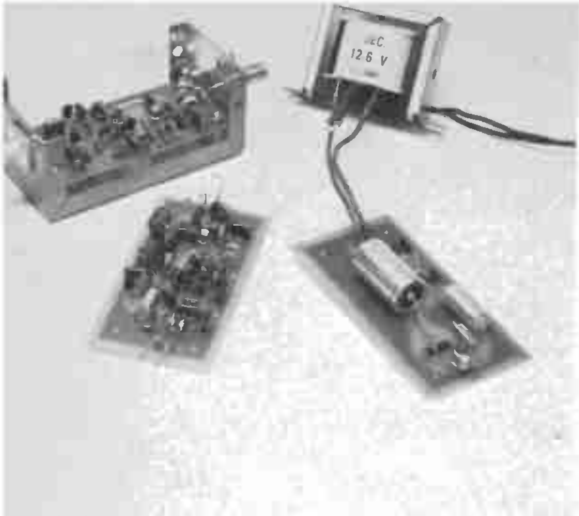


Fig. 1 — Completed six- and two-meter converters (left and center) with power supply.

CONVERTERS FOR 50 AND 144 MHz

The converters described here are designed by the Rochester VHF Group and details are presented by W2DUC and K2YCO.

Because of the nature of the project, a universal circuit-board design is used. One circuit board serves for either band, with only slight modification. Other specific design goals were:

1) Low noise figure, less than 3 dB.

2) State-of-the-art freedom from cross modulation.

3) Sufficient gain to override the front-end noise of most receivers.

4) Double-tuned bandpass interstage and output circuits to achieve a flat response over a two-MHz portion of either band.

5) Filtering of the local oscillator chain in the two-meter model to reduce spurious responses.

6) Small size and low power consumption.

7) Freedom from accidental mistuning during the life of the converter.

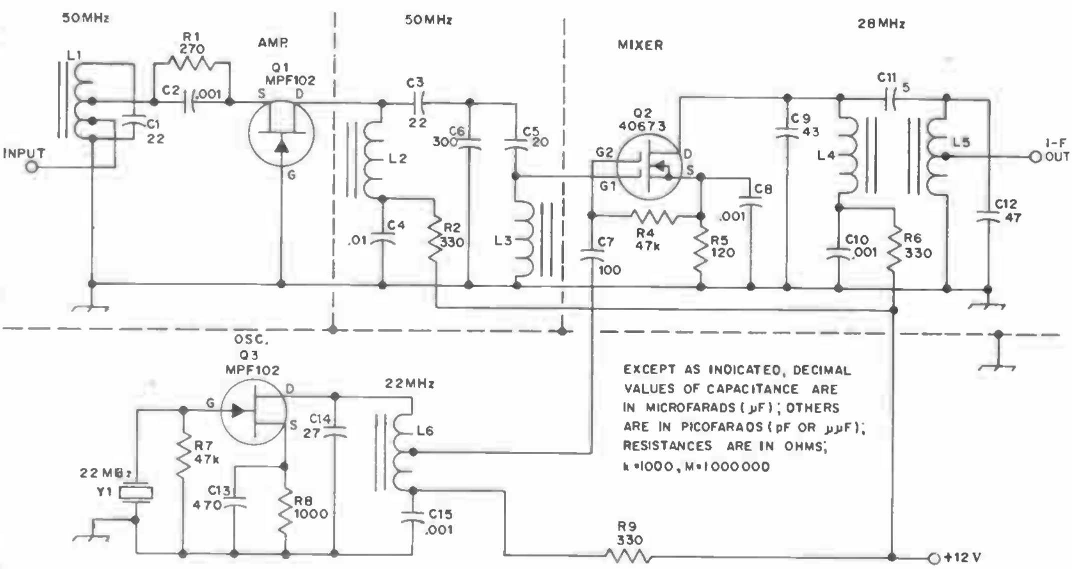


Fig. 2 — Schematic diagram of the six-meter converter. All resistors are 1/4-watt composition. C2, C8, C10 and C15 are .001μF disk ceramic. C4 is .01-μF disk ceramic. All other capacitors are dipped mica.

L1-L6, incl. — All No. 28 enam. wire wound on Amidon T-30-6 cores as follows: L1, 14 turns tapped at 4 turns and 6 turns; L2, 13 turns; L3, 12 turns; L4, 18 turns; L5, 18 turns tapped at 4 turns from cold end; L6, 26 turns tapped at 6 turns from hot end.

Y1 — 22-MHz crystals. International Crystal Mfg. Co. type EX.

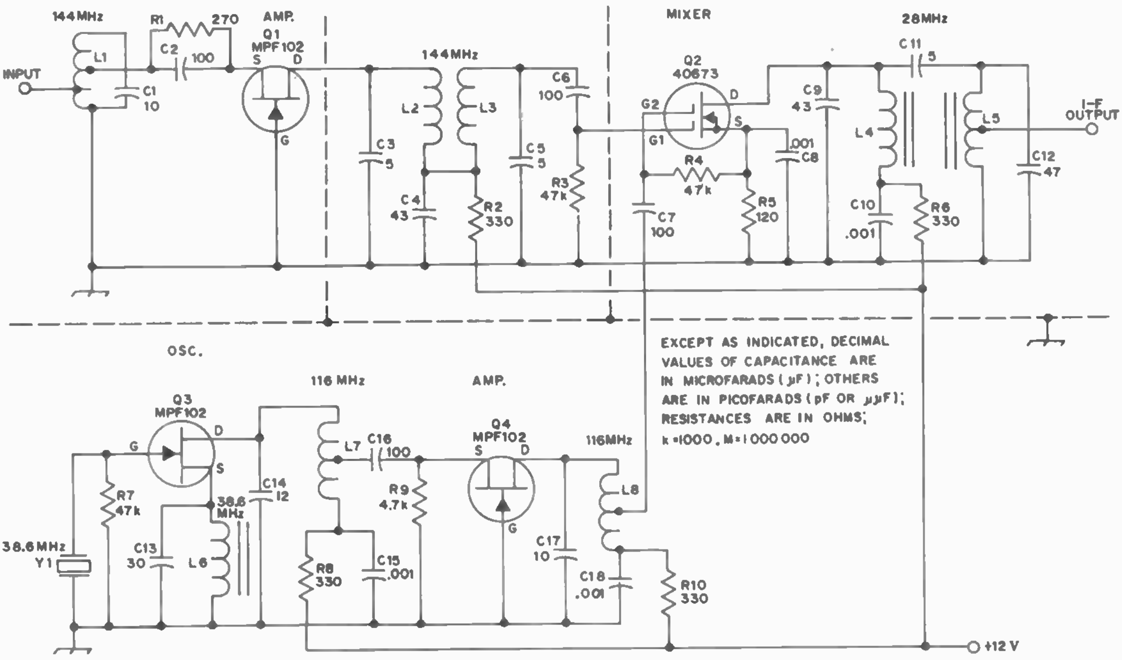


Fig. 3 — Schematic diagram of the two-meter converter. All resistors are 1/4-watt composition. C8, C10, C15 and C18 are .001-μF disk ceramic. All other capacitors are dipped mica units.

L1, L2, L3, L7, L8 — All No. 20 enam. wire formed by using the threads of a 1/4-20 bolt as a guide. L1, 5 turns tapped at 1-3/4 turns and 3/4 turn from cold end; L2, 5 turns; L3, 4 turns; L7, and L8, 5 turns tapped at 2 turns from hot end.

L4 — 18 turns No. 28 enam. wound on Amidon T-30-6 core.

L5 — 18 turns like L4, tapped at 4 turns from cold end.

L6 — 0.68 μH miniature inductor. Delevan 1025 series or J. W. Miller 9230-16.

Y1 — 38.666-MHz crystal. International Crystal Mfg. Co. type EX.

Other points considered were such things as freedom from the necessity of neutralization and the use of moderately priced transistors.

Several breadboard models were constructed and tested as the design evolved. Fig. 1 shows two completed converters and a power supply.

Circuit Design

A schematic diagram for the six-meter converter is shown in Fig. 2, and for the two-meter model in Fig. 3. The configuration of the rf and mixer portions of the circuit are virtually identical for six and two meters, with the values of the frequency-determining components being scaled appropriately. The major difference between the two converters is a change in the local oscillator chain. A minor change in the method of interstage coupling was necessary to prevent stray-capacitance effects from making the alignment critical on the six-meter converter.

All inductors in the six-meter model and the two-meter output circuit are wound on Amidon T-30-6 toroid cores. The tuned circuits are aligned by spreading or compressing the turns around the toroid core. After alignment the coils are glued in place with Silastic compound (sold as bathtub caulk).

The rf amplifier, Q1, is used in a grounded-gate configuration. The input circuit is tapped to provide a proper match between the antenna and source of the FET while maintaining a reasonable *Q*. The six-meter interstage coupling network consists of C3, C5, L2, and L3. Band-pass coupling is controlled by the capacitive T network of C3 and C5 in ratio with C6. A 40673 dual-gate MOSFET is used in the mixer circuit (Q2). Gate 1 receives the signal, while gate 2 has the local oscillator injection voltage applied to it through C7. A slight amount of positive bias is applied to gate 2 through R2. A top-coupled configuration, using toroid inductors, serves as the 28-MHz output circuit of both converters.

The oscillator circuit in the six-meter model is straightforward, relying on the drain-to-gate capacitance of the FET for feedback. A tap at four turns from the hot end of the toroid winding provides the injection to the mixer through capacitor C7. In the two-meter converter, Fig. 3, the rf stage is identical to the six-meter version except for

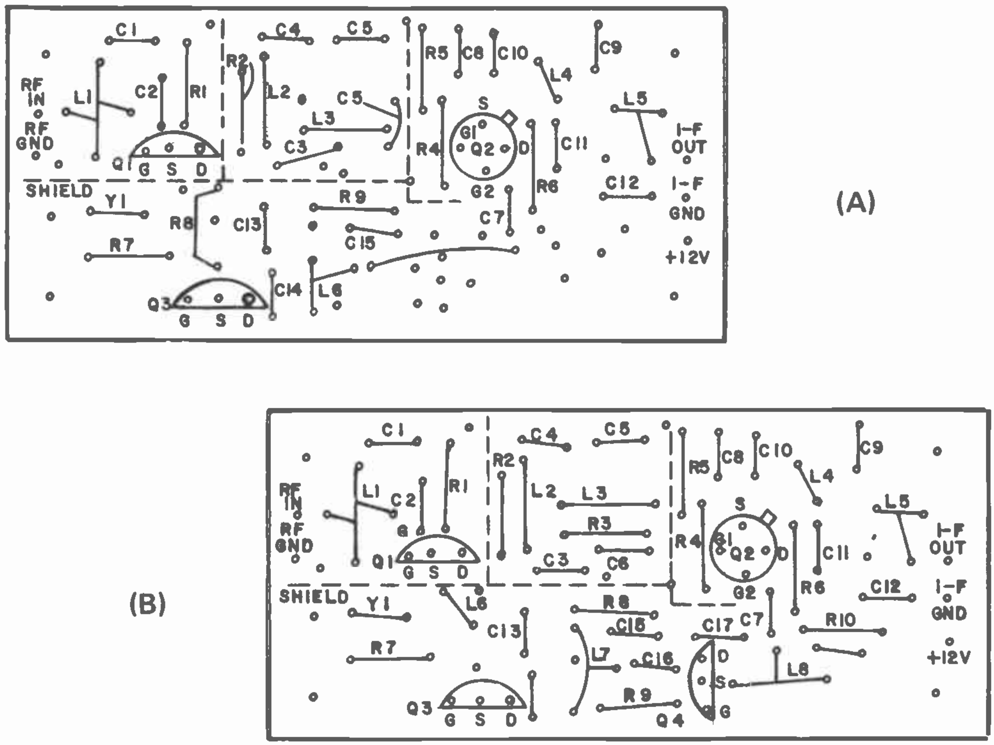


Fig. 4 — Parts-placement guide for the six-meter converter, A, and the two-meter converter, B. View is from the foil side of the board. Dashed lines show the location of shields that are soldered to short pieces of wire which project through holes in the pc board. The shields may be fabricated from sheet brass or copper, or scraps of copper-clad board material.

the tuning networks. L1, L2, and L3 are air wound, self-supporting, and are formed initially by winding wire around the threads of a 1/4-20 bolt. The turns of L1 are spread to permit adding taps prior to mounting on the board. The degree of interstage coupling in the two-meter model is controlled by the positions of L2 and L3. Since they are mounted at right angles, the coupling is very light. By changing the angle between these two coils, the passband may be optimized.

In the two-meter oscillator stage, Q3 is changed to an oscillator/tripler by replacing the source bias resistor with L6. Replace bypass capacitor, C13,

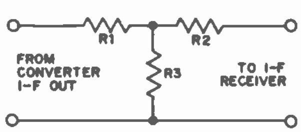


Fig. 5 — An i-f attenuator may be necessary if the receiver following the converter is exceptionally hot. Values for 6 dB: R1, R2 – 18 ohms; R3 – 68 ohms. For 10 dB: R1, R2 – 27 ohms; R3 – 39 ohms.

with a 30-pF value to resonate L6 near the crystal frequency. Source-to-gate capacitance provides the feedback in this case. The drain tank is modified to provide output at the third harmonic, thus eliminating the need for a separate tripler stage. Q4 is used as an isolation amplifier running at very low current level (as controlled by R9) to provide attenuation of the adjacent harmonics. This stage is not needed for amplification of the oscillator signal but without the additional filtering, severe "birdies" may result from nearby fm or TV stations. In both the six- and two-meter versions, a number of printed-circuit pads will be left over when construction is completed. These are the result of providing both bands on a common pc layout. For example, the isolation amplifier following the oscillator is not used on six meters. Therefore, this stage is bypassed by a jumper wire from L6 to C7. Five additional holes are located in the ground area along the centerline of the board and between rf and mixer stages. Component lead clippings are soldered into these holes to provide a mounting for the shield partitions, which are soldered to the wires where they extend through the board. Fig. 4 shows the parts layout for the six- and two-meter converters. Notice that one lead of

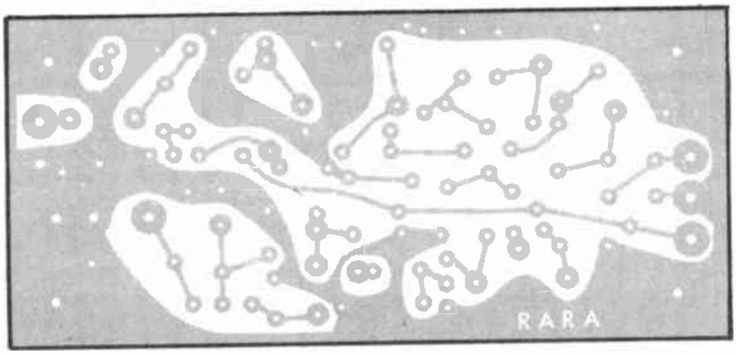


Fig. 6 — Scale-size layout for the pc board. The same pattern is used for either band. Foil side shown here.

Table I — Performance Specifications

Parameter	*6 Meters*	*2 Meters*
Noise figure, dB	1.8 – 2.3	2.0 – 2.4
Conversion gain, dB	22 – 28	17 – 24
Spurious responses, dB	–80*	–60*
	* Has a response at 6 MHz	* Responses at 107 & 181 MHz
Freq. response, ± 1 dB	49.8 – 51.5 MHz	143.9 – 146.4 MHz
Current at 12 V dc	12 – 18 mA	14 – 20 mA

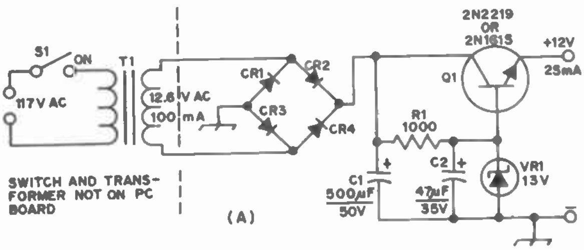
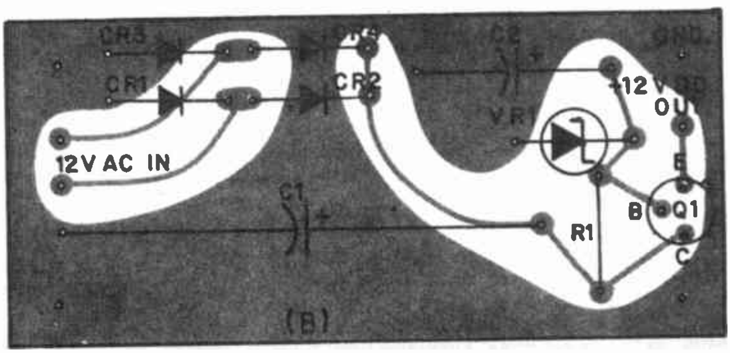


Fig. 7 — Schematic diagram and parts-placement guide for the power supply to the converters. The transformer is mounted external to the board. Pc board size is identical to the one used for the converters.



C3 must reach past the ground hole and connect to the foil. R3 is not used on the six-meter converter.

Alignment and Test

Perhaps the most difficult task in the project was the test and tune-up of the finished converter. A single test setup using a sweep generator, diode probe, and oscilloscope was a necessity to assure the flat response over the tuning range. Commercial attenuators were used to calibrate each converter by the substitution method.

Tuning of the air-wound rf circuit for two meters was accomplished by spreading or compressing the turns of the coils. After alignment, the windings were secured by a bead of Silastic compound along the oil to hold the turns in place. The noise figure of each converter was checked using the Monode noise-generator technique.[1] A final sensitivity check using a receiver (NC300) and a model 80 calibrated signal generator completed the checkout.

[1] Guentzler, "The Monode Noise Generator," *QST*, April 1967.

The transistors used in the rf stage were also subject to some variation in noise figure. When this occurred, an rf FET was carefully traded with an oscillator FET, since performance of the FET as an oscillator was always satisfactory.

The performance specification range for the converters is seen in Table I.

Small ceramic trimmers can be used in place of the fixed-value mica capacitors in the tuned circuits of these converters. The midrange of the trimmer should be approximately the value of the mica capacitors replaced. This procedure may simplify the tuning process of the converters where a sweep generator setup is not available. A little careful tweaking should give a reasonably flat response.

If trimmers are used, the rf input circuit should be tuned to the center of the desired response, 50.5 MHz as an example. This circuit tunes broadly and is not too critical. The rf interstage circuits should be stagger tuned, one at 50.0 MHz and the other at 51.0 MHz, as an example, the output i-f circuits can be tuned in a manner similar to the interstage circuits.

DOUBLE BALANCED MIXER

Advances in technology have, in recent years, provided the amateur builder with many new choices of hardware to use in the building of receivers, converters, or preamplifiers. The broadband double-balanced mixer package is a fine example of this type of progress, and as amateurs gain an understanding of the capabilities of this device, they are incorporating this type of mixer in many pieces of equipment, especially receiving mixers. The combined mixer/amplifier described here was presented originally in *QST* for March, 1975, by K1AGB,

Mixer Comparisons

Is a DBM really better than other types? What does it offer, and what are its disadvantages? To answer these questions, a look at more conventional "active" (voltages applied) mixing techniques and some of their problems is in order. The reader is referred to a recent article in *QST*[3] dealing with mixers. Briefly reiterated, common single-device active mixers with gain at vhf and uhf are beset with problems of noise, desensitization and small local-oscillator (LO) isolation from the r-f and i-f "ports." As mixers, most devices have noise figures in excess of those published for them as rf amplifiers and will not provide sufficient sensitivity for weak-signal work. To minimize noise, mixer-device current is generally maintained at a low level. This can reduce dynamic range, increasing overload potential, as defined in the terminology appendix. Gain contributions of rf amplifiers (used to establish a low system noise figure) further complicate the overload problem. LO-noise leakage to the rf and i-f ports adversely affects system performance. Mixer dynamic range can be limited by conversion of this noise to i-f, placing a lower limit on mixer system sensitivity. Generally 20 dB of mixer midband, inter-port, isolation is required, and most passive DBM can offer greater than 40 dB.

A commercially manufactured double-balanced diode mixer offers performance predictability, circuit simplicity and flexibility. Closely matched Schottky-barrier hot-carrier diodes, commonly used in most inexpensive mixers of this type, provide outstanding strong-signal mixer performance (up to about 0 dBm at the rf input port) and add little (0.5 dB or so) to the mixer noise figure. Essentially, diode conversion loss from rf to i-f, listed in Table I, represents most of the mixer

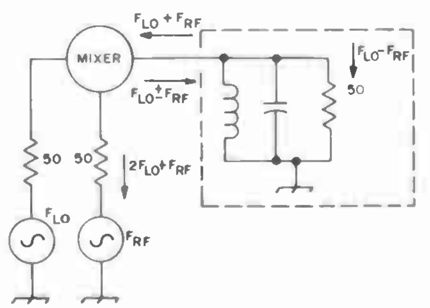


Fig. 1 — The i-f port of a double-balanced mixer is matched at fLO − $f$rf and reactive at fLO + $f$rf. In this configuration conversion loss, rf compression and desensitization levels can vary ±3 dB while harmonic modulation and third-order IMD products can vary ±20 dB.

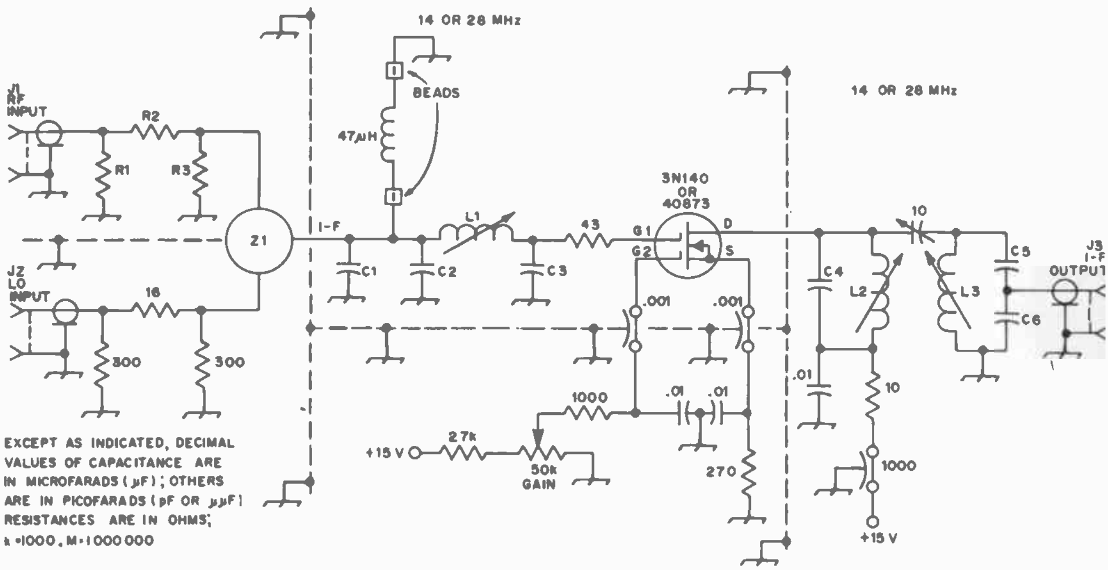


Fig. 2 – A schematic diagram for the double-balanced mixer and i-f post amplifier. The i-f can be either 14 or 28 MHz. Parts values are given in Table II.

contribution to system noise figure.† Midband isolation between the LO port and the rf and i-f ports of a DBM is typically > 35 dB – far greater than that achievable with conventional single-device active-mixing schemes. This isolation is particularly advantageous in dealing with low-level local-oscillator harmonic and noise content. Of course, selection of LO devices with low audio noise figures, and proper rf filtering in the LO output, will reduce problems from this source.

Often listed disadvantages of a diode DBM are (a) conversion loss, (b) LO power requirements, and (c) i-f-interface problems. The first two points are closely interrelated. Conversion loss necessitates some low-noise r-f amplification to establish a useful weak-signal system noise figure. Active mixers also have this requirement, as will be demonstrated later. Additional LO power is fairly easy to generate, filter, and measure. If we accept the fact that more LO power is necessary for the DBM than is used in conventional single-device active mixing circuits, we leave only two real obstacles to be overcome in the DBM, those of conversion loss and i-f output interfacing.

To minimize conversion loss in a DBM, the diodes are driven by the LO beyond their square-law region, producing an output spectrum which in general includes the terms:

- Fundamental frequencies fLO and $f$rf
- All of their harmonics
- The desired i-f output, fLO ± $f$rf
- All higher order products of nfLO ± mfrf where n and m are integers.‡

The DBM, by virtue of its symmetry and internal transformer balance, suppresses a large number of the harmonic modulation products. In the system described here, fLO is on the low side of $f$rf; therefore, numerically, the desired i-f output is $f$rf – fLO. Nonetheless, the term fLO ± $f$rf appears at the i-f-output port equal in amplitude to the desired i-f signal, and this unused energy must be effectively terminated to obtain no more than the specified mixer-conversion loss. This is not the image frequency, fLO – fi-f, which will be discussed later.

In any mixer design, all rf port signal components must be bypassed effectively for best conversion efficiency (minimum loss). Energy not "converted" by mixing action will reduce conversion gain in active systems, and increase conversion loss in passive systems such as the diode DBM. Rf bypassing also prevents spurious resonances and other undesired phenomena from affecting mixer performance. In this system, rf bypassing at the i-f-output port will be provided by the input capacitance of the i-f interface. The DBM is not a panacea for mixing ills, and its effectiveness can be reduced drastically if all ports are not properly terminated.

DBM Port Terminations

Most DBM-performance inconsistencies occur because system source and load impedances presented to the mixer are not matched at all frequencies encountered in normal operation. The terminations (attenuator pads) used in conjunction with test equipment by manufacturers to measure published performance characteristics are indeed "broadband" matched. Reactive mixer terminations can cause system problems, and multiple reactive terminations can usually compound these problems to the point where performance is very

†See appendix on noise figure.

‡See appendix for mixer terminology.

TABLE I

Manufacturer:	*Relcom*	*Anzac*	*MCL*	*MCL*	*MCL*	*MCL*
Model	M6F*	MD-108	SRA-1	SRA-1H	RAY-1	MA-1 ***
Frequency range, LO	2-500	5-500	.5-500	.5-500	5-500	1-2500
MHz rf	2-500	5-500	.5-500	.5-500	5-500	1-2500
i-f	DC-500	DC-500	DC-500	DC-500	DC-500	1-1000
Conversion loss, Mid-range	9 dB Max.	7.5 dB Max.	6.5 dB Typ.	6.5 dB Typ.	7.5 dB Typ.	8.0 dB Typ. 1-2.5 GHz.
Isolation, LO – RF	35-40 dB Min.	40 dB Min.	45 dB Typ.	45 dB Typ.	40 dB Typ.	40 dB Typ.
mid-range LO – i-f	25-35 dB Min.	35 dB Min.	40 dB Typ.	40 dB Typ.	40 dB Typ.	40 dB Typ.
Total input power:	50 mW	400 mW	500 mW	500 mW	1 W	50 mW
LO power requirement:	+7 dBm (5 mW)	+7 dBm (5 mW)	+7 dBm (5 mW)	+17 dBm (50 mW)	+23 dBm (200 mW)	+10 dBm (10 mW)
Signal 1-dB compression level:	Not spec.	Not spec.	+1 dBm	+10 dBm	+15 dBm	+7 dBm
Impedance, all ports:	50 ohms	50 ohms	50 ohms	50 ohms	50 ohms	50 ohms
Price class:	**	$7 Single unit	$9.95 Single unit	$15.95 5+ units	$31.95 4+ units	$99.95 Single unit

All specifications apply only at stated LO power level.

* 1968 data.
** Units provided by a second source.
*** SMA connectors standard.
Relcom, Division of Watkins-Johnson, 3333 Hillview Ave., Palo Alto, CA 94304.
Anzac Electronics, 39 Green Street, Waltham, MA 02154.
MCL – Mini-Circuits Laboratory, 837–843 Utica Ave., Brooklyn, NY 11203.

difficult to predict. Let's see how we can deal with reactive terminations.

The I-F Port

The i-f port is very sensitive to mismatch conditions. Reflections from the mixer/i-f amplifier interface (the pi network in Fig. 2) can cause the conversion loss to vary as much as 6 dB. Also greatly affected are third-order intermodulation-product ratio and the suppression of spurious signals, both of which may vary ±10 dB or more. It is ironic that the i-f port is the most sensitive to a reactive termination, as this is a receiving system point where sharp-skirted filters are often desired.

Briefly, here is what happens with a reactive i-f-port termination. Fig. 1 shows a DBM with "high side" LO injection and an i-f termination matched at fLO – $f$rf but reactive to fLO + $f$rf. The latter term re-enters the mixer, again combines with the LO and produces terms that exit at the rf port, namely 2fLO + $f$rf, a dc term, and fLO + $f$rf – fLO (the original rf-port input frequency). This condition affects conversion loss, as mentioned earlier, in addition to rf-port VSWR, depending on the phase of the reflected signal. The term 2fLO + $f$rf also affects the harmonic modulation-products spectrum resulting in spurious responses.

One solution to the i-f-interface problem is the use of a broadband 50-ohm resistive termination, like a pad, to minimize reflections. In deference to increased post-conversion system noise figure, it seemed impractical to place such a termination at the mixer i-f output port. While a complimentary filter or diplexer (high-pass/low-pass filters appropriately terminated) can be used to terminate both $f$rf + fLO and $f$rf – fLO §, a simpler method can be used if $f$rf + fLO is less than 1 GHz and ($f$rf + fLO) / ($f$rf – fLO) > 10. Place a short-circuit termination to $f$rf + fLO, like a simple lumped capacitance, directly at the mixer i-f terminal. This approach is easiest for the amateur to implement and duplicate, so a form of it was tried – with success. In our circuit, C1 serves a dual purpose. Its reactance at $f$rf + fLO is small enough to provide a low-impedance "short-circuit" condition to this term for proper mixer operation. Additionally, it is part of the input reactance of the mixer i-f-amplifier interface. Fortunately the network impedance-transformation ratio is large enough, and in the proper direction, to permit a fairly large amount of capacitance (low reactance) at the mixer i-f-output port. The capacitor, in its dual role, must be of good quality at vhf/uhf (specifically $f$rf + fLO), with short leads, to be effective. The mixer condition ($f$rf + fLO) / ($f$rf – fLO) > 10 is met at 432 and 220 MHz with a 404/192-MHz LO (28-MHz i-f) and on 14 MHz with a 130-MHz LO (14-MHz i-f). At 50 MHz, with a 36 MHz LO, we are slightly shy of the requirement, but no problems were encountered in an operating unit. The pi-type interface circuit assures a decreasing impedance as i-f operation departs from midband, thereby lessening IMD problems.

The LO Port

The primary effect of a reactive LO source is an increase in harmonic modulation and third-order IMD products. If the drive level is adequate, no effect is noted on conversion loss, rf compression and desensitization levels. A reactive LO source can be mitigated by simply padding the LO port with a 3- or 6-dB pad and increasing the LO drive a like amount. If excess LO power is not available, matching the LO source to the mixer will improve

§ **Presentation and calculation format of these terms is based on "low side" LO injection. See the appendix for explanation.**

performance. This method is acceptable for single-frequency LO applications, when appropriate test equipment is available to evaluate matching results. For simplicity, a 3-db pad was incorporated at the LO-input port as an interface in both versions of the mixer. Thus the LO port is presented with a reasonably broadband termination, and is relatively insensitive to applied frequency, as long as it is below about 500 MHz. This implies that frequencies other than amateur assignments may be covered – and such is indeed the case when appropriate LO frequencies and rf amplifiers are used. Remotely located LOs, when adjusted for a 50-ohm load, can be connected to the mixer without severe SWR and reflective-loss problems in the transmission line.

Broadband mixers exhibit different characteristics at different frequencies, due to circuit resonances and changes in diode impedances resulting from LO power-level changes. Input impedances of the various ports are load dependent, even though they are isolated from each other physically, and by at least 35 dB electrically. At higher frequencies, this effect is more noticeable, since isolation tends to drop as frequency increases. For this reason, it is important to maintain the LO power at its appropriate level, once other ports are matched.

The RF Port

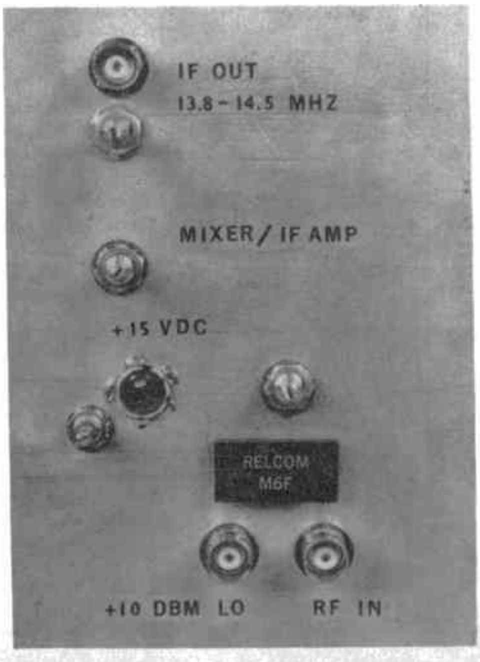
A reactive rf source is not too detrimental to system performance. This is good, since the output impedance of most amateur preamplifiers is seldom 50 ohms resistive. A 3-dB pad is used at the rf port in the 50- and 144-MHz mixer to 14 MHz, and a 2-dB pad is used in the 220/432-MHz mixer to 28 MHz, although they add directly to mixer noise figure. Rf inputs between about 80 and 200 MHz are practical in the 14-MHz i-f-output model, while the 28-MHz-output unit is most useful from 175 to 500 MHz. Mixer contribution to system noise figure will be almost completely overcome by a low-noise rf amplifier with sufficient gain and adequate image rejection.

Image Response

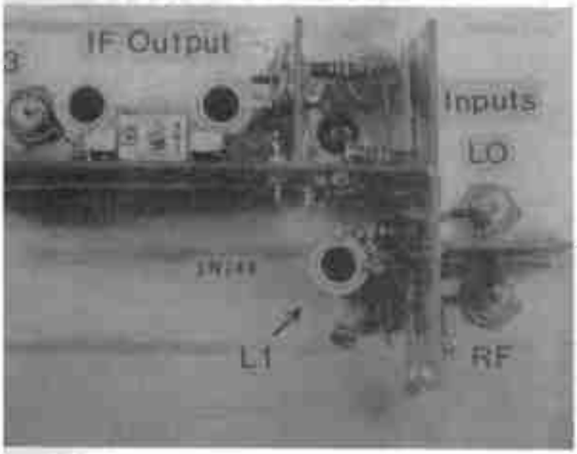
Any broadband mixing scheme will have a potential image-response problem. In most amateur vhf/uhf receiver systems (as in these units) single-conversion techniques are employed, with the LO placed below the desired rf channel for non-inverting down-conversion to i-f. Conversion is related to both i-f and LO frequencies and, because of the broadband nature of the DBM, input signals at the rf image frequency (numerically fLO – fi-f in our case) will legitimately appear inverted at the i-f-output port, unless proper filtering is used to reduce them at the mixer rf-input port. For example, a 144-MHz converter with a 28-MHz i-f output (116-MHz LO) will have rf image-response potential in the 84 to 88-MHz range. TV channel 6 wideband-fm audio will indeed appear at the i-f-output port near 28 MHz unless appropriate rf-input filtering is used to eliminate it. While octave-bandwidth vhf/uhf "imageless mixer" techniques can improve system noise performance by about 3 dB (image noise reduction), and image signal rejection by 20 dB – and much greater with the use of a simple gating scheme – such a system is a bit esoteric for our application. Double or multiple-conversion techniques can be used to advantage, but they further complicate an otherwise simple system. Image noise and signal rejection will depend on the effectiveness of the filtering provided in the rf-amplifier chain.

Mixer Selection

The mixer used in this system is a Relcom M6F, with specifications given in Table I. Suitable substitute units are also presented. The M6F is designed for printed-circuit applications (as are the recommended substitutes), and the lead pins are rather short. While mixers are available with connectors attached, they are more expensive. The simple package is suggested as, aside from less expense, improved interface between mixer and i-f amplifier is possible because of the short leads. The combining of mixer and i-f amplifier in one converter package was done for that reason. Along these lines, the modular-construction approach



This top view of the DBM/i-f amplifier shows the plastic mixer package plus rf/LO inputs and i-f output jacks clearly marked for cabling. The unit is mounted on the open face of a standard 6 × 4 × 2-inch aluminum chassis. This shielding is necessary to prevent the 3N140 from picking up external signals in the 14-MHz region.



The bottom view of the DBM/i-f amplifier shows component and shielding layout. L1, the mixer-amplifier interface inductance and associated components are indicated. C1, with its wide silver-strap leads, is connected directly between the mixer i-f output pin and the copper-clad ground plane with essentially zero lead length. Connection between the mixer output pin and other components (L1, C2 and the rf choke for d-c return) is made by using excess lead from C1. The 43-ohm, 1/4-W resistor in the 3N140 gate 1 lead is connected between the high-impedance end of L1 and a spare terminal on the coil form. The device gate No. 1 lead and resistor are joined at this point. It is important that input/output isolation of the 3N140 be maintained as it is operating at high gain. Mixer packages other than the M6F may have different pin connections and require slightly different input-circuit layout and shielding. Double-sided copper-clad board was used throughout.

permits good signal isolation and enables the mixer-amplifier/i-f system to be used at a variety of rf and LO-input frequencies, as mentioned earlier.

Most commonly available, inexpensive DBM are not constructed to take advantage of LO powers much above +10 dBm (10 mW). To do so requires additional circuitry which could degrade other mixer characteristics, specifically conversion loss and inter-port isolation. The advantage of higher LO power is primarily one of improved strong-signal-handling performance. At least one manufacturer advertises moderately priced "high-level" receiving DBM which can use up to +23 dBm (200 mW) LO power, and still retain excellent conversion loss and isolation characteristics, shown in Table I. The usefulness of mixers with LO power requirements above the commonly available +7 dBm (5 mW) level in amateur receiving applications may be a bit moot, as succeeding stages in most amateur receivers will likely overload before the DBM. Excessive overdesign is not necessary.

In general, mixer selection is based on the lowest practical LO level requirement that will meet the application, as it is more economical and results in the least LO leakage within the system. As a first-order approximation, LO power should be 10 dB greater than the highest anticipated input-signal level at the rf port. Mixers with LO requirements of +7 dBm are quite adequate for amateur receiving applications.

Application Design Guidelines

While the material just presented only scratches the surface in terms of DBM theory and utilization in amateur vhf/uhf receiving systems, some practical solutions to the non-ideal mixer-port-termination problem have been offered. To achieve best performance from most commercially manufactured broadband DBM in amateur receiver service, the following guidelines are suggested:

- Choose i-f and LO frequencies that will provide maximum freedom from interference problems. Don't "guesstimate," go through the numbers!
- Provide a proper i-f-output termination (most critical).
- Increase the LO-input power to rf-input power ratio to a value that will provide the required suppression of any in-band interfering products. The specified LO power (+7 dBm) will generally accomplish this.
- Provide as good an LO match as possible.
- Include adequate pre-mixer rf-image filtering at the rf port.

When the mixer ports are terminated properly performance usually in excess of published specifications will be achieved – and this is more than adequate for most amateur vhf/uhf receiver mixing applications.

This is a side view showing construction details for the double-tuned i-f output circuit. The 3N140 drain lead passes through the shield wall via a small Teflon press-fit bushing and is connected directly to L2. A dc-input isolation compartment along with device gate 2 biasing components (bias configuration modified slightly after photograph was taken), can be seen to the left of the i-f-output components. L2 and L3 are spaced 1–1/8-inch (2.9 cm) center-to-center in the 14-MHz model shown, and 1 inch (2.5 cm) apart in the 28-MHz unit.

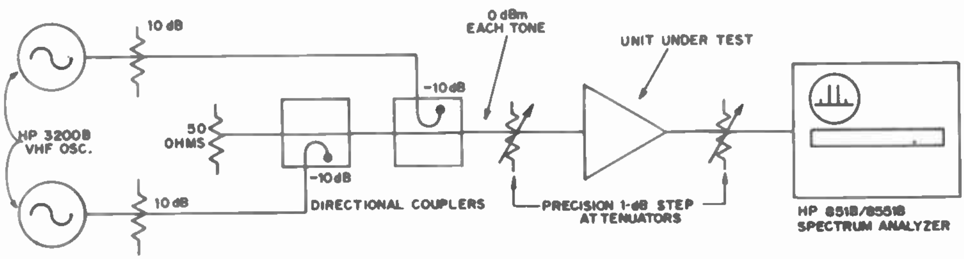


Fig. 6 — A test setup used to measure IMD. The first attenuator adjusts the input level to the unit under test. The second one provides a means of staying within the linear range or the spectrum analyzer.

The Combined DBM/I-F Amplifier

A low-noise i-f amplifier (2 dB or less) following the DBM helps ensure an acceptable system noise figure when the mixer is preceeded by a low-noise rf amplifier. A pi-network matching system used between the mixer i-f-output port and gate 1 of the 3N140 transforms the nominal 50-ohm mixer-output impedance to a 1500-ohm gate-input impedance (at 28 MHz) specifically for best noise performance. The network forms a narrow-band mixer/i-f-output circuit which serves two other important functions: It helps achieve the necessary isolation between rf-and i-f signal components, and serves as a 3-pole filter, resulting in a monotonic decrease in match imperances as the operating i-f departs from mid-band. This action aids in suppression of harmonic-distortion products.

The combined DBM/i-f amplifier is shown schematically in Fig. 2 and pictorially in the photographs. In the 14-MHz model, the 3N140 drain is tapped down on its associated inductance to provide a lower impedance for better strong-signal-handling ability. The 3N140 produces about 19 dB gain across a 700-kHz passband, flat within 1 dB between 13.8 and 14.5 MHz. A 2-MHz passband is used for the 28-MHz model, and the device drain is connected directly to the high-impedance end of its associated inductance. Both amplifiers were tuned independently of their respective mixers, and checked for noise figure as well as gain. With each i-f amplifier pretuned and connected to its mixer, signals were applied to the LO and rf-input ports. The pi-network inductance in the i-f interface was adjusted carefully to see if performance had been altered. No change was noted. I-f gain is controlled by the externally accessable potentiometer. Passband tuning adjustments in the drain circuit are best made with a sweep generator, but single-signal tuning techniques will be adequate. While there should be no difficulty with the non-gate-protected 3N140, a 40673 may be substituted directly if desired.

DBM/I-F Amplifier IMD Evaluation

Classical laboratory IMD measurements made on the DBM/i-f amplifier, using the test setup shown in Fig. 6, from both tones of a two-equal-tone rf-input test signal consisting of -10 dBm each tone. The tones were closely spaced in the 144-MHz range, and converted to 14 MHz LO. Close spacing was necessary to ensure third-order products would appear essentially unattenuated within the relatively narrow i-f-output passband. In operation, as simulated by these test conditions, equivalent output signal levels at J3 would be strong enough to severely overload most amateur receivers. Perhaps the early Collins 75A series, R390A and those systems described by Sabin[4] and Hayward[5] would still be functioning well.

A high-performance, small-signal, vhf/uhf receiving amplifier optimized for IMD reduction and useful noise figure is only as good as any succeeding receiving-system stage, in terms of overload. The DBM/i-f-amplifier combination presented significantly reduces common first-mixer overload problems, leaving the station receiver as the potentially weak link in the system. When properly understood and employed, the broadband DBM followed by a selective low-noise i-f amplifier can

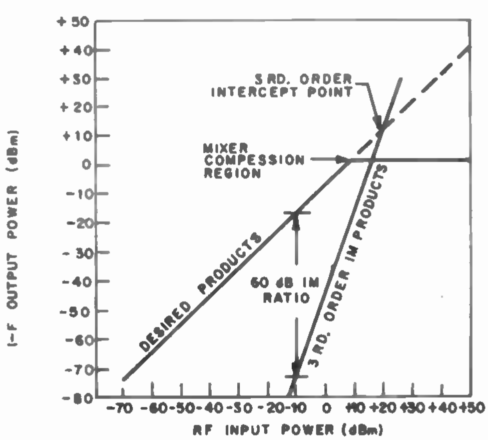


Fig. 7 — A third-order intercept point is determined by extrapolating the desired product curve beyond the mixer compression point and intersecting with the third-order IM-product curve. In this case LO power is +7 dBm, conversion loss is 5 dB.

Table II

DBM/IF AMPLIFIER

PARTS LIST

	14 MHz i-f output	*28 MHz i-f output*
C1	470 pF JFD 471J or equal.	300 pF JFD 301J or equal.
C2	390 pF SM	not used
C3	180 pF SM	51 pF SM
C4	39 pF SM	18 pF SM
C5	56 pF SM	27 pF SM
C6	300 pF SM	150 pF SM
L1	9 turns No. 18 enam., close-wound on a 3/8-inch (.95 cm) diameter red-slug coil form.	9 turns No. 24 enam., close-wound on a 1/4-inch (.65 cm) diameter green-slug coil form.
L2	18 turns No. 26 enam., close-wound on a 3/8-inch (.95 cm) diameter red-slug coil form. Tap down 7 turns from top for 3N140 drain connection. See text.	12 turns No. 26 enam., close-wound on a 1/4-inch (.65 cm) diameter green-slug coil form. No tap used.
L3	Same as L2 but no tap, spaced 1-1/8 inch (2.9 cm) center-to-center with L2.	Same as L2, spaced 1 inch (2.5 cm) center-to-center with L2.
R1, R3	300 ohm, 1/4 W carbon.	430 ohm, 1/4 W carbon.
R2	16 ohm, 1/4 W carbon.	11 ohm, 1/4 W carbon.
Note:	Ferrite beads can be replaced by a 10-ohm, 1/4 W carbon resistor at one end of the choke, if desired.	
	SM = Silver Mica.	

be a useful tool for the amateur vhf/uhf receiver experimenter.

Appendix

Mixer Terminology

ƒrf – rf input frequency
ƒLO – local-oscillator input frequency
ƒi-f – i-f output frequency

By convention, mixing signals and their products are referred to the LO frequency for calculations. In the mixer system presented, ƒrf is always above ƒLO, so we will refer our signals to ƒrf, with the exception of Fig. 1 which uses the ƒLO reference.

Overload

A generic term covering most undesired operating phenomena associated with device nonlinearity.

Harmonic Modulation Products

Output responses caused by harmonics of ƒLO and ƒrf and their mixing products.

RF Compression Level

The absolute single-signal rf input-power level that causes conversion loss to increase by 1 dB.

RF Desensitization Level

The rf input power of an interfering signal that causes the small-signal conversion loss to increase by 1 dB, i.e. reducing a weak received signal by 1 dB.

Intermodulation Products

Distortion products caused by multiple rf signals and their harmonics mixing with each other and the LO, producing new output frequencies.

Mixer Intermodulation Intercept Point

Because mixers are nonlinear devices, all signals applied will generate others. When two signals (or tones), *F*1 and *F*2, are applied simultaneously to the rf-input port, additional signals are generated and appear in the output as ƒLO ± (nF1 + mF2). These signals are most troublesome when $n \pm m$ is a low odd number, as the resulting product will lie close to the desired output. For n-1 (or 2) and m–2 (or 1), the result is three (3), and is called the two-tone/third-order intermodulation products When *F*1 and *F*2 are separated by 1 MHz, the third-order products will lie 1 MHz above and below the desired outputs. Intermodulation is generally specified under anticipated operating conditions since performance varies over the broad mixer-frequency ranges. Intermodulation products may be specified at levels required (i.e. 50 dB below the desired outputs for two 0-dBm input signals) or by the intercept point.

The intercept point is a fictitious point determined by the fact that an increase of level of two input tones by 10 dB will cause the desired output to increase by 10 dB, but the third-order output will increase by 30 dB. If the mixer exhibited no compression, there would be a point at which the level of the desired output would be equal to that of the third-order product. This is called the third-order intercept point and is the point where the desired-output slopes and third-order slopes intersect (Fig. 7).

Noise Figure

Noise figure is a relative measurement based on excess noise power available from a termination (input resistor) at a particular temperature (290 degrees K). When measuring the NF of a double balanced mixer with an automatic system, such as the HP 342A, a correction may be necessary to make the meter reading consistant with the accepted definition of receiver noise figure.

In a broadband DBM, the actual noise bandwidth consists of two i-f passbands, one on each side of the local-oscillator frequency (ƒLO + ƒi-f and ƒLO – ƒi-f). This double sideband (dsb) i-f response includes the rf channel and its image. In general, only the rf channel is desired for further amplification. The image contributes nothing but receiver and background noise.

When making an automatic noise-figure measurement using a wideband noise source, the excess noise is applied through both sidebands in a broadband DBM. Thus the instrument meter indicates NF as based on both sidebands. This means that the noise in the rf and image sidebands is combined in the mixer i-f-output port to give a double contribution (3 dB greater than under ssb conditions). For equal rf-sideband responses, which is a reasonable assumption, and in the absence of preselectors, filters, or other image rejection elements, the automatic NF meter readings are 3 dB lower than the actual NF for DBM measurements.

The noise figure for receivers (and most DBM) is generally specified with only one sideband for the useful signal. As mentioned in the text, most DBM diodes add no more than 0.5 dB (in the form of NF) to conversion loss, which is generally measured under single-signal rf-input (ssb) conditions. Assuming DBM conversion efficiency (or loss) to be within specifications, there is an excellent probability that the ssb NF is also satisfactory. Noise figure calculations in the text were made using a graphical solution of the well known noise-figure formula:

$$fT = f_1 + \frac{f_2 - 1}{g_1}$$

converted to dB.

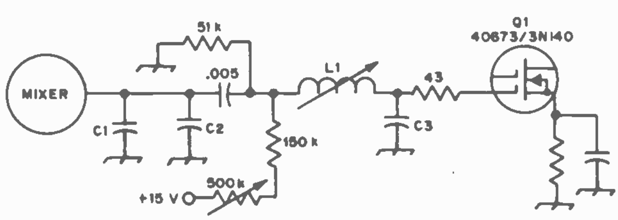
Improved Wide Band I-F Responses

The following information was developed in achieving broad-band performance in the mixer-to-amplifier circuitry. In cases where only a small portion of a band is of interest the original circuit values are adequate. For those who need to receive over a considerable portion of a band, say one to two MHz, a change of some component will provide improved performance over a broad range while maintaining an acceptable noise figure.

The term "nominal 50-ohm impedance" applied to diode DBM ports is truly a misnomer, as their reflective impedance is rarely 50 ohms $+j0$ and a VSWR of 1 is almost never achieved. Mixer performance specified by the manufacturer is measured in a 50-ohm broadband system, and it is up to the designer to provide an equivalent termination to ensure that the unit will meet specifications. Appropriate matching techniques at the rf and LO ports will reduce conversion loss and low-power requirements. Complex filter synthesis can improve the i-f output match. However, if one does not have the necessary equipment to evaluate his efforts, they may be wasted. Simple, effective, easily reproduced circuitry was desired as long as the trade-offs were acceptable, and measurements indicate this to be the case.

The most critical circuit in the combined unit is the interface between mixer and i-f amplifier. It must be low-pass in nature to satisfy vhf signal component bypassing requirements at the mixer i-f port. For best mixer IMD characteristics and low conversion loss, it must present to the i-f port a nominal 50-ohm impedance at the desired frequency, and this impedance value must not be allowed to increase as i-f operation departs from midband. The impedance at the i-f amplifier end of the interface network must be in the optimum region for minimum cross-modulation and low noise. A dual-gate device offers two important advantages over most bi-polars. Very little, if any, power gain is sacrificed in achieving best noise figure, and both parameters (gain and NF) are relatively independent of source resistance in the optimum region. As a result, the designer has a great deal of flexibility in choosing a source impedance. In general, a 3:1 change in source resistance results in only a 1-dB change in NF. With minimum cross-modulation as a prime system consideration, this 3:1 change (reduction) in source resistance implies a 3:1 improvement in cross-modulation and total harmonic distortion.

Tests on the 3N201 dual-gate MOSFET have shown device noise performance to be excellent for source impedances in the 1-kΩ to 2-kΩ region. For optimum noise and good cross-modulation



14 MHz	28 MHz
C1 – 300 pF (JFD 301).	100 pF (JFD 101).
C2 – 51 pF S.M.	Not used
C3 – 68 pF S.M.	7.5 pF S.M.
L1 – 15t No. 24 enameled on 3/8-inch dia red-slug form. 1.5-2.5 μH range, 1.95 μH for network.	19t No. 26 enameled on 1/4-inch dia red-slug form.

Fig. 1 – Suggested changes in the mixer-to 3N140 pi-network interface circuit, producing lower Q_L and better performance. See the original article for additional circuit details.

performance, the nominal 50-ohm mixer i-f output impedance is stepped up to about 1500 ohms for i-f amplifier gate 1, using the familiar low-pass pi network. This is a mismatched condition for gate 1, as the device input impedance for best gain in the hf region is on the order of 10 kΩ network loaded-Q values in the article are a bit higher than necessary, and a design for lower Q_L is preferred. Suggested modified component values are listed in Fig. 8. High-frequency attenuation is reduced somewhat, but satisfactory noise and bandwidth performance is more easily obtained. Coil-form size is the same, so no layout changes are required for the modification. Components in the interface must be of high Q and few in number to limit their noise contribution through losses. The 28-MHz values provide satisfactory interface network performance over a 2-MHz bandwidth. A higher Q_L in the 28-MHz interface can be useful if one narrows the output network and covers only a few hundred kilohertz bandwidth, as is commonly done in 432-MHz weak-signal work.

Device biasing and gain control methods were chosen for simplicity and adequate performance. Some sort of gain adjustment is desirable for drain-circuit overload protection. It is also a handy way to "set" the receiver S meter. A good method for gain adjustment is reduction of the gate-2 bias voltage from its initial optimum-gain bias point (greater than +4 V dc), producing a remote-cutoff characteristic (a gradual reduction in drain current with decreasing gate bias). The initial gain-reduction rate is higher with a slight forward bias on gate 1, than for $Vg_1s = 0$. Input and output circuit detuning resulting from gain reduction (Miller effect) is inconsequential as the gate-1 and drain susceptances change very little over a wide range of Vg_2s and I_D at both choices of i-f. Best intermodulation figure for the 3N201 was obtained with a small forward bias on gate 1, and the bias-circuit modification shown may be tried, if desired.

References

[1] Fisk, "Double-Balanced Mixers," *Ham Radio*, March, 1968.

[2] Ress, "Broadband Double-balanced Modulator," *Ham Radio*, March, 1970.

[3] DeMaw and McCoy; "Learning to Work With Semiconductors," Part IV, *QST*, July, 1974.

[6] Sabin, "The Solid-State Receiver," *QST*, July, 1970.

[7] Hayward, "A Competition-Grade CW Receiver," *QST*, March and April 1974.

AN OSCAR UP-CONVERTER

Many amateur operators who wish to receive the 10-meter signals from the Oscar satellites do so with an average receiver that is already "at hand" in the shack, sometimes adding a preamplifier for improved performance. Others use a converter to translate the signals to a lower frequency where the station receiver is more stable or more sensitive.

However, there is another approach that should be explored – that of converting the 10-meter signals *up* to a higher band. Just a very short time ago this system would have been impractical, if not ridiculous, because of the complexity and size of the equipment involved. Recent developments in two-meter transceivers make the up-conversion scheme practical and attractive. A nearly "ideal" Oscar package can be obtained by the addition of a small converter which allows a normal transceive style of operation with the vhf equipment.

There are several makes of 144-MHz ssb transceivers available, but only a few are beginning to appear on the market in the Western Hemisphere. The KLM Echo II was used here to evaluate the technique and test the performance of the converter that was assembled. This particular transceiver had been modified to permit cw operation as well as the usual ssb – a desirable feature to look for in any equipment being contemplated.

In some instances it may seem a bit redundant to convert a 28-MHz signal to 144 MHz, only to have it converted back down to 28 MHz in the receiver first mixer. However, there are reasons why this scheme is not all that bad, and a chief one is that the frequencies do not translate directly in all cases. A secondary, but important, consideration is that it may not be desirable or possible to modify the equipment to accept a 28-MHz input. And of course not all transceivers have 28 MHz as a first i-f.

The Converter

If the pc board and parts-placement layout appears familiar, it is because an existing design was modified to serve our purpose. Rather than go through the entire process of developing a new board the "Rochester" converter was rehabilitated for this project. See pages 300-304 for more details of these converters. Most suppliers of amateur radio pc boards have this pattern on hand, and many have etched and drilled boards in stock. There have been a few changes to some parts of the circuit, necessitating the placing of one capacitor on the foil side of the board. In operation the converter reverses the process of the original in that it first amplifies the 10-meter signals in Q1 (Fig. 1), then mixes them with 116.45-MHz energy in Q2, to provide an output between 145.85 and 145.95 MHz. The original oscillator and harmonic-generator circuit proved adequate with a slight modification; a third-overtone crystal with a frequency of 58.225 was used instead of the 38.6-MHz unit specified earlier. A buffer stage (Q4) is necessary to allow some rejection of unwanted harmonics while maintaining a suitable injection voltage for the mixer.

Construction

The assembly of the converter is greatly simpli-

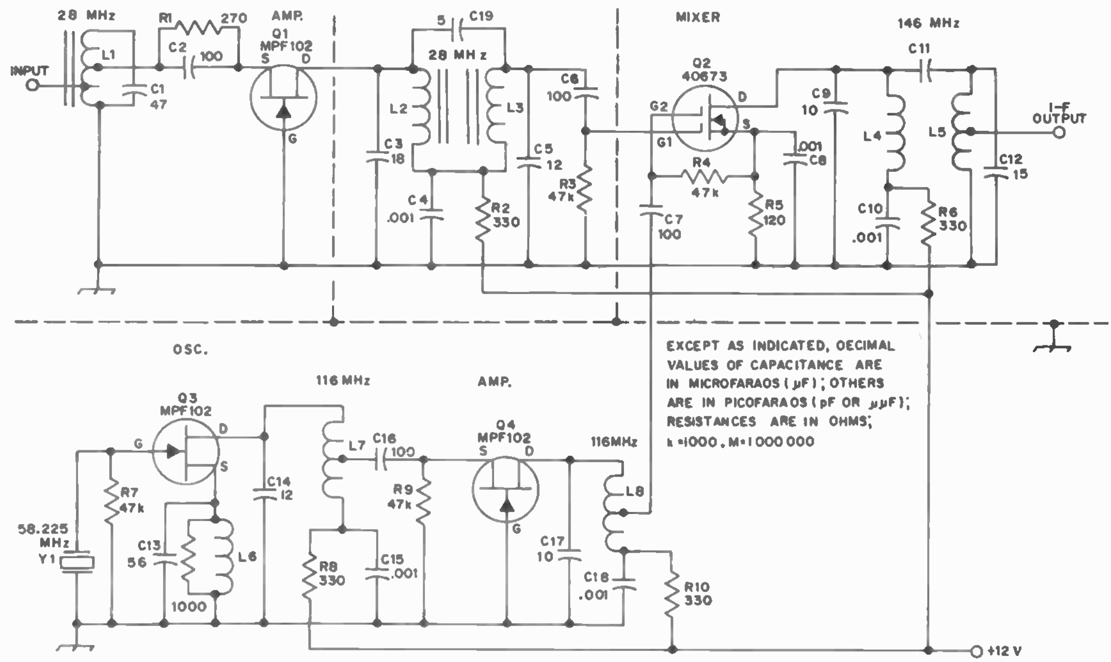


Fig. 1 – The schematic diagram of the "Rochester" converter as modified for up-conversion.

C11 – Two 1/2-inch pieces No. 18 insulated hookup wire, twisted together 1/2 turn.
L1, L2, L3 – 18 turns No. 28 enam. wound on Amidon T-30-6 core. L1 tapped at 6 turns and 11 turns from ground end.
L4, L5 – 5 turns No. 20 enam., formed by using threads of 1/4-20 bolt as a guide. L5 is tapped 2 turns from the ground end.
L6 – 10 turns No. 24 enam. close wound on the body of a 1000-ohm 1/2-watt resistor.
L7, L8 – 5 turns No. 20 enam., formed the same as L4. Both are tapped 2 turns from the hot end.
Y1 – 58.225-MHz crystal. International Crystal third-overtone type in FM-1 (wire leads) or FM-2 (pins) holder. Calibration tolerance of .0025% at a load capacitance of 20 pF used here. General-purpose (.01%) tolerance may be adequate in most instances.

fied by the use of a ready-made board, since for the most part the work consists of placing the component leads through the holes and soldering them in place.

Because of slight differences in materials used in toroids, it will be necessary to adjust the input and interstage tuned circuits to resonance after the converter is completed. This was done by the substitution method, placing different values of capacitance across the windings while monitoring the signal on a receiver. There are miniature trimmers available that would fit into the space, but the cost of these devices is a bit high. Alternatively, larger trimmers could be mounted below the pc board, if the builder allows enough space between the board and the enclosure to which it is fastened. Proper operation of the oscillator and buffer stages can be ascertained by using a grid-dip meter to indicate output on the correct harmonic of Y1.

There is sufficient space in some transceivers to allow the converter to be mounted internally, as was the case in the Echo II unit shown in the photographs. A small bracket was fastened to the rear apron of the equipment to provide a mount for the 28-MHz input connector. Where a transceiver is too compact to allow this style of assembly, the converter could be fastened in a small Minibox for shielding. A source of +12 V. and a small-diameter coax for input and output connections makes the wiring job simple. In some equipment it may be necessary to disconnect the receiver input wiring from the T/R relay but this was not done in the Echo II. The output from the converter was wired in parallel with the receiver input lead. Any small amount of noise picked up by the receiver was masked by the output from the converter. However, in areas of high local activity or high ambient noise levels, it will be necessary to disable the normal receiver input.

Performance

Sensitivity of the converter was sufficient that the residual output of a Model 80 signal generator

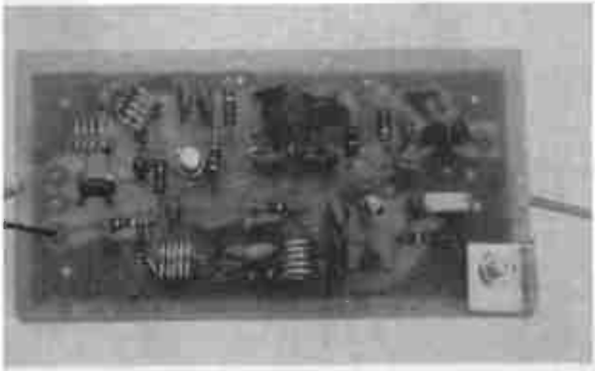
In this particular model transceiver, room was available to mount the converter in an inverted position just below the speaker. A small bracket is fastened under one of the bolts that also holds a transformer to the bottom sidewall. Cables are routed along with existing wiring harnesses and tied in place. The 28-MHz input connector is fastened to the rear apron of the equipment.

could be heard clearly. A 1-μV signal was loud enough that it evoked the immediate reaction of turning down the audio gain control on the transceiver. Ignition noise picked up by the variety of antennas tried was strong enough to be bothersome at times, further attesting to the sensitivity of the converter – it also pointed out the usefulness of the noise blanker in the Echo II. Considering the performance of the converter/transceiver combination, the addition of a preamplifier or an i-f post amplifier was not considered necessary. Additional gain could even be detrimental by causing overloading or intermodulation problems – there was no evidence of these problems during several tests.

Since the first i-f in the transceiver is at 28 MHz, the question of possible "leak-through" of local signals was raised. No indication of this type of interference was found during receiving tests, but admittedly it could happen. The output circuit of the mixer (L4, L5, C11) has a band-pass characteristic centered on 146 MHz and should provide a high degree of attenuation to hf-band signals.

In areas where strong local operation does cause such leakage of signals through the converter it will be necessary to install a high-pass filter between the converter output and the receiver input. Designs for such filters can be found in the ARRL *Handbook*. Because the filter will be used at essentially zero power level, it can be made physically quite compact. Of course good shielding and high-quality coaxial cable is a must in any effort to keep unwanted signals out – the best filters in the world will do no good if there is a path around them.

[EDITOR'S NOTE: The parts placement for this up-converter is virtually identical to that used in the "Rochester" converters, found elsewhere in this chapter. The reader can follow that layout, keeping in mind the differences in tuned-circuit frequencies throughout.]



Top view of the modified Rochester Converter. A mounting bracket has been fastened to the lower right corner. The 28-MHz input is to the upper right, with the rf amplifier along the top of the board. Oscillator and buffer stages are located along the bottom portion. The i-f output coils, L4 and L5, are at the upper left with C11 (twisted wires) just below L4, adjacent to the resistor. The shields between stages have been omitted for a better view.

INTERDIGITAL CONVERTER FOR 1296 OR 2304 MHz

In a world where rf spectrum pollution is becoming more serious, even into the microwave region, it is almost as important to keep unwanted signals out of a receiver as it is to prevent radiation of spurious energy. An interdigital filter was described some years ago, featuring low insertion loss, simplicity of construction, and reasonable rejection to out-of-band signals.[1] It could be used in either transmitters or receivers.

This twice-useful principle has now been put to work again – as a mixer. Again, the ease of construction and adaptation leads many to wonder that it had not been thought of before. It was first described by W2CQH in *QST* for January, 1974.

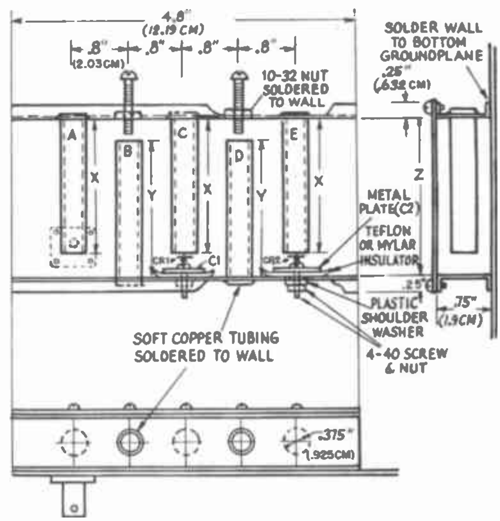
A Filter and Mixer

A layout of the microwave portions of both converters is shown in Fig. 1. The structure consists of five interdigitated round rods, made of 3/8-inch OD brass or copper tubing. They are soldered to two sidewalls and centrally located between two ground-planes made of 1/16-inch sheet brass or copper-clad epoxy fiber glass. One ground plane is made larger than the microwave assembly and thus provides a convenient mounting plate for the remainder of the converter components.

The sidewalls are bent from .032-inch thick sheet brass or they can be made from 1/4 × 3/4-inch brass rod. One edge of each sidewall is soldered to the larger ground plane. The other edge is fastened to the smaller ground plane by 4-40 machine or self-tapping screws, each located over the centerline of a rod. The sidewall edges should be sanded flat, before the ground plane is attached, to assure continuous electrical contact. Note that no end walls are required since there are no electric fields in these regions.

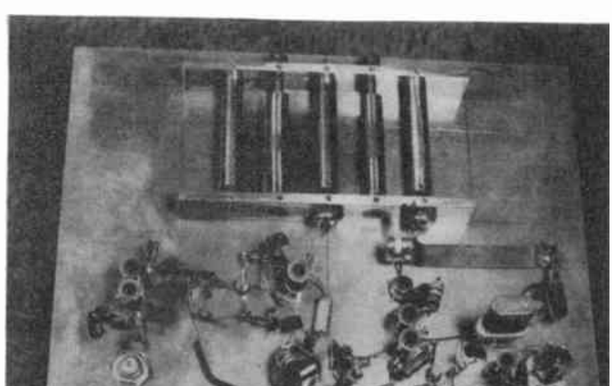
Electrically, rods A, B, and C comprise a one-stage, high-loaded-Q (Q_L = 100), interdigital filter[1] which is tuned to the incoming signal frequency near 1296 or 2304 MHz. The ungrounded end of rod A is connected to a BNC coaxial connector and serves as the coupling section to the filter input. Rod B is the high-Q

[1] Fisher, "Interdigital Bandpass Filters for Amateur VHF/UHF Applications," *QST* March, 1968.



ENCLOSURE DIMENSIONS			
FREQ.	X	Y	Z
1296 MHz	2.00" (5.08CM)	2.00" (5.08CM)	2.25" (5.71CM)
2304 MHz	.9375" (2.38CM)	1.00" (2.54CM)	1.25" (3.17CM)

Fig. 1 — Dimensions and layout for the filter and mixer portions of the interdigital converters. The signal input is to the left rod, labelled "A." Local-oscillator injection is through the diode to rod "E." CR1 is the mixer diode, connected to the center rod in the assembly.



The converter for 1296 MHz. This unit was built by R.E. Fisher, W2CQH. While the mixer assembly (top center) in this model has solid brass walls, it can be made from lighter material as explained in the text and shown in Fig. 1. The i-f amplifier is near the center, just above the mixer-current-monitoring jack, J1. A BNC connector at the lower left is for 28-MHz output. The local oscillator and multiplier circuits are to the lower right. Note that L6 is very close to the chassis, just above the crystal. The variable capacitor near the crystal is an optional trimmer to adjust the oscillator to the correct frequency.

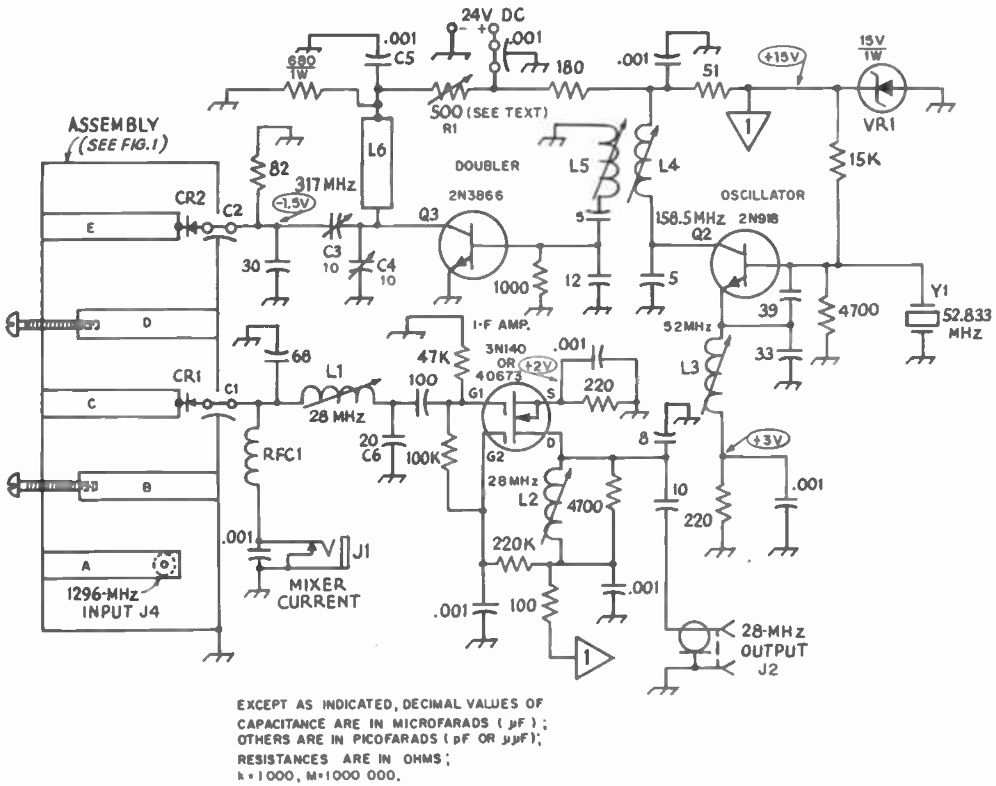


Fig. 2 — Schematic diagram of the 1296-MHz converter with oscillator and multiplier sections included. Dimensions for the filter and mixer assembly are given in Fig. 1.

C1, C2 — 30-pF homemade capacitor. See text and Fig. 1.
C3, C4 — 0.8- to 10-pF glass trimmer, Johanson 2945 or equiv.
C5 — .001-µF button mica.
C6 — 2- to 20-pF air variable, E. F. Johnson 189-507-004 or equiv.
CR1 — Hewlett Packard 5082-2577 or 5082-2835.
CR2 — Hewlett Packard 5082-2811 or 5082-2835.
J1 — Closed-circuit jack.
J2 — Coaxial connector, chassis mount. Type BNC acceptable.
L1, L2 — 18 turns No. 24 enam. on 1/4-inch OD slug-tuned form (1.5 µH nominal).
L3 — 10 turns like L1 (0.5 µH).
L4, L5 — 6 turns like L1 (0.2 µH).
L6 — Copper strip, 1/2-inch wide × 2-1/2-inches (1.27 × 6.35 cm) long. See text and photographs.
RFC1 — 33 µH, J. W. Miller 74F33SAI or equiv.

resonator and is tuned by a 10-32 machine screw. Rod C provides the filter output-coupling section to the mixer diode, CR1.

The mixer diode is a Hewlett-Packard 5082-2577 Schottkey-barrier type which is available from distributors for about $4. The cheaper 5082-2835, selling for 90 cents, can be used instead, but this substitution will increase the 2304-MHz mixer noise figure by approximately 3 dB.

One pigtail lead of the mixer diode is tack-soldered to a copper disk on the ungrounded end of rod C. Care should be taken to keep the pigtail lead as short as possible. If rod C is machined from solid brass stock, then it is feasible to clamp one of the mixer-diode leads to the rod end with a small setscrew. This alternative method facilitates diode substitution and was used in the mixer models shown in the photographs.

Fig. 1 also shows that the other end of CR1 is connected to a homemade 30-pF bypass capacitor, C1, which consists of a 1/2-inch-square copper or brass plate clamped to the sidewall with a 4-40 machine screw. The dielectric material is a small screw passes through an oversize hole and is insulated from the other side of the wall by a small plastic shoulder washer.

In the first converter models constructed by the author and shown in the photographs, C1 was a 30-pF button mica unit soldered to the flange of a 3/8-inch diameter threaded panel bearing (H.H. Smith No. 119). The bearing was then screwed into a threaded hole in the sidewall. This provision made it convenient to measure the insertion loss and bandwidth of the interdigital filters since the capacitor assembly could be removed and replaced with a BNC connector.

Rods C, D, and E comprise another high loaded-Q (Q_L = 100) interdigital filter tuned to the local oscillator (LO) frequency. This filter passes only the fourth harmonic (1268 or 2160 MHz) from the multiplier diode, CR2. The two filters have a common output-coupling section (rod C) and their loaded Qs are high enough to prevent much unwanted coupling of signal power from the antenna to the multiplier diode and LO power back out to the antenna.

The multiplier diode is connected to the driver circuitry through C2, a 30-pF bypass capacitor identical to C1. CR2 is a Hewlett-Packard 5082-2811 although the 5082-2835 works nearly as well. Fifty milliwatts drive at one quarter of the LO frequency is sufficient to produce 2 mA of mixer diode current, which represents about 1 milliwatt of the local-oscillator injection. A Schottkey-barrier was chosen over the more familiar varactor diode for the multiplier because it is cheaper, more stable, and requires no idler circuit.

Table I

Converter Specifications

	1296 MHz	*2304 MHz*
Noise figure	5.5 dB	6.5 dB
Conversion gain	20 dB	14 dB
3-dB bandwidth	2 MHz	7 MHz
Image rejection	18 dB	30 dB
I-f output	28 MHz	144 MHz

Fig. 2 shows the schematic diagram of the 1296 to 28 MHz converter. All components are mounted on a 7 × 9-inch (17.8 × 22.9 cm) sheet of brass or copper-clad epoxy-fiber glass board. As mentioned earlier, this mounting plate also serves as one ground plane for the microwave mixer. When completed, the mounting plate is fastened to an inverted aluminum chassis which provides a shielded housing.

Oscillator and Multipliers

The nonmicrowave portion of the converter is rather conventional. Q1, a dual-gate MOSFET, was chosen as the 28-MHz i-f amplifier since it can provide 25 dB of gain with a 1.5-dB noise figure. The mixer diode is coupled to the first gate of Q1 by a pi-network matching section. It is most important that the proper impedance match be achieved between the mixer and i-f amplifier if a low noise figure is to be obtained. In this case, the approximately 30-ohm output impedance of the mixer must be stepped up to about 1500 if Q1 is to yield its rated noise figure of 1.5 dB. It is for this reason that a remote i-f amplifier was not employed as is the case with many contemporary uhf converters.

Q2 functions in a oscillator-tripler circuit which delivers about 10 milliwatts of 158.5-MHz drive to the base of Q3. The emitter coil, L3, serves mainly as a choke to prevent the crystal from oscillating at its fundamental frequency. Coils L4 and L5, which are identical, should be spaced closely such that their windings almost touch.

Q3 doubles the frequency to 317 MHz, providing about 50 milliwatts drive to the multiplier diode. It is important that the emitter lead of Q3 be kept extremely short; 1/4-inch (6.36 mm) is probably too long. L6, the strip-line inductor in the collector circuit of Q3, consists of a 1/2 × 2-1/2-inch (1.27 × 6.35 cm) piece of flashing copper spaced 1/8-inch (3.18 mm) above the ground plane. The cold end of L6 is bypassed to ground by C5, a .001-μF button mica capacitor.

The multiplier circuits are tuned to resonance in the usual manner by holding a wavemeter near each inductor being tuned. Resonance in the Q3 collector circuit is found by touching a VTVM probe (a resistor must be in the probe) to C2 and adjusting the Johanson capacitors until about −1.5 volts of bias is obtained. The 317- to 1268-MHz multiplier cavity is then resonated by adjusting the 10-32 machine screw until maximum mixer current is measured at J1. When resonance is found, R1 should be adjusted so that about 2 mA of mixer current is obtained. As an alternative to mounting a potentiometer in the converter, once a value of resistance has been found that provides correct performance it can be measured and the nearest standard fixed-value resistor substituted. Some means of adjusting the collector voltage on the multiplier stage must be provided initially to allow for the nonuniformity of transistors.

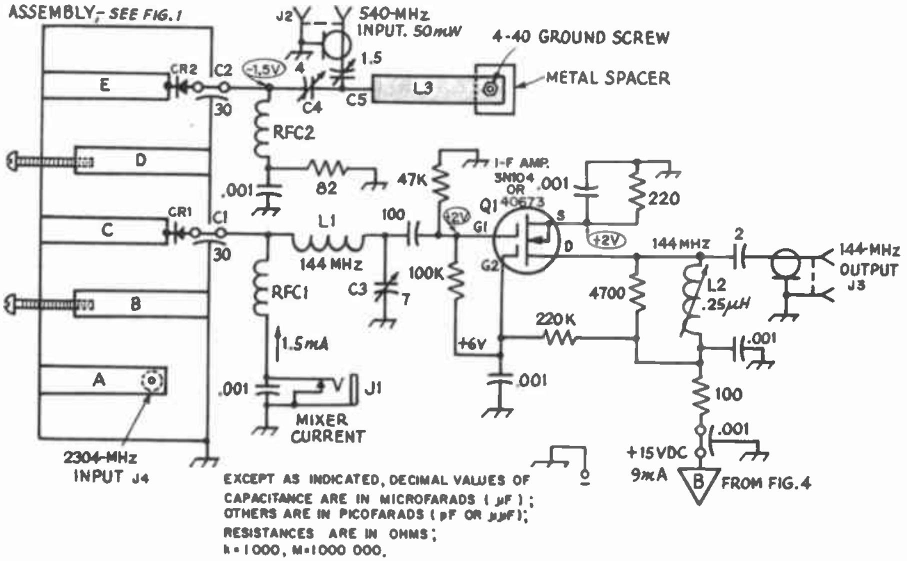
A 2304-MHz Version

Fig. 3 and 4 show the schematic diagrams of the 2304-MHz converter and multiplier. The mixer and i-f preamplifier was built on a separate chassis since, at the time of their construction, a multiplier chain from another project was available. An i-f of 144 MHz was chosen although 50 MHz would work as well. An i-f output of 28 MHz, or lower, should not be used since this would result in undesirable interaction between the mixer and multiplier interdigital filters.

The 2304-MHz mixer and i-f amplifier section, shown in Fig. 3, is very similar to its 1296-MHz counterpart. Q1, the dual-gate MOSFET, operates at 144 MHz and thus has a noise figure about 1 dB higher than that obtainable at 28 MHz.

The multiplier chain, Fig. 4, has a separate oscillator for improved drive to the 2N3866 output stage. Otherwise the circuitry is similar to the 1296-MHz version.

Fig. 3 – Schematic diagram of the 2304-MHz version of the converter, with the i-f amplifier. The oscillator and multiplier circuits are constructed separately.
C1, C2 – 30-pF homemade capacitor. See text.
C3, C4, C5 – 0.8- to 10-pF glass trimmer, Johanson 2945 or equiv.
CR1 – Hewlett Packard 5082-2577 or 5082-2835.
CR2 – Hewlett Packard 5082-2811 or 5082-2835.
J1 – Closed-circuit jack.
J2, J3, J4 – Coaxial connector, chassis mount. Type BNC.
L1 – 5 turns No. 20 enam., 1/4-inch ID × 1/2-inch long. (6.35 × 12.7 mm).
L2 – 6 turns No. 24 enam., on 1/4-inch OD slug tuned form (0.25 μH).
L3 – Copper strip 1/2-inch wide × 2-11/16 inches (1.27 × 6.86 cm) long. See text and photographs.
RFC1 – Ohmite Z-144 or equiv.
RFC2 – Ohmite Z-460 or equiv.



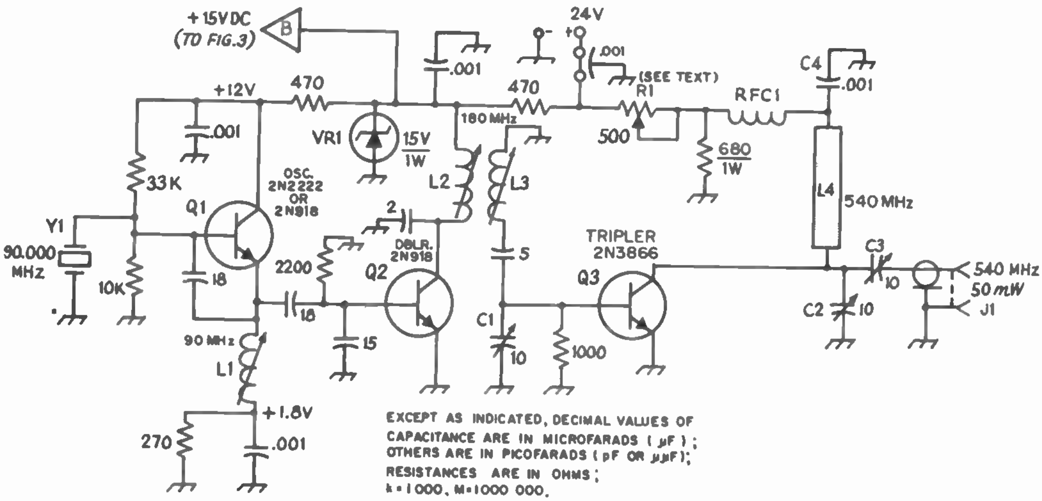


Fig. 4 – Schematic diagram of the oscillator and multiplier for the 2304-MHz converter. As explained in the text and shown in the photographs, a fixed-value resistor may be substituted for R1 after the value that provides proper performance has been found.
C1, C2, C3 – 0.8- to 10-pF glass trimmer, Johanson 2945 or equiv.
C4 – .001-μF button mica.
J1 – Coaxial connector, chassis mount, type BNC or equiv.
L1 – 10 turns No. 24 enam. on 1/4-inch OD slug-tuned form.
L2, L3 – 3 turns like L1.
L4 – Copper strip 1/2-inch wide × 1-1/2 inches (1.27 × 3.81 cm) long. Space 1/8 inch (3.18 mm) from chassis.
RFC1 – 10 turns No. 24 enam. 1/8-inch ID, closewound.

Mobile and Portable/Emergency Equipment and Practices

MOBILE AND PORTABLE EQUIPMENT

Amateur mobile and portable operation provides many opportunities for one to exercise his skill under less than ideal conditions. Additionally, the user of such equipment is available for public-service work when emergencies arise in his community – an important facet of amateur-radio operation. Operating skill must be better than that used at most fixed locations because the mobile/portable operator must utilize inferior antennas, and must work with low-power transmitters in many instances.

Most modern-day hf-band mobile work is done while using the ssb mode. Conversely, the fm mode is favored by mobile and portable vhf operators, though ssb is fully practical for vhf service. Some amateurs operate cw mobile, much to the consternation of local highway patrolmen, but cw operation from a *parked car* should not be overlooked during emergency operations.

High-power mobile operation has become practical on ssb because of the low duty cycle of voice operation, and because low-drain solid-state mobile power supplies lessen battery drain over that of dynamotors or vibrator packs. Most mobile a-m and fm operation is limited to 60 watts for reasons of battery drain.

Portable operation is popular on ssb, cw and fm while using battery-powered equipment. Ordinarily, the power of the transmitter is limited to less than five-watts dc input for practical reasons. Solid-state equipment is the choice of most modern amateurs because of its compactness, reliability, and low power consumption. High-power portable operation is practical and desirable when a gasoline-powered ac generator is employed.

The secret of successful operation from portable sites is much the same as that from a fixed station – a good antenna, properly installed. Power levels as low as 0.5 watt are sufficient for covering thousands of miles during hf-band ssb and cw operation. In the vhf and uhf region of operation it is common to work distances in excess of 100 miles – line of sight – with less than one watt of transmitter output power. Of course it is important to select a high, clear location for such operation on vhf, and it is beneficial to use an antenna with as much gain as is practical. Low-noise receiving equipment is the ever-constant companion of any low-power portable transmitter that provides successful long-distance communications. Careful matching of the portable or mobile antenna to obtain the lowest possible SWR is another secret of the successful operator.

All portable and mobile equipment should be assembled with more than ordinary care, assuring that maximum reliability under rough-and-tumble conditions will prevail. All solder joints should be made well, stranded hookup wire should be used for cabling (and in any part of the equipment subjected to stress). The cabinets for such gear should be rugged, and should be capable of protecting the components from dust, dirt, and moisture.

ELECTRICAL-NOISE ELIMINATION

One of the most significant deterrents to effective signal reception during mobile or portable operation is electrical impulse noise from the automotive ignition system. The problem also arises during the use of gasoline-powered portable ac generators. This form of interference can completely mask a weak signal, thus rendering the station ineffective. Most electrical noise can be eliminated by taking logical steps toward suppress-

Fig. 10-1 – Effective portable operation can be realized when using lofty locations for vhf or uhf. Here, W1CKK is shown operating a battery-powered, 150-mW output, 2-meter transceiver. With only a quarter wavelength antenna it is possible to communicate with stations 25 miles or more away. Low-power transistor equipment like this unit will operate many hours from a dry-cell battery pack.

Fig. 10-2 – High-power portable/emergency operation can be made possible on all amateur bands by using vacuum-tube transmitters, and powering them from a gasoline-operated ac generator of one or more kW rating. (Shown here is VE7ARV/7 during a Field Day operation.)

ing it. The first step is to clean up the *noise source* itself, then utilize the receiver's built-in noise-reducing circuit as a last measure to knock down any noise pulses from passing cars, or from other man-made sources.

Spark-Plug Noise

Spark-plug noise is perhaps the worst offender when it comes to ignition noise. There are three methods of eliminating this type of interference – resistive spark-plug suppressors, resistor spark plugs, or resistance-wire cabling. By installing Autolite resistor plugs a great deal of the noise can be stopped. Tests have proved, however, that suppressor cable between the plugs and the distributor, and between the distributor and ignition coil, is the most effective means of curing the problem. Distributed-resistance cable has an approximate resistance of 5000 ohms per foot, and consists of a carbon-impregnated sheath followed by a layer of insulation, then an outer covering of protective plastic sheathing. Some cars come equipped with suppressor cable. Those which do not can be so equipped in just a matter of minutes. Automotive supply stores sell the cable, and it is not expensive. It is recommended that this wiring be used on all mobile units. The same type of cable can be installed on gasoline-powered generators for field use. A further step in eliminating plug noise is the addition of shielding over each spark-plug wire, and over the coil lead. It should be remembered that each ignition cable is an antenna by itself, thus radiating those impulses passing through it. By fitting each spark-plug and coil lead with the shield braid from a piece of RG-59/U coax line, grounding the braid at each end to the engine block, the noise reduction will be even greater. An additional step is to encase the distributor in flashing copper, grounding the copper to the engine block. This copper is quite soft and can be form-fit to the contour of the distributor. (Commercially-manufactured shielded ignition cable kits are also available.) The shield braid of the spark-plug wires should be soldered to the distributor shield if one is used. Also, the ignition coil should be enclosed in a metal shield since the top end of many of these coils is made of plastic. A small tin can can often be used as a top cover for the coil or distributor. It should be soldered to the existing metal housing of the coil. Additional reduction in spark-plug noise can be effected by making certain that the engine hood makes *positive* contact with the frame of the car when it is closed, thus offering an additional shield over the ignition system The engine block should also be bonded to the frame at several points. This can be done with the shield braid from coax cable. Feedthrough (hi-pass) capacitors should be mounted on the coil shield as shown in Fig. 10-6 to filter the two small leads leaving the assembly.

Other Electrical Noise

The automotive generator system can create an annoying type of interference which manifests itself as a "whine" when heard in the receiver. This noise results from the brushes sparking as the commutator passes over them. A dirty commutator is frequently the cause of excessive sparking, and can be cleaned up by polishing its surface with a fine grade of emery cloth. The commutator grooves should be cleaned out with a small, pointed instrument. A coaxial feedthrough capacitor of 0.1- to 0.5-μF capacitance should be mounted on the generator frame and used to filter the generator *armature* lead. In stubborn cases of generator noise a parallel *L/C* tuned trap can be used in place of the capacitor, or in addition to it, tuned to the receiver's operating frequency. This is probably the most effective measure used for curing generator noise.

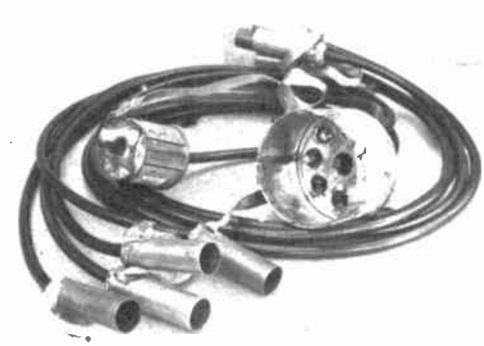


Fig. 10-3 – A typical homemade shielding kit for an automotive ignition system. Tin cans have been put to use as shields for the spark coil and distributor. Additional shields have been mounted on the plug ends of the wires for shilding the spark plugs. The shield braid of the cabling protrudes at each end of the wires and is grounded to the engine block.

Fig. 10-4 — A close-up view of the distributor shield can. The shield braid over each spark-plug wire is soldered to the top of the can, and the can is grounded to the engine block.

Voltage regulators are another cause of mobile interference. They contain relay contacts that jitter open and closed when the battery is fully charged. The noise shows up in the receiver as a ragged, "hashy" sound. Coaxial feedthrough capacitors can be mounted at the *battery* and *armature* terminals of the regulator box to filter those leads. The *field* terminal should have a small capacitor and resistor, series-connected, from it to chassis ground. The resistor prevents the regulator from commanding the generator to charge constantly in the event the bypass capacitor short-circuits. Such a condition would destroy the generator by causing overheating.

Alternators should be suppressed in a similar manner as dc generators. Their slip rings should be kept clean to minimize noise. Make sure the brushes are making good contact inside the unit. A coaxial feedthrough capacitor and/or tuned trap should be connected to the output terminal of the alternator. Make certain that the capacitor is rated to handle the output current in the line. The same rule applies to dc generators. *Do not connect a capacitor to the alternator or generator field terminals.* Capacitor values as high as 0.5 μF are suitable for alternator filtering.

Some alternator regulator boxes contain solid-state circuits, while others use single or double contact relays. The single-contact units require a coaxial capacitor at the *ignition* terminal. The double-contact variety should have a second such capacitor at the *battery* terminal. If noise still persists, try shielding the field wire between the regulator and the generator or alternator. Ground the shield at both ends.

Instrument Noise

Some automotive instruments are capable of creating noise. Among these gauges and senders are the heat- and fuel-level indicators. Ordinarily, the addition of a 0.5-μF coaxial capacitor at the sender element will cure the problem.

Other noise-gathering accessories are turn signals, window-opener motors, heating-fan motors and electric windshield-wiper motors. The installation of a 0.25-μF capacitor will usually eliminate their interference noise.

Frame and Body Bonding

Sections of the automobile frame and body that come in contact with one another can create additional noise. Suspected areas should be bonded together with flexible leads such as those made from the shield braid of RG-8/U coaxial cable. Trouble areas to be bonded are:

1 – Engine to frame.
2 – Air cleaner to engine block.
3 – Exhaust lines to car fame and engine block.
4 – Battery ground terminal to frame.
5 – Steering column to frame.
6 – Hood to car body.
7 – Front and rear bumpers to frame.
8 – Tail pipe to frame.
9 – Trunk lid to frame.

Wheel and Tire Static

Wheel noise produces a ragged sounding pulse in the mobile receiver. This condition can be cured by installing static-collector springs between the spindle bolt of the wheel and the grease-retainer cap. Insert springs of this kind are available at automotive supply stores.

Tire static has a ragged sound too, and can be detected when driving on hard-surface highways. If the noise does not appear when driving on dirt roads it will be a sure indication that tire static exists. This problem can be resolved by putting

Fig. 10-5 — Gasoline-powered ac generators used for portable/emergency operation should be treated for ignition noise in the same manner as automobile engines are. The frame of the gas generator should be connected to an earth ground, and the entire unit should be situated as far from the operating position as possible. This will not only reduce ignition noise, but will minimize ambient noise from the power unit. (Shown here is K1GTK during Field Day operations.)

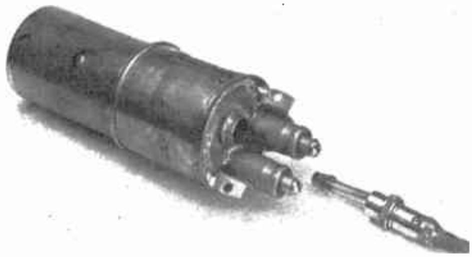


Fig. 10-6 – The automobile ignition coil should be shielded as shown here. A small tin can has been soldered to the metal coil case, and coaxial feed through capacitors have been soldered to the top of the can. The "hot" lead of the coil enters the shield can through a modified audio connector.

antistatic powder inside each tire. This substance is available at auto stores, and comes supplied with an injector tool and instructions.

Corona-Discharge Noise

Some mobile antennas are prone to corona build-up and discharge. Whip antennas which come to a sharp point will sometimes create this kind of noise. This is why most mobile whips have steel or plastic balls at their tips. But, regardless of the structure of the mobile antenna, corona build-up will frequently occur during or just before a severe electrical storm. The symptoms are a high-pitched "screaming" noise in the mobile receiver, which comes in cycles of one or two minutes duration, then changes pitch and dies down as it discharges through the front end of the receiver. The condition will repeat itself as soon as the antenna system charges up again. There is no cure for this condition, but it is described here to show that it is not of origin within the electrical system of the automobile.

Electronic Noise Limiters

Many commercially built mobile transceivers have some type of built-in noise clipping or cancelling circuit. Those which do not can be modified to include such a circuit. The operator has a choice of using af or rf limiting. Circuits of this type are described in the theory section of the hf receiving chapter.

Simple superregenerative receivers, by nature of their operation, provide noise-limiting features, and no additional circuit is needed. Fm receivers, if operating properly, do not respond to noise pulses of normal amplitude; hence no additional circuitry is required.

THE MOBILE ANTENNA

The antenna is perhaps the most important item in the successful operation of the mobile installation. Mobile antennas, whether designed for single or multiband use, should be securely mounted to the automobile, as far from the engine compartment as possible (for reducing noise pickup), and should be carefully matched to the coaxial feed line which connects them to the transmitter and receiver. All antenna connections should be tight and weatherproof. Mobile loading coils should be protected from dirt, rain, and snow if they are to maintain their Q and resonant frequency. The greater the Q of the loading coil, the better the efficiency, but the narrower will be the bandwidth of the antenna system.

Though bumper-mounted mobile antennas are favored by some, it is better to place the antenna mount on the rear deck of the vehicle, near the rear window. This locates the antenna high and in the clear, assuring less detuning of the system when the antenna moves to and from the car body. *Never use a base-loaded antenna on a bumper mount* if an efficient system is desired. Many operators avoid cutting holes in the car body for fear of devaluation when selling the automobile. Such holes are easily filled, and few car dealers, if any, lower the trade-in price because of the holes.

The choice of base or center loading a mobile antenna has been a matter of controversy for many years. In theory, the center-loaded whip presents a slightly higher base impedance than does the base-loaded antenna. However, with proper impedance-matching techniques employed there is no discernible difference in performance between the two methods. A base-loading coil requires fewer turns of wire than one for center loading, and this is an electrical advantage because of reduced coil

Fig. 10-7 – Here a mobile station is used as a portable/emergency station. As such, it can be connected to a full-size stationary antenna for maximum effectiveness. The engine should be noise-suppressed, and should be kept running during operation of the station to assure full battery power. (WA3EQK operating.)

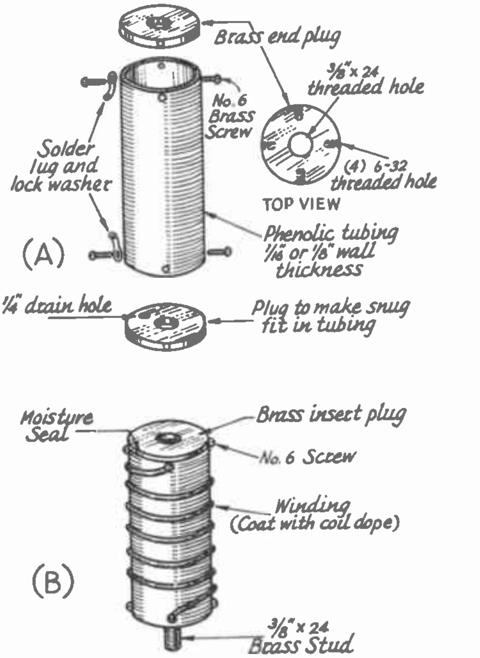


Fig. 10-8 – Details for making a home-built mobile loading coil. A breakdown view of the assembly is given at A. Brass end plugs are snug-fit into the ends of the phenolic tubing, and each is held in place by four 6-32 brass screws. Center holes in the plugs are drilled and tapped for 3/8-24 thread. The tubing can be any diameter from one to four inches. The larger diameters are recommended. Illustration B shows the completed coil. Resonance can be obtained by installing the coil, applying transmitter power, then pruning the turns until the lowest SWR is obtained. Pruning the coil for maximum field-strength-meter indication will also serve as a resonance indication. The chart in Fig. 10-10 will serve as a guide in determining the number of turns required for a given frequency of operation.

losses. A base-loaded antenna is more stable from the standpoint of wind loading and sway. If a homemade antenna system is contemplated, either system will provide good results, but the base-loaded antenna may be preferred for its mechanical advantages.

Loading Coils

There are many commercially built antenna systems available for mobile operation, and some manufacturers sell the coils as separate units. Air-wound coils of large wire diameter are excellent for use as loading inductors. Large Miniductor coils can be installed on a solid phenolic rod and used as loading coils. Miniductors, because of their turns spacing, are easy to adjust when resonating the mobile antenna, and provide excellent circuit *Q*. Phenolic-impregnated paper or fabric tubing of large diameter is suitable for making homemade loading coils. It should be coated with liquid fiber glass, inside and out, to make it weather proof. Brass insert plugs can be installed in each end, their centers drilled and tapped for a 3/8 X 24 thread to accommodate the mobile antenna sections. After the coil winding is pruned to resonance it should be coated with a high-quality, low-loss compound to hold the turns securely in place, and to protect the coil from the weather. Liquid polystyrene is excellent for this. It can be made by dissolving chips of solid polystyrene in carbon-tetrachloride. *Caution*: Do not breathe the chemical fumes, and do not allow the liquid to come in contact with the skin. *Carbon tetrachloride is hazardous to health.* Dissolve sufficient polystyrene material in the liquid to make the remaining product the consistency of Q-dope or pancake syrup. Details for making a home-built loading coil are given in Fig. 10-8.

Impedance Matching

Fig. 10-9 illustrates the shunt-feed method of obtaining a match between the antenna and the coaxial feed line. For operation on 75 meters with a center-loaded whip, L2 will have approximately 18 turns of No. 14 wire, spaced one wire thickness between turns, and wound on a 1-inch diameter form. Initially, the tap will be approximately 5 turns above the ground end of L2. Coil L2 can be inside the car body, at the base of the antenna, or it can be located at the base of the whip, outside the car body. The latter method is preferred. Since L2 helps determine the resonance of the overall antenna, L1 should be tuned to resonance in the desired part of the band with L2 in the circuit. The

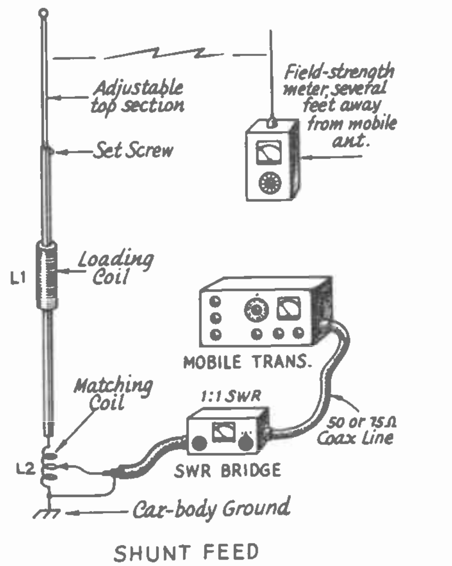


Fig. 10-9 – A mobile antenna using shunt-feed matching. Overall antenna resonance is determined by the combination of L1 and L2. Antenna resonance is set by pruning the turns of L1, or adjusting the top section of the whip, while observing the field-strength meter or SWR indicator. Then, adjust the tap on L2 for lowest SWR.

Approximate Values for 8-foot Mobile Whip

f kHz	Loading LµH	Rc(Q50) Ohms	Rc(Q300) Ohms	RR Ohms	Feed R* Ohms	Matching LµH*
Base Loading						
1800	345	77	13	0.1	23	3
3800	77	37	6.1	0.35	16	1.2
7200	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						
1800	700	158	23	0.2	34	3.7
3800	150	72	12	0.8	22	1.4
7200	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.

Suggested coil dimensions for the required loading inductance are shown in a following table.

Fig. 10-10 — Chart showing inductance values used as a starting point for winding homemade loading coils. Values are based on an approximate base-loaded whip capacitance of 25 pF, and a capacitance of 12 pF for center-loaded whips. Large-diameter wire and coils, plus low-loss coil forms, are recommended for best Q.

TABLE 10-I

Suggested Loading-Coil Dimensions

Req'd L µH	Turns	Wire Size	Dia In.	Length In.
700	190	22	3	10
345	135	18	3	10
150	100	16	2 1/2	10
77	75	14	2 1/2	10
77	29	12	5	4 1/4
40	28	16	2 1/2	2
40	34	12	2 1/2	4 1/4
20	17	16	2 1/2	1 1/4
20	22	12	2 1/2	2 3/4
8.6	16	14	2	2
8.6	15	12	2 1/2	3
4.5	10	14	2	1 1/4
4.5	12	12	2 1/2	4
2.5	8	12	2	2
2.5	8	6	2 3/8	4 1/2
1.25	6	12	1 3/4	2
1.25	6	6	2 3/8	4 1/2

adjustable top section of the whip can be telescoped until a maximum reading is noted on the field-strength meter. The tap is then adjusted on L2 for the lowest reflected-power reading on the SWR bridge. Repeat these two adjustments until no further increase in field strength can be obtained; this point should coincide with the lowest SWR. The number of turns needed for L2 will have to be determined experimentally for 40- and 20-meter operation. There will be proportionately fewer turns required.

MATCHING WITH AN *L* NETWORK

Any mobile antenna that has a feed-point impedance less than the characteristic impedance of the transmission line can be matched to the line by means of a simple L network, as shown in Fig. 10-11. The network is composed of C_M and L_M. The required values of C_M and L_M may be determined from the following:

$$C_M = \frac{\gamma\, R_A(R_0 - R_A) \times 10^9}{2\pi f\,\mathrm{kHz}\, R_A R_0}\ \mathrm{pF\ and}$$

$$L_M = \frac{\gamma\, R_A(R_0 - R_A) \times 10^3}{2\pi f\,\mathrm{kHz}}\ \mu\mathrm{H}$$

where R_A is the antenna feed-point impedance and R_0 is the characteristic impedance of the transmission line.

As an example, if the antenna impedance is 20 ohms and the line is 50-ohm coaxial cable, then at 4000 kHz,

$$C_M = \frac{\sqrt{20\,(50-20)} \times 10^9}{(6.28)\,(4000)\,(20)\,(50)}$$

$$= \frac{\sqrt{600} \times 10^4}{(6.28)\,(4)\,(2)\,(5)}$$

$$= \frac{24.1}{251.2} \times 10^4 = 974 \text{ pF}$$

$$L_M = \frac{\sqrt{20\,(50-20)} \times 10^3}{(6.28)\,(4000)}$$

$$= \frac{\sqrt{600}}{25.12} = \frac{24.5}{25.12} = 0.97\ \mu\text{H}$$

The chart of Fig. 10-12 shows the capacitive reactance of C_M, and the inductive reactance of L_M necessary to match various antenna impedances to 50-ohm coaxial cable.

In practice, L_M need not be a separate inductor. Its effect can be duplicated by adding an equivalent amount of inductance to the loading coil, regardless of whether the loading coil is at the base or at the center of the antenna.

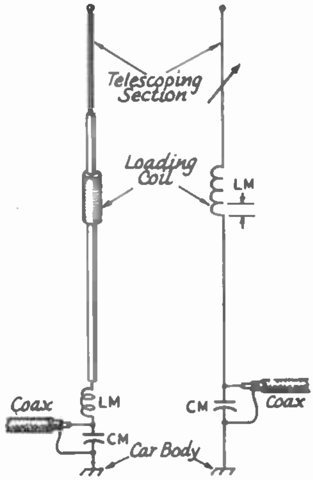


Fig. 10-11 – A whip antenna may also be matched to coax line by means of an L network. The inductive reactance of the L network can be combined in the loading coil, as indicated at the right.

Adjustment

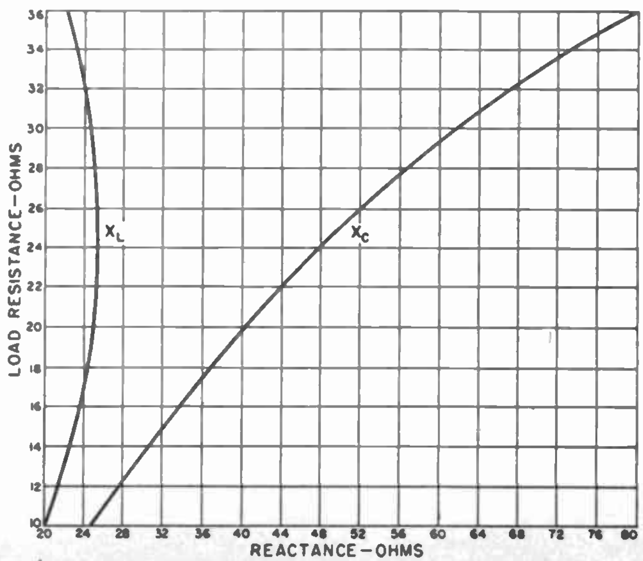
In adjusting this system, at least part of C_M should be variable, the balance being made up of combinations of fixed mica capacitors in parallel as needed.

A small one-turn loop should be connected between C_M and the chassis of the car, and the loading coil should then be adjusted for resonance at the desired frequency as indicated by a GDO coupled to the loop at the base. Then the transmission line should be connected, and a check made with an SWR bridge connected at the transmitter end of the line.

With the line disconnected from the antenna again, C_M should be readjusted and the antenna returned to resonance by readjustment of the loading coil. The line should be connected again, and another check made with the SWR bridge. If the SWR is less than it was on the first trial, C_M should be readjusted in the same direction until the point of minimum SWR is found. Then the coupling between the line and the transmitter can be adjusted for proper loading. It will be noticed from Fig. 10-12 that the inductive reactance varies only slightly over the range of antenna resistances likely to be encountered in mobile work. Therefore, most of the necessary adjustment is in the capacitor.

The one-turn loop at the base should be removed at the conclusion of the adjustment and slight compensation made at the loading coil to maintain resonance.

Fig. 10-12 – Curves showing inductive and capacitive reactances required to match a 50-ohm coax line to a variety of antenna resistances.



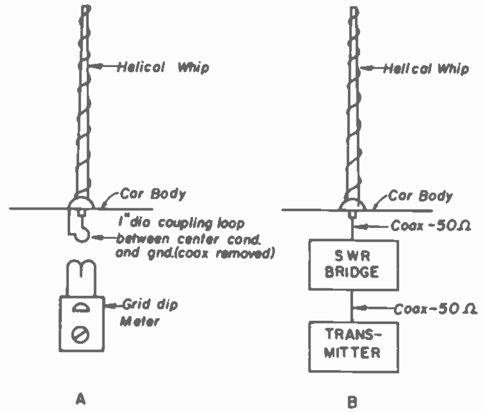


Fig. 10-13 – The resonant frequency of the antenna can be checked (A) with a grid-dip meter or (B) by finding the frequency at which minimum feed-line SWR occurs. The latter method is more accurate at high frequencies because it eliminates the effect of the coupling loop required in A.

CONTINUOUSLY-LOADED HELICAL WHIPS

A continuously-loaded whip antenna of the type shown in Fig. 10-14 is thought to be more efficient than a center- or base-loaded system (*QST*, May 1958, W9KNK). The feed-point impedance of the helically-wound whip is somewhat greater than the previously described mobile antennas, and is on the order of 20 ohms, thus providing an SWR of only 2.5 when 50-ohm coaxial feed line is used. The voltage and current distribution is more uniform than that of lumped-constant antennas. The low SWR and this feature make the antenna more efficient than the center- or base-loaded types. Antennas of this variety can be wound on a fiber glass fishing rod, then weatherproofed by coating them with liquid fiber glass, or by encapsulating them with shrinkable vinyl-plastic tubing.

Tapered Pitch

On frequencies below 28 MHz the radiation resistance falls off so rapidly that for the desired 4- and 6-foot whip lengths the resistance values are not suitable for direct operation with 50-ohm lines. It is desirable to raise the feed-point R to a value, approaching 50 ohms so that a matched line condition will exist. Based on extensive experimentation, a *tapered*-pitch continuous-loading antenna is recommended. Since it is not feasible to wind the helix with continuously varying pitch, a "step-tapered" design is best. A typical step-tapering technique for a variable-pitch helical whip antenna is to divide the total length of the radiator, say 4 feet, into 6 equal parts of 8 inches each. The helix is then wound with a 2-inch pitch for the first 8 inches, pitches of 1, 1/2, 1/4 and 1/8 inch, respectively, for the next four 8-inch sections, and finished with close winding of the final section. The resonant frequency will depend upon the rod diameter, wire size and number of turns. However, the variable-pitch 6-step taper approaches the ideal continuously-variable condition closely enough to give a good 50-ohm match with a 4-foot antenna at frequencies between 20 and 30 MHz.

Adjustment

With this design it is difficult to adjust the resonant frequency by changing the turns near the base; however, the frequency may be adjusted very readily by cutting off sections of the tightly-wound portion near the top of the whip. The technique to follow is to design for a frequency slightly lower than desired and then to bring the unit in on frequency by cutting small sections off the top until it resonates at the desired frequency. Resonance can be checked either by the use of a grid-dip meter or by the use of a transmitter and SWR bridge. Reflected power as low as 2 to 5 percent can easily be obtained with the units properly resonated even though it may mean cutting an inch or two off the top closely wound section to bring the unit in on frequency. These values can be obtained in the 10- and 15-meter band with overall lengths of 4 feet and in the 20- and 40-meter bands with a length of 6 feet. In the 75-meter band it has been possible to obtain an SWR of 1.5 using a 6-foot tapered-pitch helical winding, although the bandwidth is restricted to about 60 kHz. This affords operation comparable to the center coil loaded 12-foot whips. In general, the longer the radiator (in wavelengths), the greater the bandwidth. By arbitrarily restricting the physical length to 6 feet, or less, we obtain the following results:

Band	*Length*	*Resonant Freq.*	*SWR*	*Bandwidth for SWR = 2.0*
10 meters	4 feet	29.00 MHz	1.3	800 kHz
15 meters	4 feet	21.30 MHz	1.4	500 kHz
20 meters	6 feet	14.25 MHz	1.3	250 kHz
40 meters	6 feet	7.25 MHz	1.5	100 kHz
75 meters	6 feet	3.90 MHz	1.5	60 kHz

In the 15-, 20- and 40-meter bands the bandwidths of the taper-pitch designs are good enough to cover the entire phone portions of the bands. The bandwidths have been arbitrarily

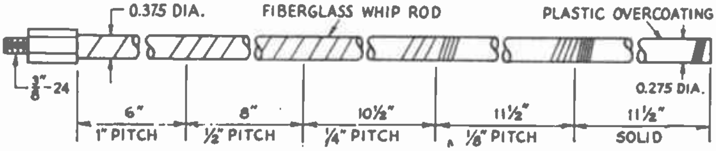


Fig. 10-14 – Dimensions for a 15-meter stepped-pitch whip, wound with No. 20 enameled wire.

Fig. 10-15 – K1MET prunes a capacity hat for antenna resonance at the low end of the 160-meter band. The Webster Big-K antenna is first tuned for the high segment of the band. The capacity hat is clipped on when operation on the "low end" is desired. Fine adjustments can be made by increasing or decreasing the spacing between the two No. 10 wires.

selected as that frequency spread at which the SWR becomes 2 on a 50-ohm line, although with most equipment SWR values up to 2.5 can be tolerated and loading accomplished with ease.

Top-Loading Capacitance

Because the coil resistance varies with the inductance of the loading coil, the resistance can be reduced, beneficially, by reducing the number of turns on the coil. This can be done by adding capacitance to that portion of the mobile antenna that is *above* the loading coil. To achieve resonance, the inductance of the coil is reduced proportionally. Capacity "hats," as they are often called, can consist of a single stiff wire, two wires or more, or a disk made up from several wires, like the spokes of a wheel. A solid metal disk can also be used. The larger the capacity hat, in terms of mass, the greater the capacitance. The greater the capacitance, the smaller the amount of inductance needed in the loading coil for a given resonant frequency.

There are two schools of thought concerning the attributes of center-loading and base-loading. It has not been established that one system is superior to the other, especially in the lower part of the hf spectrum. For this reason both the base- and center-loading schemes are popular. Capacity-hat loading is applicable to either system. Since more inductance is required for center-loaded whips to make them resonant at a given frequency, capacity hats should be particularly useful in improving their efficiency.

REMOTE ANTENNA RESONATING

Fig. 10-17 shows circuits of two remote-control resonating systems for mobile antennas. As shown, they make use of surplus dc motors driving a

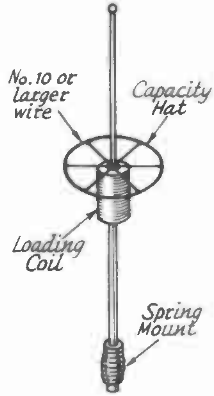


Fig. 10-16 – A capacitance "hat" can be used to improve the performance of base- or center-loaded whips. A solid metal disk can be used in place of the skeleton disk shown here.

loading coil removed from a surplus ARC-5 transmitter. A standard coil and motor may be used in either installation at increased expense.

The control circuit shown in Fig. 10-17A is a three-wire system (the car frame is the fourth conductor) with a double-pole double-throw switch and a momentary (normally off) single-pole single-throw switch. S2 is the motor reversing switch. The motor runs so long as S1 is closed.

The circuit shown in Fig. 10-17B uses a latching relay, in conjunction with microswitches, to reverse automatically the motor when the roller reaches the end of the coil. S3 and S5 operate the relay, K1, which reverses the motor. S4 is the

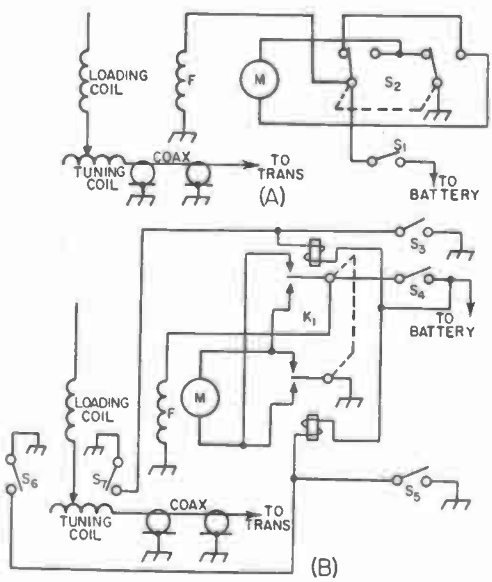


Fig. 10-17 – Circuit of the remote mobile-whip tuning systems.
K1 – Dpdt latching relay.
S1, S3, S4, S5 – Momentary-contact spst, normally open.
S2 – Dpdt toggle.
S6, S7 – Spst momentary-contact microswitch, normally open.

motor on-off switch. When the tuning coil roller reaches one end or the other of the coil, it closes S6 or S7, as the case may be, operating the relay and reversing the motor.

The procedure in setting up the system is to prune the center-loading coil to resonate the antenna on the highest frequency used without the base-loading coil. Then, the base-loading coil is used to resonate at the lower frequencies. When the circuit shown in Fig. 10-17A is used for control, S1 is used to start and stop the motor, and S2, set at the "up" or "down" position, will determine whether the resonant frequency is raised or lowered. In the circuit shown in Fig. 10-17B, S4 is used to control the motor. S3 or S5 is momentarily closed (to activate the latching relay) for raising or lowering the resonant frequency. The broadcast antenna is used with a wavemeter to indicate resonance. (Originally described in *QST*, December 1953.)

VHF MOBILE ANTENNAS

The three most popular vhf mobile antennas are the so-called halo, the turnstile, and the 1/4-wavelength vertical. The same rules apply to the installation and use of these antennas as for antennas operated in the hf bands – mounted as high and in the clear as possible, and with good electrical connections throughout the system.

The polarization chosen – vertical or horizontal – will depend upon the application and the area of the USA where operation will take place. It is best to use whatever polarity is in vogue for your region, thus making the mobile signal compatible with those of other mobiles or fixed stations. Vertically polarized mobile antennas are more subject to pattern disturbance than horizontal types. That is to say, considerably more flutter will be inherent on the signal than with horizontal antennas. This is because such objects as trees and power poles, because of their vertical profile, tend to present a greater path obstacle to the vertical antenna. It is becoming common practice, however, to use omnidirectional, vertically polarized vhf mobile antennas in connection with fm/repeater mobile service, even in areas where horizontal antennas are favored.

Both the turnstile and halo antennas are horizontally polarized. The halo is physically small, but is less effective than a turnstile. It is a half-wavelength dipole bent into a circle, and because the ends are in close proximity to one another, some signal cancellation occurs. This renders the antenna less efficient than a straight center-fed dipole. Halos do not offer a perfectly circular radiation pattern, though this has been a popular belief. Tests indicate that there is definite directivity, though broad, when a halo is rotated 360 degrees over a uniform plane surface.

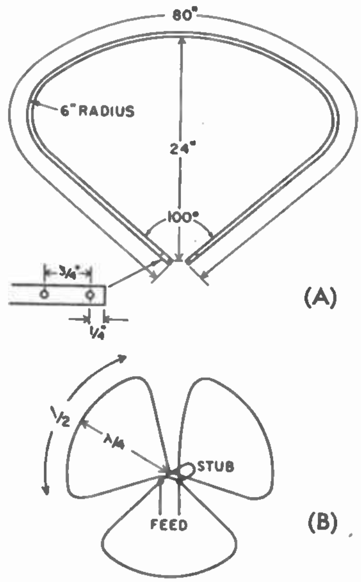


Fig. 10-19 – Schematic representation of the Big Wheel at B. Three one-wavelength elements are connected in parallel. The resulting low feed impedance is raised to 52 ohms with an inductive stub. Illustration A shows the bend details of one element for 144-MHz use.

5/8-WAVELENGTH 220-MHz MOBILE ANTENNA

This antenna was developed to fill the gap between a homemade 1/4-wavelength mobile antenna and a commercially made 5/8-wavelength model. There have been other antennas made using modified citizens band models. This still presents the problem of cost in acquiring the original antenna. The major cost in this setup is the whip portion. This can be any tempered rod that will not take a set easily.

Construction

The base insulator portion is constructed of 1/2-inch Plexiglas rod. A few minutes work on a lathe was sufficient to shape and drill the rod. The bottom 1/2-inch of the rod is turned down to a diameter of 3/8-inch. This portion will now fit into a PL-259 uhf connector. A hole, 1/8-inch diameter, is drilled through the center of the rod. This hole will contain the wires that make the connections between the center conductor of the connector and the coil tap. The connection between the whip and the top of the coil is also run through this opening. A stud was force-fitted into the top of the Plexiglas rod. This allows the whip to be detached from the insulator portion.

The coil should be initially wound on a form slightly smaller than the base insulator. When the coil is transferred to the Plexiglas rod it will keep its shape and will not readily move. After the tap

As can be seen in the photograph, the bottom end of the coil can be soldered directly to the connector.

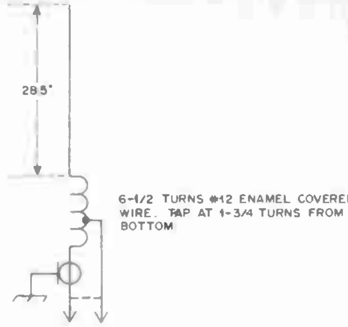


Fig. 10-20 – Diagram of 5/8-wavelength 220 MHz mobile antenna.

point has been determined, a longitudinal hole is drilled into the center of the rod. A No. 22 wire can then be inserted through the center of the insulator into the connector. This method is also used to attach the whip to the top of the coil. After the whip has been fully assembled a coating of epoxy cement was applied. This sealed the entire assembly and provided some additional strength. During a full winter's use, there was not any sign of cracking or mechanical failure.

Adjustment

Prior to final assembly of the whip antenna, the correct tap point should be determined. The correct point will produce the least reflected power. The whip length should be cut initially for the desired operating frequency.

TWO-METER 5/8-WAVELENGTH VERTICAL

Probably the most popular antenna used by the fm group is the 5/8-wavelength vertical. As stated previously, this antenna has some gain when compared to a dipole. The antenna can be used in either a fixed location with radials or in a mobile installation. An inexpensive antenna of this type can be made from a modified CB whip. The antenna shown in Figs. 1 and 2 is a 5/8-wavelength, 2-meter whip.

There are a number of different types of CB mobile antennas available. This particular antenna to be modified consists of a clamp-on trunk mount, a base loading coil, and a 39-inch spring-mounted, stainless-steel whip.

The modification consists of removing the loading-coil inductance, winding a new coil, and mounting a 3-30 pF trimmer in the bottom housing. The capacitor is used for obtaining a precise match in conjunction with the base coil tap.

The first step is to remove the weatherproof phenolic covering from the coil. Remove the base housing and clamp the whip side of the antenna in a vise. Insert a knife blade between the edge of the whip base and the phenolic covering. Gently tap the knife edge with a hammer to force the housing away from the whip section.

Next, remove the coil turns and wind a new coil using No. 12 wire. The new coil should have nine turns, equally spaced. The tap point is two turns up from the base (ground) end on the antenna as

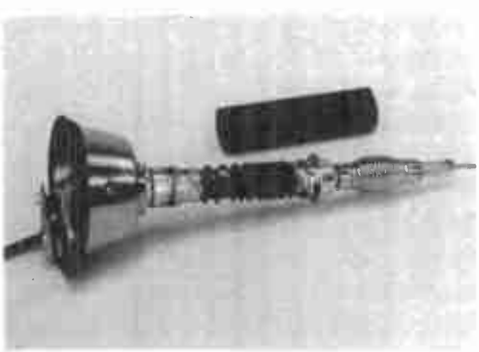


Fig. 1 – The new coil is tapped two turns from the base end. It may be necessary to file the coil ends so that the assembly will fit in the phenolic covering.

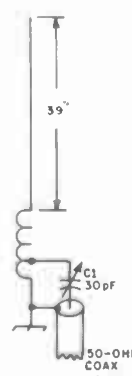
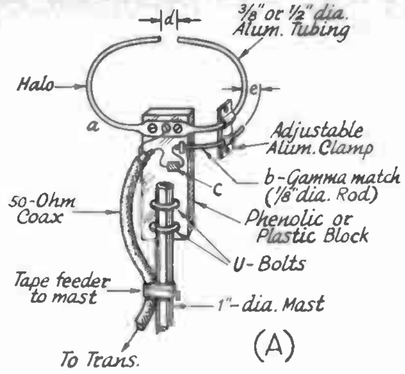
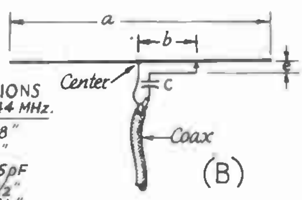


Fig. 2 – Circuit diagram of the whip antenna. C1 is a 3- to 30-pF trimmer.





TYPICAL DIMENSIONS

	50 MHz.	144 MHz.
a =	105 1/2"	38"
b =	18"	6"
c =	50pF	25pF
d =	3 1/2"	1 1/2"
e =	2 1/2"	1 3/4"

Fig. 10-22 — Details for building a halo antenna for 6- or 2-meter use are shown at A. Other mechanical methods are possible, and the construction technique used will be up to the builder. The open end of the coax cable should be sealed against the weather. At B, a schematic representation of the halo. Dimension *a* is set for 1/2 wavelength at the operating frequency. The chart gives approximate dimensions in inches, and will serve as a guide in building a halo.

modified. The trimmer capacitor is mounted on a terminal strip which is installed in the base housing. A hole must be drilled in the housing to allow access to the capacitor adjustment screw.

Initially, the tap on the coil was tried at three turns from the bottom. The antenna was mounted on the car, an SWR indicator was inserted in the feed line, and C1 and the whip height were adjusted for a match. A match was obtained, but when the phenolic sleeve was placed over the coil, it was impossible to obtain an adjustment that proved a match. Apparently the dielectric material used in the coil cover has an effect on the coil. After some experimenting it was found that with the tap two turns up from the bottom, and with the cover over the coil, it was possible to get a good match with 50-ohm line.

This antenna can be used in a fixed location by adding radials. The radials, three or four, should be slightly longer than 1/4-wave and should be attached to the base mounting section.

THE QUARTER-WAVELENGTH VERTICAL

Ideally, the vhf vertical antenna should be installed over a perfectly flat plane reflector to assure uniform omnidirectional radiation. This suggests that the center of the automobile roof is the best place to mount it. Alternatively, the flat portion of the auto's rear trunk deck can be used, but will result in a directional pattern because of car-body obstruction. Fig. 10-23 illustrates at A and B how a Millen high-voltage connector can be used as a roof mount for a 144-MHz whip. The hole in the roof can be made over the dome light, thus providing accessability through the upholstery. RG-59/U and matching section *L*, Fig. 10-23C, can be routed between the car roof and the ceiling upholstery and brought into the trunk compartment, or down to the dashboard of the car. Some operators install an SO-239-type coax connector on the roof for mounting the whip. The method is similar to that of drawing A.

VHF HALO ANTENNAS

The antenna of Fig. 10-22 can be built from aluminum tubing of medium tensile strength. The one-half-wavelength dipole is bent into a circle and fed with a gamma match. Capacitor *c* is shown as a fixed value, but a variable capacitor mounted in a weatherproof box will afford more precise

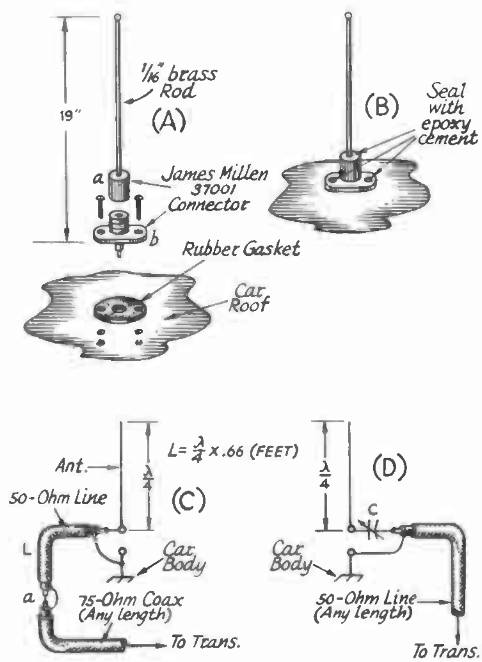
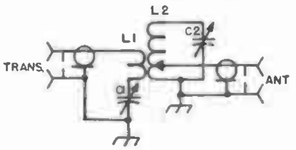


Fig. 10-23 — At A and B, an illustration of how a quarter-wavelength vertical antenna can be mounted on a car roof. The whip section should be soldered into the cap portion of the Millen connector, then screwed to the base socket. This handy arrangement permits removing the antenna when desired. Epoxy cement should be used at the two mounting screws to prevent moisture from entering the car. Diagrams C and D are discussed in the text.



50 MHz	144 MHz
C1 = 100 pF	35 pF
C2 = 25 pF	15 pF
L1 = 2 Ts NO.16 ENAM., 2-1/2" DIA.	2 Ts NO.16 ENAM., 1-1/2" DIA.
L2 = 5 Ts NO.10 1-1/2" DIA.	4 Ts NO. 10 1" DIA.

Fig. 10-24 – Schematic diagram of the 6- or 2-meter antenna-matching circuit for use at the base of the quarter-wavelength vertical antenna. It can be housed in a Minibox and mounted permanently at the antenna base, inside or outside the car. If used outside, it should be sealed against dirt and moisture.

adjustment of the SWR. Or, a variable capacitor can be used initially for obtaining a 1:1 match, then its value can be measured at that setting to determine the required value for fixed capacitor *c*. Fixed-value capacitor *c* should be a dipped silver mica. A 75-pF variable should be used for 6-meter antennas, and a 35-pF variable will suffice for 144 MHz.

The tubing of *a* can be flattened to provide a suitable mounting surface for attachment to the insulating block of Fig. 10-22A. Gamma rod *b* can be secured to the same block by flattening its end and bolting it in place with 4-40 brass hardware. The spacing at *d* can be varied during final adjustment to secure the lowest SWR. Better physical stability will result if a high-dielectric insulator is connected across area *d*. Steatite material is recommended if an insulator/stabilizer is used.

If 75-ohm transmission line is used for the vertical, a quarter-wavelength matching transformer, L, can be used to match the feed impedance of the whip – approximately 30 ohms – to that of the feed line. A section of 50-ohm coax inserted as shown provides a close match to the antenna. Coax fittings can be used at junction *a* to assure a flat line, and to provide mechanical flexibility. BNC connectors are ideal for use with small coax lines. Illustration D shows how a series capacitor can be used to tune the reactance out of the antenna when using 50-ohm feed line. For 144-MHz use it should be 35 pF. A 75-pF variable will suffice for 6-meter antennas. An SWR bridge should be connected in the line while *c* is tuned for minimum reflected-power indication.

A more precise method of matching the line to the antenna is shown in Fig. 10-24. This antenna coupler can match 50- or 75-ohm lines to any antenna impedance from 20 ohms to several hundred ohms. It should be installed *at* the base of the vertical, and with an SWR bridge in the line C1 and C2 should be adjusted for the lowest SWR possible. The tap near the ground end of L2 should then be adjusted for the lowest SWR, readjusting C1 and C2 for minimum reflected power each time the tap is moved. A very compact tuner can be built by scaling down the coil dimensions appropriately. Trimmer capacitors can be used for C1 and C2 if power levels of less than 50 watts are used.

MOBILE POWER SUPPLIES

Most modern-day mobile installations utilize commercially-built equipment. This usually takes the form of a transceiver for ssb on the hf bands, and ssb or a-m for vhf operation. For fm operation in the vhf bands, most transceivers are surplus units which were originally used by commercial land-mobile services. Some home-built equipment is still being used, and it is highly recommended that one consider building his own mobile installation for the technical experience and satisfaction such a project can afford.

Many mobile transceivers contain their own power supplies for 6- and 12-volt dc operation. Some internal power supplies will also work off the 117-V mains. Vibrator power supplies are quite popular for low and medium power levels, but solid-state supplies are more reliable and efficient. Dynamotors are still used by some operators, but are bulky, noisy, and inefficient. The latter imposes an extremely heavy drain on the car battery, and does not contribute to long-term mobile or emergency operation without having the engine running at fairly high rpm to maintain the charge level of the battery.

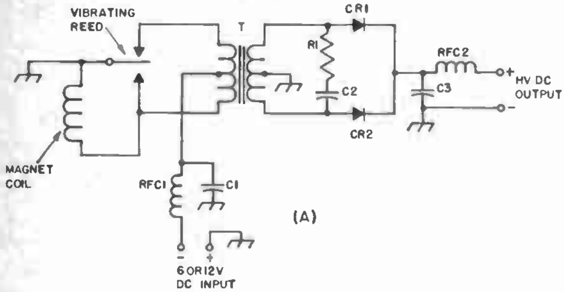
Dynamotors

A dynamotor differs from a motor generator in that it is a single unit having a double armature winding. One winding serves for the driving motor, while the output voltage is taken from the other. Dynamotors usually are operated from 6-, 12-, 28- or 32-volt storage batteries and deliver from 300 to 1000 volts or more at various current ratings.

Commutator noise is a common cause of poor reception when dynamotors are used. It can usually be cured by installing .002-μF mica bypass capacitors from the dynamotor brushes (high-voltage end of armature) to the frame of the unit, preferably inside the cover. The high-voltage output lead from the dynamotor should be filtered by placing a .01-μF capacitor in shunt with the line (a 1000-V disk), followed by a 2.5-mH rf choke (in series with the line) of adequate current rating for the transmitter or receiver being powered by the dynamotor. This network should be followed by a smoothing filter consisting of two 8-μF electrolytic capacitors and a 15- or 30-H choke having a low dc resistance. The commutator and its grooves, at both ends of the armature, should be kept clean to further minimize noise. Heavy, direct leads should be used for connecting the dynamotor to the storage battery.

Vibrator Power Supplies

The vibrator type of power supply consists of a special step-up transformer combined with a



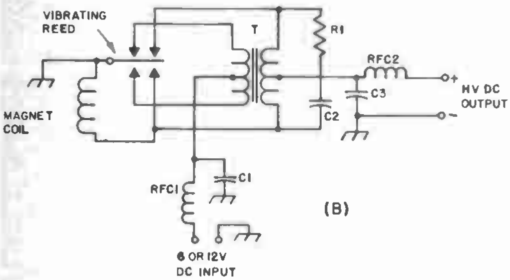


Fig. 10-25 — Basic types of vibrator power supplies. A — Nonsynchronous. B — Synchronous.

vibrating interrupter (vibrator). When the unit is connected to a storage battery, plate power is obtained by passing current from the battery through the primary of the transformer. The circuit is made and reversed rapidly by the vibrator contacts, interrupting the current at regular intervals to give a changing magnetic field which induces a voltage in the secondary. The resulting square-wave dc pulses in the primary of the transformer cause an alternating voltage to be developed in the secondary. This high-voltage ac in turn is rectified, either by silicon diode rectifiers or by an additional synchronized pair of vibrator contacts. The rectified output is pulsating dc, which may be filtered by ordinary means. The smoothing filter can be a single-section affair, but the output capacitance should be fairly large – 16 to 32 μF.

Fig. 10-25 shows the two types of circuits. At A is shown the nonsynchronous type of vibrator. When the battery is disconnected the reed is midway between the two contacts, touching neither. On closing the battery circuit the magnet coil pulls the reed into contact with one contact point, causing current to flow through the lower half of the transformer primary winding. Simultaneously, the magnet coil is short-circuited, de-energizing it, and the reed swings back. Inertia carries the reed into contact with the upper point, causing current to flow through the upper half of the transformer primary. The magnet coil again is energized, and the cycle repeats itself.

The synchronous circuit of Fig. 10-25B is provided with an extra pair of contacts which rectifies the secondary output of the transformer, thus eliminating the need for a separate rectifier tube. The secondary center tap furnishes the positive output terminal when the relative polarities of primary and secondary windings are correct. The proper connections may be determined by experiment.

The buffer capacitor, C2, across the secondary of T, absorbs spikes that occur on breaking the current, when the magnetic field collapses almost instantly and hence causes high voltages to be induced in the secondary. Without C2 excessive sparking occurs at the vibrator contacts, shortening the vibrator life. Resistor R1 is part of the buffer and serves as a fuse if C2 should short out, thus protecting the vibrator and transformer from damage. Values between 1000 and 5600 ohms, 1 watt, are commonly used. Correct values for C2 lie between .005 and .03 μF, and for 220-350-V supplies the capacitor should be rated at 2000 V or better, dc. The exact capacitance is critical, and should be determined experimentally while observing the output waveform on an oscilloscope for the least noise output. Alternatively, though not as effective a method, the capacitor can be selected for least sparking at the vibrator contacts.

Vibrator-transformer units are available in a variety of power and voltage ratings. Representative units vary from one delivering 125 to 200 volts at 100 mA to others that have a 400-volt output rating at 150 mA. Most units come supplied with "hash" filters, but not all of them have built-in ripple filters. The requirements for ripple filters are similar to those for ac supplies. The usual efficiency of vibrator packs is in the vicinity of 70 percent, so a 300-volt 200-mA unit will draw approximately 15 amperes from a 6-volt storage battery. Special vibrator transformers are also available from transformer manufacturers so that the amateur may build his own supply if he so desires. These have dc output ratings varying from 150 volts at 40 mA to 330 volts at 135 mA.

"Hash" Elimination

Sparking at the vibrator contacts causes rf interference ("hash," which can be distinguished from hum by its harsh, sharper pitch) when used with a receiver. To minimize this, rf filters are incorporated, consisting of RFC1 and C1 in the battery circuit, and RFC2 with C3 in the dc output circuit.

Equally as important as the hash filter is thorough shielding of the power supply and its connecting leads, since even a small piece of wire or metal will radiate enough rf to cause interference in a sensitive amateur receiver.

TRANSISTORIZED POWER SUPPLIES

Most present-day mobile equipment is powered by solid-state dc-to-dc converters. They are somewhat similar to vibrator supplies in that they use power transistors to switch the primary voltage of the transformer. This technique eliminates sparking in the switching circuit, and offers greater reliability and efficiency. The switching transistors can be made to oscillate, by means of a feedback winding on the transformer, and by application of forward bias on the bases of the switching transistors. The switching rate can be set for any frequency between 50 Hz and several thousand Hz and depends to a great extent upon the inductance of the transformer windings. The switching waveform is a square wave. Therefore, the supply is capable of causing a buzzing sound in transmitter or receiver output in much the same fashion as with a vibrator supply. Rf filtering should be employed as a corrective measure. At higher switching rates the buzz becomes a whine which sounds like that from a dynamotor. High-frequency switching rates are preferred for dc-to-dc converters because smaller transformer cores can be used, and because less output filtering is required. The efficiency of a well-designed solid-state power suuply is on the order of 80 percent, an improvement over the usual 60 to 70 percent of vibrator supplies, or the miserable 30 to 40 percent of dynamotors.

A typical transistorized supply is shown in Fig. 10-26. The supply voltage is fed into the emitter circuit of Q1-Q2. A resistive divider is used to obtain forward bias for the transistors through base-feedback-winding 1. The primary switching takes place between the emitter and collector of each transistor. Q1 and Q2 are connected in push-pull and conduct on alternate half cycles. As each transistor is driven into conduction it saturates, thus forming a closed contact in that leg of the circuit. The induced voltage is stepped up by T, and high-voltage appears across winding 3. Zener diodes CR1 and CR2 protect Q1 and Q2 from voltage spikes. They should be rated at a voltage slightly lower than the Vce of the transistors. Diodes CR3 through CR6 form a bridge rectifier to provide dc output from winding 3. Some supplies operate at a switching rate of 2000 to 3000 Hz. It is possible to operate such units without using output rectifiers, but good filtering is needed to remove the ripple from the dc output.

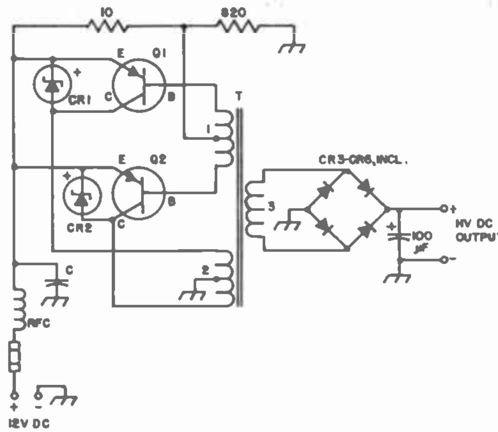


Fig. 10-26 – Typical dc-to-dc converter. Ratings for CR3-CR6, and the 100-µF filter capacitor can be selected from data in the power-supply chapter.

Transistor Selection

The switching transistors should be able to handle the primary current of the transformer. Since the feedback will diminish as the secondary load is increased, the beta of the transistors, plus the design of the feedback circuit, must be sufficient to sustain oscillation under full-load conditions. During no-load conditions, the feedback voltage will reach its highest peak at the bases of Q1 and Q2. Therefore, the transistors must be rated for whatever base-emitter reverse voltage that occurs during the cutoff period. Since the transistors must be able to handle whatever peak voltage occurs during the switching process, it is wise to stay on the safe side. Choose transistors that have a Vceo rating of three or four times the supply voltage, keeping in mind that fully charged automobile batteries can deliver as much as 14 volts. Heat sinks should be used on Q1 and Q2 to prevent damage from excessive heating. The larger the heat sink, the better. Under full-load conditions the transistors should only be slightly warm to the touch. If they are running hot, this will indicate inadequate heat sinking, too great a secondary load, or too much feedback. *Use only enough feedback to sustain oscillation under full loading*, and to assure rapid starting under the same conditions.

MOBILE POWER SUPPLY FOR TRANSCEIVERS

Transceivers, such as the Heath SB-102, and the Drake TR-4 require a separate power supply when operated from 12-volts dc. Additionally, linear amplifiers can be run from a separate dc supply to allow increased power operation from relatively low-power transceivers. The unit described here, when operated from 12-volts dc, will deliver approximately 900-volts dc at 300 mA, 250-volts dc at 200 mA, negative 150-volts dc at 40 mA, and an adjustable bias voltage from 10 to 150 volts of dc.

The Circuit

A common-emitter configuration is used with diodes to provide a return path for the feedback winding, as shown in Fig. 10-28. Assuming that Q2 conducts first, the base is driven negative by the feedback winding (connections 6 and 7 on T1). CR15 then conducts, thereby protecting the base of Q1. CR14 is back-biased to an open circuit when Q2 is conducting. When T1 saturates producing a square wave, the voltage at pins 6 and 7 of T1 reverses turning on Q1. When Q2 conducts, current flows through the primary of T2 in one direction and as Q1 conducts, current flows through the primary of T2 in the other direction. This reversal of current in the primary of T2 provides an alternating square-wave voltage which is stepped up by the secondary winding. Full-wave rectification with current limiting is used with each secondary winding.

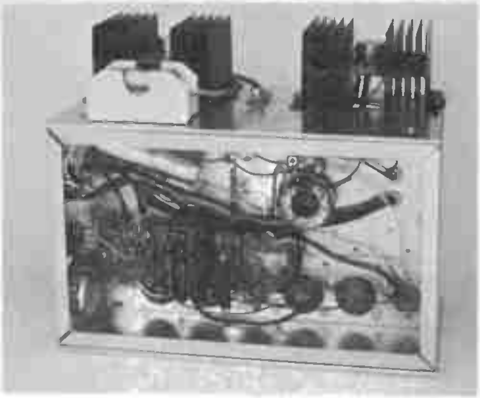


Fig. 10-27 — The heat sinks are mounted on an aluminum panel. When installing a power supply of this type, be sure to keep the heat sink fins in a vertical position to provide best air circulation. All of the filter capacitors are mounted in a row across the front of the chassis. RFC1 is located next to the transformer. Two sockets are mounted on the chassis side wall to accept an interconnecting cable from the transceiver. To the left of these sockets is the bias voltage adjustment control, R3. This model of the power supply was built by W8HS and assistance was given by W8DDO, W9IWJ (of Delco Radio Corp.), and Jim Osborne (of Osborne Transformer Co.).

The supply oscillates at about 1000 Hz and audible noise is low. The main power to the supply is applied through K1B. K1A can be connected in parallel with the filament supply in the transceiver.

Hash filtering is provided by RFC1 and its associated bypass capacitors in the primary lead. Transient suppression is assured by CR13, CR16 and CR17. Bleeder resistors are used on each supply leg to provide a constant minimum load for the circuit. The supply can be operated without being connected to its load without fear of damaging the diodes or transistors, although this is not considered good practice. Input and output connectors for interconnection to the battery and the transceiver can be selected to meet the needs of the particular installation.

Construction

This circuit requires that the transistors be insulated from the heat sink. Suitable insulators are included with the devices. Silicone grease should be used to help conduct the heat away from the transistors.

No attempt has been made to make the supply small. It is built on a 12 X 6 X 3-inch chassis which allows plenty of room for the heavy conductors. The capacitors are mounted in a row along one side

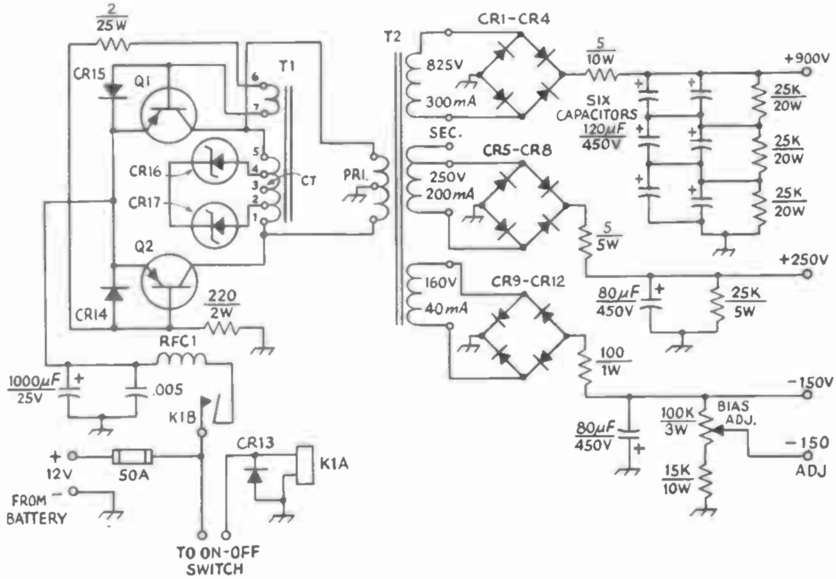


Fig. 10-28 — Circuit diagram of the mobile power supply. Polarized capacitors are electrolytic, others are paper or mica. Resistances are in ohms. Component designations not listed below are for text reference.

CR1-CR13, incl. — 1000-PRV, 1.5-A silicon diode (Mallory MR 2.5 A or equiv.).
CR14, CR15 — 50-PRV, 3-A silicon diode (G.E. A15F).
CR16, CR17 — 18-volt, 1-watt, Zener diode (Motorola 1N4746).
K1 — Spst contactor relay, 60-A, 12-volt dc coil (Potter and Brumfield MB3D).
Q1, Q2 — Delco 2N1523 transistor (substitutions not recommended). Delco insulator kits (No. 7274633) are required. The heat sinks are Delco part No. 7281366.
R3 — 100,000-ohm, 3-watt, linear-taper control.
RFC1 — 20 turns, No. 10 enam. wire on a 1/4-inch dowel.
T1 — Feedback transformer, 1000-Hz (Osborne 6784).
T2 — Hipersil transformer, 1000-Hz (Osborne 21555).

of the chassis. The heat sinks, shown in the photograph, are mounted on a 1/8-inch-thick aluminum back plate.

The leads from the battery to the relay, and from the relay to the transistors and T1, should be No. 6 or No. 8 conductors. All ground leads should be connected to one point on the chassis. The wiring layout is uncritical and no other special precautions are necessary.

Operation

The power supply should be mounted as close to the battery terminals as possible to minimize voltage drop. If the supply is trunk mounted, 1/4-inch conductors should be used to connect it to the battery. A 300-volt tap is available on the secondary of T2. If the transciever requires more than 250 volts for proper operation, this tap can be used.

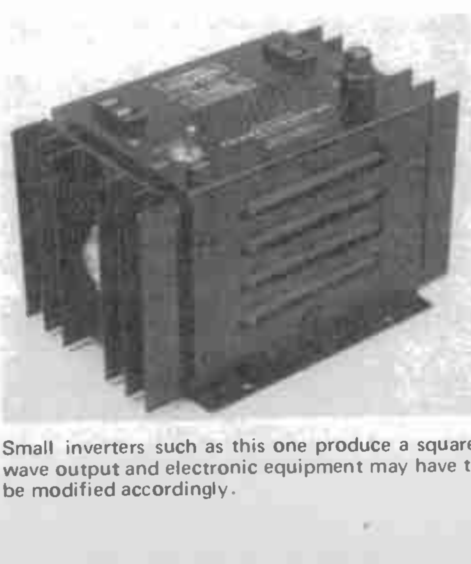
AMATEUR RADIO AFLOAT

Many amateurs like to combine two hobbies . . . radio and boating. While operating an amateur station from a boat is similar to other forms of portable operation, there are a few important differences. For one thing, equipment must be protected from occasional contact with salt water. Even in large boats with suitable cabins, it always seems an "accident" of some sort manages to occur. Someone handles a piece of equipment with wet hands or a hand-held transceiver is operated in an area of the boat subject to spray and spume. The combination of salt, air and time then do a job that experienced boaters might equate to the action of the most corrosive acid.

Since portable equipment is the most likely to encounter such hazards, protecting this kind of gear with a plastic bag is one measure that can be taken as shown in the photograph. Insert the transceiver into a suitable bag and securely tie the top opening around the whip antenna. Usually, the controls can still be manipulated with only minor inconvenience. While the method isn't waterproof, it does provide adequate protection for most purposes.

Low-voltage connectors, especially those subject to the weather, may benefit from a liberal coating of marine lube both inside and on the threads. (The same method is also useful for the sockets in trailer lights.) When possible, coaxial cable should be looped so that a small section is *above* the connection to the antenna. This will prevent water from running down inside the cable jacket after a period of time.

Low-cost insurance is provided by weatherproofing portable gear in plastic bags. Be sure to remove bag when taken ashore and discard bags at first sign of wear.



Small inverters such as this one produce a square-wave output and electronic equipment may have to be modified accordingly.

Dc-to-ac inverters can be built and are available commercially, including a model that comes in kit form. However, the output of such inverters is usually a square wave. As a consequence, some noise or buzz may occur in electronic equipment and additional filtering may be required. Also, even though the rms voltage from a square-wave inverter is rated at the same value as an ordinary residential service, the peak values are different. The peak voltage from a square wave is equal to the rms value while the peak voltage of a sine wave is 1.707 times the rms value. Since the output of most power supplies lies between the rms and peak value, a lower dc output voltage may occur when an inverter is used.

Mobile-like antennas such as whips are frequently employed at hf and the question arises of a

Unless you intend to go "First Class," as in this U.S. Navy patrol craft, be prepared for some tough technical problems (WB2VYU//MMR2 operates his SB-102 into a 28-foot whip and his "shack" would be the envy of many boaters).

ground return for the rf currents. In the case of a metal boat, or an antenna with a high feed-point impedance (such as a half-wavelength vertical whip), the problem is simple and either a connection to the metal boat or a short conductor immersed in the water should suffice. A piece of metal sheet fastened to the bottom might be a suitable substitute for low-impedance types such as quarter-wavelength whips. If the sheet is large enough, paint and other coatings shouldn't cause ill effects since the capacitance between the plate and the water may be sufficient to provide adequate coupling.

Bands to Use

In most cases, excursions are limited to a few miles offshore and vhf operation should prove sufficient in regards to activity and range. Since many boaters tend to favor a certain area, a check on the local repeater situation might be in order if continued operation in one region is the rule. On the other hand, bands such as 10 meters are ideal if activity is sufficient. Getting antennas of electrical length such that efficiency is not a problem is relatively easy and "local" propagation on this band is excellent especially over salt water. Ranges are extended by a considerable margin over propagation distances on land. The lack of atmospheric noise and QRM means low power should prove effective.

Offshore cruising is ordinarily limited to larger craft, and the hf range from 40 meters might be considered for this type of work. If 117 V ac, 60 Hz, is available, fixed-station operation can be employed. Diesel engines have no ignition system in the normal sense and electrical noise should not be a problem. However, if an auxiliary gasoline generator is to supply the power, adequate noise-suppression measures should be taken.

If the power source consists of some other type than 117 V ac, 60 Hz, one might as well be resigned to the fact conversion problems may be practically insurmountable. Rotary converters are large, heavy, expensive and inefficient. Solid-state models are good for only low-power applications and ones suitable for high-power work are apt to be hard to obtain. In addition, if operation from lead-acid storage batteries is contemplated, another difficulty exists. While such cells are useful in delivering large amounts of current for short periods of time, the voltage drop becomes severe with moderate loads over longer periods. This is because bubbles tend to form on the plates and temporarily increase the internal resistance of the cell. As a consequence, inverters and other solid-state equipment do not function as efficiently since such gear is usually designed for a nominal voltage of 13.8. Unfortunately, there are no easy solutions to this particular problem and plans in such matters as purchasing equipment should be made accordingly.

Other Uses of Radio

Even if amateur operation from a small boat is not contemplated, there are other applications of radio that might be of interest. In many areas, continuous broadcast of weather is sent out over special fm stations on approximately 162.4 MHz. Receivers that cover this range could be built but they are also readily available commercially at low cost. The Coast Guard maintains a system of radio-direction-finding stations in the lf range. Location and frequency of these stations are contained in charts and in lists published by the government. While such equipment is usually unnecessary for inshore excursions, having RDF equipment is handy for offshore trips. However, noise difficulties with gasoline engines are apt to be a problem unless adequate suppression is employed. Since RDF gear is usually battery powered, the engine can be shut off while readings are taken.

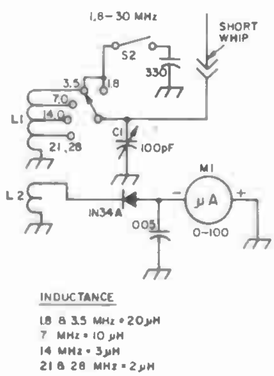


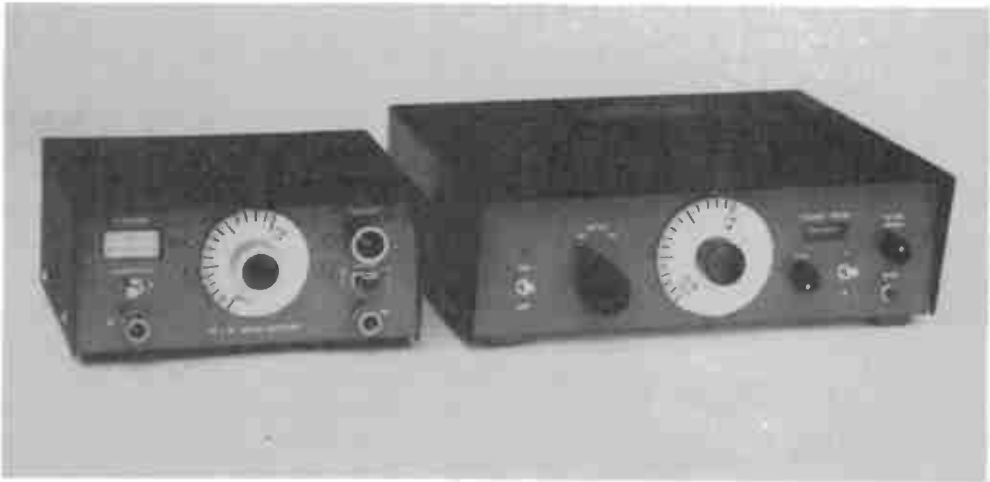
Fig. 10-32 – A band-switched field-strength meter for tuning up the hf-band mobile antenna. It should be assembled in a metal box. In use, it should be placed several feet from the antenna under test. C1 is tuned for a peak meter reading at the operating frequency. It can be detuned for varying the sensitivity.

A BAND-SWITCHING FIELD-STRENGTH METER

The circuit of Fig. 10-32 can be used for tuning the mobile antenna system to resonance. It covers a range from 1.8 to 30 MHz. A single toroidal inductor is used in the tuned circuit. The coil is tapped to provide band switching by means of S1. C1 is tuned for a peak meter reading at the transmitter's output frequency. The unit should be housed in a metal utility box. A banana jack can be used for attaching the short whip antenna.

An Amidon Associates E-core, No. T-68-2, is wound with 50 turns of No. 26 enamel wire. It is tapped 10 turns from ground for 15- and 10-meter use, 18 turns from ground for 20 meters, and 36 turns above ground for 40 meters. The entire 50 turns are used for 80 and 160 meters. S2 adds a 330-pF capacitor for 160-meter operation. S1 can be a single wafer, single-pole, 5-position rotary switch of phenolic or ceramic insulation. S2 can be a spst slide switch. C1 is a Hammarlund HF-100 capacitor, or equivalent. (Amidon cores can be obtained from Amidon Associates, 12033 Otsego St., N. Hollywood, CA 91607.)

A DIRECT-CONVERSION KILOGRAM FOR 20 AND 40 METERS



When portability, low current drain, and simplicity are required in a receiver, it is hard to beat the technique of direct conversion. The unit described here covers the cw portion of both 20 and 40 meters. As total power consumption is on the order of 0.6 watt, battery operation is practical. Packaged in an aluminum box only 6 × 7 × 3 inches (HWD), the receiver weighs in at about one kilogram (2.2 pounds) and fits easily inside a suitcase. The receiver is designed to be compatible with the low-power solid-state transmitter described in Chapter 6.

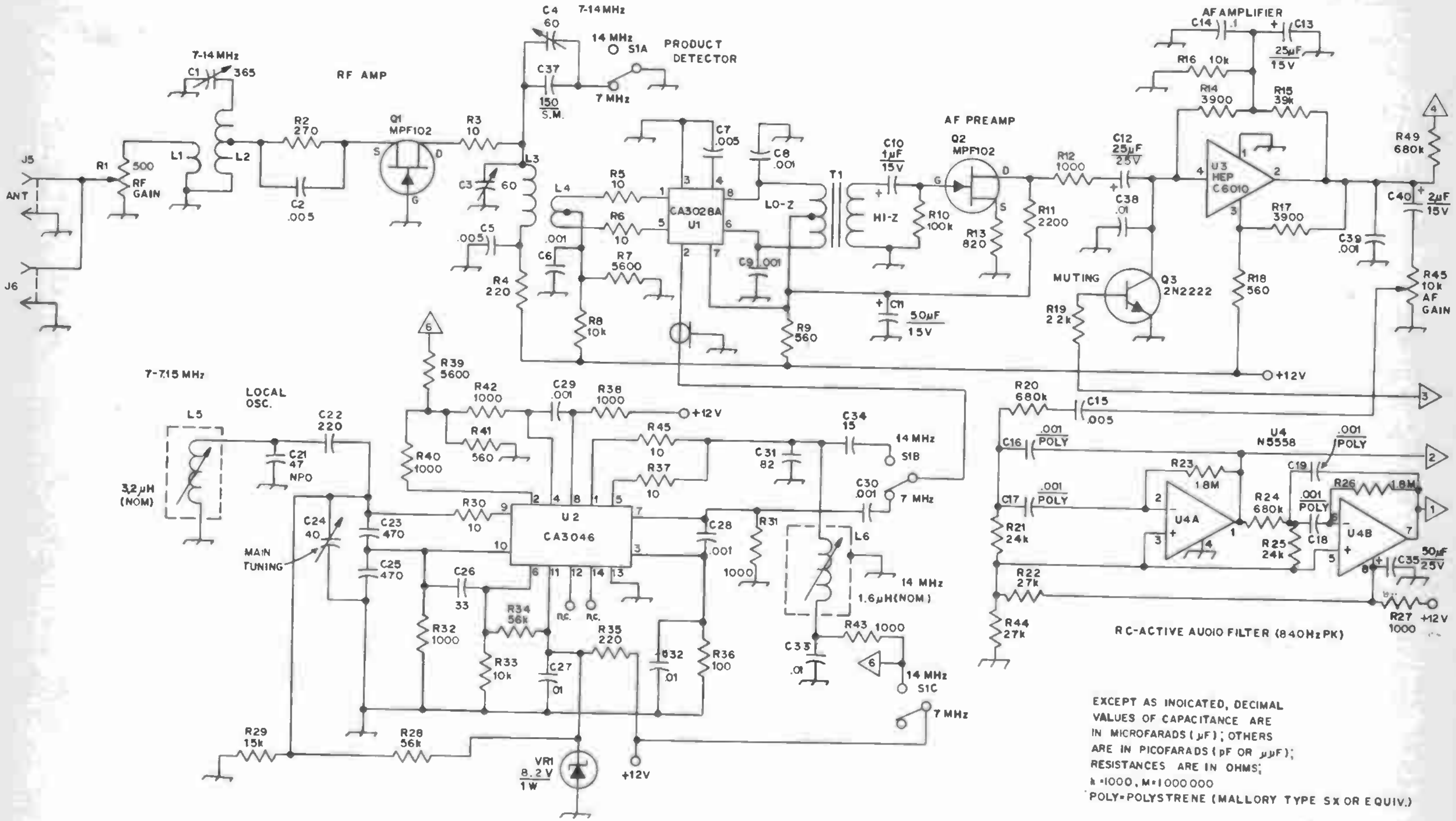
Circuit Overview

This approach to direct conversion uses an FET as a fixed-tuned rf amplifier, switchable between 20 and 40 meters. An IC transistor array serves as the heart of a band-switched local oscillator. A differential amplifier IC functions as a product detector. The audio channel uses an FET to establish a low noise figure, followed by a high-gain wide-band IC amplifier. Audio selectivity is achieved through the use of a two-stage active filter. A signal strength indication is obtained through the use of an audio-derived S-meter circuit.

Circuit Description

Q1 operates as a grounded-gate rf amplifier. C1, a front-panel mounted broadcast type variable capacitor, peaks the input of the stage for 40- or 20-meter reception. The output of the stage is tuned to 20 meters by L3-C3. One pole of S1 switches additional capacitance in parallel with C3 for 40-meter operation. L4 couples rf energy to the input of the product detector, U1. Local-oscillator injection is applied to pin 2 of U1, the base of the internal constant-current source transistor. The local oscillator is built around a CA3046 IC transistor array, and is identical to the VFO used in the companion transmitter described in Chapter 6.

The CA3046 contains three independent NPN


RF AMP
PRODUCT DETECTOR
AF PREAMP
AF AMPLIFIER
LOCAL OSC.
MUTING
MAIN TUNING
7-14 MHz
7-7.15 MHz
ANT
J5
J6
R1 500 RF GAIN
C1 365
L1
L2
R2 270
C2 .005
Q1 MPF102
R3 10
C3 60
L3
C4 60
C37 150 S.M.
14 MHz S1A
7 MHz
C5 .005
R4 220
L4
R5 10
R6 10
R7 5600
R8 10k
C6 .001
CA3028A U1
C7 .005
C8 .001
C9 .001
LO-Z
T1
HI-Z
C10 1µF 15V
R10 100k
Q2 MPF102
R13 820
C11 50µF 15V
R9 560
R12 1000
R11 2200
C12 25µF 25V
C38 .01
Q3 2N2222
R19 22k
C14 .1
C13 25µF 15V
R16 10k
R14 3900
R15 39k
U3 HEP C6010
R17 3900
R18 560
R49 680k
C40 2µF 15V
C39 .001
R45 10k AF GAIN
+12V
R39 5600
R42 1000
C29 .001
R38 1000
R41 560
R40 1000
R45 10
R37 10
C31 82
C34 15
14 MHz S1B
C30 .001
7 MHz
L5 3.2µH (NOM)
C22 220
C21 47 NPO
C24 40
C23 470
C25 470
R30 10
U2 CA3046
C28 .001
R31 1000
L6 1.6µH (NOM)
14 MHz
C26 33
R32 1000
R34 56k
R33 10k
C27 .01
R35 220
C32 .01
R36 100
n.c.
R43 1000
C33 .01
14 MHz S1C
R29 15k
R28 56k
VR1 8.2V 1W
R20 680k
C15 .005
C16 .001 POLY
U4 N5558
C19 .001 POLY
R23 1.8M
C17 .001 POLY
R26 1.8M
U4A
U4B
R24 680k
C18 .001 POLY
R21 24k
R25 24k
C35 50µF 25V
R22 27k
R44 27k
R27 1000
RC-ACTIVE AUDIO FILTER (840Hz PK)
EXCEPT AS INDICATED, DECIMAL VALUES OF CAPACITANCE ARE IN MICROFARADS (µF); OTHERS ARE IN PICOFARADS (pF OR µµF); RESISTANCES ARE IN OHMS; k=1000, M=1000 000
POLY=POLYSTRENE (MALLORY TYPE SX OR EQUIV.)

silicon transistors plus one differentially connected transistor pair, and is available in a 14 pin dual inline package. The RCA CA3045 is directly interchangeable with the CA3046, and they are available surplus very inexpensively. The ICs have identical pin connections and electrical characteristics, and the only difference between the two is that the CA3046 is packaged in a ceramic case, while the CA3045 is packaged in a plastic case.

In this application, one transistor is used in a Colpitts VFO circuit operating at 7 MHz, one transistor is used as a phase splitter, and the differential pair is used as a push-push doubler to produce output at 14 MHz. One transistor is unused. The 7-MHz output is taken from the emitter of the phase splitter. The devices in the

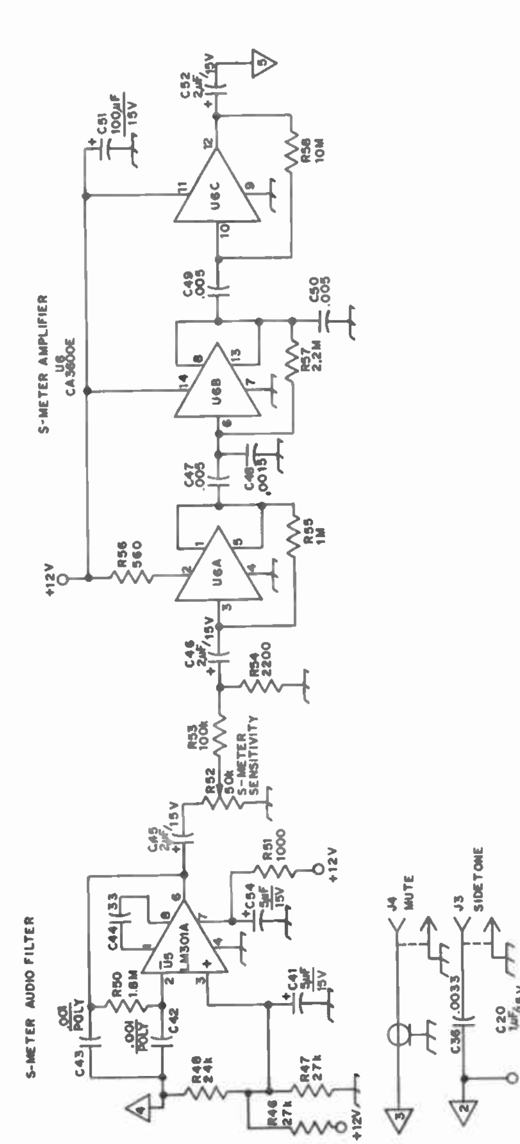


Fig. 1 — Schematic diagram of the direct-conversion receiver. Numbered parts in this diagram which are not listed below are so numbered for circuit-board identification and text discussion. Fixed-value resistors may be 1/2- or 1/4-watt composition. Polarized capacitors are electrolytic. Fixed-value capacitors are disk ceramic, unless otherwise indicated.

C1 — 365-pF miniature variable.
C3, C4 — 60-pF trimmer (Erie 538-011F-15-60 or equiv.).
C16, C17, C18, C19, C42, C43 — 1000-pF polystyrene.
C21 — 47-pF NPO.
C24 — three-section, 20-pF-per-section variable, two sections used (J.W. Miller 1460 or similar).
CR1 — Silicon diode.
J1 — Phone jack.
J2 — Miniature phone jack.
J3, J4, J6 — Phono jack.
J5 — SO-239 connector.
L1 — 2 turns small diameter hookup wire over L2.
L2 — 18 turns No. 22 enam. wound on Amidon T50-6 core (tap 4 turns above ground).
L3 — 25 turns No. 24 enam. wound on T50-6 core.
L4 — 6 turns small diameter hookup wire, center tapped, over L3.
L5 — 3.0-7.0 μH shielded variable inductor (J.W. Miller 9051 or equiv.).
L6 — 1.5-3.0 μH shielded variable inductor. (J.W. Miller 9050 or equiv.).
M1 — 100 microamperes full scale.
Q1, Q2 — MPF-102.
Q3 — 2N2222 or equiv.
S1 — 4pdt slide switch (Radio Shack 275-405).
S2 — spdt toggle switch.
T1 — Miniature interstage transformer, 2000-ohm ct to 10,000 ohms (Radio Shack 273-1378).
U1 — CA3028A.
U2 — CA3046 or CA3045.
U3 — HEP C6010 or MFC 4010A.
U4 — Dual 741 op amp (Radio Shack 276-038 or equiv.).
U5 — LM301A op amp.
U6 — CA3600E.
VR1 — 8.2-volt, 1-watt Zener diode.

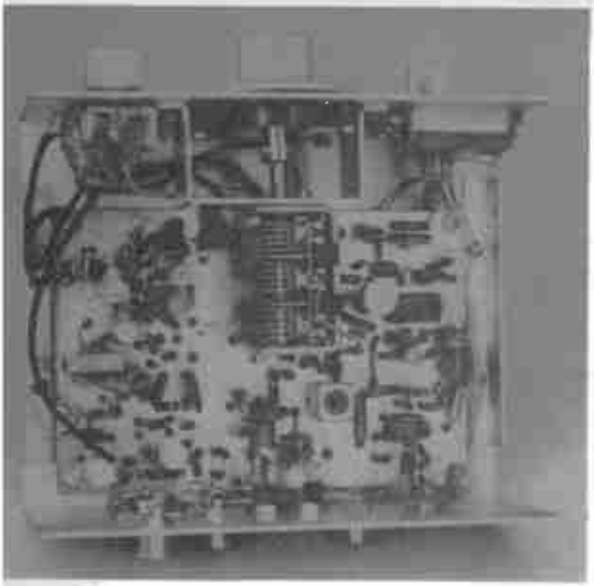


Fig. 2 — Inside view of the receiver. Most of the components are mounted on a 4- × 6-inch printed-circuit board. The rf amplifier and product detector components are grouped together at the upper left corner of the board, while the audio channel occupies the left foreground. The VFO main-tuning capacitor, C24, is centered in the cabinet, and is positioned directly over the VFO. The S-meter amplifier circuitry is visible at the right.

CA3046 have a rated F_T of 550 MHz. R30, R37, and R45 are used to prevent vhf parasitic oscillations from occurring. S1B may be used to select either 40- or 20-meter output from the local oscillator, and S1C applies 12 volts dc to the frequency doubler portion of the oscillator for 20-meter operation. With the component values shown, the receiver covers 7.0 to 7.15 MHz and 14.0 to 14.3 MHz. A miniature interstage transformer, T1, is used to couple the output of the product detector to the audio channel. Q2 functions as a moderate-gain, low-noise audio preamplifier, which is followed by the integrated high-gain amplifier U3. Q3 has been included to allow for muting of the receiver during transmitting periods by the application of 12 volts dc to R19 (in series with the base of Q3) by means of a contact on the transmitter's T-R relay. This biases Q3 into conduction and effectively breaks the circuit path between Q2 and U3. A miniature pot is used at R45, which serves as the af gain control. The setting of R45 determines the input level to U4A-U4B. A small part of the output of U3 (taken off before the af gain control) is used to drive the S-meter circuitry. Audio selectivity for the receiver is provided by two cascaded active filter sections consisting of a dual 741 op amp, U4A-U4B, plus associated passive components. With the values shown in the schematic each individual filter section will have a peak response at 840 Hz and a bandwidth of 375 Hz (measured at 6 dB below peak response). When two sections are cascaded, a narrower filter with a bandwidth of 200 Hz results. It is advantageous to match the peak frequency response of each filter section closely by hand picking component values[1] to achieve optimum filter performance. Provision is made for using either a broad or a narrow response by using S2 to switch the headphones to either the output of U4A or U4B, respectively. The active filter drives a pair of high-impedance headphones directly. Provision for monitoring the transmitter sidetone is included by introducing the tone into the audio channel beyond the muting transistor, Q3. The audio-derived S-meter circuit uses a single-stage audio filter, followed by a meter amplifier. The filter is similar to one section of the audio channel filter and is necessary to assure that the meter indication is a function of the signal being monitored and not the result of extraneous signals at the output of U3. Thus, the S-meter reading does not vary with the setting of the af gain control, and does not operate while the sidetone is being monitored. The output of the S-meter filter drives a three-section integrated-circuit amplifier, U6. The rectified amplifier output drives a 100 micorampere full-scale meter for the signal strength indication. Pc mounted pots are used at R52 and R60 for S-meter adjustment (sensitivity and zero).

Construction

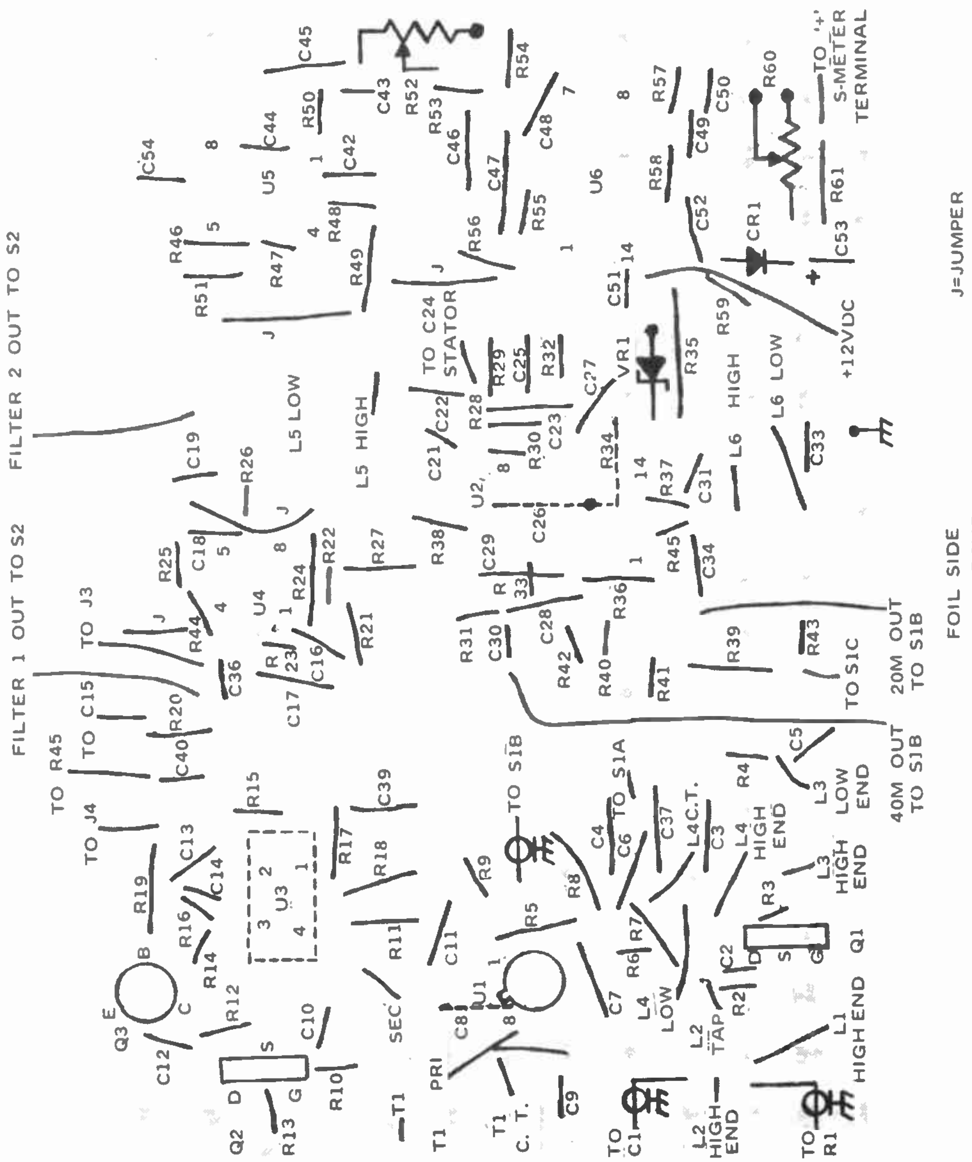
Construction of the receiver is greatly simplified by the use of an etched printed-circuit board for mounting most of the parts. A template is available from ARRL for 50 cents and an s.a.s.e. The entire receiver fits on a 4- × 6-inch board. If no S-meter is desired, it is a simple matter to adjust the layout to fit on a 4- × 4-1/2-inch board leaving out U5, U6, and their associated components. A

[1] Components plus a circuit board for the audio filter may be obtained from MFJ Enterprises, P.O. Box 494, Mississippi State, MS 39762

Fig. 3 — Front view of the direct-conversion receiver. The cabinet is homemade from two U-shaped pieces of .040-inch thick sheet aluminum. The front panel is painted battleship gray, and white "press-on" labels mark the function of the controls.

photo-etch process was used to produce the original board although it may be duplicated by other methods as long as sufficient care is taken in the vicinity of the IC pins and other high-density areas of the pattern to avoid the appearance of unwanted foil bridges. The prototype receiver was built on double-sided G-10 glass epoxy board, 1/16-inch thick. The circuit pattern is etched on the bottom of the board while the top is left as a continuous ground plane broken only where component leads project through the board. The ground plane is an aid to stability and interstage isolation. An easy technique for removing the ground plane around the component leads (after the bottom of the board has been drilled) is to use a large diameter drill (1/4-inch is satisfactory) and make a shallow hole in the top side of the board at every lead location. This may be done by hand, or very carefully with a drill press. The prototype receiver board was silver plated before component assembly, a step which while not required, makes soldering easier, and improves the appearance of the final product. Part of the key to building a compact receiver is the use of parts which are physically small. The use of small 50-volt disk ceramic capacitors can go a long way toward increasing packing density. Using miniature low-voltage electrolytic capacitors, toroidal inductors,

Fig. 4 — Parts placement and board layout for the receiver. R34, C8 and C26 are mounted on the foil side.



and 1/4-watt instead of 1/2-watt resistors where applicable will make construction much easier for the builder of portable equipment.

The placement of the front- and rear-panel-mounted parts is determined as much by symmetry as by the criterion of short leads. In the author's receiver, the S-meter, controls for af and rf gain, the band switch, input trimmer capacitor, selectivity switch, and the main tuning knob are located on the front panel. The rear panel includes banana jacks for the 12-volt dc input, phono jacks for sidetone and muting inputs from the transmitter, and a phone jack for headphones. For operating convenience, an SO-239 coax receptacle and a female phono jack wired in parallel were used as antenna connectors. The VFO tuning capacitor, C24, is mounted on an aluminum bracket from the front panel to achieve mechanical stability. The capacitor drive is a modified imported vernier dial. The original escutcheon, calibrated 0-100, was replaced with a homemade plastic dial. Fifty- and 25-kHz markers on the dial are made with thin black tape of the type usually used for printed-circuit artwork. The homemade U-shaped cabinet is spray painted with battleship gray enamel and white pressure-sensitive labels were added to identify the controls. The top cover is painted flat black. Professional-looking front-panel knobs complete the mechanical assembly.

Initial Adjustment

The local oscillator should be checked first for proper operation on 40 meters. With C24 almost fully meshed, L5 may be adjusted with a nonmetallic tool to set the output frequency to 7.000 MHz. A drop of melted wax will be sufficient to hold the slug in place. The frequency doubler portion of the local oscillator should be checked with an oscilloscope. L6 may be adjusted to provide the cleanest 14-MHz waveform. A signal generator and an oscilloscope may be used to verify that the audio channel is functioning. Assuming that parts tolerances were adhered to closely, the audio filter width and center frequency in both the broad and narrow positions should be comparable to the results mentioned above. With the aid of a signal generator or a weak on-the-air signal, the front-end response should be peaked first on 20 meters by adjusting C1 and C3, and then on 40 meters by adjusting C1 and C4. In actual operation, C1 is a front-panel control, and is tuned to the frequency band of interest. After C3 and C4 are set, it should not be necessary to repeak them. If the product detector is working correctly it should be possible to hook a pair of headphones to J1 or J2 and make these adjustments by ear. The S-meter zero and sensitivity controls can be adjusted according to operator preference.

THE MINI-MISER'S DREAM RECEIVER

A receiver that featured good performance with a modest outlay of building time and components appeared in the Forty-Fifth Edition of *The Radio Amateur's Handbook*. The original design was that of Byron Goodman, W1DX, and incorporated vacuum tubes in the construction.

This updated version by Doug DeMaw, W1CER, utilizes the same principle of going directly into a mixer in the front end which is followed by a simple crystal filter. Solid-state construction is used throughout the design and audio output is sufficient to drive a small speaker. This receiver should fill the need for a simple, compact unit where low current drain is also a requirement.

Circuit Description

There are some departures from the W1DX design, mainly to minimize cost and package size.

The major compromise was the elimination of agc and multiband coverage. There is ample room inside the cabinet of this receiver to accommodate one or two small converters for reception of bands other than 40 meters. This main frame is designed for 7- to 7.175-MHz coverage.

Fig. 1 shows an IC being used as the receiver front end – a CA3028A which is configured as a balanced mixer. The input tuned circuit, T1, is designed to match a 50-ohm antenna to the 2000-ohm base-to-base impedance of the mixer IC. The transformer is broadbanded in nature (300 kHz at the 3-dB points), and has a loaded Q of 23. This eliminates the need for a front-panel peaking control – a cost-cutting aid to simplicity.

The output tuned circuit, L1, is a bifilar-wound toroid which is tuned approximately to resonance by means of a mica trimmer, C2. The actual setting of C2 will depend upon the degree of i-f selectivity desired, and typically the point of resonance will not be exactly at 3300.5, the i-f center frequency.

Goodman used a half-lattice filter (two crystals) in his design, but this requires two crystals which are related properly in addition to a BFO crystal. For this reason, an older circuit was employed – a single crystal filter with a phasing capacitor, C3. The latter approach provides reasonably good single-signal reception (at least 30 dB rejection of the unwanted response), and assures much better performance than is possible with simpler direct-conversion receivers.

A single i-f amplifier, U2, is used to provide up to 40 dB of gain. R1 serves as a manual i-f gain control, and will completely cut off the signal output when set for minimum i-f gain. No audio gain control is used. T2 is designed to transform the 8000-ohm collector-to-collector impedance of U2 down to 500 ohms, and has a bandwidth of 100 kHz. The loaded Q is 33.

A two-diode product detector converts the i-f energy to audio. BFO injection voltage is obtained by means of a crystal-controlled oscillator, Q2. RFC2 and the 1-μF bypass capacitor filter the rf, keeping it out of the audio line to U3.

Audio-output IC U3 contains a preamplifier and power-output system. It will deliver approximately 300 mW of af energy into an 8-ohm load. RFC5 is used to prevent rf oscillations from occurring and being radiated to the front end and i-f system of the receiver. The 0.1-μF bypass at RFC5 also helps prevent oscillations.

A three-terminal voltage regulator, VR1, supplies the required operating voltage to U3. It also provides regulated voltage for the VFO and buffer stages of the local oscillator (Q2 and Q3). The latter consists of a stable series-tuned Clapp VFO and an emitter-follower buffer stage. A single-section pi network is placed between the emitter of Q3 and the injection terminal of U1. It has a loaded Q of 1, and serves as a filter for the VFO output energy. It is designed for a bilateral impedance of approximately 500 ohms. The recommended injection-voltage level for a CA3028A mixer is 1.5 rms. Good performance will result with as little as 0.5 volt rms. A 1-volt level is available with the circuit shown in Fig. 1.

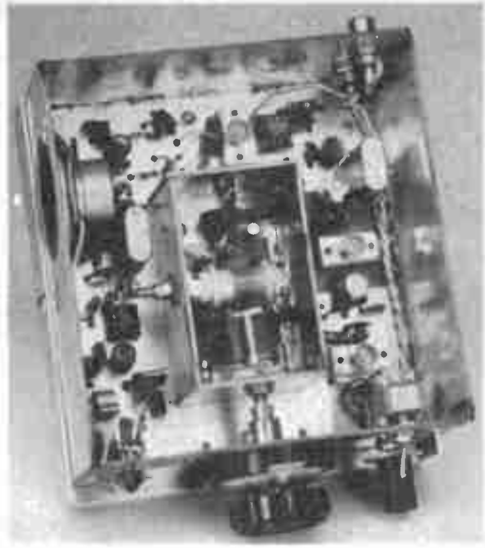
A red LED is used at DS1 as an on-off indicator. Since it serves mainly as "window dressing," it need not be included in the circuit.

Construction Notes

Front panel, rear panel, side brackets, and chassis are made from double-sided circuit-board material. The chassis is an etched circuit board, the pattern for which is given in Fig. 2. There is no reason why the top and bottom covers for the receiver can not be made of the same material by soldering six pieces of pc board together to form two U-shaped covers.

The local oscillator is housed in a compartment made from pc-board sections. It measures (HWD) 1-3/8 x 1-5/8 x 2-3/4 inches. A 1/4-inch high pc-board fence of the same width and depth is soldered to the bottom side of the pc board (opposite the top partition) to attenuate rf energy from entering or leaving the local oscillator section of the receiver. Employment of the top and bottom shields stiffens the main pc board, and that helps prevent mechanical instability of the oscillator, which can result from stress on the main assembly.

Silver plating has been applied to the main pc board, and to the front and rear panels. This was done to enhance the appearance and discourage tarnishing of the copper. It is not a necessary step in building the receiver. The front panel has been sprayed with green paint, then baked for 30 minutes by means of a heat lamp. A coarse grade of sandpaper was used to abrade the front panel before application of the paint. The technique will



Interior view of the receiver. The front end is at the lower right. The leads of U1 are bent to align with an 8-pin dual-in-line IC socket. The rim of the speaker is tack-soldered to the pc-board side wall at two points. The 20-meter converter mounts on the rear wall inside the receiver (upper left corner).

prevent the paint from coming off easily when the panel is bumped or scratched. Green Dymo tape labels are used to identify the panel controls.

There is ample room inside the cabinet, along the rear inner panel surface, to install a small crystal-controlled converter for some other hf band. A switch, S1, is located on the front panel to accommodate a planned 20-meter converter. A suitable circuit is given in Fig. 3.

All of the toroidal inductors are coated several times with Q dope after they are installed in the circuit. The VFO coil is treated in a like manner. The polystyrene VFO capacitors should be cemented to the pc board after a circuit is tested. This will help prevent mechanical instability. Hobby cement or epoxy glue is OK for the job. Use only a drop or two of cement at each capacitor – just enough to affix it to the pc board.

Alignment and Operation

The VFO should be aligned first. This can be done by attaching a frequency counter to pin 2 of U1. Coverage should be from 3699.5 to 3874.5 kHz for reception from 7.0 to 7.175 MHz. Actual coverage may be more or less than the spread indicated, depending on the absolute values of the VFO capacitors and stray circuit inductance and capacitance. Greater coverage can be had by using a larger capacitance value at C5, the main tuning control. Those individuals interested only in phone-band coverage can align the VFO accordingly and change Y2 to 3400.8 kHz.

Final tweaking is effected by attaching an antenna and peaking C1, C2 and C4 for maximum signal response at 7085 kHz. To obtain the selectivity characteristics desired (within the ca-

Fig. 1 — Schematic diagram of the 40-meter receiver. Fixed-value capacitors are chip or disk ceramic unless noted otherwise. Capacitors with polarity marked are electrolytic. S.M. indicates silver mica, and P is for polystyrene. Fixed-value resistors are 1/4- or 1/2-W composition.

C1, C2, C4 — 170- to 600-pF mica trimmer (Arco 4213).

C3 — 10-pF subminiature trimmer. Ceramic or pc-mount air variable suitable.

C5 — Miniature air variable, 30 pF maximum (Millen 25030E or similar).

CR1-CR3, incl. — High-speed silicon switching diode.

J1, J3 — Single-hole-mount phono jack.

J2 — Closed-circuit phone jack.

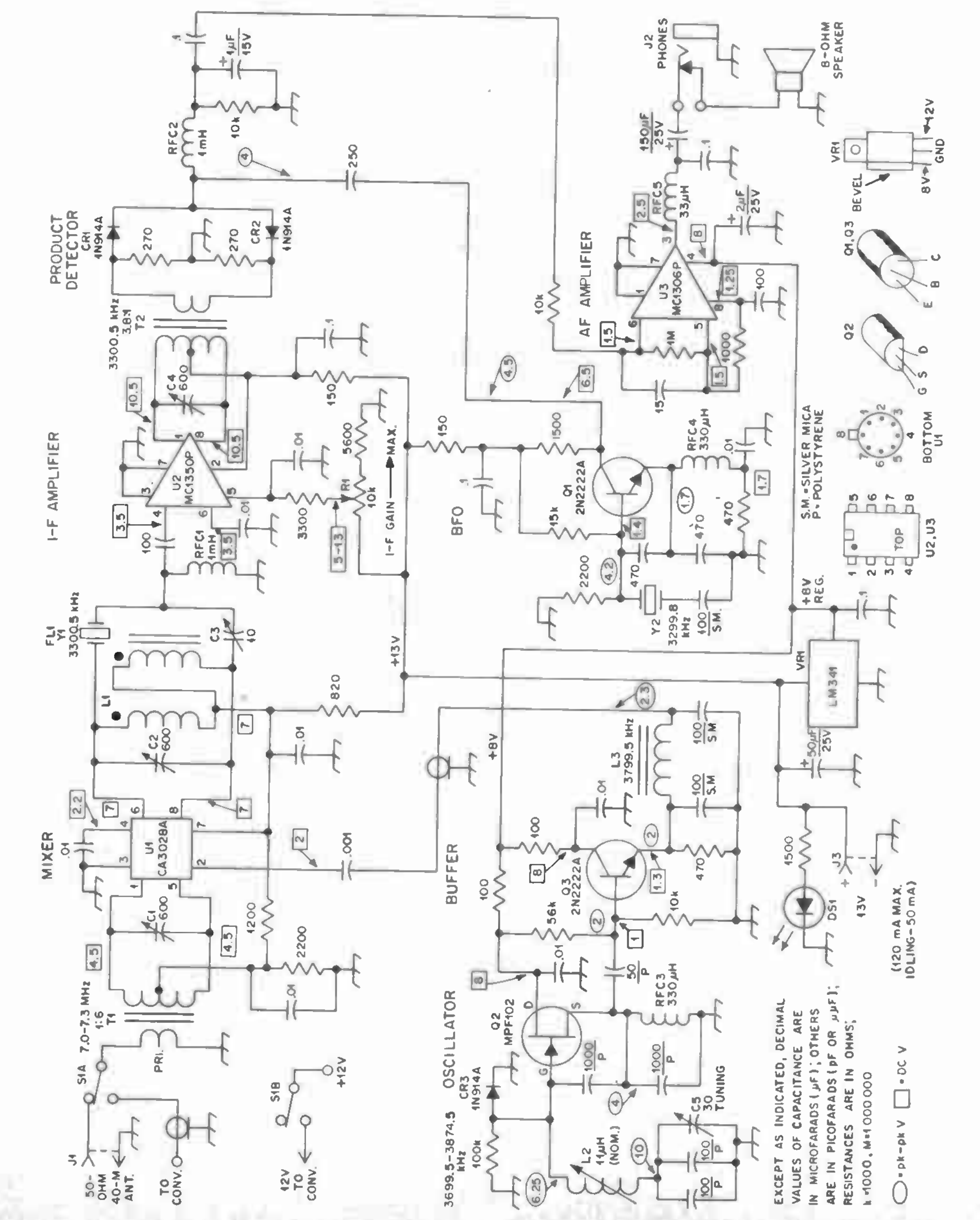
L1 — Toroidal bifilar-wound inductor. L = 5.8 μH. 8 turns No. 28 enam., bifilar wound on Amidon FT-37-61 ferrite core. Note polarity marks.

L2 — Slug-tuned inductor (see text), 11 μH nominal. J. W. Miller 42A105CBI or equiv. Q_u = 125.

L3 — Toroidal inductor, 17 μH, 19 turns No. 26 enam. wire on Amidon FT-50-61 ferrite core.

R1 — 10,000-ohm miniature composition control, linear taper.

RFC1, RFC2 — Miniature 1-mH rf choke (Millen



J302-1000 or equiv.).
RFC3, RFC4 — Miniature 330-μH rf choke (Millen J302-330 or equiv.).
RFC5 — Miniature rf choke, 33 μH (Millen J302-33 or equiv.).
S1 — Miniature dpdt toggle.
T1 — Toroidal transformer. Primary has 2 turns No. 24 enam. wire. Secondary has 14 turns No. 24 enam. wire on Amidon T-50-2 core.
T2 — Toroidal transformer. Primary has 9 turns No. 26 enam. wire on Amidon FT-37-61 core. Secondary has 3 turns No. 26 enam. wire. Primary winding has center tap.
U1 — RCA IC. Bend pins to fit 8-pin dual-in-line IC socket.
U2, U3 — Motorola IC.
VR1 — Three-terminal 8-volt regulator IC (National Semiconductor).
Y1, Y2 — Surplus crystal in HC-6/U case or International Crystal Co. type GP with 32-pF load capacitance.

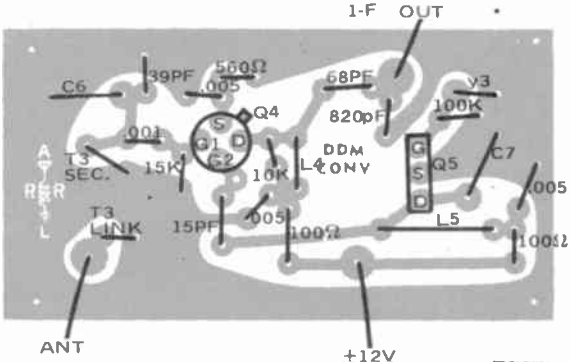
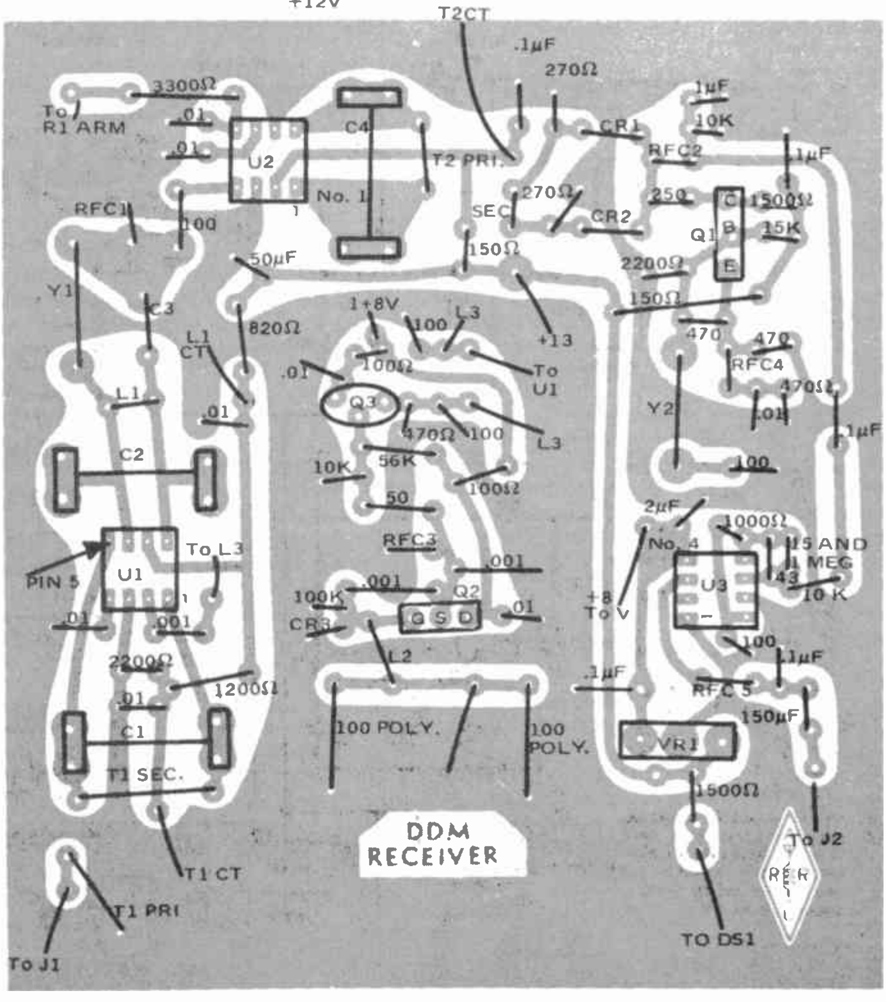


Fig. 2 — Foil-side scale pattern of the pc board. Circuit board is double-sided glass-epoxy material. Ground-plane copper should be removed directly opposite Q2 and related components (oscillator) for an area of 1-1/2 X 1-1/2 inches. Remove copper in similar manner on ground-plane side of board opposite L1, C3 and Y1 (1 X 1-1/4 inch area). Removal of foil will prevent unwanted capacitive effects in those critical parts of the circuit. Ground-plane side of board should be electrically common to the ground foils on opposite side of board at several points.



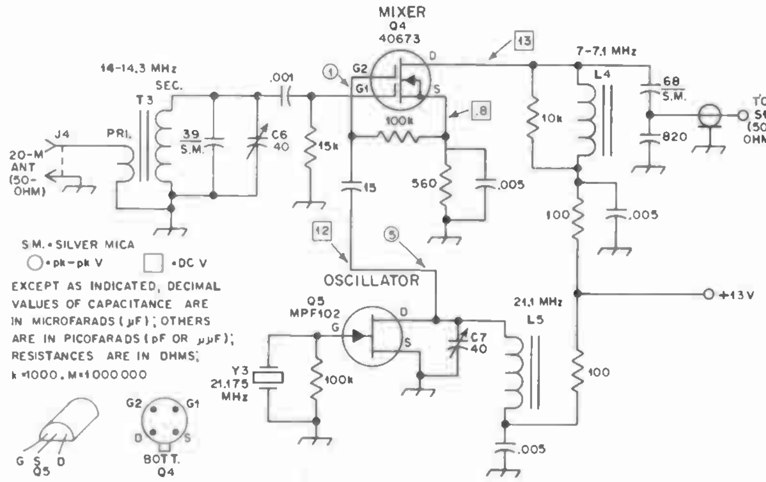


Fig. 3 — Schematic diagram of the 20-meter converter. Fixed-value capacitors are disk ceramic unless noted otherwise. Resistors are 1/4- or 1/2-W composition. Rough pc template and layout available for 25¢ and an s.a.s.e.

C6, C7 — 40-pF subminiature ceramic trimmer.

J4 — Single-hole-mount phono jack on rear panel of main receiver.

L4 — Toroidal inductor. 12 turns No. 26 enam. wire on Amidon FT-37-61 core. L = 8 μH.

L5 — Toroidal inductor. 24 turns No. 26 enam. wire on Amidon T-50-6 core. L = 2.4 μH.

Q4 — RCA transistor.

Q5 — Motorola transistor, MPF102, 2N4416, or HEP 802.

T3 — Toroidal transformer, 10:1 turns ratio. L = 1.85 μH. Pri. has 2 turns No. 26 enam. wire. Sec. contains 21 turns No. 26 enam. wire on Amidon T-50-6 core.

Y3 — 21.175-MHz fundamental crystal in HC-18/U case (International Crystal Co. type GP with 32-pF load capacitance).

pability of the circuit), adjust C2 and C3 experimentally. C2 will provide the major effect. C3 should be set for minimum response on the unwanted side of zero beat. A fairly strong signal will be needed to hear the unwanted response.

For reception of lower sideband, it will be necessary to use a different BFO frequency — 3400.5 kHz. The crystal indicated in Fig. 1 was used because it was the only one available at the time of construction. Those wishing to shift the BFO frequency a few hundred Hz can place a trimmer in series with Y2 rather than use the 100-pF capacitor shown.

Because there is no agc in this receiver, the i-f gain should be set low, for comfortable listening. Too much gain will cause the audio circuit to be overdriven, and distortion will result. To prevent ear-splitting signal levels, one can install a pair of 1N34A diodes (back to back) across the output jack, J2.

Bits and Pieces

The photograph shows some fancy looking components on the circuit board. Tantalum capacitors are seen where electrolytics are indicated on the diagram. Either type will work nicely. Tantalums were found at a flea market for 10 cents each, so they were used. Similarly, the 0.1-μF capacitors used are the high-class kind (Aerovox CK05BX), which sell for roughly 70 cents each. Mylar or disk ceramic 0.1-μF units will be fine as substitutes.

The polystyrene capacitors were obtained from Radio Shack in an assortment pack. New units are made by Centralab, and they sell for less than 20 cents each in single lots. Since they are more stable than silver micas, they are recommended for the VFO circuit.

All of the toroid cores were purchased by mail from Amidon Associates. A J. W. Miller 42-series coil is used in the VFO, but any slug-tuned ceramic form can be used if it has good high-frequency core material. The unloaded Q of the inductor should be at least 150 at 3.5 MHz. L2 in this design has 3/8-inch diameter body. The winding area is 5/8 inch long.

The metal cases of both crystals should be connected to ground by means of short lengths of wire. This will prevent unwanted radiation from the BFO crystal, and will help keep the filter crystal from picking up stray energy. A metal cover should be placed on the VFO compartment for reasons of isolation.

James Millen encapsulated rf chokes are used in the receiver. Any subminiature choke of the approximate inductance indicated will be suitable and it need not be encapsulated. The VFO tuning capacitor is also a Millen part. Ample room exists between the VFO box and the front panel to allow making the box longer. That will permit use of a larger variable capacitor. A double-bearing capacitor is recommended for best mechanical stability of the VFO.

The i-f system and BFO can be tailored to frequencies other than those indicated. The VFO, mixer, and i-f amplifier tuned circuits will have to be altered accordingly, if crystals of other frequencies in the 2- to 3-MHz range are chosen.

Performance of this receiver is quite good. A 0.1-μV signal from a generator is plainly audible. No hum or distortion is heard in the output of the receiver at normal listening levels. VFO drift is 45 Hz from a cold start to stabilization, and strong signals do not pull the oscillator.

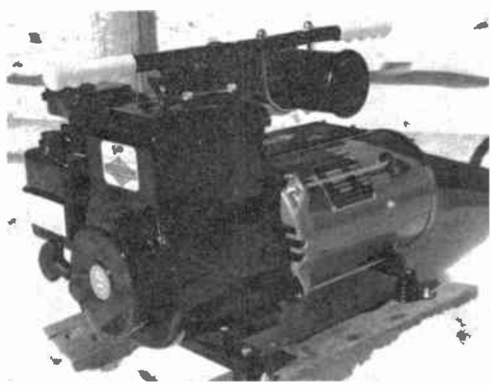
Extremely strong local signals (1000 μV or greater) will cause desensitization of the receiver when they appear off frequency from where the operator is listening. Under ordinary conditions this will not be a problem. At some sacrifice in noise figure and sensitivity, those living in areas where other amateurs are nearby can modify T1 to aid the situation. C1 should remain across all of the T1 secondary, and a 2200-ohm resistor should be connected across C1. Pins 1 and 5 of U1 should be connected to two turns each side of the center tap of the secondary. This will require cutting the pc board elements to divorce pins 1 and 5 from C1. This design trade-off is quite acceptable at 40 meters, as the atmospheric noise level will mask the reduction in receiver noise performance. With the circuit change there was no desensing evident below approximately 8000 μV.

Agc could be used in this receiver by applying the audio-derived "hang" type used by Goodman in his *Handbook* design. If the feature was adopted, agc voltage would be applied to pin 5 of U2 and the manual gain control would be eliminated. In such a case, it would be necessary to add an af gain control between the product detector and U3. It should be remembered that minimum gain results when 13 volts are applied to pin 5 of U2. The lower the voltage at that point, the greater the gain.

This Mini Miser's Dream may be just what you've been wanting for that next camping trip. Since it measures only 2-5/8 X 4-3/4 X 5 inches, it should fit easily into a rucksack, along with a battery pack (maximum current is 120 mA). Or, maybe you're trying to get that code speed peaked for a higher license class. If so, this little fella might be the right size for the night stand by the bed – assuming the XYL doesn't object!

PORTABLE AC GENERATORS

Alternators powered by internal-combustion engines have been used for years to supply 117/234 ac independently of the commercial mains. Such combinations range from tiny units powered by 2-cycle gasoline engines in the "model-airplane" class to giant multicylinder diesels capable of supplying megawatts of power. Perhaps the most practical power range for most purposes would be in the neighborhood of 2 kW. Larger units tend to become too heavy for one person to lift and handle easily while smaller generators lack sufficient power output for many applications. A 2-kW alternator is no fun to lift by any means, but it is capable of supplying power for just about any tool such as a belt sander, saw or electric drill. It is roughly the equivalent of having a single 15-A outlet in an ordinary electric service. Of course, it will handle moderate-power amateur equipment with ease.



Tough when it comes to delivering power, alternators such as this one need pampering in their maintenance. Long life and maximum performance are assured if a few simple rules are observed.

The Neglected Tool

Since an alternator represents a considerable investment, proper maintenance is important in avoiding premature failure because of neglect. Typical amateur operations requiring a generator often tend to be of an intermittent nature which is the worse kind of service imposed on an internal-combustion engine. For instance, relegating the generator to the garage after Field Day is over and forgetting about it till next year is asking for trouble. This is why it is desirable to obtain a unit that can be used for a variety of purposes rather than just a single one such as for portable amateur activities. Through frequent use, many tough technical problems related to prolonged storage can be avoided. While the manufacturer's instructions should be read and observed faithfully, the following checklist should be of aid in either supplementing the instructions or as a general guide in cases where such information is unavailable.

MAINTENANCE CHECKLIST

Although more complicated maintenance chores should be performed by qualified service personnel, many simple measures that will prolong the useful life of the alternator can be done at home. Perhaps the best plan is to keep a log of the dates when the unit was used, and the operating time in hours. Also included in the log would be dates of maintenance and type of service performed. Oil changes, when gasoline was purchased for emergency purposes, and similar data would fall under this category.

Important points that are common to all types of generators are indicated for a typical one in Fig. 1. (Consult the manufacturer's manual for additional instructions that might apply to a particular

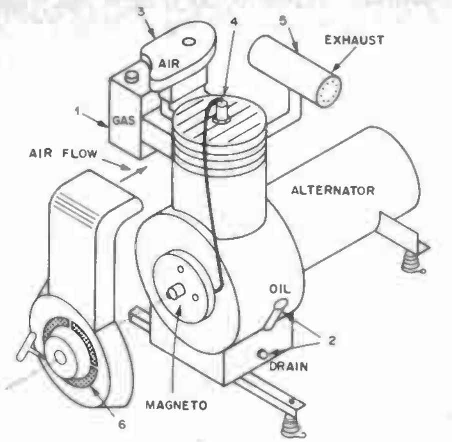


Fig. 1 — Critical points for routine generator maintenance. See text for details.

model.) The following checklist relates to the numbers on the drawing.

1) Use the proper grade of fuel. Newer models will burn either "no-lead" or regular leaded gasoline. Do not use premium or so-called "hi-test" grades unless the owner's manual recommends it. Such fuels have a high lead content for proper burning in high-compression automobile engines and are generally unsuitable for small, low-compression engines found in most alternator combinations. Check the owner's manual to determine whether or not oil must be mixed with the gasoline. While *two-cycle* models require an oil-gas mixture, most generators have a *four-cycle* engine that burns ordinary gasoline with no extra additives. Gasoline for emergency purposes should only be stored in small amounts and rotated on a regular basis. Older stock can be burned in a car (that uses the same grade of gas as the generator) since storing gasoline for any length of time is inadvisable. The more volatile components evaporate leaving excess amounts of a varnish-like substance that will clog carburator passages. Also, be sure gasoline containers are of an approved type with a clean interior, free of rust or other foreign matter. Similar considerations apply to the gas tank on the engine itself.

The majority of difficulties with small engines are related to fuel problems in some way. Dirty fuel or water in the gasoline are one source with carburetor trouble because of the use of old gas being another common cause. Except for minor adjustments recommended in the instruction manual, it is seldom necessary to touch the carburetor controls. Avoid the temptation to make such adjustments in the case of faulty operation. *Do follow* the recommendations in this guide so that more complicated maintenance procedures (such as carburetor overhaul) are not required.

2) Another important factor often neglected in maintenance of alternator engines is oil. While lubrication is one job oil has to perform, there are other considerations as well. The engine oil in the crankcase also collects a large amount of solid combustion products, bits of metal worn away by the moving parts, and any dust or other foreign matter that enters the carburetor intake. For instance, it is especially important to observe the manufacturer's recommendations concerning the length of time the engine may be operated before an oil change is required during the break-in period. If you ever have the opportunity to examine the oil from a new engine, you will note a metallic sheen to it. This is from the excessive amount of metal that is worn away. After the break-in period, much less metal is abraded and the oil doesn't have to be replaced as often.

The oil level should be checked frequently during engine operation. Each time fuel has to be added the oil should be checked also. When storing an alternator, it is also wise to drain the oil and replace it with fresh stock. This is because one of the combustion products is sulfur which forms sulfuric acid with water dispersed in the oil. The acid then attacks the special metal in the bearing surfaces causing pitting and premature replacement.

Also note the grade and weight of oil recommended by the manufacturer. Unlike their larger counterparts in the automobile, most small engines do not have oil filters which is another reason why required changes are more frequent. Some manufacturers recommend a high-detergent oil that comes in various service grades such as MS, SD, and similar types. Examine the top or side of the cans in which the oil is sold and see if the letters correspond to those recommended by the engine manufacturer.

3) Gasoline is mixed with air in the carburetor which is then burned in the engine. Before entering the carburetor, the air must be filtered so that it is free of dust and other foreign matter that might otherwise be drawn into cylinder(s). Particles that do get by the air filter are picked up by the oil which should be changed more often if the alternator is operated in a dusty location. Also, it is important to clean the air filter frequently. It contains a foam-like substance which can be cleaned in kerosene and then soaked in fresh motor oil. Squeeze excess oil from the filter before replacing. Also consult the instruction manual for further recommendations.

4) Once the gas/air mixture enters the cylinder, it is compressed by the piston into a very small volume where it is then ignited by the spark plug. During the rapid burning that then occurs, the expansion caused by the resulting heat forces the piston down and delivers the mechanical power to the alternator.

As might be expected, proper operation of the ignition system is an important factor in engine performance. Power for the spark is supplied by a device called a magneto that is normally installed on the front of the engine. The magneto seldom requires servicing and such work should only be done by those qualified to do so. (This is one reason why the magneto is often located under a flywheel that is difficult to remove by the inexperienced.)

On the other hand, faulty spark plugs are the usual cause of ignition problems. Special equipment is required to test a spark plug properly, but an easier solution is to have a new one handy. In fact, keep *two* spare plugs on hand. Spark-plug life can be notoriously short on occasion. However, repeated plug failure is also abnormal and other causes such as a poor gas/air mixture might be the culprit.

Replace the spark plug with a type similar to the one that came with the alternator or a substitute recommended by the manufacturer. Some models have resistor-type plugs which are desirable for ignition-noise suppression. Resistor plugs are usually indicated by an R prefix. For instance, the resistor version of a Champion CJ-8 would be an RC-J8

5) Little maintenance is required in regards to the exhaust system. In some forested areas, a spark-arrester type of muffler is required so be sure that your unit is so equipped *before* contemplating operation in such a location. "Quiet hours" may also be imposed in some places during the nighttime hours if generator exhaust noise is too loud.

Two very important safety precautions should be observed in regards to the exhaust system. *Never* operate an alternator in closed surroundings such as a building. Dangerous gases are emitted from the exhaust which are highly toxic. Secondly, *never* refuel an engine while it is running or if the exhaust system is still very hot. Unfortunately, this last precaution is disregarded by many which is extremely foolish. The time saved by not observing proper precautions in the handling of gasoline might be made up in the severe-burn ward of the local hospital – usually a painful prelude to the funeral home. (Experienced service-station operators will refuse to refuel an automobile with the motor running which is often *prohibited by law*.) Don't become an unnecessary statistic.

6) Most alternators are air-cooled as opposed to the water-cooled radiator system of the automobile. A fan on the front of the engine forces air over the cylinder and an unobstructed entrance for this air flow is necessary. Avoid operating the alternator in areas where obstruction to this flow might result (such as in tall grass). Alternators should be operated such that a sufficient amount of air circulation is present for cooling, carburetion and exhaust.

Storage

Proper maintenance of an alternator when it is not being used is just as important as during the time it is in operation. The usual procedure is to run the engine dry of gasoline, drain the crankcase and fill it with fresh oil, and remove the spark plug. Then pour a few tablespoons of oil into the cylinder and turn the engine over a few times with the starter and replace the plug. But *never* crank the engine with the plug removed and the *ignition* or *start* switch in the *on* or *run* position. The resulting no-load high-voltage might cause damage to the magneto. It is also a good idea to ground the spark-plug wire to the engine frame with a clip lead in case the switch is accidently activated.

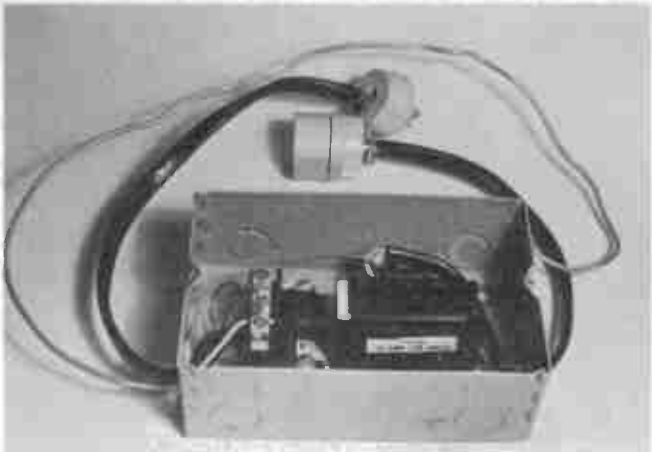
Check these areas for safe and trouble-free operation (also see Fig. 1 and text).

Moisture is the greatest enemy of an iron product in storage such as a generator. This is especially true with the precision-machined parts such as the piston(s) and cylinder wall(s). The coating of oil helps retard rust formation here which might actually weld the two surfaces together at worst or cause pitting. Also, rust particles abrade the surfaces when the engine is restarted resulting in premature wear. Consequently, it is important to store the alternator in an area of low humidity.

Although the maintenance procedures outlined may seem like a chore, the long-term benefits include low repair costs and like-new performance. Engines for alternator combinations must be able to handle a variety of loads while maintaining a constant speed in order to keep the output frequency constant. A mechanical governor performs this latter function by metering the fuel supplied to the engine under different load conditions. However, the system cannot function properly with an engine in poor mechanical condition because of lack of proper maintenance.

ELECTRICAL CONSIDERATIONS

If the mechanical aspects of generator maintenance seem a bit complicated, newer models make up for the fact by simplicity in electrical operation. Outside of observing wattage limitations of a particular model, little else has to be done. As mentioned earlier, a mechanical governor controls the speed (and frequency) while a voltage regulator in the alternator adjusts for variations in the load.



A circuit breaker and junction box is a convenient way of providing outlet and ground connections along with overload protection.

Some models have a variable voltage regulator but many others now have a sealed unit that is unadjustable. No adjustments should be made to the governor except by a service shop with personnel experienced in its operation.

Circuit-Breaker and Ground Accessory

Many alternators do not have provisions for overload protection such as fuses or circuit breakers. It is advisable to add external circuitry that will fulfill these functions and a diagram of such a unit is shown in Fig. 2. The enclosure can be either homemade or a commercially manufactured junction box. The latter is a convenient method since there is usually a mount for the fuse or circuit breaker built in with plenty of points where connections can be made for other wires.

Like fuses, circuit breakers come in standard sizes such as 15 and 20 A. A plug on one side of the junction box is connected to the generator output once it has come up to speed. Disconnect the plug or open the circuit breaker and allow the alternator to run for several minutes before turning it off (especially after using it to supply power continuously to a heavy load). At the output of the junction box, an outlet is installed either on the end of a short cable or a longer extension cord. But be sure the cord has a wire size that is large enough for the current and cable run contemplated.

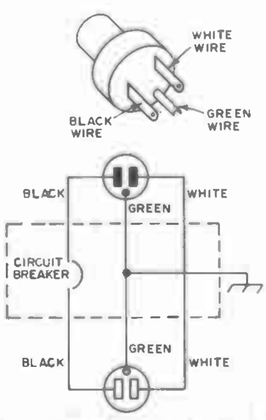


Fig. 2 — A simple accessory that provides overload protection for generators that do not have such provisions built in.

Conductor diameter depends upon the length of the cable and the intended current-carrying capacity. The longer the cable run, the greater the voltage drop. This voltage drop is just equal to twice the conductor resistance for a given length times the current. The resistance can be computed from various wire tables; however, Table 1 can also be used to compute the voltage drop directly. Figures under the conductor lengths are the voltage drops for a one-ampere ac or dc current. Voltage drops for other values of current can be found merely by multiplying the figures in the column by the desired current. For instance, a general rule is that the voltage drop in a given conductor run should be no greater than two percent. For 117-V operation, this would correspond to 2.34 volts.

Suppose it was desired to operate a piece of equipment that required 15 A and that the cable length was 50 feet. First, it must be determined if the desired current is within the maximum current limitation for the particular conductor size under consideration. These values are shown under the column labeled I_{max} in Table 1. Note that 15 A exceeds the ratings of No. 18 and No. 16 wire so these two choices are eliminated immediately.

The voltage drop across 50 feet of No. 14 wire would be 15 times 0.257 or 3.86 volts which exceeds the two-percent value of 2.34 V. The use of No. 12 wire reduces this drop to 2.43 V which would probably be close enough to 2.34 even though it exceeds the specified value by a slight amount. However, the drop in 50 feet of No. 10 wire would only be 1.53 volts which is well within the two-percent tolerable limit. Actually, a longer cable run would be possible with No. 10 wire

TABLE 1

Voltage Drop for a 1-A Current (77° F) in a Two-Conductor Copper Line as a Function of Wire Size and Length in Feet.

Length / *Wire Size*	*(I_{max})*	*25*	*50*	*75*	*100*
18	6	0.325	0.651	0.977	1.302
16	10	0.205	0.410	0.615	0.820
14	20	0.129	0.257	0.385	0.514
12	30	0.081	0.162	0.243	0.324
10	35	0.051	0.102	0.153	0.204
8	50	0.032	0.064	0.0961	0.128
6	70	0.0205	0.0410	0.0615	0.0820
4	90	0.0130	0.0259	0.0389	0.0518

which might be an advantage to using it because a later application might require a greater length. Since the variation of resistance with length is linear, the voltage drop varies linearly also. Therefore, it is possible to find the maximum permissable length of No. 10 wire that would be suitable for 15-A service merely by multiplying 50 by 2.34/1.53. This gives a length of 76.5 feet which of course is within the limit of the result if the figure in the 75-foot column is used in the initial computation. (The drop in a 75-foot length of cable would be 0.153 times 15 or 2.3 V.)

While there is no exact relation between the gauge number and actual wire size, a rule-of-thumb sort of approximation can be useful. A difference of ten gauge numbers is roughly equivalent of a change in cross-sectional area of 10 times. Thus, if 75 feet of No. 10 wire was sufficient for a 15-A service, the wire size would have to be increased to No. 0 if the desired length was 750 feet. As can be seen from the foregoing, there is a good incentive for transmitting power over long distances at as high a voltage as possible in order to keep the current low.

Grounds

Newer generators are supplied with a three-wire outlet and the ground connection should go to the plug as shown in Fig. 2. On older types, the ground would have to be connected separately to the generator frame and then to the common terminal in the junction box. A pipe or rod can then be driven into the ground and a wire connection made to either a clamp supplied with the rod or by means of a C-clamp for larger sizes of pipe. From an ignition-noise-suppression standpoint, the ground is desirable along with safety considerations when power tools are being used.

The ground connection goes to the green wire in commercially made three-wire conduit. Conduit purchased from an electrical store comes with a color-coded insulation and the colored wires should be connected as shown in Fig. 2. Consult the owner's manual for the generator for further details on power hookup that might apply to your particular model.

Code Transmission

Keying a transmitter properly involves much more than merely turning it on and off with a fast manually operated switch (the key). If the output is permitted to go from zero to full instantaneously (zero "rise" time), side frequencies, or key clicks, will be generated for many kilohertz either side of the transmitter frequency, at the instant the key is closed. Similarly, if the output drops from full to zero instantaneously (zero "decay" time), side frequencies will be generated at the instant of opening the key. The amplitude of the side-frequency energy decreases with the frequency separation from the transmitter frequency. To avoid key clicks and thus to comply with the FCC regulations covering spurious radiations, the transmitter output must be "shaped" to provide finite rise and decay times for the envelope. The longer the rise and decay times, the less will be the side-frequency energy and extent.

Since the FCC regulations require that " . . . the frequency of the emitted wave shall be as constant as the state of the art permits," there should be no appreciable change in the transmitter frequency while energy is being radiated. A *slow* change in frequency is called a frequency drift; it is usually the result of thermal effects on the oscillator. A *fast* frequency change, observable during each *dit* or *dah* of the transmission, is called a chirp. Chirp is usually caused by a nonconstant load on the oscillator or by dc voltage changes on the oscillator during the keying cycle. Chirp may or may not be accompanied by drift.

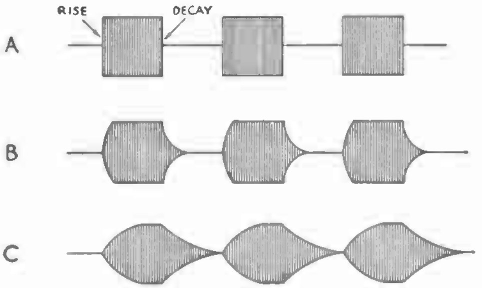


Fig. 11-1 — Typical oscilloscope displays of a code transmitter. The rectangular-shaped dots or dashes (A) have serious key clicks extending many kHz either side of the transmitter frequency. Using proper shaping circuits increases the rise and decay times to give signals with the envelope form of B. This signal would have practically no key clicks. Carrying the shaping process too far, as in C, results in a signal that is too "soft" and is not quite as easy to copy as B.

Oscilloscope displays of this type are obtained by coupling the transmitter rf to the vertical plates and using a slow sweep speed synchronized to the dot speed of an automatic key.

If the transmitter output is not reduced to zero when the key is up, a backwave (sometimes called a "spacing wave") will be radiated. A backwave is objectionable to the receiving operator if it is readily apparent; it makes the signal slightly harder to copy. However, a slight backwave, 40 dB or more below the key-down signal, will be discernible only when the signal-to-noise ratio is quite high. Some operators listening in the shack to their own signals and hearing a backwave think that the backwave can be heard on the air. It isn't necessarily so, and the best way to check is with an amateur a mile or so away. If he doesn't find the backwave objectionable on the S9+ signal, you can be sure that it won't be when the signal is weaker.

When any circuit carrying dc or ac is closed or opened, the small or large spark (depending upon the voltage and current) generates rf during the instant of make or break. This rf click covers a frequency range of many megahertz. When a transmitter is keyed, the spark at the key (and relay, if one is used) causes a click in the receiver. *This click has no effect on the transmitted signal.* Since it occurs at the same time that a click (if any) appears on the transmitter output, it must be eliminated if one is to listen critically to his own signal within the shack. A small rf filter is required at the contacts of the key (and relay); typical circuits and values are shown in Fig. 11-2. To check the effectiveness of the rf filter, listen on a band lower in frequency than the one the transmitter is tuned to, with a short receiving antenna and the receiver gain backed off.

What Transmitter Stage To Key

A satisfactory code signal, free from chirp and key clicks, can be amplified by a *linear* amplifier without affecting the keying characteristics in any way. If, however, the satisfactory signal is amplified by one or more nonlinear stages (e.g., a Class C multiplier or amplifier), the signal envelope will be modified. The rise and decay times will be decreased, possibly introducing significant key clicks that were not present on the signal before amplification. It is possible to compensate for the effect by using longer-than-normal rise and decay times in the excitation and letting the amplifier(s) modify the signal to an acceptable one.

Many two-, three- and even four-stage VFO-controlled transmitters are incapable of chirp-free output-amplifier keying because keying the output stage has an effect on the oscillator frequency and "pulls" it. Keying the amplifier presents a variable load to its driver stage, which in turn is felt as a variable load on the previous stage, and so on back

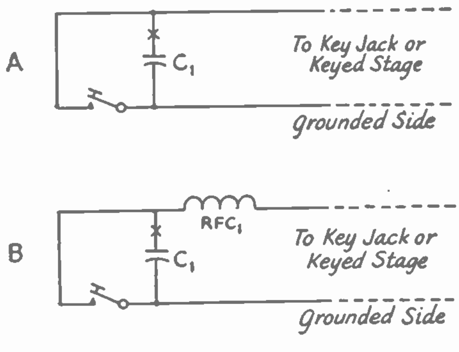


Fig. 11-2 – Typical filter circuits to apply at the key (and relay, if used) to minimize rf clicks. The simplest circuit (A) is a small capacitor mounted at the key. If this proves insufficient, an rf choke can be added to the ungrounded lead (B). The value of C1 is .001 to .01 μF; RFC1 can be 0.5 to 2.5 mH, with a current-carrying ability sufficient for the current in the keyed circuit. In difficult cases another small capacitor may be required on the other side of the rf choke. In all cases the rf filter should be mounted right at the key or relay terminals; sometimes the filter can be concealed under the key. When cathode or center-tap keying is used, the resistance of the rf choke or chokes will add cathode bias to the keyed stage, and in this case a high-current low-resistance choke may be required, or compensating reduction of the grid-leak bias (if it is used) may be needed. Shielded wire or coaxial cable makes a good keying lead.

A visible spark on "make" can often be reduced by the addition of a small (10 to 100 ohms) resistor in series with C1 (inserted at point "x"). Too high a value of resistance reduces the arc-suppressing effect on "break."

to the oscillator. Chances of pulling are especially high when the oscillator is on the same frequency as the keyed output stage, but frequency multiplication is no guarantee against pulling. Another source of reaction is the variation in oscillator supply voltage under keying conditions, but this can usually be handled by stabilizing the oscillator supply with a VR tube. If the objective is a completely chirp-free transmitter, the first step is to make sure that keying the amplifier stage (or stages) has no effect on the frequency. This can be checked by listening on the oscillator frequency while the amplifier stage is keyed. Listen for chirp on either side of zero beat, to eliminate the possibility of a chirpy receiver (caused by line-voltage changes or BFO pulling).

An amplifier can be keyed by any method that reduces the output to zero. Neutralized stages can be keyed in the cathode circuit, although where powers over 50 or 75 watts are involved it is often desirable to use a keying relay or vacuum tube keyer, to minimize the chances for electrical shock. Tube keying drops the supply voltages and adds cathode bias, points to be considered where maximum output is required. Blocked-grid keying is applicable to many neutralized stages, but it presents problems in high-powered amplifiers and requires a source of negative voltage. Output stages that aren't neutralized, such as many of the tetrodes and pentodes in widespread use, will usually leak a little and show some backwave regardless of how they are keyed. In a case like this it may be necessary to key two stages to eliminate backwave. They can be keyed in the cathodes, with blocked-grid keying, or in the screens. When screen keying is used, it is not always sufficient to reduce the screen voltage to zero; it may have to be taken to some negative value to bring the key-up plate current to zero, unless fixed negative control-grid bias is used. It should be apparent that where two stages are keyed, keying the earlier stage must have no effect on the oscillator frequency if completely chirp-free output is the goal.

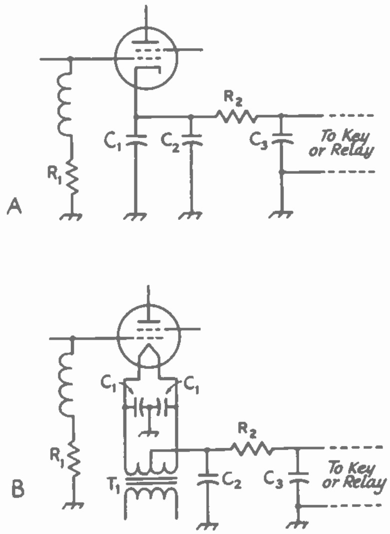


Fig. 11-3 – The basic cathode (A) and center-tap (B) keying circuits. In either case C1 is the rf return to ground, shunted by a larger capacitor, C2, for shaping. Voltage ratings at least equal to the cutoff voltage of the tube are required. T1 is the normal filament transformer. C1 and C3 can be about .01 μF.

The shaping of the signal is controlled by the values of R2 and C2. Increased capacitance at C2 will make the signal softer on break; increased resistance at R2 will make the signal softer on make.

Values at C2 will range from 0.5 to 10 μF, depending upon the tube type and operating conditions. The value of R2 will also vary with tube type and conditions, and may range from a few to one hundred ohms. When tetrodes or pentodes are keyed in this manner, a smaller value can sometimes be used at C2 if the screen-voltage supply is fixed and not obtained from the plate supply through a dropping resistor. If the resistor decreases the output (by adding too much cathode bias) the value of R1 should be reduced.

Oscillators keyed in the cathode can't be softened on break indefinitely by increasing the value of C2 because the grid-circuit time constant enters into the action.

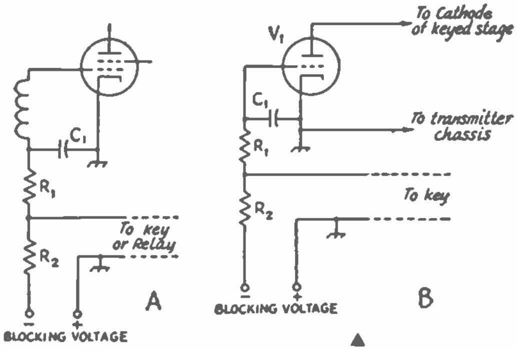


Fig. 11-4 — The basic circuit for blocked-grid keying is shown at A. R1 is the normal grid leak, and the blocking voltage must be at least several times the normal grid bias. The click on make can be reduced by making C1 larger, and the click on break can be reduced by making R2 larger. Usually the value of R2 will be 5 to 20 times the resistance of R1. The power supply current requirement depends upon the value of R2, since closing the key circuit places R2 across the blocking voltage supply.

An allied circuit is the vacuum-tube keyer of B. The tube V1 is connected in the cathode circuit of the stage to be keyed. The values of C1, R1 and R2 determine the keying envelope in the same way that they do for blocked-grid keying. Values to start with might be 0.47 megohm for R1, 4.7 megohms for R2 and .0047 μF for C1.

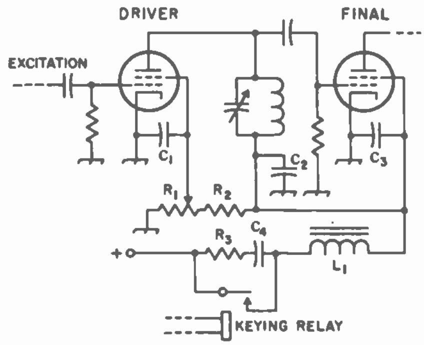
The blocking voltage supply must deliver several hundred volts, but the current drain is very low. A 6Y6 or other low plate-resistance tube is suitable for V1. To increase the current-carrying ability of a tube keyer, several tubes can be connected in parallel.

A vacuum-tube keyer adds cathode bias and drops the supply voltages to the keyed stage and will reduce the output of the stage. In oscillator keying it may be impossible to use a VT keyer without changing the oscillator dc grid return from ground to cathode.

Fig. 11-5 — When the driver-stage plate voltage is roughly the same as the screen voltage of a tetrode final amplifier, combined screen and driver keying is an excellent system. The envelope shaping is determined by the values of L1, C4, and R3, although the rf bypass capacitors C1, C2 and C3 also have a slight effect. R1 serves as an excitation control for the final amplifier, by controlling the screen voltage of the driver stage. If a triode driver is used, its plate voltage can be varied for excitation control.

The inductor L1 will not be too critical, and the secondary of a spare filament transformer can be used if a low-inductance choke is not available. The values of C4 and R3 will depend upon the inductance and the voltage and current levels, but good starting values are 0.1 μF and 50 ohms.

To minimize the possibility of electrical shock, it is recommended that a keying relay be used in this circuit, since both sides of the circuit are "hot." As in any transmitter, the signal will be chirp-free only if keying the driver stage has no effect on the oscillator frequency. (The Sigma 41FZ-35-ACS-SIL 6-volt ac relay is well-suited for keying applications.)



Shaping of the keying is obtained in several ways. Vacuum-tube keyers, blocked-grid and cathode-keyed systems get suitable shaping with proper choice of resistor and capacitor values, while screen-grid keying can be shaped by using inductors or resistors and capacitors. Sample circuits are shown in Figs. 11-3, 11-4, and 11-5, together with instructions for their adjustment. There is no "best" adjustment, since this is a matter of personal preference and what you want your signal to sound like. Most operators seem to like the make to be heavier than the break. All of the circuits shown here are capable of a wide range of adjustment.

If the negative supply in a grid-block keyed stage fails, the tube will draw excessive key-up current. To protect against tube damage in this eventuality, an overload relay can be used or, more simply, a fast-acting fuse can be included in the cathode circuit.

OSCILLATOR KEYING

One may wonder why oscillator keying hasn't been mentioned earlier, since it is widely used. A sad fact of life is that excellent oscillator keying is infinitely more difficult to obtain than is excellent amplifier keying. If the objective is no detectable chirp, it is probably *impossible* to obtain with oscillator keying, particularly on the higher frequencies. The reasons are simple. Any keyed-oscillator transmitter requires shaping at the oscillator, which involves changing the operating conditions of the oscillator over a significant period of time. The output of the oscillator doesn't rise to full value immediately so the drive on the following stage is changing, which in turn may reflect a variable load on the oscillator. No oscillator has been devised that has no change in frequency over its entire operating voltage range and with a changing load. Furthermore, the shaping of the keyed-oscillator envelope usually has to be exaggerated, because the following stages will tend to sharpen up the keying and introduce clicks unless they are operated as linear amplifiers.

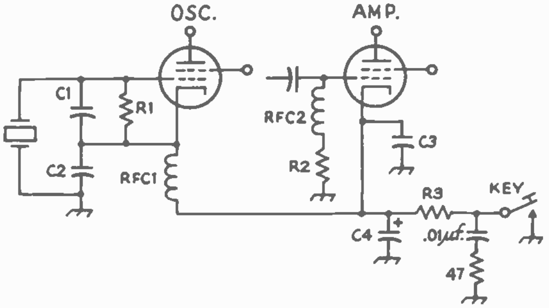


Fig. 11-6 — Simple differential-keying circuit for a crystal-controlled oscillator and power-amplifier transmitter.

Most simple crystal-controlled transmitters, commercial or home-built, return the oscillator grid-leak resistor, R1, to chassis, and "cathode keying" is used on the oscillator and amplifier stages. By returning the oscillator grid leak to the cathode, as shown here, negative power-supply-lead keying is used on the oscillator. A good crystal oscillator will operate with only 5 to 10 volts applied to it.

Using the above circuit, the signal is controlled by the shaping circuit, C4-R3. Increasing the value of R3 will make the signal "softer" on make; increasing the capacitance of C4 will make the signal softer on make and break. The oscillator will continue to operate after the amplifier has cut off, until the charge in C4 falls below the minimum operating voltage for the oscillator.

The .01-μF capacitor and 47-ohm resistor reduce the spark at the key contacts and minimize "key clicks" heard in the receiver and other nearby receivers. They *do not* control the key clicks associated with the signal miles away; these clicks are reduced by increasing the values of R3 and C4.

Since the oscillator may hold in between dots and dashes, a back wave may be present if the amplifier stage is not neutralized.

C1, C2 — Normal oscillator capacitors.
C3 — Amplifier rf cathode bypass capacitor.
C4 — Shaping capacitor, typically 1 to 10 μF, 250 volts, electrolytic
R1 — Oscillator grid leak; return to cathode instead of chassis ground.
R2 — Normal amplifier grid leak; no change.
R3 — Typically 47 to 100 ohms.
RFC1, RFC2 — As in transmitter, no change.

Break-in Keying

The usual argument for oscillator keying is that it permits break-in operation (see subsequent sections, also Chapter 23). If break-in operation is not contemplated and as near perfect keying as possible is the objective, then keying an amplifier or two by the methods outlined earlier is the solution. For operating convenience, an automatic transmitter "turner-onner" (see Campbell, *QST*, Aug., 1956), which will turn on the power supplies and switch antenna relays and receiver muting devices, can be used. The station switches over to the complete "transmit" condition where the first dot is sent, and it holds in for a length of time dependent upon the setting of the delay. It is equivalent to voice-operated phone of the type commonly used by ssb stations. It does not permit hearing the other station whenever the key is up, as does full break-in.

Full break-in with excellent keying is not easy to come by, but it is easier than many amateurs think. Many use oscillator keying and put up with a second-best signal.

Differential Keying

The principle behind "differential" keying is to turn the oscillator on fast before a keyed amplifier stage can pass any signal and turn off the oscillator fast after the keyed amplifier stage has cut off. A number of circuits have been devised for accomplishing the action. The simplest, which should be applied *only* to a transmitter using a voltage-stable (crystal-controlled) oscillator is shown in Fig. 11-6. Many "simple" and kitted Novice transmitters can be modified to use this system, which approaches the performance of the "turner-onner" mentioned above insofar as the transmitter performance is concerned. With separate transmitting and receiving antennas, the performance is comparable.

A simple differential-keying circuit that can be applied to any grid-block keyed amplifier or tube-keyed stage by the addition of a triode and a VR tube is shown in Fig. 11-7. Using this keying

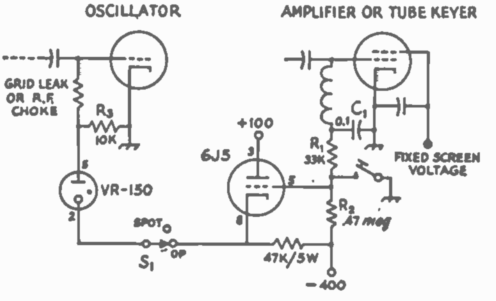


Fig. 11-7 — When satisfactory blocked-grid or tube keying of an amplifier stage has been obtained, this VR-tube break-in circuit can be applied to the transmitter to furnish differential keying. The constants shown here are suitable for blocked-grid keying of a 6146 amplifier; with a tube keyer the 6J5 and VR tube circuitry would be the same.

With the key up, sufficient current flows through R3 to give a voltage that will cut off the oscillator tube. When the key is closed, the cathode voltage of the 6J5 becomes close to ground potential, extinguishing the VR tube and permitting the oscillator to operate. Too much shunt capacity on the leads to the VR tube and too large a value of grid capacitance in the oscillator may slow down this action, and best performance will be obtained when the oscillator (turned on and off this way) sounds "clicky." The output envelope shaping is obtained in the amplifier, and it can be made softer by increasing the value of C1. If the keyed amplifier is a tetrode or pentode, the screen voltage should be obtained from a fixed voltage source or stiff voltage divider, not from the plate supply through a dropping resistor.

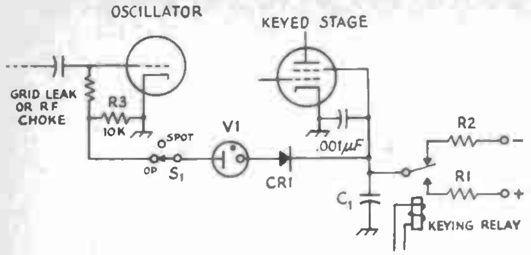


Fig. 11-8 – VR-tube differential keying in an amplifier screen circuit.

With key up and current flowing through V1 and CR1, the oscillator is cut off by the drop through R3. The keyed stage draws no current because its screen grid is negative. C1 is charged negatively to the value of the – source. When the relay is energized, C1 charges through R1 to a + value. Before reaching zero (on its way +) there is insufficient voltage to maintain ionization in V1, and the current is broken in R3, turning on the oscillator stage. As the screen voltage goes positive, the VR tube cannot reignite because the diode, CR1, will not conduct in that direction. The oscillator and keyed stage remain on as long as the relay is closed. When the relay opens, the voltage across C1 must be sufficiently negative for V1 to ionize before any bleeder current will pass through R3. By this time the screen of the keyed stage is so far negative that the tube has stopped conducting. (See Fig. 11-5 for suitable relay.)

system for break-in, the keying will be chirp-free if it is chirp-free with the VR tube removed from its socket to permit the oscillator to run all of the time. If the transmitter can't pass this test, it indicates that more isolation is required between keyed stage and oscillator.

Another VR-tube differential-keying circuit, useful when the screen-grid circuit of an amplifier is keyed, is shown in Fig. 11-8. The normal screen keying circuit is made up of the shaping capacitor C1, the keying relay (to remove dangerous voltages from the key), and the resistors R1 and R2. The + supply should be 50 to 100 volts higher than the normal screen voltage, and the – voltage should be sufficient to ignite the VR tube, V1, through the drop in R2 and R3. Current through R2 will be determined by the voltage required to cut off the oscillator; if 10 volts will do it the current will be 1 mA. For a desirable keying characteristic, R2 will usually have a higher value than R1. Increasing the value of C1 will soften both "make" and "break."

The tube used at V1 will depend upon the available negative supply voltage. If it is between 120 and 150, a 0A3/VR75 is recommended. Above this a 0C3/VR105 can be used. The diode, CR1, can be any unit operated within its ratings. A type 1N4005, for example, may be used with screen voltages under 600 and with far greater bleeder currents than are normally encountered – up to 1 ampere.

Clicks in Later Stages

It was mentioned earlier that key clicks can be generated in amplifier stages following the keyed stage or stages. This can be a puzzling problem to an operator who has spent considerable time adjusting the keying in his exciter unit for clickless keying, only to find that the clicks are bad when the amplifier unit is added. There are two possible causes for the clicks: low-frequency parasitic oscillations and amplifier "clipping."

Under some conditions an amplifier will be momentarily triggered into low-frequency parasitic oscillations, and clicks will be generated when the amplifier is driven by a keyed exciter. If these clicks are the result of low-frequency parasitic oscillations, they will be found in "groups" of clicks occurring at 50- to 150-kHz intervals either side of the transmitter frequency. Of course low-frequency parasitic oscillations can be generated in a keyed stage, and the operator should listen carefully to make sure that the output of the exciter is clean before he blames a later amplifier. Low-frequency parasitic oscillations are usually caused by poor choice in rf choke values, and the use of more inductance in the plate choke than in the grid choke for the same stage is recommended.

When the clicks introduced by the addition of an amplifier stage are found only near the transmitter frequency, amplifier "clipping" is indicated. It is quite common when fixed bias is used on the amplifier and the bias is well past the "cut-off" value. The effect can usually be minimized by using a combination of fixed and grid-leak bias for the amplifier stage. The fixed bias should be sufficient to hold the key-up plate current only to a low level and not to zero.

A linear amplifier (Class AB1, AB2 or B) will amplify the excitation without adding any clicks, and if clicks show up a low-frequency parasitic oscillation is probably the reason.

KEYING SPEEDS

In radio telegraphy the basic code element is the dot, or unit pulse. The time duration of a dot and a space is that of two unit pulses. A dash is three unit pulses long. The space between letters is three unit pulses; the space between words or groups is seven unit pulses. A speed of one baud is one pulse per second.

Assuming that a speed key is adjusted to give the proper dot, space and dash values mentioned above, the code speed can be found from

$$Speed\ (wpm) = \frac{dots/min.}{25} = 2.4 \times dots/sec.$$

E.g.: A properly adjusted electronic key gives a string of dots that count to 10 dots per second. Speed = 2.4 × 10 = 24 wpm.

Many modern electronic keyers use a clock or pulse-generator circuit which feeds a flip-flop dot generator. For these keyers the code speed may be determined directly from the clock frequency

$$Speed\ (wpm) = 1.2 \times clock\ frequency\ (Hz).$$

For a quick and simple means of determining the code speed, send a continuous string of dashes and count the number of dashes which occur in a 5-second period. This number, to a close approximation, is the code speed in words per minute.

BREAK-IN OPERATION

Smooth cw break-in operation involves protecting the receiver from permanent damage by the transmitter power and assuring that the receiver will "recover" fast enough to be sensitive between dots and dashes, or at least between letters and words.

Separate Antennas

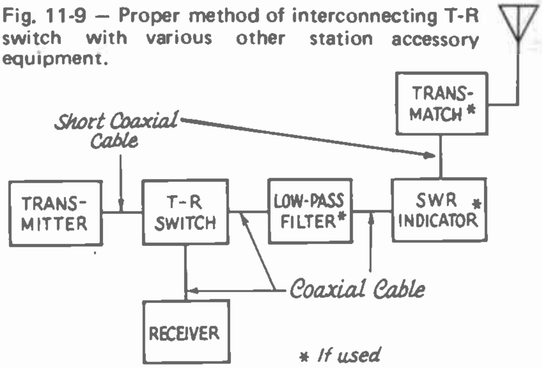
Few of the available antenna transfer relays are fast enough to follow keying, so the simplest break-in system is the use of a separate receiving antenna. If the transmitter power is low (25 or 50 watts) and the isolation between transmitting and receiving antennas is good, this method can be satisfactory. Best isolation is obtained by mounting the antennas as far apart as possible and at right angles to each other. Feed-line pickup should be minimized, through the use of coaxial cable or 300-ohm Twin-Lead. If the receiver recovers fast enough but the transmitter clicks are bothersome (they may be caused by the receiver overload and so exist only in the receiver) their effect on the operator can be minimized through the use of input and output limiters (see Chapter 8).

ELECTRONIC TRANSMIT-RECEIVE SWITCHES

When powers above 25 or 50 watts are used, where two antennas are not available, or when it is desired to use the same antenna for transmitting and receiving (a "must" when directional antennas are used), special treatment is required for quiet break-in operation on the transmitter frequency. A means must be provided for limiting the power that reaches the receiver input. This can be either a direct short-circuit, or may be a limiting device like an electronic switch used in the antenna feed line. The word "switch" is a misnomer in this case; the transmitter is connected directly to the antenna at all times. The receiver is connected to the antenna through the T-R switch, which functions to protect the receiver's input from transmitted power. In such a setup, all the operator need do is key the transmitter, and all the switching functions are taken care of by the T-R switch.

With the use of a T-R switch some steps should be taken to prevent receiver blocking. Turn off the agc or avc, decrease the rf gain setting, and advance the audio gain control. Use the rf gain control for obtaining the desired listening level. A little experimenting with the controls will provide the receiver settings best suited to individual operating preferences. A range of settings can usually be found, just on the threshold of receiver blocking, where comfortable levels of received signals are heard, and where, without adjusting the controls, the receiver can be used as a monitor during transmission. Usually no modification to the

Fig. 11-9 — Proper method of interconnecting T-R switch with various other station accessory equipment.



receiver is required, but if annoying clicks and thumps or excess volume occur at all settings of the receiver controls during transmission, their effect can be reduced with output audio limiting (see Chapter 8).

TVI and T-R Switches

T-R switches generate harmonics of the transmitted signal because of rectification of the energy reaching the input of the switch. These harmonics can cause TVI if steps are not taken to prevent it. Any T-R switch should be very well shielded, and should be connected with as short as possible a cable length to the transmitter. In addition, a low-pass filter may be required in the transmission line between the T-R switch and the antenna. Fig. 11-9 shows the proper method of interconnecting the various station accessory equipment.

Reduction of Receiver Gain During Transmission

For absolutely smooth break-in operation with no clicks or thumps, means must be provided for momentarily reducing the gain through the receiver. The system shown in Fig. 11-10 permits quiet break-in operation of high-powered stations. It may require a simple operation on the receiver, although many commercial receivers already provide the connection and require no internal modification. The circuit is for use with a T-R switch and a single antenna. R1 is the regular receiver rf and i-f gain control. The ground lead is run to chassis ground through R2. A wire from the junction runs to the keying relay, K1. When the key is up, the ground side of R1 is connected to ground through the relay arm, and the receiver is in its normal operating condition. When the key is closed the relay closes, which breaks the ground connection from R1 and applies additional bias to the tubes in the receiver: This bias is controlled by R2. When the relay closes, it also closes the circuit to the transmitter keying circuit. A simple rf filter at the key suppresses the local clicks caused by the

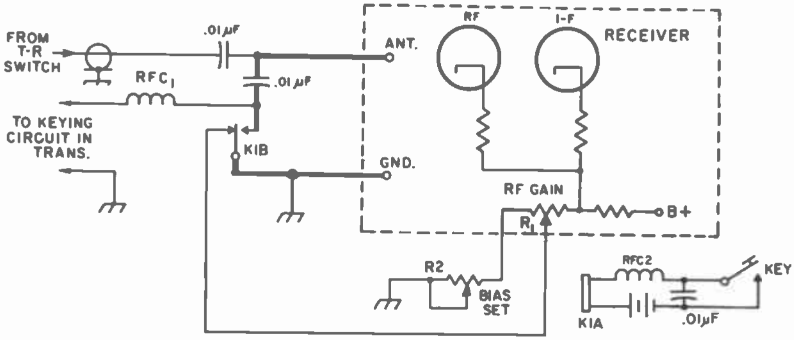


Fig. 11-10 – circuit for smooth break-in operation, using an electronic T-R switch. The leads shown as heavy lines should be kept as short as possible, to minimize direct transmitter pickup.

K1 – Spdt keying relay (Sigma 41FZ-10000-ACS-SIL or equiv.). Although battery and dc relay are shown, any suitable ac or dc relay and power source can be used.

R1 – Receiver manual gain control.

R2 – 5000- or 10,000-ohm wire-wound potentiometer.

RFC1, RFC2 – 1- to 2 1/2-mH rf choke, current rating adequate for application.

relay current. This circuit is superior to any working on the agc line of the receiver because the cathode circuit(s) have shorter time constants than the agc circuits and will recover faster. A similar circuit may be used in the emitters or source leads of transistorized receivers.

TESTING AND MONITORING OF KEYING

In general, there are two common methods for monitoring one's "fist" and signal. The first type involves the use of an audio oscillator that is keyed simultaneously with the transmitter.

The second method is one that permits receiving the signal through one's receiver, and this generally requires that the receiver be tuned to the transmitter (not always convenient unless working on the same frequency) and that some method be provided for preventing overloading of the receiver, so that a good replica of the transmitted signal will be received. Except where quite low power is used, this usually involves a relay for simultaneously shorting the receiver input terminals and reducing the receiver gain.

An alternative is to use an rf-powered audio oscillator. This follows the keying very closely (but tells nothing about the quality – chirps or clicks – of the signal).

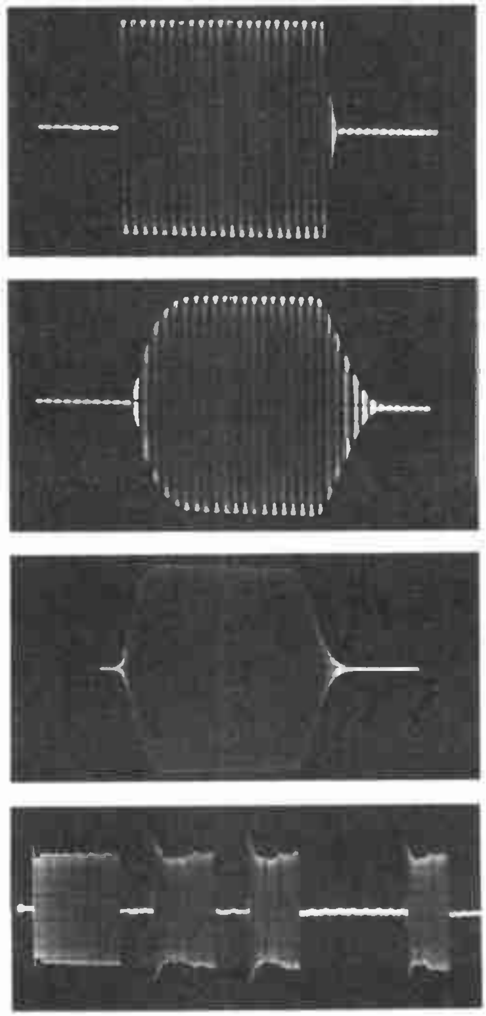
The easiest way to find out what your keyed signal sounds like on the air is to trade stations with a near-by ham friend some evening for a short QSO. If he is a half mile or so away, that's fine, but any distance where the signals are still S9 will be satisfactory.

After you have found out how to work his rig, make contact and then have him send slow dashes, with dash spacing (the letter "T" at about 5 wpm). With minimum selectivity, cut the rf gain back just enough to avoid receiver overloading (the condition where you get crisp signals instead of mushy ones) and tune slowly from out of beat-note range on one side of the signal through to zero and out the other side. Knowing the tempo of the dashes, you can readily identify any clicks in the vicinity as yours or someone else's. A good signal will have a thump on "make" that is perceptible only where you can also hear the beat note, and the click on "break" should be practically negligible at any point. If your signal is like that, it will sound good, provided there are no chirps. Then have your friend run off a string of fast dots with the bug – if they are easy to copy, your signal has no "tails" worth worrying about and is a good one for any speed up to the limit of manual keying. Make one check with the selectivity in, to see that the clicks off the signal frequency are negligible even at high signal level.

If you don't have any friends with whom to trade stations, you can still check your keying, although you have to be a little more careful. The transmitter output should be fed into a shielded dummy load. Ordinary incandescent lamps are unsatisfactory as lamp resistance varies too much with current. The thermal lag may cause the results to be misleading.

The first step is to get rid of the rf click at the key. This requires an rf filter (mentioned earlier). With no clicks from a spark at the key, disconnect the antenna from your receiver and short the antenna terminals with a short piece of wire. Tune in your own signal and reduce the rf gain to the point where your receiver doesn't overload. Detune any antenna trimmer the receiver may have. If you can't avoid overload with the rf gain-control range, pull out the rf amplifier tube and try again. If you still can't avoid overload, listen to the second harmonic as a last resort. An overloaded receiver can generate clicks.

Describing the volume level at which you should set your receiver for these "shack" tests is a little difficult. The rf filter should be effective with



(A)

(B)

(C)

(D)

These photos show cw signals as observed on an oscilloscope. At A is a dot generated at a 46-baud rate with no intentional shaping, while at B the shaping circuits have been adjusted for approximately 5-ms rise and decay times. Vertical lines are from a 1-kHz signal applied to the Z or intensity axis for timing. Shown at C is a shaped signal with the intensity modulation of the pattern removed. For each of these photos, sampled rf from the transmitter was fed directly to the deflection plates of the oscilloscope.

At D may be seen a received signal having essentially no shaping. The spike at the leading edge is typical of poor power-supply regulation, as is also the immediately following dip and rise in amplitude. The clicks were quite pronounced. This pattern is typical of many observed signals, although not by any means a worst case. The signal was taken from the receiver's i-f amplifier (before detection) using a hand-operated sweep circuit to reduce the sweep time to the order of one second. (Photos from *QST* for October and November 1966.)

the receiver running wide open and with an antenna connected. When you turn on the transmitter and take the steps mentioned to reduce the signal in the receiver, run the audio up and the rf down to the point where you can just hear a little "rushing" sound with the BFO off and the receiver tuned to the signal. This is with the selectivity in. At this level, a properly adjusted keying circuit will show no clicks off the rushing-sound range. With the BFO on and the same gain setting, there should be no clicks outside the beat-note range. When observing clicks, make the slow-dash and dot tests outlined previously.

Now you know how your signal sounds on the air, with one possible exception. If keying your transmitter makes the lights blink, you may not be able to tell too accurately about the chirp on your signal. However, if you are satisfied with the absence of chirp when tuning *either side of zero beat*, it is safe to assume that your receiver isn't chirping with the light flicker and that the observed signal is a true representation. No chirp either side of zero beat is fine. Don't try to make these tests without first getting rid of the rf click at the key, because clicks can mask a chirp.

The least satisfactory way to check your keying is to ask another ham on the air how your keying sounds. It is the least satisfactory because most hams are reluctant to be highly critical of another amateur's signal. In a great many cases they don't actually know what to look for or how to describe any aberrations they may observe.

A MEMORY FOR THE DELUXE KEYER

The system described below permits storage of up to 200 letters of text organized in one, two, three, or four messages. A digital display provides an indication of the message being sent or loaded (No. 1, 2, 3, or 4) and the message bit being addressed (0 to 512). Any number of pauses may be programmed into a message to allow manual insertion of changeable text (such as RST or

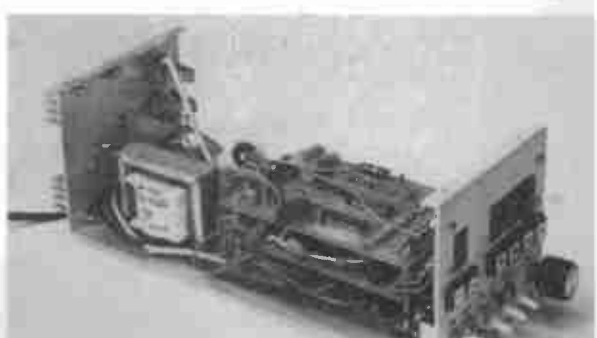


Fig. 1 — A look at the inside of the Accu-Memory. The power supply components may be seen at the left, and the three "stacked" circuit boards to their right. The fourth circuit board, containing the readout, is mounted behind the sloping portion of the front panel. The board at the bottom of the "stack" is that of the original Accu-Keyer.

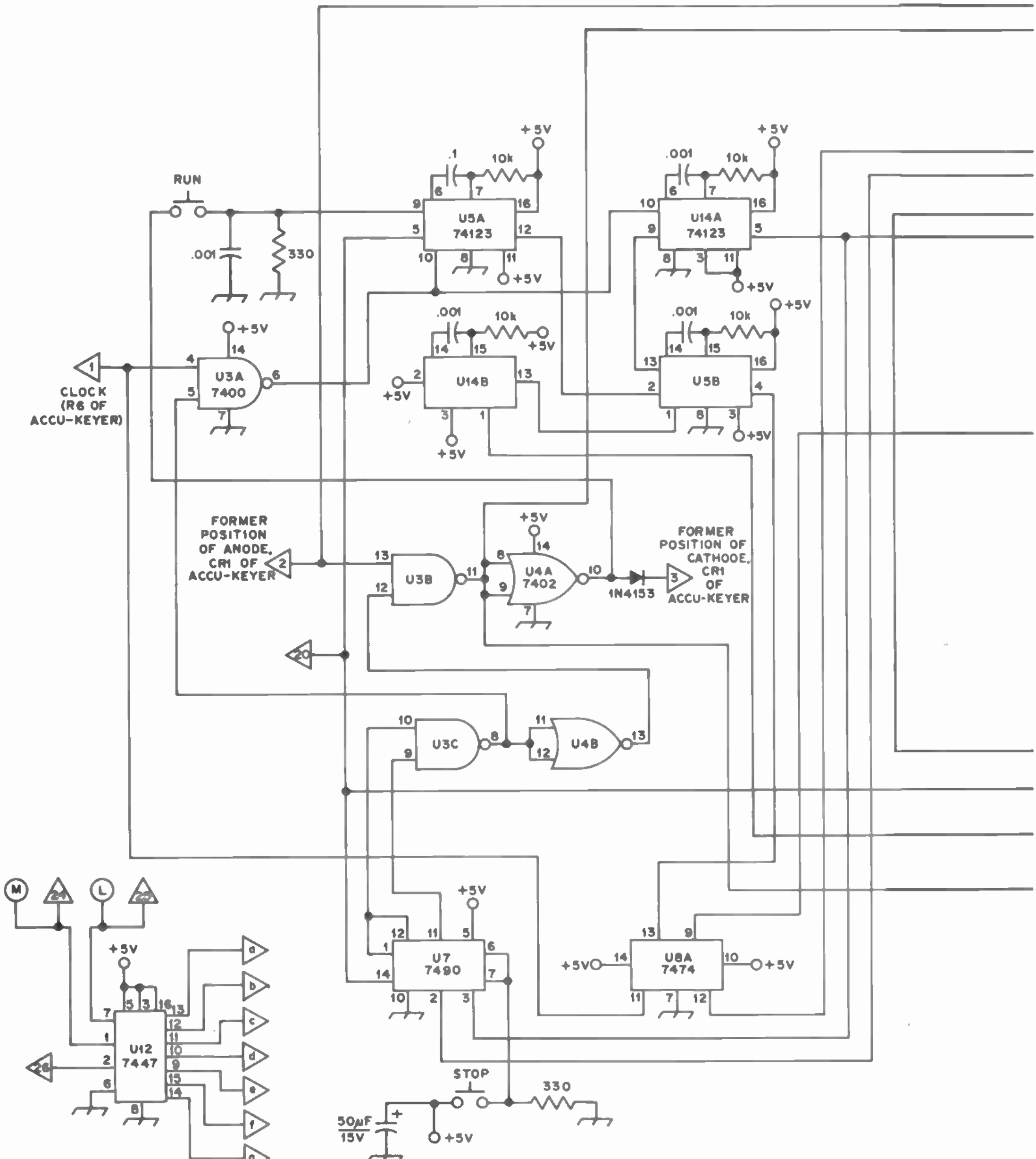
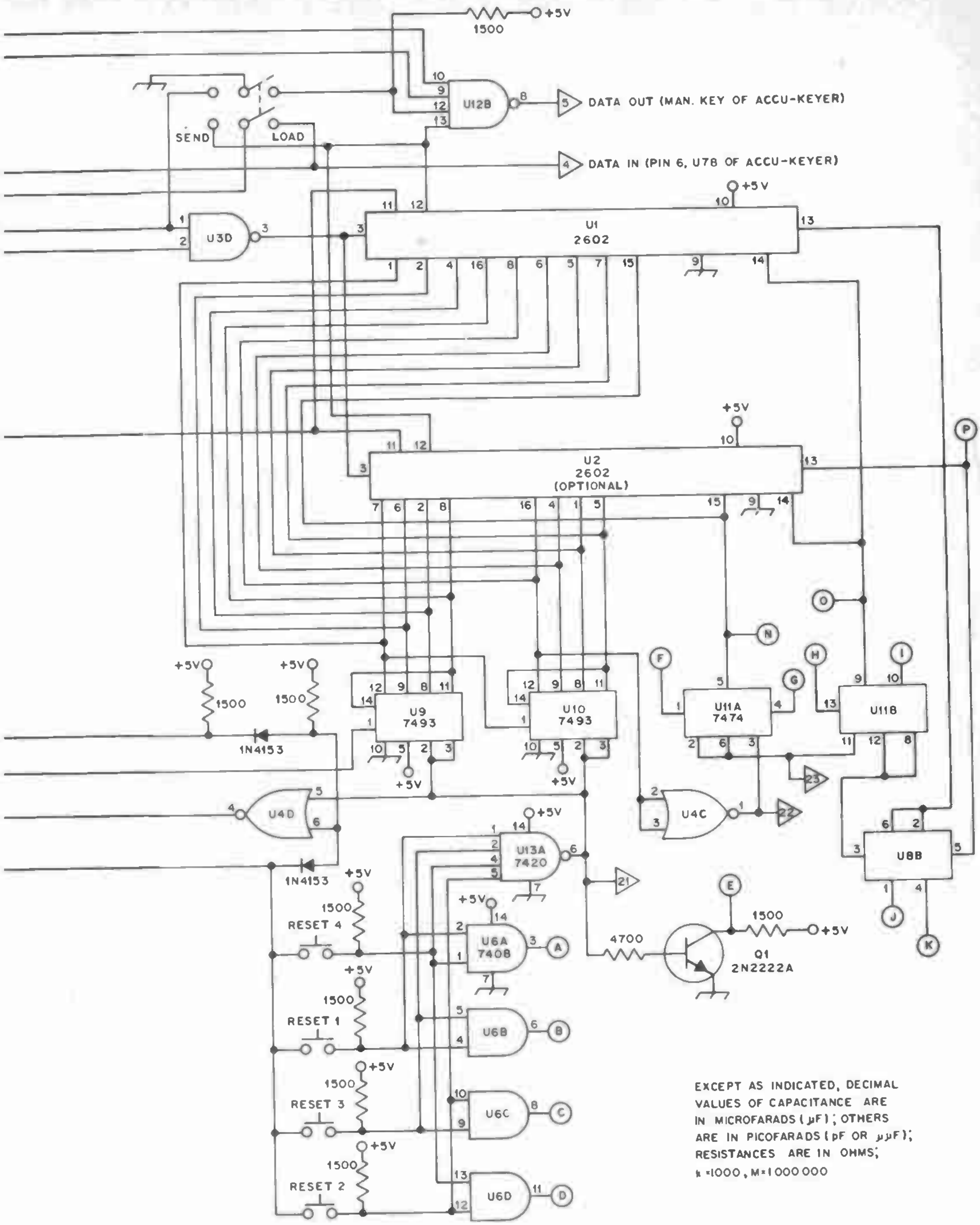


Fig. 2 — Diagram of memory circuitry of the Accu-Memory. See Table II for list of parts. Numbers and letters in triangles identify inter-connections to other parts of the Accu-Memory, as listed in Table I. Letters in circles indicate terminals for jumpers to be wired for either one or two RAM ICs. This wiring information is also listed in Table I.

contest serial number). After manual insertion a touch of the RUN button allows the remainder of the programmed message to continue. The message being sent may be aborted by pressing the STOP button (the "I didn't mean to press the button!" button). Unlike some programmable keyers, the use of a free-running (asynchronous) clock in the load mode has been avoided, greatly simplifying the loading process. All features of the original Accu-Keyer have been retained. The dot and dash memories of the Accu-Keyer and its automatic character-space feature are used to good advantage in the Accu-Memory.

In addition to the Accu-Keyer board, three printed circuit boards make up the Accu-Memory: a memory board, a display board, and a display-



driver board. The power supply provides 5 volts at 0.8 ampere to power all the circuitry.

The Accu-Memory has been "battle tested" in contests and has been found to be very effective in reducing operator fatigue. It is of use whenever there is a requirement for repeatedly sending the same cw sequences such as in contests, DX pileups, and net-control operations. Experience has shown that the digital displays are far more useful than originally anticipated.

Construction

As shown in Fig. 1, the Accu-Memory is constructed in an aluminum box made by cutting and bending sheet aluminum. The front-panel dimensions are deliberately made small because

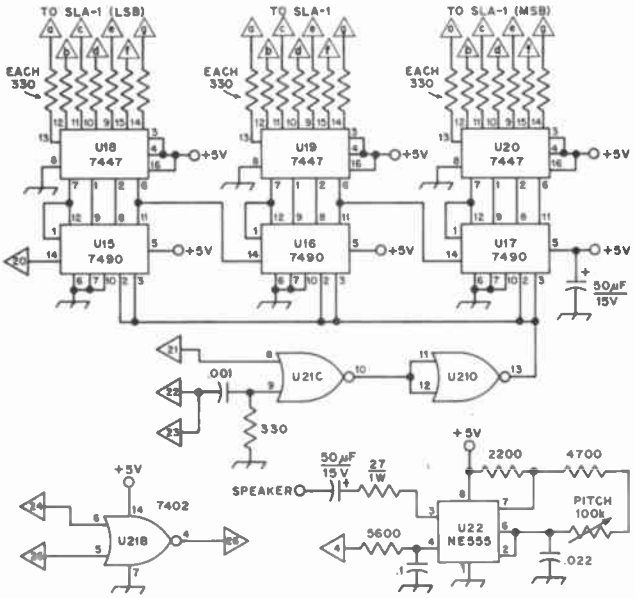


Fig. 3 – Diagram of driver and display. See Table II for list of parts. Numbers inside triangles identify interconnections to other parts of the Accu-Memory, as listed in Table I.

depth in most ham shacks is more abundant than frontal-area space. This method also gives a neat, streamlined appearance. The overall outside dimensions are 4-¼ × 3-½ × 10-½ with the length dimension measured across the bottom plate, less knobs and heat sink. The heat sink for the LM309 is attached to the rear panel,[1] along with the key jack, the output jack, and a fuse holder (Safety First!). Power supply components are located on the bottom plate near the rear. Two terminal strips are used to mount the power supply diodes and filter capacitor. All the other electronic parts are mounted on four printed-circuit cards.

The push buttons are sold by Solid State Systems (see footnote [1]). One word of caution: do not increase the value of the filter capacitor in the power supply. It has been chosen for minimum dissipation by the LM309 regulator.

Fig. 2 is a schematic diagram of the circuitry on the driver and display board. Wires that interconnect the boards are shown as numbers or lower case letters in triangles on the figures. Selectable jumpers allow the use of one or two RAM IC's.

[1] **The heat sink for the LM-309 is available from Solid State Systems, Inc., Box 773, Columbia, MO 65201.**

The jumper points are shown as capital letters in circles. Table I is a list of interconnecting wires. Table II gives a parts list for each board. Fig. 4 is a diagram of the power supply.

To send, place the LOAD/SEND switch in the SEND position and press the proper message button. The STOP button will halt sending, but the message can be continued from the halted point if the RUN button is depressed.

If it is desired to use the insert feature, load the first part of the message as described above. Then after the memory stops advancing, press the RUN button once, wait until the count stops, and then load the second half of the message. In the SEND mode the memory will send the first part, stop and allow insertion of manual input such as signal reports, and then, when the RUN button is depressed, continue with the second half. This procedure may be repeated as many times as necessary.

The readout indicates the message number and the location within the message starting at 000 and continuing through either 256 or 512, depending on whether one or two memories are installed. A decimal point lights when the keyer is sending either manually or automatically.

Helpful Advice

After a lot of correspondence with amateurs who built the Accu-Keyer, it is apparent that some do not know that there is a difference between a 7400, a 74H00, a 74L00, and a 74C00. These are all members of a family of quad two-input gates that are different internally and are not interchangeable (in almost all cases) with each other. Some IC distributors tend to be haphazard about which type they send.

As with the Accu-Keyer, ready-made boards are available for the memory through Garrett.[2] A business-size self-addressed stamped envelope is mandatory to reduce addressing time to a minimum. If any problems develop or changes occur in the circuit, a data sheet showing corrections will be included with the boards.

[2] **As a service to those who wish to avail themselves, ready-made circuit boards may be obtained through James Garrett, WB4VVF, 126 W. Buchanon Ave., Orlando, FL 32809. All boards are glass epoxy and drilled. At the time of this printing, the Accu-Keyer board is $3.50. The memory, readout, and readout-driver boards are $12 as a set. The memory board, if ordered alone, is $6.**

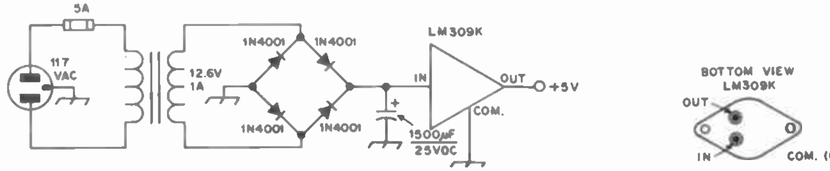


Fig. 4 -- Power supply for Accu-Memory and

TABLE I – INTERCONNECTIONS

WIRE NUMBER	FUNCTION	
Keyer-to-memory interconnections.		
1	Clock – connect to R6	
2	Anode CR1	
3	Cathode CR1 (Remove CR1 in keyer and connect as shown.)	
4	Data in (Connect to U7B in keyer and tone oscillator on driver board.)	
5	Data out (Connect to manual key input, U7 pin 5 in keyer.)	
Memory-to-control switches		
6	Send 1	Dpdt switch
7	Send 2	
8	Load 1	
9	Load 2	
10	Common 1	
11	Common 2	
Memory to readout		
12	Insert	Push buttons
13	Insert return	
14	Reset 1	
15	Reset 2	
16	Reset 3	
17	Reset 4	

WIRE NUMBER	FUNCTION
18	Reset common
19	Stop
a, b, c, d, e, f, g	Quadrant readout
Memory to driver	
20	Readout count
21	Readout reset
22	Readout quadrant reset (use with one 2601)
23	Readout quadrant reset (use with two 2602s)
24	NOR 1
25	NOR 2
26	NOR out
Driver to readout	
27-33 (a-g, LSB)	Least-significant digit
34-40 (a-g, CSB)	Center-significant digit
41-47 (a-g, MSB)	Most-significant digit
48-49	Pitch control (short if no control desired)
50	Speaker

Memory interconnections

For one memory IC connect: A to H, B to G, C to I, D to F, J to ground, K to + 5 V, L to N, and M to O.

For two memory ICs connect: A to J, B to I, C to K, D to H, E to F, G to + 5 V, L to O. and M to P.

Connect DP (decimal point) on readout board to wire 13.

TABLE II – Accu-Memory Parts List

Memory Board

2	7474 ICs	U8, U11
2	7493 ICs	U9, U10
1	7408 IC	U6
2	74123 ICs	U5, U14
1	7400 IC	U3
1	7490 IC	U7
1	7402 IC	U4
1	7420 IC	U13
1	7447 IC	U12 (optional)
2	2102 or 2602 ICs	U1, U2 (U2 optional)
1	2N2222A transistor	
3	1N4148 silicon diodes or equivalent	
8	1500-Ω ¼-W resistors	
4	10-kΩ ¼-W resistors	
2	330-Ω resistors	
1	4700-Ω resistor	
4	.001-μF disk ceramic capacitors	
4	.1-μF disk ceramic capacitors	
1	50-μF 15-V electrolytic capacitor	

Driver Board

3	7490 ICs
3	7447 ICs
1	7402 IC
1	NE555 IC
22	330-Ω resistors
1	5600-Ω resistors
1	2200-Ω resistor
1	4700-Ω resistor (33kΩ with no pitch control)
1	27-Ω resistor
1	.001-μF disk ceramic capacitor
1	.022-μF disk ceramic capacitor
2	.1 μF disk ceramic capacitors
2	50-μF 15-V electrolytic capacitors

Readout Board

4	SL A-1 readouts
6	Push buttons (see text)
8	330-Ω resistors

DELUXE ALL-SOLID-STATE KEYER

The Accu-Keyer is a modern keying device with deluxe features available on only the most expensive of commercially available instruments, but it may be built for less than $25.

The basic circuit uses seven TTL integrated circuits which may be purchased at "bargain" suppliers for less than $3. Optional features which may be incorporated at the builder's discretion are a stiffly regulated power supply, a keying monitor, and provisions for solid-state keying of cathode-keyed transmitters.

The Accu-Keyer was designed with these features in mind:

1) Self-completing dots and dashes
2) Dot and dash memories
3) Iambic operation
4) Dot and dash insertion
5) Automatic character space (with switching provided to defeat this feature)
6) 5-50 wpm speed range
7) Low cost

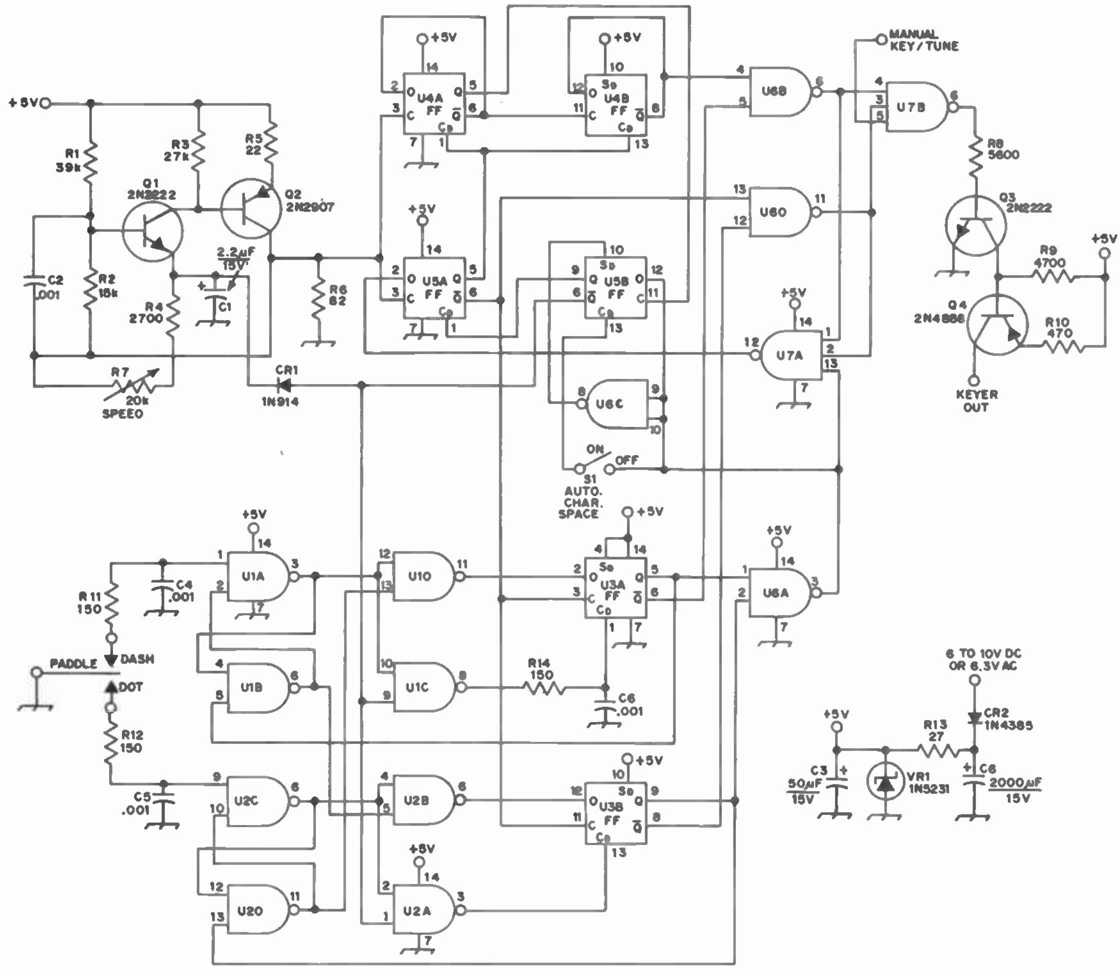
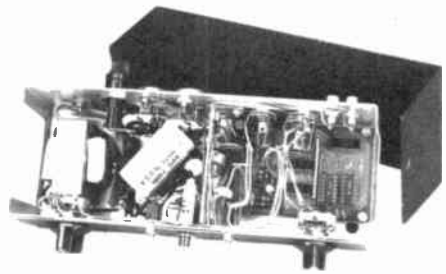


Fig. 2 — Schematic diagram of the Accu-Keyer. Resistances are in ohms; k = 1000. All capacitances are in microfarads. All resistors may be 1/4 watt except R13, which should have a 2-W rating. Capacitors with polarity indicated are electrolytic; all others are disk ceramic. Parts not listed below are for text reference and circuit-board identification.

CR1 — Small-signal silicon diode.
CR2 — Rectifier diode, 1/2 A or greater.
Q1, Q3 — Silicon npn, 250-mW, high-speed switching or rf-amplifier transistor.
Q2 — Silicon pnp, 250-mW, high-speed switching or rf-amplifier transistor.
Q4 — Silicon pnp, 250-mW, high-voltage af-amplifier transistor.
R7 — Reverse-log-taper control; Mallory U-28 suitable.
S1 — Spst toggle.
U1, U2, U6 — Quad 2-input NAND gate, type 7400.*
U3, U4, U5 — Dual type *D* flip-flop, type 7474.*
U7 — Triple 3-input NAND gate, type 7410.*
VR1 — 5.1-V, 0.5-W Zener diode.

* All ICs are dual-in-line package, 14 pin. Note: All ICs are available from various manufactuerers or as surplus. Motorola part numbers are prefixed by *MC* and suffixed by *P*. Texas Instruments parts have an *SN* prefix and *N* suffix. Signetics ICs have an *N* prefix and an *A* suffix. For example, Motorola's MC7400P is equivalent to Texas Instruments' SN7400N or Signetics' N7400A.



A peek inside the Accu-Keyer shows compact construction in this deluxe version built by W1RML. The ac-operated power supply components are located at the left, and the basic keyer board at the right. The keying monitor is constructed on a separate vertically mounted circuit board positioned near the center of the enclosure. The pitch control is mounted inside the keyer on this circuit board, as it is not adjusted frequently. The speaker is mounted over a "grille" formed by drilling many holes at the bottom of the enclosure, and is nearly hidden by the filter capacitor in this view. On the rear panel, in TO-3 style cases, are the 5-volt regulator IC and the cathode keying transistor.

The Circuit

The schematic diagram of the Accu-Keyer is shown in Fig. 2. The voltage applied to CR2 for powering the keyer may be either 8 to 10 volts dc or 6.3 volts ac, such as from the filament supply of a transmitter or receiver. If dc is applied, C6 is not required. If ac is applied to CR2, VR1 functions more to protect the ICs from overvoltage by limiting the amplitude of the ripple than it does for voltage regulation. If a well-filtered and regulated supply is desired, the circuit of Fig. 3A may be used in place of CR2, R13, and VR1 and associated capacitors. Constructed with the components shown, that supply will handle the keyer requirements with power to spare.

Should a keying monitor be desired, the diagram of Fig. 3B may be used to construct a circuit which will afford plenty of volume and a stable, pleasing tone. The circuit is a modified version of the code-practice oscillator appearing in Chapter 1. Equipped with such a monitor, the Accu-Keyer becomes ideal for conducting code practice sessions for small and medium-sized groups.

Fig. 3C shows a circuit which may be used for cathode-keyed or solid-state "QRP" transmitters. The Delco keying transistor will safely handle two amperes of current and a collector-to-emitter potential of 800 V, and yet its cost is less than that of a new mercury-wetted relay. The use of a transistor offers advantages over both vacuum-tube keying and relay keying of cathode-keyed rigs; the voltage drop across the transistor when saturated introduces negligible grid-cathode bias to the keyed stage, and the keying is softened somewhat over relay keying because the transistor cannot go from cutoff to saturation (or vice versa) instantaneously. For QRP transmitters, Q6 may be a 300- or 500-mW silicon npn transistor, such as a 2N2222 or 2N4123.

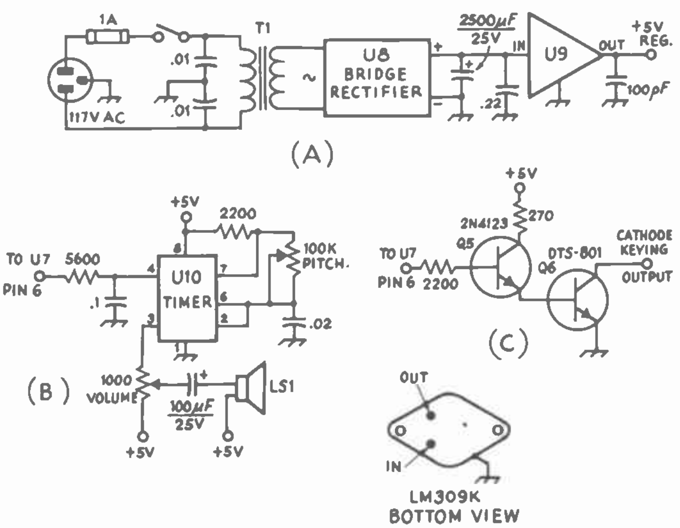


Fig. 3 — At A, optional ac-operated power supply circuit for the Accu-Keyer; At B, an optional monitor, and at C a circuit for cathode keying.

LS1 — Miniature speaker, 4-, 8- or 16-ohm impedance.

Q6 — High-voltage high-current silicon npn power transistor (Delco DTS-801, -802, or -804 or equiv.).

T1 — Surplus filament transformer, 12.6-V 1-A secondary rating.

U8 — Full-wave rectifier bridge, 1-A 50-V (Motorola 920-2, HEP 175, or equiv.). Four rectifier diodes in a bridge arrangement may be used instead.

U9 — Voltage-regulator IC, 5-volt (National Semiconductor LM309K or equiv.).

U10 — Signetics NE555 timer IC.

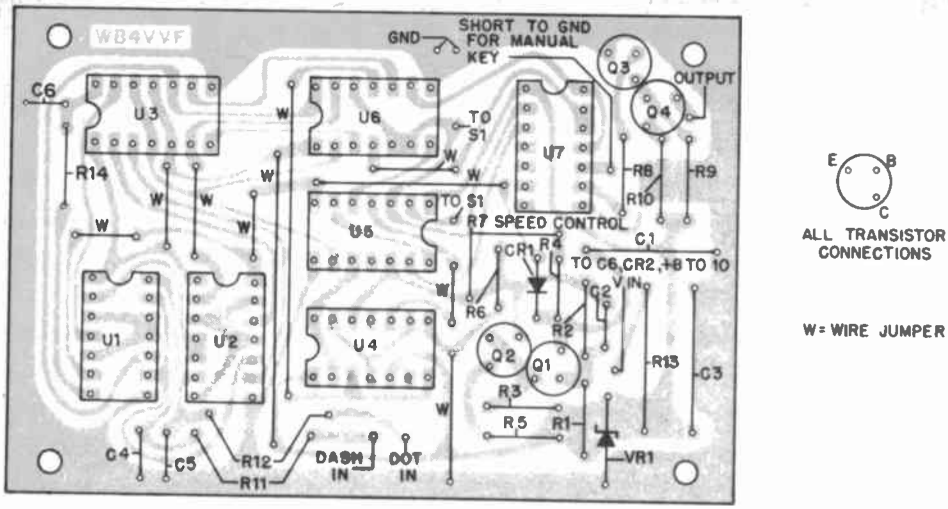


Fig. 4 — Etching pattern and parts-layout diagram for the Accu-Keyer. Pattern is actual size, shown from *foil side* of board.

Construction and Operation

A ready-made circuit board is available for the basic circuit of the Accu-Keyer.[1] Fig. 4 is an actual-size board layout and parts-placement guide. If the builder elects to use none of the optional circuit features of Fig. 3, the complete keyer may be built into a 3 × 2 × 5-inch Minibox. The board pattern in Fig. 4 contains all parts of Fig. 2 except the controls, the filter capacitor, and the rectifier in the power supply.

It is essential that all leads to the keyer be shielded from rf. RG-174/U coax may be used. A .01-μF bypass capacitor is provided on the power input to remove rf. As shown on the diagram, the inputs from the paddle are filtered by 150-ohm resistors bypassed by .001-μF capacitors. In stubborn cases it may be necessary to bypass the paddle contacts at the paddle itself.

Substitution of transistors for Q1 and Q2 may require changing the value of R5 to make the first clock pulse the same length as the rest. Both should be transistors with a beta of at least 60. Q3 is noncritical, and any good silicon transistor should work. Q4 should be capable of withstanding the transmistter key-up voltage. Any pnp silicon device having a reasonable beta and meeting this requirement should work. The value of C1 may be juggled to change the range of the speed control. The value specified gives a range of approximately 5 to 50 wpm.

[1]A glass-epoxy board, pre-drilled, is available for $3.50 from James M. Garrett, WB4VVF, 126 W. Buchanon, Orlando, FL 32809.

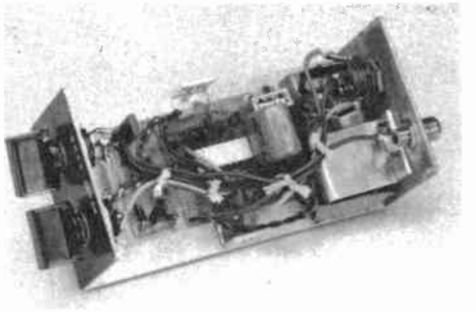
A SINGLE IC KEYER

The 8043 Integrated Electronic Keyer IC is a space-age component, designed specifically for the cw operator. A product of the latest design and processing technology available in the integrated circuit industry, the 8043 represents the same advancements which make possible the one chip electronic calculator and the digital electronic wrist watch. The 8043 is available from Curtis Electro Devices, Mountain View, California.

A good keyer exhibits no idiosyncrasies. The 8043 incorporates filters which eliminate the effects of key bounce on both make and break. Another intangible quality is rf immunity. To protect false triggering by rf on the paddle leads, the 8043 dot and dash inputs are equipped with active pull-up resistors (a system which reflects a few hundred ohms during quiescence and zero ohms during key down) which exhibit only a few hundred ohms impedance to the power supply when the key is open. In order to assure that the 8043 dissipates as little power as possible, CMOS (complementary metal-oxide semiconductor) circuit techniques are used. As a result, the quiescent current required is only 50 microamperes at 0.5 volts. This makes an on-off switch unnecessary even when using a battery supply. The key-down current is about 30 mA with about 99.9% of this current being required for sidetone output and drive for the output transistor.

Once a dot, dash or space is commenced, there is no way to prevent it from transmitting at the exact standard length. It may be neither cut short nor extended by improper key action. When the dot paddle is depressed, a continuous string of dots is produced; when the dash paddle is depressed, continuous dashes are produced. When both paddles are closed, an alternate (iambic) series of dots and dashes is made. The series can be started with either a dot or a dash, depending on which key side is closed first. Iambic operation allows squeeze keying if a twin-lever paddle is used.

The self-completing function of electronic keyers can cause dots to get lost because the operator, attempting to initiate a dot before the last character has been completed, tends to lead the keyer. Since dashes are naturally held longer, they seldom get lost. To prevent lost dots, the 8043 employs a memory to remember when a dot is called for and to insert it at the proper time. The dot memory also helps in squeeze keying, where a tap on the dot paddle will insert a dot into a series of dashes. Although a dot-space ratio of 1-to-1 is correct timing, some operators like heavier keying. The 8043 has provision for a weight control if desired.



Inside view of the one IC keyer.

The 8043 has a built-in sidetone generator with pitch adjustable to your preference. This sidetone also functions when a straight key is used or the tune switch is closed. The keying output voltage from the 8043 is low for a key-up condition; key down is represented by high output of sufficient level to drive an npn keying transistor. The speed range is normally about 8 to 50 words per minute, but by selecting timing components, you can get almost any speed range you like.

The 8043 is contained in a 1-3/4 X 2-1/4 X 4-3/4-inch metal box. There is no on-off switch because of the low quiescent current drain. Jacks for the paddle and output keying are located on the rear of the enclosure and are not visible in the photograph.

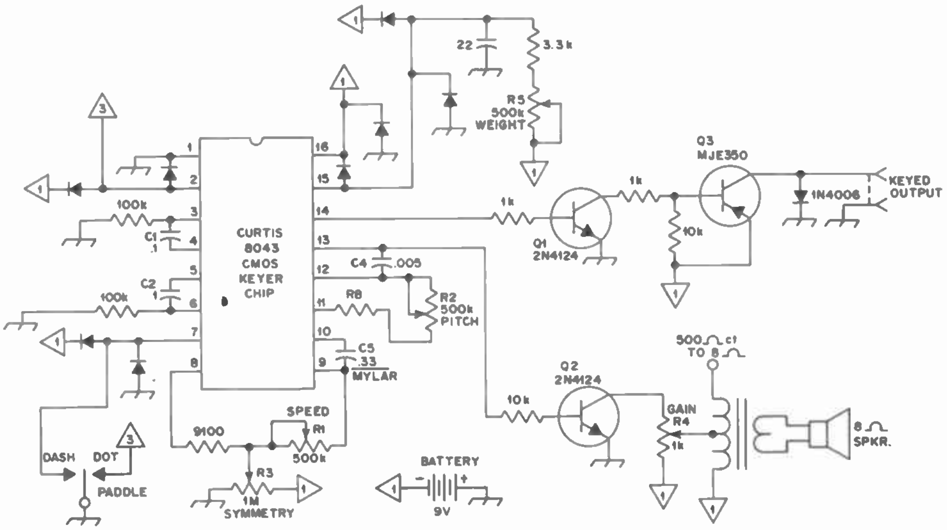
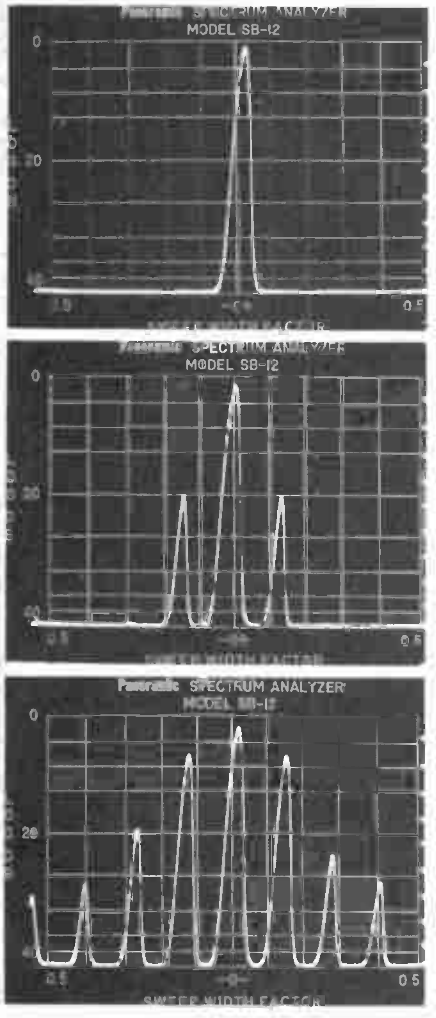


Fig. 1 — Circuit diagram for the keyer. All diodes are 1N4006 or equiv. All potentiometers are linear taper and resistors are 1/2 or 1/4 watt.

Chapter 12

Amplitude Modulation and Double-Sideband Phone



(A)

(B)

(C)

Fig. 12-1 — Spectrum-analyzer display of the rf output of an a-m transmitter. Frequency is presented on the horizontal axis (7-kHz total display width) versus relative amplitude of the signal component on the vertical axis. Shown at A is the unmodulated carrier, which occupies but a single frequency. At B the carrier is 20-percent modulated with a 1000-Hz tone. Each sideband may be seen to be at a level approximately 20 dB below the carrier. The signal bandwidth in this case is twice the modulating frequency, or 2 kHz. Shown at C is the widened channel bandwidth resulting from splatter caused by overmodulation. New frequencies, audio harmonics of the 1000-Hz modulating tone, extend for several kilohertz either side of the carrier.

As described in the chapter on circuit fundamentals, the process of modulation sets up groups of frequencies called sidebands, which appear symmetrically above and below the frequency of the unmodulated signal or carrier. If the instantaneous values of the amplitudes of all these separate frequencies are added together, the result is called the modulation envelope. In amplitude modulation (a-m) the modulation envelope follows the amplitude variations of the signal that is used to modulate the wave.

For example, modulation by a 1000-Hz tone will result in a modulation envelope that varies in amplitude at a 1000-Hz rate. The actual rf signal that produces such an envelope consists of three frequencies – the carrier, a side frequency 1000 Hz higher, and a side frequency 1000 Hz lower than the carrier. See Fig. 12-1. These three frequencies easily can be separated by a receiver having high selectivity. In order to reproduce the original modulation the receiver must have enough bandwidth to accept the carrier and the sidebands simultaneously. This is because an a-m detector responds to the modulation envelope rather than to the individual signal components, and the envelope will be distorted in the receiver unless all the frequency components in the signal go through without change in their amplitudes.

In the simple case of tone modulation the two side frequencies and the carrier are constant in amplitude – it is only the envelope amplitude that varies at the modulation rate. With more complex modulation such as voice or music the amplitudes and frequencies of the side frequencies vary from instant to instant. The amplitude of the modulation envelope varies from instant to instant in the same way as the complex audio-frequency signal causing the modulation. Even in this case the *carrier* amplitude is constant if the transmitter is properly modulated.

A-M Sidebands and Channel Width

Speech can be electrically reproduced, with high intelligibility, in a band of frequencies lying between approximately 100 and 3000 Hz. When these frequencies are combined with a radio-frequency carrier, the sidebands occupy the frequency spectrum from about 3000 Hz below the carrier frequency to 3000 Hz above – a total band or channel of about 6 kHz.

Actual speech frequencies extend up to 10,000 Hz or more, so it is possible to occupy a 20-kHz channel if no provision is made for reducing its width. For communication purposes such a channel width represents a waste of valuable spectrum space, since a 6-kHz channel is fully adequate for intelligibility. Occupying more than the minimum channel creates unnecessary interference.

THE MODULATION ENVELOPE

In Fig. 12-2 the drawing at A shows the unmodulated rf signal, assumed to be a sine wave of the desired radio frequency. The graph can be taken to represent either voltage or current.

In B, the signal is assumed to be modulated by the audio frequency shown in the small drawing above. This frequency is much lower than the carrier frequency, a necessary condition for good modulation. When the modulating voltage is "positive," (above its axis) the envelope amplitude is increased *above* its unmodulated amplitude; when the modulating voltage is "negative," the envelope amplitude is *decreased*. Thus the envelope grows larger and smaller with the polarity and amplitude of the modulating voltage.

The drawing at C shows what happens with stronger modulation. The envelope amplitude is doubled at the instant the modulating voltage reaches its positive peak. On the negative peak of the modulating voltage the envelope amplitude just reaches zero.

Percentage of Modulation

When a modulated signal is detected in a receiver, the detector output follows the modulation envelope. The stronger the modulation, therefore, the greater is the useful receiver output. Obviously, it is desirable to make the modulation as strong or "heavy" as possible. A wave modulated as in Fig. 12-2C would produce more useful audio output than the one shown at B.

The "depth" of the modulation is expressed as a percentage of the unmodulated carrier amplitude. In either B or C, Fig. 12-2, X represents the unmodulated carrier amplitude, Y is the maximum envelope amplitude on the modulation uppeak, and Z is the minimum envelope amplitude on the modulation downpeak.

In a properly operating modulation system the modulation envelope is an accurate reproduction of the modulating wave, as can be seen in Fig. 12-2 at B and C by comparing one side of the outline with the shape of the modulating wave. (The lower outline duplicates the upper, but simply appears upside down in the drawing.)

The percentage of modulation is

$$\%\ \text{Mod.} = \frac{Y - X}{X} \times 100 \text{ (upward modulation), or}$$

$$\%\ \text{Mod.} = \frac{X - Z}{X} \times 100 \text{ (downward modulation)}$$

If the two percentages differ, the larger of the two is customarily specified. If the wave shape of the modulation is such that its peak positive and negative amplitudes are equal, then the modulation percentage will be the same both up and down, and is

$$\%\ \text{Mod.} = \frac{Y - Z}{Y + Z} \times 100.$$

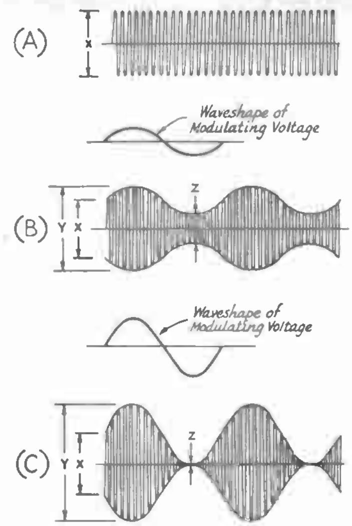


Fig. 12-2 — Graphical representation of (A) rf output unmodulated, (B) modulated 50 percent, (C) modulated 100 percent. The modulation envelope is shown by the thin outline on the modulated wave.

Power in Modulated Wave

The amplitude values shown in Fig. 12-2 correspond to current and voltage, so the drawings may be taken to represent instantaneous values of either. The power in the wave varies as the *square* of either the current or voltage, so at the peak of the modulation upswing the instantaneous power in the envelope of Fig. 12-2C is four times the unmodulated carrier power (because the current and voltage both are doubled). At the peak of the downswing the power is zero, since the amplitude is zero. These statements are true of 100-percent modulation no matter what the wave form of the modulation. The instantaneous envelope power in the modulated signal is proportional to the square of its envelope amplitude at every instant. This fact is highly important in the operation of every method of amplitude modulation.

It is convenient, and customary, to describe the operation of modulation systems in terms of sine-wave modulation. Although this wave shape is seldom actually used in practice (voice wave shapes depart very considerably from the sine form) it lends itself to simple calculations and its use as a standard permits comparison between systems on a common basis. With sine-wave modulation the *average* power in the modulated signal over any number of full cycles of the modulation frequency is found to be 1-1/2 times the power in the unmodulated carrier. In other words, the power output increases 50 percent with 100-percent modulation by a sine wave.

This relationship is very useful in the design of modulation systems and modulators, because any such system that is capable of increasing the *average* power output by 50 percent with sine-wave

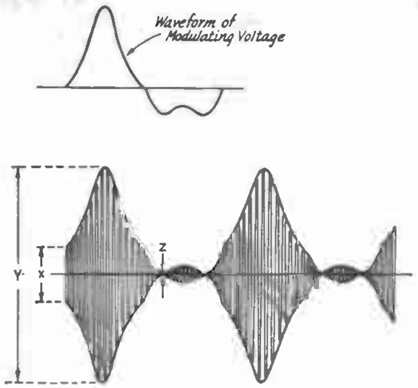


Fig. 12-3 – Modulation by an unsymmetrical wave form. This drawing shows 100-percent downward modulation along with 300-percent upward modulation. There is no distortion, since the modulation envelope is an accurate reproduction of the wave form of the modulating voltage.

modulation automatically fulfills the requirement that the *instantaneous* power at the modulation uppeak be four times the carrier power. Consequently, systems in which the additional power is supplied from outside the modulated rf stage (e.g., plate modulation) usually are designed on a sine-wave basis as a matter of convenience. Modulation systems in which the additional power is secured from the modulated rf amplifier (e.g., grid modulation) usually are more conveniently designed on the basis of peak envelope power rather than average power.

The extra power that is contained in a modulated signal goes entirely into the sidebands, half in the upper sideband and half into the lower. As a numerical example, full modulation of a 100-watt carrier by a sine wave will add 50 watts of sideband power, 25 in the lower and 25 in the upper sideband. With lower modulation percentages, the sideband power is proportional to the *square* of the modulation percentage, i.e., 50-percent modulation will add 12.5 watts of sideband power, 6.25 watts in each sideband. Supplying this additional power for the sidebands is the object of all of the various systems devised for amplitude modulation.

No such simple relationship exists with complex wave forms. Complex wave forms such as speech do not, as a rule, contain as much average power as a sine wave. Ordinary speech wave forms have about half as much average power as a sine wave, for the same peak amplitude in both wave forms. Thus for the same modulation percentage, the sideband power with ordinary speech will average only about half the power with sine-wave modulation, since it is the peak envelope amplitude, not the average power, that determines the percentage of modulation.

Unsymmetrical Modulation

In an ordinary electric circuit it is possible to increase the amplitude of current flow indefinitely, up to the limit of the power-handling capability of the components, but it cannot very well be decreased to less than zero. The same thing is true of the amplitude of an rf signal; it can be modulated *upward* to any desired extent, but it cannot be modulated *downward* more than 100 percent.

When the modulating wave form is unsymmetrical it is possible for the upward and downward modulation percentages to be different. A simple case is shown in Fig. 12-3. The positive peak of the modulating signal is about 3 times the amplitude of the negative peak. If, as shown in the drawing, the modulating amplitude is adjusted so that the peak downward modulation is just 100 percent ($Z = 0$) the peak upward modulation is 300 percent ($Y = 4X$). The carrier amplitude is represented by X, as in Fig. 12-2. The modulation envelope reproduces the wave form of the modulating signal accurately, hence there is no distortion. In such a modulated signal the increase in power output with modulation is considerably greater than it is when the modulation is symmetrical. In Fig. 12-3 the peak envelope amplitude, Y, is four times the carrier amplitude, X, so the peak-envelope power (PEP) is 16 times the carrier power. When the upward modulation is more than 100 percent the power capacity of the modulating system obviously must be increased sufficiently to take care of the much larger peak amplitudes. Such a system of modulation, often called "supermodulation," was popular among amateurs in the early 1950s. (See bibliography at the end of this chapter.)

Overmodulation

If the amplitude of the modulation on the downward swing becomes too great, there will be a period of time during which the rf output is entirely cut off. This is shown in Fig. 12-4. The shape of the downward half of the modulating wave is no longer accurately reproduced by the modulation envelope, consequently the modulation is distorted. Operation of this type is called overmodulation.

The distortion of the modulation envelope causes new frequencies (harmonics of the modulating frequency) to be generated. These combine

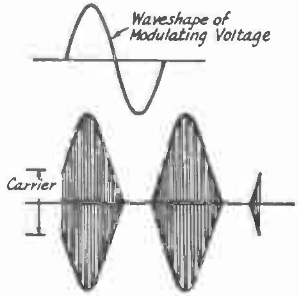


Fig. 12-4 – An overmodulated signal. The modulation envelope is not an accurate reproduction of the wave form of the modulating voltage. This, or any type of distortion occurring during the modulation process, generates spurious sidebands or "splatter."

with the carrier to form new side frequencies that widen the channel occupied by the modulated signal, as shown in Fig. 12-1C. These spurious frequencies are commonly called "splatter."

It is important to realize that the channel occupied by an amplitude-modulated signal is dependent *on the shape of the modulation envelope.* If this wave shape is complex and can be resolved into a wide band of audio frequencies, then the channel occupied will be correspondingly large. An overmodulated signal splatters and occupies a much wider channel than is necessary because the "clipping" of the modulating wave that occurs at the zero axis changes the envelope wave shape to one that contains high-order harmonics of the original modulating frequency. These harmonics appear as side frequencies separated by, in some cases, many kilohertz from the carrier frequency.

Because of this clipping action at the zero axis, it is important that care be taken to prevent applying too large a modulating signal in the downward direction. Overmodulation downward results in more splatter than is caused by most other types of distortion in a phone transmitter.

AMPLITUDE MODULATION METHODS

MODULATION SYSTEMS

As explained in the preceding section, amplitude modulation of a carrier is accompanied by an increase in power output, the additional power being the "useful" or "talk power" in the sidebands. This additional power may be supplied from an external source in the form of audio-frequency power. It is then added to the unmodulated power input to the amplifier to be modulated, after which the combined power is converted to rf. This is the method used in plate or collector modulation. It has the advantage that the rf power is generated at the high-efficiency characteristic of Class C amplifiers – of the order of 65 to 75 percent – but has the accompanying disadvantage that generating the audio-frequency power is rather expensive.

An alternative that does not require relatively large amounts of audio-frequency power makes use of the fact that the power output of an amplifier can be controlled by varying the potential of a tube or transistor element – such as a control or screen grid or a transistor base – that does not, in itself, consume appreciable power. In this case the additional power during modulation is secured by sacrificing carrier power; in other words, a tube is capable of delivering only so much total power within its ratings, and if more must be delivered at full modulation, then less is available for the unmodulated carrier. Systems of this type must of necessity work at rather low efficiency at the unmodulated carrier level. As a practical working rule, the efficiency of the modulated rf amplifier is of the order of 30 to 35 percent, and the unmodulated carrier power output obtainable with such a system is only about one-fourth to one-third that obtainable from the same amplifier with plate modulation.

PLATE OR COLLECTOR MODULATION

Fig. 12-5 shows a system of plate modulation, in this case with a triode rf tube. A balanced (push-pull Class A, Class AB or Class B) modulator is transformer coupled to the plate circuit of the modulated rf amplifier. The audio-frequency power generated by the modulator is combined with the dc power in the modulated-amplifier plate circuit by transfer through the coupling transformer, T. For 100-percent modulation the audio-frequency power output of the modulator and the turns ratio of the coupling transformer must be such that the voltage at the plate of the of the modulated amplifier varies between zero and twice and dc operating plate voltage, thus causing corresponding variations in the amplitude of the rf output. The tubes of Fig. 12-5 may be replaced with transistors, either bipolar or FET, for collector or drain modulation.

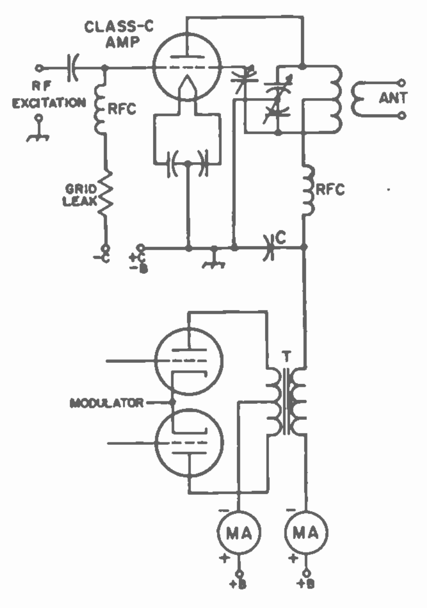


Fig. 12-5 – Plate modulation of a Class C rf amplifier. The rf plate bypass capacitor, C, in the amplifier stage should have reasonably high reactance at audio frequencies. A value of the order of .001 μF to .005 μF is satisfactory in practically all cases for vacuum-tube circuits. A considerably higher value will be required if the vacuum tubes are replaced by transistors – in the order of a few microfarads.

Audio Power

As stated earlier, the average power output of the modulated stage must increase during modulation. The modulator must be capable of supplying to the modulated rf stage sine-wave audio power equal to 50 percent of the dc input power. For example, if the dc input power to the rf stage is 100 watts, the sine-wave audio power output of the modulator must be 50 watts.

Although the total power input (dc plus audio-frequency ac) increases with modulation, the dc plate or collector current of a modulated amplifier should not change when the stage is modulated. This is because each increase in voltage and current is balanced by an equivalent decrease in voltage and current on the next half cycle of the modulating wave. Dc instruments cannot follow the af variations, and since the average dc plate or collector current and voltage of a properly operated amplifier do not change, neither do the meter readings. A change in current with modulation indicates nonlinearity. On the other hand, a thermocouple rf ammeter connected in the antenna, or transmission line, will show an increase in rf current with modulation, because instruments of this type respond to power rather than to current or voltage.

Modulating Impedance; Linearity

The modulating impedance, or load resistance presented to the modulator by the modulated rf amplifier, is equal to

$$Z_m = \frac{E_b}{I_p} \times 1000 \text{ ohms}$$

where E_b = Dc plate or collector voltage
I_p = Dc plate or collector current (mA)
E_b and I_p are measured without modulation.

The power output of the rf amplifier must vary as the square of the instantaneous plate or collector voltage (the rf output voltage must be proportional to the plate or collector voltage) for the modulation to be linear. This will be the case when the amplifier operates under Class C conditions. The linearity depends upon having sufficient grid or base excitation and proper bias, and upon the adjustment or circuit constants to the proper values.

Screen-Grid RF Amplifiers

Screen-grid tubes of the pentode or beam-tetrode type can be used in Class C plate-modulated amplifiers by applying the modulation to both the plate and screen grid. The usual method of feeding the screen grid with the necessary dc and modulation voltages is shown in Fig. 12-6. The dropping resistor, R, should be of the proper value to apply normal dc voltage to the screen under steady carrier conditions. Its value can be calculated by taking the difference between plate and screen voltages and dividing it by the rated screen current.

The modulating impedance is found by dividing the dc plate voltage by the sum of the plate and screen currents. The plate voltage multiplied by the sum of the two currents gives the power input to be used as the basis for determining the audio power required from the modulator.

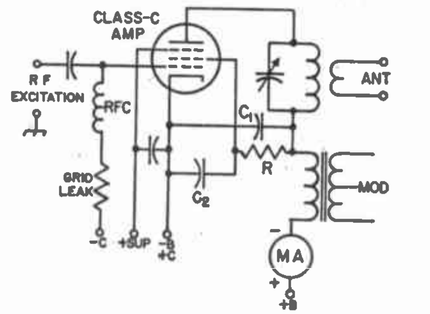


Fig. 12-6 — Plate and screen modulation of a Class C rf amplifier using a screen-grid tube. The plate rf bypass capacitor, C1, should have reasonably high reactance at all audio frequencies; a value of .001 to .005 μF is generally satisfactory. The screen bypass, C2, should not exceed .002 μF in the usual case.

Modulation of the screen along with the plate is necessary because the screen voltage has a much greater effect on the plate current than the plate voltage does. The modulation characteristic is nonlinear if the plate alone is modulated.

Choke-Coupled or Heising Modulation

One of the oldest types of plate modulating systems is the choke-coupled Class A or Heising modulator shown in Fig. 12-7. Because of the relatively low power output and plate efficiency of a Class A amplifier, the method is rarely used now except for a few special applications.

The audio power output of the modulator is combined with the dc power in the plate circuit through the modulation choke, L1, which has a high impedance at audio frequencies. This technique of modulating the rf signal is similar to the case of the transformer-coupled modulator but there is considerably less freedom in adjustment since no transformer is available for matching impedances. The dc input power to the rf stage must not exceed twice the rated af power output of the modulator, and for 100-percent modulation the plate voltage on the modulator must be higher than the plate voltage on the rf amplifier. This is because the af voltage developed by the modulator cannot swing to zero without a great deal of distortion. R1 provides the necessary dc voltage drop between the modulator and the rf amplifier. The voltage drop across this resistor must equal the minimum instantaneous plate voltage on the modulator tube under normal operating conditions. C1, an audio-frequency bypass across R1, should have a capacitance such that its reactance at 100 Hz is not more than about one-tenth the resistance of R1. Without R1-C1 the percentage of modulation is limited to 70 to 80 percent in the average case.

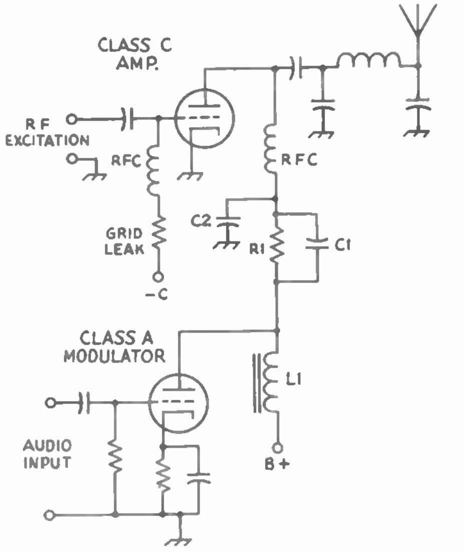


Fig. 12-7 – Choke-coupled Class A modulator. The modulation choke, L1, should have a value of 5 H or more. A value of .001 to .005 μF is satisfactory for C2. See text for discussion of C1 and R1.

GRID MODULATION

The principal disadvantage of plate modulation is that a considerable amount of audio power is necessary. This requirement can be avoided by applying the modulation to a grid element in the modulated amplifier. However, serious disadvantages of grid modulation are the reduction in the carrier power output obtainable from a given rf amplifier tube and the more rigorous operating requirements and more complicated adjustment.

The term "grid modulation" as used here applies to all types – control grid, screen, or suppressor – since the operating principles are exactly the same no matter which grid is actually modulated. (Screen-grid modulation is the most commonly used technique of the three types listed here.) With grid modulation the plate voltage is constant, and the increase in power output with modulation is obtained by making both the plate current and plate efficiency vary with the modulating signal. The efficiency obtainable at the envelope peak depends on how carefully the modulated amplifier is adjusted, and sometimes can be as high as 80 percent. It is generally less when the amplifier is adjusted for good linearity, and under average conditions a round figure of 2/3, or 66 percent, is representative. The efficiency without modulation is only half the peak efficiency, or about 33 percent. This low average efficiency reduces the permissible carrier output to about one-fourth the power obtainable from the same tube in cw operation, and to about one-third the carrier output obtainable from the tube with plate modulation.

The modulator is required to furnish only the audio power dissipated in the modulated grid under the operating conditions chosen. A speech amplifier capable of delivering 3 to 10 watts is usually sufficient.

Grid modulation does not give quite as linear a modulation characteristic as plate modulation, even under optimum operating conditions. When misadjusted the nonlinearity may be severe, resulting in considerable distortion and splatter.

Screen Grid Modulation

Screen modulation is probably the simplest and most popular form of grid modulation, and the least critical of adjustment. The most satisfactory way to apply the modulating voltage to the screen is through a transformer.

With practical tubes it is necessary to drive the screen somewhat negative with respect to the cathode to get complete cutoff of rf output. For this reason the peak modulating voltage required for 100-percent modulation is usually 10 percent or so greater than the dc screen voltage. The latter, in turn, is approximately half the rated screen voltage recommended by the manufacturer under maximum ratings for radiotelegraph operation. The audio power required for 100-percent modulation is approximately one-fourth the dc power input to the screen in cw operation, but varies somewhat with the operating conditions.

Controlled Carrier

As explained earlier, a limit is placed on the output obtainable from a grid-modulation system by the low rf-amplifier plate efficiency (approximately 33 percent) under unmodulated carrier

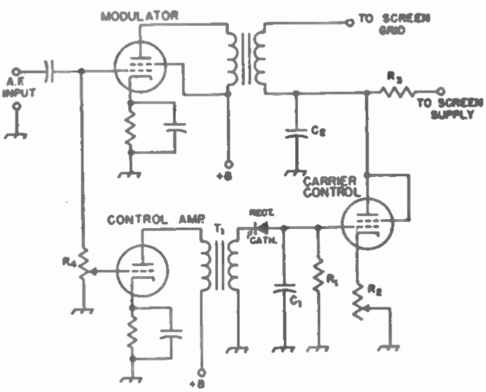


Fig. 12-8 – Circuit for carrier control with screen modulation. A small triode such as the 6C4 can be used as the control amplifier and a 6Y6G is suitable as a carrier-control tube. T1 is an interstage audio transformer having a 1-to-1 or larger turns ratio. R4 is a 0.5-megohm volume control and also serves as the grid resistor for the modulator. A germanium diode may be used as the rectifier. R3 may be the normal screen dropping resistor. C1-R1 and C2-R3 should have a time constant of about 0.1 second.

conditions. The plate efficiency increases with modulation, since the output increases while the dc input remains constant, and reaches a maximum in the neighborhood of 50 percent with 100-percent sine-wave modulation. If the power input to the amplifier can be reduced during periods when there is little or no modulation, thus reducing the plate loss, advantage can be taken of the higher efficiency at full modulation to obtain higher effective output. This can be done by varying the dc power input to the modulated stage in accordance with *average* variations in voice intensity, in such a way as to maintain just sufficient carrier power to keep the modulation high, but not exceeding 100 percent, under all conditions. Thus the carrier amplitude is controlled by the average voice intensity. Properly utilized, controlled carrier permits increasing the carrier output at maximum level to a value about equal to the rated plate dissipation of the tube, twice the output obtainable with constant carrier.

It is desirable to control the power input just enough so that the plate loss, without modulation, is safely below the tube rating. Excessive control is disadvantageous because the distant receiver's avc system must continually follow the variations in average signal level. The circuit of Fig. 12-8 permits adjustment of both the maximum and minimum power input, and separates the functions of modulation and carrier control. A portion of the audio voltage at the modulator grid is applied to a Class A "control amplifier," which drives a rectifier circuit to produce a dc voltage negative with respect to ground. C1 filters out the audio variations, leaving a dc voltage proportional to the average voice level. This voltage is applied to the grid of a "clamp" tube to control the dc screen voltage and thus the rf carrier level.

DOUBLE-SIDEBAND GENERATORS

The a-m carrier can be suppressed or nearly eliminated by using a balanced modulator. The basic principle in any balanced modulator is to introduce the carrier in such a way that it does not appear in the output but so that the sidebands will. This requirement is satisfied by introducing the audio in push-pull and the rf drive in parallel, and connecting the output in push-pull. Balanced modulators can also be connected with the rf drive and audio inputs in push-pull and the output in parallel with equal effectiveness.

Vacuum-tube balanced modulators can be operated at high power levels and the double-sideband output can be used directly into the antenna. Past issues of *QST* have given construction details on such transmitters (see, for example, Rush, "180-Watt D.S.B. Transmitter," *QST* July, 1966). A dsb signal can be copied by the same methods that are used for single-sideband signals, provided the receiver has sufficient selectivity to reject one of the sidebands. In any balanced-modulator circuit, no rf output will exist with no audio signal. When audio is applied, the balance of the modulator is upset so that sum and difference frequencies (sidebands) appear at the output. Further information on balanced modulators is presented in Chapter 13.

CHECKING A-M PHONE OPERATION

USING THE OSCILLOSCOPE

Proper adjustment of a phone transmitter is aided immeasurably by the oscilloscope. The scope will give more information, more accurately, than almost any collection of other instruments that might be named. Furthermore, an oscilloscope that is entirely satisfactory for the purpose is not necessarily an expensive instrument; the cathode-ray tube and its power supply are about all that are needed. Amplifiers and linear sweep circuits are by no means necessary.

In the simplest scope circuit, radio-frequency voltage from the modulated amplifier is applied to the vertical deflection plates of the tube, usually through blocking capacitors, and audio-frequency voltage from the modulator is applied to the horizontal deflection plates. As the instantaneous amplitude of the audio signal varies, the rf output of the transmitter likewise varies, and this produces a wedge-shaped pattern or trapezoid on the screen. If the oscilloscope has a built-in horizontal sweep, the rf voltage can be applied to the vertical plates as before, and the sweep will produce a pattern that follows the modulation envelope of the transmitter output, provided the sweep frequency is lower than the modulation frequency. This produces a wave-envelope modulation pattern.

The Wave-Envelope Pattern

The connections for the wave-envelope pattern are shown in Fig. 12-9A. The vertical deflection plates are coupled to the amplifier tank coil (or an antenna coil) through a low-impedance (coax, twisted pair, etc.) line and pickup coil. As shown in the alternative drawing, a resonant circuit tuned to the operating frequency may be connected to the vertical plates, using link coupling between it and the transmitter. This will eliminate rf harmonics, and the tuning control provides a means for adjustment of the pattern height.

If it is inconvenient to couple to the final tank coil, as may be the case if the transmitter is tightly shielded, the pickup loop may be coupled to the tuned tank of a matching circuit or antenna coupler. Any method (even a short antenna coupled to the tuned circuit shown in the "alternate input connections" of Fig. 12-9A) that will pick up enough rf to give a suitable pattern height may be used.

The position of the pickup coil should be varied until an unmodulated carrier pattern, Fig. 12-10A, of suitable height is obtained. The horizontal sweep voltage should be adjusted to make the width of the pattern somewhat more than half the diameter of the screen. When voice modulation is applied, a rapidly changing pattern of varying height will be obtained. When the maximum height of this pattern is just twice that of the carrier alone, the wave is being modulated 100 percent. This is illustrated by Fig. 12-10C.

If the height is greater than twice the unmodulated carrier amplitude, as illustrated in Fig. 12-10D, the wave is overmodulated in the upward direction. Overmodulation in the downward direction is indicated by a gap in the pattern at the reference axis, where a single bright line appears on the screen. Overmodulation in either direction may take place even when the modulation in the other direction is less than 100 percent.

The Trapezoidal Pattern

Connections for the trapezoid or wedge pattern as used for checking a-m are shown in Fig. 12-9B. The vertical plates of the CR tube are coupled to the transmitter tank through a pickup loop, preferably using a tuned circuit, as shown in the upper drawing, adjustable to the operating frequency. Audio voltage from the modulator is applied to the horizontal plates through a voltage divider, R1-R2. This voltage should be adjustable so a suitable pattern width can be obtained; a 0.25-megohm volume control can be used at R2 for this purpose.

The resistance required at R1 will depend on the dc voltage on the modulated element. The total resistance of R1 and R2 in series should be about 0.25 megohm for each 100 volts. For example, if a plate-modulated amplifier operates at 1500 volts, the total resistance should be 3.75 megohms, 0.25 megohm at R2 and the remainder, 3.5 megohms, in R1. R1 should be composed of individual resistors not larger than 0.5 megohm each, in which case 1-watt resistors will be satisfactory.

For adequate coupling at 100 Hz, the capacitance in microfarads of the blocking capacitor, C, should be at least .05/R, where R is the total resistance (R1 + R2) in megohms. In the example above, where R is 3.75 megohms, the capacitance should be .05/3.75 = .013 μF or more. The voltage rating of the capacitor should be at least twice the dc voltage applied to the modulated element.

Trapeziodal patterns for various conditions of modulation are shown in Fig. 12-10, each alongside the corresponding wave-envelope pattern. With no signal, only the cathode-ray spot appears on the screen. When the unmodulated carrier is applied, a vertical line appears; the length of the line should be adjusted, by means of the pickup-coil coupling, to a convenient value. When the carrier is modulated, the wedge-shaped pattern appears; the higher

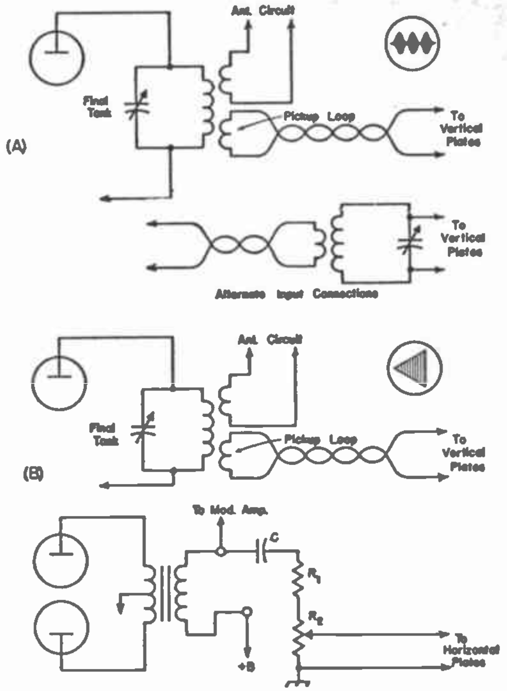


Fig. 12-9 — Methods of connecting the oscilloscope for modulation checking. A — connections for wave-envelope pattern with any modulation method; B — connections for trapezoidal pattern with plate or screen modulation.

the modulation percentage, the wider and more pointed the wedge becomes. At 100-percent modulation it just makes a point at one end of the horizontal axis, and the height at the other end is equal to twice carrier height. Overmodulation in the upward direction is indicated by increased height, at one end, and downward by an extension along the horizontal axis at the pointed end.

CHECKING A-M TRANSMITTER PERFORMANCE

The trapezoidal pattern is generally more useful than the wave-envelope pattern for checking the operation of the phone transmitter. However, both types of patterns have their special virtues, and the best test setup is one that makes both available. The trapezoidal pattern is better adapted to showing the performance of a modulated amplifier from the standpoint of inherent linearity, without regard to the wave form of the audio modulating signal, than is the wave-envelope pattern. Distortion in the audio signal also can be detected in the trapezoidal pattern, although experience in analyzing scope patterns is required to recognize it.

If the wave-envelope pattern is used with a sine-wave audio modulating signal, distortion in the

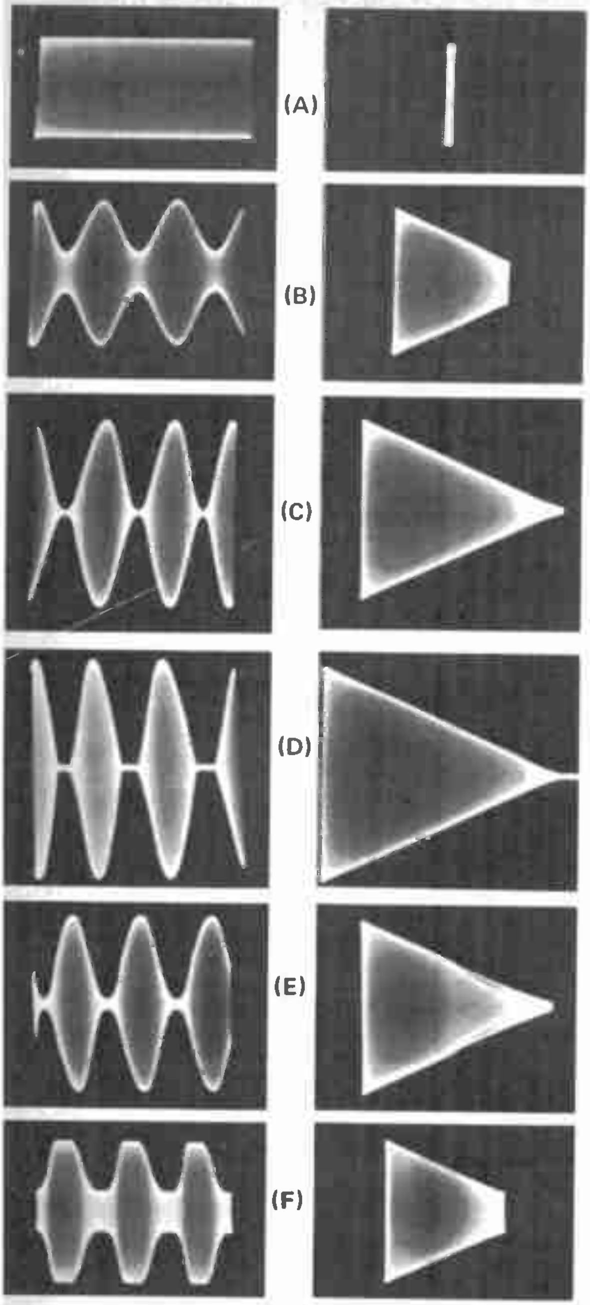


Fig. 12-10 — Oscilloscope patterns showing various forms of modulation of an rf amplifier. At left, wave-envelope patterns; at right, corresponding trapezoidal patterns. The wave-envelope patterns were obtained with a linear oscilloscope sweep having a frequency one-third that of the sine-wave audio modulating frequency, so that three cycles of the modulation envelope may be seen. Shown at A is an unmodulated carrier, at B approximately 50-percent modulation, and at C, 100-percent modulation. The photos at D show modulation in excess of 100 percent. E and F show the results of improper operation or circuit design. See text.

modulation envelope is easily recognizable; however, it is difficult to determine whether the distortion is caused by lack of linearity of the rf stage or by af distortion in the modulator. If the trapezoidal pattern shows good linearity in such a case, the trouble obviously is in the audio system. It is possible, of course, for both defects to be present simultaneously. If they are, the rf amplifier should be made linear first; then any distortion in the modulation envelope will be the result of improper operation in the speech amplifier or modulator, or in coupling the modulator to the modulated rf stage.

Rf Linearity

The trapezoidal pattern is a graph of the modulation characteristic of the modulated amplifier. The sloping sides of the wedge show the rf amplitude for every value of instantaneous modulating voltage. If these sides are perfectly straight lines, the modulation characteristic is linear. If the sides show curvature, the characteristic is nonlinear to an extent shown by the degree to which the sides depart from perfect straightness. This is true regardless of the modulating wave form. If these edges tend to bend over toward the horizontal at the maximum height of the wedge, the amplifier is "flattening" on the modulation uppeaks. This is usually caused by attempting to get too large a carrier output, and can be corrected by tighter coupling to the antenna or by a decrease in the dc screen voltage. The slight "tailing off" at the modulation downpeak (point of the wedge) can be minimized by careful adjustment of excitation and plate loading.

Several types of improper operation are shown in Fig. 12-10. The patterns at E show the effect of a too long time constant in the screen circuit, in an amplifier getting its screen voltage through a dropping resistor, both plate and screen being modulated. The "double-edged" pattern is the result of audio phase shift in the screen circuit combined with varying screen-to-cathode resistance during modulation. This effect can be reduced by reducing the screen bypass capacitance, and also by connecting resistance (to be determined experimentally, but of the same order as the screen dropping resistance) between screen and cathode.

The pictures at the bottom, F, show the effect of insufficient audio power. Although the trapezoidal pattern shows good linearity in the rf amplifier, the wave-envelope pattern shows flattened peaks (both positive and negative) in the modulation envelope even though the audio signal applied to the amplifier was a sine wave. More speech-amplifier gain merely increases the flattening without increasing the modulation percentage in such a case. The remedy is to use a larger modulator or less input to the modulated rf stage. In some cases the trouble may be caused by an incorrect modulation-transformer turns ratio, causing the modulator to be overloaded before its maximum power output capabilities are reached.

GENERAL-PURPOSE AMPLITUDE MODULATORS

The two modulator circuits shown in Figs. 12-11 and 12-12 can be employed to deliver from 3 to 70 watts of audio power. The basic designs are taken from RCA's *Audio Design Phase 2.* The complementary-symmetry circuit, Fig. 12-11, is characterized by a Class A driver and a complementary pair (npn/pnp) of output transistors. The primary advantages of this circuit are simplicity and economy. Common conduction is minimized because the transistor which is "off" during half of the audio cycle is reverse biased. The output transistors are operated at zero bias, providing excellent dc stability. Elaborate regulated power supplies are not required. The complementary-symmetry amplifier is limited to about 20 watts output because of the high level of heat that the driver stage must dissipate. Component values and transistor types are given in Table 12-I for 3-, 5-, 12-, and 20-watt designs.

For higher power levels, the quasi-complementary circuit (Fig. 12-12) is usually chosen. Here a Class A predriver feeds a Class B npn/pnp driver

TABLE 12-I

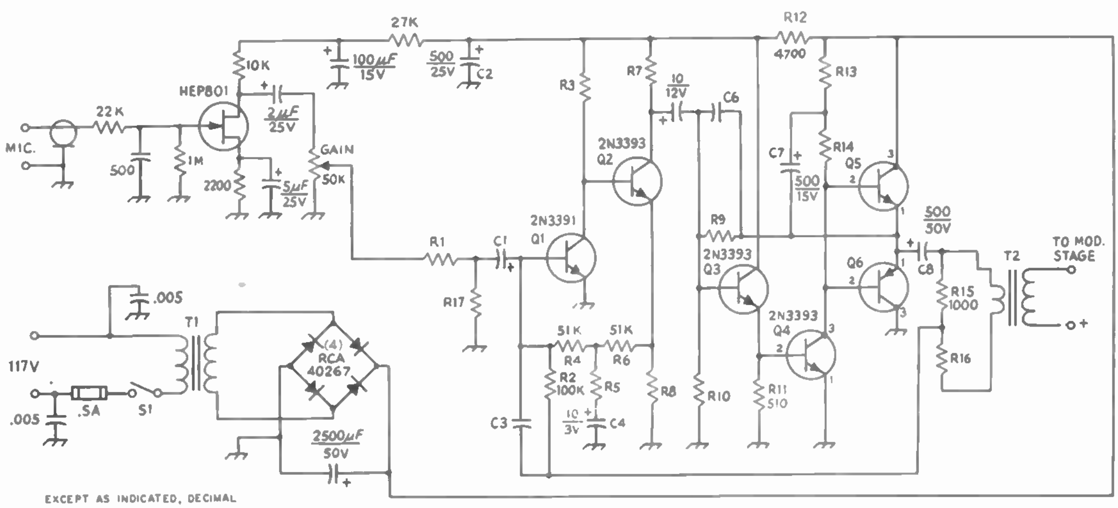
PARTS VALUES FOR COMPLEMENTARY – SYMMETRY CIRCUIT

Power (Watts)	*R1*	*R3*	*R5*	*R7*	*R8*	*R9*	*R10*	*R13*	*R14*	*R16*	*R17*	*C1 µF*	*C3 pF*	*C6 pF*	*Q4*	*Q5*	*Q6*	*T1*	
3	91k	68k	2.7k	3.9k	620	33k	5.6k	120	150	22	22k	0.1/6V	10	100	40611	40610	40609	15V, 1A	(Stancor TP-4)
5	51k	68k	3.3k	3.9k	620	27k	3.6k	110 (1W)	110 (1W)	27	22k	0.25/6V	5	150	40616	40615	40614	17V, 1A	(Stancor TP-4)
12	16k	91k	7.5k	2.7k	390	18k	1.8k	91 (2W)	91 (2W)	56		1/6V	10	220	40389	40622	40050	25V, 1A	(Stancor TP-4)
20	8.2k	91k	8.2k	2.2k	360	22k	1.3k	100 (2W)	100 (2W)	100		2/6V	10	270	40628	40627	40626	32V, 1A	(C. P. Elec. 10596)

TABLE 12-II

PARTS VALUES FOR QUASI-COMPLEMENTARY-SYMMETRY CIRCUIT

Power (Watts)	*R3*	*R7*	*R8*	*R10*	*R11*	*R22 R23*	*Q4*	*Q5*	*Q6 Q7*	*T1*		
25	12k	680	1800	2200	270	0.43(5W)	2N3568	2N3638	40632	37V	1.5A	(C. P. Elec. 10596)
40	15k	560	2200	2700	390	0.39(5W)	40635	40634	40633	46V	2A	(C. P. Elec. 10596)
70	18k	470	2700	3300	470	0.33(5W)	40594	40595	40636	60V	2.5A	(C. P. Elec. 10598)



EXCEPT AS INDICATED, DECIMAL VALUES OF CAPACITANCE ARE IN MICROFARADS (µF); OTHERS ARE IN PICOFARADS (pF OR µµF); RESISTANCES ARE IN OHMS; k = 1000, M = 1000 000

Fig. 12-11 — General-purpose amplitude modulator for 3 to 20 watts of audio power. Capacitors with polarity indicated are electrolytic. See Table 12-I for parts not listed below.
S1 — Spst toggle.
T2 — See text.

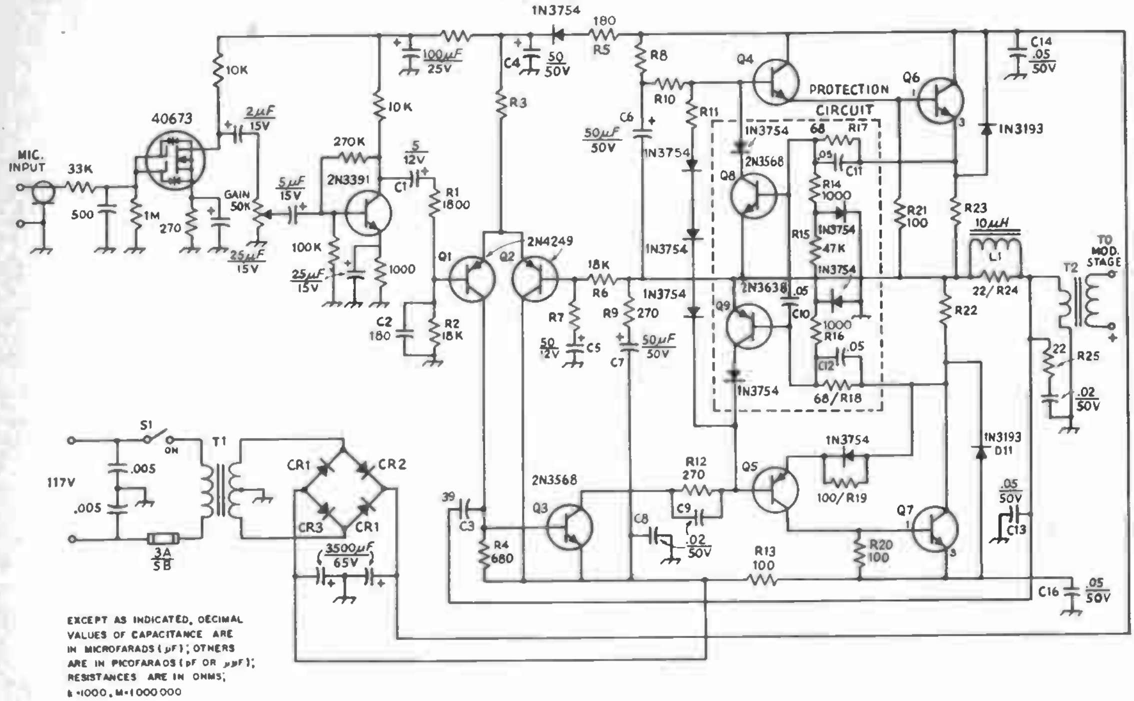


Fig. 12-12 — General-purpose amplitude modulator for 25 to 70 watts of audio power. Capacitors with polarity indicated are electrolytic. See Table 12-II for parts not listed below.

L1 — J. W. Miller 4622 or equiv.
S1 — Spst toggle.
T2 — See text.

pair, which, in turn, activates the npn output transistors. The danger of damage to the output stage from a short circuit is high, so protection is included. Table 12-II includes parts information for three power levels: 25, 40, and 70 watts.

All amplifiers are designed for an 8-ohm output, so T2 can be a standard audio output transformer in "reverse." The secondary impedance will depend on the impedance of the stage to be modulated.

Bibliography

Audio Design Phase 2, RCA, Somerville, NJ, 1970.

Preiss, "The '2-Meter QRP Mountain Topper'," *QST*, May, 1970.

Rush, "180-Watt D.S.B. Transmitter," *QST*, July, 1966.

Villard, " 'Supermodulation' – An Evaluation and Explanation," *QST*, December, 1950.

Single-Sideband Transmission

GENERATING THE SSB SIGNAL

A fully modulated a-m signal has two thirds of its power in the carrier and only one third in the sidebands. The sidebands carry the intelligence to be transmitted; the carrier "goes along for the ride" and serves only to demodulate the signal at the receiver. By eliminating the carrier and transmitting only the sidebands, or just one sideband, the available transmitter power is used to greater advantage. To recover the intelligence being transmitted, the carrier must be reinserted at the receiver, but this is no great problem with a proper detector circuit.

Assuming that the same final-amplifier tube or tubes are used either for normal a-m or for single sideband, carrier suppressed, it can be shown that the use of ssb can give an effective gain of up to 9 dB over a-m – equivalent to increasing the transmitter power 8 times. Eliminating the carrier also eliminates the heterodyne interference that so often spoils communication in congested phone bands.

Filter Method

Two basic systems for generating ssb signals are shown in Fig. 13-2. One involves the use of a bandpass filter having sufficient selectivity to pass one sideband and reject the other. Mechanical filters are available for frequencies below 1 MHz. From 0.2 to 10 MHz, good sideband rejection can be obtained with filters using four or more quartz crystals. Oscillator output at the filter frequency is combined with the audio signal in a balanced modulator, and only the upper and lower sidebands appear in the output. One of the sidebands is passed by the filter and the other rejected, so that an ssb signal is fed to the mixer. The signal is mixed with the output of a high-frequency rf oscillator to produce the desired output frequency. For additional amplification a linear rf amplifier must be used. When the ssb signal is generated around 500 kHz it may be necessary to convert twice to reach the operating frequency, since this simplifies the problem of rejecting the "image" frequencies resulting from the heterodyne process. The problem of image frequencies in the frequency conversions of ssb signals differs from the problem in receivers because the beating-oscillator frequency becomes important. Either balanced mixers or sufficient selectivity must be used to attenuate these frequencies in the output and hence minimize the possibility of unwanted radiations. (Examples of filter-type exciters can be found in various issues of *QST* and in *Single Sideband for the Radio Amateur.*)

Phasing Method

The second system is based on the phase relationships between the carrier and sidebands in a modulated signal. As shown in the diagram, the audio signal is split into two components that are identical except for a phase difference of 90 degrees. The output of the rf oscillator (which may be at the operating frequency, if desired) is likewise split into two separate components having a 90-degree phase difference. One rf and one audio component are combined in each of two separate balanced modulators. The carrier is suppressed in the modulators, and the relative phases of the sidebands are such that one sideband is balanced out and the other is augmented in the combined output. If the output from the balanced modulators is high enough, such an ssb exciter can work directly into the antenna, or the power level can be increased in a following amplifier.

Generally, the filter-type exciter is easier to adjust than is the phasing exciter. Most home built ssb equipment uses commercially made filters these days. The alignment is done at the factory, thus relieving the amateur of the sometimes tedious task of adjusting the filter for suitable bandpass characteristics. Filter-type exciters are more popular than phasing units and offer better carrier suppression and alignment stability. It is still practical for the builder to fabricate his own crystal-lattice filter by utilizing low-cost surplus crystals. This possibility should not be overlooked if the builder is interested in keeping the overall cost of the home-built exciter at a minimum.

BALANCED MODULATORS

The carrier can be suppressed or nearly eliminated by using a balanced modulator or an extremely sharp filter. In ssb transmitters it is common practice to use both devices. The basic principle of any balanced modulator is to

Fig. 13-1 – Single sideband is the most popular of all the modes for amateur hf communication.

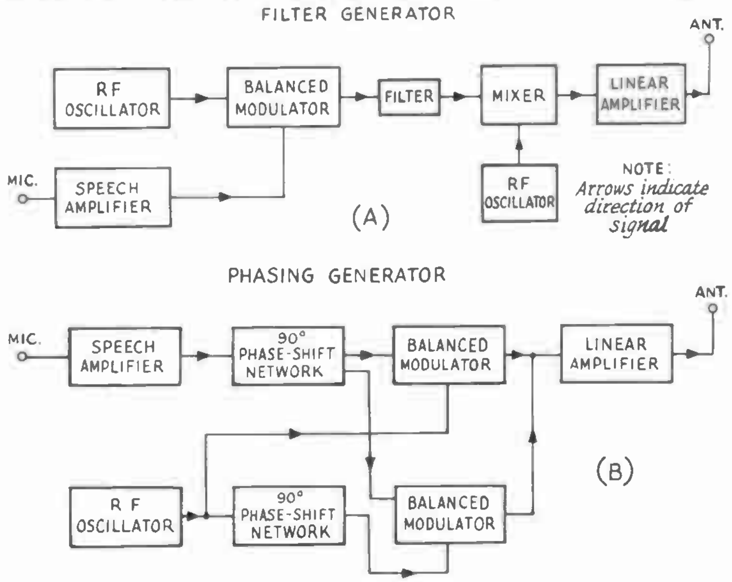


Fig. 13-2 — Two basic systems for generating single-sideband suppressed carrier signals.

introduce the carrier in such a way that it does not appear in the output, but so that the sidebands will. The type of balanced-modulator circuit chosen by the builder will depend upon the constructional considerations, cost, and whether diodes or transistors are to be employed.

In any balanced-modulator circuit there will be no output with no audio signal. When audio is applied, the balance is upset, and one branch will conduct more than the other. Since any modulation process is the same as "mixing" in receivers, sum and difference frequencies (sidebands) will be generated. The modulator is not balanced for the sidebands, and they will appear in the output.

In the rectifier-type balanced modulators shown in Fig. 13-3, at A and B, the diode rectifiers are connected in such a manner that, if they have equal forward resistances, no rf can pass from the carrier source to the output circuit via either of the two possible paths. The net effect is that no rf energy appears in the output. When audio is applied, it unbalances the circuit by biasing the diode (or diodes) in one path, depending upon the instantaneous polarity of the audio, and hence some rf will appear in the output. The rf in the output will appear as a double-sideband suppressed-carrier signal.

In any diode modulator, the rf voltage should be at least 6 to 8 times the peak audio voltage for minimum distortion. The usual operation involves a fraction of a volt of audio and several volts of rf. Desirable diode characteristics for balanced modulator and mixer service include: low noise, low forward resistance, high reverse resistance, good temperature stability, and fast switching time (for high-frequency operation). Fig. 13-4 lists the different classes of diodes, giving the ratio of forward-to-reverse resistance of each. This ratio is an important criterion in the selection of diodes. Also. the individual diodes used should have closely matched forward and reverse resistances; an ohmmeter can be used to select matched pairs or quads.

One of the simplest diode balanced modulators in use is that of Fig. 13-3A. Its use is usually limited to low-cost portable equipment in which a high degree of carrier suppression is not vital. A ring balanced modulator, shown in Fig. 13-3B, offers good carrier suppression at low cost. Diodes CR1 through CR4 should be well matched and can be 1N270s or similar. C1 is adjusted for best rf phase balance as evidenced by maximum carrier null. R1 is also adjusted for the best carrier null obtainable. It may be necessary to adjust each control several times to secure optimum suppression.

Varactor diodes are part of the unusual circuit shown in Fig. 13-3C. This arrangement allows single-ended input of near-equal levels of audio and carrier oscillator. Excellent carrier suppression, 50 dB or more, and a simple method of unbalancing the modulator for cw operation are features of this design. CR1 and CR2 should be rated at 20 pF for a bias of -4 V. R1 can be adjusted to cancel any mismatch in the diode characteristics, so it isn't necessary that the varactors be well matched. T1 is wound on a small-diameter toroid core. The tap on the primary winding of this transformer is at the center of the winding.

A bipolar-transistor balanced modulator is shown in 13-3D. This circuit is similar to one used by Galaxy Electronics and uses closely matched transistors at Q1 and Q2. A phase splitter (inverter) Q3, is used to feed audio to the balanced modulator in push-pull. The carrier is supplied to the circuit in parallel and the output is taken in push-pull. CR1 is a Zener diode and is used to stabilize the dc voltage. Controls R1 and R2 are adjusted for best carrier suppression.

The circuit at E offers superior carrier suppression and uses a 7360 beam-deflection tube as a balanced modulator. This tube is capable of

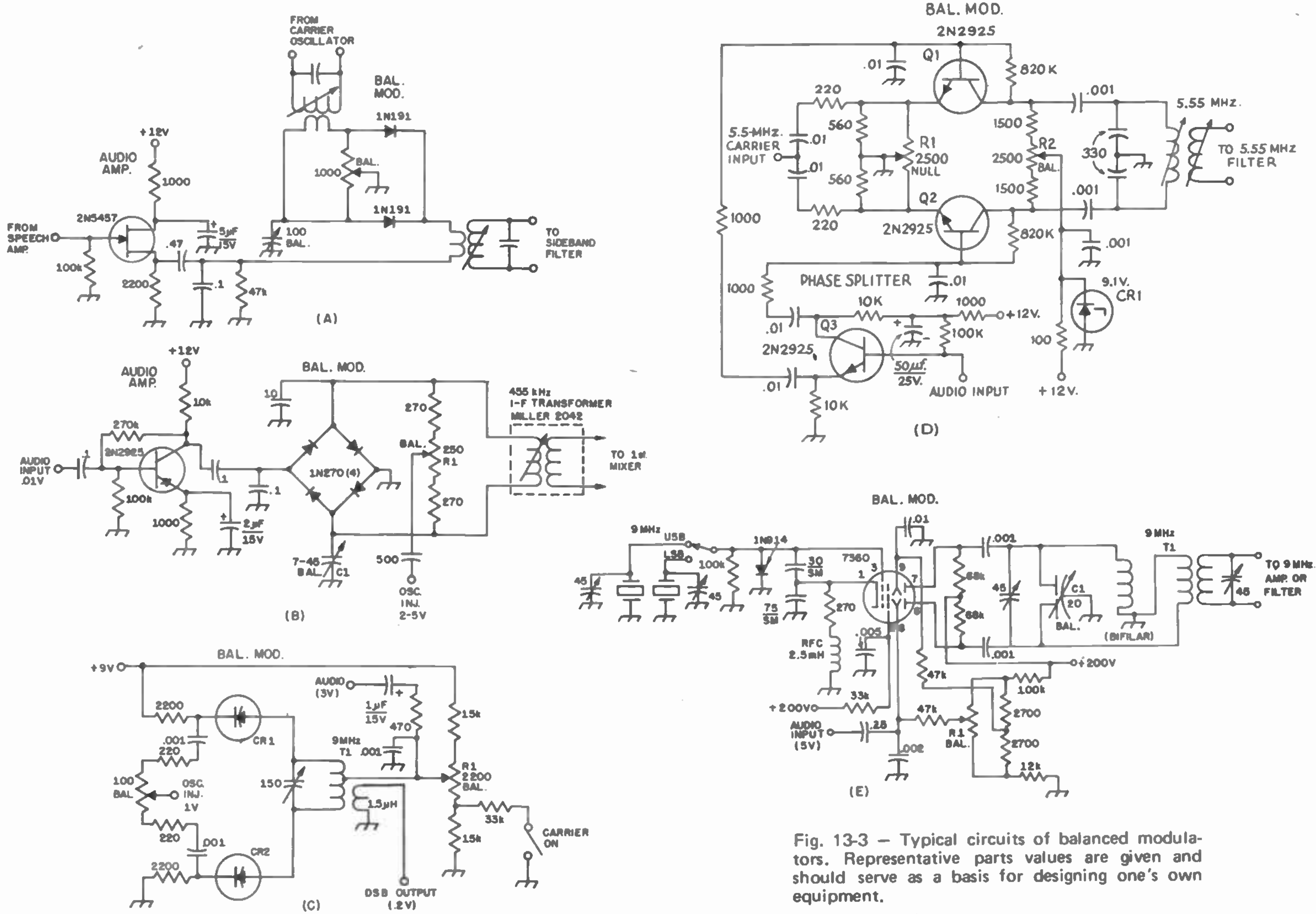


Fig. 13-3 – Typical circuits of balanced modulators. Representative parts values are given and should serve as a basis for designing one's own equipment.

Diode Type	Ratio (M = 1,000,000)
Point-contact germanium (1N98)	500
Small-junction germanium (1N270)	0.1M
Low-conductance silicon (1N457)	48M
High-conductance silicon (1N645)	480M
Hot-carrier (HPA-2800)	2000M

Fig. 13-4 — Table showing the forward-to-reverse resistance ratio for the different classes of solid-state diodes.

providing as much as 60 dB of carrier suppression. When used with mechanical or crystal-lattice filters the total carrier suppression can be as great as 80 dB. Most well-designed balanced modulators can provide between 30 and 50 dB of carrier suppression; hence the 7360 circuit is highly desirable for optimum results. The primary of transformer T1 should be bifilar wound for best results.

IC Balanced Modulators

Integrated circuits (ICs) are presently available for use in balanced-modulator and mixer circuits. A diode array such as the RCA CA3039 is ideally suited for use in circuits such as that of Fig. 13-5A. Since all diodes are formed on a common silicon chip, their characteristics are extremely well matched. This fact makes the IC ideal in a circuit where good balance is required. The hot-carrier diode also has closely matched characteristics and excellent temperature stability. Using broad-band toroidal-wound transformers, it is possible to construct a circuit similar to that of Fig. 13-6 which will have 40 dB of carrier suppression without the need for balance controls. T1 and T2 consist of trifilar windings, 12 turns of No. 32 enam. wire wound on a 1/2-inch toroid core. Another device with good inherent balance is the special IC made for modulator/mixer service, such as the Motorola MC1496G or Signetics S5596. A sample circuit using the MC1496 can be seen in Fig. 13-5B. R1 is adjusted for best carrier balance. The amount of energy delivered from the carrier generator effects the level of carrier suppression; 100 mV of injection is about optimum, producing up to 55 dB of carrier suppression. Additional information on balanced modulators and other ssb-generator circuits is given in the texts referenced at the end of this chapter.

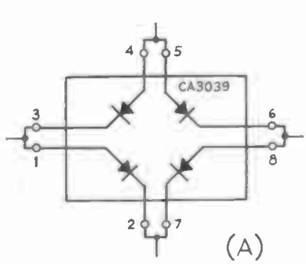
FILTERS

A home-built crystal lattice filter suitable for use in an ssb generator is shown in Fig. 13-7. This unit is composed of three half-lattice sections, with 2 crystals in each section, made with surplus hf crystals. The 330-ohm resistor between sections two and three reduces interaction and smooths the passband response. The leakage reactance between the two halves of L2 and L3 is tuned out by the capacitors connected in series with the center taps of these coils. L1 and L4, the input and output coils, resonate with the calculated value of terminating capacitance at 5060 kHz and reflect the needed inductance across the crystals. The 2000-ohm resistors complete the termination.

All the crystals were purchased as 5500-kHz FT-243s and etched to the desired frequencies with hydrofluoric acid. It is best to wash each crystal with soap and water and measure its frequency before etching. The crystals in each set of three should be as close to each other in frequency as possible, and the separation between the two groups should be about 1500 Hz.

Tuning the filter is quite simple since all four adjustements can be peaked for maximum output at a fixed alignment frequency. This frequency should be on the high side of the pass band and can be the carrier frequency used for lower-sideband transmission (5505.5 kHz in the case of the filter described). Using the carrier frequency it is only necessary to unbalance the balanced modulator to obtain a cw alignment signal. Of course, a signal generator and rf-probe-equipped VTVM can also be used. C1, C2, L1 and L4 are adjusted for maximum output.

A slightly better shape factor can be had by detuning the carrier oscillator to a lower alignment frequency corresponding to about the 4-dB-down point on the high-frequency side of the pass band. Fig. 13-8 shows the measured performance of the filter when aligned at 5505.2 kHz. The 6-dB bandwidth is 2750 Hz.



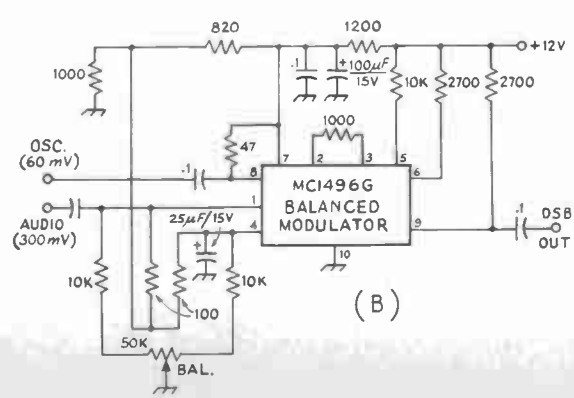


Fig. 13-5 — Additional balanced-modulator circuits in which integrated circuits are used.

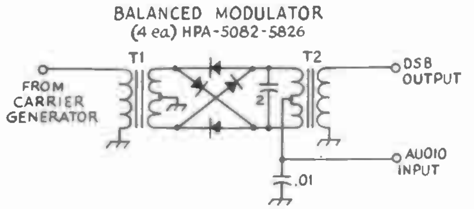


Fig. 13-6 — Balanced modulator design using hot-carrier diodes.

The (suppressed) carrier frequency must be adjusted so that it falls properly on the slope of the filter characteristic. If it is too close to the filter mid frequency the sideband rejection will be poor; if it is too far away there will be a lack of "lows" in the signal.

Ordinarily, the carrier is placed on one side of the curve, depending upon which sideband is desired, which is approximately 20 dB down from the peak. It is sometimes helpful to make provisions for "rubbering" the crystal of the carrier oscillator so that the most natural voice quality can be realized when making initial adjustments.

Using Commercial Crystal Filters

Some builders may not have adequate testing facilities for building and aligning their own filters. In such instances it is possible to purchase ready-made units which are prealigned and come equipped with crystals for upper- and lower-side-band use. Spectrum International[1] has two types for use at 9 MHz. Another manufacturer, McCoy Electronics Co.,[2] sells 9-MHz models for amateur use, and other filters are available surplus.[3]

Mechanical Filters

Mechanical filters contain elements that vibrate and establish resonance mechanically. In crystal filters the coupling between filter sections is achieved by electrical means. In mechanical filters, mechanical couplers are used to transfer the vibrations from one resonant section to the next. At the input and output ends of the filter are transducers which provide for electrical coupling to and from the filter. Most mechanical filters are designed for use from 200 to 600 kHz, the range near 455 kHz being the most popular for amateur use. Mechanical filters suitable for amateur radio circuits are manufactured by the Collins Radio Co. and can be purchased from some dealers in amateur radio equipment.

[1] McCoy Electronics Company, Mt. Holly Springs, PA.

[2] Spectrum International, Topsfield, MA.

[3] E. S. Electronic Labs, 31 Augustus, Excelsior Springs, MO.

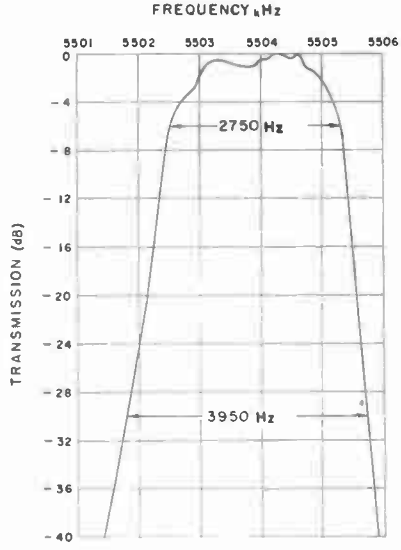


Fig. 13-8 — Measured selectivity characteristic of the filter when aligned at 5505.2 kHz. The 6-dB bandwidth is 2750 Hz and the 30-dB/6-dB shape factor is 1.44.

FILTER APPLICATIONS

Methods for using typical sideband filters are shown schematically in Fig. 13-9. In the circuit of

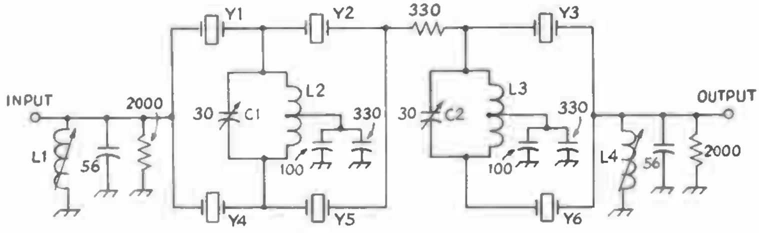


Fig. 13-7 — Circuit diagram of a filter. Resistances are in ohms, and resistors are 1/2-watt composition; capacitors are disk ceramic except as noted.

C1, C2 — Mica trimmer.
L1, L4 — 50 turns No. 38 enamel, close-wound on 17/64-inch dia ceramic slug-tuned form (CTC LS-6, National XR-81 or similar).
L2, L3 — 60 turns No. 38 enamel, close wound on 17/64-inch ceramic form (CTC LS-6, National XR-81 or similar with powered-iron core removed), center tapped.
Y1, Y2, Y3 — All same frequency (near 5500 kHz).
Y4, Y5, Y6 — All same frequency and 1500 to 1700 Hz different from Y1, Y2, Y3.

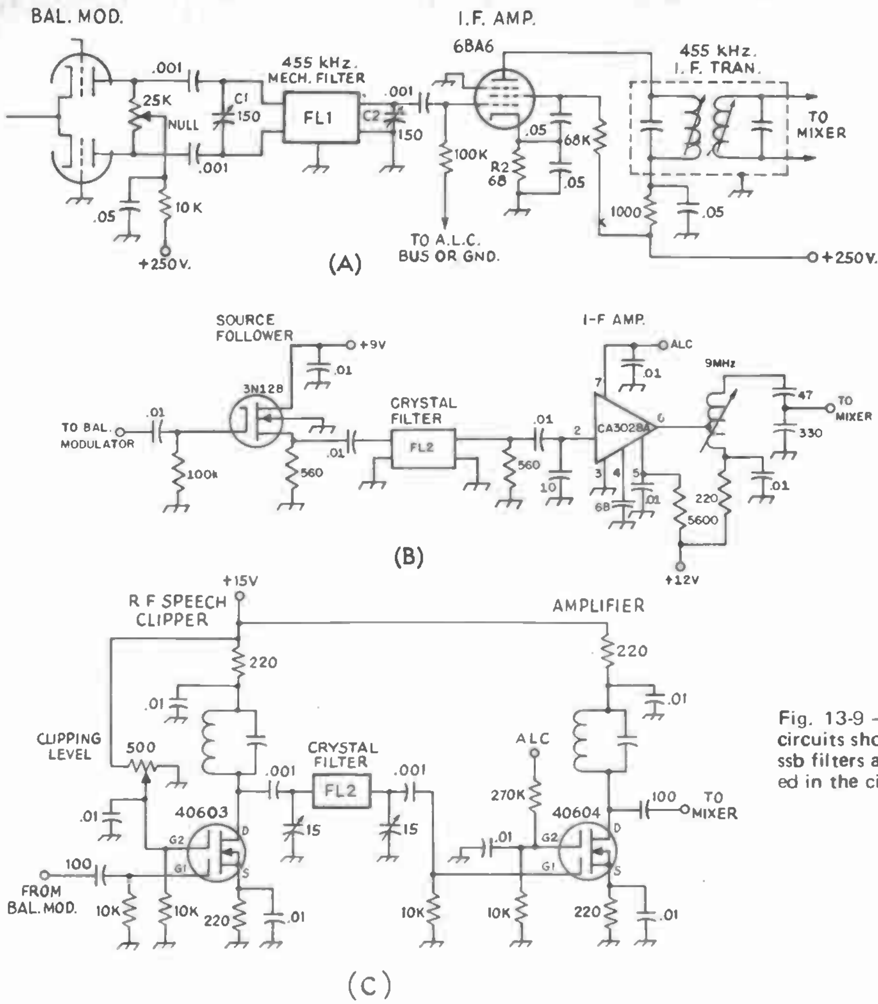


Fig. 13-9 — Typical circuits showing how ssb filters are connected in the circuit.

Fig. 13-9A a 455-kHz mechanical filter is coupled to the balanced modulator by means of two dc isolating capacitors. C1 is used to tune the input of FL1 to resonance (if a Collins type 455-FB-21 is used). Frequently, a fixed-value 120-pF capacitor will suffice at each end of the filter. C2 tunes the output of the filter. A stage of i-f amplification usually follows the filter, as shown, to compensate for the insertion loss of the filter and to provide a stage to which agc can be applied for alc (automatic level control) purposes. In the circuit shown the operator can ground R1 if alc is not used. R2 can be lifted from ground and a 5000-ohm control can be placed between it and ground to provide a means of manual gain control for providing the desired signal level to the mixer.

The circuit of Fig. 13-9B uses a 9-MHz crystal filter, followed by an IC i-f amplifier. Either the McCoy or Spectrum International filters are suitable. Most commercial ssb filters are supplied with a data sheet which shows recommended input and output circuits for matching the impedance of the filter. All are adaptable to use with tubes or transistors.

Another circuit which uses an hf crystal filter, preceded by a dual-gate MOSFET operating as an rf speech clipper, is shown in Fig. 13-9C. The advantages of rf clipping are explained later in this chapter. A second MOSFET amplifies the signal from the filter and provides a variable level of output which is controlled by the alc line.

CARRIER OSCILLATOR

The ssb-generation process starts with a crystal-controlled oscillator, as shown in Fig. 13-2. In a filter-type generator, the oscillator frequency is set on the low-frequency side of the filter bandpass to produce upper sideband and on the upper side when lower-sideband operation is desired. Suitable oscillator circuits are shown in Chapter 6.

MIXER

A single-sideband signal, unlike fm or cw, cannot be frequency multiplied. One or more mixer stages are employed in an ssb exciter to

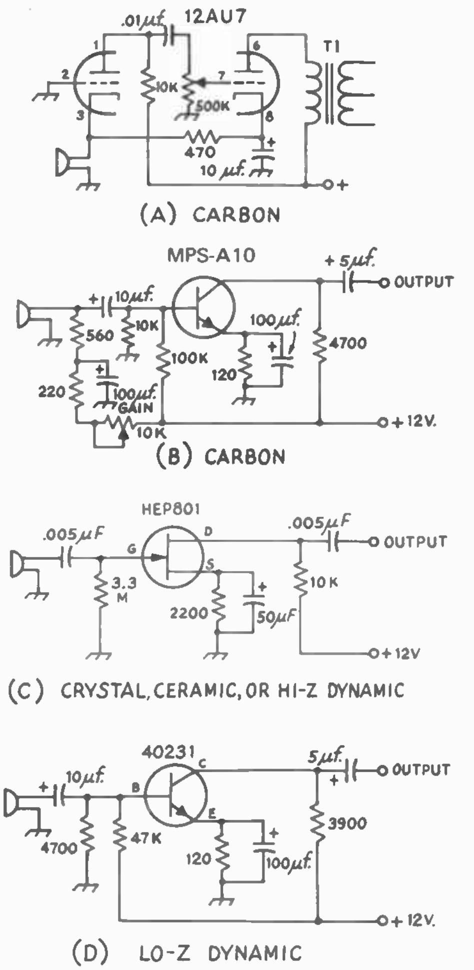


Fig. 13-10 – Speech circuits for use with standard-type microphones. Typical parts values are given.

heterodyne the output of a fixed-frequency ssb generator to the desired operating frequency. See Chapter 8 for details of mixer design and sample mixer circuits.

THE SPEECH AMPLIFIER

The purpose of a speech amplifier is to raise the level of audio output from a microphone to that required by the modulator of a transmitter. In ssb and fm transmitters the modulation process takes place at low levels, so only a few volts of audio are necessary. One or two simple voltage-amplifier stages will suffice. A-m transmitters often employ high-level plate modulation requiring considerable audio power, as described in Chapter 12. The microphone-input and audio voltage-amplifier circuits are similar in all three types of phone transmitters, however.

When designing speech equipment it is necessary to know (1) the amount of audio power the modulation system must furnish and (2) the output voltage developed by the microphone when it is spoken into from normal distance (a few inches) with ordinary loudness. It then becomes possible to choose the number and type of amplifier stages needed to generate the required audio power without overloading or undue distortion anywhere in the system.

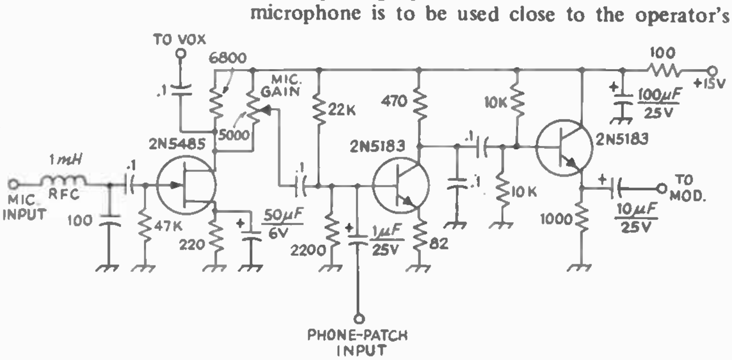
MICROPHONES

The level of a microphone is its electrical output for a given sound intensity. The level varies somewhat with the type. It depends to a large extent on the distance from the sound source and the intensity of the speaker's voice. Most commercial transmitters are designed for the median level. If a high-level mike is used, care should be taken not to overload the input amplifier stage. Conversely, a microphone of too low a level must be boosted by a preamplifier.

The frequency response (fidelity) of a microphone is its ability to convert sound uniformly into alternating current. For high articulation it is desirable to reproduce a frequency range of 200-3500 Hz. When all frequencies are reproduced equally, the microphone is considered "flat." Flat response is highly desirable as peaks (sharp rises in the reproduction curve) limit the swing or modulation to the maximum drive voltage, whereas the usable energy is contained in the flat part of the curve.

Microphones are generally omnidirectional, and respond to sound from all directions, or unidirectional, picking up sound from one direction. If a microphone is to be used close to the operator's

Fig. 13-11 – A resistance-coupled speech amplifier. Component values are representative of a typical circuit.



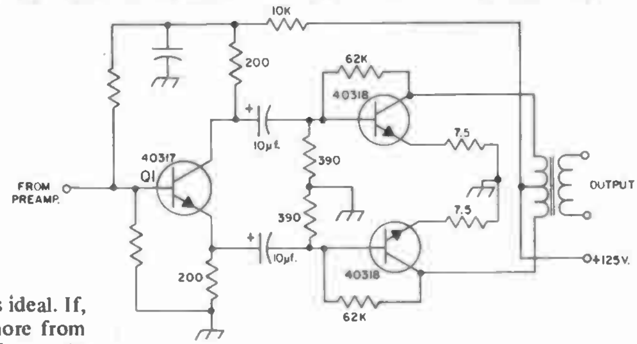


Fig. 13-12 — Typical phase-inverter circuits for transistor amplifier applications.

mouth, an omnidirectional microphone is ideal. If, however, speech is generated a foot or more from the microphone, a unidirectional microphone will reduce reverberation by a factor of 1.7:1. Some types of unidirectional microphones have proximity effect in that low frequencies are accentuated when the microphone is too close to the mouth.

Carbon Microphones

The carbon microphone consists of a metal diaphragm placed against a cup of loosely packed carbon granules. As the diaphragm is actuated by the sound pressure, it alternately compresses and decompresses the granules. When current is flowing through the button, a variable dc will correspond to the movement of the diaphragm. This fluctuating dc can be used to provide grid-cathode voltage corresponding to the sound pressure.

The output of a carbon microphone is extremely high, but nonlinear distortion and instability has reduced its use. The circuit shown in Fig. 13-10 will deliver 20-30 volts at the transformer secondary.

Piezoelectric Microphones

Piezoelectric microphones make use of the phenomena by which certain materials produce a voltage by mechanical stress or distortion of the material. A diaphragm is coupled to a small bar of material such as Rochelle salt or ceramic made of barium titanate or lead zirconium titanate. The diaphragm motion is thus translated into electrical energy. Rochelle-salt crystals are susceptible to high temperatures, excessive moisture, or extreme dryness. Although the output level is higher, their use is declining because of their fragility.

Ceramic microphones are impervious to temperature and humidity. The output level is adequate for most modern amplifiers. They are capacitive devices and the output impedance is high. The load impedance will affect the low frequencies. To provide attenuation, it is desirable to reduce the load to 0.25 megohm or even lower, to maximize performance when operating ssb, thus eliminating much of the unwanted low-frequency response.

Dynamic Microphones

The dynamic microphone somewhat resembles a dynamic loudspeaker. A lightweight coil, usually made of aluminum wire, is attached to a diaphragm. This coil is suspended in a magnetic circuit. When sound impinges on the diaphragm, it moves the coil through the magnetic field generating an alternating voltage.

Electret Microphones

The electret microphone has recently appeared as a feasable alternative to the carbon, piezoelectric or dynamic microphone. An electret is an insulator which has a quasi-permanent static electric charge trapped in or upon it. The electret operates in a condenser fashion which uses a set of biased plates whose motion, caused by air pressure variations, creates a changing capacitance and accompanying change in voltage. The electret acts as the plates would, and being charged, it requires no bias voltage. A low voltage provided by a battery used for an FET impedance converter is the only power required to produce an audio signal.

Electrets traditionally have been susceptable to damage from high temperatures and high humidity. New materials and different charging techniques have lowered the chances of damage, however. Only in extreme conditions (such as 120 degrees F at 90 percent humidity) are problems present. The output level of a typical electret is higher than that of a standard dynamic microphone.

VOLTAGE AMPLIFIERS

The important characteristics of a voltage amplifier are its voltage gain, maximum undistorted output voltage, and its frequency response. The voltage gain is the voltage-amplification ratio of the stage. The output voltage is the maximum af voltage that can be secured from the stage without distortion. The amplifier frequency response should be adequate for voice reproduction; this requirement is easily satisfied.

The voltage gain and maximum undistorted output voltage depend on the operating conditions of the amplifier. The output voltage is in terms of *peak* voltage rather than rms; this makes the rating independent of the waveform. Exceeding the peak value causes the amplifier to distort, so it is more useful to consider only peak values in working with amplifiers.

Resistance Coupling

Resistance coupling generally is used in voltage-amplifier stages. It is relatively inexpensive, good frequency response can be secured, and there

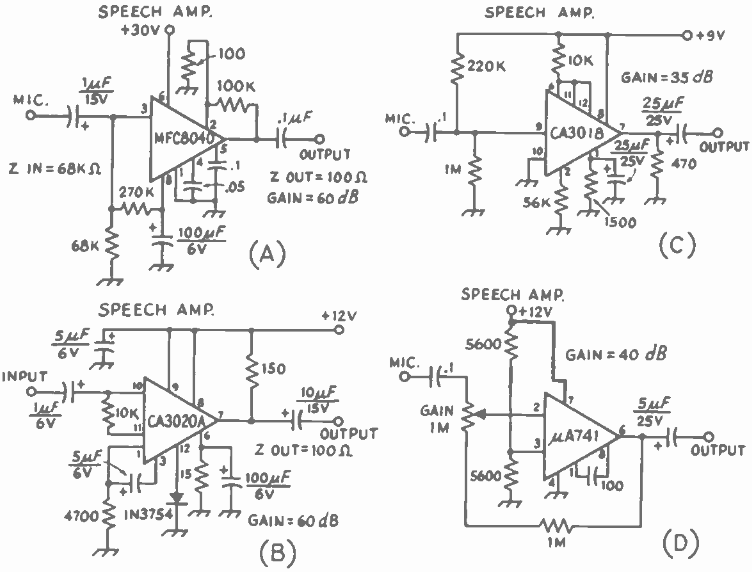


Fig. 13-13 — Typical speech amplifier using integrated circuits.

is little danger of hum pick-up from stray magnetic fields. It is the most satisfactory type of coupling for the output circuits of pentodes and high-μ triodes, because with transformers a sufficiently high load impedance cannot be obtained without considerable frequency distortion. A typical circuit is given in Fig. 13-11.

Phase Inversion

Push-pull output may be secured with resistance coupling by using phase-inverter or phase-splitter circuits as shown in Fig. 13-12. In this circuit the voltage developed across the emitter resistor of Q1 is equal to, but 180 degrees out of phase with, the voltage swing across the collector resistor. Thus, the following two stages are fed equal af voltages. The gain of Q1 will be quite low, if indeed the stage exhibits any gain at all.

Transformer Coupling

Transformer coupling between stages ordinarily is used only when power is to be transferred (in such a case resistance coupling is very inefficient), or when it is necessary to couple between a single-ended and a push-pull stage.

Several types of ICs have been developed for use in speech amplifiers. The Motorola MFC8040 features very low noise, typically 1 μV, (Fig. 13-13A), while the RCA CA3020 has sufficient power output ~ 500 mW – to drive low-impedance loads (Fig. 13-13B). A transistor IC array can also be put to work in a speech amplifier, as shown in Fig. 13-13C. This circuit uses an RCA CA3018, with a high-gain Darlington pair providing high gain and high input impedance. A second transistor within the IC functions as an emitter follower, for low-impedance output. Most of the operational-amplifier ICs will work as high-gain speech amplifiers, using a minimum of external parts as shown in Fig. 13-13D. The μA741 has internal frequency compensation, but the popular 709 series of operational amplifiers requires external frequency compensation to prevent self-oscillation.

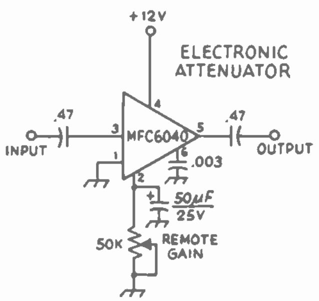


Fig. 13-14 — A dc voltage controls the gain of this IC, eliminating the need for shielded leads to the gain control.

Gain Control

A means for varying the over-all gain of the amplifier is necessary for keeping the final output at the proper level for modulating the transmitter. The common method of gain control is to adjust the value of ac voltage applied to the base or grid of one of the amplifiers by means of a voltage divider or potentiometer.

The gain-control potentiometer should be near the input end of the amplifier, at a point where the signal voltage level is so low there is no danger that the stages ahead of the gain control will be overloaded by the full microphone output. In a high-gain amplifier it is best to operate the first stage at maximum gain, since this gives the best signal-to-hum ratio. The control is usually placed in the input circuit of the second stage.

Remote gain control can also be accomplished with an electronic attenuator IC, such as the Motorola MFC6040. A dc voltage varies the gain of the IC from +6 dB to -85 dB, eliminating the need

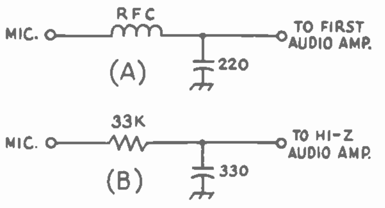


Fig. 13-15 – Rf filters using *LC* (A) and *RC* (B) components, which are used to prevent feedback caused by rf pickup on the microphone lead.

for shielded leads to a remotely located volume control. A typical circuit is shown in Fig. 13-14.

Speech-Amplifier Construction

Once a suitable circuit has been selected for a speech amplifier, the construction problem resolves itself into avoiding two difficulties – excessive hum, and unwanted feedback. For reasonably humless operation, the hum voltage should not exceed about 1 percent of the maximum audio output voltage – that is, the hum and noise should be at least 40 dB below the output level.

Unwanted feedback, if negative, will reduce the gain below the calculated value; if positive, is likely to cause self-oscillation or "howls." Feedback can be minimized by isolating each stage with decoupling resistors and capacitors, by avoiding layouts that bring the first and last stages near each other, and by shielding of "hot" points in the circuit, such as high-impedance leads in low-level stages.

If circuit-board construction is used, high-impedance leads should be kept as short as possible. All ground returns should be made to a common point. A good ground between the circuit board and the metal chassis is necessary. Complete shielding from rf energy is always required for low-level solid-state audio circuits. The microphone input should be decoupled for rf with a filter, as shown in Fig. 13-15. At A, an rf choke with a high impedance over the frequency range of the transmitter is employed. For high-impedance inputs, a resistor may be used in place of the choke.

When using paper capacitors as bypasses, be sure that the terminal marked "outside foil," often indicated with a black band, is connected to ground. This utilizes the outside foil of the capacitor as a shield around the "hot" foil. When paper or mylar capacitors are used for coupling between stages, always connect the outside foil terminal to the side of the circuit having the lower impedance to ground.

DRIVER AND OUTPUT STAGES

Few ssb transmitting mixers have sufficient output to properly drive an output stage of any significant power level. Most modern-day linear amplifiers require at least 30 to 100 watts of exciter output power to drive them to their rated power input level. It follows, then, that an intermediate stage of amplification should be used between the mixer and the pa stage of the exciter.

The vacuum-tube mixers of Chapter 8 will provide 3 to 4 peak volts of output into a high-impedance load. Since most AB_1 exciter output stages need from 25 to 50 volts of swing on their grids for normal operation, it is necessary to employ a driver stage to amplify the mixer output. There are several high-transconductance pentode tubes that work well as drivers. Among them are the 6CL6, the 12BY7, the 6EH7, and the 6GK6. Since all of these tubes are capable of high gain, instability is sometimes encountered during their use. Parasitic suppression should be included as a matter of course, and can take the form of a low-value noninductive resistor in series with the grid, or a standard parasitic choke installed directly at the plate of the tube. Some form of neutralization is recommended and is preferred to resistive loading of the tuned circuits. The latter method lowers the tuned-circuit *Q*. This in turn lowers the stage selectivity and permits spurious responses from the mixer to be passed on to the following stage of the exciter.

A typical driver and PA stage for modern exciters is shown in Fig. 13-16. The PA is set up for AB_1 amplification. The AB_1 mode is preferred because it results in less distortion than does the AB_2 or Class-B modes, and because driving power is not needed for AB_1 operation. A 6146 tube is used but an inexpensive TV sweep tube may be employed if a higher level of IMD is permissible.

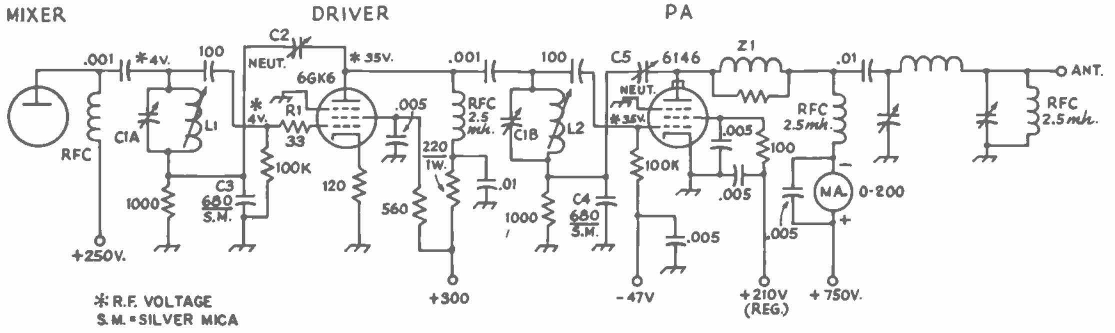


Fig. 13-16 – Schematic diagram of a typical driver and final stage for ssb exciter. Neutralization and parasitic-suppression circuits have been included.

Some sweep tubes are capable of producing less IMD than others, but if not overdriven most of them are satisfactory for ham use. Among the sweep tubes useful as AB_1 amplifiers are the following: 6DQ5, 6GB5, 6GE5, 6HF5, 6JE6, 6JS6, 6KD6, 6KG6, 6LF6 and 6LQ6.

A Practical Circuit

In the circuit of Fig. 13-16, a 6GK6 and a 6146 are shown in a typical driver-amplifier arrangement. Each stage is stabilized by means of R1 in the driver grid, and Z1 in the PA plate, both for parasitic suppression. C2 and C5 are neutralizing capacitors and can take the form of stiff wires placed adjacent to, and in the same plane as the tube anode. Varying the spacing between the neutralizing stubs and the tube envelopes provides the adjustment of these capacitors. Parallel dc feed is used in the mixer and driver stages to prevent the tuned-circuit *Q* from being lowered by dc current flow through L1 and L2. C1A and C1B are ganged, and slug-tuned inductors are used at L1 and L2 to permit tracking of the mixer and driver plate tanks. C3 and C4 form part of the neutralizing circuits. The values shown are suitable for operation on 3.5 MHz but may require modification for use on the other bands. Regulated dc voltage is recommended for the screen grids of the driver and rf stages. Typical rf voltages (measured with a diode rf probe and VTVM are identified with an asterisk. A circuit of this type is capable of up to 60 watts PEP output. For more information on linear amplifiers for sideband service, see Chapter 6.

POWER RATINGS OF SSB TRANSMITTERS

Fig. 13-17 is more or less typical of a few voice-frequency cycles of the modulation envelope of a single-sideband signal. Two amplitude values associated with it are of particular interest. One is the *maximum peak amplitude*, the greatest amplitude reached by the envelope at any time. The other is the *average amplitude*, which is the average of all the amplitude values contained in the envelope over some significant period of time, such as the time of one syllable of speech.

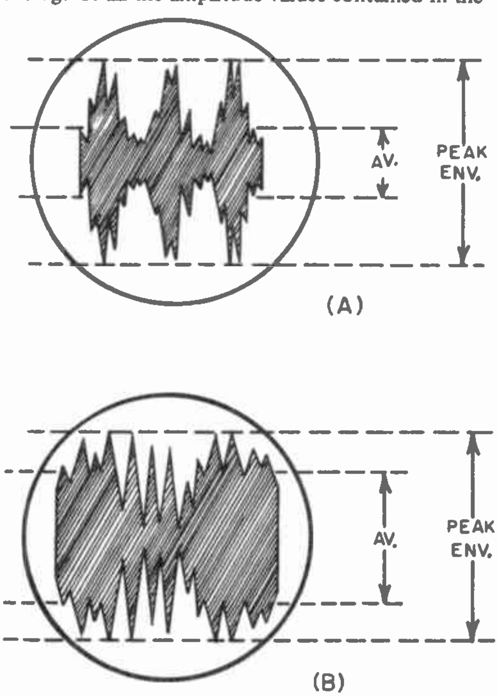


Fig. 13-17 — (A) Typical ssb voice-modulated signal might have an envelope of the general nature shown, where the rf amplitude (current or voltage) is plotted as a function of time, which increases to the right horizontally. (B) Envelope pattern after speech processing to increase the average level of power output.

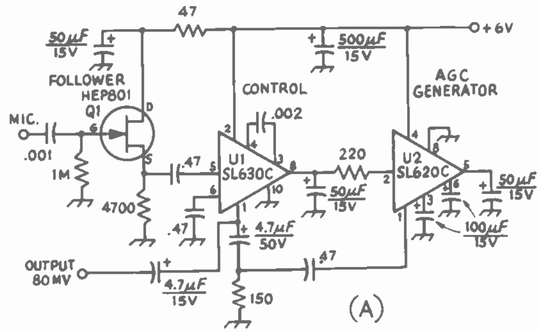
The power contained in the signal at the maximum peak amplitude is the basic transmitter rating. It is called the *peak-envelope power*, abbreviated PEP. The peak-envelope power of a given transmitter is intimately related to the distortion considered tolerable. The lower the signal-to-distortion ratio the lower the attainable peak-envelope power, as a general rule. For splatter reduction, an *S/D* ratio of 25 dB is considered a border-line minimum, and higher figures are desirable.

The signal power, *S*, in the standard definition of *S/D* ratio is the power in *one* tone of a two-tone test signal. This is 3 dB below the peak-envelope power in the same signal. Manufacturers of amateur ssb equipment usually base their published *S/D* ratios on PEP, thereby getting an *S/D* ratio that looks 3 dB better than one based on the standard definition. In comparing distortion-product ratings of different transmitters or amplifiers, first make sure that the ratios have the same base.

When the output of an ssb transmitter is viewed on a spectrum analyzer, the display shows the power in the two tones separately, so that the level of distortion products is 6 dB below the level of either tone. However, commercial analyzers usually have a scale over the display tube which is calibrated directly in dB below a single-tone test. Readings may be converted to dB below the PEP level by subtracting 6 dB from the indicated distortion levels.

Peak vs. Average Power

Envelope peaks occur only sporadically during voice transmission, and have no direct relationship with meter readings. The meters respond to the amplitude (current or voltage) of the signal averaged over several cycles of the modulation envelope. (This is true in practically all cases, even though the transmitter rf output meter may be *calibrated* in watts. Unfortunately, such a calibra-



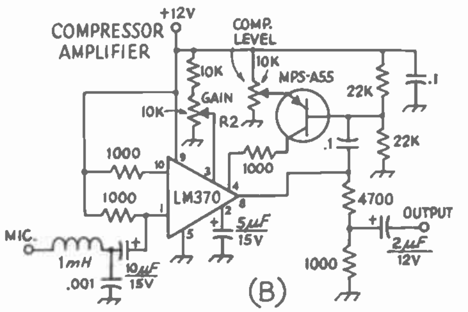


Fig. 13-18 – Typical solid-state compressor circuit.

tion means little in voice transmission since the meter can be calibrated in watts only by using a sine-wave signal – which a voice-modulated signal definitely is not.)

The ratio of peak-to-average amplitude varies widely with voice of different characteristics. In the case shown in Fig. 13-17A the average amplitude, found graphically, is such that the peak-to-average ratio of amplitudes is almost 3 to 1. The ratio of peak *power* to average *power* is something else again. There is no simple relationship between the meter reading and actual average power, for the reason mentioned earlier.

DC Input

FCC regulations require that the transmitter power be rated in terms of the dc input to the final stage. Most ssb final amplifiers are operated Class AB_1 or AB_2, so that the plate current during modulation varies upward from a "resting" or no-signal value that is generally chosen to minimize distortion. There will be a peak-envelope value of plate current that, when multiplied by the dc plate voltage, represents the instantaneous tube power input required to produce the peak-envelope output. This is the "peak-envelope dc input" or "PEP input." It does not register on any meter in the transmitter. Meters cannot move fast enough to show it – and even if they did, the eye couldn't follow. What the plate meter *does* read is the plate current averaged over several modulation-envelope cycles. This multiplied by the dc plate voltage is the number of watts input required to produce the *average* power output described earlier.

In voice transmission the power input and power output are both continually varying. The power input peak-to-average ratio, like the power-output peak-to-average ratio, depends on the voice characteristics. Determination of the input ratio is further complicated by the fact that there is a resting value of dc plate input even when there is no rf output. *No exact figues are possible.* However, experience has shown that for many types of voices and for ordinary tube operating conditions where a moderate value of resting current is used, the ratio of PEP input to average input (during a modulation peak) will be in the neighborhood of 2 to 1. That is why many amplifiers are rated for a PEP input of 2 kilowatts even though the maximum legal input is 1 kilowatt.

PEP Input

The 2-kilowatt PEP input rating can be interpreted in this way: The amplifier can handle dc peak-envelope inputs of 2 kw, presumably with satisfactory linearity. But it should be run up to such peaks if – and *only* if – in doing so the dc plate current (the current that shows on the plate meter) multiplied by the dc plate voltage does not at any time exceed 1 kilowatt. On the other hand, if your voice has characteristics such that the dc peak-to-average ratio is, for example, 3 to 1, you should not run a greater dc input during peaks than 2000/3, or 660 watts. Higher dc input would drive the amplifier into nonlinearity and generate splatter.

If your voice happens to have a peak-to-average ratio of less than 2 to 1 with this particular amplifier, you cannot run more than 1 kilowatt dc input even though the envelope peaks do not reach 2 kilowatts.

It should be apparent that the dc input rating (based on the *maximum* value of dc input developed during modulation, of course) leaves much to be desired. Its principal virtues are that it can be measured with ordinary instruments, and that it is consistent with the method used for rating the power of other types of emission used by amateurs. The meter readings offer no assurance that the transmitter is being operated within linearity limits, unless backed up by oscilloscope checks using *your* voice.

It should be observed, also, that in the case of a grounded-grid final amplifier, the 1-kilowatt dc input permitted by FCC regulations must include the input to the driver stage as well as the input to the final amplifier itself. Both inputs are measured as described above.

SPEECH PROCESSING

Four basic systems, or a combination thereof, can be used to reduce the peak-to-average ratio, and thus, to raise the average power level of an ssb signal. They are: compression or clipping of the af wave before it reaches the balanced modulator, and compression or clipping of the rf waveform after the ssb signal has been generated. One form of rf compression, commonly called alc (automatic level control) is almost universally used in amateur ssb transmitters. Audio processing is also used to increase the level of audio power contained in the sidebands of an a-m transmitter and to maintain constant deviation in an fm transmitter. Both compression and clipping are used in a-m systems, while most fm transmitters employ only clipping.

Volume Compression

Although it is obviously desirable to keep the voice level as high as possible, it is difficult to maintain constant voice intensity when speaking into the microphone. To overcome this variable output level, it is possible to use automatic gain control that follows the *average* (not instantaneous) variations in speech amplitude. This can be done by rectifying and filtering some of the audio output and applying the rectified and filtered dc to a control electrode in an early stage in the amplifier.

A practical example of an audio compressor circuit is shown in Fig. 13-18A. Q1 is employed as an impedance converter, providing coupling between a high-impedance microphone and the input terminal of the Plessey SL630C audio-amplifier IC. Low-impedance microphones can be connected directly to the input of the SL630C. U1 has an agc terminal which allows logarithmic control of the output level with a variable dc voltage. High-frequency cutoff is accomplished by connecting a .002-μF capacitor between pins 3 and 4. Manual gain control is effected by applying a dc voltage to pin 8.

Agc voltage for U1 is developed by the SL620C. A suitable time constant for voice operation is established by the capacitors connected to pins 3, 4 and 6, respectively. The IC provides a fast-attack, slow-decay characteristic for the agc voltage when voice signals are applied and a short burst of agc voltage when a short noise burst occurs. Twenty transistors and four diodes are used in U2.

The compressor will hold the output level constant within 2 dB over a 40-dB range of input signal. The nominal output level is 80 mV; the microphone used should develop at least 3 mV at the gate of Q1.

Fig. 13-18B shows an IC audio compressor circuit using the National Semiconductor LM-370. This IC has two gain-control points, pins 3 and 4; one is used for the input gain adjustment while the other receives agc voltage whenever the output level exceeds a preset norm. R2 establishes the point at which compression starts.

Speech Clipping and Filtering

In speech wave forms the average power content is considerably less than in a sine wave of the same peak amplitude. If the low-energy peaks are clipped off, the remaining wave form will have a considerably higher ratio of average power to peak amplitude. Although clipping distorts the wave form and the result therefore does not sound exactly like the original, it is possible to secure a worthwhile increase in audio power without sacrificing intelligibility. Once the system is properly adjusted *it will be impossible to overdrive* the modulator stage of the transmitter because the maximum output amplitude is fixed.

By itself, clipping generates high-order harmonics and therefore will cause splatter. To prevent this, the audio frequencies above those

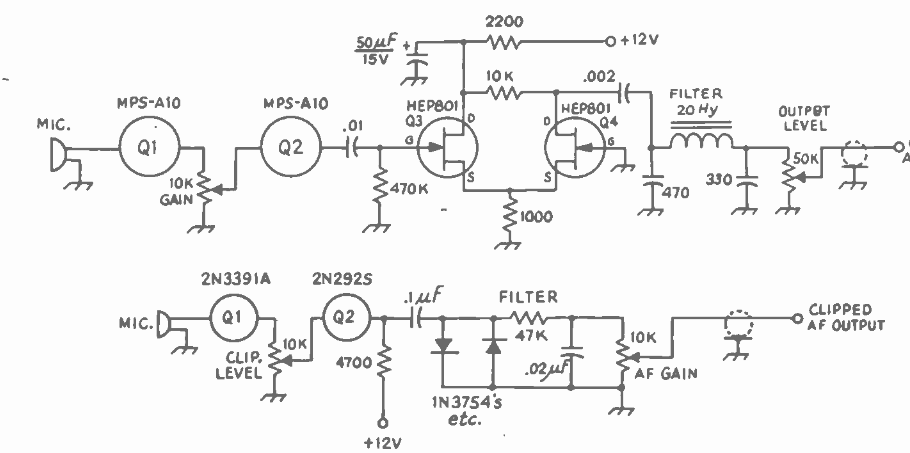


Fig. 13-19 — This drawing illustrates use of JFETs or silicon diodes to clip positive and negative voice peaks.

needed for intelligible speech must be filtered out, *after* clipping and *before* modulation. The filter required for this purpose should have relatively little attenuation below about 2500 Hz, but high attentuation for all frequencies above 3000 Hz.

The values of L and C should be chosen to form a low-pass filter section having a cutoff frequency of about 2500 Hz, using the value of the terminating resistor load resistance. For this cutoff frequency the formulas are:

$$L1 = \frac{R}{7850} \text{ and } C1 = C2 = \frac{63.6}{R}$$

where R is in ohms, L1 in henrys, and C1 and C2 in microfarads.

There is a loss in naturalness with "deep" clipping, even though the voice is highly intelligible. With moderate clipping levels (6 to 12 dB) there is almost no change in "quality" but the voice power is increased considerably.

Before drastic clipping can be used, the speech signal must be amplified several times more than is necessary for normal modulation. Also, the hum and noise must be much lower than the tolerable level in ordinary amplification, because the noise in the output of the amplifier increases in proportion to the gain.

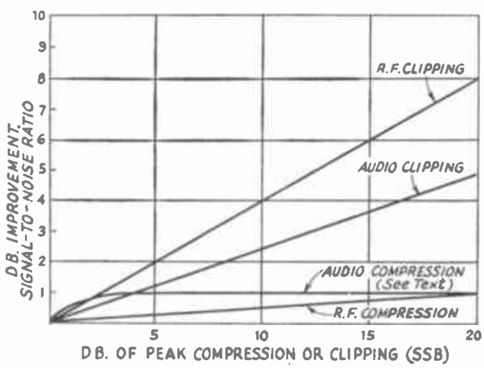


Fig. 13-20 — The improvement in received signal-to-noise ratio achieved by the simple forms of signal processing.

In the circuit of Fig. 13-19B a simple diode clipper is shown following a two-transistor preamplifier section. The 1N3754s conduct at approximately 0.7 volt of audio and provide positive- and negative-peak clipping of the speech wave form. A 47,000-ohm resistor and a .02-μF capacitor follow the clipper to form a simple R-C filter for attenuating the high-frequency components generated by the clipping action, as discussed earlier. Any top-hat or similar silicon diodes can be used in place of the 1N3754s. Germanium diodes (1N34A type) can also be used, but will clip at a slightly lower peak audio level.

SSB SPEECH PROCESSING

Compression and clipping are related, as both have fast attack times, and when the compressor release time is made quite short, the effect on the wave from approaches that of clipping. Speech processing is most effective when accomplished at radio frequencies, although a combination of af clipping and compression can produce worthwhile results. The advantage of an outboard audio speech processor is that no internal modifications are necessary to the ssb transmitter with which it will be used.

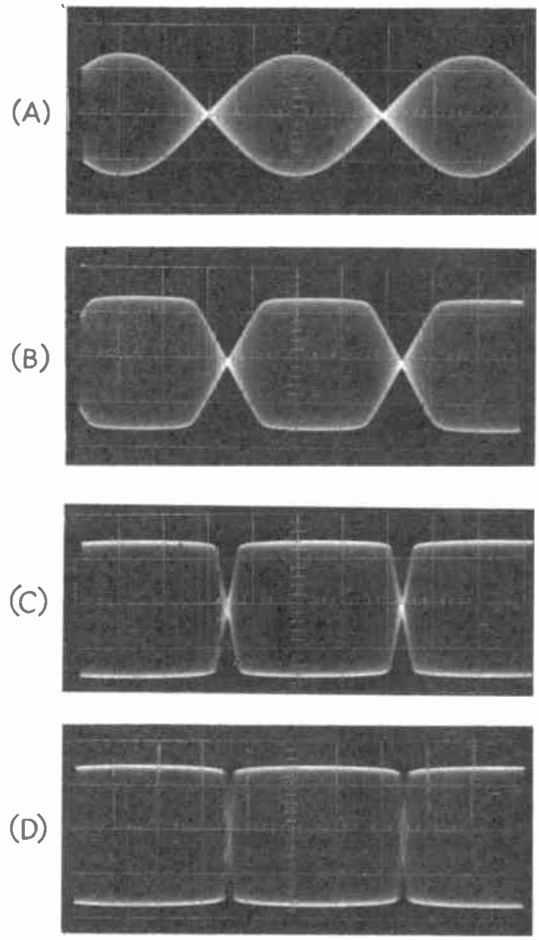


Fig. 13-21 — Two-tone envelope patterns with various degrees of rf clipping. All envelope patterns are formed using tones of 600 and 1000 Hz. (A) At clipping threshold; (B) 5 dB of clipping; (C) 10 dB of clipping; (D) 15 dB of clipping.

To understand the effect of ssb speech processing, review the basic rf waveforms shown in Fig. 13-17A. Without processing, they have high peaks but low average power. After processing, Fig. 13-17B, the amount of average power has been raised considerably. Fig. 13-20 shows an advantage of several dB for rf clipping (for 20 dB of processing) over its nearest competitor.

Investigations by W6JES reported in *QST* for January, 1969, show that, observing a transmitted signal using 15 dB of audio clipping from a remote receiver, the intelligibility threshold was improved nearly 4 dB over a signal with no clipping.

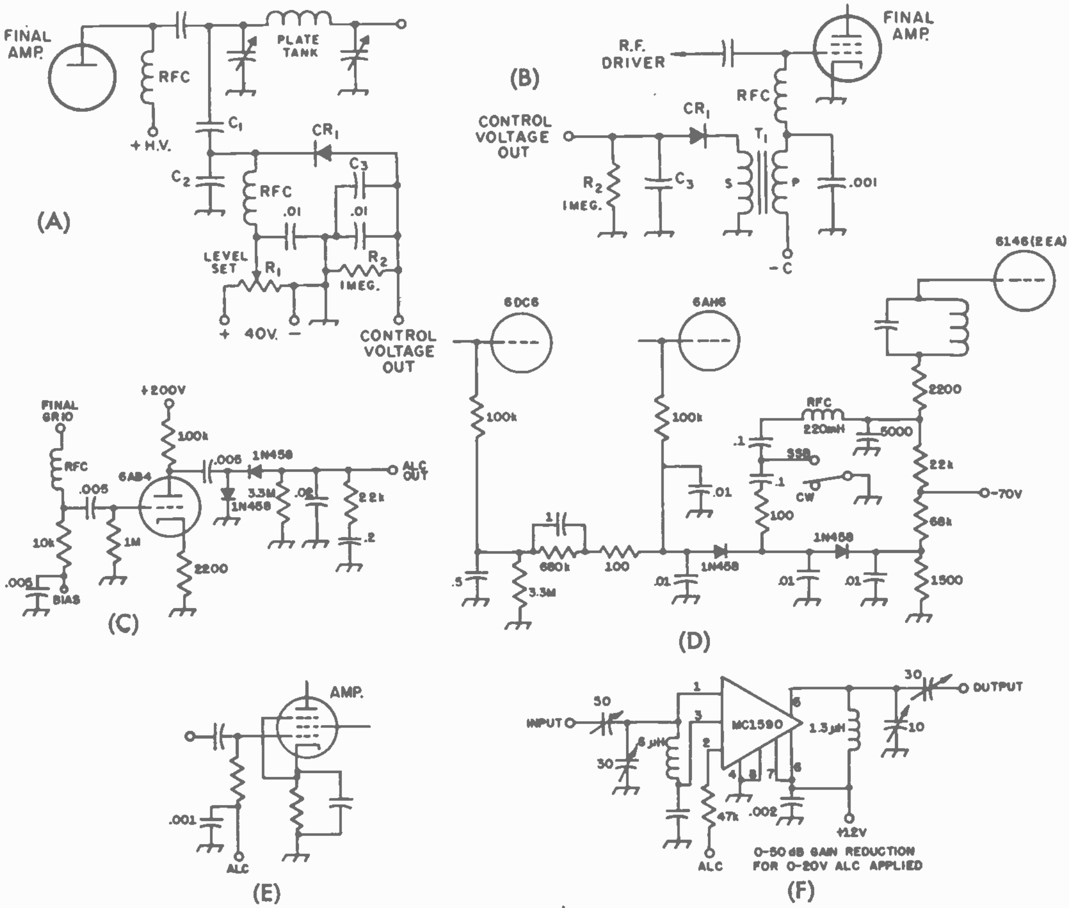


Fig. 13-22 – (A) Control voltage obtained by sampling the rf output voltage of the final amplifier. The diode back bias, 40 volts or so maximum, may be taken from any convenient positive voltage source in the transmitter. R1 may be a linear control having a maximum resistance of the order of 50,000 ohms. CR1 may be a 1N34A or similar germanium diode.
(B) Control voltage obtained from grid circuit of a Class AB_1 tetrode amplifier. T1 is an interstage audio transformer having a turns ratio, secondary to primary, of 2 or 3 to 1. An inexpensive transformer may be used since the primary and secondary currents are negligible. CR1 may be a 1N34A or similar; time constant R2C3 is discussed in the text.
(C) Control voltage is obtained from the grid of a Class AB_1 tetrode amplifier and amplified by a triode audio stage.
(D) Alc system used in the Collins 32S-3 transmitter.
(E) Applying control voltage to the tube or (F) linear IC controlled amplifier.

Increasing the af clipping level to 25 dB gave an additional 1.5 dB improvement in intelligibility. Audio compression was found to be valuable for maintaining relatively constant average-volume speech, but such a compressor added little to the intelligibility threshold at the receiver, only about 1-2 dB.

Evaluation of rf clipping from the receive side with constant-level speech, and filtering to restore the original bandwidth, resulted in an improved intelligibility threshold of 4.5 dB with 10 dB of clipping. Raising the clipping level to 18 dB gave an additional 4-dB improvement at the receiver, or 8.5-dB total increase. The improvement of the intelligibility of a weak ssb signal at a distant receiver can thus be substantially improved by rf clipping. The effect of such clipping on a two-tone test pattern is shown in Fig. 13-21.

Automatic level control, although a form of rf speech processing, has found its primary application in maintaining the peak rf output of an ssb transmitter at a relatively constant level, hopefully below the point at which the final amplifier is overdriven, when the audio input varies over a considerable range. These typical alc systems, shown in Fig. 13-22, by the nature of their design time constants offer a limited increase in transmitted average-to-PEP ratio. A value in the region of 2-5 dB is typical. An alc circuit with shorter time constants will function as an rf syllabic compressor, producing up to 6 dB improvement in the intelligibility threshold at a distant receiver. The Collins Radio Company uses an alc system with dual time constants (Fig. 13-22D) in their S/Line transmitters, and this has proven to be quite effective.

Heat is an extremely important consideration in the use of any speech processor which increases the average-to-peak power ratio. Many transmitters, in particular those using television sweep tubes, simply are not built to stand the effects of increased average input, either in the final-amplifier tube or tubes or in the power supply. If heating in the final tube is the limiting factor, adding a cooling fan may be a satisfactory answer.

AN AUDIO SPEECH PROCESSOR

Over the years, different speech processing schemes have been employed, with varying degrees of success, to raise the average-to-peak power ratio of a voice signal and improve communications effectiveness. The various methods generally fall into one of two categories – compression or clipping. Described here is a processor which represents a departure from these standard approaches. Processing is done at audio frequencies, but in a unique fashion. The unit is used between the microphone and the transmitter so that no modification to the transmitter is required.

Technical Description

Operation is a consequence of the fact that speech energy resembles an amplitude-modulated signal. The speech waveform represents multiplication of a slowly varying envelope containing energy below 100 Hz with a voice-frequency signal contained mostly between 300 and 3000 Hz. In an analagous fashion, a conventional a-m modulator multiplies an amplitude-varying low-frequency signal (the applied modulation) with a constant-amplitude higher-frequency carrier. Thus the speech waveform envelope corresponds to the a-m modulation and the voice-frequency portion to the carrier. Note that the voice carrier actually varies continuously in frequency, unlike the conventional fixed-frequency a-m case, but is constant in amplitude. The object of this speech processor is to reproduce only the carrier portion of speech. The voice envelope is separated from the voice carrier, and because their respective frequency spectrums are nonoverlapping, the envelope can be filtered leaving only the carrier. (See Fig. 1.)

To separate the envelope and carrier, the speech signal is passed through a logarithmic amplifier which performs the mathematical operation of taking logarithms. By analogy to the a-m model, this signal can be represented as the mathematical product EV, where E represents the envelope and V, the voice carrier, both of which are functions of time. Taking the logarithm produces log EV, but log EV = log E + log V (a well known property of the logarithm). The envelope and carrier components are then separated in terms of their logarithms and it is now a relatively simple matter to process the two components independently. This is something which could not be done up to this point. A high-pass filter with an appropriately chosen cutoff frequency attenuates the envelope waveform but passes the higher-frequency voice carrier. The remaining signal is log V. It goes through an inverse-logarithm amplifier which produces at its output the signal V. The result is the desired constant-amplitude voice carrier.

Circuit Description

Some additional issues arise when one tries to implement the preceding scheme. These will be considered now in a stage-by-stage operational description of the processor. The reader is referred to the block diagram given in Fig. 2 and the circuit shown in diagram Fig. 3. Speech amplifier U1 first boosts the incoming audio signal to a convenient and usable level. Before taking logarithms, however, output from U1 must be full-wave rectified to be all positive since the logarithmic amplifier works only for positive signal input. The logarithm of a number is defined only for positive numbers. U4 and U5 serve as a full-wave rectifier and precedes the logarithmic amplifier, U6 and U7. Matched silicon diodes are recommended for CR1 and CR2. If none are available, individual 1N914 diodes may be substituted. The logarithmic stage separates the voice-frequency and envelope components of the speech waveform, as described above, and the envelope is filtered by an active *RC* high-pass filter, U8.[1] A two-pole Butterworth configuration is used with the lower half-power frequency set at approximately 50 Hz. Those who are experimentally inclined may wish to try lower or higher cutoff frequencies. The expression for cutoff frequency, f_c, in terms of the filter components is:

$$f_c = \frac{1}{2\pi\sqrt{R3\,R2\,C1\,C2}}$$

[1] Because the rectification and logarithmic operations performed upon the original speech signal are nonlinear, the frequency spectrums of the actual envelope and voice carrier signals are, strictly speaking, not exactly the same as those of the signals appearing at the output of the logarithmic amplifier. The main result of these operations is to introduce additional higher-frequency components not present in the original signal. It has been determined, however, that the logarithm of the rectified speech envelope is still primarily low-frequency in nature (mostly far below 100 Hz). This is sufficient to allow the processor to operate as originally described.

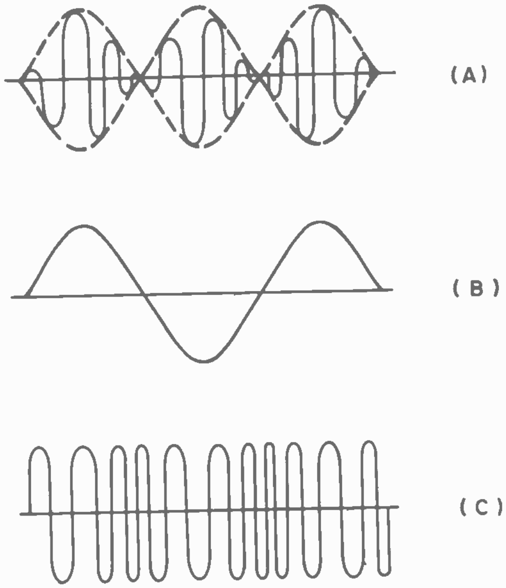


Fig. 1 — A voice signal can be represented as an a-m waveform, which results from multiplication of a relatively slowly varying envelope (B) with a carrier, (C). Note the carrier peak amplitude is constant. The speech processor separates components B and C, and filters out B, leaving the carrier portion only.

where it is required that C1 = C2 and R3 = 2 R2 for proper filter response. Varying the cutoff frequency corresponds to changing the compression level setting on a conventional speech compressor. Lower cutoff frequencies result in reduced "compression." In the original model of this processor, it was found that a filter cutoff frequency of about 400 Hz or higher produced essentially constant-amplitude output from the processor. Harmonic distortion was quite noticeable, however. Thus 50 Hz was chosen as a compromise between maximum "compression" and minimum distortion. The distortion that is inherent in this unit occurs for signals that have considerable energy in the neighborhood of the high-pass filter cutoff frequency. With a setting of 50 Hz, the distortion is quite low. The filtered signal proceeds to an exponential amplifier, U9 and U10. As with the logarithmic amplifier, matched diodes for CR3 and CR4 will produce the best results, but individual 1N914's will serve satisfactorily. The signal at the output of U10 is still in rectified form (all positive). To be converted back to its bipolar form, the signal is multiplied by the correct sign information (either positive or negative). The effect is to invert (make negative) portions of the signal which should be negative, leaving the remaining parts positive. The correct sign information is obtained by hard-limiting the voice signal at the processor input. Output from U1 is further amplified by U2 and then limited by a diode clipper, CR5 and CR6. Because of the very high gain of the U1-U2 cascade, the clipper produces almost pure square-wave output. Thus, any positive input to U1 produces a level of approximately one volt at the output of U2, and any negative input produces a level of about minus one volt. The square-wave output is multiplied with the signal from the exponential amplifier by an analog multiplier, U14. The LM1595 used at U14 produces an output voltage equal to the mathematical product of its two input signals, which in this case are the signals from U2 and U10. The result, then, is to multiply the rectified signal from the exponential amplifier by plus or minus one volt to produce the desired bipolar signal. Output is taken from buffer amplifier U11. The processed signal is passed through a low-pass filter with sharp cutoff above 3 kHz to eliminate unwanted high-frequency energy.

Because the processor is inherently sensitive to even the smallest input signals, undesired background noise or induced ac hum will be processed along with the speech and will appear as a loud disturbance at the output. To help eliminate this problem noise blanker U3 is included in the design. It consists of a free-running multivibrator with square-wave output at about 20 kHz, which is beyond audibility. When this signal is added to the output of the speech amplifier, the effect is to mask, before processing, any noise which is lower in amplitude than the 20-kHz signal.

An audio amplifier, U13, at the output, provides a convenient means of monitoring the processed audio output with low-impedance (eight-ohm) headphones. If high-impedance headphones are to be used, T1 may be omitted and output can be taken directly from pin 6 of U13 through a 5 μF coupling capacitor.

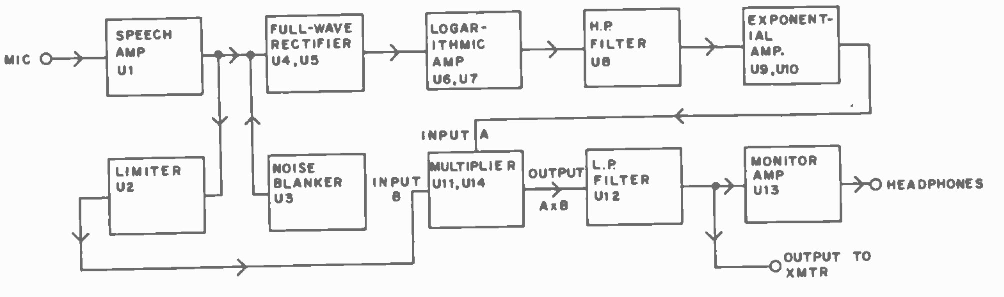


Fig. 2 — Block diagram for the processor.

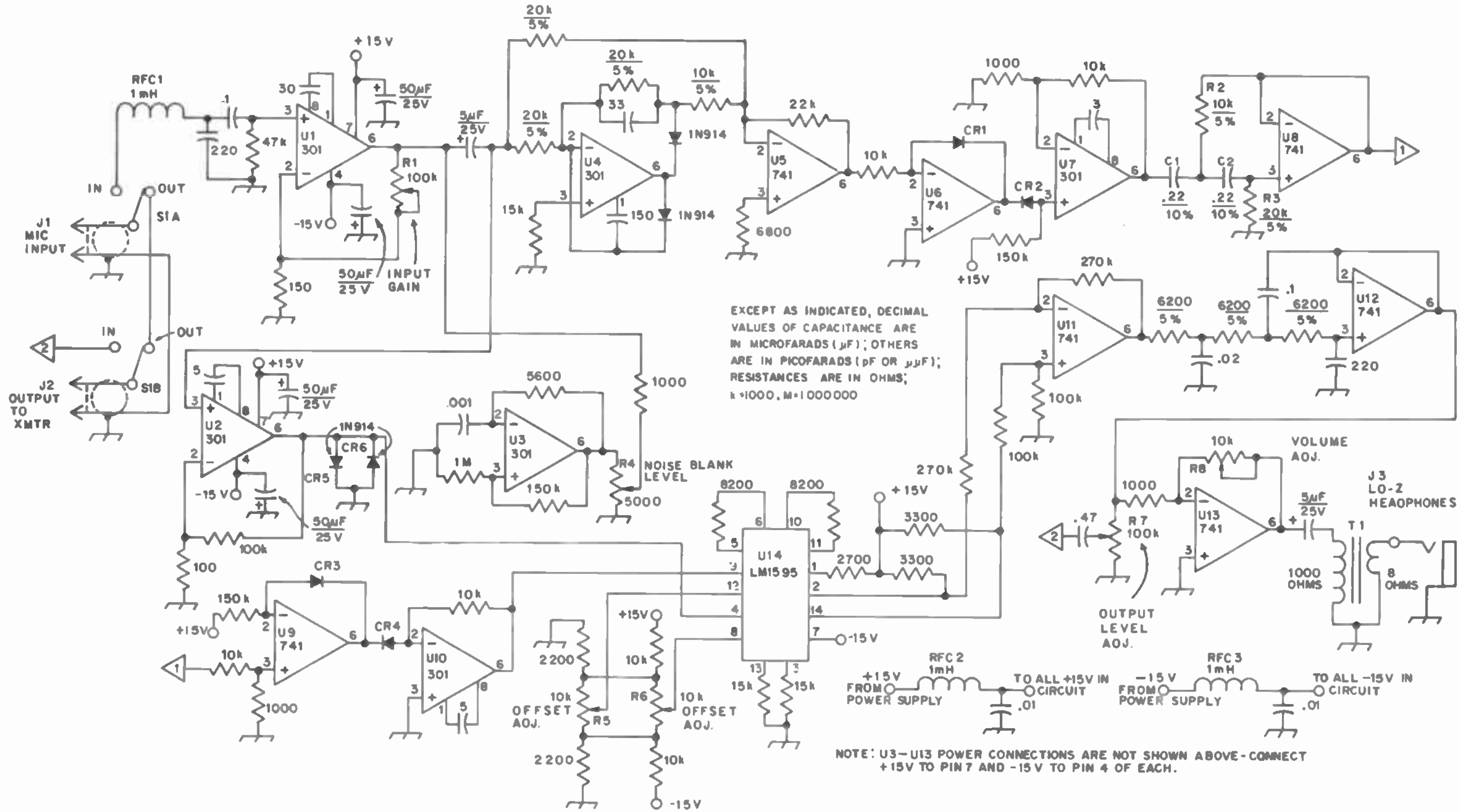


Fig. 3 — Circuit diagram for the speech processor.
CR1, CR2 — pair of matched silicon diodes (see text).
CR3, CR4 — Same as CR1, CR2.
R1, R4, R5, R6, R7 — Use circuit-board type pots.
S1 — Dpdt toggle switch.
T1 — 1000-ohm to 8-0hm audio transformer, 250 milliwatts.
U1, U5, U6, U8, U9, U11, U12, U13 — 741 op amp (Fairchild μA741, National Semiconductor LM741, Motorola MC1741, or equivalent), 8-pin mini DIP package.
U2, U3, U4, U7, U10 — National Semiconductor LM301 op amp, 8-pin mini DIP package.
U14 — National Semiconductor LM1595 (or M torola MC1595), 14-pin DIP package.

Construction Information

An etched circuit board template pattern is available from ARRL, 225 Main Street, Newington CT 06111. Please include 50 cents and a self-addressed, stamped envelope. Builders who use this layout should experience no problems. Those attempting their own layout, however, should be cautioned that because of the relatively large number of active devices, some operating with high gain, the potential for instability (oscillation) exists in a haphazard layout. Arrangement of circuit components should be generally in a straight line from input to output. The shortest possible leads should be used in all cases. Particular attention should be paid to the positions of U1 and U2 with respect to each other. Because of the very high gain the input of U1 should be kept physically as far apart as possible from the output of U2. Mounting the circuit board inside a metal chassis, such as a suitable Minibox, is recommended for rf shielding.

Procurement of parts should present no particular problems. As of the time of this writing, the 741 and LM301 operational amplifiers used in the circuit can be purchased from mail order houses for about 30 cents apiece. The LM1595 integrated circuit, probably the single most expensive item in the processor, was bought for under two dollars. Matched diodes for CR1, CR2, CR3 and CR4 cost less than a dollar.[2]

The circuit is powered by a dual dc power supply that provides plus and minus 15 volts, as is typically used with most operational amplifiers. Current consumption is approximately 50 mA from each side of the supply.

Initial Adjustments

If an oscilloscope and audio sine wave generator are available, the following alignment procedure should be followed: Set R4 to minimum resistance. Connect a microphone to J1 and the oscilloscope probe to pin 6 of U1. Adjust R1, the input gain control, while speaking into the microphone so that the voice peaks viewed on the oscilloscope are slightly below the output clipping level of U1 (approximately 14 volts peak). Remove the microphone and connect the signal generator to J1. Set the generator frequency to about 1000 Hz and adjust its output level to produce about 10 volts peak at pin 6 of U1. Place the oscilloscope probe on pin 6 of U12. Adjust offset controls R5 and R6 for the best-looking sine wave. It should be possible to produce a nearly perfect sine wave. Disconnect the generator, reconnect the microphone, and plug headphones into J3. Advancing volume control R8, one should now be able to hear himself talking, although background noise and ac hum will probably be very high. Adjust noise blanking control R4 for the desired degree of background noise suppression.

Those who do not have access to test equipment may do the following: Set R4 to the center of its range. Connect a microphone to J1 and headphones to J3. Speaking into the microphone, advance input gain control R1 and monitor volume control R8 to the point where the speech becomes audible in the headphones. Adjust offset controls R5 and R6 for minimum distortion as monitored in the headphones. The final setting of R1 is not critical. It should be high enough so that the circuit functions properly (if set too low, the audio output will sound broken up and "grainy") but not so high that the speech amplifier itself distorts the signal by clipping. Adjust R4 to suppress background noise as desired.

Finally, connect the processor output at J2 to the transmitter's microphone jack. Switch the processor out of the line by means of S1. If a Monitorscope is available to view rf output, speak into the microphone and note the level of the voice peaks. Switch the unit "in" and adjust output level control R7 for the same peak voice output level. If a monitorscope is not available, the transmitter's alc meter readings may be used. With the processor switched "out," speak into the microphone and increase the transmitter's microphone gain control until the alc meter starts to deflect. Note the peak readings. Switch the processor "in" and adjust R7 to obtain the same peak reading.

[2]One source for this item: Tri-tek, Inc., Box 14206, Phoenix, AZ 85031.

SINGLE-SIDEBAND TRANSCEIVERS

A "transceiver" combines the functions of transmitter and receiver in a single package. In contrast to a packaged "transmitter-receiver," it utilizes many of the active and passive elements for both transmitting and receiving. Ssb transceiver operation enjoys widespread popularity for several justifiable reasons. In most designs the transmissions are on the same (suppressed-carrier) frequency as the receiver is tuned to. The only practical way to carry on a rapid multiple-station "round table" or net operation is for all stations to transmit on the same frequency. Transceivers are ideal for this, since once the receiver is properly set the transmitter is also. Transceivers are by nature more compact than transmitter-receivers, and thus lend themselves well to mobile and portable use.

Although the many designs available on the market differ in detail, there are of necessity many points of similarity. All of them use the filter type of sideband generation, and the filter unit furnishes the receiver i-f selectivity as well. The carrier oscillator doubles as the receiver (fixed) BFO. One or more mixer or i-f stage or stages will be used for both transmitting and receiving. The receiver S meter may become the transmitter plate-current or output-voltage indicator. The VFO that sets the receiver frequency also determines the transmitter frequency. The same signal-frequency tuned circuits may be used for both transmission and reception, including the transmitter pi-network output circuit.

Usually the circuits are switched by a

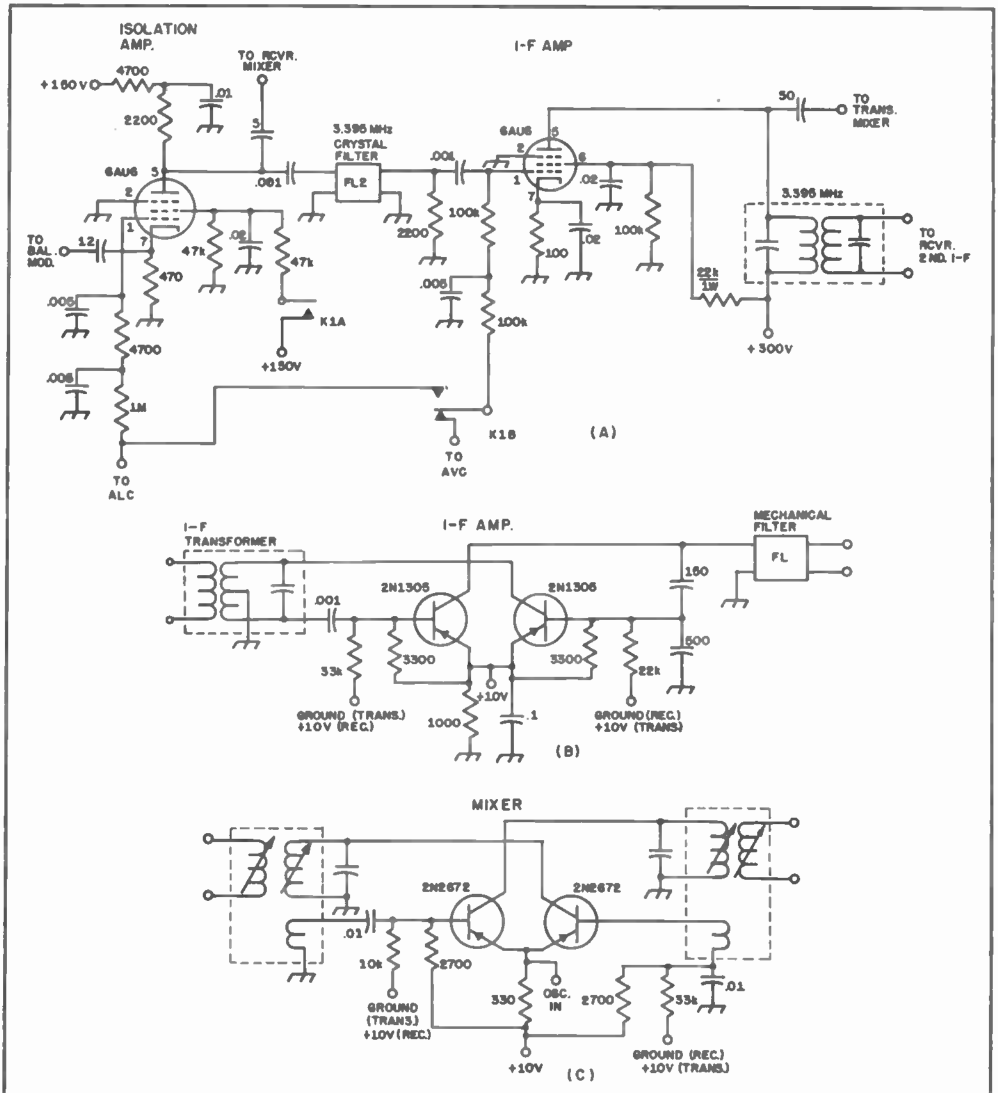


Fig. 13-28 – Transceiver circuits where a section is made to operate on both transmit and receive. See text for details.

multiple-contact relay, which transfers the antenna if necessary and also shifts the biases on several stages. Most commercial designs offer VOX (voice-controlled operation) and MOX (manual operation). Which is preferable is a controversial subject; some operators like VOX and others prefer MOX.

Circuits

The use of a filter-amplifier combination common to both the transmitter and receiver is shown in Fig. 13-28A. This circuit is used by the Heath Company in several of their transceiver kits. When receiving, the output of the hf mixer is coupled to the crystal filter, which, in turn, feeds the first i-f amplifier. The output of this stage is transformer coupled to the second i-f amplifier. During transmit, K1 is closed, turning on the isolation amplifier that links the balanced modulator to the band-pass filter. The single-sideband output from the filter is amplified and capacitance-coupled to the transmitter mixer. The relay contacts also apply alc voltage to the first i-f stage and remove the screen voltage from the second i-f amplifier, when transmitting.

Bilateral amplifier and mixer stages, first used by Sideband Engineers in their SBE-33, also have found application in other transceiver designs. The circuits shown in Fig. 13-28B and C are made to work in either direction by grounding the bias divider of the input transistor, completing the bias network. The application of these designs to an amateur transceiver for the 80-10 meter bands is given in the 5th Edition of *Single Sideband for the Radio Amateur.*

The complexity of a multiband ssb transceiver is such that most amateurs buy them fully built and tested. There are, however, some excellent designs available in the kit field, and any amateur able to handle a soldering iron and follow instructions can save himself considerable money by assembling an ssb transceiver kit.

Some transceivers include a feature that permits the receiver to be tuned a few kHz either side of the transmitter frequency. This consists of a voltage-sensitive capacitor, which is tuned by varying the applied dc voltage. This can be a useful device when one or more of the stations in a net drift slightly. The control for this function is usually labled RIT for *receiver independent tuning.* Other transceivers include provision for a crystal-controlled transmitter frequency plus full use of the receiver tuning. This is useful for "DXpeditions" where net operation (on the same frequency) may not be desirable.

SSB Bibliography

Single Sideband for the Radio Amateur, by the American Radio Relay League, 5th Edition, 1970.

Hennebury, *Single Sideband Handbook*, Technical Material Corporation, 1964.

Pappenfus, Bruene and Schoenike, *Single Sideband Principles and Circuits*, McGraw-Hill, 1964.

Amateur Single Sideband, by Collins Radio Company, 1962.

TESTING A SIDEBAND TRANSMITTER

There are three commonly used methods for testing an ssb transmitter. These include the wattmeter, oscilloscope, and spectrum-analyzer techniques. In each case, a two-tone test signal is fed into the mic input to simulate a speech signal. From the measurements, information concerning such quantities as PEP and intermodulation-distortion-product (IMD) levels can be obtained. Depending upon the technique used, other aspects of transmitter operation (such as hum problems and carrier balance) can also be checked.

As might be expected, each technique has both advantages and disadvantages and the suitability of a particular method will depend upon the desired application. The wattmeter method is perhaps the simplest one but it also provides the least amount of information. Rf wattmeters suitable for single-tone or cw operation may not be accurate with a two-tone test signal. A suitable wattmeter for the latter case must have a reading that is proportional to the actual power consumed by the load. The reading must be independent of signal waveform. A thermocouple ammeter connected in series with the load would be a typical example of such a system. The output power would be equal to I^2R, where I is the current in the ammeter and R is the load resistance (usually 50 ohms). In order to find the PEP output with the latter method (using a two-tone test input signal), the power output is multiplied by 2.

A spectrum analyzer is capable of giving the most information (of the three methods considered here), but it is also the most costly method and the one with the greatest chance of misinterpretation. Basically, a spectrum analyzer is a receiver with a readout which provides a plot of signal amplitude vs. frequency. The readout could be in the form of a paper chart but usually it is presented as a trace on a CRT. A sweep voltage which is applied to the horizontal-deflection amplifiers of the CRT is also used to control the frequency of the LO for the first mixer (there may be other mixers but these are fixed-frequency types) in the analyzer. (See the block diagram in Fig. 1.) In order to give a meaningful output waveform, the first mixer has to have a broadband and "flat" response. It also has to have very good IMD suppression characteristics, otherwise, the mixer may generate spurious signals itself. Unfortunately, these signals fall on the same frequencies as those being measured in the transmitter output and it may be difficult to tell whether or not the spurious products are being generated in the transmitter or in the spectrum analyzer. Other precautions that should be taken would be to ensure that good RFI-prevention practices are observed. In effect, the problem is similar to trying to listen to one's own signal in the station receiver. Quite often, a signal may not be as bad *or as good* in the latter as it is at a distant station.

Two-Tone Tests and Scope Patterns

A method which is a very practical one for amateur applications is to use a two-tone test signal

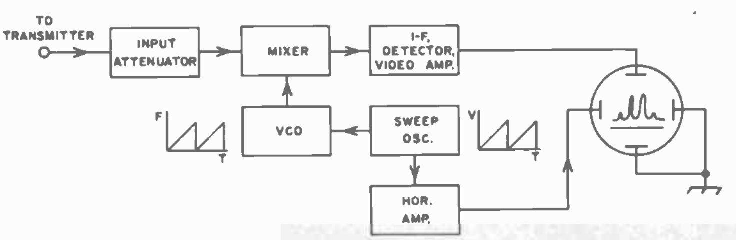


Fig. 1 — Block diagram of a spectrum analyzer.

(usually audio) and sample the transmitter output. The waveform of the latter is then applied directly to the vertical-deflection plates in an oscilloscope. An alternative method is to use an rf probe and detector to sample the waveform and apply the resulting audio signal to the vertical-deflection amplifier input.

If there are no appreciable nonlinearities in the amplifier, the resulting envelope will approach a perfect sine-wave pattern (see Fig. 2A). As a comparison, a spectrum-analyzer display for the same transmitter and under the same conditions is shown in Fig. 2B. In this case, spurious products can be seen which are approximately 30 dB below the amplitude of each of the tones.

As the distortion increases, so does the level of the spurious products and the resulting waveform departs from a true sine-wave function. This can be seen in Fig. 2C. One of the disadvantages of the scope and two-tone test method is that a relatively high level of IMD-product voltage is required before the waveform seems distorted to the eye.

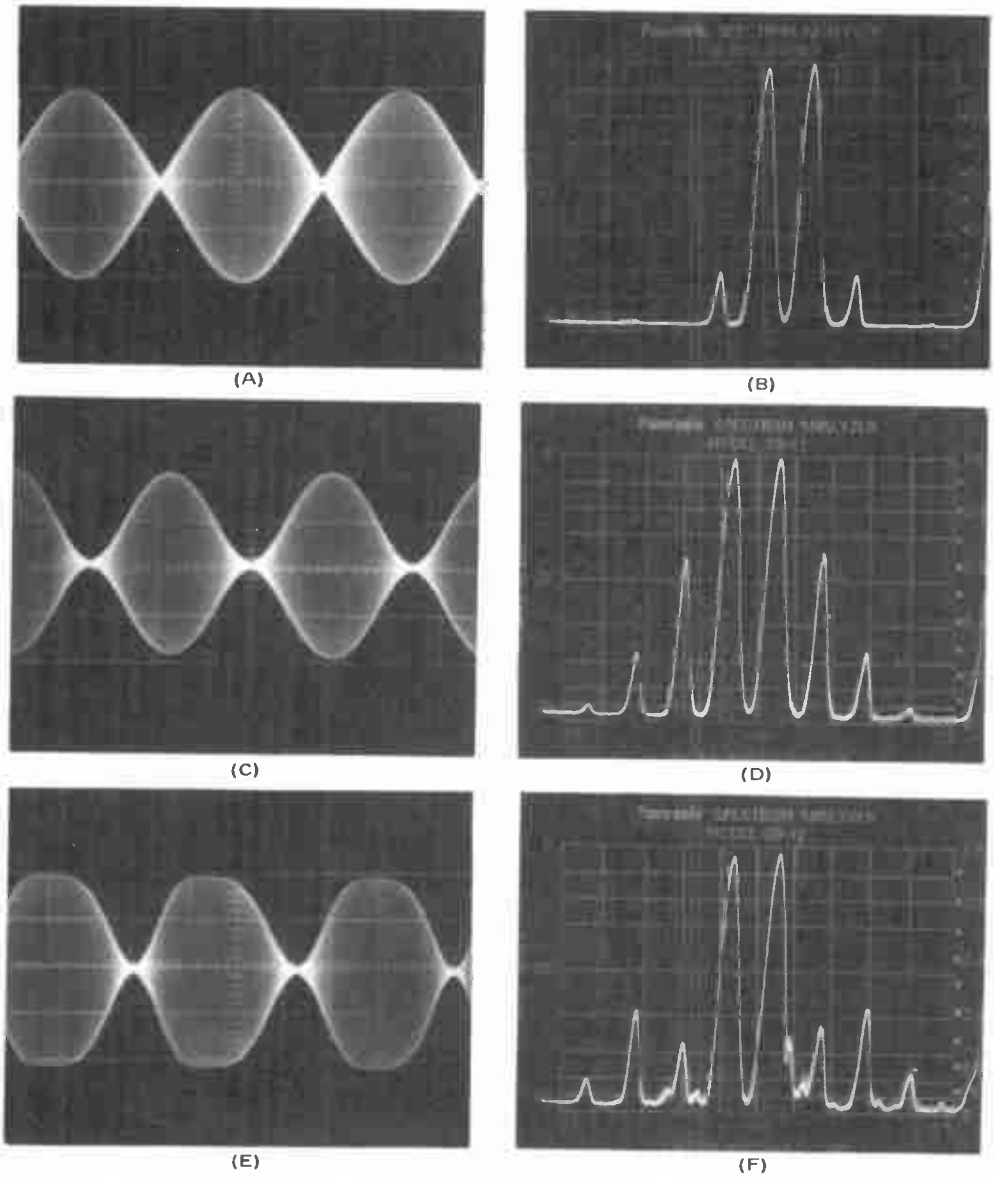


Fig. 2 — Scope patterns for a two-tone test signal and corresponding spectrum-analyzer displays. The pattern in A is for a properly adjusted transmitter and consequently the IMD products are relatively low as can be seen on the analyzer display. At C, the PA bias was set to zero idling current and considerable distortion can be observed. Note how the pattern has changed on the scope and the increase in IMD level. At E, the drive level was increased until the flattopping region was approached. This is the most serious distortion of all since the width of the IMD spectrum increases considerably causing splatter (F).

For instance, the waveform in Fig. 2C doesn't seem too much different than the one in Fig. 2A but the IMD level is only 17 dB below the level of the desired signal (see analyzer display in Fig. 2D). A 17- to 20-dB level corresponds to approximately ten-percent distortion in the voltage waveform. Consequently, a "good" waveform means the IMD products are at least 20 dB below the desired tones. Any *noticeable* departure from the waveform in Fig. 2A should be suspect and the transmitter operation should be checked.

The relation between the level at which distortion begins for the two-tone test signal and an actual voice signal is a rather simple one. The maximum deflection on the scope is noted (for an acceptable two-tone test waveform) and the transmitter is then operated such that voice peaks are kept below this level. If the voice peaks go above this level, a type of distortion called "flattopping" will occur and the results are shown for a two-tone test signal in Fig. 2E. IMD-product levels raise very rapidly when flattopping occurs. For instance, third-order product levels will increase 30 dB for every 10-dB increase in desired output as the flattopping region is approached, and fifth-order terms will increase by 50 dB (per 10 dB).

Interpreting Distortion Measurements

Unfortunately, considerable confusion has grown concerning the interpretation and importance of distortion in ssb gear. Distortion is a very serious problem when high spurious-product levels exist at frequencies removed from the passband of the desired channel but is less serious if such products fall within the bandwidth of operation. In this former case, such distortion may cause needless interference to other channels ("splatter") and should be avoided. This can be seen quite dramatically in Fig. 2F when the flattopping region is approached and the fifth and higher order terms increase drastically.

On the other hand, attempting to suppress in-band products more than necessary is not only difficult to achieve but may not result in any noticeable increase in signal quality. In addition, measures required to suppress in-band IMD often cause problems at the expense of other qualities such as efficiency. This can lead to serious difficulties such as shortened tube life or transistor heat-dissipation problems.

The two primary causes of distortion can be seen in Fig. 3. While the waveform is for a single-tone input signal, similar effects occur for the two-tone case. As the drive signal is increased, a point is reached where the output current (or voltage) cannot follow the input and the amplifier saturates. This condition is often referred to as flattopping (as mentioned previously). It can be prevented by ensuring that excessive drive doesn't occur and the usual means of accomplishing this is by alc action. The alc provides a signal that is used to lower the gain of earlier stages in the transmitter.

The second type of distortion is called "crossover" distortion and occurs at low signal levels. (See Fig. 3.) Increasing the idling plate or collector

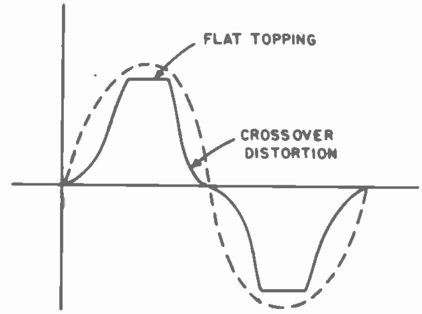


Fig. 3 — Waveform of an amplifier with a single-tone input showing flattopping and cross-over distortion.

current is one way of reducing the effect of cross-over distortion in regards to producing undesirable components near the operating frequency. Instead, the components occur at frequencies considerably removed from the operating frequency and can be eliminated by filtering.

As implied in the foregoing, the effect of distortion frequencywise is to generate components which add or subtract in order to make up the complex waveform. A more familiar example would be the harmonic generation caused by the nonlinearities often encountered in amplifiers. However, a common misconception which should be avoided is that IMD is caused by fundamental-signal components beating with harmonics. Generally speaking, no such simple relation exists. For instance, single-ended stages have relatively poor 2nd-harmonic suppression but with proper biasing to increase the idling current, such stages can have very good IMD-suppression qualities.

However, a definite mathematical relation does exist between the desired components in an ssb signal and the "distortion signals." Whenever nonlinearities exist, products between the individual components which make up the desired signal will occur. The mathematical result of such multiplication is to generate other signals of the form (2f1 – f2), (3f1), (5f2 – f1) and so on. Hence the term intermodulation-distortion products. The "order" of such products is equal to the sum of the multipliers in front of each frequency component. For instance, a term such as (3f1 – 2f2) would be called a fifth-order term since 3+2 is equal to 5. In general, the 3rd, 5th, 7th, and similar "odd order" terms are the most important ones since some of these fall near the desired transmitter output frequency and can't be eliminated by filtering. As pointed out previously, such terms do not normally result from fundamental components beating with harmonics. An exception would be when the fundamental signal along with its harmonics is applied to another nonlinear stage such as a mixer. Components at identical frequencies as the IMD products will result.

When two equal tones are applied to an amplifier and the result is displayed on a spectrum analyzer, the IMD products appear as "pips" off to the side of the main signal components (Fig. 2). The amplitudes associated with each tone and the

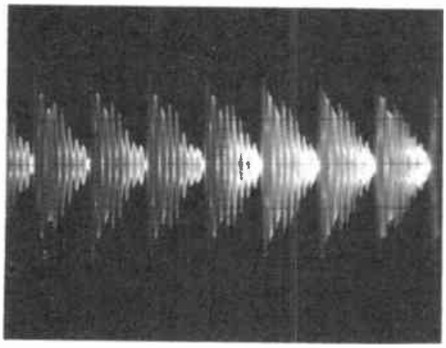


Fig. 4 – Speech pattern of the word "X" in a properly adjusted ssb transmitter.

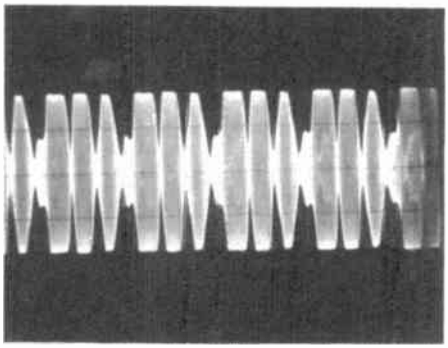


Fig. 5 – Severe clipping (same transmitter as Fig. 4 but with high drive and alc disabled).

IMD products are merely the dB difference between the particular product and one tone. However, each desired tone is 3 dB down from the *average* power output and 6 dB down from the *PEP* output.

Since the PEP represents the most important quantity as far as IMD is concerned, relating IMD-product levels to PEP is one logical way of specifying the "quality" of a transmitter or amplifier in regard to low distortion. For instance, IMD levels are referenced to PEP in Recent Equipment reviews of commercially made gear in *QST*. PEP output can be found by multiplying the PEP input by the efficiency of the amplifier. The input PEP for a two-tone test signal is given by:

$$PEP = E_p \, I_p \, [1.57 - 0.57 \frac{I_o}{I_p}]$$

where E_p is the plate voltage and I_p is the average plate current. I_o is the idling current.

Generally speaking, most actual voice patterns will look alike (in the presence of distortion) except in the case where severe flattopping occurs. This condition is not too common since most rigs have an alc system which prevents overdriving the amplifiers. However, the voice pattern in a properly adjusted transmitter usually has a "Christmas tree" shape when observed on a scope and an example is shown in Fig. 4.

Mathematical Relation Between Amplifier Nonlinearity and IMD Products

The term intermodulation-distortion product is often used and the following derivation shows how it is related to amplifier nonlinearity. The output of an amplifier can be related to the input by means of a power series of the form:

$$S = A + Br + Cr^2 + Dr^3 + Er^4 + \ldots\ldots$$

where *s* represents some parameter such as output voltage or current and *r* represents some input quantity (voltage or current). *A*, *B*, *C*, and other constants are primarily determined by the amplifier nonlinearity. *A* represents a dc term and can be neglected. In an ideal amplifier with no distortion, *C*, *D*, and the constants for the higher exponent terms will be zero and only the constant for the "linear" term, *B*, will exist. Consequently, the output quantity will be an exact replica of the input.

If the output is plotted against the input, a straight line will result, hence the term "linear operation." On the other hand, if distortion is present, the *C*, *D*, and other constants will not be zero. The values of the constants will be such that as *r* increases, the higher order terms will add (or subtract) so that *s* follows the input-output curve of the amplifier.

For a two-tone test signal, *r* can be represented by the following formula:

$$r = Ro\,(Cos\, w_1\, t + Cos\, w_2\, t);\quad w_1 = 2\pi f_1 \quad w_2 = 2\pi f_2$$

where *f*1 and *f*2 are the frequencies of the two tones. If this equation for *r* is substituted into the power series, many terms will result and the algebra involved to find each one would be rather tedious. However, the purpose here is only to show how the IMD products come about. For instance, some terms will yield products such as:

$$(Cos^2\, w_1\, t)\,(Cos\, w_2\, t)$$

The squared term can be expanded by a trigonometric formula:

$$Cos\, 2\, w_1 t = \frac{1 + Cos\, 2\, w_1 t}{2}$$

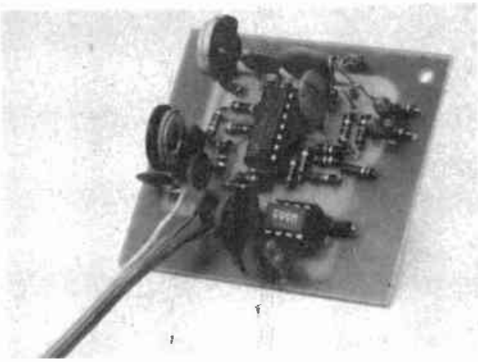
This gives rise to a term $Cos\, 2\, w_1 t\, Cos\, w_2 t$ which can be expanded by another trigonometric formula to give:

$$Cos\, 2\, w_1\, t\, Cos\, w_2\, t = \frac{1}{2}[Cos\,(2\, w_1 + w_2)\, t + Cos\,(2\, w_1 - w_2)\, t]$$

The second term in the bracket represents a third-order IMD "product" which falls close to the ssb passband. Notice that the exponents of the product functions which gave rise to this term are 2 and 1, respectively, hence the term "third-order" product. The manner in which the terms increase will depend upon the distortion curve but generally speaking, the amplitude will follow a law which is proportional to *r* raised to a power *x*, where *x* is the order of the term.

A MODERN SOLID-STATE VOX

In *QST* for March, 1964, Campbell[1] described a simple and inexpensive voice-operated relay (VOX) device which he called "a VOX in a box." Today, most manufactured ssb transceivers include a VOX function. It would seem that insufficient time or effort has gone into the design of some commercial VOX circuits, because performance is often poor. The VOX unit described here retains Campbell's concepts of a small, low-cost device, but ICs and modern circuit techniques have been employed to improve operational characteristics.



The unit shown here was designed and built by W1KLK. It originally appeared in *QST* for March, 1976.

Circuit Description

Two integrated circuits, an LM3900 (a quad of Norton operational amplifiers) and an NE555 timer, have been used in the VOX circuit. Each IC is currently priced at one dollar. A description of the NE555 has previously appeared in *QST*[2] and will not be repeated here. The Norton circuit is an unusual type of operational amplifier in which the differential input transistors of a conventional op amp have been replaced by a "current mirror" circuit to obtain a non-inverting input directly from the inverting input. One advantage of this circuit arrangement is operation from a single supply voltage. A simplified diagram of the input portion of the Norton amplifier is shown in Fig. 1A. CR1 and the base-emitter junction of Q2 clamp the maximum input voltages at approximately 0.5 volt. External series resistors are employed to convert voltage changes to current differences which are applied to the input terminals. CR1 and Q1 form the mirror circuit which assures that the bias current at both inputs will be the same; whatever bias voltage is applied to the noninverting input will be reproduced for the inverting input.

The basic design data for the LM3900 amplifiers are given in Fig. 1B. At audio frequencies, the maximum gain of a single stage is 40 dB, while the open-loop gain is specified at approximately 70 dB. Input bias current is rated at 30 nanoamperes (nA). Each of the four op amps in the LM3900 may be operated independently.

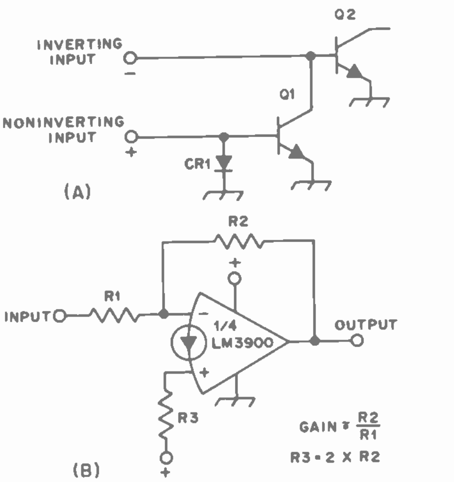


Fig. 1 — (A) Simplified input circuit for a Norton operational amplifier such as used in the National Semiconductor LM3900. (B) Design equations for an audio amplifier using the LM3900.

The schematic diagram of the VOX device is shown in Fig. 2. Three of the LM3900 sections have been configured as high-gain audio amplifiers. U1A and U1B amplify the signal from the microphone. For applications where a high-output microphone is employed, or when the audio signal is sampled after a preamplifier stage, the builder may wish to reduce the gain of U1A. This can be accomplished by changing the values of R1 and R2 in accordance with the design equations in Fig. 1B. If R2 is changed, the value of R3 must be modified so that R3 is twice the resistance of R2. U1C functions as an amplifier for audio sampled at the station speaker. Coupling capacitors in the audio stages have been chosen to reduce response below 300 Hz. This will minimize hum problems.

Outputs from the microphone and speaker amplifiers are capacitively coupled to rectifier stages which convert the audio signals to varying dc voltages. Germanium diodes, because of the lower threshold voltage, have been used as audio rectifiers instead of silicon units. The outputs of the two rectifier stages are summed resistively by means of R6 and R7, and applied to the inverting input of a voltage comparator, U1D. The output of U1D remains high (approximately 0.5 volt less than the supply voltage) so long as the voltage at the noninverting input is less than the 0.2-volt reference applied to the inverting input. Whenever the input exceeds the reference, the output of the comparator goes low – to near the ground or common potential. Voltage output from the microphone-signal rectifier is positive and, thus, will cause the comparator to switch as soon as the reference is exceeded. Because the speaker-signal rectifier produces negative voltage, it will not trigger the comparator. If the outputs of the two rectifiers are equal, as will happen when the microphone is picking up audio from the speaker, the resulting voltage from the summing network

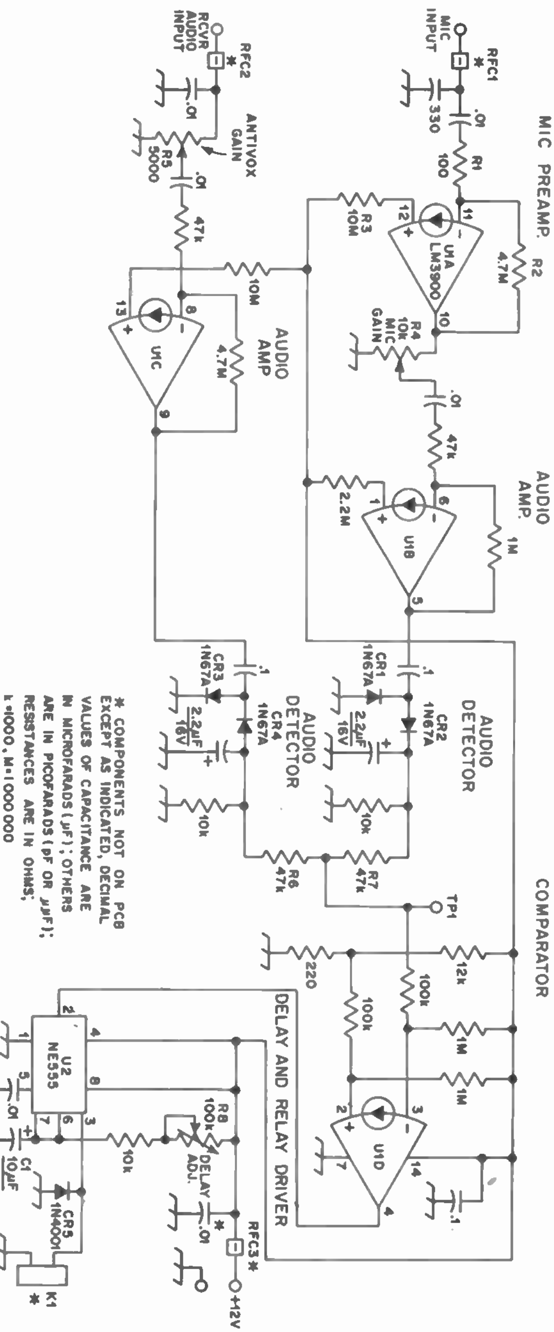


Fig. 2 — Schematic diagram of the VOX unit. Unless otherwise specified, resistors are 1/4-watt composition. Capacitors with polarity marked are plastic-encapsulated tantalum; others are disk ceramic.
C1 — For text reference.
CR1 to CR4, incl. — Germanium diode, 1N34A, 1N67 or equivalent.
CR5 — Silicon diode, 50 PRV or more, 1N4001 or similar.
K1 — Miniature type, 12-volt coil (see text).
R1-R3, incl., R6, R7 — For text reference.
R4, R5, R8 — Miniature control (see text).
RFC1-RFC3, incl. — Ferrite bead.
U1 — National Semiconductor LM3900.
U2 — Signetics NE555 or Motorola MC1455.

will be zero and the comparator will not trigger. The ability to reject speaker audio is usually called the ANTIVOX function.

The positive-to-ground transition of the comparator output starts the timing cycle of the NE555. The length of the time cycle is determined by the values used for R9 and C1. The time delay produced is identical each time the microphone signal stops. One of the major difficulties of earlier VOX circuits was that capacitor discharge circuits were used where the capacitor would not always be fully charged, so the time delay produced would vary. Operators compensated for the uncertain time delay by using a modified version of the English language where an ahhh, oouh, or errr was inserted after each word to assure that the VOX relay would not drop out prematurely.

The NE555 has a current-switching capability of 200 mA, sufficient to directly drive either a relay or a solid-state switching arrangement. CR5 is included to protect the IC from transients generated when switching an inductive load such as a relay coil.

Components and Construction

The VOX unit is constructed on a 2-3/8 X 2-3/4-inch etched circuit board. The photo indicates that one third of the board real estate is unused, so a smaller version is possible. The type of controls and relay employed will be determined by the builder's individual requirements. This writer's unit uses pc-mount controls which are aligned on the board so that they may be accessed through small holes in the rear panel of the transceiver. If panel-mount controls are desired, Mallory MLC units may be used for R4, R5 and R8.

The VOX device is small enough so it can be mounted inside most rigs. If a separate VOX unit is needed, a small utility or Minibox will make an appropriate housing. Rf interference can cause trouble, so the unit should be shielded in any application where rf fields may be present. The bypass capacitors for the audio inputs are located on the circuit board. If the leads from the audio connectors are more than a few inches long, the bypass capacitors and their associated ferrite-bead chokes should be mounted at the connectors.

No provision has been made for mounting the relay on the circuit board, as the type of relay will

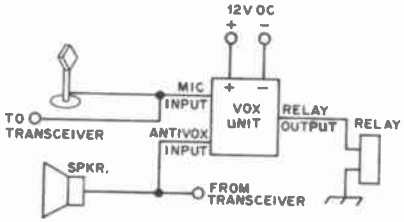


Fig. 3 – Typical connections to the VOX unit.

depend on how the VOX device will be used. Any 12-volt relay which requires less than 200 mA of current can be employed. When the VOX relay must drive a second relay, such as the antenna relay in a transceiver, the fast operating time of a reed relay is needed to prevent clipping of the first syllable spoken. The total close time of all relays connected in tandem should be 10 milliseconds or less. If the VOX relay will perform all switching functions directly, a miniature control relay such as the Potter & Brumfield R10 series is appropriate. These relays are available in 2-, 4- and 6-pole versions, part numbers R10-E1-Y2-185, R10-E1-Y4-V185 and R10-E1-Y6-V90, respectively.

The circuit board is designed for 1/4-watt resistors which are mounted flat. If 1/2-watt units are used, they must be positioned vertically. Care must be employed when mounting and soldering the germanium diodes. If the leads are bent too close to the body of the diode, breakage can result. If excessive heat is applied to the diode, it can be damaged, so use a heat sink (such as a small alligator clip) when soldering. Assure that proper polarity is observed when installing the diodes and tantalum capacitors.

Installation and Operation

Typical connections for the VOX unit are shown in Fig. 3. Shielded cable should be used for all audio connections. Audio for the ANTIVOX function can be sampled at the station speaker or at the phone-patch output (which is a feature of many commercial transceivers). If VOX operation of a cw rig is desired, connect the output of a sidetone monitor to the microphone input of the VOX unit. The mic gain control should be set so that the VOX relay closes each time a word is spoken. The delay control should be adjusted to fit individual speech patterns and operating habits. The delay time must be long enough that the VOX relay will drop out only during a pause in speech. There are two methods of setting the ANTIVOX gain control. The first way is simply to advance the control until audio from the speaker does not trip the VOX unit. A more scientific approach is to connect a voltmeter to TP1. With no audio input, the meter should read only the comparator reference voltage, approximately 0.1 volt. Tune the receiver to provide a steady tone signal, such as the heterodyne note from a crystal calibrator. Advance the ANTIVOX control until the voltmeter registers only the reference voltage. The ANTIVOX gain should be set with the audio from the speaker slightly louder than is necessary during normal operation.

VOX control can be a helpful operating aid. It can speed up traffic handling and contest operation. It might make round-table discussions more natural. The unit described here will help to eliminate some of the operator idiosyncrasies that have become associated with VOX operation. The operators themselves can cure the rest.

References

[1] Campbell, "A VOX in a Box," *QST* for March, 1964.

[2] Technical Topics, "Time – IC Controlled," *QST* for June, 1972.

[3] Blakeslee, "Some Thoughts on Station Control," *QST* for January, 1966.

[4] Blakeslee, "A Solid-State VOX," *QST* for September, 1970.

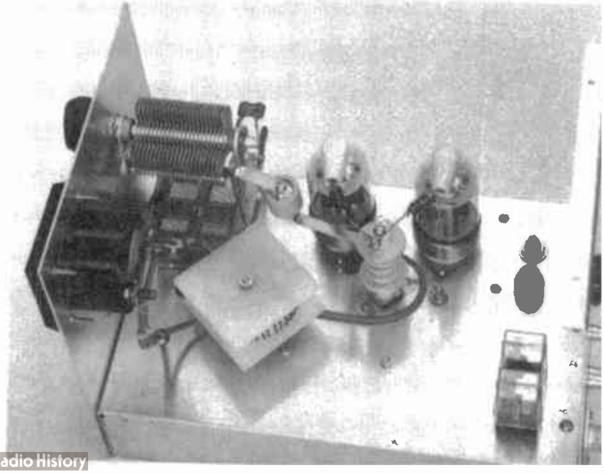
A TRANSVERTER FOR 1.8 MHZ

Owners of five-band transceivers often get the urge to try "top band." Converting a transceiver to cover a frequency range for which the rig was not designed is difficult indeed. A far better approach is to build an outboard transverter, such as the one described here. This particular system requires one watt of drive power at either 21 or 28 MHz. Many transceivers can provide this low-level output along with the power supply voltages through an accessory socket.

The Circuit

A schematic diagram of the transverter is given in Fig. 1. Q1 operates as a crystal oscillator, to produce the local oscillator energy for the receive (Q5) and transmit (Q2) mixing stages, which runs continuously. During transmit 21.1 MHz ssb or cw energy is supplied to the emitter of Q2 through a power divider network. This signal is mixed with

Top view of transverter with cover removed. Final amplifier circuit is at the left. The rear apron has an accessory socket for an external power supply (transceiver), rf, and remote-keying connectors. The plate meter is at the lower left.



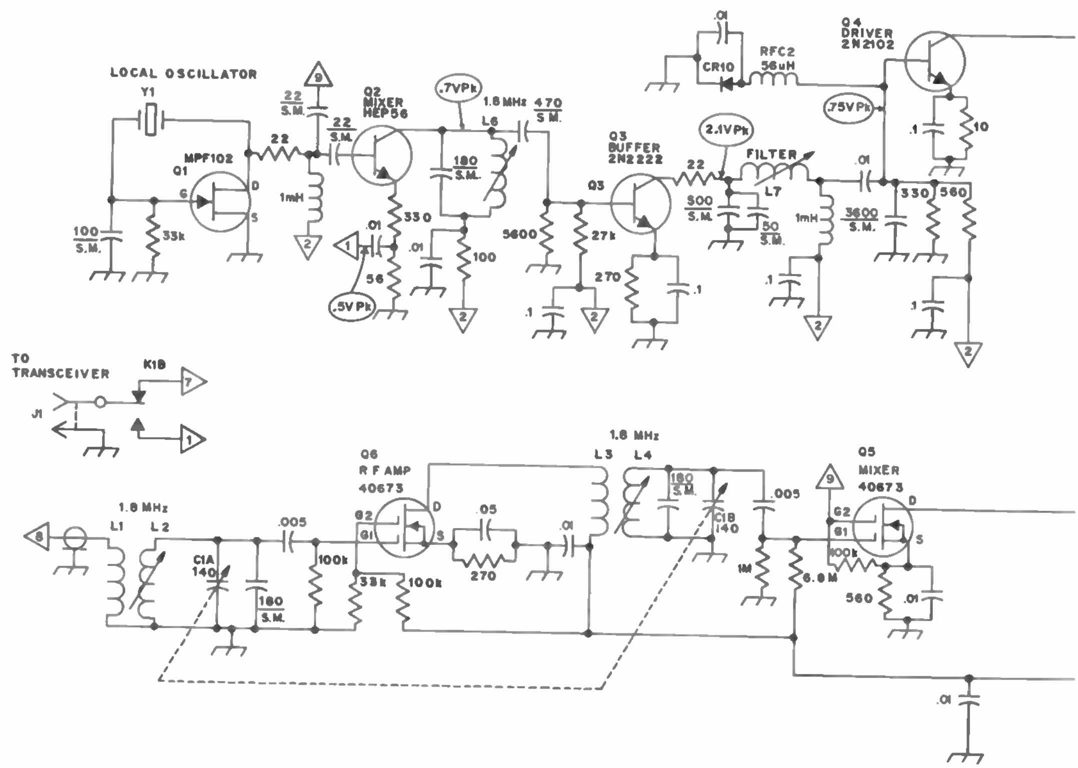
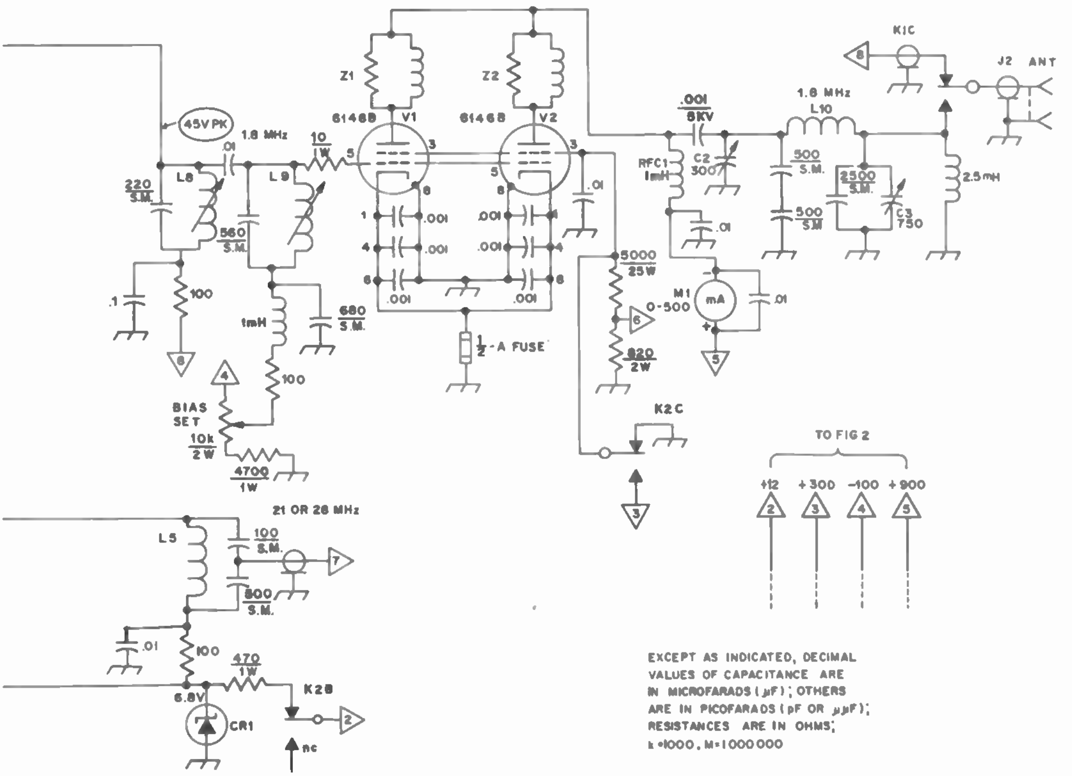


Fig. 1 — Schematic diagram for the transverter. Resistors are 1/2-watt composition and capacitors are disk ceramic, unless otherwise noted.

C1 — Dual-section air variable, 140 pF per section, or two 150 pF air variable units.
C2 — Air variable (Millen 19280).
C3 — Dual-section broadcast variable, 365 pF per section, both sections connected in parallel.
CR1 — Zener diode, 6.8-volt, 1-watt (1N4736).
CR10 — Silicon, 50 PRV, 100 mA.
J1 — Phono type, chassis mount.
J2 — Coaxial receptacle, chassis mount.
K1, K2 — 12 V dc, 2-A contacts, dpdt relay (Radio Shack 275-206).
L1 — 11 turns of No. 28 enam. wire wound over L2.
L2, L4 — 19.5-24.3 μH variable inductor (Miller 46A225CPC).
L3 — 22 turns of No. 38 enam. wire wound on L4 coil form.
L5 — 18.8-41.0 μH variable inductor (Miller 42A335CPC).
L6, L8 — 35-43.0 μH variable inductor (Miller 46A395CPC).
L7 — 13.2-16.5 μH variable inductor (Miller 46A155CPC).
L9 — 10.8-18.0 μH adjustable coil (Miller 21A155RBI).
L10 — 42 turns, No. 16 enam. wire equally spaced on a T-200 Amidon core.
M1 — 500 mA, panel mount (Simpson 17443 or similar).
Q5, Q6 — RCA MOSFET.
RFC1 — 1 mH. 500 mA rf choke (Johnson 102-572).
RFC2 — 56 μH rf choke (Millen J-302-56).
Y1 — 19.3-MHz crystal is used for a 21-MHz i-f, 26.5-MHz crystal for a 28-MHz i-f.
Z1, Z2 — 2 turns, No. 18 enam. wound over 47-ohm, 2-watt composition resistor.

the 19.3-MHz output from the LO producing 1.8 MHz power which is amplified by Q3, followed by a filter network. Q4 provides adequate drive to the pair of 6146Bs. The PA stage operates class AB1 which will deliver in excess of 100 watts PEP output.

During receive, an incoming signal is amplified by Q6, a dual-gate, diode-protected MOSFET. The output from the rf amplifier is mixed with local-oscillator energy at Q5 to produce a receiving i-f of 21 MHz. The frequency of the crystal is the only change required to make this system useable at 28 MHz. Changeover from transmit to receive is accomplished by K1 and K2 which are controlled by the associated transceiver. If the LO frequency is 19.3 MHz, the 1.8 to 2.0 MHz band will correspond with 21.1 to 21.3 MHz on the transceiver dial. Likewise, with a 26.5 MHz crystal in the LO circuit, the 160-meter band will appear between 28.3 and 28.5 MHz.



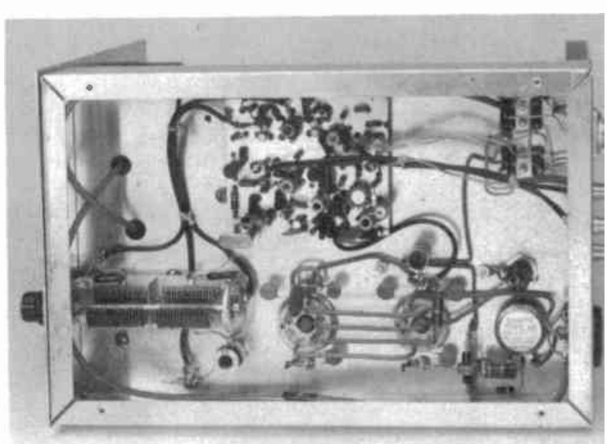
If the various supply voltages can not be obtained from the transceiver, an economy power supply shown in Fig. 2 can be used. The 6.3- and 5-volt windings of T1 are series-connected to provide 11.5 volts to power K1 and K2, the receiving converter and the predriver stages of the transmitting section. The windings must be phased properly to prevent cancellation of the voltages. If no output is obtained when the windings are connected, reverse the leads of one winding. The 11.3-volt ac is rectified by CR6.

Bias voltage is obtained for V1 and V2 by connecting a 6.3-volt filament transformer in back-to-back fashion with the 6.3-volt winding of T1. The 125-volt ac output from T2 is rectified, filtered, and then routed to the bias-adjust control, R1, to establish a PA resting plate current of 50 mA.

The metering circuit consists of a 500 mA meter connected in the plate voltage line. Other meters may be employed by using the proper shunts, as described in the Measurements Chapter.

Construction

An aluminum chassis which measures 7 × 11 × 2 inches is used as the base for the transverter. A homemade panel and cabinet enclose the unit. The front panel is 8 × 7-1/4 inches. The layout employed should be apparent from the photographs. All long runs of rf wiring should be made with subminiature coaxial cable (RG-174/U or similar).



The bottom view of the chassis, the sockets for the 6146B tubes are at the lower center. The etched-circuit board is above the final amplifier tube sockets and the T-R relays at the upper right. The different supply voltages are obtained from the associated transceiver.

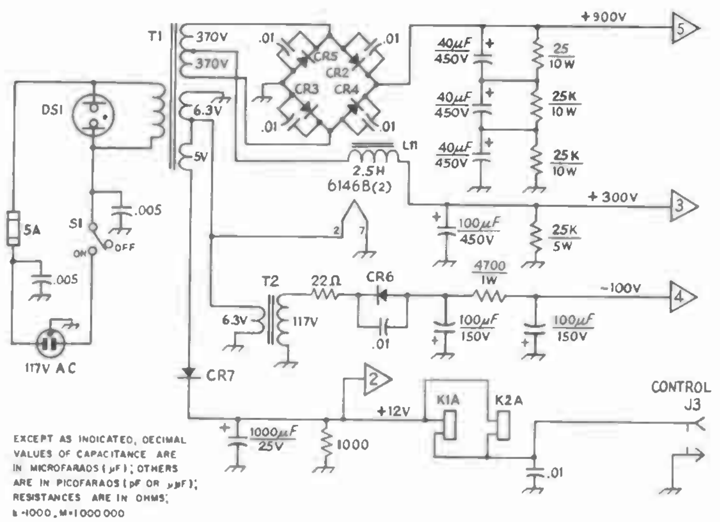


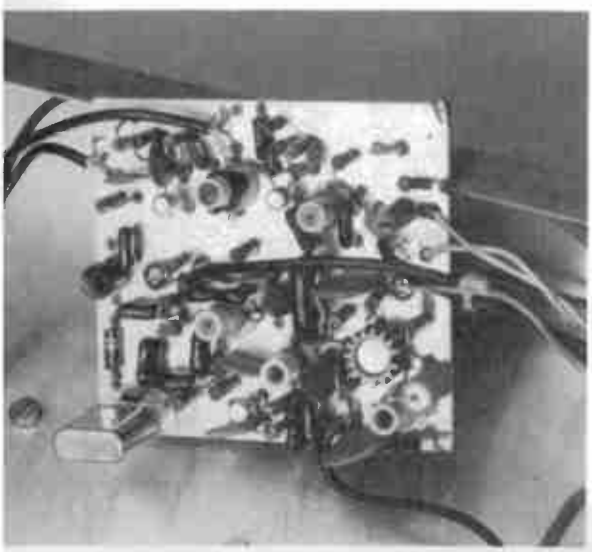
Fig. 2 — Diagram of the power-supply section. Resistors are 1/2-watt composition. Capacitors are disk ceramic, except those with polarity marked which are electrolytic.
CR2-CR5, incl. — Silicon, 1000 PRV, 1 A.
CR6, CR7 — Silicon, 400 PRV, 1 A.
J3 — Phono type, chassis mount.
K1, K2 — see Fig. 1.
L11 — Power choke, 130 mA (Allied 6X24HF or equiv.).
S1 — Spst toggle.
T1 — Power transformer, 117-V primary; secondary windings 740 V ct at 275 mA, 6.3 V at 7 A, and 5 V at 3 A (Stancor P-6315 or equiv.).
T2 — Filament transformer, 117-V primary; 6.3-V, 1-A secondary.

The receiver section, driver stages and local oscillator are constructed on a double-sided printed-circuit board measuring 3 × 3-1/2 inches. Inductors L1 and L2 are mounted on the chassis close to C1. Short leads are used from the circuit board to the PRESELECTOR capacitor and L1-L2 which are located on the underside of the chassis. The final tank inductor is wound on an Amidon T-200 toroid core. It is supported above the chassis by a ceramic standoff insulator and two pieces of Plexiglas.

Tune Up

Provision must be made to reduce the power output of most transceivers used with the transverter since only about one watt of drive power is required. Too much rf voltage can damage the HEP 56 and will "smoke" the input resistors. Some transceivers are capable of delivering sufficient drive by removing the screen voltage from the PA stage. Or, it may be practical to disable the PA and obtain a sample of driver output by a link-coupling circuit.

Before testing the transverter, assure that the changeover relays, K1 and K2, are connected to the remote-keying terminals of the transceiver. Then connect an antenna to J2 and listen for signals. Peak the incoming signals with the PRE-



Close-up view of the printed-circuit board. This board has the local oscillator, receiver, and low-level driver stages. The crystal socket and crystal for the LO are shown at the lower left.

SELECTOR control. The slugs of L2 and L4 should be adjusted for the highest S-meter reading on the transceiver. L5 should be set for maximum output at 21 or 28 MHz. If the receiving converter is functioning properly, it will be possible to copy a 0.1 μV signal without difficulty in areas where atmospheric and man-made noise are at a minimum. If no signals can be heard, check Q1 to make certain that it is working properly. A wavemeter or general-coverage receiver can be employed to see if the crystal oscillator is operating.

Attach a 50-ohm load to J2 before testing the transmitter section. Set R1 for an indicated resting plate current of 50 mA on M1. This adjustment should be made without drive applied but with K1 and K2 energized. Next, apply about one watt of 21.1-MHz cw drive power at J1. Tune L6, L7, L8 and L9 for maximum meter reading. While monitoring the plate current, tune C2 for a dip. C3 is the PA LOADING control. When the PA capacitors are properly adjusted, the plate current will be about 220 mA.

A LOW-POWER SSB/CW TRANSMITTER FOR 80 OR 20 METERS

A number of QRP transmitter designs have appeared in the past, mostly for cw-only operation. The unit to be described operates in both the ssb and cw modes. Using solid-state devices throughout, the transmitter is capable of delivering up to nine watts PEP output into a 50-ohm load. A 9-MHz i-f in conjunction with a VFO that tunes 5.0 to 5.5 MHz results in single-band operation on either 80 or 20 meters. A regulated 12-volt dc supply that can furnish at least two amperes is required to power the transmitter.

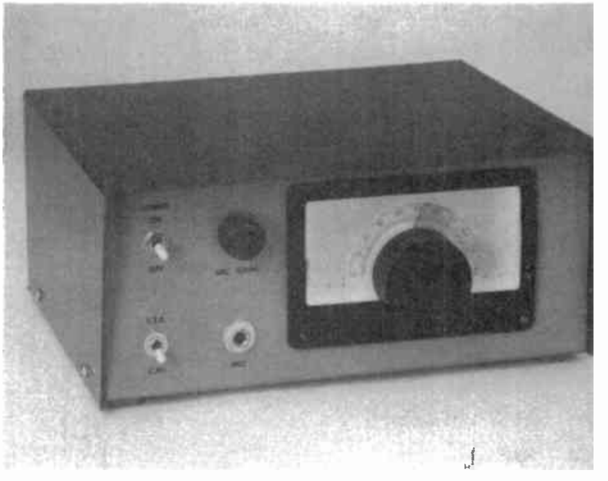


Fig. 1 — Front view of the transmitter with cover in place. The two-piece chassis is made from sheet aluminum. The front panel, which measures 9-1/2 by 4 inches, is spray-painted orange and the cover is finished in brown. White decals are used to identify the power switch, mode switch, microphone gain control, and jack. For cw operation, the key plugs into a jack at the rear of the chassis. A two-speed vernier dial is employed for VFO tuning and stick-on rubber feet are fitted on the bottom of the chassis.

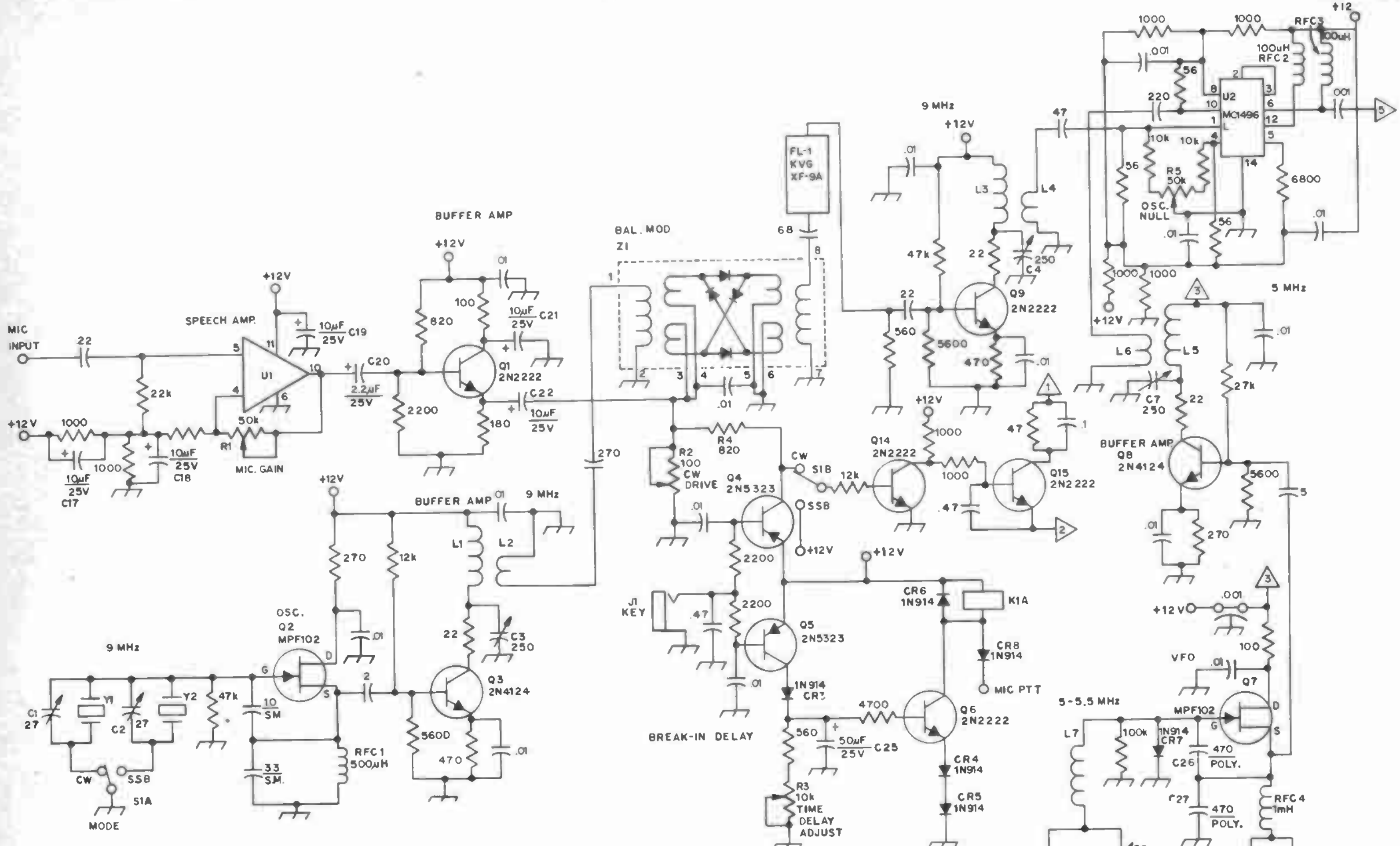
Construction Details

Four separate circuit board assemblies are used. Two boards, measuring 6 by 2-3/8 inches and 5-3/4 by 2-3/4 inches, contain most of the transmitter circuitry. The VFO and power output amplifier are included on separate boards, measuring 2 by 3 inches and 2-1/2 by 4 inches respectively. Double-clad circuit board should be used for all except the VFO board. The copper plane on the component side of each board provides a good rf ground and thus enhances stability in the unit. Component leads which are soldered to ground should be soldered to both the ground plane on the component side and the ground foil on the reverse side. To prevent other leads from shorting to the ground plane, their holes on the component side are drilled out slightly with a 1/4-inch drill before mounting. As can be seen in Fig. 2, shields, made from pieces of double-clad circuit board soldered to the main circuit boards, are used to isolate stages which are susceptible to stray rf pickup.

The VFO is housed in a four-walled enclosure formed by four pieces of circuit board soldered together at their common seams. The VFO board fits snugly inside and is soldered along its edges to the walls of the enclosure. The tuning capacitor, C6, mounts firmly against the front wall from which its shaft protrudes. Dc power connection is made via a feedthrough capacitor. A short length of subminiature RG-174/U coax connects the VFO output to U2.

If the 80-meter version of the unit is being constructed, Q11 and associated components in this amplifier stage are to be omitted from the board since the stage is required only for 20-meter operation. Instead, then, a jumper connection is made from common connection 4 (as indicated in the schematic diagram) directly to the base of Q12. In the 20-meter unit this stage is included and the jumper omitted.

The broadband power output stage employs a 2N6367 rf power transistor rated for 9 watts PEP output with a −30 dB IMD specification. All rf-carrying leads associated with the base circuit of Q13 should be absolutely as short as possible. Because of the low base input impedance – two or three ohms – even small amounts of stray reactance cannot be tolerated. Leads as short as one inch can contribute a considerable amount of inductive reactance in the base circuit.



EXCEPT AS INDICATED, DECIMAL VALUES OF CAPACITANCE ARE IN MICROFARADS (µF); OTHERS ARE IN PICOFARADS (pF OR µµF); RESISTANCES ARE IN OHMS; k = 1000, M = 1000 000.

POLY. = POLYSTRENE (MALLORY TYPE SX OR EQUIV.)

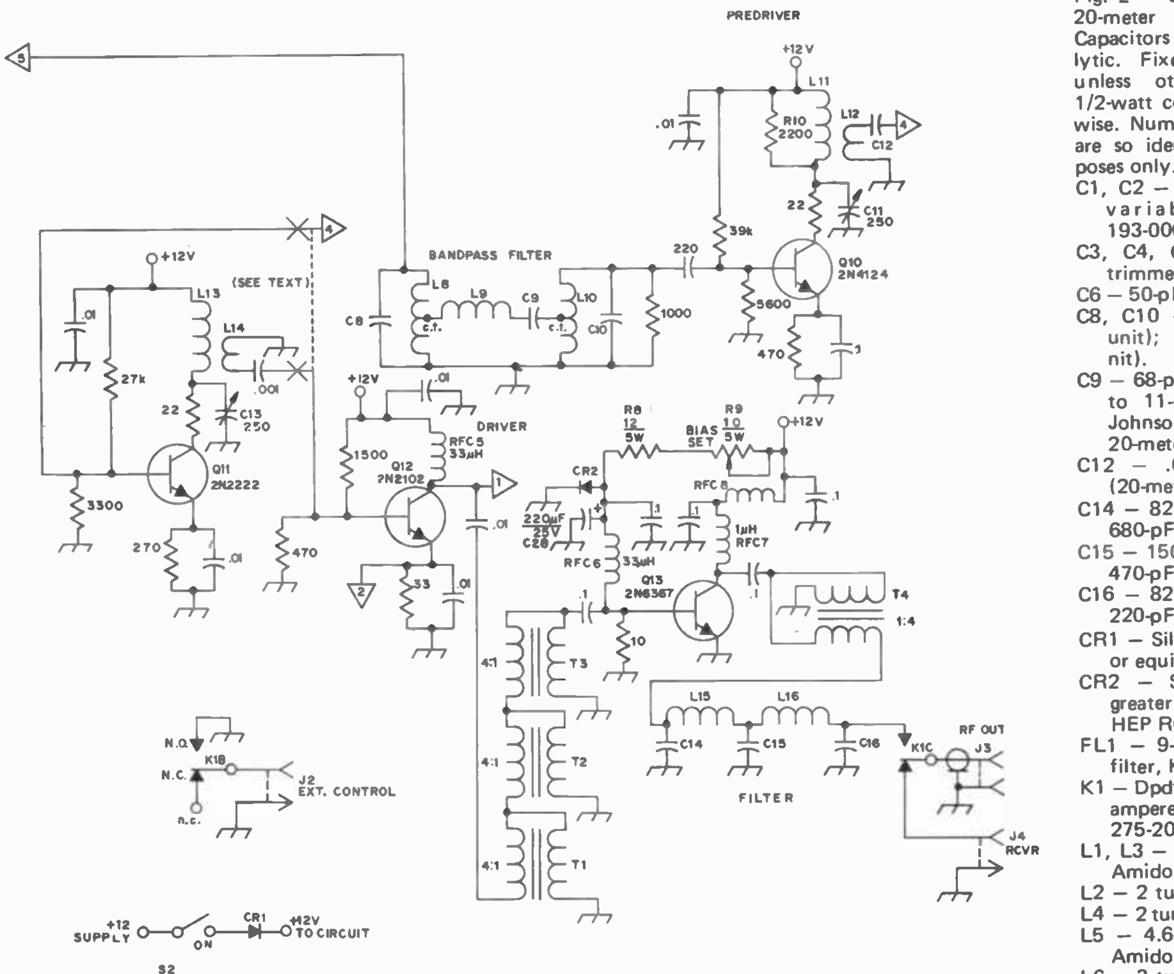


Fig. 2 — Schematic diagram of the 80- or 20-meter low-power ssb/cw transmitter. Capacitors with polarity marked are electrolytic. Fixed-value capacitors are ceramic unless otherwise noted. Resistors are 1/2-watt composition unless marked otherwise. Numbered parts not appearing below are so identified for pc-board layout purposes only.

C1, C2 — 2.0- to 27-pF printed-circuit air variable (E.F. Johnson No. 193-0008-005 or equiv.).

C3, C4, C7, C11, C13 — 250-pF max. trimmer (Arco Elmenco 426).

C6 — 50-pF air variable.

C8, C10 — 220-pF silver mica (80-meter unit); 100-pF silver mica (20-meter unit).

C9 — 68-pF silver mica (80-meter unit); 1.7- to 11-pF miniature air variable (E.F. Johnson No. 187-0106-005 or equiv. for 20-meter unit).

C12 — .001-μF (80-meter unit); 68-pF (20-meter unit).

C14 — 820-pF silver mica (80-meter unit); 680-pF silver mica (20-meter unit).

C15 — 1500-pF silver mica (80-meter unit); 470-pF silver mica (20-meter unit).

C16 — 820-pF silver mica (80-meter unit); 220-pF silver mica (20-meter unit).

CR1 — Silicon diode, 50 PRV, 1A (1N4001 or equiv.).

CR2 — Silicon diode, 50 PRV, 3A or greater, stud-mounting type (Motorola HEP R0130 or equiv.).

FL1 — 9-MHz, 2.5-kHz bandwidth crystal filter, KVG type XF-9A.

K1 — Dpdt 12 Vdc relay, contact rating of 1 ampere or greater (Radio Shack cat. No. 275-206 or equiv.).

L1, L3 — 2.5-μH, 25 turns No. 24 enam. on Amidon T-50-6 toroid core.

L2 — 2 turns No. 24 enam. wound over L1.

L4 — 2 turns No. 24 enam. wound over L3.

L5 — 4.6-μH, 34 turns No. 24 enam. on Amidon T-50-6 core.

L6 — 3 turns No. 24 enam. wound over L5.

(Continued on next page)

L7 — 7.8- to 12.0-μH slug-tuned coil (Miller 4309).
L8, L10 — 8.9-μH, 49 turns No. 26 enam. on Amidon T-50-6 core (80-meter unit); 1.2-μH, 17 turns No. 24 enam. on Amidon T-50-6 (20-meter unit).
L9 — 26.5-μH, 68 turns No. 28 enam. on Amidon T-68-2 toroid core (80-meter unit); 17.6-μH, 55 turns No. 28 enam. on Amidon T-68-2 core (20-meter unit).
L11 — 8.1-μH, 45 turns No. 26 enam. on Amidon T-50-6 core (80-meter unit); 0.68-μH, 13 turns No. 24 enam. on Amidon T-50-6 core (20-meter unit).
L12 — 3 turns No. 24 enam. wound over L11.
L13 — 0.68-μH, 13 turns No. 24 enam. on Amidon T-50-6 core (used in the 20-meter unit only).
L14 — 1 turn No. 24 enam. wound over L13.
L15, L16 — 2.6-μH, 25 turns No. 24 enam. on Amidon T-50-6 core (80-meter unit); 0.28-μH, 8 turns No. 24 enam. on Amidon T-50-6 core (20-meter unit).
RFC8 — 6 turns No. 28 enam. using Amidon Jumbo Ferrite Bead as toroid core.
S1 — Dpdt subminiature toggle switch (Radio Shack 275-1546 or equiv.).
S2 — Spst subminiature toggle switch (Radio Shack 275-324 or equiv.).
T1, T4 — 4:1 broadband transformer; 12 turns of 1 twisted pair of No. 24 enam. wire (6 turns per inch, not critical) wound on Amidon FT-61-301 toroid core.
T2 — 4:1 broadband transformer; 6 turns of 2 twisted pairs of No. 26 enam. wire (6 turns per inch) wound on Amidon FT-61-301 toroid core. In twisting the wires, a single turn consists of a full twist of all wires. The two wires at the ends of each pair are soldered together and each pair then comprises one winding of the transformer.
T3 — 4:1 broadband transformer; 4 turns of 4 twisted pairs of No. 26 enam. wire (6 turns per inch) wound on Amidon FT-61-301 toroid core. A single turn consists of a full twist of all wires. The two wires at the ends of each pair are soldered together, the ends of two pairs are soldered together, resulting in two 4-conductor wires. Each 4-conductor wire comprises one winding of the transformer.
U1 — 741 operational amplifier (Motorola MC1741, National Semiconductor LM741, Fairchild μA741 or equiv.), 14-pin DIP used here.
U2 — Balanced modulator IC (Motorola MC1496L or National Semiconductor LM1496L, Signetics 5596). 14-pin DIP used here.
Y1 — 8999.0-kHz crystal (KVG type XF 903).
Y2 — 8998.5-kHz crystal (KVG type XF 901).
Z1 — Double balanced mixer, model SRA-1 manufactured by Mini-Circuits Laboratory, 2913 Quentin Rd., Brooklyn, NY 11229.

Some care must be taken in mounting Q13. The power amplifier board is drilled out to allow the flange for heat-sink mounting to pass through the board. The transistor leads, which are short straps, then lie flush with the top surface of the board and the flange lies beneath the board. The leads are soldered to the board as close to the body of the transistor as possible and should not be bent. The heat sink consists of a 1/16-inch thick rectangular piece of sheet aluminum cut to the same size as the circuit board. The transistor requires two bolts for mounting to the heat sink. The use of silicone grease to improve thermal conductivity between the transistor and heat sink is recommended. CR2, a stud-mounting type diode, also mounts on the heat sink near the power transistor since good thermal contact with Q13 is necessary for CR2 to provide thermal compensation for the output transistor. Note also that in the driver stage, Q12 requires a heat sink.

The chassis is formed from a piece of sheet aluminum bent into a U-shape. The front panel measures 9-1/2 by 4 inches and the chassis is 7 inches deep. Except for the VFO enclosure, the circuit boards mount vertically on the chassis by means of spade bolts. An aluminum cover is formed to fit over the chassis. A Jackson Bros. No. 4103 dial mechanism is used for VFO tuning. Any similar mechanism with a slow tuning rate should be satisfactory. Front-panel controls are the microphone-gain control, VFO tuning, mode switch, and on-off switch. The key jack, rf-output connector, and dc-power connector are mounted at the rear of the chassis. The transmit-receive relay socket mounts inside the chassis and the spare set of contacts are brought out to a connector at the back of the unit.

Initial Adjustments and Operation

A well-calibrated general-coverage receiver, a VOM, and a VTVM with an rf probe (or better yet, a good rf oscilloscope) are required for making the initial adjustments on the transmitter. Connect a 12-volt dc supply to the transmitter (except the final power amplifier stage). With power turned on, check to see that the unit draws no more than about 200 mA from the supply. Any reading well in excess of this value indicates a wiring error or defective components.

Tune the receiver to 9 MHz to locate the heterodyne oscillator signal. A few feet of hookup wire placed near the circuit board should suffice for a receiving antenna. Switching the mode switch to change the oscillator crystals Y1 and Y2 should shift the oscillator frequency accordingly. Adjust C3 for maximum signal at the collector of Q3 as displayed on an oscilloscope or VTVM with rf probe.

Plug a key into the key jack and set cw drive control, R2, for minimum resistance. Place the mode switch in the cw position. Depress the key while monitoring the collector of Q9 for rf output with the oscilloscope or rf probe. Slowly increase the cw drive level until a 9-MHz signal appears at Q9. Adjust C4 for maximum signal at the collector. Tuning C1 allows a small variation of the cw frequency. Adjustment is not critical but it should be tuned to place the cw signal inside the passband of FL1. If the signal is too close to the edge of the passband, keying may cause the filter to "ring," producing key clicks at the output.

Tune the receiver between 5.0 and 5.5 MHz and vary the VFO tuning capacitor, C6, until the VFO signal is located. Slug-tuned coil L7, which resonates with C6, should then be adjusted so that an entire sweep of C6 covers the range of 5.0 to 5.5 MHz or slightly greater. While monitoring the collector of Q8 with the rf probe or the oscilloscope, adjust C7 for maximum signal.

At this point, an 80- or 20-meter signal should be generated when the transmitter is keyed. By tuning the monitor receiver to the appropriate amateur band, it should be possible to hear this signal. Tune C6 to place the signal in the center of the amateur band. If an oscilloscope is being used, adjust C11 for maximum signal at the base of Q11. Be careful not to peak this capacitor to a harmonic of the 80- or 20-meter signal or to another undesired frequency output from the mixer. If the 20-meter system is being tested, it is also necessary to peak the band-pass filter at the balanced mixer output for optimum frequency response. With the VFO set to the center of the 20-meter band, adjust C9 for maximum response, using an insulated tool so as not to introduce stray capacitance which could upset the filter tuning. Place the oscilloscope probe at the base of Q12. Next tune C13 for maximum signal amplitude. If an oscilloscope is not available, the above steps may be performed by listening to the radiated signal in a general-coverage receiver and using the receiver's S meter as a peak indicator.

The carrier null pot, R5, is adjusted to minimize any 5-MHz VFO signal appearing at the output of the balanced mixer. First disable the 9-MHz heterodyne oscillator by removing Y1 or Y2 from the socket. With the oscilloscope attached to the mixer output at pin 6 of the MC1496L, adjust R5 for minimum 5-MHz output. If an oscilloscope is not available R5 should be set to the middle of its range. Good suppression of the 5-MHz signal will result at the transmitter output in conjunction with the filtering in the following stages. It is difficult to perform this step with just a receiver alone since direct pickup from the VFO will obscure any null indication. Y1 or Y2 should be replaced in its socket.

A check should be made with the VOM or VTVM to assure that the dc voltage drop across R6 is approximately 4.4 volts. This indicates proper quiescent collector current at Q12 for linear operation.

Connect the power amplifier stage to the 12-volt supply but do not key the transmitter. R9 should temporarily be set at maximum resistance first. The VOM is temporarily inserted in series with the collector lead of Q13 to measure collector current. R9 should slowly be decreased to the point where the static collector current (no rf signal input) reaches 35 mA. The VOM is removed, the collector lead reconnected, and R9 left at this setting. If a silicon diode other than an HEP RO130 is used for CR2, it may be necessary to change the values of R8 and R9 to achieve the proper quiescent collector current. Some experimentation will determine the correct values, keeping in mind the resistor power dissipation requirements.

Finally, connect a 50-ohm dummy load to the antenna connector and a high-impedance micro-

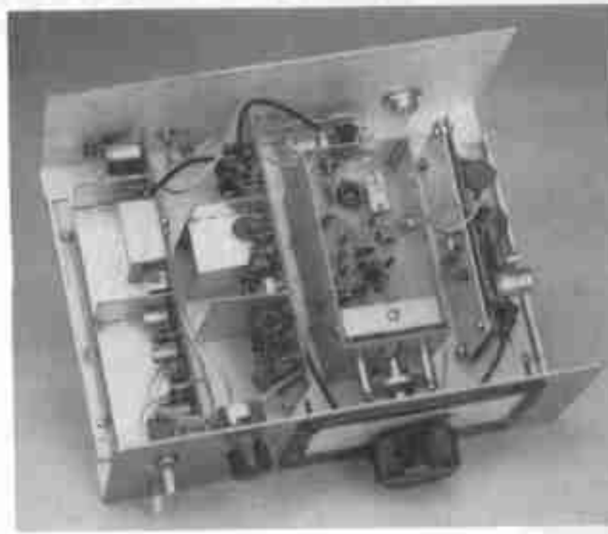


Fig. 3 – View of the inside of the transmitter. The two main circuit boards are mounted vertically on the left side. Note the use of shields (made from double-clad circuit board) to provide isolation between stages on the boards. The VFO, the third circuit board assembly from the left, is housed in a four-sided enclosure fashioned from four pieces of circuit board. The power output amplifier board mounts over the sheet aluminum heat sink which is also mounted vertically on the chassis. Transmit-receive relay K1 is just visible behind the VFO.

phone to the microphone jack. The gain control is adjusted in the same manner as with any conventional ssb transmitter. The setting is best determined with the aid of a Monitorscope at the transmitter output. R1 should never be advanced beyond the point where rf peak flattening begins. A slight readjustment of C2 may be necessary to center the ssb signal properly in the passband of the crystal filter. Improper centering will impair the audio frequency response. This adjustment can be made by listening to oneself in a receiver and tweaking C2 for proper response. Note that the ssb frequency will shift, too, as the setting of C2 is changed. The audio should be clean and free from distortion.

In cw operation, the cw drive control R2 should be brought up to the point where no further increase in rf output results. The break-in delay circuit provides semibreak-in operation for cw and push-to-talk operation for ssb. The setting of time delay adjustment R3 determines the length of time the transmit-receive relay K1 remains closed after the key is released. Adjust R3 for a delay time suitable for the keying speed being used. The spare set of contacts of K1 may be used for muting a receiver during transmit or for switching an outboard power amplifier. The transmitter should never be operated without a proper load at the output.

A set of templates for the four circuit boards is available from ARRL Headquarters, Newington, Conn. 06111 for $1.

A SOLID-STATE TRANSCEIVER FOR 160 METERS

This ssb transceiver is suitable for QRP operation from batteries or as a main frame for fixed-station use. Its circuitry is simple enough to permit easy duplication (or substitution of components where necessary) by proficient builders with only limited experience in solid-state design.

Some 160 Notes

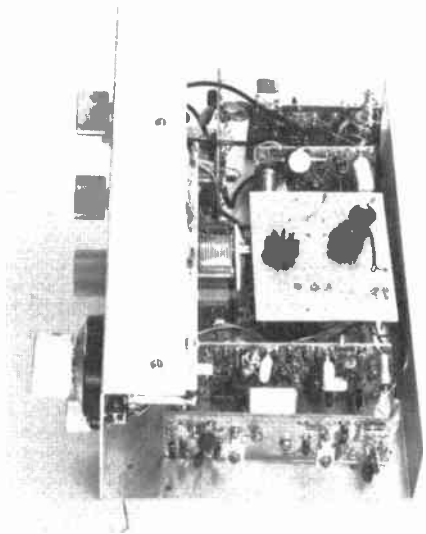
Technically speaking, 160 meters is interesting since it is the only amateur band in the mf range. Phone operation is similar to that encountered on the hf bands but the use of cw is somewhat different. Split-frequency operation is common and one should avoid transmitting within the DX "window" from 1825 to 1830 kHz when the band is open. While cw operation is possible with a transceiver, the above precaution should be noted. Because of the LORAN (Long Range Navigation) service, the band is split up according to geographical area and one should observe the frequency range and power limit for his region (See Chapter 1).

LORAN, proximity to the broadcast band, QRN, and interference from TV sets often imposes severe requirements on receiving devices for this band. While little can be done with sky-wave signals, experimentation with various antenna systems can reduce local interference to a great extent. Proximity and orientation of the antenna to the interfering source are the prime factors here. Because of latter consideration, separate transmitting and receiving antennas may be necessary. Hf-band dipoles, even though they may be electrically short on 160 meters, can still make excellent receiving antennas if a balancing network is used. The balancing transformer (T1) shown in Fig. 1 can be used for both transmitting and receiving, thus reducing ground-loop currents. A simple loading coil in one side of the feed line can be used to tune out the antenna capacitive reactance.

Adequate front-end selectivity is also necessary to assure that unwanted rf energy is rejected *before* it reaches the active elements in the receiving section of the transceiver. The preselector shown in Fig. 1 may be built from readily available parts. Some experimentation with the number of turns on L1 in receive-only applications may be necessary. Use the minimum number of turns that give sufficient sensitivity without signs of overloading. This preselector could also be used with existing receivers with inadequate front-end selectivity on 160.

Circuit Details

The circuit diagram of the transceiver is shown in Fig. 1 and Figs. 3 through 8 incl. The block diagram and switching logic of the transceiver are shown in Fig. 2. This arrangement eliminates the need for relays and provides excellent isolation around the 9-MHz filter board. The full capabilities of a good receiving filter may be reduced considerably by undesirable stray paths. Rf energy rejected by the filter goes around it through the unwanted paths. In the receive position, signals from 1.8 to 2 MHz are mixed with the LO (10.8 to 11 MHz) to give a 9-MHz i-f. Greater bandspread can be achieved by using a smaller value for C10 and increasing L5 or C11. This would reduce the band coverage, however. In the transmit position,



Interior view of the transceiver.

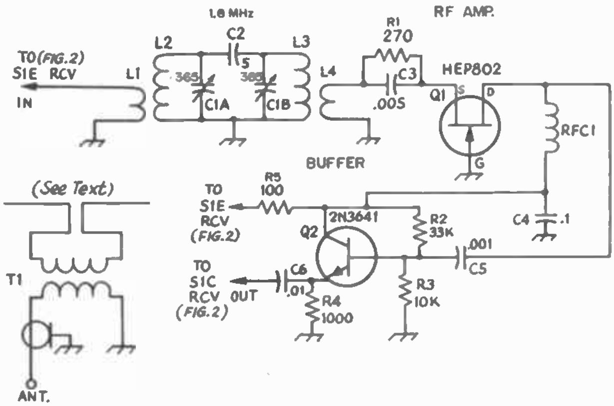


Fig. 1 — Schematic diagram of the rf amplifier and preselector. In this and succeeding diagrams, component designations not mentioned in the captions are for text and layout references only. Unless otherwise noted, resistors are 1/4- or 1/2-watt composition and capacitors are disk ceramic.

C1 — Air variable, 365 pF per section (J.W. Miller 2112 or equiv.).

L1, L4 — 2 turns of plastic-coated wire over cold ends of L2 and L3 respectively.

L2, L3 — Modified Ferri-Tenna Coil (Radio Shack No. 270-1430). Remove coupling coil and all but 35 turns of fine wire on core (see text).

RFC1 — 2.5 mH rf choke pc-board mounting type (Millen J302-2500).

T1 — 40 turns over Amidon T-68-3 toroid (gray core) of bifilar-wound No. 26 enamel wire.

the same mixer is used but rf energy from the balanced modulator and filter board at 9-MHz is converted to the 1.8-MHz band.

Because of the relationship between the LO and the i-f, a sideband inversion occurs. This means that the carrier oscillator crystals will be opposite that usually marked on the filter package. Cw operation is in the usb mode and both carrier-oscillator and VFO offset is used. The carrier-oscillator offset pulls the crystal frequency into the passband of the filter slightly, while the VFO offset can be adjusted for the desired tone on receiving. Keying is accomplished by unbalancing the 1496 IC balanced modulator. Waveshape is determined by the time constant of R62 and C59 in Fig. 7.

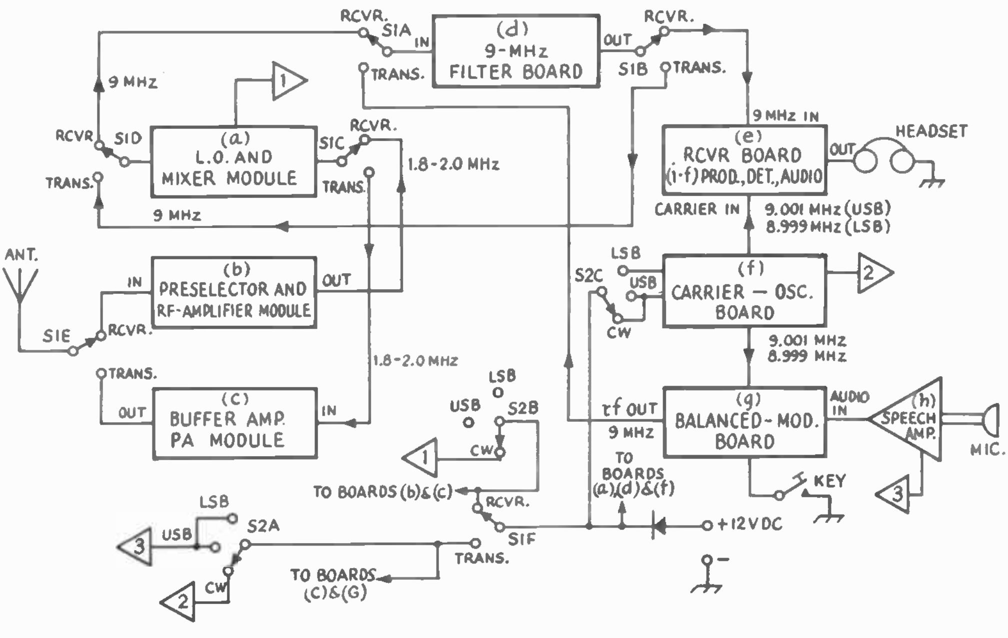


Fig. 2 — Block diagram & switching logic of the transceiver.

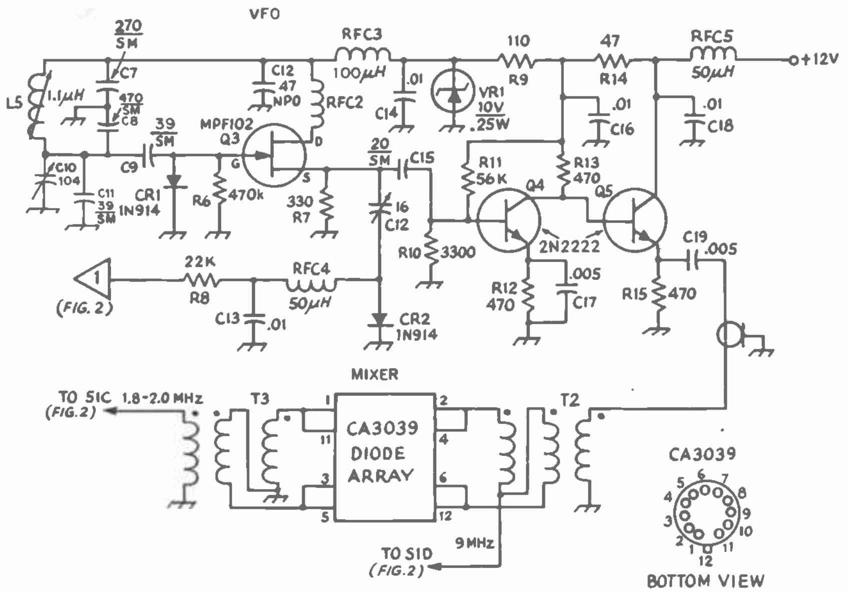


Fig. 3 — Schematic diagram of LO and mixer module. If greater bandspread is desired, a smaller value capacitor could be substituted for C10 with C11 increased by an appropriate amount to set the low-frequency end of the tuning range to 10.8 MHz.

C10 — Air variable, 104 pF maximum (J.W. Miller 2101 or equiv.).

L5 — 1.1-μH slug tuned (Millen 69054-0.91 or equiv.).

RFC2 — Three Amidon ferrite beads at drain terminal of Q3. Install on 1/2-inch length of No. 24 bare wire.

RFC3, RFC4, RFC5 — Miniature 50-μH choke (Millen Co. J300-50).

T2, T3 — 25 turns No. 28 (trifilar wound) on Amidon T-50-3 toroid core.

The low-pass filter shown in Fig. 8 is used to eliminate unwanted rf energy (LO, carrier oscillator, and other products) above 2 MHz before going to the buffer transistor Q11. While various transistors are suitable for cw service in the hf range, many will not perform well as linear power amplifiers. The variation in transistor current gain over a large dynamic range is too great. This results in distortion or imposes severe biasing problems. Generally speaking, uhf types are the best ones to use. The amplifier used with the transceiver is capable of approximately one-watt output with good IMD characteristics.

Construction

A modular-type layout was used that allows the builder to pretest various sections of the transceiver *before* installation in the cabinet. Single-sided pc board or Vectorbord construction should be avoided since unwanted capacitive and inductive coupling may cause spurious oscillations. Use double-sided pc board, or, as in the case of the unit shown, isolated-pad construction. The latter is highly recommended. The individual boards are then mounted in the cabinet with small "L" brackets or in the case of the VFO module, with screws.

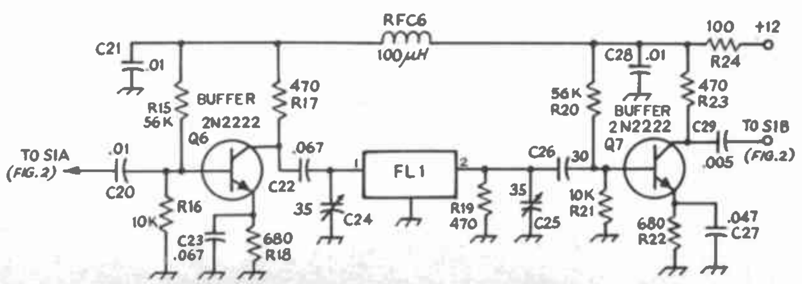


Fig. 4 — The 9-MHz filter board. Physical layout should keep input and output leads separated.

C22, C25 — 3- to 35- pF mica compression trimmer.

RFC6 — Miniature 100-μH choke (Millen Co. J300-100).

FL1 — 9-MHz crystal filter, 2.1-kHz bandwidth (KVG XF-9B Spectrum International, Box 87, Topsfield, MA 01983).

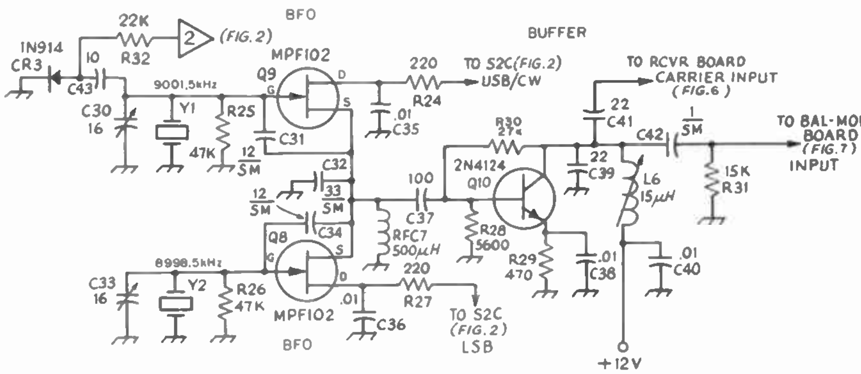


Fig. 5 — Carrier oscillator board.
C30, C33 — Miniature pc-mount air variable (Johnson 189-506-5, Allied Electronics 828-1219).
L6 — 15 µH nominal (Miller 4506 or equiv.).
RFC7 — 500-µH rf choke (Millen J300-500).
Y1, Y2 — KVG matching crystals for FL1.

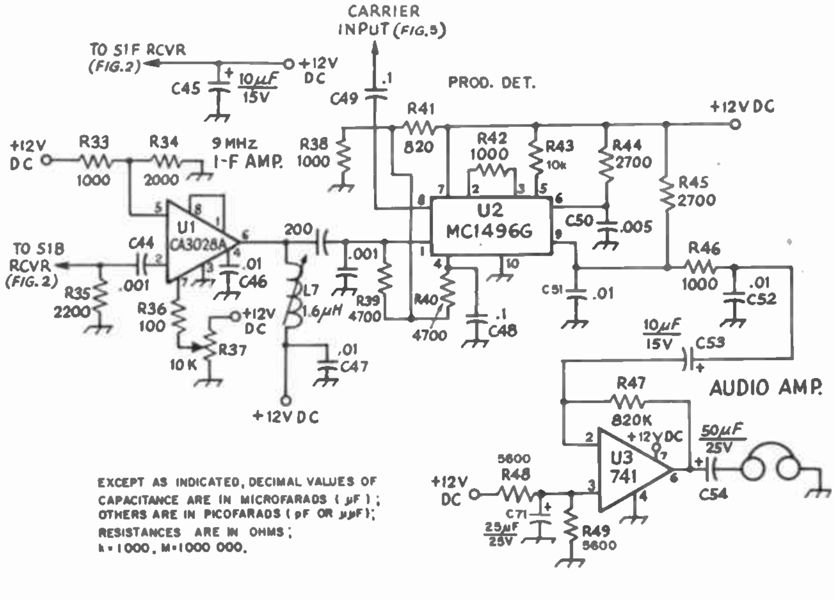
Where interconnecting shielded cables are used (such as the connections on S1 and other rf leads), small coaxial cable is ideal. RG-174/U was used in the unit shown and it is good practice to tie the ground leads to one point where two or more cables come together. An example would be the switch connections at S1. Regular hook-up wire can be used for the power-supply leads going to each board.

While the general layout should not be critical, the one shown in the photograph is suggested. The cabinet is a Ten-Tec MW-10 and the dial assembly can be obtained from Allied/Radio Shack. The rootary switches for S1 and S2 are surplus miniature types with glass-epoxy insulation. The size of the various components available will determine the final layout but care should be taken to keep all leads as short as possible.

It is a good idea to start with the receiver portion of the transceiver (the rf amplifier and preselector is the simplest module to build). Carefully unwind (and save) the wire from the two ferrite-loop antenna coils.

Wind a one-layer coil (35 turns) back on each form and solder it in place. Paint each coil with *Q* dope to keep the turns from unwinding. Mount the completed coils (L2 and L3) using heavy wire leads on the 365-pF capacitor as shown in the photograph. L1 and L4 consist of 2 turns of hook up wire wound on the cold end of L2 and L3 respectively. Next, lay out the circuit board for the

Fig. 6 — Receiver board. This includes the i-f amplifier, product detector, and audio amplifier. Audio power is sufficient for high-impedance earphones.
L7 — Slug-tuned inductor, 1.6 µH nominal, 13 turns No. 26 enam. on 1/4-inch form.



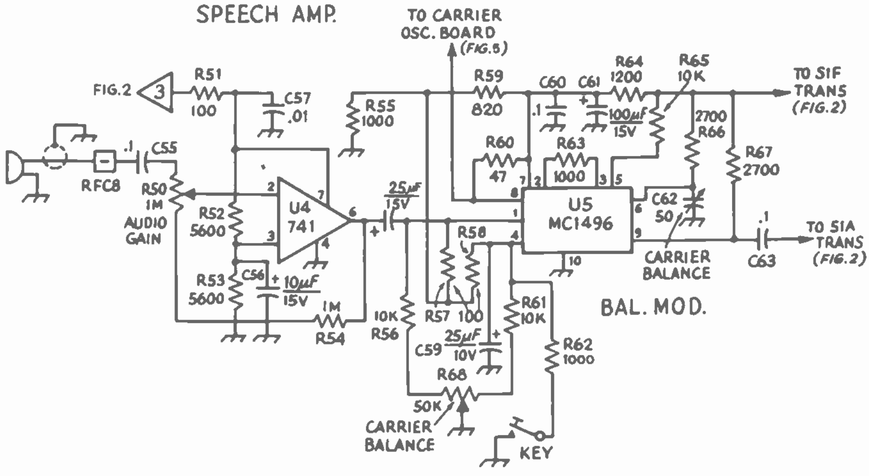


Fig. 7 — Schematic diagram of the speech amplifier and the balanced modulator boards.
C62 — Mica Compression trimmer, 50 pF.
R52, R68 — Control, pc-mounting type.
RFC8 — 3 ferrite beads over microphone-input lead.

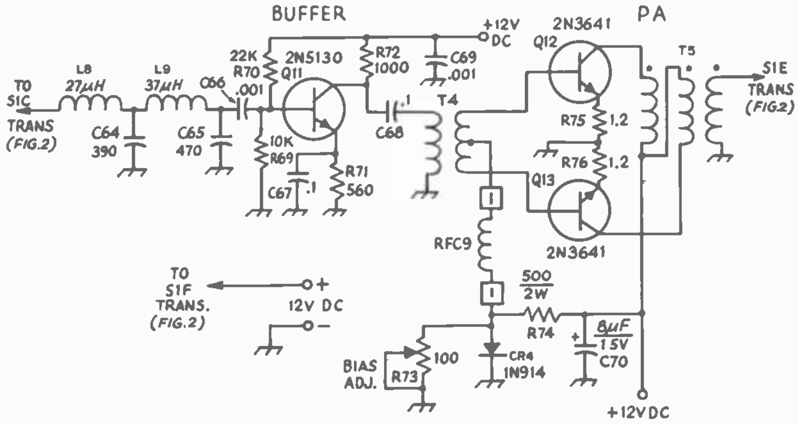
rf amplifier, making it small enough to mount on the back of the capacitor with spacers and screws. Layout for this board (and the remaining ones) will be successful if the following rules are observed. First, keep all component leads as short as possible (especially IC leads) and second, lay out the stages in a straight line as shown in the photograph. Also assure that input and output leads are kept as far away from each other as is practical. If the isolated-pad construction technique is used, a drill press (bench style) is handy. However, either a hand-held electric drill or a crank-type hand drill may be used. Once the preselector module is completed, perform the alignment procedure before going on to the next board. Complete and test the remaining boards before mounting them permanently in the cabinet.

Alignment

While the transceiver could be tested after it is completed, the procedure outlined here will assure each module is working before the next one is mounted in the cabinet. Necessary test equipment includes a signal source and receiver covering 1.8 to 2.0 MHz, and 9 to 11 MHz. The receiver should be capable of receiving ssb signals. Other suggested equipment would be a VTVM, a monitor scope which can be used with the receiver to check modulation, and a frequency counter.

The preselector module should be aligned first. Connect a signal source to the general-coverage receiver and tune in the signal. Next, connect the preselector between the generator and the receiver and adjust the slugs until the signals peak. For correct alignment, C1 should be fully meshed at the low end (1.8 MHz) of the band. The VFO should be adjusted by setting its range for 10.8 to 11 MHz as indicated on either a general-coverage receiver or a frequency counter. The preselector and LO/mixer modules may be mounted inside the cabinet and interconnected. See blocks (a) and (b) in Fig. 2. The external receiver should be connected to the output of T2. When power is applied to the transceiver and S1 is set for RCV, signals and noise should be heard at 9 MHz as the VFO and preselector are tuned. The 9-MHz filter board should be installed and the receiver connection moved from T2 to the output of the filter. See block (d) in Fig. 2. Peak C24 and C25 for maximum signal. The carrier-oscillator board may be checked by listening with the general-coverage receiver to the two crystal frequencies (8.999 and 9.001). Mount the carrier-oscillator and receiver boards, connect a headphone set and adjust L7 for maximum receiver sensitivity. This completes the alignment of the receiver. See block (c) in Fig. 2 for details.

Refer to block (h) in Fig. 2 and mount the speech amplifier. Install the appropriate power, input and output connections. Couple a headset to the output of this circuit through a 0.5-µF capacitor and speak into the microphone. Speech should be heard. Install and connect the balanced-modulator board. Refer to block (g) in Fig. 2. Ssb signals should be detected at the output terminal of T3. Adjust R68 and C62 for minimum carrier. Interconnect the buffer and PA modules, and connect a dummy load (with an output indicator) to the antenna jack. A small pilot light (No. 47) will suffice if the PA shown in Fig. 8 is used. R73 should be set for minimum collector current. A short whistle into the microphone should produce



an output signal. Clear-sounding ssb signals should be heard when listening to the general-coverage receiver. This completes the ssb alignment.

Place a jumper from either the USB or LSB position of S2A to the CW position of S2A. Set the general-coverage receiver to the USB position. Turn the transceiver to the CW position and tune until a readable ssb signal is heard. Key the transceiver and depending upon the settings of C12 and C30, a tone should be heard. C30 will determine the amount of output. Adjust C12 until the desired sidetone is obtained. This will require retuning the receiver for readable usb after each adjustment. When the adjustment is correct, a proper-sounding ssb signal can be heard in the CW position and the desired note will also be heard when the transmitter is keyed. Remove the jumper from S2A. This completes alignment of the transceiver.

Fig. 8 — Schematic diagram of buffer and PA. If a broad-band amplifier or antenna circuit is to follow T5, a low-pass filter may be necessary to reduce unwanted harmonic energy.
L8 — 27 µH, 66 turns of No. 30 enam. wire on Amidon T-50-3 (gray) toroid core.
L9 — 37 µH, 76 turns No. 30 enam. on T-50-3 core.
R73 — Control, pc-mounting type.
RFC9 — 2.7 µH minimum. Slip a ferrite bead over each end of a small rf choke (Millen 34300).
T4 — Stack two Amidon Husky (7 mm) beads and wind a 5-turn primary and a 3 turn secondary through both cores. Use No. 26 enam. wire. Make a second transformer similar to the first one. Parallel the primaries, and series connect the secondaries observing the polarities shown on the diagram.
T5 — 24 turns No. 26 enam. wire (trifilar wound) on Amidon T-68-3 core.

Frequency Modulation and Repeaters

Methods of radiotelphone communication by frequency modulation were developed in the 1930s by Major Edwin Armstrong in an attempt to reduce the problems of static and noise associated with receiving a-m broadcast transmissions. The primary advantage of fm, the ability to produce a high signal-to-noise ratio when receiving a signal of only moderate strength, has made fm the mode chosen for mobile communications services and quality broadcasting. The disadvantages, the wide bandwidth required and the poor results obtained when an fm signal is propagated via the ionosphere (because of phase distortion), has limited the use of frequency modulation to the 10-meter band and the vhf/uhf section of the spectrum.

Fm has some impressive advantages for vhf operation, especially when compared to a-m. With fm the modulation process takes place in a low-level stage. The modulation equipment required is the same, regardless of transmitter power. The signal may be frequency multiplied after modulation, and the PA stage can be operated Class C for best efficiency, as the "final" need not be linear.

In recent years there has been increasing use of fm by amateurs operating around 29.6 MHz in the 10-meter band. The vhf spectrum now in popular use includes 52 to 53 MHz, 146 to 147.5 MHz, 222 to 225 MHz, and 440 to 450 MHz.

FREQUENCY AND PHASE MODULATION

It is possible to convey intelligence by modulating any property of a carrier, including its frequency and phase. When the frequency of the carrier is varied in accordance with the variations in a modulating signal, the result is frequency modulation (fm). Similarly, varying the phase of the carrier current is called phase modulation (pm).

Frequency and phase modulation are not independent, since the frequency cannot be varied without also varying the phase, and vice versa.

Fig. 14-1 — The use of vhf fm mobile rigs in conjunction with repeaters has improved the communications of many amateur emergency groups. Here F2BQ relays traffic being received via 2-meter fm on a 40-meter ssb link.

The effectiveness of fm and pm for communication purposes depends almost entirely on the receiving methods. If the receiver will respond to frequency and phase changes but is insensitive to amplitude changes, it will discriminate against most forms of noise, particularly impulse noise such as is set up by ignition systems and other sparking devices. Special methods of detection are required to accomplish this result.

Modulation methods for fm and pm are simple and require practically no audio power. There is also the advantage that, since there is no amplitude variation in the signal, interference to broadcast reception resulting from rectification of the transmitted signal in the audio circuits of the bc receiver is substantially eliminated.

Frequency Modulation

Fig. 14-2 is a representation of frequency modulation. When a modulating signal is applied, the carrier frequency is increased during one half cycle of the modulating signal and decreased during the half cycle of opposite polarity. This is indicated in the drawing by the fact that the rf cycles occupy less time (higher frequency) when the modulating signal is positive, and more time (lower frequency) when the modulating signal is negative. The change in the carrier frequency (frequency deviation) is proportional to the instantaneous amplitude of the modulating signal, so the deviation is small when the instantaneous amplitude of the modulating signal is small, and is greatest when the modulating signal reaches its peak, either positive or negative.

As shown by the drawing, the amplitude of the signal does not change during modulation.

Phase Modulation

If the phase of the current in a circuit is changed there is an instantaneous frequency change during the time that the phase is being shifted. The amount of frequency change, or deviation, depends on how rapidly the phase shift is accomplished. It is also dependent upon the total amount of the phase shift. In a properly operating

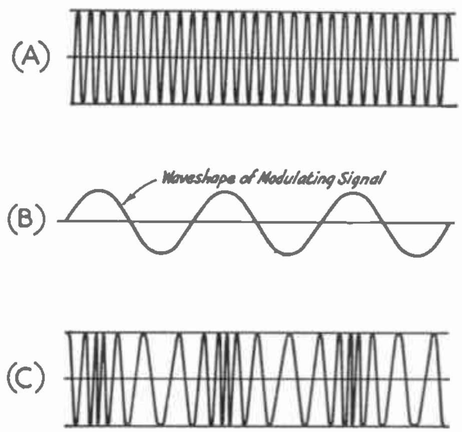


Fig. 14-2 — Graphical representation of frequency modulation. In the unmodulated carrier at A, each rf cycle occupies the same amount of time. When the modulating signal, B, is applied, the radio frequency is increased and decreased according to the amplitude and polarity of the modulating signal.

pm system the amount of phase shift is proportional to the instantaneous amplitude of the modulating signal. The rapidity of the phase shift is directly proportional to the frequency of the modulating signal. Consequently, the frequency deviation in pm is proportional to both the amplitude and frequency of the modulating signal. The latter represents the outstanding difference between fm and pm, since in fm the frequency deviation is proportional only to the amplitude of the modulating signal.

FM and PM Sidebands

The sidebands set up by fm and pm differ from those resulting from a-m in that they occur at integral multiples of the modulating frequency on either side of the carrier rather than, as in a-m, consisting of a single set of side frequencies for each modulating frequency. An fm or pm signal therefore inherently occupies a wider channel than a-m.

The number of "extra" sidebands that occur in fm and pm depends on the relationship between the modulating frequency and the frequency deviation. The ratio between the frequency deviation, in Hertz, and the modulating frequency, also in Hertz, is called the modulating index. That is,

$$\textit{Modulation index} = \frac{\textit{Carrier frequency deviation}}{\textit{Modulating frequency}}$$

> ***Example:*** **The maximum frequency deviation in an f.m. transmitter is 3000 Hz. either side of the carrier frequency. The modulation index when the modulating frequency is 1000 Hz. is**
>
> $$\textbf{Modulation index} = \frac{3000}{1000} = 3$$
>
> **At the same deviation with 3000-Hz. modulation the index would be 1; at 100 Hz. it would be 30, and so on.**

In pm the modulation index is constant regardless of the modulating frequency; in fm it varies with the modulating frequency, as shown in the above example. In an fm system the ratio of the *maximum* carrier-frequency deviation to the *highest* modulating frequency used is called the deviation ratio.

Fig. 14-3 shows how the amplitudes of the carrier and the various sidebands vary with the modulation index. This is for single-tone modulation; the first sideband (actually a pair, one above and one below the carrier) is displaced from the carrier by an amount equal to the modulating frequency, the second is twice the modulating frequency away from the carrier, and so on. For example, if the modulating frequency is 2000 Hz and the carrier frequency is 29,500 kHz, the first sideband pair is at 29,498 kHz and 29,502 kHz, the second pair is at 29,496 kHz and 29,504 kHz, the third at 29,494 kHz and 29,506 kHz, etc. The amplitudes of these sidebands depend on the modulation index, not on the frequency deviation.

Note that as shown by Fig. 14-3, the carrier strength varies with the modulation index. (In amplitude modulation the carrier strength is constant; only the sideband amplitude varies.) At a modulation index of approximately 2.4 the carrier disappears entirely. It then becomes "negative" at a higher index, meaning that its phase is reversed as compared to the phase without modulation. In fm and pm the energy that goes into the sidebands is taken from the carrier, the *total* power remaining the same regardless of the modulation index.

Since there is no change in amplitude with modulation, an fm or pm signal can be amplified without distortion by an ordinary Class C amplifier. The modulation can take place in a very low-level stage and the signal can then be amplified by either frequency multipliers or straight-through amplifiers.

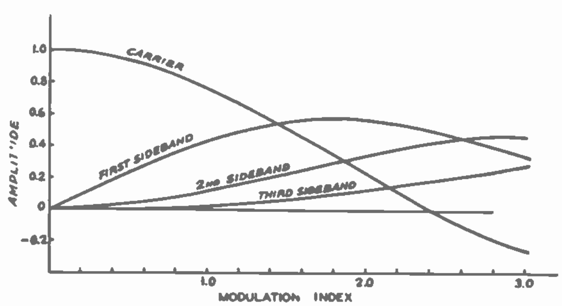


Fig. 14-3 — How the amplitude of the pairs of sidebands varies with the modulation index in an fm or pm signal. If the curves were extended for greater values of modulation index it would be seen that the carrier amplitude goes through zero at several points. The same statement also applies to the sidebands.

If the modulated signal is passed through one or more frequency multipliers, the modulation index is multiplied by the same factor that the carrier frequency is multiplied. For example, if modulation is applied on 3.5 MHz and the final output is on 28 MHz, the total frequency multiplication is 8 times, so if the frequency deviation is 500 Hz at 3.5 MHz it will be 4000 Hz at 28 MHz. Frequency multiplication offers a means for obtaining practically any desired amount of frequency deviation, whether or not the modulator itself is capable of giving that much deviation without distortion.

Bandwidth

FCC amateur regulations (Part 97.61) limit the bandwidth of F3 (frequency and phase modulation) to that of an a-m transmission having the same audio characteristics below 29.0 MHz and in the 50.1- to 52.5-MHz frequency segment. Greater bandwidths are allowed from 29.0 to 29.7 MHz and above 52.5 MHz.

If the modulation index (with single-tone modulation) does not exceed 0.6 or 0.7, the most important extra sideband, the second, will be at least 20 dB below the unmodulated carrier level, and this should represent an effective channel width about equivalent to that of an a-m signal. In the case of speech, a somewhat higher modulation index can be used. This is because the energy distribution in a complex wave is such that the modulation index for any one frequency component is reduced as compared to the index with a sine wave having the same peak amplitude as the voice wave.

The chief advantage of fm or pm for frequencies below 30 MHz is that it eliminates or reduces certain types of interference to broadcast reception. Also, the modulating equipment is relatively simple and inexpensive. However, assuming the same unmodulated carrier power in all cases, narrow-band fm or pm is not as effective as a-m *with the methods of reception used by many amateurs*. To obtain the benefits of the fm mode, a good fm receiver is required. As shown in Fig. 14-3, at an index of 0.6 the amplitude of the first sideband is about 25 percent of the unmodulated-carrier amplitude; this compares with a sideband amplitude of 50 percent in the case of a 100 percent modulated a-m transmitter. When copied on an a-m receiver, a narrow-band fm or pm transmitter is about equivalent to a 100-percent modulated a-m transmitter operating at one-fourth the carrier power. On a suitable (fm) receiver, fm is as good or better than a-m, watt for watt.

Three deviation amounts are now standard practice: 15, 5 and 2.5 kHz, which in the current vernacular of fm users, are known as wide band, narrow band, and sliver band, respectively. (See box above.) The 2.5-3 kHz deviation (called nbfm by OTs) was popular for a time on the vhf bands and 10 meters after World War II. Deviation figures are given for the frequency swing in one direction. The rule-of-thumb for determination of bandwidth requirements for an fm system is:

$$2\,(\Delta F) + F_{\mathrm{Amax}}$$

where ΔF is one half of the total frequency deviation, and F_{Amax} is the maximum audio frequency (3 kHz for communications purposes). Thus, for narrow-band fm, the bandwidth equals (2) 5 + 3 or 13 kHz. Wide-band systems need a 33-kHz receiver bandwidth.

Comparison of FM and PM

Frequency modulation cannot be applied to an amplifier stage, but phase modulation can; pm is therefore readily adaptable to transmitters employing oscillators of high stability such as the crystal-controlled type. The amount of phase shift that can be obtained with good linearity is such that the maximum practicable modulation index is about 0.5. Because the phase shift is proportional to the modulating frequency, this index can be used only at the highest frequency present in the modulating signal, assuming that all frequencies will at one time or another have equal amplitudes. Taking 3000 Hz as a suitable upper limit for voice work, and setting the modulation index at 0.5 for 3000 Hz, the frequency response of the speech-amplifier system above 3000 Hz must be sharply attenuated, to prevent excess splatter. (See Fig. 14-4.) Also, if the "tinny" quality of pm as received on an fm receiver is to be avoided, the pm must be changed to fm, in which the modulation index decreases in inverse proportion to the modulating frequency. This requires shaping the speech-amplifier frequency-response curve in such a way that the output voltage is inversely proportional to frequency over most of the voice range. When this is done the maximum modulation index can only be used to some relatively low audio frequency, perhaps 300 to 400 Hz in voice transmission, and must decrease in proportion to the increase in frequency. The result is that the maximum linear frequency deviation is only one or two hundred Hz, when pm is changed to fm. To increase the deviation for narrow band requires a frequency multiplication of 8 times or more.

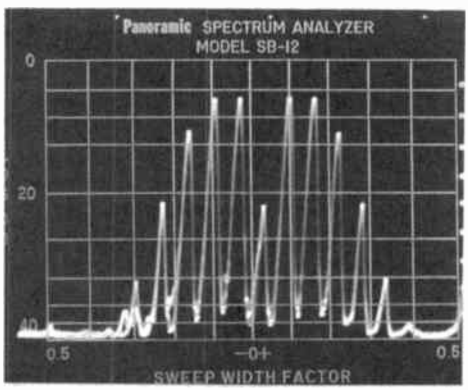


Fig. 14-4 — Output frequency spectrum of a narrow-band fm transmitter modulated by a 1-kHz tone.

It is relatively easy to secure a fairly large frequency deviation when a self-controlled oscillator is frequency-modulated directly. (True frequency modulation of a crystal-controlled oscillator results in only very small deviations and so requires a great deal of frequency multiplication.) The chief problem is to maintain a satisfactory degree of carrier stability, since the greater the inherent stability of the oscillator the more difficult it is to secure a wide frequency swing with linearity.

METHODS OF FREQUENCY MODULATION

Direct FM

A simple and satisfactory device for producing fm in the amateur transmitter is the reactance modulator. This is a vacuum tube or transistor connected to the rf tank circuit of an oscillator in such a way as to act as a variable inductance or capacitance.

Fig. 14-5A is a representative circuit. Gate 1 of the modulator MOSFET is connected across the oscillator tank circuit, C1L1, through resistor R1 and blocking capacitor C2. C3 represents the input capacitance of the modulator transistor. The resistance of R1 is made large compared to the reactance of C3, so the rf current through R1C3 will be practically in phase with the rf voltage appearing at the terminals of the tank circuit. However, the voltage across C3 will lag the current by 90 degrees. The rf current in the drain circuit of the modulator will be in phase with the grid voltage, and consequently is 90 degrees behind the current through C3, or 90 degrees behind the rf tank voltage. This lagging current is drawn through the oscillator tank, giving the same effect as though an inductance were connected across the tank. The frequency increases in proportion to the amplitude of the lagging plate current of the modulator. The audio voltage, introduced through a radio-frequency choke, varies the transconductance of the transistor and thereby varies the rf drain current.

The modulated oscillator usually is operated on a relatively low frequency, so that a high order of carrier stability can be secured. Frequency multipliers are used to raise the frequency to the final frequency desired.

A reactance modulator can be connected to a crystal oscillator as well as to the self-controlled type as shown in Fig. 14-5B. However, the resulting signal can be more phase-modulated than it is frequency-modulated, for the reason that the frequency deviation that can be secured by varying the frequency of a crystal oscillator is quite small.

The sensitivity of the modulator (frequency change per unit change in grid voltage) depends on the transconductance of the modulator transistor. It increases when R1 is made smaller in comparison with C3. It also increases with an increase in L/C ratio in the oscillator tank circuit. However, for highest carrier stability it is desirable to use the largest tank capacitance that will permit the desired deviation to be secured while keeping within the limits of linear operation.

A change in *any* of the voltages on the modulator transistor will cause a change in rf drain current, and consequently a frequency change. Therefore it is advisable to use a regulated power supply for both modulator and oscillator.

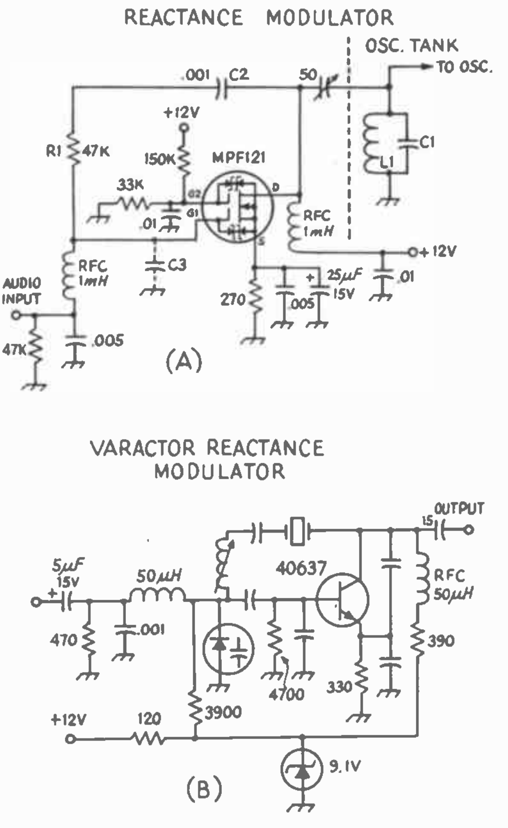


Fig. 14-5 — Reactance modulators using (A) a high-transconductance MOSFET and (B) a varactor diode.

Indirect FM

The same type of reactance-tube circuit that is used to vary the tuning of the oscillator tank in fm can be used to vary the tuning of an amplifier tank and thus vary the phase of the tank current for pm. Hence the modulator circuit of Fig. 14-5A or 14-6A can be used for pm if the reactance transistor or tube works on an amplifier tank instead of directly on a self-controlled oscillator. If audio shaping is used in the speech amplifier, as described above, fm instead of pm will be generated by the phase modulator.

The phase shift that occurs when a circuit is detuned from resonance depends on the amount of detuning and the Q of the circuit. The higher the

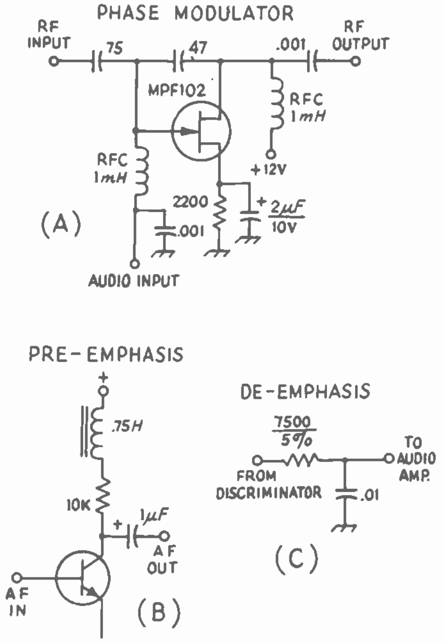


Fig. 14-6 – (A) The phase-shifter type of phase modulator. (B) Pre-emphasis and (C) de-emphasis circuits.

Q, the smaller the amount of detuning needed to secure a given number of degrees of phase shift. If the Q is at least 10, the relationship between phase shift and detuning (in kHz either side of the resonant frequency) will be substantially linear over a phase-shift range of about 25 degrees. From the standpoint of modulator sensitivity, the Q of the tuned circuit on which the modulator operates should be as high as possible. On the other hand, the effective Q of the circuit will not be very high if the amplifier is delivering power to a load since the load resistance reduces the Q. There must therefore be a compromise between modulator sensitivity and rf power output from the modulated amplifier. An optimum figure for Q appears to be about 20; this allows reasonable loading of the modulated amplifier and the necessary tuning variation can be secured from a reactance modulator without difficulty. It is advisable to modulate at a low power level.

Reactance modulation of an amplifier stage usually results in simultaneous amplitude modulation because the modulated stage is detuned from resonance as the phase is shifted. This must be eliminated by feeding the modulated signal through an amplitude limiter or one or more "saturating" stages – that is, amplifiers that are operated Class C and driven hard enough so that variations in the amplitude of the input excitation produce no appreciable variations in the output amplitude.

For the same type of reactance modulator, the speech-amplifier gain required is the same for pm as for fm. However, as pointed out earlier, the fact that the actual frequency deviation increases with the modulating audio frequency in pm makes it necessary to cut off the frequencies above about 3000 Hz before modulation takes place. If this is not done, unnecessary sidebands will be generated at frequencies considerably away from the carrier.

SPEECH PROCESSING FOR FM

The speech amplifier preceding the modulator follows ordinary design, except that no power is taken from it and the af voltage required by the modulator grid usually is small – not more than 10 or 15 volts, even with large modulator tubes, and only a volt or two for transistors. Because of these modest requirements, only a few speech stages are needed; a two-stage amplifier consisting of two bipolar transistors, both resistance-coupled, will more than suffice for crystal ceramic or hi-*Z* dynamic microphones. For more information on speech amplifiers see Chapter 13.

Several forms of speech processing produce worthwhile improvements in fm system performance. It is desirable to limit the peak amplitude of the audio signal applied to an fm or pm modulator, so that the deviation of the fm transmitter will not exceed a preset value. This peak limiting is usually accomplished with a simple audio clipper which is placed between the speech amplifier and modulator. The clipping process produces high-order harmonics which, if allowed to pass through to the modulator stage, would create unwanted sidebands. Therefore, an audio low-pass filter with a cut-off frequency between 2.5 and 3 kHz is needed at the output of the clipper. Excess clipping can cause severe distortion of the voice signal. An audio processor consisting of a compressor and a clipper, such as described in Chapter 13, has been found to produce audio with a better sound (i.e., less distortion) than a clipper alone.

To reduce the amount of noise in some fm communications systems, an audio shaping network called pre-emphasis is added at the transmitter to proportionally attenuate the lower audio frequencies, giving an even spread to the energy in the audio band. This results in an fm signal of nearly constant energy distribution. The reverse is done at the receiver, called de-emphasis, to restore the audio to its original relative proportions. Sample circuits are shown in Fig. 14-6.

FM EXCITERS

Fm exciters and transmitters take two general forms. One, shown at Fig. 14-7A, consists of a reactance modulator which shifts the frequency of an oscillator to generate an fm signal directly. Successive multiplier stages provide output on the desired frequency, which is amplified by a PA stage. This system has a disadvantage in that, if the oscillator is free running, it is difficult to achieve sufficient stability for vhf use. If a crystal-controlled oscillator is employed, unless the amount that the crystal frequency is changed is kept small, it is difficult to achieve equal amounts of frequency swing.

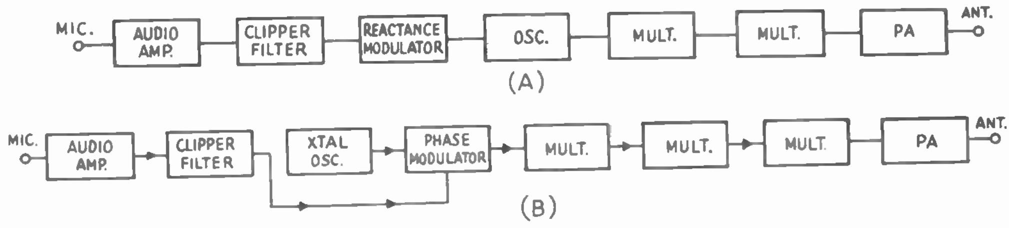


Fig. 14-7 — Block diagrams of typical fm exciters.

The indirect method of generating fm shown in Fig. 14-7B is currently popular. Shaped audio is applied to a phase modulator to generate fm. As the amount of deviation produced is very small, then a large number of multiplier stages is needed to achieve wide-band deviation at the operating frequency. In general, the system shown at A will require a less complex circuit than that at B, but the indirect method (B) often produces superior results.

TESTING AN FM TRANSMITTER

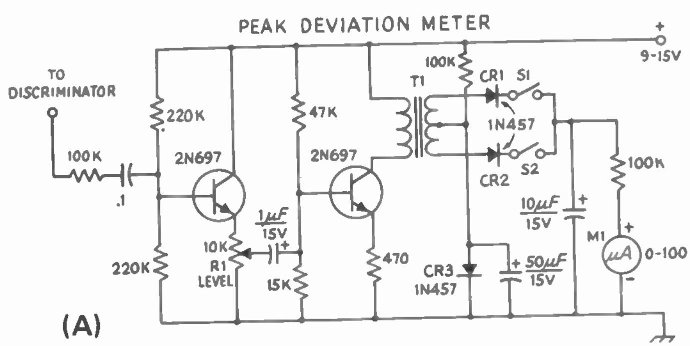
Accurate checking of the operation of an fm or pm transmitter requires different methods than the corresponding checks on an a-m or ssb set. This is because the common forms of measuring devices either indicate amplitude variations only (a milliammeter, for example), or because their indications are most easily interpreted in terms of amplitude.

The quantities to be checked in an fm transmitter are the linearity and frequency deviation and the output frequency, if the unit uses crystal control. The methods of checking differ in detail.

Frequency Checking

The crystal-controlled, channelized operation that is now popular with amateur fm users requires that a transmitter be held close to the desired channel, at least within a few hundred Hertz, even in a wide-band system. Having the transmitter on the proper frequency is particularly important when operating through a repeater. The rigors of mobile and portable operation make a frequency check of a channelized transceiver a good idea at three-month intervals.

Frequency meters generally fall in two categories, the hererodyne type and the digital counter. For amateur use, the vhf/uhf counterparts of the



(B)

Audio Frequency	Deviation Produced 1st Null	2nd Null	3rd Null
905.8 Hz	±2.18 kHz	± 5.00 kHz	± 7.84 kHz
1000.0 Hz	±2.40 kHz	± 5.52 kHz	± 8.65 kHz
1500.0 Hz	±3.61 kHz	± 8.28 kHz	±12.98 kHz
1811.0 Hz	±4.35 kHz	±10.00 kHz	±15.67 kHz
2000.0 Hz	±4.81 kHz	±11.04 kHz	±17.31 kHz
2079.2 Hz	±5.00 kHz	±11.48 kHz	±17.99 kHz
2805.0 Hz	±6.75 kHz	±15.48 kHz	±24.27 kHz

Fig. 14-8 — (A) Schematic diagram of the deviation meter. Resistors are 1/2-watt composition and capacitors are ceramic, except those with polarity marked, which are electrolytic. CR1-CR3, incl. are high-speed silicon switching diodes. R1 is a linear-taper composition control, and S1, S2 are spst toggle switches. T1 is a miniature audio transformer with a 10,000-ohm primary and 20,000-ohm center-tapped secondary (Triad A31X). (B) Chart of audio frequencies which will produce a carrier null when the deviation of an fm transmitter is set for the values given.

popular BC-221 frequency meter, the TS-174 and TS-175, will provide sufficient accuracy. Frequency counters that will work directly up to 500 MHz and higher are available, but their cost is high. The less expensive low-frequency counters can be employed using a scaler, a device which divides an input frequency by a preset ratio, usually 10 or 100. The Heathkit IB-102 scaler may be used up to 175 MHz, using a counter with a 2-MHz (or more) upper frequency limit. If the counting system does not have a sufficient upper frequency limit to measure the output of an fm transmitter directly, one of the frequency-multiplier stages can be sampled to provide a signal in the range of the measurement device. Alternatively, a crystal-controlled converter feeding an hf receiver which has accurate frequency readout can be employed, if a secondary standard is available to calibrate the receiving system.

Deviation and Deviation Linearity

A simple deviation meter can be assembled following the diagram of Fig. 14-8A. This circuit was designed by K6VKZ. The output of a wide-band receiver discriminator (before any de-emphasis) is fed to two amplifier transistors. The output of the amplifier section is transformer coupled to a pair of rectifier diodes to develop a dc voltage for the meter, M1. There will be an indication on the meter with no signal input because of detected noise, so the accuracy of the instrument will be poor on weak signals.

To calibrate the unit, signals of known deviation will be required. If the meter is to be set to read 0-15 kHz, then a 7.5-kHz deviation test signal should be employed. R1 is then adjusted until M1 reads half scale, 50 μA. To check the peak deviation of an incoming signal, close both S1 and S2. Then, read the meter. Opening first one switch and then the other will indicate the amount of positive and negative deviation of the signal, a check of deviation linearity.

Measurement of Deviation Using Bessel Functions

Using a math. relationship known as the Bessel Function it is possible to predict the points at which, with certain audio-input frequencies and predetermined deviation settings, the carrier output of an fm transmitter will disappear completely. Thus, by monitoring the carrier frequency with a receiver, it will be possible by ear to identify the deviation at which the carrier is nulled. A heterodyne signal at either the input or receiver i-f is required so that the carrier will produce a beat note which can easily be identified. Other tones will be produced in the modulation process, so some concentration is required by the operator when making the test. With an audio tone selected from the chart (Fig. 14-8B), advance the deviation control slowly until the first null is heard. If a higher-order null is desired, continue advancing the control further until the second, and then the third, null is heard. Using a carrier null beyond the third is generally not practical.

For example, if a 905.8-Hz tone is used, the transmitter will be set for 5-kHz deviation when the second null is reached. The second null achieved with a 2805-Hz audio input will set the transmitter deviation at 15.48 kHz. The Bessel-function approach can be used to calibrate a deviation meter, such as the unit shown in Fig. 14-8A

RECEPTION OF FM SIGNALS

Receivers for fm signals differ from others principally in two features – there is no need for linearity preceding detection (it is, in fact, advantageous if amplitude variations in signal and background noise can be "washed out") and the detector must be capable of converting frequency variations in the incoming signal into amplitude variations.

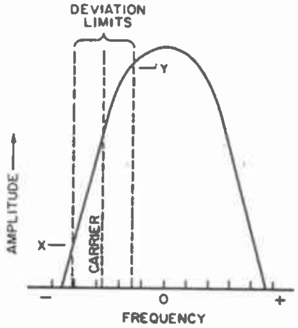


Fig. 14-9 – Fm detector characteristics. Slope detection, using the sloping side of the receivers selectivity curve to convert fm to a-m for subsequent detection.

Frequency-modulated signals can be received after a fashion on any ordinary receiver. The receiver is tuned to put the carrier frequency partway down on one side of the selectivity curve. When the frequency of the signal varies with modulation it swings as indicated in Fig. 14-9, resulting in an a-m output varying between X and Y. This is then rectified as an a-m signal.

With receivers having steep-sided selectivity curves, the method is not very satisfactory because the distortion is quite severe unless the frequency deviation is small, since the frequency deviation and output amplitude is linear over only a small part of the selectivity curve.

The FM Receiver

Block diagrams of an a-m/ssb and an fm receiver are shown in Fig. 14-10. Fundamentally, to achieve a sensitivity of less than one microvolt, an fm receiver requires a gain of several million – too much total gain to be accomplished with stability on a single frequency. Thus, the use of the

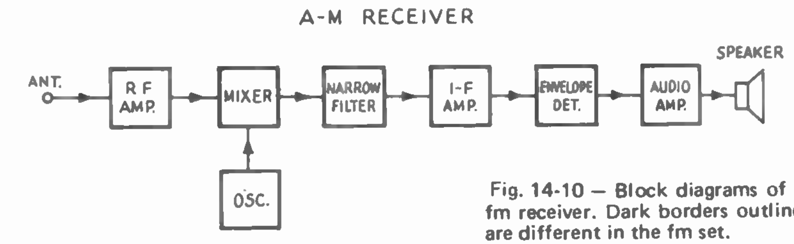
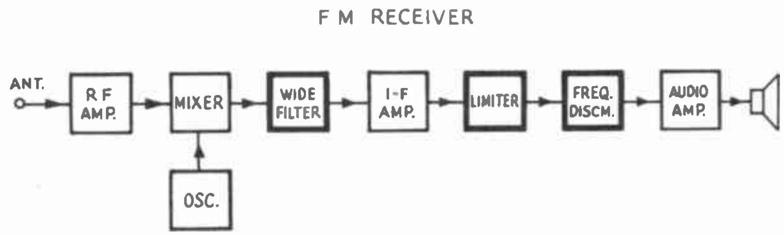


Fig. 14-10 — Block diagrams of (A) an a-m (B) an fm receiver. Dark borders outline the sections that are different in the fm set.



superheterodyne circuit has become standard practice. Three major differences will be apparent from a comparison of the two block diagrams. The fm receiver employs a wider-bandwidth filter, a different detector, and has a limiter stage added between the i-f amplifier and the detector. Otherwise the functions, and often the circuits, of the rf, oscillator, mixer and audio stages will be the same in either receiver.

In operation, the noticeable difference between the two receivers is the effect of noise and interference on an incoming signal. From the time of the first spark transmitters, "rotten QRM" has been a major problem for amateurs. The limiter and discriminator stages in an fm set can eliminate a good deal of impulse noise, except that noise which manages to acquire a frequency-modulation characteristic. Accurate alignment of the receiver i-f system and phase tuning of the detector are required to achieve good noise suppression. Fm receivers perform in an unusual manner when QRM is present, exhibiting a characteristic known as the *capture effect.* The loudest signal received, even if it is only two or three times stronger than other stations on the same frequency, will be the only transmission demodulated. By comparison, an S9 a-m or cw signal can suffer noticeable interference from an S2 carrier.

Bandwidth

Most fm sets that use tubes achieve i-f selectivity by using a number of overcoupled transformers. The wide bandwidth and phase-response characterisitic needed in the i-f system dictate careful design and alignment of all interstage transformers.

F M FILTERS

Manufacturer	*Model*	*Center Frequency*	*Nominal Bandwidth*	*Ultimate Rejection*	*Impedance (r) In*	*Out*	*Insertion Loss*	*Crystal Discriminator*
KVG (1)	XF-9E	9.0 MHz	12 kHz	90 dB	1200	1200	3 dB	XD9-02
KVG (1)	XF-107A	10.7 MHz	12 kHz	90 dB	820	820	3.5 dB	XD107-01
KVG (1)	XF-107B	10.7 MHz	15 kHz	90 dB	910	910	3.5 dB	XD107-01
KVG (1)	XF-107C	10.7 MHz	30 kHz	90 dB	2000	2000	4.5 dB	XD107-01
Heath Dynamics (2)	–	21.5 MHz	15 kHz	90 dB	550	550	3 dB	–
Heath Dynamics (2)	–	21.5 MHz	30 kHz	90 dB	1100	1100	2 dB	–
E.S. (3)	FB-6D	10.7 MHz	15 kHz	80 dB	950	950	2 dB	AB-1C
E.S. (3)	10-MA	10.7 MHz	30 kHz	80 dB	2000	2000	4 dB	AB-1C
E.S. (3)	EL-3A	11.5 MHz	36 kHz	70 dB	50	50	4 dB	AL-1
E.S. (3)	DR-9	21.4 MHz	20 kHz	40 dB	750	750	5 dB	AR-10
Clevite (4)	TCF4-12D3CA	455 kHz	12 kHz	60 dB	40k	2200	6 dB	–
Clevite (4)	TCF4-18G45A	455 kHz	18 kHz	50 dB	40k	2200	6 dB	–
Clevite (4)	TCF6-30D55A	455 kHz	30 kHz	60 dB	20k	1000	5 dB	–

Fig. 14-11 — A list of fm-bandwidth filters that are available to amateurs. Manufacturer's addresses are as follows:

1) Spectrum International, P. O. Box 87, Topsfield, MA 01983.
2) Heath Dynamics, Inc., 6050 N. 52nd Avenue, Glendale, AZ 85301.
3) E. S. Electronic Labs, 301 Augustus, Excelsior Springs, MO 64024.
4) Semiconductor Specialists, Inc., P. O. Box 66125, O'Hare International Airport, Chicago, IL 60666. (Minimum order $5.00.)

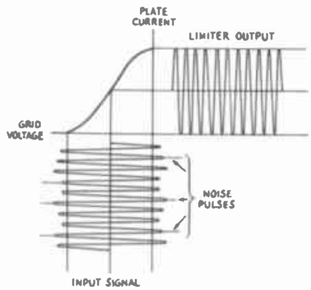


Fig. 14-12 — Representation of limiter action. Amplitude variations on the signal are removed by the diode action of the grid- and plate-current saturation.

For the average ham, the use of a high-selectivity filter in a homemade receiver offers some simplification of the alignment task. Following the techniques used in ssb receivers, a crystal or ceramic filter should be placed in the circuit as close as possible to the antenna connector – at the output of the first mixer, in most cases. Fig. 14-11 lists a number of suitable filters that are available to amateurs. Prices for these filters are in the range of $10 to $30. Experimenters who wish to "roll their own" can use surplus hf crystals, as outlined in ARRL's *Single Sideband for the Radio Amateur*, or ceramic resonators.

One item of concern to every amateur fm user is the choice of i-f bandwidth for his receiver, as both 15- and 5-kHz deviation are now in common use on the amateur bands. A wide-band receiver can receive narrow-band signals, suffering only some loss of audio in the detection process. However, a wideband signal will be badly distorted when received on a narrow-band rig. At this point it seems reasonable to assume that increasing fm activity and continued production of commercial narrow-band transceivers may gradually shift amateur operation to a 5-kHz deviation standard. But, as with the a-m operators, the wide-band enthusiasts will be around for some time to come, lured by inexpensive surplus wide-band gear.

Limiters

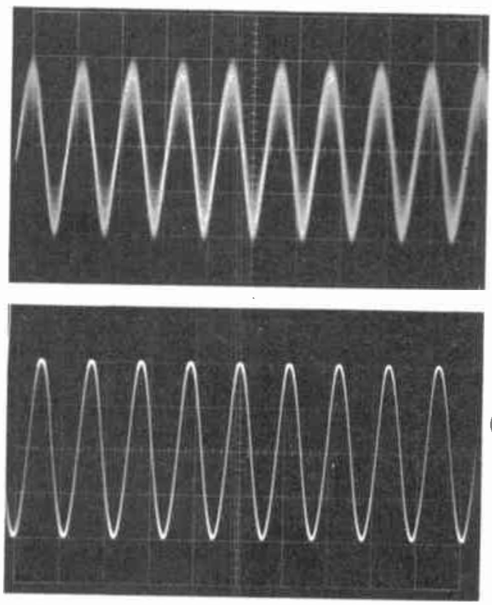
When fm was first introduced, the main selling point used for the new mode was the noise-free reception possibilities. The circuit in the fm receiver that has the task of chopping off noise and amplitude modulation from an incoming signal is the *limiter*. Most types of fm detectors respond to both frequency and amplitude variations of the signal. Thus, the limiter stages preceding the detector are included to "cleanse" the signal so that only the desired frequency modulation will be demodulated. This action can be seen in Fig. 14-13.

Limiter stages can be designed using tubes, transistors, or ICs. For a tube to act as a limiter, the applied B voltages are chosen so that the stage will overload easily, even with a small amount of signal input. A sharp-cutoff pentode such as the 6BH6 is usually employed with little or no bias applied. As shown in Fig. 14-12, the input signal limits when it is of sufficient amplitude so that diode action of the grid and plate-current saturation clip both sides of the input signal, producing a constant-amplitude output voltage.

Obviously, a signal of considerable strength is required at the input of the limiter to assure full clipping, typically several volts for tubes, one volt for transistors, and several hundred microvolts for ICs. Limiting action should start with an rf input of 0.2 μV or less, so a large amount of gain is required between the antenna terminal and the limiter stages. For example, the Motorola 80D has eight tubes before the limiter, and the solid-state MOTRAC receivers use nine transistor stages to get sufficient gain before the first limiter. The new ICs offer some simplification of the i-f system as they pack a lot of gain into a single package.

When sufficient signal arrives at the receiver to start limiting action, the set *quiets* – that is, the background noise disappears. The sensitivity of an fm receiver is rated in terms of the amount of input signal required to produce a given amount of quieting, usually 20 dB. Current practice using the new solid-state devices can produce receivers which achieve 20 dB quieting with 0.15 to 0.5 μV of input signal.

A single tube or transistor stage will not provide good limiting over a wide range of input signals. Two stages, with different input time constants, are a minimum requirement. The first stage is set to handle impulse noise satisfactorily while the second is designed to limit the range of signals passed on by the first. At frequencies below 1 MHz it is useful to employ untuned *RC*-coupled limiters which provide sufficient gain without a tendency toward oscillation.



(A)

(B)

Fig. 14-13 — (A) Input wave form to a limiter stage shows a-m and noise. (B) The same signal, after passing through two limiter stages, is devoid of a-m components.

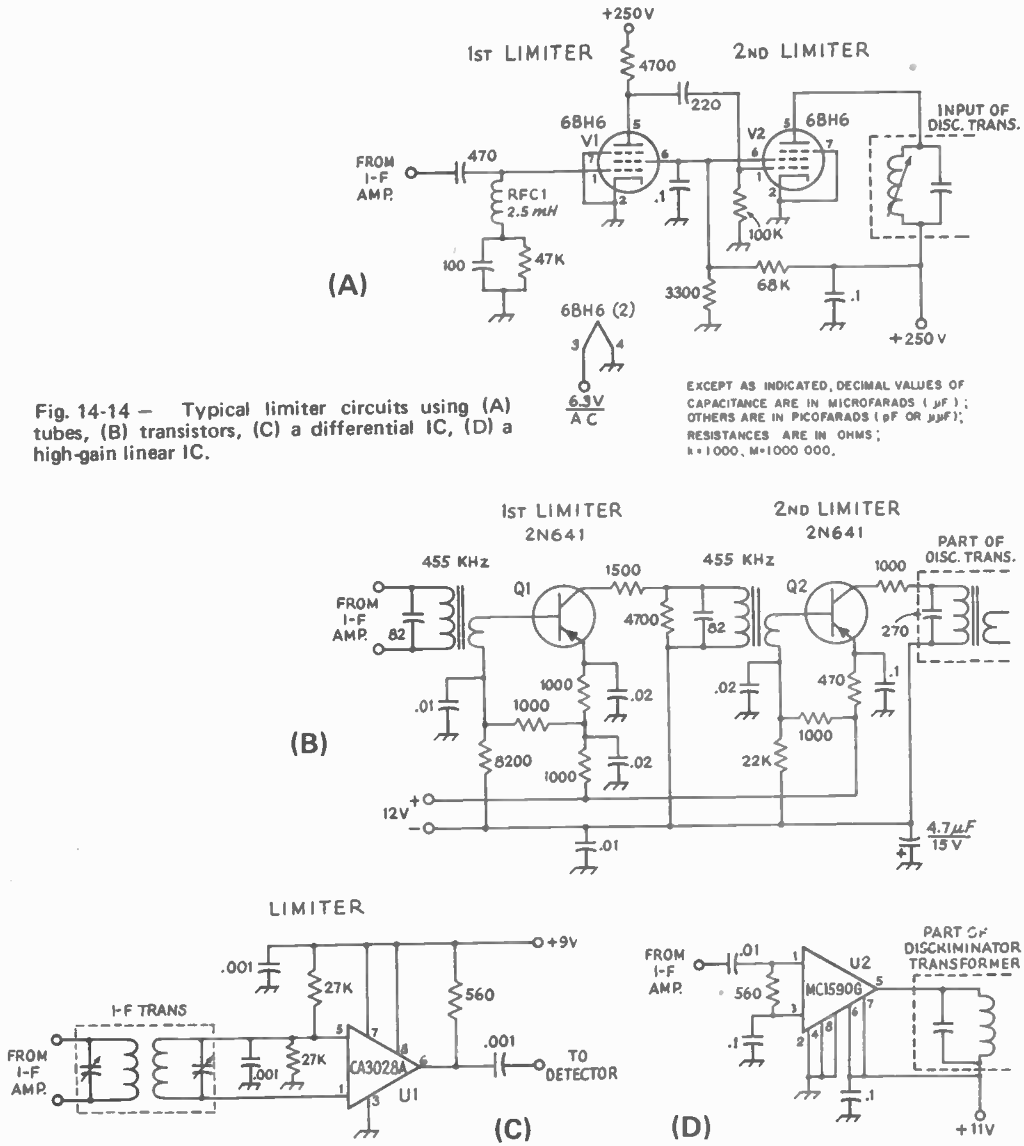


Fig. 14-14 — Typical limiter circuits using (A) tubes, (B) transistors, (C) a differential IC, (D) a high-gain linear IC.

Fig. 14-14A shows a two-stage limiter using sharp-cutoff tubes, while 14-14B has transistors in two stages biased for limiter service. The base bias on either transistor may be varied to provide limiting at a desired level. The input-signal voltage required to start limiting action is called the *limiting knee*, referring to the point at which collector (or plate) current ceases to rise with increased input signal. Modern ICs have limiting knees of 100 mV for the circuit shown in Fig. 14-14C, using the CA3028A or MC1550G, or 200 μV for the Motorola MC1590G of Fig. 14-14D. Because the high-gain ICs such as the CA3076 and MC1590G contain as many as six or eight active stages which will saturate with sufficient input, one of these devices provides superior limiter performance compared to a pair of tubes or transistors.

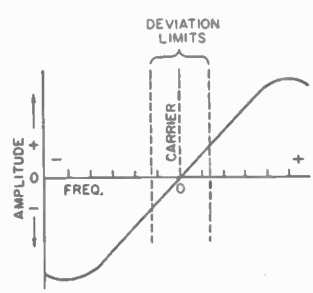


Fig. 14-15 — The characteristic of an fm discriminator.

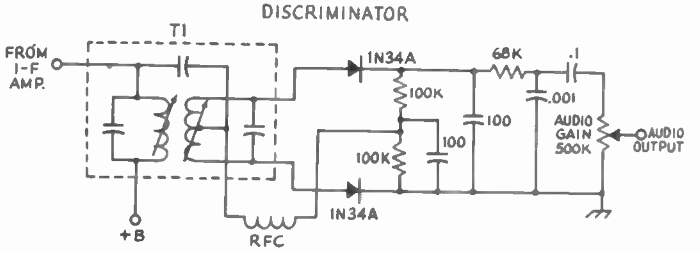


Fig. 14-16 – Typical frequency-discriminator circuit used for fm detection. T1 is a Miller 12-C45.

Detectors

The first type of fm detector to gain popularity was the frequency discriminator. The characteristic of such a detector is shown in Fig. 14-15. When the fm signal has no modulation, and the carrier is at point O, the detector has no output. When audio input to the fm transmitter swings the signal higher in frequency, the rectified output increases in the positive direction. When the frequency swings lower the output amplitude increases in the negative direction. Over a range where the discriminator is linear (shown as the straight portion of the line), the conversion of fm to a-m which is taking place will be linear.

A practical discriminator circuit is shown in Fig. 14-16. The fm signal is converted to a-m by transformer T1. The voltage induced in the T1 secondary is 90 degrees out of phase with the current in the primary. The primary signal is introduced through a center tap on the secondary, coupled through a capacitor. The secondary voltages combine on each side of the center tap so that the voltage on one side leads the primary signal while the other side lags by the same amount. When rectified, these two voltages are equal and of opposite polarity, resulting in zero-voltage output. A shift in input frequency causes a shift in the phase of the voltage components that results in an increase of output amplitude on one side of the secondary, and a corresponding decrease on the other side. The differences in the two changing voltages, after rectification, constitute the audio output.

In the search for a simplified fm detector, RCA developed a circuit that has now become standard in entertainment radios which eliminated the need for a preceding limiter stage. Known as the *ratio detector*, this circuit is based on the idea of dividing a dc voltage into a ratio which is equal to the ratio of the amplitudes from either side of a discriminator-transformer secondary. With a detector that responds only to ratios, the input signal may vary in strength over a wide range without causing a change in the level of output voltage – fm can be detected, but not a-m. In an actual ratio detector, Fig. 14-17, the dc voltage required is developed across two load resistors, shunted by an electrolytic capacitor. Other differences include the two diodes, which are wired in series aiding rather than series opposing, as in the standard discriminator circuit. The recovered audio is taken from a tertiary winding which is tightly coupled to the primary of the transformer. Diode-load resistor values are selected to be lower (5000 ohms or less) than for the discriminator.

The sensitivity of the ratio detector is one half that of the discriminator. In general, however, the transformer design values for Q, primary-secondary coupling, and load will vary greatly, so the actual performance differences between these two types of fm detectors are usually not significant. Either circuit can provide excellent results. In operation, the ratio detector will not provide sufficient limiting for communications service, so this detector also is usually preceded by at least a single limiting stage.

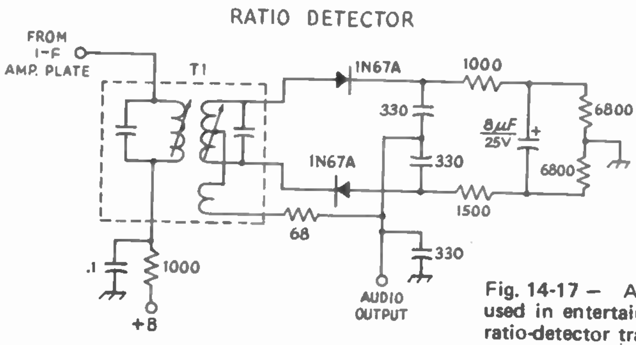
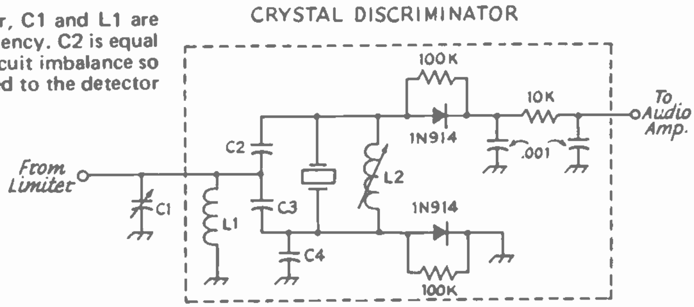


Fig. 14-17 – A ratio detector of the type often used in entertainment radio and TV sets. T1 is a ratio-detector transformer such as the Miller 1606.

Fig. 14-18 – Crystal discriminator, C1 and L1 are resonant at the intermediate frequency. C2 is equal in value to C3. C4 corrects any circuit imbalance so that equal amounts of signal are fed to the detector diodes.



New Detector Designs

The difficulties often encountered in building and aligning *LC* discriminators have inspired research that has resulted in a number of adjustment-free fm detector designs. The *crystal discriminator* utilizes a quartz resonator, shunted by an inductor, in place of the tuned-circuit secondary used in a discriminator transformer. A typical circuit is shown in Fig. 14-18. Some commercially-made crystal discriminators have the input-circuit inductor, L1, built in (C1 must be added) while in other types both L1 and C1 must be supplied by the builder. Fig. 14-18 shows typical component values; unmarked parts are chosen to give the desired bandwidth. Sources for crystal discriminators are listed in Fig. 14-11.

The PLL

Now that the phase-locked loop (PLL) has been reduced to a single IC package, this circuit is destined to revolutionize some facets of receiver design. Introduction by Signetics of a PLL in a single flat-pack IC, followed by Motorola and Fairchild (who are making the PLL in separate building-block ICs), allows a builder to get to work with a minimum of bother.

A basic phase-locked loop (Fig. 14-19A) consists of a phase detector, a filter, a dc amplifier, and a voltage-controlled oscillator (VCO). The VCO runs at a frequency close to that of an incoming signal. The phase detector produces an error voltage if any difference in frequency exists between the VCO and the i-f signal. This error voltage is applied to the VCO. Any changes in the frequency of the incoming signal are sensed at the detector and the error voltage readjusts the VCO frequency so that it remains locked to the intermediate frequency. The bandwidth of the system is determined by a filter on the error-voltage line.

Because the error voltage is a copy of the audio variations originally used to shift the frequency of the transmitter, the PLL functions directly as an fm detector. The sensitivity achieved with the Signetics NE565 PLL is good – about 1 mV for the circuit shown in Fig. 14-19B. No transformers or tuned circuits are required. The PLL bandwidth is usually two to ten percent of the i-f for fm detection. Components R1-C1 set the VCO to near the desired frequency. C2 is the loop-filter capacitor which determines the capture range – that range of frequencies over which the loop will acquire lock with an input signal, initially starting

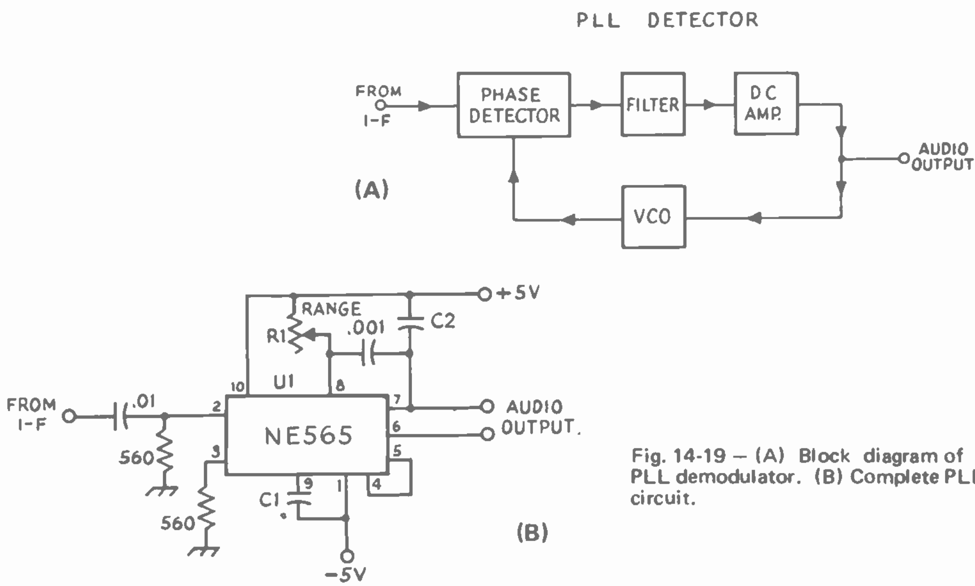
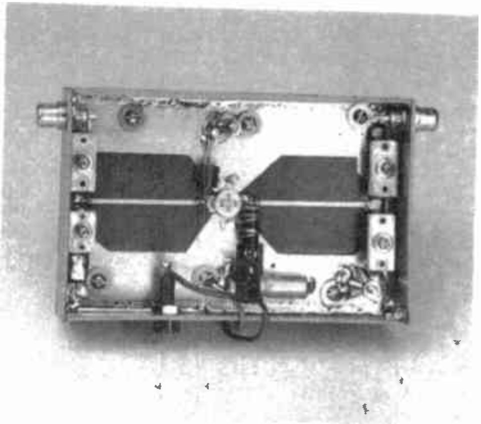


Fig. 14-19 – (A) Block diagram of a PLL demodulator. (B) Complete PLL circuit.

out of lock. The NE565 has an upper frequency limit of 500 kHz; for higher frequencies, the NE561, which is usable up to 30 MHz, can be employed.

220 VISITED

There are several different models of 220-MHz transceivers on the market today. The power levels range from 8 to 15 watts output. This amplifier was developed to make all the power levels the same. It also provides a common base for large amplifiers. Since the selection of 12-volt transistors for 220 MHz is limited, the MRF 226 is used here.



Completed amplifier. Phone plugs were used for this first model. In actual use better vhf connectors should be installed.

Circuit Description

The schematic diagram of the amplifier is shown in Fig. 1. This amplifier was designed for Class C operation only. RFC1 provides the necessary ground and isolation for the base lead. The matching network consisting of C1, C2, and L1 is used to match the low input impedance of the base. This impedance, 1.7 + j0.2 ohms, is transformed to 50 ohms. L1 and L2 are microstrip transmission lines. Their characteristic impedance is 62.7 ohms and this value of impedance was selected for convenience of circuit-board layout. Width of the lines is .062 inch. The output impedance, 6.6-j3.7 ohms, is also transformed to 50 ohms. The combination of L2, C3 and C4 provides the output matching. RFC2, C5, C6, and C7 provide collector isolation and decoupling.

The matching networks were all developed with the aid of a Smith chart. An in-depth discussion of microstrip impedance and matching network calculations can be found in the *Motorola Application Notes* AN-548A and AN-267.

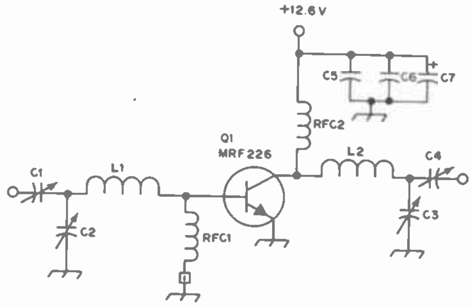


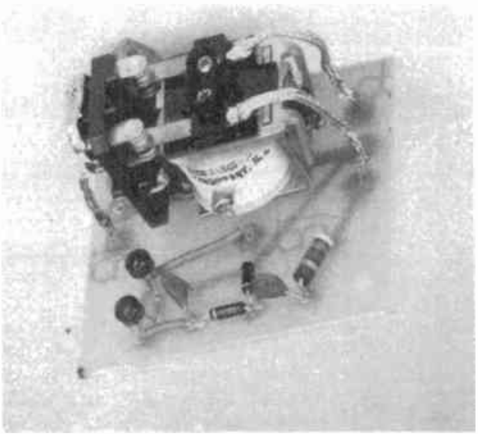
Fig. 1 — Schematic diagram of the 220-MHz amplifier.

C1 — Arco 402 trimmer (1.5 to 20 pF).
C2, C3, C4 — Arco 421 trimmer (2 to 25 pF).
C5 — 220 pF, disk ceramic.
C6 — .01 μF, disk ceramic.
C7 — 10 μF, 50-V electrolytic.
L1, L2 — Strip-line indicator 5 cm long X .062 in. wide.
RFC1 — Molded rf choke 0.15 μH (J. W. Miller 9250-15) with ferrite bead (Ferroxcube 56-590-65/3B).
RFC2 — 2 turns No. 22 enam. wire on 330-Ω, 1-watt resistor.

Construction

Double-sided pc board was used for the construction of the amplifier. The board is 1 oz. copper-clad glass epoxy. Thickness of the copper and type of material are factors in the impedance calculations of the strip lines. The microstrip lines were laid out using .062-inch drafting tape. Using this method, errors in the width are then very small. Pads were etched at the ends of the strip lines to provide areas for capacitor and transistor connections. There are several places on the board that were used for ground connections between the two sides. A No. 16 tinned wire is inserted in a No. 55 drill hole. The wire was soldered on both sides and then cut off flush with the board.

The next step is to mount the transistor to the heat sink and circuit board. Insert the stud of the transistor through the top of the board and attach it to the heat sink by lightly tightening the stud nut after applying heat-sink compound. Now examine the spacing between the heat sink and the bottom of the circuit board with the transistor flush against the top of the board. Insert a number of No. 4 washers (approximately four) so that the transistor tabs are flush against the board. There should be no upward pressure that would tend to lift them off. After the correct number of No. 4 washers have been inserted, the transistor leads are soldered to the board. The stud should be tightened to 6 inch-pounds. Do not exceed the maximum of 6.5. It is better to undertighten than to overtighten the stud. A good heat-sink compound, Dow Corning 340, should be applied between the device and the heat sink. After the amplifier has been tuned the compression trimmers could be replaced with high quality capacitors.



Antenna transfer relay.

Adjustment and Operation

The amplifier should be connected as illustrated in Fig. 2. Adjustment of the amplifier is very easy. A small amount of drive is applied to the input and the output matching capacitors are then adjusted for maximum output. Next adjust the input capacitors for minimum reflected power. Readjust the output capacitors for maximum power. Now increase the drive power to the amplifier. The input and output capacitors will have to be readjusted for minimum reflected power and maximum output respectively. The maximum collector current of 2.5 amperes should not be exceeded.

If maximum gain is achieved, this amplifier will deliver 13 watts output with 1.5 watts of drive. Most of the transceivers on the market have a low power position that provides between 0.5 and 1.5 watts output. This low power could be used to drive the amplifier. If the lower-power position output is very small, the transceiver could be used in the high-power mode and an attenuator inserted before the amplifier.

The antenna-transfer relay was added after the amplifier was developed. If desired, the relay could be mounted on the circuit board. Relay circuitry is similar to the one described in a previous article (*QST*, May, 1972).

A SOLID-STATE ADAPTER

Tubes are seldom used in current designs. For those builders who prefer to be "up with the times," a solid-state version of the 455-kHz adapter was constructed. Using IC limiter/amplifier, and miniature i-f transformers, the unit requires only 25 mA at 12 V for power. See Fig. 14-24A. The Motorola MC1590G provides 70 dB gain, and hard limiting action superior to that obtained with the tube version.

The unit is built on a 2 X 6-1/2-inch circuit board; a template is given in Fig. 14-24B. Because of the high gain of the IC stage, a shield is required across pins 4 and 6 to isolate the input from the output. Alignment and installation are the same as for the tube version. The bandwidth of the miniature transformers restricts this adapter to narrow-band reception. However, builders wishing a wideband version can use the J. W. Miller 8811 miniature coils which are combined with a 12-pF coupling capacitor to form a wide-band transformer.

Fig. 2 — Test setup for adjusting amplifier.

FM COMMUNICATIONS

Although information on fm theory and construction has been available to the amateur for a number of years, this mode has been largely neglected. But now large quantities of used commercial fm mobile equipment have become available for amateur use, creating new interest. Originally designed to cover frequency ranges adjacent to amateur bands, this equipment is easily retuned for amateur use.

One feature of fm is its noise-suppression capability. For signals above the receiver threshold, wideband fm has a signal-to-noise ratio advantage over a-m as a result of its greater "intelligence bandwidth." This same increased bandwidth, however, results in a much more abrupt signal threshold effect, causing weak signals to suddenly disappear. The generality can be made that a-m has a greater range in weak signal work but that wideband fm will provide greater noise suppression in local work. However, in practice, vhf fm mobiles experience greater range than previously found on a-m due to the output powers employed which are considerably higher than those common on a-m.

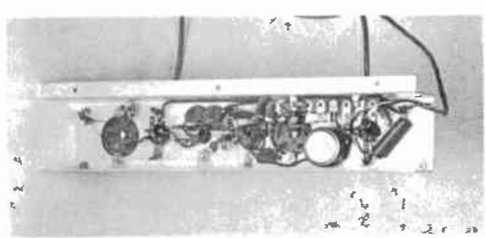


Fig. 14-22 — In this bottom view, the input transformer is to the left, followed by the i-f amplifier, limiter and detector. On the far right are the audio amplifier stage and gain control.

Fig. 14-23 – The solid-state fm adapter is constructed on a 6 X 2-inch etched-circuit board, mounted on a homemade chassis.

Operating Practices

Amateur fm practice has been to retain the fixed-frequency channelized capability of the commercial equipment. VFOs and tunable receivers have not proven satisfactory because of the requirement for precise frequency netting. An off-frequency signal will be received with distortion and will not have full noise rejection. Channelized operation with squelched receivers permits continuous monitoring of the active frequencies. Long, time-consuming calls and CQs are not necessary (or appreciated) to establish communications, as all receivers on the channel "come alive" with the operator's first word. Natural, short transmissions are usually encour-

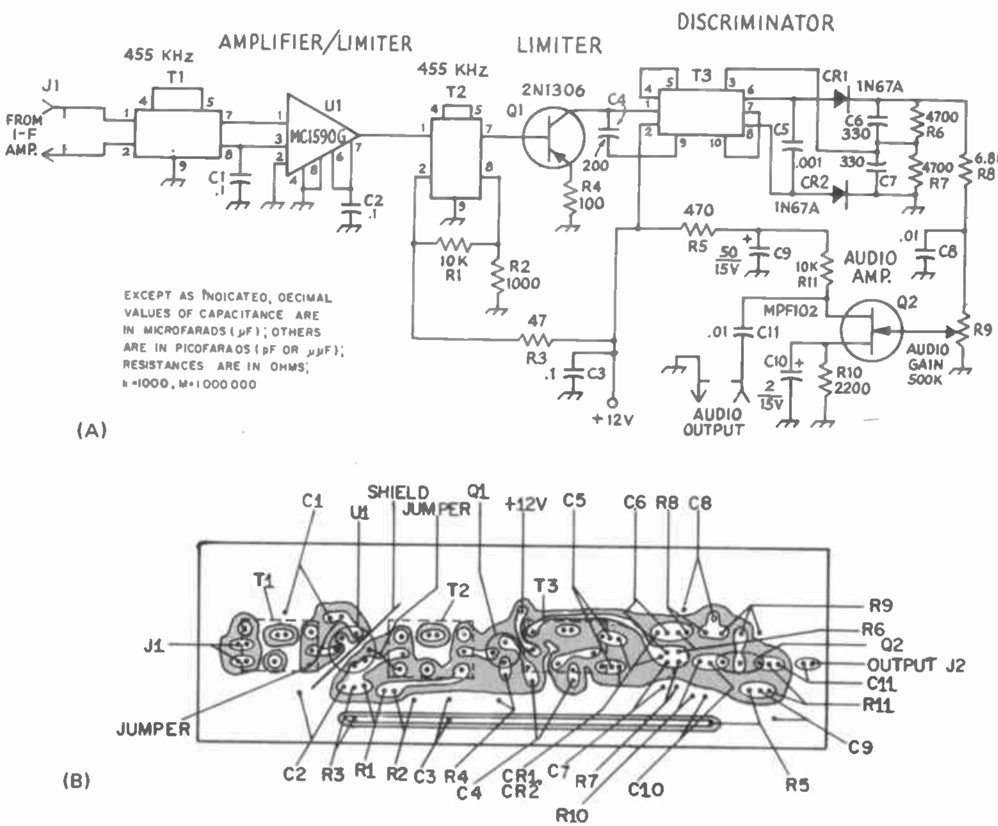


Fig. 14-24 – (A) Diagram of the 455-kHz narrow-band adapter. Resistors are 1/4- or 1/2-watt composition and capacitors are disk ceramic, except those with polarity marked, which are electrolytic. Components with reference numbers that are not listed below are noted for circuit-board location.

J1, J2 – Phono receptacle, panel mount.

R1 – Miniature 1/2-watt composition control.

T1 – Miniature 455-kHz i-f transformer (Miller 8807).

T2 – Miniature discriminator transformer, 455 kHz (Miller 8806).

U1 – Motorola MC1590G.

(B) Template for the solid-state adapter (not to scale).

aged. The old monopoly switch routine, where the operator gabs to himself for 10 minutes at a time, will get him invited off a busy fm channel. Some channels are calling channels on which extended ragchewing is discouraged, whereas other channels, or the same channel in another area, may be alive with chatter. This is a matter of local determination, influenced by the amount of activity, and should be respected by the new operators and the transient mobile operator alike. Some groups have adopted the use of the "10 code" which was originated for law enforcement communications. However, plain language in most cases is as fast and requires no clarification or explanation to anyone.

Fig. 14-25 — A homemade fm transceiver. The transmitter section uses the solid-state exciter and amplifier shown in Chapter 10.

Standards

Standard channel frequencies have been agreed upon to permit orderly growth and to permit communications from one area to another. On two meters, it has been agreed that any frequency used will fall on increments of 60 kHz, beginning at 146.01 MHz. 146.94 MHz (or "nine-four") is the national calling frequency. On six meters, the national calling frequency is 52.525 MHz, with other channels having a 40-kHz spacing beginning at 52.56 MHz. Ten-meter fm activity can be found on 29.6 MHz. Recommendations for 10 meters and 220 MHz are for 40 kHz channel spacing starting at 29.04 and 220.02 MHz. Usage of the 420-MHz band varies from area to area, as it is used for control channels, repeaters, and remote bases, as will be discussed later. In the absence of any other local standard, usage should begin at 449.95 MHz and proceed downward in 50-kHz increments.

Two deviation standards are commonly found. The older standard, "wide band," calls for a maximum deviation of 15 kHz. The newer standard, "narrow band," imposed on commercial users by the splitting of their assigned channels, is 5 kHz. The deviation to be employed by amateurs on frequencies where fm is permitted is not limited to a specific value by the FCC, but it is limited by the bandpass filters in the fm receivers. In general, a receiver with a filter for 5-kHz deviation will not intelligibly copy a signal with 15-kHz deviation. In some areas, a compromise deviation of 7 or 8 kHz is used with some success with both wide and narrow receivers. When necessary, receiver filters can be exchanged to change the bandpass.

REPEATERS

A repeater is a device which retransmits received signals in order to provide improved communications range and coverage. This communications enhancement is possible because the repeater can be located at an elevated site which has coverage that is superior to that obtained by most stations. A major improvement is usually found when a repeater is used between vhf mobile stations, which normally are severely limited by their low antenna heights and resulting short communications range. This is especially true where rough terrain exists.

The simplest repeater consists of a receiver with its audio output directly connected to the audio input of an associated transmitter tuned to a second frequency. In this way, everything received on the first frequency is retransmitted on the second frequency. But, certain additional features are required to produce a workable repeater. These are shown in Fig. 14-28A. The "COR" or carrier-operated relay is a device connected to the receiver squelch circuit which provides a relay contact closure to key the transmitter when an input signal of adequate strength is present. As all amateur transmissions require a licensed operator to control the emissions, a "control" switch is provided in the keying path so that the operator may exercise his duties. This repeater, as shown, is

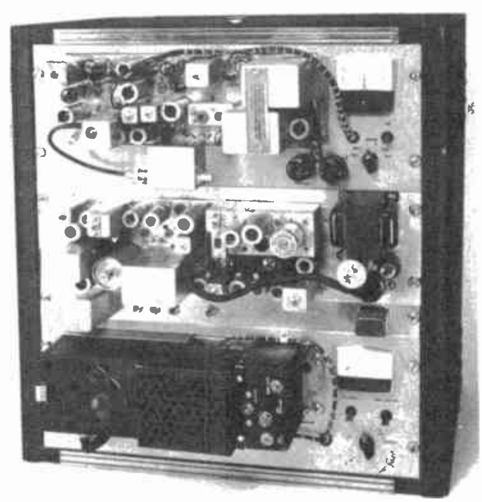


Fig. 14-26 — This typical 144-MHz amateur repeater uses GE Progress-Line transmitter and receiver decks. Power supplies and metering circuits have been added. The receiver located on the middle deck is a 440-MHz control receiver, also a surplus GE unit. A preamplifier, similar to that shown in Fig. 14-44, has been added to the 2-meter receiver to improve the sensitivity so that 0.2 μV of input signal will produce 20 dB quieting.

FM JARGON (Fig. 14-27)

Duplex – Simultaneous transmissions between two stations using two frequencies.
Simplex – Alternating transmission between two or more stations using one frequency.
Low band – 30 to 50 MHz. Also, the six-meter amateur band.
High band – 148 to 174 MHz. Also, the two-meter amateur band.
Remote base – A remotely controlled station, usually simplex (see text).
Machine – Either a repeater or a remote base. Also called a "box."
Vault – Building that houses the machine.
COR – Carrier-operated relay (see text).
CTCSS – Continuous tone-controlled squelch system. Continuous subaudible tone (250 Hz or lower) transmitted along with the audio to allow actuation of a repeater or receiver only by transmitters so equipped. More frequently referred to by various trade names such as Private Line, Channel Guard, and Quiet Channel.
Down channel – Communications circuit from the machine to the control point.
Up channel – Communications and/or control circuit from the control point to the machine.
Open repeater – A machine where transient Operators are welcome.
Closed repeater – A machine where use by non-members is not encouraged. (When heavy expenditures are involved, free-Loaders are not popular.)

suitable for installation where an operator is present, such as the home of a local amateur with a superior location, and would require no special licensing under existing rules.

In the case of a repeater located where no licensed operator is available, a special license for remote control operation must be obtained and provisions made to control the equipment over a telephone line or a radio circuit on 220 MHz or higher. The licensed operator must then be on hand at an authorized control point. Fig. 14-28B shows the simplest system of this type. The control decoder may be variously designed to respond to simple audio tones, dial pulsed tones, or even "Touch-Tone" signals. If a leased telephone line with dc continuity is used, control voltages may be sent directly, requiring no decoder. A 3-minute timer to disable the repeater transmitter is provided for fail-safe operation. This timer resets during pauses between transmissions and does not interfere with normal communications. The system just outlined is suitable where all operation is to be through the repeater and where the frequencies to be used have no other activity.

Remote Base Stations

The remote base, like the repeater, utilizes a superior location for transmission and reception, but is basically a simplex device. That is, it transmits and receives on a single frequency in order to communicate with other stations also operating on that frequency. The operator of the remote base listens to his hilltop receiver and keys his hilltop transmitter over his 220-MHz or higher control channels (or telephone line). Fig. 14-29A shows such a system. Control and keying features have been omitted for clarity. In some areas of high activity, repeaters have all but disappeared in favor of remote bases because of the interference to simplex activity caused by repeaters unable to minitor their output frequency from the transmitter location.

Complete System

Fig. 14-29B shows a repeater that combines the best features of the simple repeater and the remote base. Again, necessary control and keying features have not been shown in order to simplify the drawing, and make it easier to follow. This repeater is compatible with simplex operation on the output frequency because the operator in control monitors the output frequency from a receiver at the repeater site between transmissions. The control operator may also operate the system as a remote base. This type of system is almost mandatory for operation on one of the national calling frequencies, such as 146.94 MHz, because it minimizes interference to simplex operation and permits simplex communications through the system with passing mobiles who may not have facilities for the repeater-input frequency.

The audio interface between the repeater receivers and transmitters can, with some equipment, consist of a direct connection bridging the transmitter microphone inputs across the receiver speaker outputs. This is not recommended, however, because of the degradation of the audio quality in the receiver-output stages. A cathode

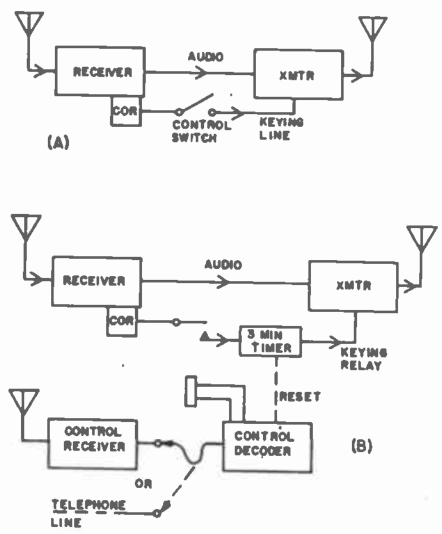


Fig. 14-28 – Simple repeaters. The system at A is for local control. Remote control is shown at B.

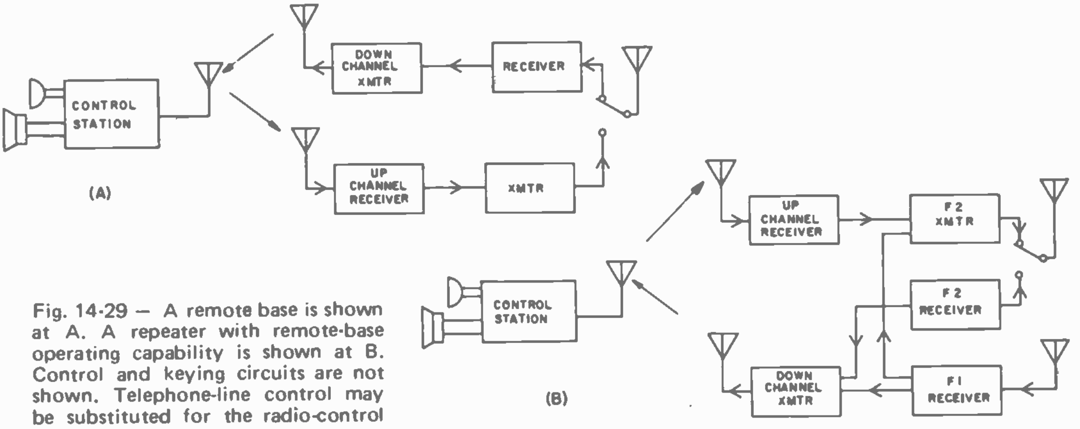


Fig. 14-29 — A remote base is shown at A. A repeater with remote-base operating capability is shown at B. Control and keying circuits are not shown. Telephone-line control may be substituted for the radio-control channels shown.

follower connected to each receiver's first squelch-controlled audio amplifier stage provides the best results. A repeater should maintain a flat response across its audio passband to maintain the repeater intelligibility at the same level as direct transmissions. There should be no noticeable difference between repeated and direct transmissions. The intelligibility of some repeaters suffers because of improper level settings which cause excessive clipping distortion. The clipper in the repeater transmitter should be set for the maximum system deviation, for example, 10 kHz. Then the receiver level driving the transmitter should be set by applying an input signal of known deviation below the maximum, such as 5 kHz, and adjusting the receiver audio gain to produce the same deviation at the repeater output. Signals will then be repeated linearly up to the maximum desired deviation. The only incoming signal that should be clipped in a properly adjusted repeater is an overdeviated signal.

The choice of repeater input and output frequencies must be carefully made. On two meters, 600-kHz spacing between the input and output frequencies is common. Closer spacing makes possible interference problems between the repeater transmitter and receiver more severe. Greater spacing is not recommended if the user's transmitters must be switched between the two frequencies, as happens when the output frequency is also used for simplex operation, either for short-range communications, or to maintain communications when the repeater is not functioning. A 5-MHz spacing is recommended on 440 MHz.

Careful consideration of other activity in the area should be made to prevent interference to or from the repeater. Many "open" or general-use repeaters have been installed on one of the national calling frequencies. On two meters, a 146.94 MHz output is usually paired with a 146.34-MHz input, and many travelers have made good use of this combination where it is found. Where 146.94-MHz simplex activity has not permitted a repeater on this frequency, 146.76 MHz has been used as an alternative. On six meters, several choices of input frequencies have been paired with 52.525 MHz. The choice and usage is a matter for local agreement.

In some cases where there is overlapping geographical coverage of repeaters using the same frequencies, special methods for selecting the desired repeater have been employed. One of the most common techniques requires the user to transmit automatically a 0.5-second burst of a specific audio tone at the start of each transmission. Different tones are used to select different repeaters. Standard tone frequencies are 1800, 1950, 2100, 2250, and 2400 Hz.

PRACTICAL REPEATER CIRCUITS

Because of their proven reliability, commercially made transmitter and receiver decks are generally used in repeater installations. Units designed for repeater or duplex service are preferred because they have the extra shielding and filtering necessary to hold mutual interference to a minimum when both the receiver and transmitter are operated simultaneously.

Wideband noise produced by the transmitter is a major factor in the design of any repeater. The use of high-Q tuned circuits between each stage of the transmitter, plus shielding and filtering throughout the repeater installation, will hold the wideband noise to approximately 80 dB below the output carrier. However, this is not sufficient to prevent desensitization – the reduction in sensitivity of the receiver caused by noise or rf overload from the nearby transmitter – if the antennas for the two units are placed physically close together.

Desensitization can easily be checked by monitoring the limiter current of the receiver with the transmitter switched off, then on. If the limiter current increases when the transmitter is turned on, then the problem is present. Only physical isolation of the antennas or the use of high-Q tuned cavities in the transmitter and receiver antenna feedline will improve the situation.

Antenna Considerations

The ultimate answer to the problem of receiver desensing is to locate the repeater transmitter a

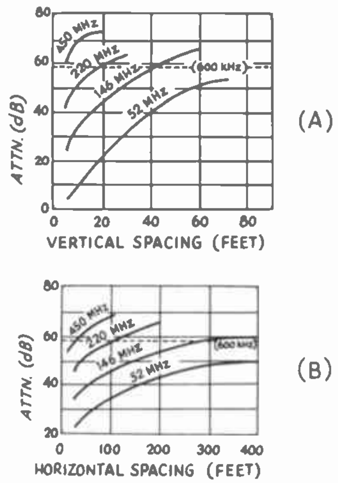


Fig. 14-30 — Charts to calculate the amount of isolation achieved by (A) vertical and (B) horizontal spacing of repeater antennas. If 600-kHz separation between the transmitted and received frequencies is used, approximately 58-dB attenuation (indicated by the dotted line) will be needed.

mile or more away from the receiver. The two can be interconnected by telephone line or uhf link. Another effective approach is to use a single antenna with a duplexer, a device that provides up to 120 dB of isolation between the transmitter and receiver. High-*Q* cavities in the duplexer prevent transmitted signal energy and wideband noise from degrading the sensitivity of the receiver, even though the transmitter and receiver are operating on a single antenna simultaneously. A commercially made duplexer is very expensive, and constructing a unit requires extensive metal-working equipment and test facilities.

If two antennas are used at a single site, there will be a minimum spacing of the two antennas required to prevent desensing. Fig. 14-30 indicates the spacing necessary for repeaters operating in the 50-, 144-, 220-, and 420-MHz bands. An examination of 14-30 will show that vertical spacing is far more effective than is horizontal separation. The chart assumes unity-gain antennas will be used. If some type of gain antenna is employed, the pattern of the antennas will be a modifying factor. A rugged repeater antenna was described in *QST* for January, 1970.

Control

Two connections are needed between the repeater receiver and transmitter, audio and transmitter control. The audio should be fed through an impedance-matching network to insure that the receiver output circuit has a constant load while the transmitter receives the proper input impedance. Filters limiting the audio response to the 300- to 3000-Hz band are desirable, and with some gear an audio-compensation network may be required. A typical COR (carrier-operated relay) circuit is shown in Fig. 14-31A. This unit may be operated by the grid current of a tube limiter or the dc output of the noise detector in a solid-state receiver.

Normally a repeater is given a "tail"; a timer holds the repeater transmitter on for a few seconds after the input signal disappears. This delay prevents the repeater from being keyed on and off by a rapidly fading signal. Other timers keep each transmission to less than three minutes duration (an FCC requirement), turn on identification, and control logging functions. A simple timer circuit is shown in Fig. 14-31B.

Logging and Identification

Current FCC rules require that a log be kept of repeater operations showing each time the repeater is placed in (or taken out of) service. Individual transmissions, however, need not be entered. Although regulations do not require logging of individual transmissions through a repeater, some repeater committees have tape recording equipment connected to the repeater system in order to record a small portion of each transmission. The tapes provide an "unofficial" record concerning repeater usage. A two track tape recorder may have one of the tracks connected to a receiver tuned to WWV or CHU if the repeater committee is interested in having time information.

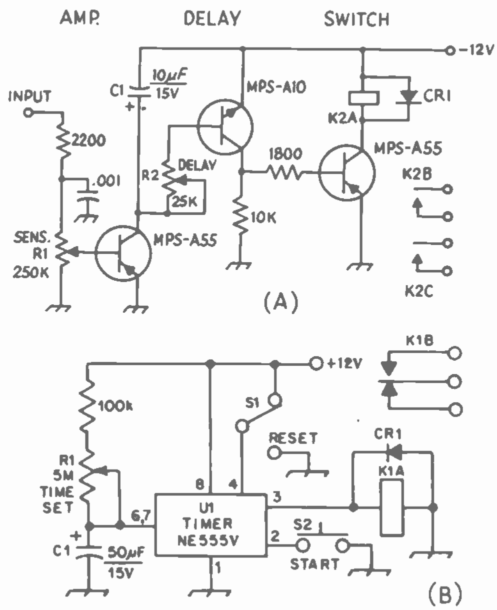
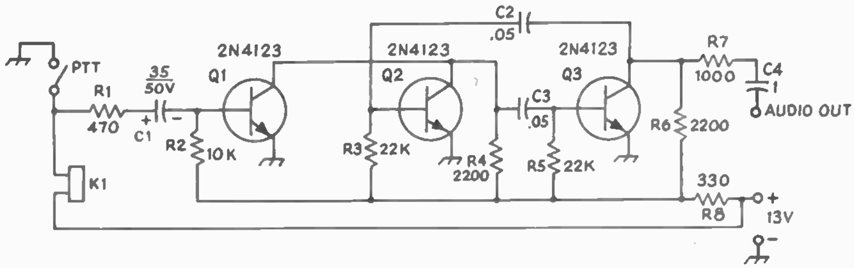
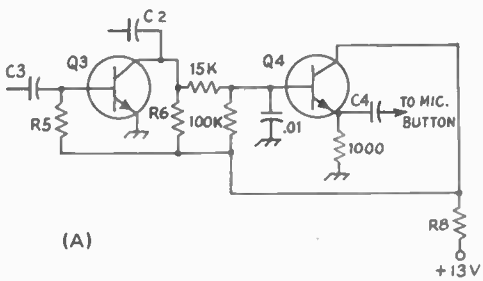


Fig. 14-31 — (A) COR circuit for repeater use. R2 sets the length of time that K1 will stay closed after the input voltage dissappears. K1 may be any relay with a 12-volt coil, although the long-life reed type is preferred. CR1 is a silicon diode. (B) Timer circuit using a Signetics NE555. R1, C1 sets the timers range. C1 should be a low-leakage type capacitor. S1, S2 could have their contacts paralleled by the receiver COR for automatic *START* and *RESET* controlled by an incoming signal.





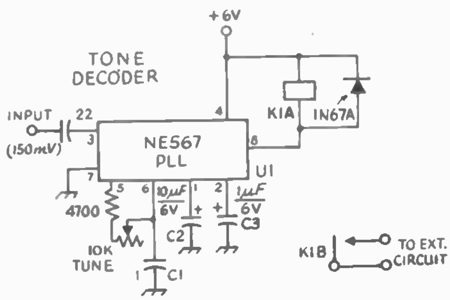
(A)

Fig. 14-32 — (A) Schematic diagram of the "electronic whistle." The main diagram is for high-impedance output. The lower portion has an emitter-follower added, for use with transmitters having low-impedance speech input circuits. All values of capacitance are in μF; polarity indicates electrolytic. (B) Tone-burst decoder. Resistors are 1/2-watt composition and capacitors are mylar. K1 is an spst reed relay with a 6-volt coil (C. P. Clare PRA-2010).

Identification of the repeater itself may be done by users, but lest a forgetful operator leave the repeater unknown, some form of automatic ID is preferred; A tape deck with a short loop tape for voice ID or a digital cw generator has proven to be effective. A suitable solid-state cw generator was described in *QST* for June, 1970.

Many repeaters use a form of tone control so that a carrier on the input frequency will not inadvertently key the transmitter. The most popular form of tone control is known as tone burst, often called whistle on because an operator with a good ear for frequency can use a short whistle instead of an electronically generated tone to key the repeater. A better approach, however, is a simple transistor tone generator, such as shown in Fig. 14-32A.

The whistle-on device was built for use with a Motorola 30-D transmitter, on a 1 1/2 X 2 1/2-inch piece of Vectorbord. It is nothing more than an astable multivibrator, triggered by a one-shot. When the push-to-talk switch is closed, actuating the transmitter relay, K1, Q1 goes from saturation to cutoff, and the multivibrator, Q2-Q3, begins oscillating with a period dependent on the values of R3, R5, C2 and C3. Values given result in a "whistle" of roughly 650 Hz.



(B)

Low Tone (Hz)	*High Tone 1209 Hz*	*1336 Hz*	*1477 Hz*	*1633 Hz*
697	1	2	3	cFO
770	4	5	6	F
852	7	8	9	I
941	*	0	#	P

Fig. 14-33 — Standard Touch-Tone frequencies for the 12-digit pad.

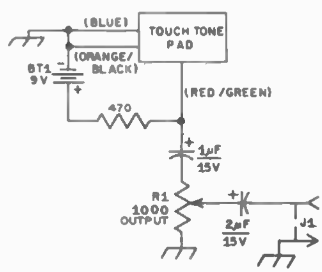


Fig. 14-34 — Typical connections to use a Touch-Tone pad for repeater control. Resistances are in ohms. R1 is a linear-taper composition control and J1 is a panel-mounted phono jack. Capacitors are electrolytic; color coding on the wire leads from the pad is shown in parentheses.

Oscillation ceases when Q1 turns on again. This is regulated by the values of R2 and C1, and is roughly 0.25 second with the values shown. The 470-ohm resistor, R1, protects the base of Q1 from current surges when the PTT is released.

The lower right portion of Fig. 14-32A shows an emitter-follower added, for use with transmitters employing carbon microphones. The value of C4 can be adjusted to give the appropriate output level.

Most of the component values are not critical, except the *RC* products which determine timing. Since the frequency is low, almost any bipolar transistors can be used. Npn types are shown, but pnp will work with opposite voltage polarity. The beta rating should be at least twice R3/R4, to insure saturation.

Most narrow-bandwidth tone decoders currently used in amateur repeater and remote-station applications employ several bulky *LC* circuits to achieve the required audio selectivity. The phase-locked loop (PLL) ICs, pioneered by Signetics, have simplified the design and reduced the size of tone decoders so that a complete Touch-Tone demodulator can be built on a 3 X 5 1/2-inch etched circuit board (about the size for a single-tone decoder using *LC* components).

A typical PLL single-tone decoder, such as might be employed for tone-burst entry control at a repeater, is shown in Fig. 14-32B. One *RC* network establishes the frequency to which the PLL is tuned, according to the relationship:

$$frequency = \frac{1}{R1C1}$$

The PLL, a Signetics NE567, may be operated from 0.1 Hz to 500 kHz. C2 establishes the bandwidth of the decoder, which can be set between one and fourteen percent of the operating frequency. C3 smooths the output signal, and, when this capacitor is made a high value, provides a delay in the turn-on function when a tone is received. Up to 100 mA may be drawn by the '567 output circuit, enough to key a relay directly or to drive TTL logic. The PLL contains 62 transistors.

Autopatch and Touch Tone

Some repeater groups have provided an interconnection to the public telephone network through a device called an autopatch. Details on all phases of phone patching are contained in Chapter 15. Such interconnection has led to the widespread use of the telephone company's Touch Tone system of tone signaling for repeater control functions, as well as telephone dialing. Because all of the Touch-Tone frequencies are within the voice band, they can be transmitted by any amateur voice transmitter.

The Touch-Tone control system consists of pairs of tones (see Fig. 14-33) for each of 10 numbers and the two special functions. One tone from the high-frequency group is generated simultaneously with one tone from the low-frequency group to represent each number or function. The Touch-Tone generator pad from a standard telephone instrument is usually employed. See Fig. 14-34 for connections. A simple Touch-Tone decoder using ICs throughout was described in July 1971 *QST*.

A SCANNING TOUCH-TONE DIGIT AND WORD DECODER

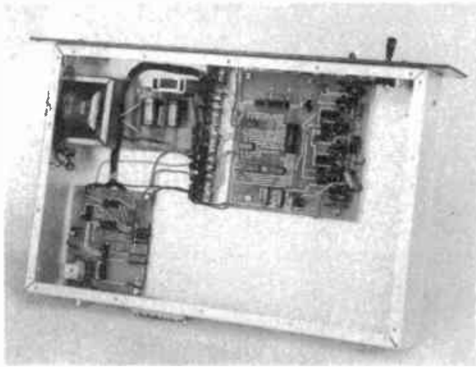
The Touch-Tone encoding system, used extensively in auto-patch operations on fm repeaters across the country, offers a ready-made source for dual-tone codes, and advances in micro-circuitry design have produced a single device that can be used to decode these dual-tone codes for a variety of remotely controlled functions. However, one device is required to decode each tone. In this article the writer shows how a scanning decoder evolved as an attempt to avoid using seven of these decoder ICs, and how a simple counter circuit can recognize specific four-digit-word sequences to provide a unique approach to a remote-control decoder.

There are sixteen tone pairs possible – selecting one from the low group, 697, 770, 852 and 941 Hz, and one from the high group, 1209, 1336, 1477 and 1633 Hz. Two phase-locked-loop types of tone decoders should therefore be sufficient if each one sequentially scans the four tones of one group. In this way two decoders with some added scanning circuits take the place of eight. Parts of the scanning circuit such as the clock oscillator and digit decoder would be required in any case for word decoding, and the parts' cost of the present system using primarily low cost TTL logic is reasonable. One disadvantage of the scanning decoder is the slow response time resulting from the need to wait for each decoder to find the received tone. Also, a delay is built-in which requires both decoders to halt for at least one full clock period before a digit is registered. The operation is, thus, relatively immune to spurious responses from voice signals, yet takes 1/2 second or less to respond properly to any digit.

The type 567 tone decoder is not satisfactory for use in this circuit because neither side of the frequency determining R-C network is grounded. A Motorola MC1310P was tried because one had been used previously for tone decoding and was found to work well in this frequency range. Its intended use is as a phase-locked-loop fm stereo decoder. In this application it locks onto the 19-kHz pilot tone which is present, along with the audio signal, and turns on an open-collector output to light a stereo indicator lamp.[1] Its internal oscillator runs at 76 kHz, and an internal frequency divider gives the 19 kHz for the pilot tone

[1] Gay, *Electronics 44* (24), p. 62, November 22, 1971.

detection. A 19-kHz monitor output is provided. In the present circuit the oscillator is run at four times the Touch-Tone frequencies, and the stereo decoder function (except for the indicator lamp output) is ignored. Since the frequency determining resistor runs to ground, it is programmed easily for scanning operation by using four resistance values and four, open-collector, 15-volt NAND gates (7426).



Shown here is the decoder built by W1GNP as described in *QST* for January, 1976.

Digit Decoding.

The digit decoder is shown in Fig. 1. A 12-volt power supply is required for the MC1310P circuits, and the voltage on the programming resistance network is somewhat high for the usual open-collector NAND gates, so a 7426 is used. The .01- and .015-μF capacitors should be mylar or silver mica for temperature stability, and some experimentation with the resistance values to achieve the correct frequencies may be necessary. A fixed resistance of 10k-Ω or 12k-Ω was used and a jumper-wire or selected value of fixed resistance was inserted between all of the 1-kΩ potentiometers. The circuit time constants resulting from the use of 1.0-μF ceramic capacitors appear to be the correct value for the present system. The 820-Ω pull-up resistor from the +5-volt supply makes the output TTL compatible; the monitor output is correct for driving TTL devices.

The 555 timer U9 and inverting gate U7D provide a positive clock pulse for all the 7473 flip-flops which toggle on the trailing edge of the pulse. When no tone inputs are received, U3 and U4 count through four states each and cause the open collector gates U5 and U6 to conduct in sequence A through D, thereby sweeping the frequencies of U1 and U2 upward through the low and high tone groups respectively. When either tunes to an incoming tone, it becomes phase locked to it, its output at the test point goes low, the counter is stopped because its *J* and *K* inputs are low and the detector remains locked on the incoming frequency. Also, the monitor output can pass through gate U7B or U7C and can be used for exact measurement of incoming tone frequencies. When both tones are so detected, a logic-one condition appears at the output of NOR gate U7A and counter U8 is permitted to advance from its cleared condition.

The *J-K* flip-flops in U8 are wired to advance in count through states 0, 1, 3, 2, 2 and become stopped in state *2* (U8A off, U8B on) until reset when one or the other tone detector drops out. Its purpose is to provide a two-clock-period double check on the decoder operation and yield a single clock pulse (CLK) just before the end of state 3 if the tone signal is so validated. Also, during state 3 (U8A and U8B on) the decoder U11 is enabled, and one of the digit outputs from U12 or U13 comes on.

The decoder makes use of the counter states of U3 and U4 when they are stopped by an incoming two-tone signal. Since the tones are scanned from low to high and the low group (top to bottom rows on a standard pad) is wired to the two least-significant-digit inputs on U11, the output states of U11 would correspond to the tone button assignments of a standard Touch-Tone pad. In order that the digit outputs are correct for the actual assignment on the tone pad, the U11 outputs are reordered and the twelve corresponding to the commonly used twelve button pad are inverted to the positive logic form by U12 and U13. The four gates U14, U15, U16 and U17 are optional and are used to obtain the binary equivalent of the standard digit assignment of the Touch-Tone pad.

Word Decoding

The word-decoding circuitry shown in Fig. 2 consists of two, three-digit prefix decoders and an output flip-flop U22. Each prefix decoder consists of a dual *J-K* flip-flop (U18) two AND gates (U20A and B) and two NAND gates (U21A and B). The prefix or first three digits of the four-letter word being decoded are selected by connecting inputs digit 1 (DG1), digit 2 (DG2), digit 3 (DG3) and digit 4 (DG4) to the desired outputs from U12 and U13. Likewise, the other word-decoder inputs DH1, DH2, DH3 and DH4 are connected to four outputs from U12 and U13. The first four-letter word such as the sequence 4639 would turn U22 on, and the second such as 1 * 8 # would turn it off.

This sequence detection is achieved by the gating used on the *J-K* inputs of U18 and U19. Each is a two-stage counter which will advance in the state sequence 0 – 1 – 3 – 2 – 0 only if the correct digit input is on in proper sequence. That is, in order to advance from 0 – 1 DG1 must be on, to advance from 1 – 3 DG2 must be on, and to advance from 3 – 2 DG3 must be on. If any are off when they should be on, the state goes directly to zero. When state 2 is reached, U18A is off and U18B is on, and two of the three AND *J* inputs of U22 are on. At this point the three-digit prefix has been received successfully. If the fourth digit received corresponds to DG4, the clock pulse (CLK) will also turn on U22 since all its *J* inputs will then be on. Similarly, the second four-digit-word sequence will turn off U22. Further interfacing between the TTL output of U22 and a controlled system will depend upon its nature. A simple relay driver using two parallel-connected 15-volt open-collector buffer inverters (1/3 of a 7416) and a 12-volt, 150-ohm relay is shown. A small silicon diode connected as shown helps to avoid transient problems.

The on-off function of Fig. 2 can be simplified by using the same prefix decoder for both turn on and turn off: only the fourth digit need be different. One must be sure the word decoder is reset before it will respond properly to a four-digit word. This is just a matter of being sure that any code such as 4639 is not preceeded by a 4, a 46, or a 463. If in doubt about what the last digit may have been in the system, an extra random digit other than 4 is generally sufficient. Alternatively, the reset inputs of U18 and U19 can be wired to some completely independent source of a reset such as the carrier-input detector.

A few words about the choice of codes. For most amateur radio applications the four-digit word provides adequate security. If a great deal of phone patch activity is present on the channel, the characters * or # should be used in the code as these do not appear in phone numbers. The four additional characters generated only by a sixteen-button pad can be used if four additional inverters are added to the group in U12 and U13. When a number of four-digit codes are used to operate a remote system of some sort, it becomes difficult to remember them all, and ease of use becomes an important factor in choosing codes. Often a single digit is better to turn something off because it's quicker and less likely to be forgotten. In any case, all system codes must be mutually compatible.

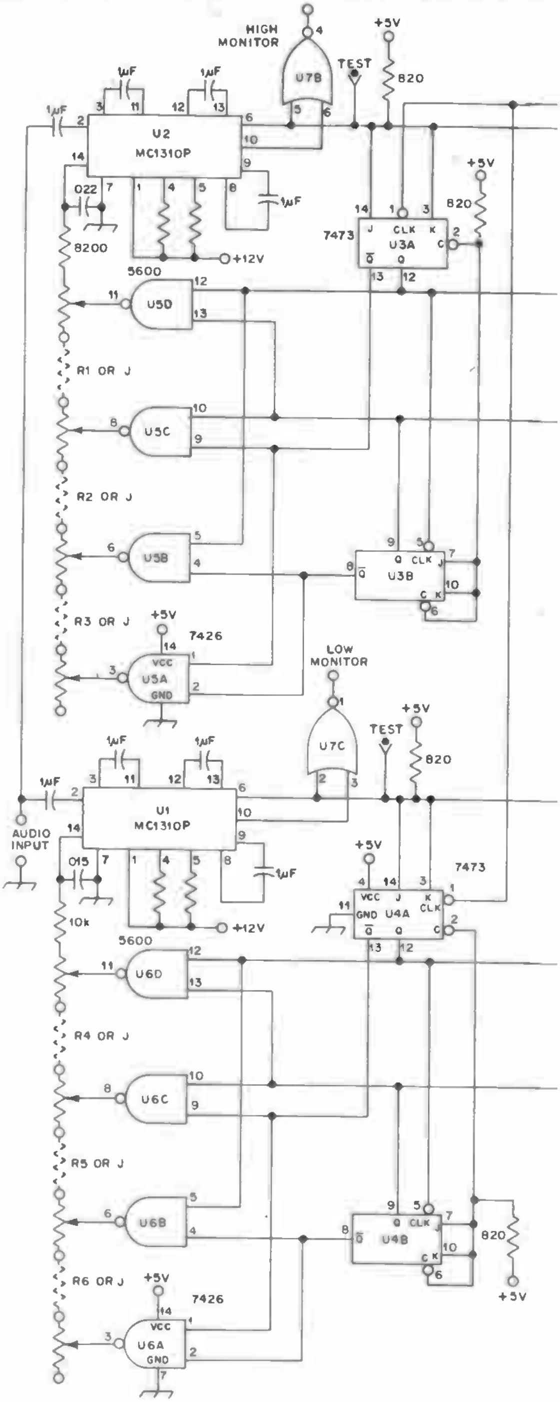
Construction

The unit shown in the title photograph was constructed on double-sided, copper-clad pc board. The layout and fabrication of the boards was done by Chuck Carroll, W1GQO, in the ARRL laboratory.

The tone and digit decoder circuit of Fig. 1 was constructed on a 6 X 6-inch pc board. All of the components are mounted on the top side of the board and are soldered on both sides of the pc board. The value for R1 through R6 in each tone-selection line should be selected so that the potentiometer will tune the circuit to the proper tone in the middle of its resistance range. The values shown in the circuit diagram of Fig. 1 are typical and can be used as a starting point for selecting the final value. The word decoders are also constructed on pc board with a double-sided layout. The decoder board is 4 X 4 inches with all of the components installed on the top side. Several of these decoder boards can be stacked and will make the addition of control functions a simple task.

Alignment

Alignment is a matter of setting the scanned frequencies to the correct values using a frequency



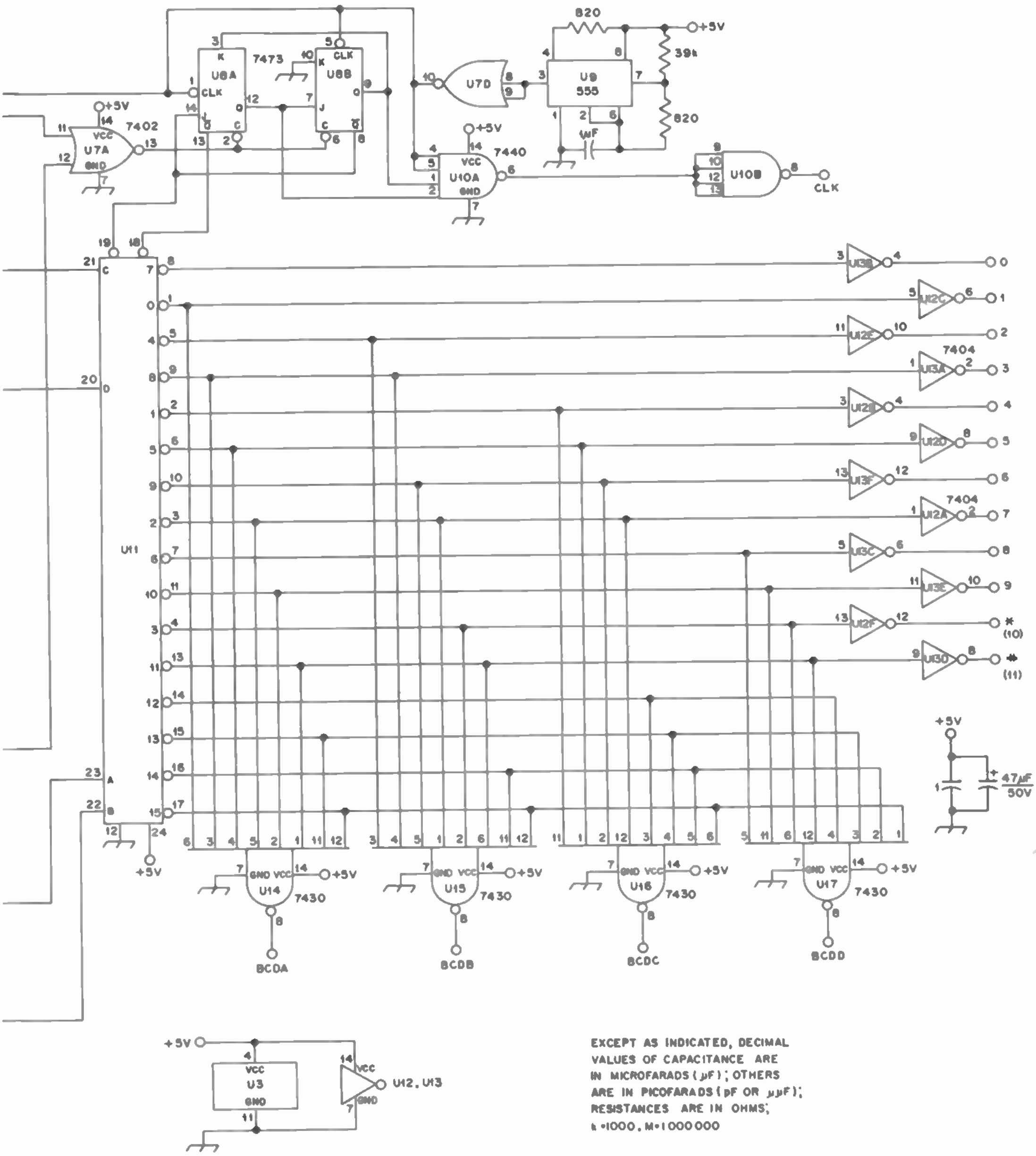


Fig. 1 — Schematic diagram of the tone and digit decoder. Parts placement is not critical, but standard construction practice should be followed when fabricating these circuits.

counter connected to a monitor point. The associated test point is grounded, and with no input to the decoder one of the frequencies can be adjusted, depending on which of the four tones in any group happened to be on when it was stopped manually. It is best to stop the highest tone first and adjust the first potentiometer, along with R9 if necessary, to get 941 Hz. Then select the second potentiometer and value of R1 until 852 Hz can be tuned. In a like manner, adjust all eight frequencies to the correct values. After several months of operation the response became sluggish and finally the unit stopped working, but original performance was restored by retuning R1 through R6. Satisfactory operation is obtained with input levels between 0.1 and 1 volt ac.

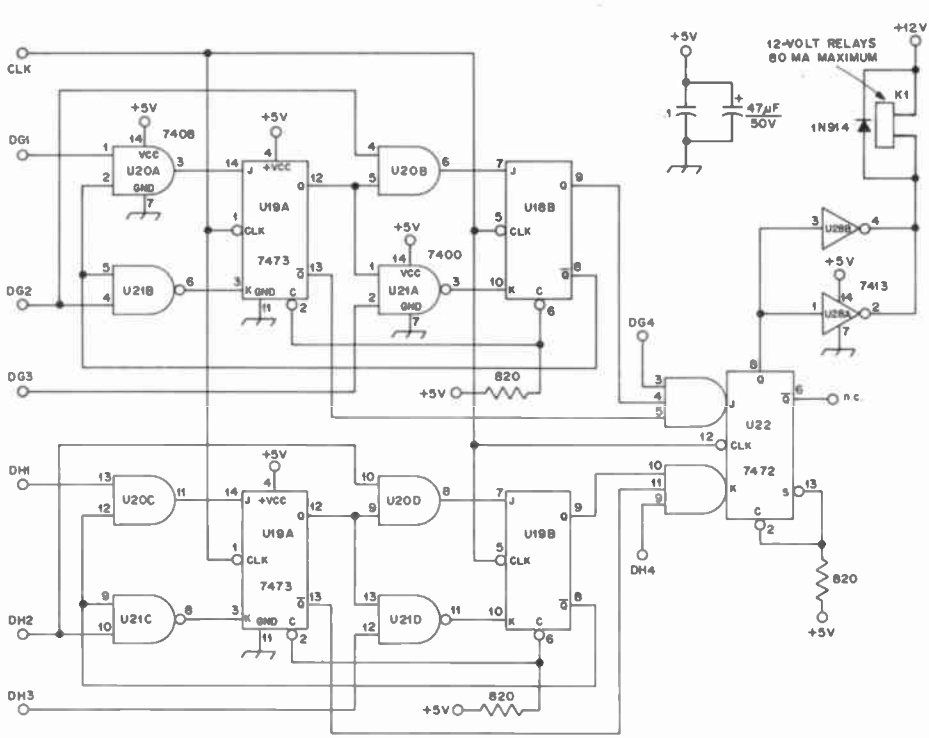


Fig. 2 — Schematic diagram of the relay-control pc board. Be sure to select relays that have contact-current capability for use in the desired application.

A TONE BEEP KEYER FOR REPEATERS

This simple telemetry circuit was designed for the WR6ABN repeater. Earlier uses of tones and tone bursts reminded users to allow time for breaking stations, and to indicate that the time-out timer had been reset. This latter indication was by means of transmitting two tones simultaneously.

This system is designed to inhibit one of the two tones, selectively, and allow either the high or low tone to indicate the position of the user's carrier in the receiver passband.

The sensors were adjusted to trip the relays at 1 kHz above or below the center frequency; this appears to be a practical value for narrow-band receivers. Thus, the "on-channel" slot is 2-kHz wide, centered about the receiver input frequency.

The 741 op amp is set for a dc gain of 1000. The ac gain of the circuit is very low, as set by the 1-µF bypass capacitor across the 1M-Ω resistor in the feedback loop, and 1 µF across the 50-kΩ control in the input circuit. The output of the 741 feeds two transistors and a zero-center meter.

The steering diodes, CR1 and CR2, allow the op amp to drive Q1 or Q2 into conduction and to charge C1 or C2 to the value of the op-amp output voltage. R1 and R2 allow capacitors C1 and C2 to charge above the base voltage of the transistors and to cause them to conduct for about 5 seconds after the drive voltage from the op amp is removed. This delay acts as a memory.

Note that the poor ac frequency response of the op-amp means that the input to it must remain for approximately 3 seconds in order for it to load C1 or C2 for the readout.

The input to the op amp is shorted to ground when a carrier is not present. This prevents noise from loading up the sensor prior to a reading. It also allows the adjustment of the dc offset control, R5. The calibrate potentiometer, R6, is adjusted to a point where signals 1 kHz above or below the center frequency of the receiver will just trip relays K4 or K5. (Note that the receiver should be adjusted so that the discriminator voltage is zero with no signal.) This adjustment of R6 to ±1 kHz determines the slot width. The center frequency is determined by the usual crystal-oscillator adjustments in the receiver.

K1 can be the normal COR or a separate relay keyed by the COR. This relay keys both the input to the op amp and the delay relay, K2. Because of the discharge time of C3, K2 will have a delayed release. When K2 releases, it keys K3 for a short period as determined by C4 and R8. The values needed for C3, C4, R7 and R8 will vary, depending upon the characteristics of K2 and K3.

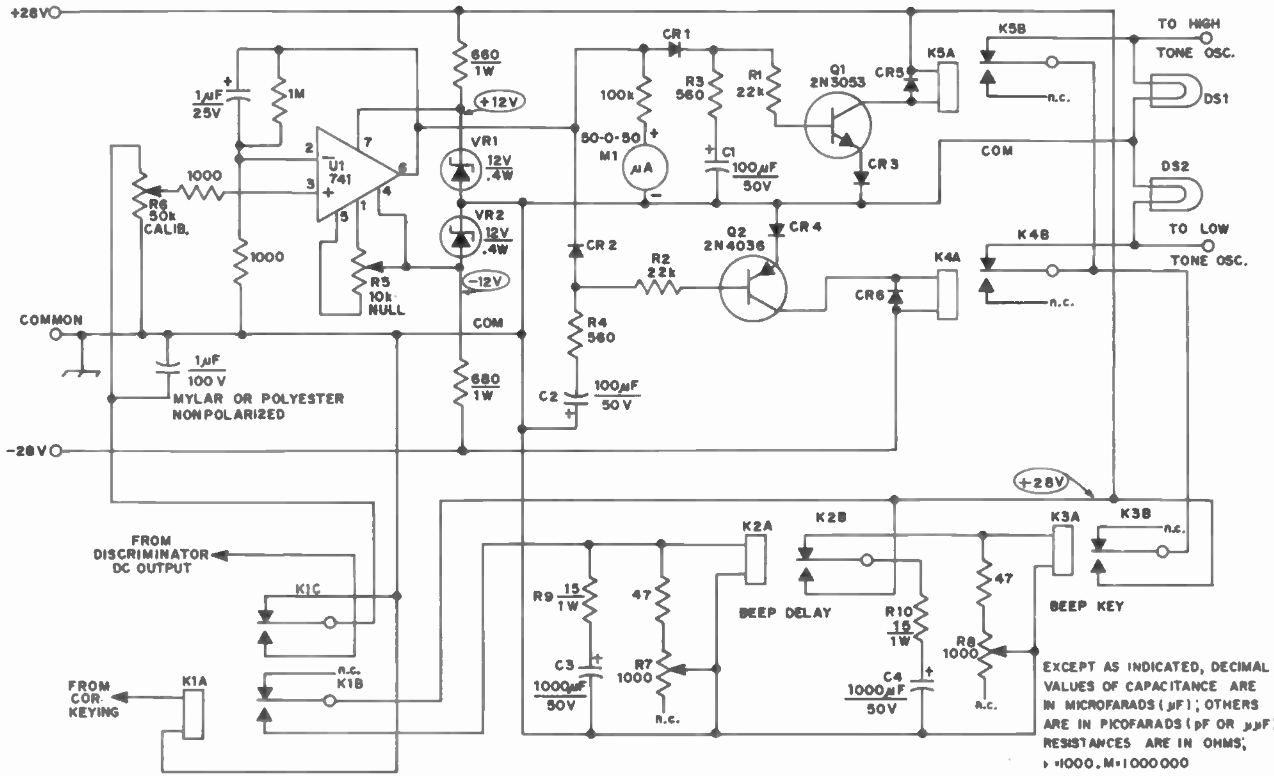


Fig. 1 — The schematic diagram of the tone-beep keyer. A dual 28-V supply is used in this system, but there should be no difficulty in revising values to make use of lower voltages. The charging current of C1 through C4 is limited to a safe value by means of the series resistor in each case. If the meter is omitted, tip jacks should be provided to aid in adjusting the circuit.

CR1-CR6, incl. — Silicon diodes, 1N2069 or equiv.

DS1, DS2 — 28-V pilot lamps. Lower-voltage units or LEDs with suitable dropping resistors may be used.

K1 — Dpdt relay. Coil voltage and current must be compatible with voltages available from receiver COR circuitry.

K2-K5, incl. — Spdt relays, 450- to 700-ohm coil for 24 V dc. Allied Control T154-2C or equiv.

U1 — Operational amplifier IC. Fairchild µA741 (U5B7741312), Signetics µA741T or µA741CV, Motorola MC1741G or MC1741P1 or equiv.

Operational Notes

1) K2 establishes length of delay between end of carrier and keying of tone beep.

2) K3 establishes length of tone beep.

3) K4 or K5 select the desired tone to be keyed.

4) Adjust R5, NULL control, for zero dc volts at the output of the 741 with K1 deenergized.

5) The trip point of K4 or K5 is adjusted by means of R6, the CALIBRATE control. It should be adjusted while monitoring a carrier set to the desired frequency offset value.

This unit was built by W6MEP and described originally in *QST* for May, 1974.

IMPROVING FM RECEIVER PERFORMANCE

Many older fm receivers, and some new models, do not have sufficient sensitivity or limiting capability. Also, the transceivers designed for the mobile telephone service do not have a squelch or audio power-amplifier circuit. Suitable accessory units can be easily constructed to improve the performance of a rig deficient in any of these areas.

A simple preamplifier, such as shown in Fig. 14-45 for 146 MHz and in Fig. 14-47 for 440 MHz, may be added to a receiver to increase its sensitivity and to improve limiting (as the overall gain before the limiter will be increased by 10-15 dB). The 2-meter version uses a dual-gate MOSFET while the 440-MHz unit employs two JFETs in a grounded-gate circuit. Both amplifiers are adjusted by peaking all tuned circuits for maximum limiter current while receiving a weak signal.

A receiver will have a poor limiting characteristic if the gain before the limiter circuit is insufficient, or if the limiter itself is of poor design. The circuit of Fig. 14-48 can be added to a receiver to replace an existing limiter stage. The new limiter uses an RCA CA3011 integrated circuit. Care must be used in the installation and layout of this high-gain IC to insure stability. The CA3011 will provide a "hard" limiting characteristic with about 100 mV of signal input.

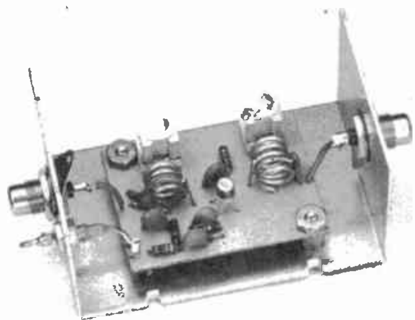


Fig. 14-44 – The 2-meter preamp. may be mounted in a small Minibox or connected directly inside an fm receiver.

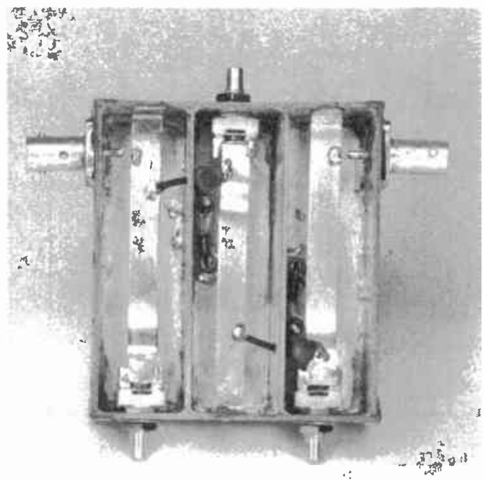
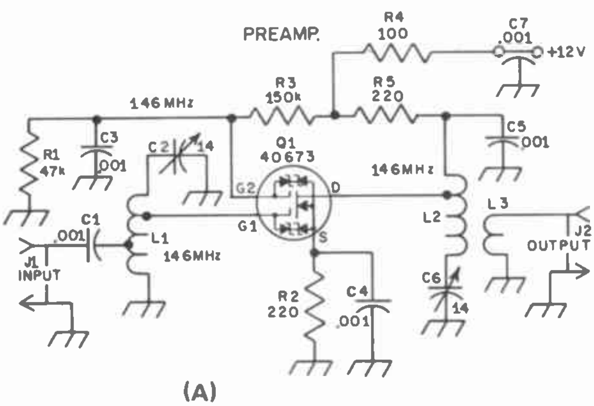


Fig. 14-46 – The 440-MHz preamplifier is constructed in a 3 X 3 1/2 X 1-inch box made of double-sided circuit board. All abutting edges are soldered to complete the enclosure. Two 3 X 15/16-inch shields separate the tuned lines.



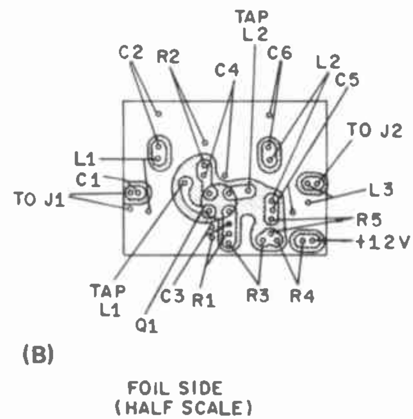


Fig. 14-45 – Circuit diagram (A) and pc-board layout (B) for the 2-meter preamplifier. Resistors are 1/4-watt composition and capacitors are disk ceramic unless otherwise noted. Components not listed below are given designators for circuit-board location purposes.

C2, C6 – Air variable (Johnson 189-506-5).
J1, J2 – Phono type, panel mount

L1 – 5 turns, No. 16, 5/16 inch dia, 1/2 inch long. Tapped at 2 turns for the antenna connection, and 4 turns for G1.
L2 – 4 turns, No. 16, 5/16 inch dia, 3/8 inch long. Tapped at 2 turns.
L3 – 1 turn, plastic-covered hookup wire, 5/16 inch dia, placed between two turns of L2.

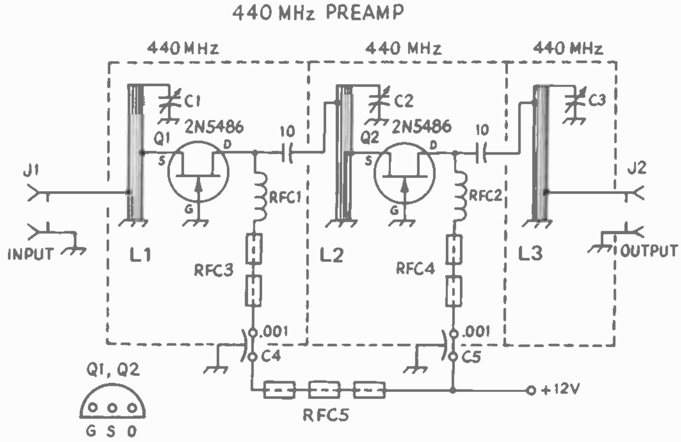


Fig. 14-47 – Schematic diagram of the uhf preamplifier. Capacitors are disk ceramic unless otherwise noted.
C1-C3, incl. – 1.4 to 9.2-pF miniature variable (Johnson 189-0563-001).
C4, C5 – Feedthrough type.
J1, J2 – BNC type, chassis mount.
L1-L3, incl. – 2 5/8 X 1/4-inch strip of brass, soldered to the enclosure on one end and to the capacitor at the other. Input and output taps (on L1 and L3) are 1/2-inch up from the ground end. Drain taps for Q1 and Q2 on L2 and L3, respectively, are made just below C2 and C3.
RFC1, RFC2 – 420-MHz choke (Miller 4584).
RFC3, RFC4 – Two ferrite beads on a short piece of No. 20 hookup wire. (Beads are available from Amidon Associates, 12033 Otsego St., N. Hollywood, CA 91607.)
RFC5 – Three ferrite beads on No. 20 hookup wire.
Q1, Q2 – Motorola JFET.

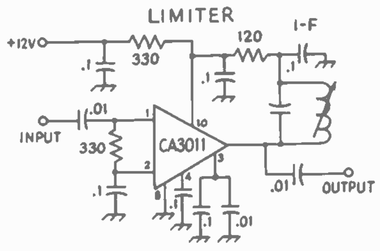


Fig. 14-48 – Diagram of a limiter which may be added between the last i-f stage and the detector of a receiver.

FM Bibliography

Goldsmith *et al*, *Frequency Modulation*, in two volumes, RCA Review, RCA, 1948.

Rider and Uslan, *FM Transmission and Reception*, John F. Rider Publisher, 1948.

Wolf, *FM Schematic Digest*, Two-Way Radio Engineers, 1970.

Pre-Progress Line Diagrams, in two volumes, Mobile Radio Department, General Electric Company, 1968.

Hund, *Frequency Modulation*, McGraw-Hill Book Company, 1942.

Lytel, *Two-Way Radio*, McGraw-Hill Book Company, 1959.

A SOLID-STATE FM TRANSMITTER FOR 146 MHz

In an effort to shrink the dimensions of the solid-state fm transmitter treated earlier in *QST*, and in the 1972 *ARRL Handbook*, it became necessary to eliminate one stage of the rf section, and to reduce the size of the speech amplifier and clipper. The product of that effort is shown schematically in Fig. 1.

A slightly different electrical approach was taken, wherein the oscillator was called upon to deliver a fair amount of power. The increased output from Q1 permitted the deletion of a driver stage ahead of the PA. The change made it necessary to pay particular attention to the design of all networks between stages, providing adequate selectivity to assure suppression of unwanted output frequencies. The criterion was met, as evidenced by a spectral display of the output energy. The MK-II version is as clean as was the MK-I model.

A logical approach to reducing the area occupied by the speech amplifier and clipper was the employment of a transistor-array IC as opposed to the use of discrete components. The latter technique was used in the MK-I example.

Circuit Highlights

Generally, the circuit of Fig. 1 follows the classic sonobuoy format given in RCA's *Power Circuits, DC to Microwaves.*[1] Some of the circuit changes made are radical; others are subtle. The

[1] Recommended for amateur libraries. Order from local radio store, or write RCA Electronic Components, Harrison, NJ 07029. Price: $2.

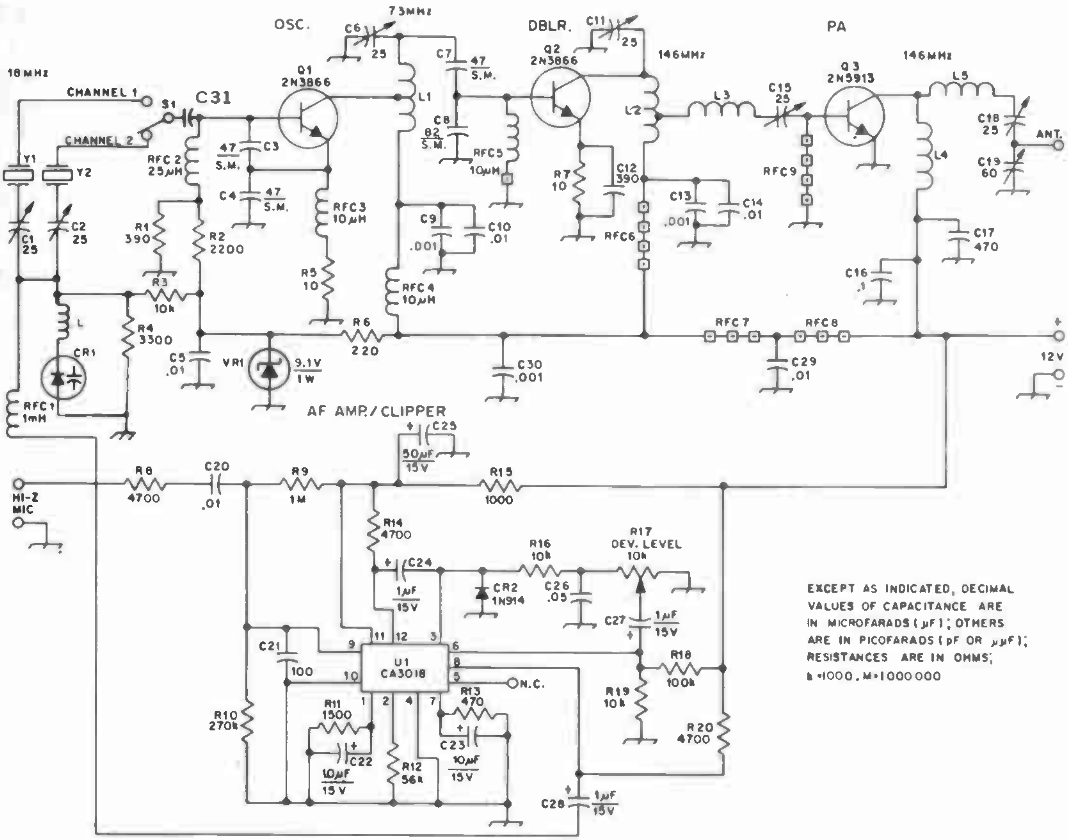


Fig. 1 — Schematic diagram of the 2-meter fm transmitter, Fixed-value capacitors are disk ceramic unless otherwise marked. Polarized capacitors are electrolytic. Fixed-value resistors are 1/2-watt composition. Numbered components not appearing in parts list are so numbered for pc-board layout purposes only. Use crown type heat sink on Q1, larger style on Q2 and Q3.

C1, C2, C6, C11, C15, C18 — 7- to 25-pF miniature ceramic trimmer (Erie 538-002B-7-25 or equiv. Avail. new from Newark Electronics. Avail. surplus from Reliance Merchandising Co., Phila. PA).

C19 — 15- to 60-pF miniature ceramic trimmer (Erie 538-002F-15-60 or equiv.).

C31 — 100-pF silver mica.

CR1 — Voltage-variable capacitor (Varicap) diode.

CR2 — High-speed silicon switching diode.

L — 1 to 2 μH inductor. 20 turns No. 30 enam. close-wound on 100,000 ohm, 1-watt resistor.

L1 — 5 turns No. 16 tinned bus wire, 1/4-inch ID × 5/8 inch long. Tap at 1-1/2 turns from 12-volt end.

L2 — 3 turns No. 16 tinned bus wire, 1/4-inch ID × 3/8 inch long. Tap at 1/2 turn from C13 end.

L3 — 4 turns No. 22 enam. wire, close-wound, 1/4-inch ID.

L4 — 25 turns No. 28 enam. wire, close-wound on body of 100,000-ohm, 1-watt resistor. Use resistor pigtails as anchor points for ends of winding.

L5 — 5 turns No. 16 tinned bus wire, 5/16 ID × 1/2 inch long.

Q1-Q3, incl. — RCA transistor.

R17 — 10,000-ohm pc-board carbon control linear taper (Mallory MTC 14L1 or equiv.).

RFC1 — 1-mH miniature rf choke (Millen J300-25).

RFC3, RFC4 — 10-μH miniature rf choke (Millen J300-10).

RFC5 — 10-μH miniature rf choke (Millen J300-10) with one Amidon ferrite bead over ground-end pigtail.

RFC6, RFC9 — 4 Amidon ferrite beads on 1/2-inch length of No. 24 wire (Amidon Associates, 12033 Otsego St., No. Hollywood, CA 91607).

RFC7, RFC8 — Same as RFC6 but with three beads on 3/8-inch length of wire.

S1 — Spdt slide or rotary switch.

U1 — RCA integrated circuit.

VR1 — 9.1 volt, 1-watt Zener diode.

Y1, Y2 — 18-MHz crystal (International Crystal Co. ground for 20-pF load capacitance. HC-25/U holder. Use International FM-2 pc-board crystal socket). High accuracy .002 percent temperature-tolerance crystal recommended.

boiled-down version is based on amateur-band performance criteria and the more commonly available supply voltage of 12. Emphasis has been placed on good frequency stability, narrow-band deviation (up to 6 kHz), and relative freedom from spurious output.

Low-cost transistors are used at Q1 and Q2. A ballasted transistor (mismatch protected) is used at Q3 to prevent burnout resulting from temporary open- or short-circuit conditions in the antenna system. The current OEM price (single lot) for the 2N5913 is $3.63. Over-the-counter prices will be slightly higher, but it is recommended that the builder use the '5913 if he wishes to have the circuit perform as specified here. Substitutes for any of the devices used in the circuit should be employed only by those who are experienced in semiconductor work. The wrong choice can lead to dismal results with the circuit – instability, low output, or destruction of one or more of the transistors.

Ferrite beads are used generously in the circuit, for decoupling of the dc bus and as rf chokes.[2] The beads provide low-Q impedances and are superior to solenoid-wound inductors in preventing circuit instability caused by tuned-base-tuned-collector conditions. A further aid to stability is provided through the use of high and low values of capacitance (combined) in various parts of the circuit. This standard technique helps to assure stability at hf and vhf, and is necessary because of the high f_T of the transistors used.[3]

Transistor sockets should not be used at Q1, Q2 or Q3. The additional lead lengths resulting from the use of sockets could lead to instability problems. Those wishing to use a socket at U1 may do so by redesigning the pc board to allow a socket to be installed (bringing the twelve holes for the IC closer together). Alternatively, one might employ an IC socket which has fairly long lugs, bending the lugs outward to mate with the holes in the pc-board.

Speech Amplifier

U1 consists of four bipolar transistors on a common substrate. Two of the transistors are connected for use as a Darlington pair. The remaining two are separate from one another. In the circuit of Fig. 1 the Darlington pair serves as a preamplifier for a high-impedance crystal, ceramic, or dynamic microphone. One of the separate transistors is used as a diode in the clipper circuit (an outboard silicon diode is used to clip the opposite side of the af sine wave) and the remaining transistor amplifies the clipped audio after it is filtered by an R-C network. Deviation is set by adjustment of a pc-board potentiometer, R17.

The processed audio is fed to CR1, the varactor diode modulator. Some reverse bias is used on CR1 to assure greater linearity of modulation (3 volts dc taken from the junction of R3 and R4). As the audio voltage is impressed across CR1, the junction capacitance of the diode shifts above the steady-state value which exists when no af voltage is present. The change in capacitance shifts the crystal frequency above and below its nominal value to provide fm.

Construction

There are no special instructions provided the builder follows the template pattern offered.[4] However, it is worth mentioning that the *QST*

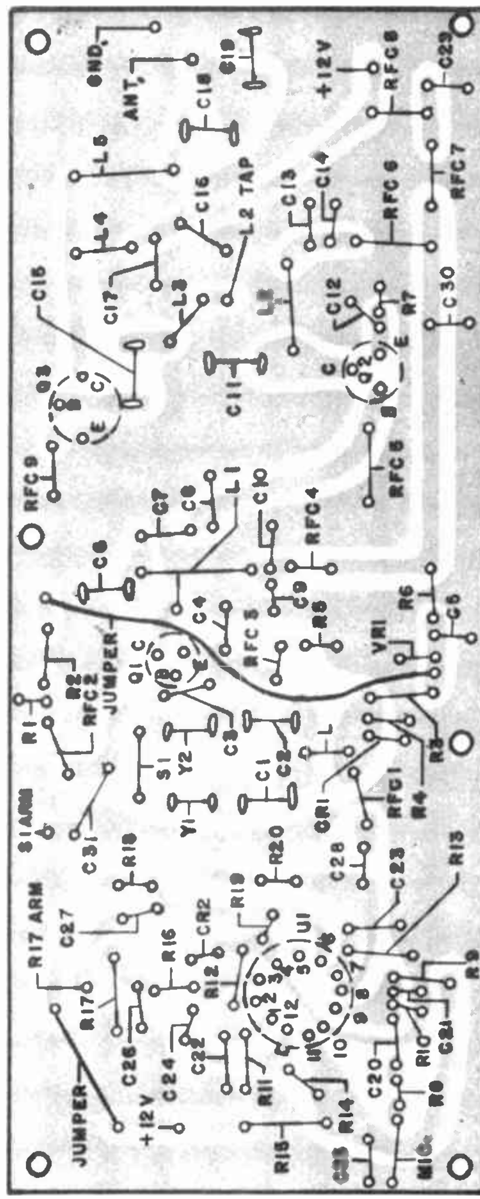


Fig. 2 — Template and parts layout for the transmitter drawn to full scale. Foil-side view.

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[2] See parts list for ordering information.

[3] The higher the f_T (upper-frequency rating) of a transistor, the greater will be its gain capability at lower frequencies, thus giving rise to unwanted hf or lf oscillations.

model was built on glass-epoxy circuit board. Those attempting to use phenolic or other types of pc board may encounter difficulty in obtaining proper circuit performance. The dielectric properties of the various board materials are different, thereby causing different values of capacitance to exist between pc-board foil strips. The condition can cause instability, unwanted coupling, and tuned circuits that will not hit resonance. Some builders of the MK-1 transmitter learned this the hard way!

Transistors Q2 and Q3 require fairly hefty heat sinks if good efficiency and longevity of the devices is to be realized. Homemade sinks are shown in the photo. Each consists of a piece of 1/16-inch thick aluminum (brass or copper is ok) formed over a drill bit slightly smaller in diameter than a TO-5 transistor case. The aluminum can be crimped in a bench vise until it fits snugly around the drill body. Silicone grease should be used to coat the transistor bodies prior to installation of the heat sinks. The height of the sinks is 1 inch. The ID is approximately 1/4 inch.

Lead lengths of the wires going from the pc board to S1 should be kept short – preferably less than 1-1/2 inches long. Coaxial cable (50-ohm impedance) should be used between the antenna terminals on the pc board and the antenna connector. The shield braid must be grounded at each end of the cable. Similarly, shielded cable should be employed between the microphone jack and the audio-input terminals on the pc board.

Checkout and Use

Initial checkout should be undertaken at reduced supply voltage. Apply a voltage of between 6 and 12, making certain that a dummy load of approximately 50 ohms is connected to the output of Q3. A 56-ohm 2-watt resistor or a No. 47 pilot lamp will suffice. Using a wavemeter tuned to 73 MHz, adjust the collector tank of Q1 for a peak reading on the wavemeter. Next, set the wavemeter for operation at 146 MHz and adjust the collector tuned circuit of Q2 for maximum meter indication. The tank circuit of Q3 should be adjusted for maximum power output as observed on an rf wattmeter or Monimatch-type SWR indicator. A rough check can be made by using a No. 47 lamp as a load, adjusting for maximum bulb brilliancy. The next step is to raise the supply voltage to 12 and repeat the tweaking procedure outlined above. If all stages are functioning normally, a No. 47 lamp should illuminate to slightly more than normal brilliance. Power output into a 50-ohm load should be between 1-1/2 and 2 watts Current drain will be between 200 and 250 mA, speech amplifier included.

Adjustment of the transmitter frequency and deviation can best be done while using a vhf frequency counter and deviation meter. Alternatively, one can put the transmitter in service and ask one of the other fm operators in the area to observe his receiver's discriminator meter while you adjust your crystal trimmer for a zero reading. Deviation can be set reasonably close to the desired amount by comparing your modulation against that of other local stations, having a third operator report the comparisons.

This transmitter is well suited as a companion unit to the fm receiver described in Chapter 3, and in *QST*.[5] The two units can be packaged to form a trans-receiver for portable, mobile, or fixed-station use. The transmitter can be used to drive a high-power solid-state 2-meter amplifier, described later in this chapter, if one wishes to put on a pair of "boots."[6]

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[5] DeMaw, "A Single-Conversion 2-Meter FM Receiver," *QST*, August, 1972.

[6] Hejhall, "Some 2-Meter Solid-State RF Power-Amplifier Circuits," *QST*, May, 1972, p. 40.

[7] Write: Spectrum Research Laboratory, Box 5824, Tucson, AZ 85703.

2-METER SOLID-STATE RF POWER-AMPLIFIERS

The majority of the commercially made 2-meter fm transceivers available today have rf power-output levels of 1 to 15 watts. There are many occasions when an fm operator would like to have a little more power to be able to work over greater distances. Described here are two amplifiers, one for 25 watts and another for 50 watts output for the 2-meter band. Both amplifiers use a single transistor and operate directly from a 13.6-volt vehicular electrical system.

Circuit Description

The amplifier circuit shown in Fig. 14-53 utilizes a single 2N6084 transistor operated in a Class-C, zero-bias configuration. This mode of operation has the advantages of high collector efficiency at full output and zero dc current drain when no rf driving signal is applied. The reader should note that zero-bias operation yields an amplifier that is not a "linear." It is designed for

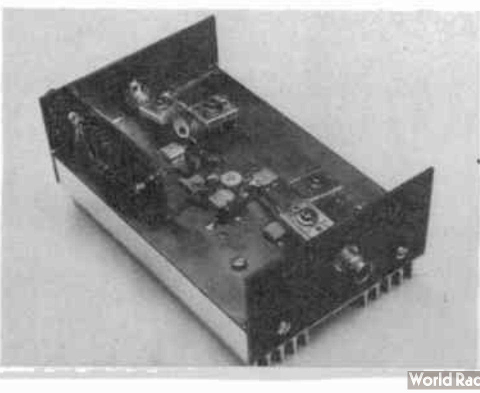
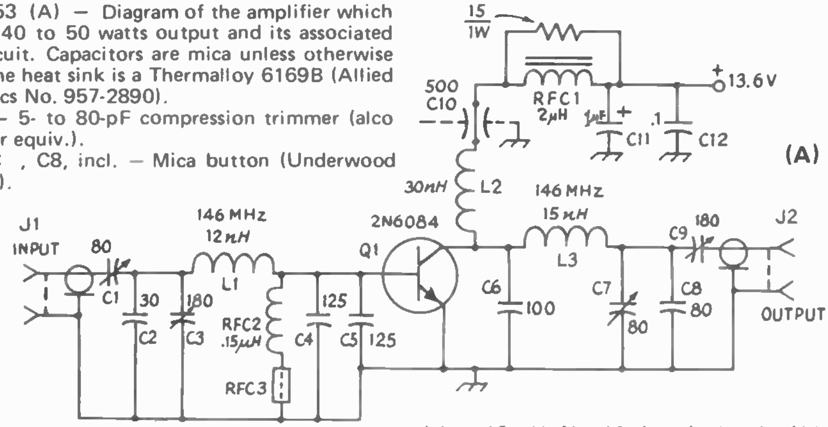


Fig. 14-52 — An end view of the breadboard version of the 50-watt 2-meter amplifier. The input circuit is at the lower right, and the output network is at the upper left.

Fig. 14-53 (A) — Diagram of the amplifier which provides 40 to 50 watts output and its associated COR circuit. Capacitors are mica unless otherwise noted. The heat sink is a Thermalloy 6169B (Allied Electronics No. 957-2890).

C1, C7 — 5- to 80-pF compression trimmer (alco 462 or equiv.).

C2, C4-C , C8, incl. — Mica button (Underwood J-101).



C3, C9 — 9- to 180-pF compression trimmer (Arco 463 or equiv.).

C10 — Feedthrough type.

C11 — Tantalum.

C12, C15, C16 — Ceramic disk.

C13, C14 — 39-pF mica (Elmenco 6ED390J03 or equiv.).

CR1 — 100-PRV or more, 500-mA or more silicon diode (Motorola 1N4001 or equiv.).

CR2,CR3 — High-speed, low capacitance 100-PRV silicon diode (Motorola MSD7000 dual package used here).

J1,J2 — Coaxial connector, panel mount.

K1 — 4pdt open-frame relay, 12-V contacts (Comar CRD-1603-4S35 or equiv., Sigma 67R4-12D also suitable), modified as described below.

L1 — 12 nH, No. 10 tinned wire, 1 1/4-inch long straight conductor.

L2 — 30 nH, 1 3/4 turns, No. 10 tinned wire, 3/8 inch ID, 3/4 inch long.

L3 — 15 nH, No. 14 tinned wire, 3/4-inch long straight conductor.

L4 — 2 turns of No. 18 tinned wire 1/4-inch ID, 0.2 inch long (approximately 44 nH).

Q1 — Motorola silicon power transistor.

Q2 — Npn silicon Darlington transistor, h_{FE} of 5000 or more (Motorola MPS-A13 or equiv.).

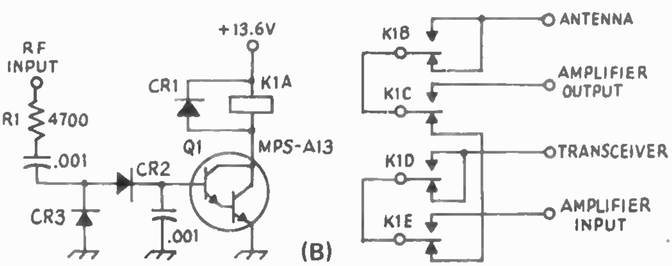
R1 — 15 ohm, 1-watt composition.

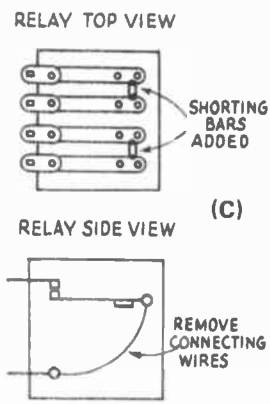
R2 — 4700 ohm 1/2-watt composition

RFC1 — 17 turns, No. 16 enam. wire wound on Amidon T-80-2 toroid core.

RFC2 — Molded rf choke (J. W. Miller 9250-15).

RFC3 — Ferrite bead (Ferroxcube 56-590-65/3B or equiv.).





(B) COR circuit. Capacitors are disk ceramic.

The COR relay is modified by removing the connecting wires from all four wiper arms and adding two shorting bars, as shown. Only the stationary-contact connections are used.

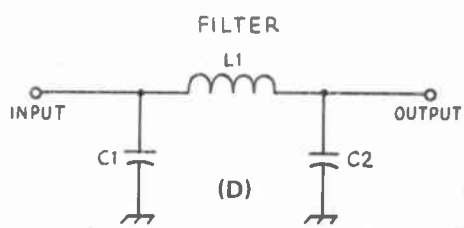
CR1 — 100-PRV or more, 500-mA or more silicon diode (Motorola 1N4001 or equiv.).

CR2, CR3 — High-speed, low-capacitance 100-PRV silicon diode (Motorola MSD7000 dual package used here).

K1 — 4pdt open-frame relay, 12-V contacts (Comar CRD-1603-4S35 or equiv., Sigma 67R4-12D also suitable), modified as described below.

Q2 — Npn silicon Darlington transistor, h_{FE} of 5000 or more (Motorola MPS-A13 or equiv.).

(C) The COR relay is modified by removing the connecting wires from all four wiper arms and adding two shorting bars, as shown. Only the stationary-contact connections are used.



(D) Pi-section output filter, C1 and C2 are 39-pF mica capacitors (Elmenco 6ED390J03 or equiv.), and L1 consists of 2 turns of No. 18 tinned wire, 1/4 inch ID, 0.2 inch long (approximately 44 nH).

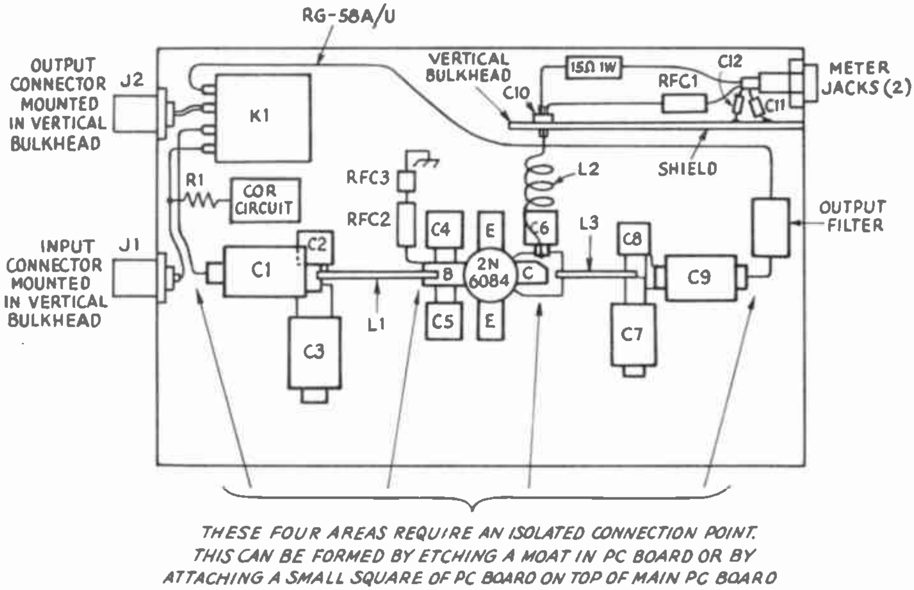


Fig. 14-54 — Parts-layout diagram for the 50-watt amplifier (not to scale). A 4 × 6-inch pc board is used as the base.

fm (or cw– operation only, and would produce objectional distortion and splatter if used to amplify either a-m or ssb signals.

The amplifier operates directly from an automobile electrical system, so no additional power supply is required for mobile operation. The input and output-tuned circuits are designed to match the impedances of the transistor to a 50-ohm driving source and to a 50-ohm antenna system, respectively. Since both the input and output impedances of the transistor are extremely low (in the 1- to 5-ohm region), the matching networks employed are somewhat different than those used with tubes. The networks chosen for the amplifier are optimized for low-impedance matching, and they perform their tasks efficiently. The network designs for this amplifier were done with the aid of a computer.

The elaborate decoupling network used in the collector dc feed is for the purpose of assuring amplifier stability with a wide variety of loads and tuning conditions. The 2N6084 transistor is conservatively rated at 40 watts output (approximately 60 watts dc input). The amplifier can readily be driven to power output levels considerably higher than 40 watts, but it is recommended that it be kept below 50 watts output. If your transmitter or transceiver has greater than 10 watts output, an attenuator should be used at the amplifier input to keep the output from the amplifier below 50 watts.

Construction Details

Construction of the amplifier is straightforward. The usual precautions that must be observed when building a solid-state final amplifier are followed. These precautions include proper mechanical mounting of the transistor, emitter grounding, heat sinking, and decoupling of the supply-voltage leads. Most of the components used are conventional items which are readily available, with two exceptions. The fixed mica capacitors, Underwood type J-101, are a special mica unit designed for high-frequency applications. The core for RFC1 and the rf bead used for RFC3 are available from Elna Ferrite Labs, Inc., 9 Pine Grove St., Woodstock, NY 12498.

The amplifier is constructed on a pc board which is bolted to a heat sink. A few islands can be etched on the board for tie points, at the builder's discretion; a complex foil pattern is not required. In the amplifier shown in the photo, islands were etched only for input and output tie points. Circuit-board islands may also be etched for the transistor base and collector leads. However, an interesting alternative method was used in the author's breadboard amplifier. The base and collector islands were formed by attaching small pieces of pc board to the top of the main board. This procedure added a few tenths of a pF of capacitance at the connection points, so if you choose to etch islands directly on the main board you may want to increase the value of C6 slightly. (The values of C4 and C5 are not critical.)

A word about the care of a stud-mount rf power transistor: Two of the most important mounting precautions are (1) to assure that there is no upward pressure (in the direction of the ceramic cap) applied to the leads, and (2) that the nut on the mounting stud is not over-tightened. The way to accomplish item 1 is to install the nuts *first* and solder the leads to the circuit later. For item 1, the recommended stud torque is 6 inch-pounds. For those who don't have a torque wrench in the

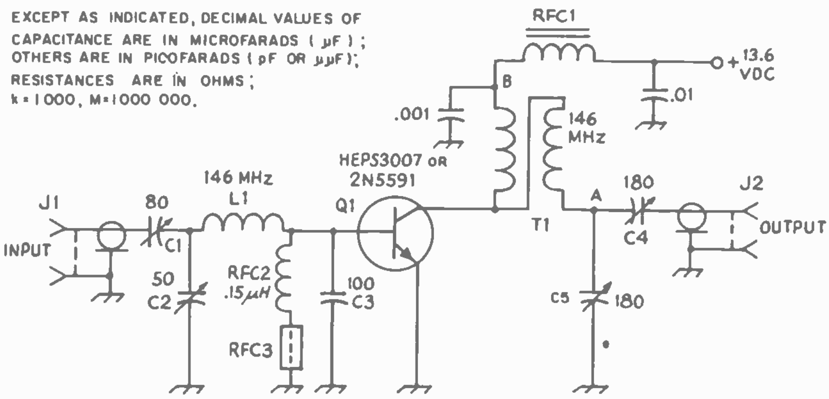


Fig. 14-55 — Circuit diagram of the 25-watt amplifier. Capacitors are disk ceramic unless otherwise noted.

C1 — 5- to 80-pF compression trimmer (Arco 462 or equiv.).
C2 — 2- to 50-pF compression trimmer (Arco 461 or equiv.).
C3 — Button mica (Underwood J-101).
C4, C5 — 9- to 180-] -pF compression trimmer (Arco 463 or equiv.).
J1, J2 — Coaxial connector, panel mount.
L1 — 1-inch length of No. 14 tinned wire.
Q1 — Motorola silicon power transistor (2N5591 or HEP S3007 for 25 W output, 2N5590 or HEP S3006 for 10 W output).
RFC1 — Ferroxcube VK200-19/4B ferrite choke.
RFC2 — Molded rf choke (J. W. Miller 9250-15).
RFC3 — Ferrite bead (Ferroxcube 56-590-65/3B or equiv.).
T1 — See Fig. 14-56.

shack, remember that it is better to under tighten than to over tighten the mounting nut.

The transistor stud is mounted through a hole drilled in the heat sink. A thermal compound, such as Dow Corning 340 heat-sink grease, should be used to decrease the thermal resistance from transistor case to heat sink. See the excellent article by White in *QST* for April, 1971, for details of heat-sink design.

Series impedance in the emitter circuit can drastically reduce the gain of the amplifier. Both transistor emitter leads should be grounded as close to the transistor body as is practical.

The wiring for the dc voltage feeder to the collector should have extremely low dc resistance. Even a drop of one volt can significantly reduce the power output of the amplifier. A good goal is less than 0.5 volt drop from the car battery to the transistor collector. With operating currents of several amperes, a total dc resistance of only a fraction of an ohm is needed. A standard commercially made heat sink is used for the 50-watt amplifier, and it is adequate for amateur communications. Forced-air cooling across the heat sink should be used for any appliation requiring long-term key-down operation at 40 watts or more of output.

Tune-Up Procedure

Generally, the best way to tune a transistor final is for maximum rf power output. If this approach results in exceeding the power ratings of the transistor, then the power output should be reduced by reducing the drive-level, not by detuning the final. In the case of an outboard PA stage, such as described here, both the input and output networks can be tuned for maximum rf output, if the driving source has an output impedance of approximately 50 ohms. However, a better procedure consists of tuning the output tank circuit for maximum rf output and tuning the input circuit for minimum SWR as measured between the exciter and the final amplifier. This tune-up procedure has the added advantage of assuring that the amplifier presents a 50-ohm load to the exciter. A dc ammeter to check collector current is a useful tune-up aid. Since tuning is for peak output, a Monimatch-type SWR bridge is adequate for the job. Also, the wattmeter described in Chapter 22 would be an excellent choice. The best tuning procedure is to monitor simultaneously both output power (absolute or relative) and the SWR between the exciter and amplifier.

First, apply dc voltage with no rf drive. No collector current should flow. Then apply a low level of rf drive – perhaps 25 percent or less of the rated 10 watts maximum drive – and tune the input network for maximum indicated collector current. The networks may not tune to resonance at this low drive level, but you should at least get an indication of proper operation by smooth tuning and lack of any erratic behavior in the collector-current reading. Gradually increase the drive, retuning as you go, until the rated 7-10 watts input and 40 to 50 watts output are obtained. As power input is increased, use the recommended tuning procedure of maximum output from the output tank and minimum input SWR for the input circuit.

There is danger of low-frequency oscillations with most transistor amplifiers. A scope of 5-MHz or more bandwidth connected to the dc feeder at point A makes an excellent indicator of any low-frequency oscillation. It is possible to have

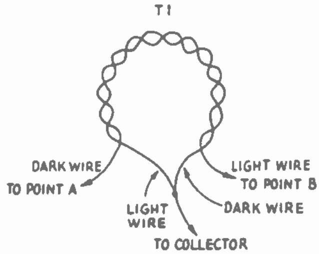


Fig. 14-56 – Transmission-line output transformer consisting of 2 4-inch long conductors, No. 20 enam. wire, twisted to 16 crests per inch, using an electric drill. The conductors should be color coded, one with one color and one with a second color. Form the twisted pair into a 1/2-inch dia circle. Unwind the leads so that only the portion of the pair forming the circle remains twisted. Connect the leads of each color as shown.

signal output on all hf and vhf amateur bands and all TV channels, simultaneously, when a bad case of parasitic oscillation occurs. For those who may have access to one, the best indicator of parasitic oscillation is a wide-band spectrum analyzer.

An Additional Design

For those who own a low-power fm transceiver, an intermediate amplifier stage or a final amplifier providing 10 to 25 watts may be desired. The circuit of Fig. 14-55 is suitable for the 2N5591 or HEP S3007 transistors (25 watts), and the 2N5590 or HEP 3006 transistors (10 watts). An unusual feature of this circuit is the use of a transmission-line transformer in the output network. The construction and tune-up procedures for the amplifiers of Fig. 14-55 is similar to that described earlier for the 50-watt amplifier.

Accessories

When an amplifier stage is used with an fm transceiver, a method of automatic transmit/receive switching is needed. A simple carrier-operated relay (COR), such as shown in Fig. 14-16 can be employed for the amplifiers described in this article. The level of input rf required to operate the COR is determined by the value of R1. One to two watts of 2-meter energy will operate K1 when a 4700-ohm resistor is employed. The rf signal is rectified by two high-speed switching diodes; the dc output from the rectifier is applied to Q1, a Darlington-connected transistor pair. When sufficient current is developed in the base circuit, Q1 will turn on, activating K1. A transient-suppression diode is included across the relay coil to prevent voltage-spike damage to Q1.

The switching circuits needed to take the amplifier in and out of the circuit- are somewhat complex. The cost of four coaxial relays would be prohibitive. But, an open-frame relay can cause sufficient loss at 146 MHz to severely degrade the sensitivity of the associated receiver. To get around this problem the author modified an inexpensive relay. The long leads to the wiper arms were removed and discarded. Two shorting bars were added, as shown in the drawing. External connections were made only to the stationary contacts. Received signal loss through the modified relay measured 0.4 dB – an insignificant amount.

Second-harmonic output from the 50-watt amplifier measured 34 dB down from the level of the 146-MHz energy. Thus, the computer-design output network compares favorably with the pi-section tank circuits often used in hf transmitters. To assure that harmonic energy didn't cause a problem to other services, a simple pi-section output filter was added. This filter is designed for 50-ohm input and output impedances; it can be used with any two-meter amplifier. The insertion loss of the filter at 146 MHz is 0.2 dB, while it provides 46 dB attenuation at 292 MHz and 25 dB at 438 MHz.

Appendix A

1) Amidon toriod cores are available from Amidon Associates, 12033 Otsego Street, No. Hollywood, CA 91607.

2) Ferroxcube components can be purchased from Elna Ferrite Laboratories, Inc., 9 Pine Grove Street, Woodstock, NY 12498.

3) J. W. Miller chokes are available from distributors, or directly from J. W. Miller, 19070 Reyes Ave., Compton, CA 90224.

4) Underwood mica capacitors must be ordered directly from the manufacturer, Underwood Electric and Manufacturing Company, Inc., P. O. Box 188, Maywood, IL 60153. Price for the J-101 units specified in this article is approximately $1.20 each (specify the value — in pF — desired).

5) A circuit board for the 50-watt amplifier will be available from Spectrum Research Labs, P. O. Box 5824, Tucson, AZ 85703.

2-METER FM RECEIVER

An fm purist is not likely to settle for second-rate receiver performance in this day of vhf-band saturation. A satisfactory fm receiver must be able to separate the various repeater output frequencies without being affected by IMD and overload problems. The sensitivity must be good, and so should the limiting characteristics. Few low-cost designs satisfy the foregoing criteria. The circuit of Fig. 14-58 represents a practical compromise between cost and circuit complexity, yet provides performance which is comparable to that of many commercial fm receivers in use by amateurs.

The single-conversion solid-state fm receiver described here is intended as a mate for the transmitter shown in Fig. 14-50. This design centers around a multifunction IC, the CA3089E. Circuit simplicity, good performance, and low cost are the keynotes in this project.

Circuit Highlights

A JFET was chosen for rf amplifier Q1, Fig. 14-58. Neutralization is unnecessary provided the gate and drain elements are tapped down on their

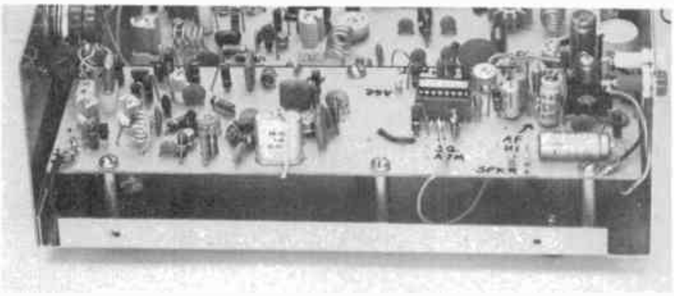


Fig. 14-57 – This photo shows the final breadboard version of the fm receiver. Some of the bypass capacitors are located on the foil side of the pc board in this example. The template and parts-layout sheet provides for topside mounting of the capacitors. The differences between the receiver shown here and the final model are quite minor.

respective tuned circuits. For simplicity s sake only two tuned circuits are used ahead of the mixer, which uses a dual-gate MOSFET. The combination of FETs Q1 and Q2 assures low IMD and provides good immunity to overloading. Output from the mixer is supplied to FL1. This is a four-pole 10.7-MHz i-f filter which is fed from a 900-ohm tap point on tuned circuit C9-C10-L3.

The oscillator/multiplier stage, Q3, is a carbon copy of that used by Pearce-Simpson in their Gladding 25 fm transceiver. It is one of the simplest circuits one can use, yet it performs well. Injection to the mixer is supplied at 157 MHz (10.7-MHz i-f plus the frequency of the received signal). The oscillator crystal frequency is one half the injection frequency – 78 MHz in this example. No netting trimmers are necessary if crystals for the Gladding circuit are ordered and used. Frequency doubling from 78 MHz is accomplished in the collector circuit of Q3.

I-f amplifier U1 is a CA3028A wired for cascode operation. FL1 connects to input terminal 2 through a .01-μF blocking capacitor. Terminating resistor R7 is selected for the characteristic impedance of the filter used. The KVG filter has a 910-ohm bilateral impedance, so if precise matching is desired one can use a 910-ohm unit at R7. Output from U1 is fed to multifunction chirp U2, across R11.

Audio output from U2 is amplified by Q4 before being routed to U3, a transformerless 1-watt output IC. Though the MC1454 is designed to work into a 16-ohm speaker, good results can be had when using an 8-ohm speaker.

Construction

How the receiver is packaged can best be decided by the builder. Two choices are offered: dividing the board in two parts and stacking one section above the other on standoff posts. If this is done it will be necessary to cut the board midway between U1 and U2. If compactness is not necessary the constructor can follow a one-piece assembly format, keeping the board its 8 × 2-11/16-inch size.

Those who desire additional crystal positions can make the board slightly longer. This will provide room for more crystal sockets, but will require that a switch with more positions be used for S1.

It is recommended that transistor and IC sockets be avoided except at Q4 and U3. Short leads between the bodies of the devices and the pc board must be maintained to prevent unstable operation. The use of sockets will cause instability unless low-profile receptacles are used. Similarly, the pigtails on the bypass capacitors should be kept as short as possible in all parts of the rf circuit.

The wiring which connects the audio and squelch controls to the circuit board should be of the shielded variety. If the board is cut into two sections, as mentioned earlier, use shielded cable between U1 and U2, routing the i-f signal from pin 6 of U1 to pin 1 of U2. *Don't leave out C16.*

The leads from S1 to the crystal sockets must be kept as short as possible – less than 1-1/2 inches each. As a further aid to circuit stability mount the pc board on a metal cabinet wall or chassis by means of four or six metal standoff posts. This technique is beneficial in preventing rf ground loops.

Checkout and Alignment

It should be stressed that there is no simple way to align an fm receiver. A stable signal generator will be required, preferably one with fm capability. Initial alignment cannot be properly effected by using off-the-air fm signals. A weak-signal source can be built by using the modulator and crystal oscillator stage of the low power transmitter described in Fig. 14-50. Whatever method is used, make certain that the test signal is no farther off frequency than 200 Hz from the desired frequency of reception. Ideally, the signal source should be *exactly* on the chosen input frequency of the receiver.

Connect the signal generator to J1. Attach a meter across J2 and J3. Make certain that a speaker is hooked to the output of U3. Assuming that an ohmmeter check shows no shorted or open circuits in the completed assembly, connect a 12-volt dc supply to the receiver. With the squelch turned off (maximum hiss noise) adjust C2, C4, and C44 for an upward deflection of the relative-signal-strength meter (at J2 and J3). Next, adjust C10 for maximum meter reading. Repeat these steps two more times. All tuning adjustments should provide fairly sharp peaks when the circuits are tuned to resonance.

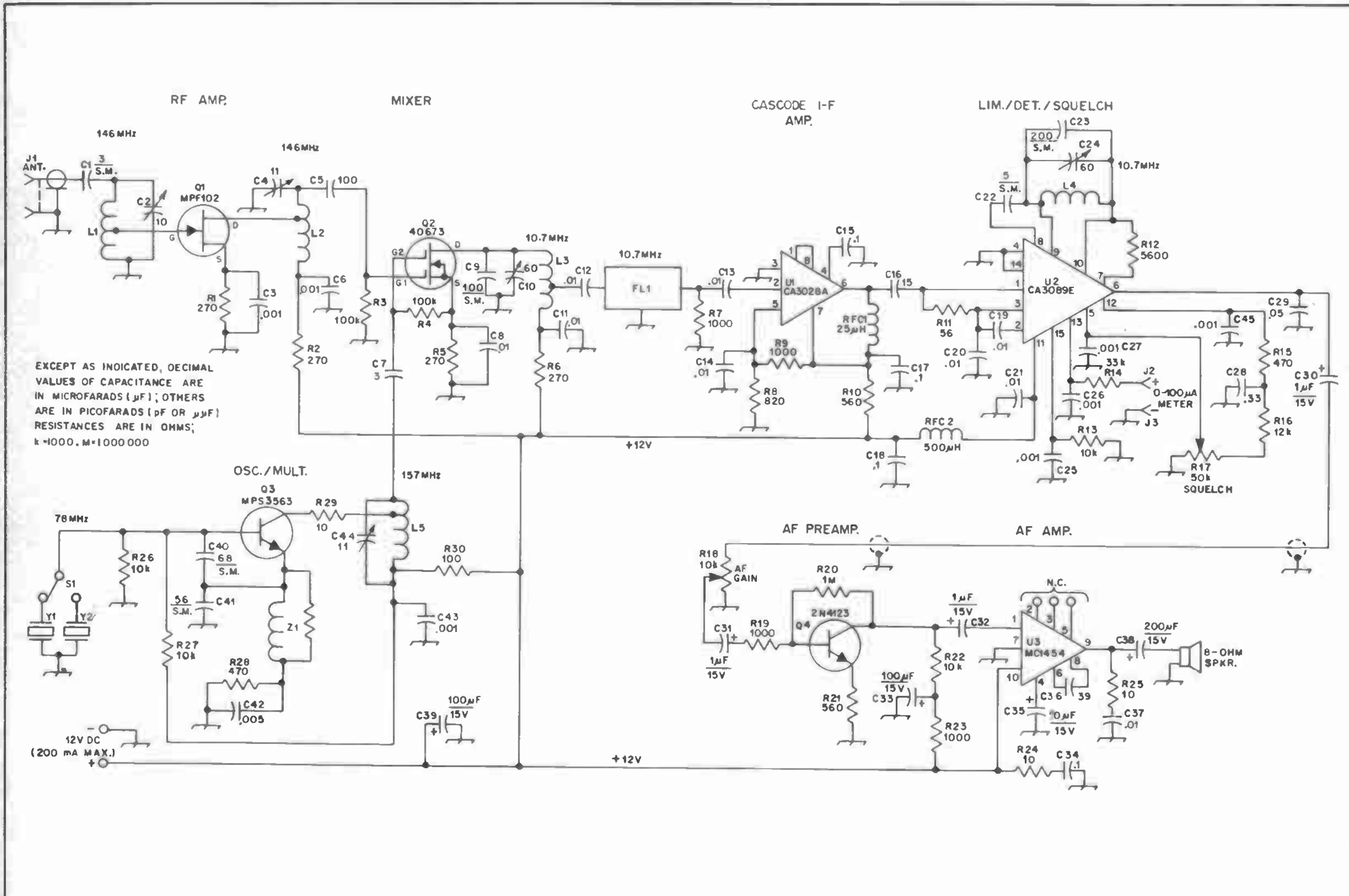

RF AMP.
MIXER
CASCODE I-F AMP.
LIM./DET./SQUELCH
OSC./MULT.
AF PREAMP.
AF AMP.
146 MHz
10.7 MHz
157 MHz
78 MHz
J1 ANT.
Q1 MPF102
Q2 40673
Q3 MPS3563
Q4 2N4123
U1 CA3028A
U2 CA3089E
U3 MC1454
FL1
R17 50k SQUELCH
R18 10k AF GAIN
0-100μA METER
8-OHM SPKR.
12V DC (200 mA MAX.)
+12V
EXCEPT AS INDICATED, DECIMAL VALUES OF CAPACITANCE ARE IN MICROFARADS (μF); OTHERS ARE IN PICOFARADS (pF OR μμF); RESISTANCES ARE IN OHMS; k=1000, M=1000 000

Fig. 14-58 — Schematic diagram of the fm receiver. Fixed-value capacitors are disk ceramic unless otherwise noted. Polarized capacitors are electrolytic. Fixed-value resistors are 1/2-watt composition types. Numbered components not in parts list are identified for pc-board layout purposes only. A template and parts layout diagram are available from ARRL Hq. for $0.50. A circuit board is available from Spectrum Research Labs., Box 5824, Tucson, AZ 85703.

C1 — 3-pF silver mica. For precise matching substitute a 10-pF trimmer for the fixed-value capacitor.
C2, C4, C44 — 11-pF pc-mount miniature air variable (E.F. Johnson 187-0106-005, avail. from Newark Electronics).
C10, C24 — Subminiature ceramic trimmer, 15 to 60 pF (Erie 538-011F-15-60, available from Newark Electronics, or surplus from Reliance Merchandising, 2223 Arch St., Phila., PA 19103).
FL1 — See text and footnotes.
J1 — Coax receptacle of builder's choice.
J2, J3 — Binding posts of pin jacks.
L1 — 5 turns No. 18 tinned bus wire, 1/4-inch ID x 1/2-inch long. Tap gate of Q1 at 2-1/2 turns from hi-Z end.
L2 — 4 turns tap at 1 turn from the top (high-impedance) end.
L3 — 16 turns No. 22 enam. wire to occupy entire circumference of Amidon T-50-2 toroid core. Tap at 6 turns from hi-Z end for FL1.
L4 — 1-µH high-Q inductor, unloaded Q 150 or greater. 18 turns No. 24 enam. wire to occupy entire circumference of Amidon T-37-2 toroid core (Amidon Assoc., 12033 Otsego St., N. Hollywood, CA 91607).
L5 — 5 turns No. 18 tinned bus wire, 1/4-inch ID x 1/2 inch long. Tap collector via R29 at 1 turn from hi-Z end.
Q1, Q3, Q4 — Motorola Transistor.
Q2 — RCA MOSFET
R17 — 50,000-ohm linear-taper carbon control.
R18 — 10,000-ohm audio-taper carbon control.
RFC1 — 25-µH choke (Millen J300-25 or equivalent).
RFC2 — 500-µH choke (Millen J300-500 or equivalent).
S1 — Spdt miniature slide or wafer switch, nonshorting.
U1, U2 — RCA integrated circuit.
U3 — Motorola integrated circuits.
Y1, Y2 — International Crystal Co. receiving crystal ground to Pearce-Simpson Gladding 25 specs. Nylon crystal sockets available from International Crystal Co.
Z1 — Rf choke consisting of 8 turns No. 24 enam. wire, close-wound on body of 1000-ohm 1/2-watt carbon resistor. Solder coil leads to resistor pigtails.

A frequency-modulated signal will be required for on-the-nose adjustment of the detector (L4 and C24). C24 should be adjusted slowly until the point is found where best audio quality occurs. Audio recovery will be the lowest at this point, creating the illusion of reduced receiver sensitivity. If no fm signal is available for this part of the alignment, tune the detector for minimum hiss noise as heard in the speaker. After the detector is aligned, readjust C10 for best audio quality of a received fm signal. It may be necessary to go back and forth between C10 and C24, carefully tweaking each capacitor for the best received-signal audio quality. The detector should be adjusted while a strong signal (100 µV or greater) is being supplied at J1.

Adjustment of the squelch control should provide complete muting of the hiss noise (no signal present) as approximately midrange in its rotation. If the audio channel is functioning properly one should find that plenty of volume occurs at less than a midrange setting of R18.

Performance

In two models built, both identical to the circuit of Fig. 14-58, sensitivity checked out at roughly 0.8 µV for 20 dB of quieting. This sensitivity figure is by no means spectacular, but is quite ample for work in the primary signal contour of any repeater. The addition of a dual-gate MOSFET preamplifier ahead of Q1 resulted in a sensitivity of 0.25 µV for 20 dB of quieting. The barefoot receiver requires approximately 0.5 µV of input signal to open the squelch. A more elaborate circuit would have provided greater sensitivity, but at increased cost and greater circuit complexity.

Hard limiting occurs at signal input levels in excess of 10 µV, with 3 dB of limiting exhibited at 1 µV. Addition of an outboard preamplifier will greatly improve the limiting characteristics, and this would benefit those who are dealing primarily with weak signals.

A KVG XM 107S04 i-f filter (FL1)[1] is used in the circuit of Fig. 14-58. However, any 10.7-MHz filter with suitable handwidth characteristics for amateur fm reception can be substituted for the unit specified. During the development period a Piezo Technology Comline filter was used at FL1.[2] The model tried was a PTI 2194F, which sells for $10 per unit in single lots. Club groups may wish to take advantage of the 5 to 9 price break . . $5.95 each. The PTI 2194F gave performance similar to that of the KVG unit.

Each brand of filter has its own characteristic impedance, so if substitutions are made it will be necessary to change the tap position on L3 to assure a proper match between Q2 and FL1. Similarly, the ohmic value of R7 will have to be changed.

1 A product review describing the filter's characteristics was given in *QST* for June,, 1972, p. 56. The filter sells for $15.95 and can be ordered from Spectrum International, Box 87, Topsfield, MA 01983. A drilled printed circuit board is available for $5 from: D.L. McClaren, W8URX, 19721 Maplewood Avenue, Cleveland, OH 44135.

2 Piezo Technology Inc., Box 7877, Orlando, FL 32804.

Specialized Communications Systems

The field of specialized amateur communications systems includes radioteletype, amateur television, amateur facsimile, phone patching, and space and satellite communications. Radio control of models is not a "communications" system in the amateur (two-way) sense. The specialized hobby of radio control does have a large following, but "citizen-band" provisions for frequency allocations and operator registrations divorce it from the strictly ham-radio field (unless one wishes to avoid the QRM).

By far the greatest activity in the specialized fields is to be found in radioteletype (RTTY). Operation using frequency-shift keying techniques is permitted on all amateur bands except 160 meters.

Activity in amateur TV (ATV) can be found primarily in a number of population centers around the country. Most of the work is based on converted entertainment receivers and manufacturer's-surplus camera tubes (vidicons). ATV is permitted on the amateur bands above 420 MHz, and this and the broadband nature of the transmissions precludes extensive DX work.

Slow-scan TV (SSTV) is a narrow-band system that is permitted in any of the phone bands except 160 meters. It is a completely electronic system, however; no photographic techniques are required. Pictures are transmitted in 8 seconds or less.

Amateur facsimile operation, under present U.S. regulations, is permitted only above 50.1 MHz. Operation in the 6- and 2-meter bands is restricted to the use of shifting audio tones with an amplitude-modulated carrier (A4 emission), so operation through an fm repeater on these bands is prohibited. Facsimile operation is undertaken primarily by groups in heavily populated areas.

Amateur satellites – called *Oscars* for Orbiting Satellites Carrying Amateur Radio – offer another way of extending the range of vhf and uhf stations. Satellites can also operate in the hf region to provide communication during times of poor ionospheric conditions.

Phone patches permit third parties to communicate via amateur radio, through an interconnection between the amateur's station equipment and his telephone line. With voice operation in use, phone patching may be conducted in any amateur voice band between domestic stations, or between stations of any two countries permitting third-party communications.

RADIOTELETYPE (RTTY)

Radioteletype (abbreviated RTTY) is a form of telegraphic communication employing typewriter-like machines for (1) generating a coded set of electrical impulses when a typewriter key corresponding to the desired letter or symbol is pressed, and (2) converting a received set of such impulses into the corresponding printed character. The message to be sent is typed out in much the same way that it would be written on a typewriter, but the printing is done at the distant receiving point. The teletypewriter at the sending point may also print the same material.

The teleprinter machines used for RTTY are far too complex mechanically for home construction, and if purchased new would be highly expensive. However, used teletypewriters in good mechanical condition are available at quite reasonable prices. These are machines retired from commercial service but capable of entirely satisfactory operation in amateur work. They may be obtained from several sources on condition that they will be used purely for amateur purposes and will not be resold for commercial use.

Some dealers and amateurs around the country make it known by advertising that they handle parts or may be a source for machines and accessory equipment. *QST*'s Ham-Ads and other publications often show good buys in equipment as amateurs move about, obtain newer equipment, or change interests.

Periodic publications are available which are devoted exclusively to amateur RTTY. Such publications carry timely technical articles and operating information, as well as classified ads.

The Teletype Corp. Model 28ASR teleprinter is used by many amateurs. In addition to the keyboard and page printer, this model contains facilities for making and sending perforated tapes.

Over the years *QST* has carried a number of articles on all aspects of RTTY, including a detailed series by Hoff in 1965 and 1966. For a list of surplus-equipment dealers, information on publishers of RTTY periodicals, and a bibliography of all articles on RTTY which have appeared in *QST*, write to RTTY T.I.S., ARRL Headquarters, 225 Main Street, Newington, CT 06111. U.S. residents should enclose a stamped business-size envelope bearing a return address with their request.

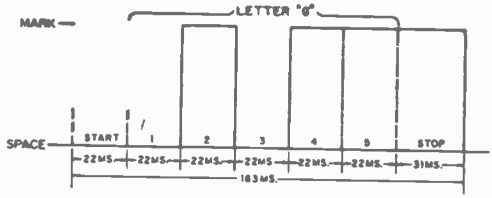


Fig. 15A-1 — Pulse sequence in the teleprinter code. Each character begins with a start pulse, always a "space," and ends with a "stop" pulse, always a "mark." The distribution of marks and spaces in the five elements between start and stop determines the particular character transmitted.

Types of Machines

There are two general types of machines, the page printer and the tape printer. The former prints on a paper roll about the same width as a business letterhead. The latter prints on paper tape, usually gummed on the reverse side so it may be cut to letter-size width and pasted on a sheet of paper in a series of lines. The page printer is the more common type in the equipment available to amateurs.

The operating speed of most machines is such that characters are sent at the rate of either 60, 67, 75 or 100 wpm depending on the gearing ratio of a particular machine. Current FCC regulations allow amateurs the use of any of these four speeds. Interchangeable gears permit most machines to operate at these speeds. Ordinary teletypewriters are of the start-stop variety, in which the pulse-forming mechanism (motor driven) is at rest until a typewriter key is depressed. At this time it begins operating, forms the proper pulse sequence, and then comes to rest again before the next key is depressed to form the succeeding character. The receiving mechanism operates in similar fashion, being set into operation by the first pulse of the sequence from the transmitter. Thus, although the actual transmission speed cannot exceed about 60 wpm (or whatever maximum speed the machine is geared for), it can be considerably slower, depending on the typing speed of the operator.

It is also possible to transmit by using perforated tape. This has the advantage that the complete message may be typed out in advance of actual transmission, at any convenient speed; when transmitted, however, it is sent at the machine's normal maximum speed. A special tape reader, called a transmitter-distributor, and tape perforator are required for this process. A reperforator is a device that may be connected to the conventional teletypewriter for punching tape when the machine is operated in the regular way. It may thus be used either for an original message or for "taping" an incoming message for later retransmission.

Teleprinter Code

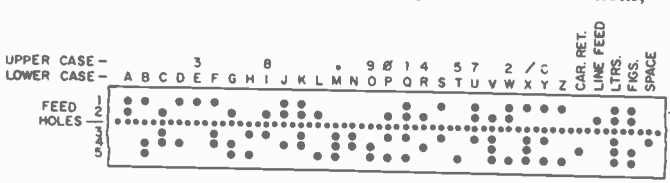
In the special code used for teleprinter operation, every character has five "elements" sent in sequence. Each element has two possible states, either "mark" or "space," which are indicated by different types of electrical impulses (i.e., mark might be indicated by a negative voltage and space by a positive voltage). At 60 wpm each element occupies a time of 22 milliseconds. In addition, there is an intial "start" element (space), also 22 ms long, to set the sending and receiving mechanisms in operation, and a terminal "stop" element (mark) 31 ms long, to end the operation and ready the machine for the next character. This sequence is illustrated in Fig. 15A-1, which shows the letter G with its start and stop elements.

At maximum machine speed, it takes 163 ms to send each character. This is the equivalent of 368 operations per minute. At 75 wpm with this same code, 460 operations per minute result, and 600 for 100 wpm. The letter code as it appears on perforated tape is shown in Fig. 15A-2, where the black dots indicate marking pulses. Figures and arbitrary signs – punctuation, etc. – use the same set of code impulses as the alphabet, and are selected by shifting the carriage as in the case of an ordinary typewriter. The carriage shift is accomplished by transmitting either the "LTRS" or "FIGS" code symbol as required. There is also a "carriage return" code character to bring the carriage back to the starting position after the end of the line is reached on a page printer, and a "line feed" character to advance the page to the next line after a line is completed.

Additional System Requirements

To be used in radio communication, the pulses (dc) generated by the teletypewriter must be utilized in some way to key a radio transmitter so they may be sent in proper sequence and usable form to a distant point. At the receiving end the incoming signal must be converted into dc pulses suitable for operating the printer. These functions,

Fig. 15A-2 — 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.



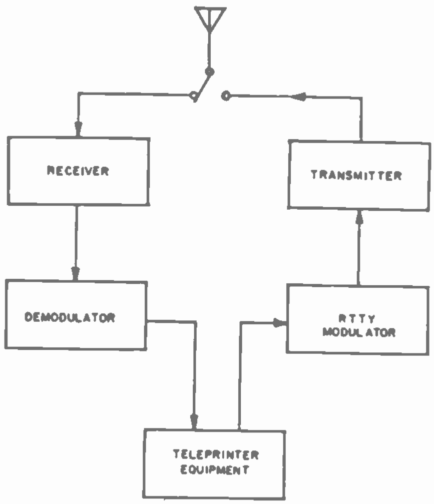


Fig. 15A-3 – Block diagram showing the basic equipment required for amateur RTTY operation.

shown in block form in Fig. 15A-3, are performed by electronic units known respectively as the frequency-shift keyer or RTTY modulator and receiving converter or RTTY demodulator.

The radio transmitter and receiver are quite conventional in design. Practically all the special features needed can be incorporated in the keyer and converter, so that most ordinary amateur equipment is suitable for RTTY with little or no modification.

Transmission Methods

It is quite possible to transmit teleprinter signals by ordinary "on-off" or "make-break" keying such as is used in regular hand-keyed cw transmission. In practice, however, frequency-shift keying is preferred because it gives definite pulses on both mark and space, which is an advantage in printer operation. Also, since fsk can be received by methods similar to those used for fm reception, there is considerable discrimination against noise, both natural and manmade, distributed uniformly across the receiver's passband, when the received signal is not too weak. Both factors make for increased reliability in printer operation.

Frequency-Shift Keying

On the vhf bands where A2 transmission is permitted, audio frequency-shift keying (afsk) is generally used. In this case the rf carrier is transmitted continuously, the pulses being transmitted by frequency-shifted tone modulation. The audio frequencies used have been more-or-less standardized at 2125 and 2975 Hz, the shift being 850 Hz. (These frequencies are the 5th and 7th harmonics, respectively, of 425 Hz, which is half the shift frequency, and thus are convenient for calibration and alignment purposes.) With afsk, the lower audio frequency is customarily used for mark and the higher for space.

Below 50 MHz, F1 or fsk emission must be used. The carrier is on continuously, but its frequency is shifted to represent marks and spaces. General practice with fsk is to use a frequency shift of 850 Hz, although FCC regulations permit the use of any value of frequency shift up to 900 Hz. The smaller values of shift have been shown to have a signal-to-noise-ratio advantage, and 170-Hz shift is currently being used by a number of amateurs. The nominal transmitter frequency is the mark condition and the frequency is shifted 850 Hz (or whatever shift may have been chosen) lower for the space signal.

RTTY with SSB Transmitters

A number of amateurs operating RTTY in the hf bands, below 30 MHz, are using audio tones fed into the microphone input of an ssb transmitter. With properly designed and constructed equipment which is correctly adjusted, this provides a satisfactory method of obtaining F1 emission. The user should make certain, however, that audio distortion, carrier, and unwanted sidebands are not present to the degree of causing interference in receiving equipment of good engineering design. The user should also make certain that *the equipment is capable of withstanding the higher-than-normal average power involved.* The RTTY signal is transmitted with a 100-percent duty cycle, i.e., the average-to-peak power ratio is 1, while ordinary speech waveforms generally have duty cycles in the order of 25 percent or less. Many ssb transmitters, such as those using sweep-tube final amplifiers, are designed only for low-duty-cycle use. Power-supply components, such as the plate-voltage transformer, may also be rated for light-duty use only. As a general rule when using ssb equipment for RTTY operation, the dc input power to the final PA stage should be no more than twice the plate dissipation rating of the PA tube or tubes.

FREQUENCY-SHIFT KEYERS

The keyboard contacts of the teletypewriter actuate a direct-current circuit that operates the printer magnets. In the "resting" condition the contacts are closed (mark). In operation the contacts open for "space." Because of the presence of dc voltage across the open keyboard contacts in such an arrangement, they cannot normally be used directly to frequency-shift-key another circuit. Isolation in the form of a keying relay or electronic switching is ordinarily used.

Saturated-Diode Keying

Perhaps the simplest satisfactory circuit for frequency-shift keying a VFO is the one shown in Fig. 1. This uses a diode to switch a capacitor in and out of the circuit, and is intended for use in a transmitter which heterodynes the VFO signal to the operating frequency. Because of the small number of parts required for the modification, they can often be mounted on a small homemade subchassis, which in turn is mounted alongside the VFO tube. Connection to the VFO circuit can be made by removing the tube from its socket,

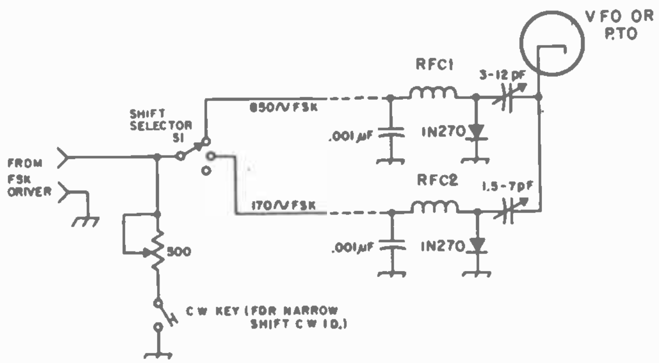


Fig. 1 — Frequency-shift keyer using saturated diodes

RFC1, RFC2 — 2.5 mH (National R-100 or equiv.).
S — Spdt rotary, toggle, or slide.

wrapping the connecting lead around the tube's cathode pin, and reinserting the tube in its socket. The variable capacitors are adjusted for the desired shifts. Once set, the shifts will remain constant for all bands of operation. With this circuit the VFO frequency will be lower on space when the fsk driver of the RTTY demodulator shown later is used. If VFO "sideband inversion" takes place in a mixer stage of the transmitter, it will be necessary to key from the afsk driver output of the demodulator to send a signal which is "right side up."

Be sure to use an NPO type miniature ceramic trimmer for best stability. Use only an rf choke wound on a ceramic form. Ferrite or iron-core types are not suitable because of excessive internal capacitance, so the National type R-100 is recommended. Use only the 1N270 diode specified. This diode is a special high-conductance computer type which provides maximum circuit Q, avoiding variations in oscillator output level.

"Shift-Pot" Keying Circuit

The circuit of Fig. 2 may be used with transmitters having a VFO followed by frequency-multiplying stages. The amount of frequency multiplication in such transmitters changes from one amateur band to another, and to maintain a constant transmitted frequency shift readjustment is necessary during band changes. In this circuit the natural VFO frequency is used for mark, and for space the frequency is lowered somewhat depending on the current flowing through CR1. R1 adjusts this current, and therefore controls the amount of frequency shift. As shown, the circuit may be keyed by the fsk driver stage of the RTTY demodulator shown later. If a keying relay is used, Q1 may be omitted and the keying contacts (closed on mark, open on space) connected directly from the junction of R1 and R2 to ground.

Leads inside the VFO compartment should be kept as short as possible. Lead length to the remainder of the circuit is not critical, but to avoid inducing rf or 60-Hz hum into the circuit, shielded wiring should be used for runs longer than a few inches. Positive voltages other than 150 may be used for the bias supply; the value and wattage of R3 should be chosen to supply a current of 2 mA or more to the 6.5-V Zener diode.

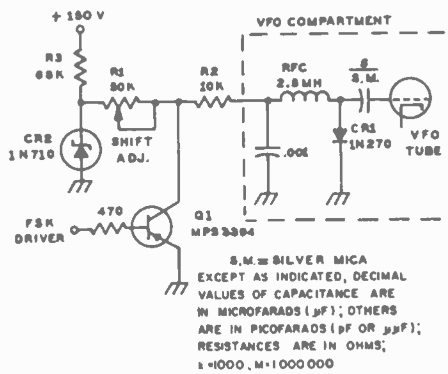


Fig. 2 — "Shift-pot" frequency-shift keyer circuit. The shift-adjustment control may be remoted from the VFO circuit.
CR2 — Zener, 6.5-V 400 mW (1N710 or equiv.).
R1 — Linear-taper control, low wattage.
Q1 — Audio transistor, npn silicon (Motorola MPS3394 or equiv.).

AN RTTY DEMODULATOR

Fig. 1 on page 462 shows the diagram of a solid-state demodulator which can be built for approximately $60. Using surplus 88-mH toroidal inductors,[1] the discriminator filters operate with audio tones of 2125 and 2295 Hz for copying 170-Hz shift, which is used almost exclusively on the amateur bands these days.

The demodulator is intended to be operated from a 500-ohm source. If only a 4- or 8-ohm speaker output is available at the receiver, a small line to voice-coil transformer should be used between the receiver and the demodulator to provide the proper impedance match. An integrated-circuit operational amplifier, having very high-gain capability, is used for the limiter. The discriminator filters and detectors convert the shifting audio tones into dc pulses which are amplified in the slicer section. The keyer transistor,

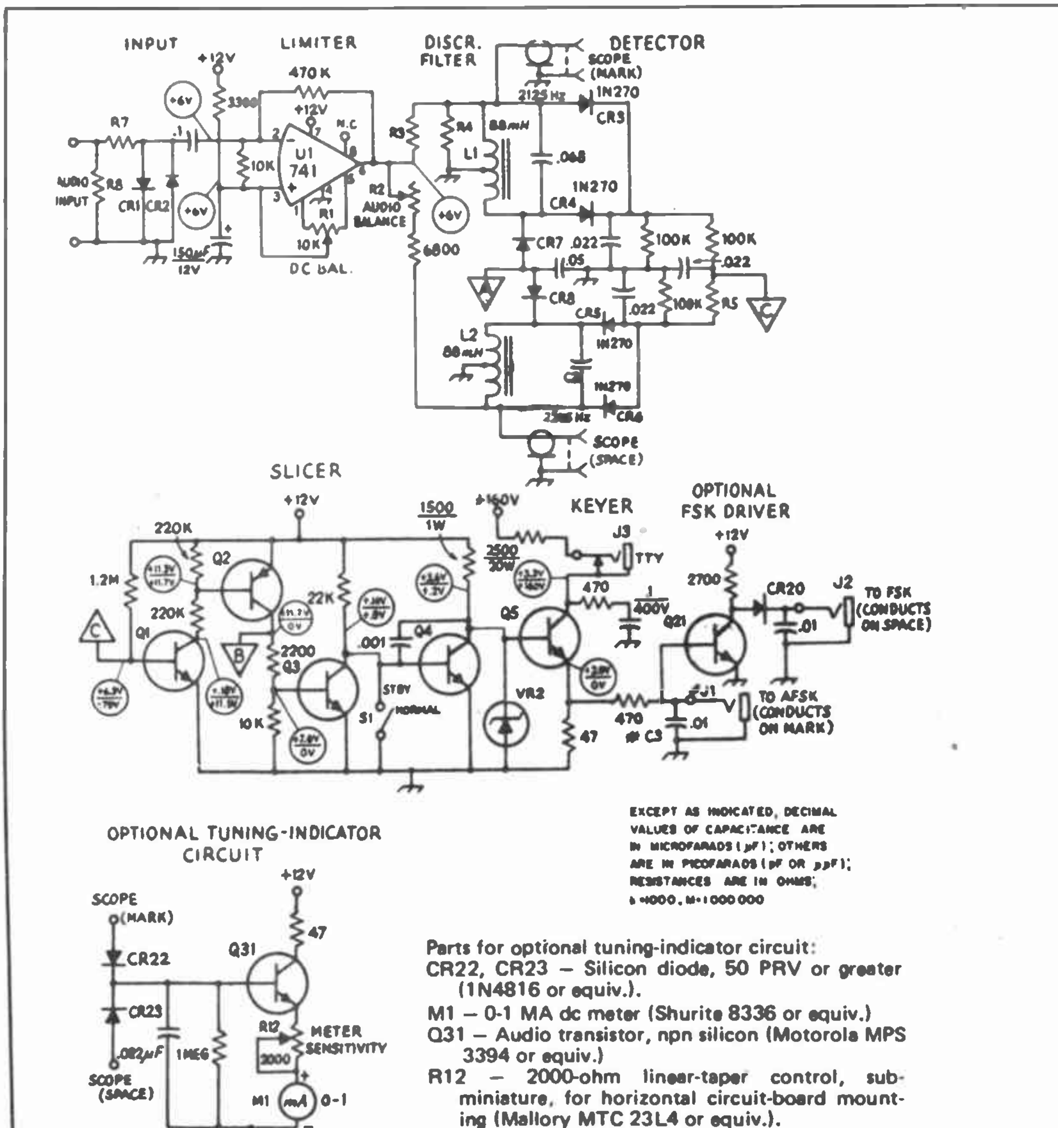
[1] See *QST* Ham-Ads for suppliers of 88-mH toroids.

Q5, controls the printer selector magnets, which should be wired for 60-mA operation. The teleprinter keyboard is to be connected in series with the printer magnets, both being connected to the demodulator via J3. Typing at the keyboard will then produce local copy on the printer and will also produce voltages at J1 and J2 for frequency-shift keying a transmitter or an audio oscillator.

The autoprint and motor-delay section provides optional features which are not necessary for basic operation. This section provides a simulated mark signal at the keyer when no RTTY signal is being received, preventing cw signals and random noise from printing "garble" at the printer. The motor-control circuit energizes the teleprinter motor in the presence of an RTTY signal, but turns off the motor should there be no RTTY signal present for approximately 30 seconds.

Adjustments

With a VTVM, measure the +12-V supply potential. Ground the audio input to the demodulator, and connect the VTVM to pin 3 of the IC. Adjust R1 through its total range, and note that the voltage changes from approximately 1.6 V at either extreme to about +6 V at the center setting of R1. Perform a coarse adjustment of R1 by setting it for a peak meter reading, approximately +6 V. Now move the VTVM lead to pin 6 of the IC. Slowly adjust R1 in either direction, and note that adjustment of just a *small fraction* of a



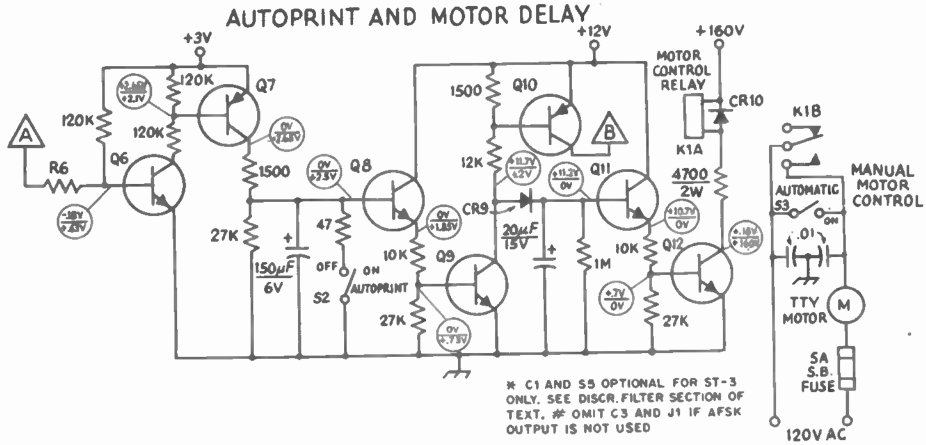
Parts for optional tuning-indicator circuit:

CR22, CR23 – Silicon diode, 50 PRV or greater (1N4816 or equiv.).

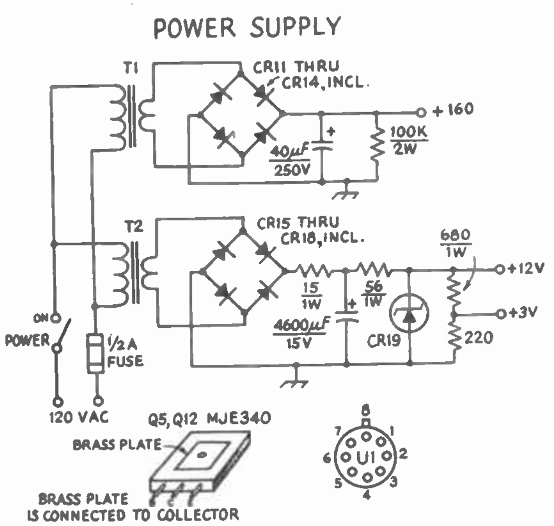
M1 – 0-1 MA dc meter (Shurite 8336 or equiv.)

Q31 – Audio transistor, npn silicon (Motorola MPS 3394 or equiv.)

R12 – 2000-ohm linear-taper control, subminiature, for horizontal circuit-board mounting (Mallory MTC 23L4 or equiv.).



EXCEPT AS INDICATED, DECIMAL VALUES OF CAPACITANCE ARE IN MICROFARADS (µF), OTHERS ARE IN PICOFARADS (pF OR µµF); RESISTANCES ARE IN OHMS; k=1000, M=1000 000



C1 — Optional
C2 — .033 µF, paper or Mylar, 75- or 100-volt rating.
C3 — .01 µF Mylar or disk, 600 volt. Omit if af keying output is not used.
CR1, CR2, CR7, CR8, CR9, CR15-CR18, incl., CR20 — Silicon diode, 50 PRV or greater (1N4816 or equiv.)
CR3-CR6, incl. — Germanium diode, type 1N270.
CR10-CR14, incl. — Silicon rectifier, 400 PRV or greater (1N4004 or equiv.).
CR19 — Zener diode, 12-V, 1-W (Sarkes-Tarzian VR-12 or equiv.).
CR21 — Zener diode, 4.3-V, 400-mW (1N4731 or equiv.).
J1, J2 — Phone jacks. Omit J1 if af keying output is not used.
J3 — Phone jack, single circuit, shorting.
K1 — 110-V dc relay, dpdt contacts with 10-A minimum rating (Potter and Brumfield type KA11DG or equiv.).
L1, L2 — 88 mH toroid.
Q1, Q6, Q8, Q9, Q11, Q21 — Audio transistor, npn silicon (Motorola MPS3394 or equiv.).
Q2, Q7, Q10 — Audio transistor, pnp silicon (Motorola MPS3702 or equiv.).
Q3, Q4 — General-purpose transistor, npn silicon (Motorola MPS2926 or equiv.).
Q5, Q12 — Audio transistor, npn silicon, 300-V collector-emitter rating (Motorola MJE340 or equiv.).
R1, R2 — 10,000-ohm linear taper control, subminiature, for horizontal circuit-board mounting (Mallory MTC-14L4 or equiv.).
R3 — 5600 ohms.
R4 — 18,000 ohms.
R5 — 82,000 ohms.
R6 — 0.1 megohm.
R7 — 1000 ohms.
R8 — 560 ohms.
S1-S5, incl. — Spst toggle. S5 optional.
T1 — Power; primary 120 V; secondary 125 V (Chicago-Stancor PA-8421 or Triad N51-X or equiv.).
T2 — Power; primary 120 V; secondary 12 V, 350 mA (Chicago-Stancor P8391 or equiv.).
U1 — Integrated-circuit operational amplifier, µA741, TO-5 package.

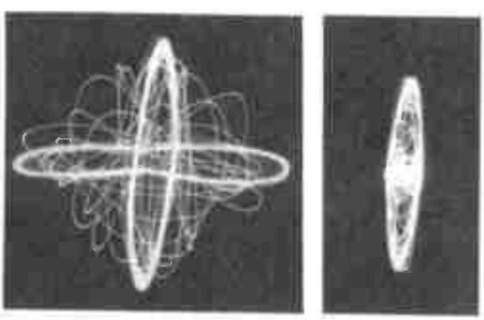
Fig. 1 — The ST-4 RTTY demodulator (by Hoff — from *QST*, April 1970). Unless otherwise indicated, resistors are 1/4-watt 10-percent tolerance. Capacitors with polarity indicated are electrolytic. Dc operating voltages are indicated in the limiter, slicer, keyer, and autoprint and motor delay circuits. All voltages are measured with respect to chassis ground with a VTVM. In the slicer and keyer stages, voltage values above the line should appear with a mark tone present at the demodulator input, while values below the line appear with a space tone present. In the autoprint and motor delay circuit, voltage values above the line occur with a mark or space tone present while those values below the line are present with only receiver noise applied at the demodulator input.

The RTTY demodulator may be constructed on a large circuit board which is mounted inside a standard aluminum chassis, as shown here. A decorative self-adhesive paper provides the grained-wood appearance. The meter is optional and provides a tuning indication for use in the hf amateur bands.

turn causes the voltage to swing from approximately +1 V to +11 V. Carefully perform a fine adjustment of R1 by setting it for a voltmeter reading of half the supply voltage, approximately +6 V. Next, again measure the voltage at pin 3. If the potential is approximately +6 V, R1 is properly set. If the potential is in the range of +2 V or less, R1 is misadjusted, and the procedure thus far should be repeated.

Next connect the VTVM to point A. With a mark-tone input, adjust the tone frequency for a maximum reading around −2.5 volts. Then change the tone for maximum reading on the space frequency. Adjust R2 until the voltages are equal.

With afsk at vhf, audio tones modulating the carrier are fed from the receiver to the RTTY demodulator. At hf, the BFO must be energized and the signal tuned as if it were a lower sideband signal for the proper pitches. If the tuning-indicator meter is used, the hf signal should be tuned for an unflickering indication. A VTVM connected at point A of Fig. 1 will give the same type of indication. An oscilloscope may be connected to the points indicated in the filter section and used for a tuning indicator, as shown in the accompanying photographs.



Oscilloscope presentations of the type obtained with the scope mark and scope space connections in the filter section are made. For these displays the mark frequency is displayed on the horizontal axis and the space frequency on the vertical axis. The signals appear as ellipses because some of the mark signal appears in the space channel and vice versa. Although only one frequency is present at a given instant, the persistence of the scope screen permits simultaneous observation of both frequencies. The photo at the left shows a received signal during normal reception, while the photo at the right shows a signal during unusual conditions of selective fading, where the mark frequency is momentarily absent.

AMATEUR TELEVISION (ATV)

Television is not exactly new to amateur radio. Enterprising amateurs have been playing with this branch of the electronics art for a matter of 45 years or more. Files of *QST* dating back to the '20s offer proof that there was amateur television before many of our present-day amateurs were born. The methods used then bore little resemblance to the techniques employed today, but hams were sending and receiving pictures (or trying to) two generations ago.

QST carried many articles on television from 1925 on, and there was plenty of interest. But the work was being done by the motor-driven scanning disk method, and it was doomed to failure. Though many dollars and man-hours were spent on the problem, nobody succeeded in developing mechanical systems that were completely practical. As early as 1928, a *QST* author was pointing out the possibilities of electronic television, using the then rare-and-expensive cathode-ray tube. The days of the scanning disk were numbered.

But predicting the coming of electronic television and bringing it about were two quite different matters. Though it had become fashionable, by 1931, to say that "Television is just around the corner," the cathode-ray tube was a laboratory curiosity, and it was to remain so for some years to come, as far as most amateurs were concerned. Not until 1937 was the subject of

An actual 440-MHz TV picture transmitted with the equipment shown in Fig. 15B-2.

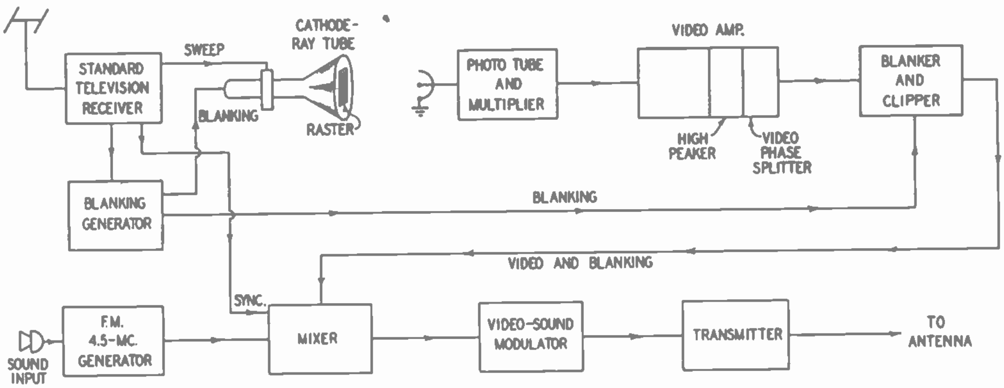


Fig. 15B-1 – Block diagram of the television system used by W2BK, formerly W2LNP. (From *QST*, June, 1950.)

television to appear again in *QST*. By then the problems involved in electronic television were gradually being solved. Usable components were beginning to appear, and television experimental work loomed as a possible field for the more advanced amateur. For more than two years almost every issue of *QST* carried something on television, but it was mostly concerned with the receiving end. The generation of a television picture for transmission was still considered to be beyond the radio amateur, until moderately priced iconoscope tubes were introduced for amateur use in 1940. Television transmitter and camera design were treated extensively in *QST* for 1940.

The highly involved and expensive process required in getting on the air for actual television communication was just too much for most amateurs, and progress in amateur television slowed to a standstill until well into the postwar period. At that time, availability of most of the needed components on the surplus market gave amateur television the push that it had always needed, and the period since 1948 has seen more amateur TV activity than existed in all previous years combined. By 1960, color-TV signals were being transmitted by amateurs.

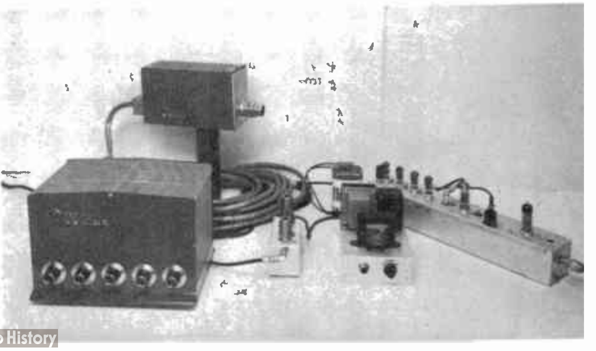
From several cities in this country has come news of activity in amateur television. Much of the effort has been concerned with transmitting. The trend in this country has been to use transmitting systems that would tie in with those employed in commercial services, so that ordinary home television receivers could be used for amateur work by the addition of a simple converter. In this country amateur TV is limited to the frequencies from 420 MHz up, because of the bandwidth involved.

A Novel Way to Get Started

The cost and complexity of TV gear has so far left most amateurs convinced that television is not for them, but ways have been found to cut corners. There have been several ideas developed for bringing the transmission of television nearer to the abilities of the average experienced ham. One such simplified system was developed by J. R. Popkin-Clurman, W2LNP, later to become W2BK. This system simplifies matters for the ham who would like to transmit transparencies (film negatives or positives, movies, diagrams, visual messages) without going into the complexities of camera design and construction. It also lets a local TV station and a standard TV receiver do some of the work, as shown in block-diagram form in Fig. 15B-1. A standard TV receiver is tuned to a local station and the lead from the receiver video amplifier to the cathode-ray tube is disconnected and the output of the amplifier is fed to a blanking generator. The output of the blanking generator is applied to the receiver cathode-ray tube, the raster of which is used as a light source.

In the simplest form of picture transmission, a transparency is placed directly on the face of the cathode-ray tube, which for this purpose can be almost any type, including those with P-7 phosphor. Light from the raster, passing through the transparency, is picked up by a photo tube and multiplier and fed to a video-amplifier unit that includes a high-frequency peaker and possibly a video phase inverter. The latter is used only if it is desired to transmit negatives in positive form. After passing through a clipper and blanking inserter and a mixer, the signal is ready for the modulator and transmitter. Sound and video are

Fig. 15B-2 – The transmitting portion of a complete ATV station. The video system utilizes a modified RCA TV Eye closed-circuit camera and control unit, shown at the left. The 440-MHz TV signal comes out the BNC connector at the end of the mixer-amplifier chassis. The power supply and bias battery are also visible in the photograph. (From *QST*, November, 1962)



transmitted on the same channel, first by frequency modulating a 4.5-MHz oscillator. The 420-MHz transmitter is modulated simultaneously with this signal and the video, by means of the video-sound modulator.

The signal thus transmitted has all the characteristics of a commercial video transmission, and may be received on any standard home television receiver equipped with a 420-MHz converter. In the absence of a local TV station it is merely necessary to derive the sync and blanking from the receiver's own sweep circuits. In this case the picture will have only 262 lines, noninterlaced. It retains the same horizontal resolution, but the vertical resolution is reduced. In this type of operation it is desirable to sync the vertical to the 60-Hz power supply, to reduce hum effects.

The photo tube may be a 931-A multiplier type, available as surplus. The output of the photo tube is fed into a series of video amplifiers, one of which is a high-frequency peaker. This is necessary to compensate for the build-up and decay times of the cathode-ray tube's phosphor screen.

The rf section of the transmitter is crystal-controlled. The receiver has a crystal mixer and a 6J6 oscillator, followed by a cascode amplifier working into a home television receiver. The channel used for the i-f should be one that is not in use locally, and should be in the low-TV band for best results.

The system may be adapted for transmission of movies. A film-projector light source is removed, and the photo tube installed in its place. A 60-Hz synchronous motor is used to drive the film sprocket and the film is run at 30 frames per second instead of 16 or 24. It is necessary to blank the raster during the film pull-down time. Pictures of live subjects may also be transmitted by projecting the light from the raster on the subject and collecting the reflected light with a condensing-lens system for the photo tube. Considerably greater light is needed than for transparencies, and a 5TP4 or a 5WP15 projection cathode-ray tube, with its associated high voltage, is suitable.

Adapting Closed-Circuit TV Systems

By adapting closed-circuit TV systems, a number of amateurs have been able to get a picture on the air without having to struggle with cut-and-try methods, not to mention the mechanical problems of camera construction. A manufactured TV camera and control unit are used, along with home-built rf sections necessary for the ATV station. Such a system is not restricted to sending slides or stills. It is capable of transmitting a moving picture of professional quality. Such a station is shown in Fig. 15B-2.

Many closed-circuit TV cameras provide a picture signal on any regular TV channel from 2 to 6, inclusive. In a typical system, the camera contains a vidicon camera tube, a three-stage video amplifier, a video output stage, a 55- to 85-MHz tunable oscillator, and a modulator stage that combines the rf, video, and sync signals from a control unit. The control unit contains the horizontal and vertical deflection circuits for the vidicon tube, a protective circuit that prevents damage to the vidicon in the event of a sweep-circuit failure, a blanking and vertical sync stage, and the power supply. For use in an ATV station, most amateurs choose to modify the camera oscillator circuit to provide crystal-controlled operation on a locally unused low TV channel. In this way, a regular TV receiver can be used as a monitor. The video-modulated rf signal from the camera is fed through amplifier and mixer stages to derive the transmitted video signal. For reception, a converter is used ahead of a regular TV receiver.

SLOW-SCAN TELEVISION (SSTV)

Because of the required bandwidth, amateur TV transmissions in this country are limited to the frequencies above 420 MHz. With essentially line-of-sight propagation of signals at these frequencies, it has always been necessary for an amateur wishing to engage in ATV to interest another local amateur in this mode, or for him to work into a local group which may already be active if he did not wish to transmit pictures merely for his own amusement. For this reason, ATV has had little to offer to the amateur who lives in a sparsely populated area, perhaps hundreds of miles from any large city. Slow-scan TV, on the other hand, offers a great deal. By using voice-channel bandwidths, SSTV transmissions may be used in any amateur band except 160 meters. The amateur in the sparsely populated area can exchange pictures with the fellows in the big city, the next state, or even with fellows in other countries.

Fig. 51C-1 — A typical slow-scan TV picture.

Work in the area of SSTV was pioneered by a group of amateurs headed by Copthorne Macdonald, W4ZII (later to become in succession, WA2BCW, WAØNLQ, WA2FLJ, and W1GNQ). The first of Macdonald's several articles on the subject appeared in *QST* in 1958. Early on-the-air tests took place in the then-available 11-meter shared band, the only hf amateur band where "facsimile" transmissions were permitted. The video information was transmitted as amplitude modulation of a 2000-Hz subcarrier tone, which in turn was fed into the speech-amplifier circuits of a conventional transmitter.

The loss to U.S. amateurs of the 27-MHz band in September 1958 did much to dampen the enthusiasm of would-be slow scanners. However, special temporary authorizations were granted by the FCC to a few amateurs for the purpose of making experimental SSTV transmissions, first on 10 meters, and later on 20 meters. Tests by WA2BCW and others in 1959 and 1960 indicated that signal fading and interfering transmissions from other stations caused considerable degrading of pictures received from subcarrier a-m (scam) transmissions. This led to experiments with subcarrier fm (scfm) transmissions, and the superiority of this technique for average propagation conditions was immediately recognized. The resulting standards proposed by Macdonald in January 1961 have since been adapted and are in use today (see Table I). In the scfm system, the frequency of the audio tone conveys the video information, with 1500 Hz representing black and 2300 Hz representing white. Intermediate shades of gray are transmitted with intermediate-frequency tones. Tones of 1200 Hz (ultrablack) are used to transmit vertical and horizontal sync pulses. The success of experiments in the mid '60s on 20 meters with scfm, and especially the fact that SSTV occupies a normal voice-channel bandwidth with no side-frequency products to cause interference on adjacent channels, led to changes in the FCC rules.

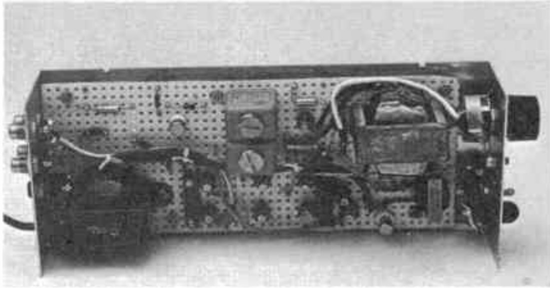
SSTV Emissions

Since August 1968, narrow-band A5 and F5 emissions (SSTV) have been permitted in the Advanced and Extra Class portions of 75, 40, 20 and 15 meters, in all but the cw-only portions of 10, 6, and 2 meters, and the entire amateur range above 220 MHz. The regulations permit the transmission of independent sidebands, with picture information contained in one sideband and voice in the other. Few amateurs today are equipped for this type of operation, however. The usual practice is to intersperse picture transmissions with voice transmissions on single sideband.

A stipulation in the U.S. regulations limits the bandwidth of A5 or F5 emissions below 50 MHz; they must not exceed that of an A3 single-sideband emission, approximately 3000 Hz. This precludes the use of an a-m transmitter with the standard SSTV subcarrier tones. Most amateurs operating in the hf bands feed the video information as a varying-frequency tone into the microphone input of an ssb transmitter, and with carrier suppression, F5 emission results. A seldom-used but quite feasible alternative is to frequency modulate an rf oscillator with video signals from the camera.

Because of the narrow bandwidth used, tape recordings of SSTV video signals can be made with an ordinary *audio* tape recorder running at 3 3/4 inches per second. Nearly every slow scanner preserves some of his on-the-air contacts on tape, and most prepare an interesting program to be transmitted. A good number of amateurs begin making two-way picture transmissions while equipped with nothing more than a receiving monitor and a tape recorder, in addition to ordinary station equipment. In lieu of a camera, they enlist the aid of a friend having the proper equipment to prepare a taped program which is sent during transmissions. Because of the slow frame rate with SSTV (one picture every 7 or 8 seconds), live pictures of anything except still subjects are impractical. Viewing a series of SSTV frames has frequently been compared to viewing a series of projected photographic slides.

Experiments are currently being made with the transmission of color pictures by SSTV. Various techniques are being used, but in essence the process involves the sending of three separate frames of the same picture, with a red, a blue, and



The SSTV Viewing Adapter with the top cover removed. The adapter may be constructed on Vectorbord, as shown. The transformer near the rear (left) is in the power supply circuit; the one near the front is in the video detector stage. On the front panel are the power switch and indicator, the manual vertical-sweep push button, and vertical sync control. Phono jacks on the rear panel are for connections to the oscilloscope and receiver. Two banana jacks are used for the CRT connections. (Originally described in *QST* for June, 1970, by W7ABW and W7FEN.)

TABLE I

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 sec.	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 millisec.	5 millisec.
Vertical	30 millisec.	30 millisec.
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

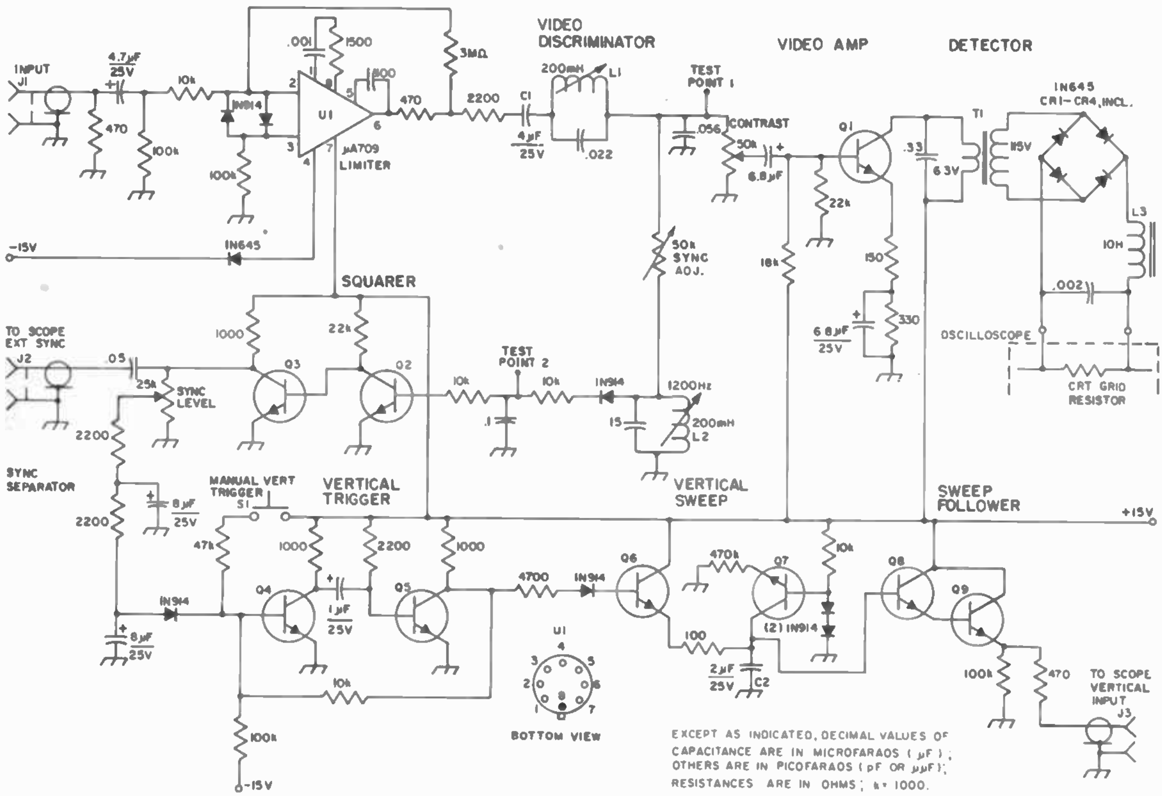


Fig. 1 — Schematic diagram of the slow-scan adapter. Capacitors with polarity indicated are electrolytic, others are ceramic or paper, except as indicated. Variable resistors are composition controls, linear taper. Resistors are 1/2-watt.

C1 — 4-µF, 25-volt, nonpolarized tantalum.
C2 — 2-µF, 25-volt, Mylar.
J1-J3, incl. — Phono jack.
L1, L2 — Variable inductor, approx. 200 mH (Miller 6330, UTC HVC-6, or Stancor WC-14).
L3 — 10-H, low-current choke, 3000-volt insulation from ground (Burstein-Applebee 18A959).
Q1-Q9, incl. — 2N718, 2N697, 2N2222, or 2N3641-3.
T1 — 6.3-volt, low current, 3000-volt insulation.
U1 — Operational amplifier (Fairchild µA709, Texas Instruments SN6715 or Motorola SC4070G).

a green filter successively placed in front of the camera lens for each of the three frames. At the receiving end of the circuit, corresponding filters are used and each frame is photographed on color film. After a tricolor exposure is made, the photograph is developed and printed in the normal manner. The use of Polaroid camera equipment with color film is popular in this work because it affords on-the-spot processing. Color reproduction by this technique can be quite good.

SLOW-SCAN TV VIEWING ADAPTER FOR OSCILLOSCOPES

The slow-scan TV adapter shown in Figs. 1-4, incl., permits the ham with an oscilloscope to view slow-scan TV with a minimum of investment and effort. The adapter has been used successfully with several oscilloscopes, including the Tektronix 514, Dumont 304, Heathkit IO-18, Heathkit IO-10, and a Navy surplus scope, OS-8B.

The oscilloscope's horizontal scan must be able to synchronize from an external trigger at 15 Hz. The scope should have a dc vertical input that will accept 10 volts. If the scope does not have a dc input, the vertical deflection amplifier may be able to be driven directly. The circuit shown in Fig. 3 was used with the Heath IO-18. This arrangement should be adaptable to other scopes not having a dc input, but R1 and R2 would have to be scaled to provide proper centering.

Most oscilloscopes have cathode-ray tubes with a P1 phosphor. The P1 phosphor is of short persistence, which is not suitable for slow-scan TV. Therefore, the P1 tube should be replaced with a P7-phosphor tube which has the long persistence required. The last two characters of the CRT type usually indicate the phosphor, and most types are available in several different phosphors. The Heath IO-18 uses a 5UP1 which was replaced with a 5UP7 at a cost of less than $15.00.[1] If a direct substitute cannot be found, it may be possible to find a

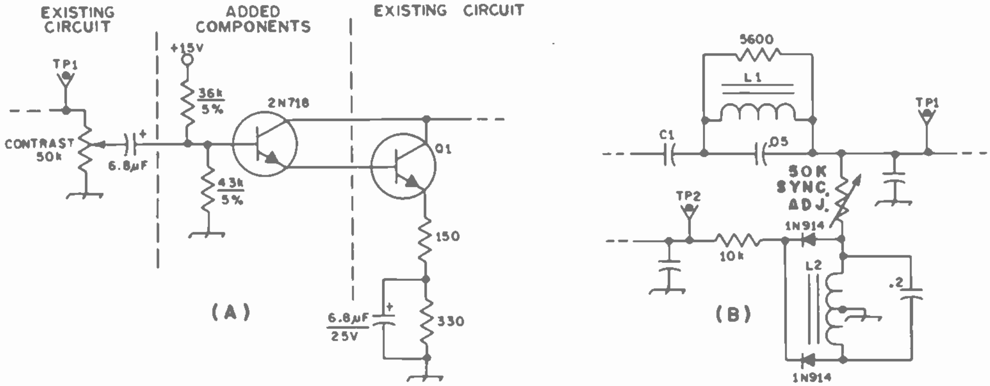


Fig. 2 — At A, a circuit which may be added to increase the contrast of the SSTV adapter, and at B, an alternative circuit using surplus 88-mH toroidal inductors for L1 and L2. If the circuit of A is used, the 18,000- and 22,000-ohm resistors shown connected to the base of Q1 in Fig. 1 are unnecessary.

surplus CRT of another type which will function. The Dumont 304 used a 5ABP1 CRT, which was replaced with a 5CP7. This CRT was obtained on the surplus market for less than $5.00.[2] If the purchase of a new oscilloscope is anticipated, a P7-phosphor cathode-ray tube should be requested.

Adapter Circuit Design

The schematic diagram of the slow-scan TV converter is shown in Fig. 1. The slow-scan signal from the audio output of a communications receiver, tape recorder, or other source is fed into the input of an integrated-circuit operational amplifier having a gain of 300. Therefore, a 0.1-volt ac peak-to-peak signal causes the amplifier to limit at the supply voltages, and the limited output will be approximately 28 volts ac peak-to-peak. The limited signal is then fed to a series video discriminator. The output of the video discriminator is fed to Q1, a video amplifier with a 6.3-volt ac filament transformer as a collector load. The transformer is used to provide voltage step-up. A transformer with 3000-volt insulation from ground is used, as the CRT grid circuit has a 1400-volt potential which must be insulated from ground. The video is then full-wave rectified and fed to a 1000-Hz filter. The output video dc is then connected across the scope CRT's series grid resistor to modulate the CRT intensity.

The output of the video discriminator is also fed to a 1200-Hz sync discriminator. This circuit passes only the 1200-Hz sync pulses. The 1200-Hz sync pulses are then rectified, filtered and fed to a two-stage amplifier, Q2 and Q3. The output of this squarer provides 15-volt sync pulses.

A 5-volt sawtooth voltage is required for vertical sweep on the oscilloscope. This voltage should have a very fast rise time and a linear decay. A sync separator circuit is used to separate the 30-ms vertical pulses from the 5-ms horizontal pulses. The vertical pulses are fed into the vertical trigger, a one-shot multivibrator. Provision is made for manually triggering the vertical sweep with a front-panel push button, S1, in case a vertical sync pulse is missed. The multivibrator triggers a transistor switch, Q6, that instantaneously charges C2 every time a vertical sync pulse is received. This capacitor is discharged at a linear rate through Q7. The base of Q7 is biased by two diodes at 1.2 volts. Thus, the current through the 0.47-megohm emitter resistor is held at a constant value, giving a linear voltage discharge across C2. This sawtooth voltage is sampled by a Darlington transistor follower, Q8 and Q9, whose output will sweep from 10 to 5 volts dc when receiving slow-scan TV. The value of 5 volts was chosen so that when a signal is not present, the dot on the scope CRT will be off the screen.

If the capability for high contrast is desired, the video signal level may be increased by adding a 2N718 transistor ahead of Q1, as shown in Fig. 2A. For those who wish to use 88-mH toroids in place of the variable inductors, L1 and L2, the circuit of Fig. 2B may be used.

Construction

The layout is relatively noncritical with the exception of the 6-volt ac filament transformer which will have high voltage on the secondary, so necessary precautions must be taken. It should be mounted away from the power transformer to minimize hum pickup. High-voltage wire is used to bring the CRT grid connection into the unit. Sockets were used for the IC amplifier and transistors; however, the components can be soldered directly into the circuit. The vertical-scan output lead should be shielded. Several types of transistors may be used; the circuit was designed for devices with a minimum beta of 50. A variety of integrated operational amplifiers may be used; however, the 709 was chosen because of its low cost and availability.

Scope Modification

The potential between the CRT's control grid and the cathode varies the intensity. The control

[1] Available from Barry Electronics, 512 Broadway, New York, NY 10012.
[2] Catalog SC2799P7, Fair Radio Sales, P. O. Box 1105, Lima, OH 45802.

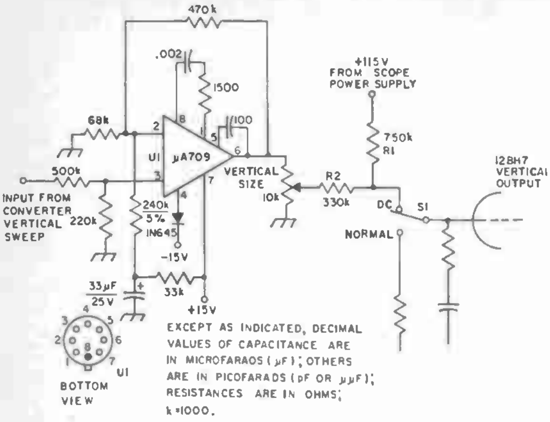


Fig. 3 — Amplifier circuit to provide a dc vertical input for ac-only oscilloscopes. Capacitors are ceramic, and resistors are 1/2-watt. The switch, S1, may be any convenient type. The operational amplifier, U1, is a Fairchild μA709. R1 and R2 should be adjusted in value to give proper centering, if necessary.

grid usually has an isolation resistor in series with the negative voltage lead. Video from the converter is connected across this resistor to vary the intensity of the CRT. This resistor should be at least 100 kΩ. If it is not this large in the existing scope circuit, it should be changed. This will have no effect on the scope's operation, since this control grid draws no current. There is usually ample room on most scopes to install two additional insulated jacks on the terminal board that has the direct deflection-plate connections.

Adjustment

1) Connect the scope's vertical input to test point 1.

2) Connect a 2350-Hz signal to the input and adjust the video discriminator coil L1 for minimum indication on the scope. This is usually with the slug fully inserted.

3) Connect the scope to test point 2. Change the input to 1200 Hz and peak the sync discriminator coil L2 for maximum indication on the scope. Connect a dc voltmeter between the collector of Q3 (sync level) and ground. With a 1300-Hz tone fed to the input of the adapter, adjust the 50,000 ohm sync adjust control to the point where the dc voltmeter just reads +15 volts.

4) Make the connections from the adpater to the oscilloscope's external sync, vertical input, and the CRT grid.

5) Connect the adapter's input to the receiver or tape recorder.

6) Set the contrast control at midposition and the sync control to maximum.

7) Adjust the scope's sweep to 15 Hz for trigger lock.

8) Adjust the size of the raster with the scope horizontal and vertical size controls until a square raster is obtained.

9) Adjust the adapter contrast and the scope intensity controls until a clear picture is obtained. If the picture is negative, the connections to the CRT grid should be reversed.

10) When a picture is obtained, the sync level should be adjusted to a point just before sync is lost. This will eliminate false triggering when copying weak signals and, if a vertical sync pulse is missed, the manual trigger can be used.

The finished adapter can be finally tested in several ways:

1) Tune to one of the SSTV frequencies listed below and look for a station transmitting SSTV. Tune the signal as you normally would for ssb. It is a good idea to tape-record a few pictures off the air — they then can be played back as often as necessary while adjusting the adapter.

2) Send a blank recording tape (with return postage) to any amateur who is equipped with an SSTV flying-spot scanner or camera. All amateurs in this field are happy to make a tape to get a newcomer going.

3) Listen to the SSTV frequencies. You may find a nearby amateur is on the air with SSTV. You can take your adapter to his shack to try it directly on a picture generator.

The slow-scan TV adapter has given good pictures on the scopes tried. A hood should be provided around the CRT face for direct viewing. Scopes with CRT tubes that have an accelerator will provide a brighter scan. The Heath IO-18 scope uses a CRT without the accelerator, and the brightness was noticeably less than others tried.

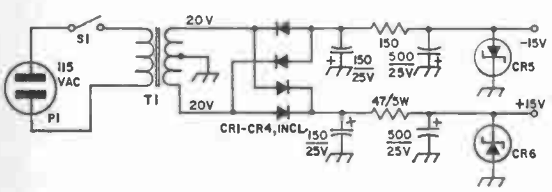


Fig. 4 — Power supply for the adapter. Capacitors are electrolytic. Resistors are 1/2-watt unless otherwise specified.
CR1-CR4, incl. — Silicon type, 200 PRV or greater (Motorola 1N4002, 1N4004 or 1N4007).
CR5, CR6 — 15-volt, 1-watt Zener (Centralab R4128-4, Unitriode Uz715).
P1 — Fused line plug.
S1 — Toggle.
T1 — 40-volt ct, 100 mA (Triad F90X).

At the present time, most SSTV operation takes place on 20 meters, on or above 14,230 kHz. Local nets operate on 3845 kHz. Other hf calling and working frequencies are 7171, 21,340, and 28,680 kHz. (In the U.S., SSTV emissions are authorized in the Advanced and Extra Class portions of all hf phone bands.) Stations from all continents are to be found on SSTV. The DX capability of SSTV is being demonstrated daily by picture exchanges between the U.S. and Canada and foreign amateurs.

SPACE COMMUNICATIONS

Moonbounce may be beyond the capabilities of some amateurs, but other forms of space communications are not. The amateur satellite program puts the excitement of vhf DX and space communications within the reach of every amateur.

It all began in 1961 with the successful launch of OSCAR 1, the world's first nongovernmental satellite. OSCAR (for Orbiting Satellite Carrying Amateur Radio) rode "piggyback" on a regular launch, with the ham satellite taking the place of ballast. The tiny satellite, built for out-of-pocket expenses of $64, transmitted HI on two meters for three weeks before its internal batteries were exhausted. The nearly identical OSCAR 2 transmitted for 18 days after its launch in June, 1962.

The first active communications satellite in the OSCAR series was OSCAR 3, launched in March, 1965. More than 100 stations in 16 countries helped make satellite history with OSCAR 3, the first free-access satellite. OSCAR 4 followed later that year, and achieved another communications first in spite of a bad orbit – the first U.S.-to-U.S.S.R. satellite contact. OSCAR 5 was constructed by a team at a university in Melbourne, Australia, and was launched in 1970. Its internal batteries provided two-way communications for more than six weeks.

AMSAT-OSCAR 6 was the first extended-lifetime amateur satellite, using solar cells to generate power. Launched in 1972 with a design lifetime of one year, OSCAR 6 was still going strong five years later. The professional satellite whose launch OSCAR 6 shared had long since died. It seems that amateurs *can* do a better job!

AMSAT-OSCAR 7 joined the series late in 1974, providing another major step forward for the

TABLE 1
AN OSCAR GLOSSARY

AMSAT – The Radio Amateur Satellite Corporation, a nonprofit organization located in Washington, DC; has overseen the development of the last three OSCAR satellites.

AMSAT-OSCAR-6 – The longest-lived amateur satellite, launched Oct. 15, 1972; has been used by more than 3,000 amateur stations in more than 100 nations.

AMSAT-OSCAR-7 – The latest and most sophisticated OSCAR; launched Nov. 15, 1974.

AOS (see also LOS) – Acquisition of signal – the time you can first hear the satellites, usually just after they rise over the horizon.

Ascending Node – The point where the satellites cross the equator traveling north.

Codestore – The special system that allows a Morse code message to be placed in a memory storage unit and automatically broadcast with telemetry data.

Decending Node – The point where the satellites cross the equator traveling south.

Doppler Effect – An apparent shift in frequency caused by the satellite's movement toward or away from your location.

Downlink – The frequency at which radio signals are received from the satellite.

ERP – Effective radiated power – transmitter output after transmission line losses, multiplied by antenna gain.

Inclination – The angle at which the satellite crosses the equator.

Increment – The number of degrees longitude the satellite appears to move westward at the equator with each orbit. Both satellites have an increment of approximately 28.7 degrees.

LOS (see also AOS) – Loss of signal – the time the satellite's signals pass out of range.

Mode A – The 2- to 10-m transponder aboard OSCAR 7.

Mode B – The 70-cm to 2-m transponder aboard OSCAR 7.

OSCAR – Orbiting Satellite Carrying Amateur Radio. There have been seven in the series and more are being built.

OSCAR Education Program – A special program that brings live demonstrations of the satellite to classrooms. Teachers use ARRL curriculum materials to apply the principles to a wide variety of subject areas.

Pass – An orbit of the satellite.

Passband – The range of frequencies handled by the satellite's transponder.

Period – The time it takes for a complete orbit. Periods of both satellites are approximately 115 minutes.

QRP Tests – Special orbits set aside for operating through the satellites while using a maximum of 10-watts erp.

Sun-Synchronous – A type of orbit that approximates the sun's apparent movement. Because their orbits are approximately sun-synchronous, OSCARs 6 and 7 can be heard at one location at about the same times each day.

Telemetry Beacon – The transmitters aboard each satellite that enable ground stations to monitor the satellites' vital functions.

Transponder – The repeater aboard the satellites (OSCAR 7 has two) that retransmits signals it receives on another frequency.

Uplink – The frequency at which radio signals are transmitted up to the satellites.

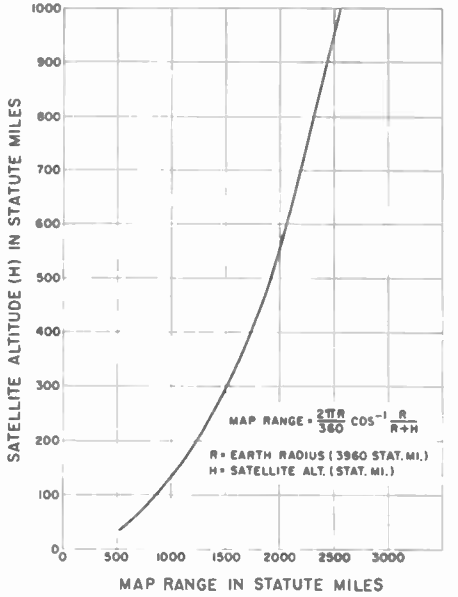


Fig. 15E-1 — Satellite altitude above earth versus ground station map range (statute miles).

amateur satellite program: It is on continuously, alternating between two different modes of operation. A cooperative international effort, OSCAR 7 was constructed in module form by amateurs in West Germany, Canada and Australia as well as in the U.S. More than 5,000 amateurs from over 100 countries have made two-way satellite contacts through OSCARs 6 and 7. And OSCAR 8 is scheduled to join the two functioning amateur satellites late in 1977 or early 1978.

More than an exotic means of communication for hams, the amateur satellites have been involved in a host of unique experiments. In the latter category, OSCARs 6 and 7 have located a hidden emergency locator transmitter similar to that carried aboard downed aircraft. Complex medical data such as electrocardiograms have been relayed from coast to coast through OSCAR, simulating the relay of such information out of a disaster-stricken area. A special authorization has permitted the use of 8-level ASCII code through the satellites, allowing amateurs to test direct computer-to-computer links through OSCAR. Finally, twice each year OSCAR 6 and OSCAR 7, in nearly identical orbits, pass close enough to each other to permit signals relayed through OSCAR 7 to be further relayed through OSCAR 6. The first such link experiment was another milestone for free-access satellites.

Communications firsts are not the only reason NASA cooperates with the OSCAR program by putting the amateur satellites in orbit. NASA has a strong interest in the OSCAR Education Program, designed to bring satellite and space technology into high school classrooms throughout the world.

Using readily available, commercial equipment, students are experiencing first-hand the unique aspects of space communications: Doppler shift, tracking, Faraday rotation, telemetry decoding, and much more.

Although each satellite has individual characteristics which distinguish it from its brethren, there are several aspects of communication which are common to all satellites.

RANGE

The determining factor in the maximum theoretical range of satellite communications is the height of the satellite. Fig. 15E-1 can be used to determine this range for the low orbits characteristic of early amateur satellites. To determine when you can hear the satellite, draw a circle with a radius equal to the map range from Fig. 15E-1. For OSCAR 6 and 7, this is 2450 miles (4000 km) for the 910-mile-high (1500 km) orbit of these two satellites. When the satellite passes over this circle, you should be able to hear it.

The maximum range of two-way communication through the satellite is illustrated in Fig. 15E-2. The more the range circles of the two stations overlap, the longer the time these two stations can remain in contact. For the farther DX stations, near the theoretical maximum limit of communications through OSCAR 6 and 7, this time is a matter of a minute or two, but for

Fig. 15E-2 — Satellite passes through the range of two stations, enabling contact.

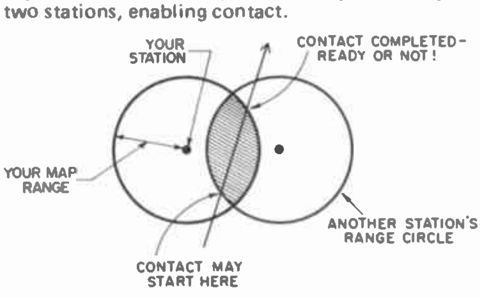


Fig. 15E-3 — Satellite transmitter frequency versus Doppler shift for satellite in 200- or 1000-statute-mile orbits. For a translator, use the difference between uplink and downlink frequencies as the "frequency."

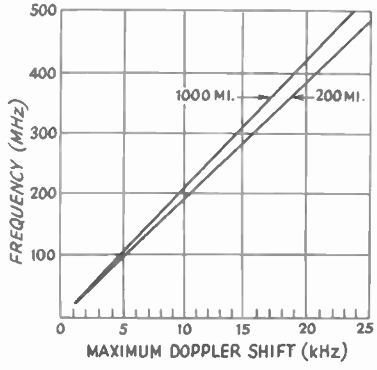


TABLE 2

SPACECRAFT FREQUENCIES

SPACECRAFT	*UPLINK*	*DOWNLINK*	*BEACON*
A-O-7			
Mode A	145.850 – 145.950 MHz	29.400 – 29.500 MHz	29.502 MHz
Mode B	432.125 – 432.175 MHz	145.975 – 145.925 MHz	145.972 MHz
A-O-8			
Mode A	145.850 – 145.950 MHz	29.400 – 29.500 MHz	29.402 MHz
Mode J	145.900 – 146.000 MHz	435.200 – 435.100 MHz	435.095 MHz
Phase III			
(2m/70cm)	145.850 – 145.990 MHz	435.150 – 435.290 MHz	435.145 MHz
(70cm/2m)	435.150 – 435.290 MHz	145.850 – 145.990 MHz	145.995 MHz

1) All date and time references are UTC.

2) OSCAR 7 operates in Mode A on odd days of the year, Mode B on even days. Wednesdays are reserved for experimental uses.

3) The Mode B and Mode J and both Phase III transponders turn upside down. A signal to the low end of the 70-cm input returns on the high end of the 2-meter output, and vice versa. Thus, an upper-sideband signal returns as lower sideband.

4) All orbits on Mondays (UTC) have been designated QRP orbits – use maximum of 10 watts of erp when transmitting.

stations separated by less than 4900 miles (7800 km), the time increases. Stations in the same town can communicate through the satellite for the entire 25 minute pass!

The length of time the satellite is within range is determined, as is the range, by the height of the orbit. The higher the orbit, the slower the satellite moves, and the longer the satellite is within range. The time it takes the satellite to make a complete revolution around the earth is called the period. Fig. 15E-3 gives the relationship between orbit height and period and speed. For OSCAR's 6 and 7, the period is about 115 minutes.

TRANSPONDERS

Present communications satellites use transponders to relay amateur communications. These transponders receive signals over a given segment of one amateur band and retransmit each signal over another segment in another band. Thus, OSCAR 6 receives signals between 145.9 and 146.0 MHz. These are re-transmitted between 29.45 and 29.55 MHz. The use of a transponder rather than a channelized repeater means that many more stations can be accommodated by the satellite at one time. In fact, the number of different stations using OSCAR simultaneously is limited only by mutual interference, and the fact that the output power of the satellite (in the order of a couple of watts) is divided among the users.

Each satellite transponder is also equipped with a telemetry beacon, which continuously transmits status reports on a variety of satellite parameters, such as internal temperature, current drains, power generation from the solar cells, and more. The telemetry information is used to determine the satellite operating schedule and to diagnose any operating difficulty. For example, early in 1976, OSCAR 6 began to have battery problems. (These problems were long overdue. Based on its one-year design lifetime, OSCAR 6 was the equivalent of a 300-year-old human!) The telemetry allowed ground command stations to shut the satellite off at regular intervals to prolong the useful life of the satellite.

The telemetry from OSCARs 6 and 7 is sent in the form of Morse code at about 20 wpm. The regular format of the telemetry makes copying the code easy for even the beginner. With the use of a decoding chart, anyone can compute the current generated by each solar panel, or any of the other 23 parameters. Looking at these figures over several frames of telemetry will allow you to determine the rate of spin of the satellite, and much more. The Radio Amateur Satellite Corporation (AMSAT), which oversees the construction of present amateur satellites, is always looking for telemetry reports from both birds. The first such report from each station is rewarded with a handsome QSL card. (Fig. 15E-4).

USING THE SATELLITES

To get started in satellite communications, you must know where the satellite is located. The simplest way to determine this is through the use of the ARRL OSCARLOCATOR, available for $1

Fig. 15E-4 – Telemetry report QSL card.

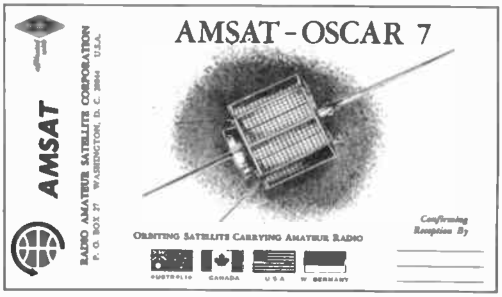


TABLE 3

AMSAT NETS AND BULLETINS SCHEDULES

The following AMSAT Nets meet regularly to disseminate information to newcomers and to keep regular satellite users in communication with one another.

USA-East Coast Net	Wednesdays	0100Z	3,850 kHz LSB	Net Control W3ZM or W3UN
USA-Mid States Net	Wednesdays	0200Z	3,850 kHz LSB	Net Control WØCY
USA-West Coast Net	Wednesdays	0300Z	3,850 kHz LSB	Net Control W6CG
International Net	Sundays	1800Z	14,280 kHz USB	Net Control W3ZM or W3UN

Bulletins of general interest to those interested in amateur satellites are transmitted regularly on OSCAR-7 reference orbits, at approximately 10 minutes after ascending node. These bulletins are transmitted on a downlink frequency of approximately 29,440 kHz or 145.960 MHz and can be received over most of eastern North America.

postpaid from ARRL hq. This device makes use of the fact that the orbit is very regular, and a single reference point anywhere along the orbit will allow you to predict the location of the satellite for as far ahead as several days.

The full details of the use of the locator are included with the package. Briefly, you need one reference point each day, which is usually the first time in a given UTC day that the satellite passes over the equator in a northerly direction. These data are reference-orbit EQX (equator crossing) data, and are available from a variety of sources, including W1AW bulletins, the AMSAT yearly orbit calendar, and the amateur publications, including *QST*. Armed with one reference point and the locator, you can determine the approximate location of, and bearing to the satellite anywhere in the Northern Hemisphere in seconds. Enterprising amateurs have programmed both programmable calculators and minicomputers to track the satellites. A sample in BASIC is available from ARRL hq. for a self addressed, stamped envelope. For those of you with Hewlett Packard HP-65 or HP-25 programmable calculators, the League will also supply the relevant programs and instructions on how to use them.

Receiving OSCAR

The need for telemetry-beacon information and constant monitoring of the OSCAR satellites has helped many amateurs get started. Most hams already have the needed equipment to receive the signals from the satellites, and they can get their feet wet in OSCAR work with little effort. Any hf receiver with adequate sensitivity in the upper part of 10 meters and any 10-meter antenna will work. What is adequate sensitivity? Compare the noise level in your receiver with the antenna connected, and with a dummy load or other 50-ohm nonreactive load. If the noise level is higher on the antenna, you have the needed sensitivity. If not, a small preamplifier for 10 meters will improve copy on the satellite signals (and terrestrial 10-meter work as well!) Hamtronics, 182 Belmont Road, Rochester, NY 14612, has an excellent, inexpensive preamp kit. Or see the receiving chapter in this book.

The ideal antenna for OSCAR work is a 10-meter turnstile. This consists of two dipoles at right angles, fed out of phase. The two antennas help to reduce the common deep fades caused by cross polarization as the satellite tumbles. However, if you are willing to put up with momentary fades (most satellite users are) any existing 10-meter antenna will work fine. The higher the gain, the more directional, and the greater difficulty you will have in aiming the antenna accurately. Also, you will experience signal loss as the satellite passes directly overhead. A typical tribander, vertical, or dipole will not have this problem, so all but the biggest 10-meter antennas are suitable for receiving OSCAR.

Transmitting

Once you have copied a few passes of the satellite, you will probably want to put your own signal through the bird. The main requirement is the ability to generate about 100 watts erp in the proper frequency range of 145.8-146.0 MHz. This 100 watts of effective radiated power (erp) can be produced in two ways: generation of about 100 watts of rf into an omnidirectional antenna, or less power into a gain antenna. The first option requires more power on 2 meters, but eliminates possibly confusing tracking problems. The second trades greater complexity in antennas and tracking for reduced cost of equipment.

Run both high power and a gain antenna? Negative, as the satellites are very sensitive to excessive power on the input. This reduces the sensitivity of the satellite receiver and increases the battery drain to dangerous levels. 100 watts of erp is plenty for satellite contacts; more is asking for trouble.

Antennas

Your choice of an antenna for accessing the satellites is related to the power you are generating on 2 meters. If you run more than 100 watts output, a simple turnstile antenna is sufficient. This is two dipoles at right angles to each other, fed 90° out of phase. The turnstile helps to reduce the effects of the changing polarizations due to Faraday rotation, satellite tumbling, and other factors.

If you run lower power and increase your erp with a gain antenna, you will need directional control. A simple solution is to aim a small (3-8 element) Yagi about 30° above the horizon. This will not seriously impair the terrestrial performance of the antenna, while greatly increasing its usefulness for satellite communications, because of the higher angle of radiation. You can even go to the ultimate in satellite antenna system: the az-el.

Az-el, or azimuth-elevation control, means controlling the antenna in both horizontal and vertical planes. This allows you to aim the antenna directly at the satellite, in a no-compromise approach. If you run only a few watts of rf into a high-gain antenna, you should consider az-el to increase your "window" of possible satellite communications time.

More details on this and many other aspects of satellite operation can be found in a reprint booklet of OSCAR articles, available from your local radio dealer or directly from ARRL. Included are descriptions of various antennas, equipment and operating procedure. More on operating and awards can be found in the OSCAR chapter of the *ARRL Ham Radio Operating Guide*, available from the same sources.

Operating

Operating through the satellites involves two unique aspects: full duplex and Doppler. In satellite communications, you can continuously monitor your signal through the satellite, as the signal from the satellite is on a different band that the one you are transmitting on. You can thus evaluate both the strength and quality of your own signal throughout the contacts. In fact, this full duplex is almost essential, because of the effects of Doppler.

Doppler shift is caused by the relative motions of you and the satellite. As the satellite is moving toward you, the frequency of the beacons and other signals is increased by a small amount. As the satellite passes overhead and begins to move away from you, you will notice a sudden drop in frequency of a few kilohertz. Just as the tone of a car horn or train whistle drops as the vehicle moves past you, so does the frequency of the satellite drop. This Doppler effect will be slightly different for stations located at different distances from the satellite, with the result that signals through the satellite move slowly around the frequency passband. Locating your own signal is more difficult than simply computing the relation between input and output frequency, as the rather hard-to-predict effects of Doppler must be taken into account.

The simplest way around this is the use of full duplex. By sending a series of dits through the satellite, and turning the keyer on and off, you can identify your signal from among the many other satellite users. Then you can determine the relationship between that frequency and the one on which you wish to transmit, either zero beat another station or calling CQ.

Say you find your own signal 12 kHz higher in the band than you wish to be. Simply lower your transmit frequency by that amount (with the transmitter off!) and you should be close. As you progress through the contact, you will notice the two frequencies changing. A hand on the transmitter tuning will help keep you on frequency. Thus it is extremely useful to be able to continuously monitor your signal from the satellite. No worry about inflated signal reports through OSCAR; you can give yourself an honest one!

Current Satellite Status

Early in 1977, two amateur satellites were operational: OSCAR 6 and OSCAR 7. Each contains the 2-to-10 transponder as described pre-

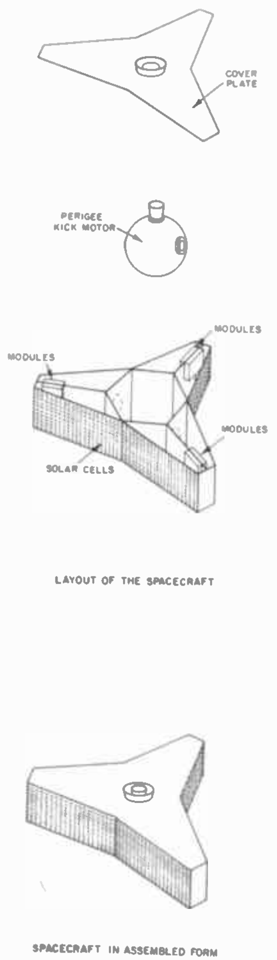


Fig. 15E-5 – Phase III Satellite: The start of worldwide satellite DX.

viously. OSCAR 7 also contains a 70-cm to 2-m transponder which functions in a very similar fashion. Signals around 432.15 MHz are received and retransmitted in the 145.9-MHz range. This Mode B is very sensitive, and even 10 watts erp will produce a fully copyable signal through OSCAR 7.

Scheduled for launch early in 1978 is OSCAR 8, which contains a similar 2-to-10 transponder. It also carries a Mode J transponder, designed and built by a team of Japanese amateurs. This transponder receives signals in the 2-meter band and retransmits them around 435 MHz. Also due to be operational soon is a Russian amateur satellite, but no details are available yet.

The amateur satellite program has a giant step forward scheduled for 1979. The Phase III satellite will be a high-power, elongated-orbit satellite, capable of providing hemisphere-wide communications for as long as 12 hours at a stretch. A built-in kick motor will boost the satellite into a high (25,000 mile) apogee, with a perigee of about 1000 miles, similar to present satellites.

Further details on the amateur satellite program, current operating schedules, details on awards and certificates, and ways to contribute to this important volunteer amateur effort are a few of the items available from satellite experts at the ARRL hq. Write for more information.

PHONE PATCHING

A phone patch is an interconnection made between a radiotelephone system and a wire-line telephone. When the patch is made properly, the radio link and the wire line will effectively extend each other. Phone patches have provided vital communication when a natural disaster has caused disruption of normal communication facilities. More commonly, phone patches permit men in service or on scientific expeditions to talk with their families. Few activities can create a more favorable public image for amateurs than to bring people together in this way. Such public service is always appreciated. Amateurs are using phone patches for their own convenience, too. A phone patch might be used to talk with a friend in a distant city or to make a phone call from a car. In the latter case, a number of clubs are equipping their repeaters with unattended phone patch arrangements.

Occasionally, a phone patch will be used at both ends of a radio link. That is sometimes the case when the radio contact is made to overseas military bases. Some bases have a special phone booth or a small studio where the serviceman can have more privacy and be at ease while in conversation. The studio may be equipped with a regular telephone or it may have a microphone and earphones or a loudspeaker. It is common, too, for the participants to be asked to end each comment with the word "over" as a cue to radio operators (who may be using push-to-talk operation) to reverse the direction of transmission.

Fig. 15F-1 – The voice coupler, to the left of the touch-tone telephone, is supplied by the telephone company. The coupler is normally fixed to a wall or desk, and contains a jack for connection of the amateur's phone patch.

A few general considerations apply to phone patching. It constitutes the handling of third-party traffic. Agreements between governments specifically permitting such traffic must be in effect if the radio link is to a foreign country. Amateurs are responsible for conforming to regulations on station identification, prohibited language and the like while a phone patch is in progress. If a repeater is involved, the arrangement should meet all applicable rules regarding repeater-control facilities. Telephone companies, too, are concerned that the interconnection arrangements be made in the proper way and that the electrical signals meet certain standards.

THE TELEPHONE SYSTEM

Telephone company regulations are published in their tariffs, which in most states must be available in the company's business offices. In the tariffs, phone patches are included under "Interconnection Arrangements" or a similar designation. Telephone employees may not be familiar with the term, "phone patch" so it should be used with caution when talking with them. Patching is accomplished with the aid of devices called "couplers" or "voice connecting arrangements." These are provided by the telephone company and are important in several ways. They protect the amateur's telephone service from interruption that might result from a malfunction in his equipment; they protect other users, too. By isolating the amateur's equipment electrically from the telephone line, they give him a great deal of freedom in the design of his circuits. The protective device also permits proper adjustment of the circuit impedance, energy levels and other operating conditions to be met by the amateur's equipment.

Several different interconnection arrangements are listed in Table I.

A telephone line normally consists of a single pair of wires which is used for both directions of transmission. At the amateur's station it will be terminated in a telephone set. A voice coupler will be connected in parallel with the telephone set when the phone patch is in progress. For design purposes, the telephone set and line are each assumed to have an impedance of about 900 ohms (in the case of residence service) and the best impedance for the phone-patch circuit is also 900 ohms. In operation, the patch will see a load of about 450 ohms. This small mismatch should not be cause for concern, however, as it is the best possible compromise. The phone patch's basic function is to connect the radio receiver's audio output circuit and the radio transmitter's audio input circuit to the telephone voice coupler. It should do this in a way that results in correct circuit impedances and voice levels. Provision should be made, too, for measuring and adjusting the voice level that is transmitted to the telephone line and for electrical filtering to the extent needed to comply with telephone company limitations.

Fig. 15F-1 shows a typical voice coupler and a related telephone set. A simplified schematic diagram of this setup is given in Fig. 15F-2. The telephone is equipped with an exclusion key and a turn button. The telephone operates in the usual way when the two switches are in their normal positions. Lifting the exclusion key causes the voice coupler to be connected to the telephone line. If it is requested when the voice coupler is ordered, the turn button will be supplied and can be wired by the telephone company to cut off the handset transmitter, the receiver, or both of them. The *transmitter* cutoff feature is preferred, as it will eliminate the pickup of room noise by the telephone while permitting the patched communication to be monitored on the handset receiver. The operator can restore the turn button as required for station identification or to break in for other purposes.

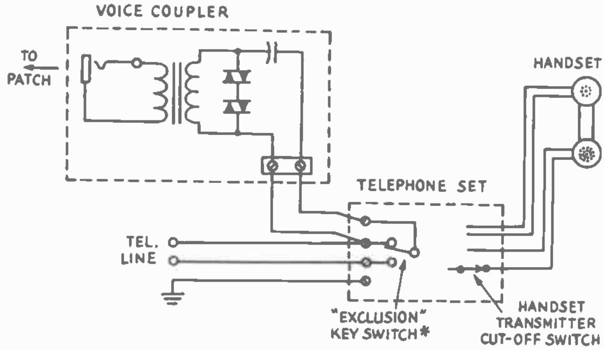


Fig. 15F-2 — Simplified diagram of voice coupler and telephone set. *Both the cutoff switch and the exclusion key switch are shown in their normal positions.

Supplemental information and pertinent telephone company technical specifications as they may apply to amateur radio are given in the appendix which appears at the end of this chapter.

PHONE PATCH CIRCUITS

Where push-to-talk operation is used, the phone patch can be as simple as a transfer switch (connecting the receiver and the transmitter, alternately, to the coupler) or it can be a resistive combining network of the kind shown in Fig. 15F-3. Included in the circuit is a 2600-Hz filter, the need for which is discussed later.

Hybrid Circuits

Where it is desirable to use voice-operated transmitter control (VOX), more elaborate arrangements are required. The VOX circuit must determine when the distant radio station is transmitting and inhibit the local transmitter. When the party

TABLE I

Voice Interconnection Arrangements of Interest to Amateurs

Applicable Bell System Publication	*Arrangement Service Code*	*Arrangement Description*
PUB42101	QKT	Provides manual connection of transmitting or receiving equipment to an exchange line by means of a telephone set; uses a 30A or L-7049A voice coupler. Telephone handset transmitter cutoff is optional. Connection to the coupler is made with a 1/4-inch tip-sleeve plug, provided by the user. Impedance, 900-ohms.
PUB 42208	STC (QX or VX)	Provides automatic (unattended) call origination and answering for one exchange line. Connection to the unit is made with a special plug to be supplied by the user. Required is a Cinch Co. No. 231-15-61-133 plug equipped with a hood, No. 239-13-99-069. Impedance, 600 ohms. Ac power is required.
PUB42402	CD8	Provides automatic (unattended) call origination for up to 14 trunks. Impedance, 600 ohms. Ac power is required.

NOTE: Publications are made available through the telephone company in local areas. Consult your telephone company about the use of these service arrangements.

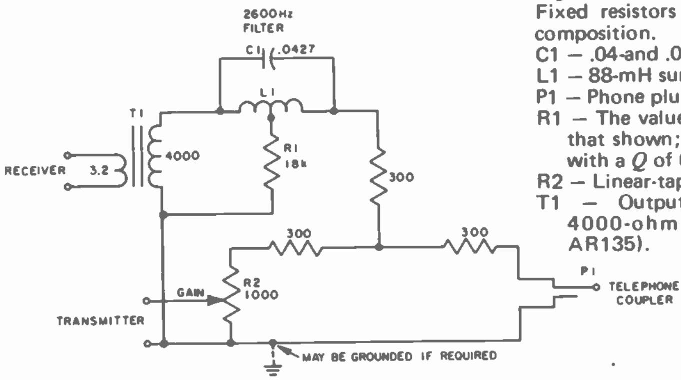


Fig. 15F-3 — Schematic of the simple phone patch. Fixed resistors are 1/2 watt, 5-percent tolerance, composition.
C1 — .04-and .0027-μF paper in parallel.
L1 — 88-mH surplus toroid.
P1 — Phone plug.
R1 — The value of this resistor may be varied from that shown; 18,000 ohms is correct for a toroid with a Q of 63.
R2 — Linear-taper composition control.
T1 — Output transformer, 3.2-ohm primary, 4000-ohm secondary (Lafayette Radio AR135).

on the land telephone is talking, the VOX circuit must activate the local transmitter. This function is made difficult by the difference in audio levels. The phone patch must transmit a voice level of approximately −5 VU toward the telephone line, whereas the level received from the distant land telephone may range from −45 VU to −10 VU.[1] The contrast in levels can be reduced considerably at the input to the local transmitter's VOX circuit by using a "hybrid" circuit. A hybrid circuit is an electrical network connecting together the transmitter, receiver and the voice coupler in such a way that the audio energy from the receiver is canceled at the input to the transmitter. Hybrids require a fourth circuit element, called a balancing network, in order to function.

Several kinds of hybrids can be constructed, the simplest of which is an adaptation of the Wheatstone bridge. Such a hybrid is shown in Fig. 15F-4.

1 Volume units (VU) are measured with an instrument which is basically an ac voltmeter of appropriate range and with dynamic characteristics which are carefully controlled to provide standardized measurement of complex wave forms. When sine-wave power is measured, a VU meter and one calibrated in dB relative to a milliwatt (dBm) should give the same numerical indication.

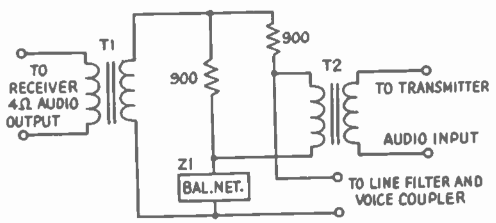


Fig. 15F-4 — Wheatstone-bridge hybrid phone-patch circuit. Resistances are in ohms. Half-watt resistors of 20-percent tolerance are adequate. Filters and level-measuring arrangements are not included in this simple circuit.
T1 — Line to voice coil; primary 1000 ohms, secondary 4 ohms, such as Allied 6W3HFL or equiv.
T2 — Audio; primary 1000 ohms, secondary as appropriate to match transmitter input impedance.
Z1 — Balancing network. See Fig. 15F-5 and text.

When the impedance of the balancing network is equal to the impedance at the input to the line filter, the bridge will be in a condition of balance. The amount of audio from the receiver that reaches the transmitter (or VOX circuit) will then be minimized.

The balancing network, shown schematically in Fig. 15F-5 is not complicated. In most cases it will consist only of a resistor and a capacitor in parallel. Typical values for a condition of balance when a voice coupler is used would be 470 ohms for R1 and .04 μF for C1. Other interface devices, such as might be used at repeaters for unattended operation, will require other values. The resistance might be between 500 and 1200 ohms and the shunt capacitance might range from .01 to 0.1 μF; in rare cases, a series capacitor in the order of 2 μF may be required. The values for a particular installation must be found by trial. The hybrid can be balanced by establishing a telephone call, and tuning in a clear voice signal on the receiver. With headphones connected to the transmitter audio circuit, adjust the hybrid balance network for minimum signal in the headset.

With the Wheatstone bridge hybrid circuit of Fig. 15F-4, losses between the receiver and the telephone line, and between the line and the transmitter, will be in the order of 6 to 10 dB. Transformer-type hybrid circuits exhibit lower losses, only 4.5 to 6 dB. A circuit for a single-transformer hybrid is shown in Fig. 15F-6. A two-transformer arrangement (giving better isola-

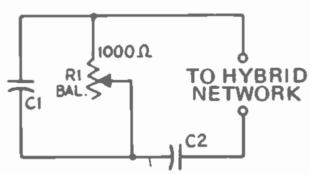


Fig. 15F-5 — Balancing network. R1 is a wire-wound control. C1 and R1 should balance a voice coupler; typical values are 470 ohms and .04 μF. C2 is ordinarily not used, but values in the order of 1 to 4 μF may be required with unattended interconnection devices.

(Use appropriate impedance ratio)

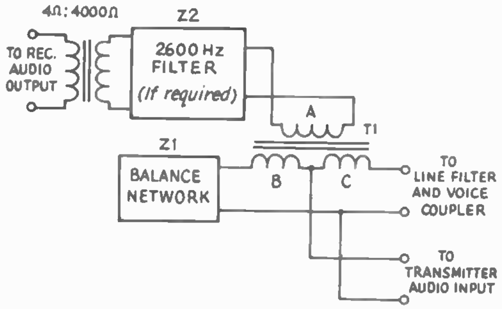


Fig. 15F-6 – Hybrid circuit made with a single audio transformer.
T1 – Windings designated "B" and "C" should be of about 900 ohms impedance each. Winding "A" may be of higher impedance if the 2600-Hz filter is used; a lower impedance may be used to match the receiver if a 2600-Hz filter is not needed.
Z1 – Balancing network. See Fig. 15F-5 and text.
Z2 – 2600-Hz filter (C1, L1, and R1 of Fig. 15F-3).

tion between elements) is shown later in this chapter.

Filters

Standards have been established for the maximum signal levels that can be connected to the input of a coupler or other interconnection device. They are listed in Table II. The limits of out-of-band energy are best met by using a low-pass line filter. Located between the coupler and the hybrid it will protect the line and also band-limit line signals to the transmitter. Filters of several types (image parameter, elliptic function, and so on) may be used. The filter should be of 600- or 900-ohm impedance (depending on the interface), passing frequencies below 3 kHz with losses rising rapidly above that point; a rejection notch should be provided at 4 kHz.

In the long distance network the telephone system uses 2600 Hz as a "disconnect" signal. If patched calls are made to telephone offices distant from your own, the need for filtering at that frequency can best be judged by experience. The filter can be made switchable, if desired. The best location for a 2600-Hz rejection filter is at the receiver output.

REPEATER PATCHES

Some interesting phone-patch possibilities exist at repeaters. Unattended interconnection devices are associated with the repeaters to provide a form of mobile telephone service for the clubs operating them. The connections to a typical unattended interface device are shown in Fig. 15F-7.

Suitable signals generated in mobile units work through a base station to activate the interconnection device, causing it to connect and pass dial pulses to the telephone line. The system may be arranged so that the base transmitter carries both sides of the conversation or only the voice of the distant telephone user. Switching of the patch's voice path between the transmitter and the receiver could be done under the control of tones or a carrier-operated relay. A simple combining circuit may be used if both sides of the conversation are to be put out over the air. To equalize audio levels, a wide-range agc amplifier might have to be provided, or an attenuator in the transmitter audio line would have to be switched in and out. A

TABLE II

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.

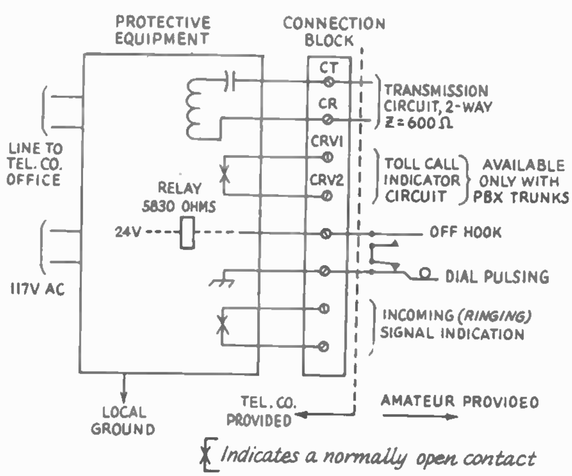


Fig. 15F-7 – Interconnection diagram for a Bell CD8 coupler, representative of connections to unattended interface devices.

hybrid circuit could be used in this case but the retransmitted audio from the mobile unit would not be as free from distortion as with the combining arrangement.

Some telephone lines and interface devices can be arranged to signal the fact that a toll call has been dialed. Such a signal might be used to disconnect the phone patch if the repeater owners do not want long distance calls to be made. Clubs would probably want to control access to the patch in any case, as they would be responsible for all telephone service charges, even if the calls were not made by their members.

A HYBRID PHONE PATCH

The photographs and Fig. 1 show a deluxe 2-transformer hybrid phone patch for home construction. Some form of hybrid circuit is necessary if VOX control of the transmitter is to be used. A third transformer matches the 3.2-ohm output of the receiver. A 2600-Hz filter is provided in the line from the receiver to reduce the possibility of unwanted disconnections resulting from heterodyning signals during use over long-distance telephone lines. The filter may be switched out for local calls for a slight improvement in voice fidelity from the received signal to the telephone line. A modified VU meter indicates the levels received from and applied to the telephone line entering the amateur station. The use of surplus or "bargain" components, especially transformers, will greatly reduce the cost of construction.

The circuit of the phone-patch unit is shown in Fig. 1. C1, L1, and R2 form the 2600-Hz receiver-line filter. Its insertion loss at 1000 Hz is negligible, but is in excess of 15 dB at 2600 Hz. T2 and T3 are the hybrid transformers, with C3 and R5 provided to balance the network. Independent level adjustments are provided for the signal coupled from the receiver to the telephone line (R1) and from the telephone line to the transmitter speech amplifier (R3).

The phone patch unit is built into a homemade aluminum enclosure measuring 3 X 3 X 6 inches. A coating of spray-on enamel, rubber feet, and wet-transfer decal labels plus shiny knobs give the unit a professional appearance.

M1 is a Calectro model DI-930A "VU" meter with its time constant modified by adding external capacitance. The "A" model is identified with the letter A appearing in a circle near the bottom of the meter-scale card. Earlier models of the DI-930 meter, without the A, are unsuitable without internal modification. The correct value of damping capacitance is 400 μF, and may be obtained by connecting four 100-μF 6-V electrolytic capacitors in parallel. These are to be connected directly across the meter terminals, observing proper polarity. This capacitance value applies only to this particular make and model of VU meter. The modified meter responds to speech signals of 3 kHz or less in a way that compares very closely with the measuring sets mentioned in the Bell interface specifications. Error should be less than 1 dB and should be found to be on the safe side. The meter, as modified, has a 1-kHz impedance of approximately 6500 ohms. It should be mounted only on a nonferrous panel.

Construction

The component layout for the phone patch is not critical, and any of several construction techniques is quite acceptable. In the model photographed all components except the modified meter, controls, and phono jacks were mounted on a piece of circuit-board material. The balance control was mounted on the front panel, but this is a "set once and forget" control so some builders may wish to include it inside the enclosure. An etched pattern in the copper foil provides a few of the circuit interconnections, but most connections, including all those to the two hybrid transformers, are made with point-to-point wiring. The UTC transformers specified have mounting studs affixed to the top of the case, and these are used to mount the transformers in an inverted position on the circuit board. This same construction idea can be used with perforated phenolic board and point-to-point wiring for all components, instead of an etched circuit board.

The only precaution to observe during construction is to keep J3 insulated from chassis ground, to reduce rf coupling into the telephone line. In the model photographed this was done by drilling a 1/2-inch hole in the rear panel where J3 was to be mounted, and then, with machine screws, fastening a small piece of phenolic board to cover the hole. Next J3 was mounted on the phenolic board, centered in the hole. Some types of phono jacks come supplied with phenolic mounting material, and if the clearance hole is large enough these types may be mounted directly on a metal panel without grounding the outer contact.

Adjustment

If one has access to an accurately calibrated audio signal generator or to an electronic fre-

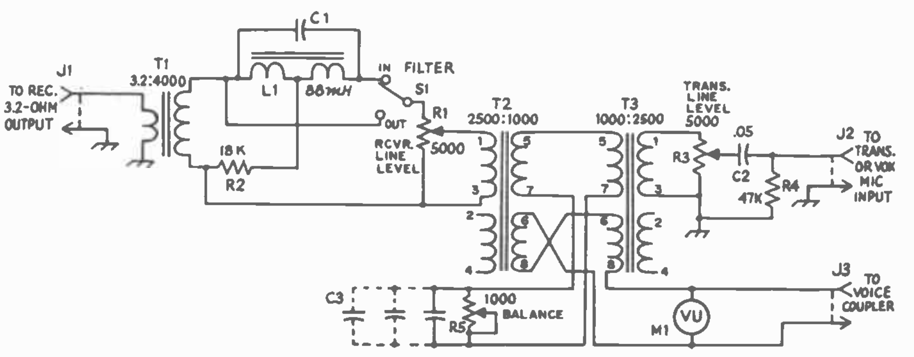


Fig. 1 — Schematic diagram of the phone-patch circuit. Resistances are in ohms, k = 1000. Fixed resistors may be 1/2 watt, 10 percent tolerance. Capacitance is in microfarads. Components not listed below are identified for text reference.

C1 — Capacitors in parallel to give required value of .0427 μF; low-voltage metalized paper or Mylar are suitable.
C3 — Typical value, .04 μF. See text and Fig. 15F-5 if hybrid network cannot be balanced.
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-930A VU meter, modified. See text.
R1, R3 — 5000-ohm audio-taper control (Mallory U12 or equiv.).
R5 — 1000-ohm linear-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 O-19 or equiv.).

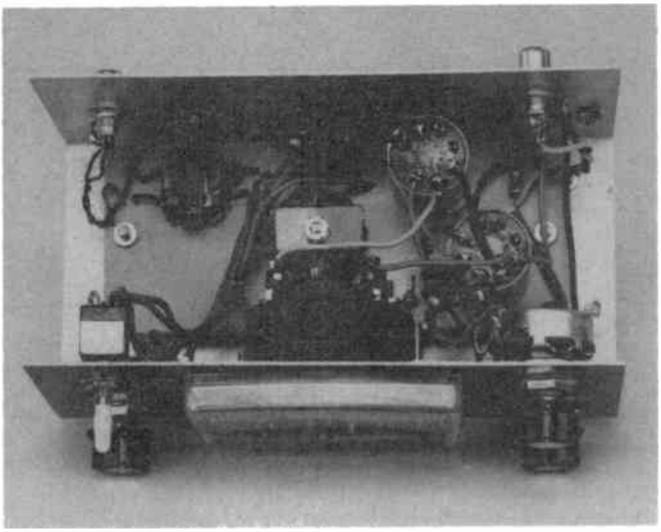
quency counter he may wish to check the notch frequency of the 2600-Hz filter, although this step is not essential. The frequency may be adjusted by using various combinations of fixed-value capacitors for C1 until the notch appears at exactly 2600 Hz. In the model photographed stock-value capacitors, selected at random to provide the specified total capacitance for C1, resulted in a notch frequency of 2621 Hz, which is quite acceptable.

Correct adjustment of the balance control, R5, will facilitate the operation of the transmitter VOX circuit by the distant party on the land telephone. Connect all station equipment to place the patch into operation. Connect a pair of headphones or an ac voltmeter to the transmitter audio circuit. If a sensitive ac VTVM is available, one which will measure in the millivolt range, it may be connected directly to the output from J2, in parallel with the line connected to the transmitter. Establish a phone call and connect the phone patch to the voice coupler. Tune in a clear voice signal on the receiver, and adjust R5 for the best null of the received signal as monitored in the transmitter audio section. If the null does not occur within the range of R5, experimentally try different capacitance values for C3 and a larger value for R5 (connect a fixed-value resistor in series with R5 to obtain a higher equivalent value). With R5 properly adjusted, the distant party should be able to trip the transmitter VOX circuit satisfactorily even though no anti-trip connection is used from the receiver. With such a connection made, VOX operation will be quite reliable.

Installation and Operation

The receiver input to the phone-patch unit may be taken in parallel with the speaker leads from the receiver. Most operators prefer to disconnect or disable the speaker, however, and to connect the patch directly to the speaker-output terminals of the receiver. The switching to and from phone-patch station operation is generally done in suitable control circuits which may be included in the phone-patch enclosure itself, if desired. Operating with the speaker disconnected will result in a 3-dB-greater audio signal being fed to the hybrid circuit, and monitoring of the receiver audio by the amateur operator may be done through the telephone handset.

The level of signal being fed from the receiver to the telephone line during reception may be adjusted either with R1 or with the receiver audio gain control. Similarly, the level of audio being fed to the transmitter from the telephone line during transmission may be adjusted with R3 and with the transmitter microphone gain control. If the distant party on the telephone line is not talking loudly enough for proper operation of the transmitter, remember that often he can be made to speak louder *simply by reducing the level of audio being sent to him.* The speech level should never be permitted to exceed −2 VU on the DI-930A scale. When the telephone connection is made to a nearby point (such as a line served out of the same telephone building as the patched line), the distant listener will receive a more comfortable listening



The layout of the phone-patch components is not critical. The two hybrid-network transformers are visible to the right of center, and in the upper left corner of the circuit board the receiver matching transformer may be seen. Two damping capacitors added during modification of the Calectro DI-930A meter are visible atop the meter case; two more are hidden beneath the meter.

level if the maximum signal is held to about −9 on the meter scale.

Many times when phone-patch operation is heard over the air, the transmitted voice quality of the distant land-telephone party seems to be as good as if he were speaking directly into the station microphone. Occasionally, however, signals will be heard with an undue amount of power-line-frequency hum present on the signal. Of course the quality and level of the voice signal coming in on the telephone line plays an important part in how that voice signal sounds over the air, but sometimes a hum problem can be traced directly to the installation of the phone-patch equipment. In particular, the phone patch (and the voice coupler) should be located away from power supply transformers in station equipment. Complete magnetic shielding may not exist even with steel enclosures for power supplies. If other equipment is mounted nearby, the 60-Hz field can induce hum into the transformers of the phone patch. Hum problems of this sort can usually be solved simply by relocating the position of the phone-patch unit.

During operation of a phone patch in the hf amateur bands it is considered good practice to avoid the transmission of operator chatter, dial tones, dial pulses, ringing and busy signals, as they are not essential to communications.

Appendix

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 first 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 (l/min).
c) Reorder (all trunks busy): 480 and 620 Hz, interrupted at 120 l/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.

Bibliography

Source material and more extended discussions of topics covered in this chapter can be found in the references given below. This listing does not include every article published in *QST* on the subjects of this chapter, however. A detailed bibliography of references in *QST* on any of the subjects amateur television, slow-scan television, radioteletype, phone patching, Oscar and moonbounce, will be sent on request to ARRL, Newington, CT 06111. Please enclose a business-size stamped self-addressed envelope.

RADIOTELETYPE

Antanaitis, "A Simple Two-Transistor A.F.S.K. Generator," *QST*, September, 1969.

Craig, "Teleprinter Selector Magnets," Technical Correspondence, *QST*, September, 1971.

Drake, "An Audio Synthesizer – A Device to Generate RTTY Tones with Crystal-Controlled Accuracy," *QST* April, 1972.

Hall, "Frequency Shift Keying the Johnson Ranger, Valiant, Navigator," *RTTY Journal*, Jan. 1968; "What is RTTY?," *QST*, Dec. 1968.

Hoff, "Transmitting Radioteletype," *QST*, May, 1965; "Audio Frequency-Shift Keying for RTTY," *QST*, June, 1965; "The Mainline TT/L F.S.K. Demodulator," *QST*, August, 1965; "The Mainline ST-3 RTTY Demodulator," *QST*, April, 1970.

Petersen, "The Mainline TT/L-2 F.S.K. Demodulator," Part I, "Construction and Adjustment," *QST*, May, 1969, and Part II, "Circuit Description, and the Mainline F.S.K. Keyer," *QST*, June, 1969.

Schecter, "First Steps in RTTY," *QST*, June, 1971.

AMATEUR TELEVISION

Campbell, "Amateur TV – The Easy Way," *QST*, November, 1962.

Keller, "An Amateur Television Camera," *QST*, November, 1953.

Tilton, "Amateur Television – A Progress Report," *QST*, June, 1950.

SLOW-SCAN TELEVISION

Briles and Gervenack, "Slow-Scan TV Viewing Adapter for Oscilloscopes," *QST*, June, 1970.

Macdonald, "S.C.F.M. – An Improved System for Slow-Scan Image Transmission," Part 1, "Slow-Scan Modulation Tests and Proposed Standards," *QST*, Jan. 1961, and Part II, "Circuit Details," *QST*, Feb. 1961; "A New Narrow-Band Image Transmission System," Part I, "Principles of Slow Scan Picture Reproduction," *QST*, August 1958, and Part II, "Circuit and Construction Details," *QST*, Sept. 1958; "A Slow-Scan Vidicon Camera," in three parts, *QST*, June, July and Aug. 1965.

Tschannen, "A Solid-State SSTV Monitor," *QST*, March, 1971.

PHONE PATCHING

Berry, "Legalize Your Phone Patch," *QST*, May 1969; "An Improved Phone Patch," Hints and Kinks, *QST*, Nov. 1970.

Hoff, "Stopping Telephone Interference," *QST*, March, 1968.

Schleicher, "Phone Patching – Legitimately," *QST*, March, 1969; "Phone Patching – One Year Later," *QST*, Nov., 1970; "Measuring Phone-Patch Levels Accurately," *QST*, February, 1972.

Interference with other Services

RADIO FREQUENCY INTERFERENCE (RFI) has probably been with us since the first amateur stations came on the air some 70 years ago. Fed by the technology that developed during and following WW II, the problem has become an increasing source of irritation between radio operators and their neighbors. Home-entertainment electronics devices now abound, with most families owning at least one television receiver, an a-m or fm radio, and any one of several audio devices (such as a phonograph, an intercom, an electronic guitar, or an electronic organ). Given the innate perversity of these objects to intercept radio signals, it should surprise no one to learn that RFI is one of the most difficult problems amateurs face in their day-to-day operations.

How Serious is the RFI Problem?

In 1974, the FCC received 42,000 RFI complaints, up 20% from the number of complaints received in 1970. Of these, 38,000 involved interference to home-entertainment equipment. Most important, 36,000 of these would never have come to the Commission's attention if the manufacturers had corrected design deficiencies in their home-entertainment products at the time of manufacture. It is of interest to note that over 60% of the interference cases reported in 1974 were related to television interference (TVI).

In the case of television interference, FCC experience shows that 90% of the problems experienced can only be cured at the television receiver. Further, when it comes to audio equipment, the *only* cure for RFI is by treatment of the audio device experiencing the interference. There is nothing an amateur can do to his transmitter which will stop a neighbor's phonograph from acting like a short-wave receiver. It should be emphasized that phonographs and Hi-Fi units are not designed to be receivers, but simply audio devices.

It is clear, therefore, that almost all RFI problems experienced with home-entertainment devices result from basic design deficiencies in this equipment. The few small components or filters which would prevent RFI are often left out of otherwise well-designed products as manufacturers attempt to reduce costs, and hence, to reduce the prices of their products.

The Solution – Consumer Protection

Given the present unacceptable situation, what can we as amateurs do to help the consumer resolve the RFI problem? One step which should certainly be taken is to advise our friends and neighbors to inquire, before they make a purchase of an electronic device, whether the product has been certified for operation in the presence of a radio transmitter. Manufacturers must be made to recognize the RFI protection of their home-entertainment equipment has become essential, and that this must be incorporated. Further, where interference is being experienced, the consumer should be encouraged to contact the manufacturer of his equipment and to request that the *manufacturer* furnish the components or services necessary to eliminate RFI.

What Are Manufacturers Doing Today?

Many responsible manufacturers have a policy of supplying filters for eliminating television interference when such cases are brought to their attention. A list of those manufacturers, and a more thorough treatment of the RFI problem, can be obtained by writing the ARRL. If a given manufacturer is not listed, it is still possible that he can be persuaded to supply a filter; this can be determined by writing either directly to him or to the Electronic Industries Association (EIA).[1]

With respect to audio devices, some manufacturers will supply modified schematic diagrams showing the recommended placement of bypass capacitors and other components to reduce rf susceptibility. One large American manufacturer of Hi-Fi equipment has in some cases supplied the necessary components free of charge, although no consistent policy has been evident and the consumer must still pay to have a serviceman install the components.

While these are encouraging developments, it appears likely that meaningful and widespread corrective action by equipment designers will require both pressure from consumers and establishment of suitable government standards.

Voluntary after-the-fact measures on the part of manufacturers simply are not enough. It is a foregone conclusion that as long as the inclusion of additional components for susceptibility reduction increases a manufacturer's cost, however slightly, there will be reluctance to take steps to improve equipment designs by the manufacturers themselves. What appears to be necessary, therefore, is federal legislation giving the FCC the authority to regulate the manufacture of home-entertainment devices and thus protect the consumer.

It's Up to Us

If requests to manufacturers of home-entertainment equipment for those components and installation services necessary to relieve RFI problems are to be successful, each of us, when faced with an RFI problem, must make known our position to

[1] Electronic Industries Association, 2001 Eye Street, N.W., Washington, DC 20006. Attention: Director of Consumer Affairs.

the manufacturers involved. While a respectful request for assistance will bring more cooperation than a blunt demand, do not hesitate to let the manufacturers know that they have a responsibility to the consumer for correcting the design deficiencies that are causing the problem. Before casting the first stone, however, make sure you're not sitting in a glass house. Certainly, if your own television receiver experiences no interference while you are on the air, it is most likely that interference to a more distant television receiver is not the fault of your transmitter.

All of the above is not to say, however, that we should not continue to assist in resolving RFI problems. Radio amateurs have typically sought to assist their neighbors in correcting RFI problems, even where those problems were in no way attributable to the performance of the transmitter. Ultimately, of course, it is the manufacturers' responsibility to correct those deficiencies which lead to the interception of radio signals. But in the interest of good neighborhood relations, we must continue to provide this assistance wherever older equipment designs are in use.

Clean House First

In approaching an RFI problem, the first step obviously is to make sure that the transmitter has no radiations outside the bands assigned for amateur use. The best check on this is your own a-m or TV receiver. It is always convincing if you can demonstrate that you do not interfere with reception in your own home.

Don't Hide Your Identity

Whenever you make equipment changes – or shift to a hitherto unused band or type of emission – that might be expected to change the interference situation, check with your neighbors. If no one is experiencing interference, so much the better; it does no harm to keep the neighborhood aware of the fact that you are operating without bothering anyone.

Should you change location, make your presence known and conduct occasional tests on the air, requesting anyone whose reception is being spoiled to let you know about it so steps may be taken to eliminate the trouble.

Act Promptly

The average person will tolerate a limited amount of interference, but the sooner you take steps to eliminate it, the more agreeable the listener will be; the longer he has to wait for you, the less willing he will be to cooperate.

Present Your Story Tactfully

Whenever a device intercepts your signals, it is natural for the complainant to assume that your transmitter is at fault. If you are certain that the trouble is not in your transmitter, explain to the listener that the reason lies in the receiver design, and that some modifications may have to be made in the receiver if he is to expect interference-free reception.

Arrange for Tests

Most listeners are not very competent observers of the various aspects of interference. If at all possible, enlist the help of another amateur and have him operate your transmitter while you see for yourself what happens at the affected receiver.

In General

In this "public relations" phase of the problem a great deal depends on your own attitude. Most people will be willing to meet you half way, particularly when the interference is not of long standing, if you as a person make a good impression. Your personal appearance is important. So is what you say about the receiver – no one takes kindly to hearing his possessions derided. If you discuss your interference problems on the air, do it in a constructive way – one calculated to increase listener cooperation, not destroy it.

VHF TELEVISION

For the amateur who does most of his transmitting on frequencies below 30 MHz, the TV band of principal interest is the low vhf band between 54 and 88 MHz. If harmonic radiation can be reduced to the point where no interference is caused to Channels 2 to 6, inclusive, it is almost certain that any harmonic troubles with channels above 174 MHz will disappear also.

The relationship between the vhf television channels and harmonics of amateur bands from 14 through 28 MHz is shown in Fig. 16-1. Harmonics of the 7- and 3.5-MHz bands are not shown because they fall in every television channel. However, the harmonics above 54 MHz from these bands are of such high order that they are usually rather low in amplitude, although they may be strong enough to interfere if the television receiver is quite close to the amateur transmitter. Low-order harmonics – up to about the sixth – are usually the most difficult to eliminate.

Of the amateur vhf bands, only 50 MHz will have harmonics falling in a vhf television channel (channels 11, 12 and 13). However, a transmitter for any amateur vhf band may cause interference if it has multiplier stages either operating in or having harmonics in one or more of the vhf TV channels. The rf energy on such frequencies can be radiated directly from the transmitting circuits or coupled by stray means to the transmitting antenna.

Frequency Effects

The degree to which transmitter harmonics or other undesired radiation actually in the TV channel must be suppressed depends principally on

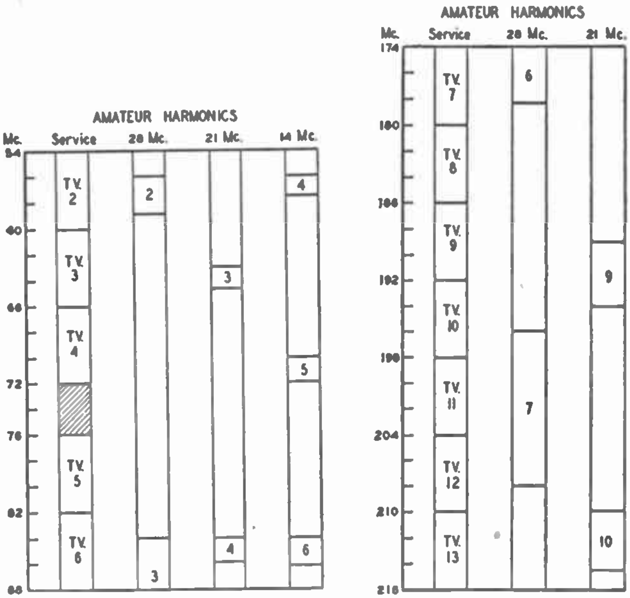


Fig. 16-1 – Relationship of amateur-band harmonics to vhf TV channels. Harmonic interference from transmitters operating below 30 MHz is likely to be serious in the low-channel group (54 to 88 MHz).

two factors, the strength of the TV signal on the channel or channels affected, and the relationship between the frequency of the spurious radiation and the frequencies of the TV picture and sound carriers within the channel. If the TV signal is very strong, interference can be eliminated by comparatively simple methods. However, if the TV signal is very weak, as in "fringe" areas where the received picture is visibly degraded by the appearance of set noise or "snow" on the screen, it may be necessary to go to extreme measures.

In either case the intensity of the interference depends very greatly on the exact frequency of the interfering signal. Fig. 16-2 shows the placement of the picture and sound carriers in the standard TV channel. In Channel 2, for example, the picture carrier frequency is 54 + 1.25 = 55.25 MHz and the sound carrier frequency is 60 – 0.25 = 59.75 MHz. The second harmonic of 28.010 kHz (56,020 kHz or 56.02 MHz) falls 56.02 – 54 = 2.02 MHz above the low edge of the channel and is in the region marked "Severe" in Fig. 16-2. On the other hand, the second harmonic of 29,500 kHz (59,000 kHz or 59 MHz) is 59 – 54 = 5 MHz from the low edge of the channel and falls in the region marked "Mild." Interference at this frequency has to be about 100 times as strong as at 56,020 kHz to cause effects of equal intensity. Thus an operating frequency that puts a harmonic near the picture carrier requires about 40 dB more harmonic suppression in order to avoid interference, as compared with an operating frequency that puts the harmonic near the upper edge of the channel.

For a region of 100 kHz or so either side of the sound carrier there is another "Severe" region where a spurious radiation will interfere with reception of the sound program and this region also should be avoided. In general, a signal of intensity equal to that of the picture carrier will not cause noticeable interference if its frequency is in the "Mild" region shown in Fig. 16-2, but the same intensity in the "Severe" region will utterly destroy the picture.

Interference Patterns

The visible effects of interference vary with the type and intensity of the interference. Complete "blackout," where the picture and sound disappear

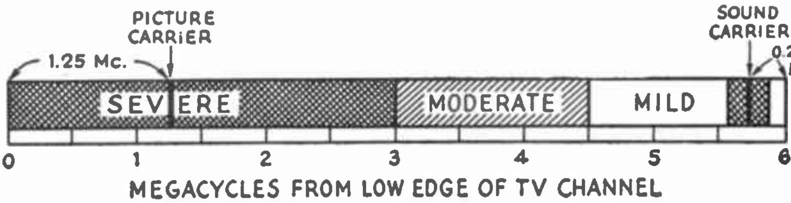


Fig. 16-2 – Location of picture and sound carriers in a monochrome television channel, and relative intensity of interference as the location of the interfering signal within the channels is varied without changing its strength. The three regions are not actually sharply defined as shown in this drawing, but merge into one another gradually.

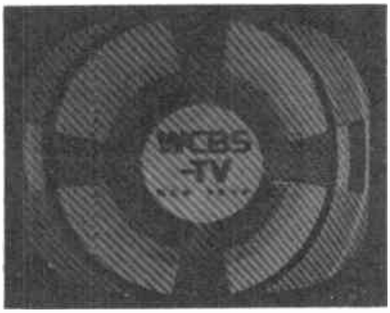


Fig. 16-3 – "Cross-hatching," caused by the beat between the picture carrier and an interfering signal inside the TV channel.

completely, leaving the screen dark, occurs only when the transmitter and receiver are quite close together. Strong interference ordinarily causes the picture to be broken up, leaving a jumble of light and dark lines, or turns the picture "negative" – the normally white parts of the picture turn black and the normally black parts turn white. "Cross-hatching" – diagonal bars or lines in the picture – accompanies the latter, usually, and also represents the most common type of less severe interference. The bars are the result of the beat between the harmonic frequency and the picture carrier frequency. They are broad and relatively few in number if the beat frequency is comparatively low – near the picture carrier – and are numerous and very fine if the beat frequency is very high – toward the upper end of the channel. Typical cross-hatching is shown in Fig. 16-3. If the frequency falls in the "Mild" region in Fig. 16-2 the cross-hatching may be so fine as to be visible only on close inspection of the picture, in which case it may simply cause the apparent brightness of the screen to change when the transmitter carrier is thrown on and off.

Whether or not cross-hatching is visible, an amplitude-modulated transmitter may cause "sound bars" in the picture. These look about as shown in Fig. 16-4. They result from the variations in the intensity of the interfering signal when modulated. Under most circumstances modulation bars will not occur if the amateur transmitter is frequency- or phase-modulated. With these types of modulation the cross-hatching will "wiggle" from side to side with the modulation.

Except in the more severe cases, there is seldom any effect on the sound reception when interference shows in the picture, unless the frequency is quite close to the sound carrier. In the latter event the sound may be interfered with even though the picture is clean.

Reference to Fig. 16-1 will show whether or not harmonics of the frequency in use will fall in any television channels that can be received in the locality. It should be kept in mind that not only harmonics of the final frequency may interfere, but also harmonics of any frequencies that may be present in buffer or frequency-multiplier stages. In the case of 144-MHz transmitters, frequency-multiplying combinations that require a doubler or tripler stage to operate on a frequency actually in a low-band vhf channel in use in the locality should be avoided.

Harmonic Suppression

Effective harmonic suppression has three separate phases:

1) Reducing the amplitude of harmonics generated in the transmitter. This is a matter of circuit design and operating conditions.

2) Preventing stray radiation from the transmitter and from associated wiring. This requires adequate shielding and filtering of all circuits and leads from which radiation can take place.

3) Preventing harmonics from being fed into the antenna.

It is impossible to build a transmitter that will not generate *some* harmonics, but it is obviously advantageous to reduce their strength, by circuit design and choice of operating conditions, by as large a factor as possible before attempting to prevent them from being radiated. Harmonic radiation from the transmitter itself or from its associated wiring obviously will cause interference just as readily as radiation from the antenna, so measures taken to prevent harmonics from reaching the antenna will not reduce TVI if the transmitter itself is radiating harmonics. But once it has been found that the transmitter itself is free from harmonic radiation, devices for preventing harmonics from reaching the antenna can be expected to produce results.

REDUCING HARMONIC GENERATION

Since reasonably efficient operation of rf power amplifiers always is accompanied by harmonic generation, good judgment calls for operating all frequency-multiplier stages at a very low power level. When the final output frequency is reached, it is desirable to use as few stages as possible in building up to the final output power level, and to use tubes that require a minimum of driving power.

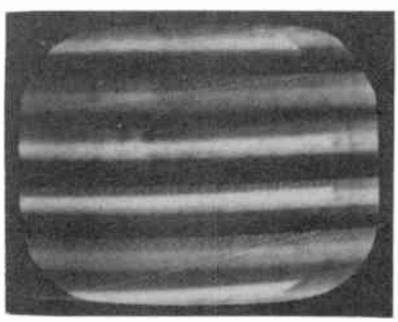


Fig. 16-4 – "Sound bars" or "modulation bars" accompanying amplitude modulation of an interfering signal. In this case the interfering carrier is strong enough to destroy the picture, but in mild cases the picture is visible through the horizontal bars. Sound bars may accompany modulation even though the unmodulated carrier gives no visible cross-hatching.

Circuit Design and Layout

Harmonic currents of considerable amplitude flow in both the grid and plate circuits of rf power amplifiers, but they will do relatively little harm if they can be effectively bypassed to the cathode of the tube. Fig. 16-5 shows the paths followed by harmonic currents in an amplifier circuit; because of the high reactance of the tank coil there is little harmonic current in it, so the harmonic currents simply flow through the tank capacitor, the plate (or grid) blocking capacitor, and the tube capacitances. The lengths of the leads forming these paths is of great importance, since the inductance in this circuit will resonate with the tube capacitance at some frequency in the vhf range (the tank and blocking capacitances usually are so large compared with the tube capacitance that they have little effect on the resonant frequency). If such a resonance happens to occur at or near the same frequency as one of the transmitter harmonics, the effect is just the same as though a harmonic tank circuit had been deliberately introduced; the harmonic at that frequency will be tremendously increased in amplitude.

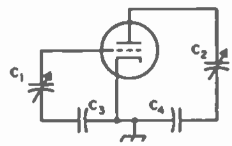


Fig. 16-5 – A vhf resonant circuit is formed by the tube capacitance and the leads through the tank and blocking capacitors. Regular tank coils are not shown, since they have little effect on such resonances. C1 is the grid tuning capacitor and C2 is the plate tuning capacitor. C3 and C4 are the grid and plate blocking or bypass capacitors, respectively.

Such resonances are unavoidable, but by keeping the path from plate to cathode and from grid to cathode as short as is physically possible, the resonant frequency usually can be raised above 100 MHz in amplifiers of medium power. This puts it between the two groups of television channels.

It is easier to place grid-circuit vhf resonances where they will do no harm when the amplifier is link-coupled to the driver stage, since this generally permits shorter leads and more favorable conditions for bypassing the harmonics than is the case with capacitive coupling. Link coupling also reduces the coupling between the driver and amplifier at harmonic frequencies, thus preventing driver harmonics from being amplified.

The inductance of leads from the tube to the tank capacitor can be reduced not only by shortening but by using flat strip instead of wire conductors. It is also better to use the chassis as the return from the blocking capacitor or tuned circuit to cathode, since a chassis path will have less inductance than almost any other form of connection.

The vhf resonance points in amplifier tank circuits can be found by coupling a grid-dip meter covering the 50-250 MHz range to the grid and plate leads. If a resonance is found in or near a TV channel, methods such as those described above should be used to move it well out of the TV range. The grid-dip meter also should be used to check for vhf resonances in the tank coils, because coils made for 14 MHz and below usually will show such resonances. In making the check, disconnect the coil entirely from the transmitter and move the grid-dip meter coil along it while exploring for a dip in the 54-88-MHz band. If a resonance falls in a TV channel that is in use in the locality, changing the number of turns will move it to a less-troublesome frequency.

Operating Conditions

Grid bias and grid current have an important effect on the harmonic content of the rf currents in both the grid and plate circuits. In general, harmonic output increases as the grid bias and grid current are increased, but this is not necessarily true of a *particular* harmonic. The third and higher harmonics, especially, will go through fluctuations in amplitude as the grid current is increased, and sometimes a rather high value of grid current will minimize one harmonic as compared with a low value. This characteristic can be used to advantage where a particular harmonic is causing interference, remembering that the operating conditions that minimize one harmonic may greatly increase another.

For equal operating conditions, there is little or no difference between single-ended and push-pull amplifiers in respect to harmonic generation. Push-pull amplifiers are frequently troublemakers on even-order harmonics because with such amplifiers the even-harmonic voltages are in phase at the ends of the tank circuit and hence appear with equal amplitude across the whole tank coil, if the center of the coil is not grounded. Under such circumstances the even harmonics can be coupled to the output circuit through stray capacitance between the tank and coupling coils. This does not occur in a single-ended amplifier having an inductively coupled tank, if the coupling coil is placed at the cold end, or with a pi-network tank.

SOME TVI TESTS

One of the difficulties in solving TVI problems, particularly in fringe areas, is the number of possible causes and their elusive nature. A "cure" seems to be found only to have the problem return with renewed severity the next day. Consequently, some tests were performed by the ARRL in order to isolate the causes, if possible. Although the results weren't encouraging in regard to certain

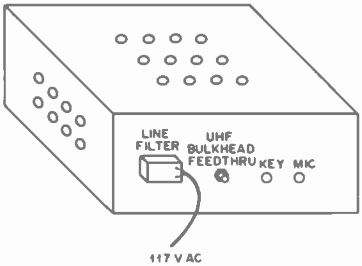


Fig. 1 — Construction details of the rf enclosure. For the model shown, thin aluminum sheet metal was used to form a box 12 X 28 X 20 inches (HWD) (30 X 71 X 51 cm). Small holes were drilled for ventilation and a fan might be advisable if temperature rise is considered a problem. Feed-through connectors can be of the builder's choice but ac conduits and control leads should be filtered. For key and mic leads, bypass with 0.001-μF disk-ceramic capacitors and install a small ferrite bead (if available). A commercially manufactured line filter was used although a homemade one of the "brute force" type would also be suitable.

Although dimensions and material are not critical, the cabinet should be deep enough to form as much of an overlap as possible between the front of the equipment panel and the front of the cabinet. It is important that all leads be run through the rear of the cabinet.

aspects, one bright spot was some "fallout" in the way of additional suppression measures previously neglected.

Test Procedures

A ham experiencing some TVI in a fringe area (on his own set) generously agreed to be the "patient" in the tests. A large screened enclosure was transported to his location. It could contain the TV set along with a smaller version of the screen enclosure (Fig. 1) for the transmitter. Other equipment included a gasoline-powered generator that could power either the TV set or transmitter. There is always the possibility that feedback of rf energy through the power line (or "conducted interference") is a factor in a TVI problem. By running the equipment on separate power systems, some idea of the importance of this type of coupling would be possible.

Various low-pass filters, high-pass filters, and power harnesses made up the rest of the equipment list. Checks on various TV channels indicated the most serious problem resulted from third-harmonic energy on channel 3 during 15-meter operation. Tests were performed with the rig inside and out of its shielded enclosure, TV set inside and out of the larger enclosure, and with either the TV set or transmitter on independent power.

Test Results

Previous checks in the lab revealed that almost all currently manufactured amateur transmitters and transceivers emit harmonics in the form of "chassis radiation" to varying degrees. It should be pointed out that no outstanding "saints" were found in this area but mostly "sinners." Once this energy escapes from the transmitter cabinet, it can be conducted to the antenna or back through the power line via a single-conductor waveguide type of mode. This mode is very similar to the propagation of rf energy over a two-dimensional conducting surface in the form of a ground wave. But the important thing to keep in mind is that devices such as filters, traps, and grounds are ineffective since the rf energy flows *around* the suppression network. The only effective measure is adequate shielding.

As might be expected, the field tests verified the importance of this aspect. *The only test that indicated appreciable reduction of TVI was the one with the transmitter placed inside of its shielded enclosure.* In all other tests, there was no improvement or the change was so slight as to be inconclusive. Some residual interference still remained with the rig enclosed in the shield. This was likely caused by rectification in the external environment. One unexpected result was that no noticeable difference was observed with the door of the enclosure opened or closed. In fact, with the door partially closed and touching the shield at only a few points the TVI became worse!

Conclusions

The transmitter power for these tests was approximately 180 watts: Considerable work has yet to be performed to determine the important

Lab simulation of TVI tests discussed in the text. A dummy load placed next to the "rabbit ears" served as the transmitting antenna.

Severe interference occurred with the setup shown above with the transmitter out of the enclosure. Interference was about the same with the leads running out the front of the cabinet instead of through the rear connectors.

The interference was either reduced considerably (as shown in this photo) or eliminated completely depending upon TV signal strength with all the leads exiting out the rear of the cabinet.

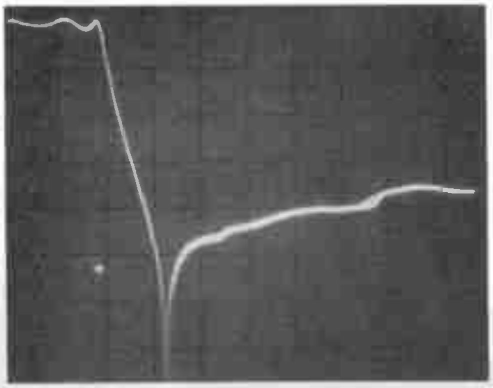
factors at higher levels. However, there is hope for the ham experiencing TVI because of chassis radiation. With the door open, it is believed that the enclosure acts as a waveguide below cutoff and still offers some measure of suppression while permitting access to the controls. In discussing the tests with other amateurs experiencing TVI, reports from the field were favorable when similar measures were tried.

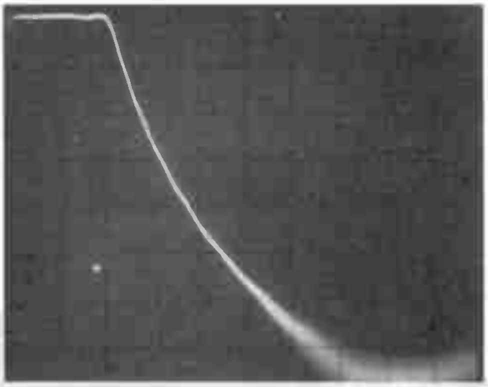
Some further experimentation along these lines is in order. For instance, former shielding theory advocated the use of *high-conductivity* materials. Newer methods often rely on the dissipation of unwanted rf energy in lower conductivity materials such as steel. Although rf energy can penetrate deeper into low-conductivity metals, and greater thicknesses are required to provide the same isolation (as that of copper for example), other problems are simplified. Unwanted rf energy must be dissipated *somewhere* and when a good conductor is used for a shielding enclosure there is a greater tendency for this energy to leak out through doors, conduits, and other points of entry. By dissipating energy internally on the shield walls, there is less chance for it to leak out. (However, if the unexpected attenuation with the door open was actually caused by a waveguide-below-cutoff effect as speculated, high-conductivity material near the door opening would be advisable.)

Other Results

With the TV set in its shielded enclosure, and with power fed through a commercially manufactured line filter, there was no difference in TVI with the rig *or* TV set on independent power. Rf signal energy from the TV antenna was fed into the shield enclosure through a commercially manufactured high-pass filter. Little change was noted when these measures were eliminated and the set operated on the same power service as the transmitter, without a high-pass filter, and outside of the shield.

It should not be concluded that such measures will be equally ineffective under all circumstances. However, the claims of some manufacturers are open to question. Items such as power supplies





Stray lead inductance of a capacitor can degrade filter performance. (See p. 492.)

that eliminate TVI, and similar nostrums have come to our attention. Consequently, common sense is in order in judging whether or not a particular device will prove effective in eliminating interference or how it should be employed.

SUPPRESSION PRACTICES

As the test results reported in the previous section reveal, complete elimination of TVI is often not a simple process. It seldom happens that a single measure such as installing a high-pass filter at the TV set will cure the problem. Rather, a number of methods must be applied simultaneously. The principal factor in any TVI situation is the ratio of TV signal strength to interference level. This includes interference of all types such as ignition noise, random or thermal noise (which isn't really interference but sets the minimum signal that permits "snow-free" reception), and unwanted signals that fall within the TV channel. A signal-to-interference ratio greater than approximately 35 to 40 dB is required for good picture quality.

In this regard, an area frequently overlooked in TVI difficulties is the TV-set antenna. A poor antenna with little gain in the direction of the TV station, old and corroded wire and connections (which can *cause* the generation of harmonics by rectification of a "clean" signal generated in a nearby amateur transmitter), may result in a TVI situation that is impossible to solve. For instance, the "simulated" lab tests illustrated in the photographs were performed with a dummy load next to a set of "rabbit ears" which comprised the TV-set antenna. With a good outdoor system, the TVI would not have been present even though there was a leakage from the transmitter cabinet. Generally speaking, if the picture quality on the TV set experiencing the interference is poor to begin with, even sophisticated suppression measures are likely to prove futile.

Grounds

Grounding of equipment has long been considered to be a first step in eliminating interference. While the method is very effective in the mf range and below, for all practical purposes it is useless in suppressing vhf energy. This is because even short lengths of wire have considerable reactance at vhf. For instance, suppose a length of wire by itself has an inductance of 1 μH. At 550 kHz, the reactance would be about 3.46 Ω. On the other hand, the same wire would have a reactance of over 300 Ω at 56 MHz, which is the frequency range of TV channel 2. (Actually, the impedance of a wire becomes a more complicated entity to define at vhf. The delay effects along the wire are similar to those on the surface of an antenna. Consequently, the wire might even appear as an open circuit rather than as a ground as the electrical length approaches a quarter wavelength.)

From a shock-hazard point of view, grounding is important. *However, never connect a ground for any reason to the chassis of a TV set.* This is because many TV sets derive their operating voltages directly from the ac-service line. Although a schematic diagram of a TV set may indicate a "power transformer" is being used, caution should be exercised to be sure it is actually being employed for this purpose. Quite often, the only voltage the transformer is supplying is for the filament of the TV picture tube.

Shielding

Effective shielding is perhaps the single most important measure in preventing or curing any RFI problem. However, as pointed out in previous sections, unwanted rf energy must be dissipated. The task becomes harder to perform when the spacing between the source of energy and the boundaries of the shield diminish. Consequently, the use of a double shield (as used in the tests) is one way of reducing residual radiation from the primary shielding surface.

In order to obtain maximum effectiveness of a particular shielding measure, no breaks or points of entry should be permitted. Small holes for ventilation purposes usually do not degrade shielding effectiveness. But even here, a honeycomb type of duct is often employed when maximum isolation is required. (A parallel bundle of small tubing has very high attenuation since each tube by itself acts as a waveguide below cutoff.)

The isolation of a coaxial cable can be degraded considerably unless the ends of the shield are terminated properly. A braid should be soldered so that it completely encloses the inner conductor(s) at the connector junction. For instance, the practice of twisting the braid and point soldering it to the base of a connector may result in a 20-dB degradation in isolation. Normally, this effect is not serious if the cable run is through an area where sensitive circuits don't exist. However, the isolation afforded by a filter can be reduced considerably in circuits where such cable breaks occur.

Once instance where a shield break causes a serious problem is in the connection between the antenna terminals on a TV set and the tuner. Newer sets have a 75-Ω coaxial input along with a balun for 300-Ω line. However, because many TV sets have direct connections to the ac line, a decoupling network is used. The shielded lead to the tuner is broken and a capacitor is connected in series with the braid. This provides a low-impedance path for rf energy while presenting a high impedance at 60 Hz. Consequently, because of the cable break, high-pass filters at the antenna input terminals are not as effective as those built into the tuner itself.

Capacitors at RF

Capacitors are common elements found in almost any piece of electronics gear. However, some precautions are necessary when they are employed in RFI-preventive purposes such as in filters and bypassing applications. In particular, lead inductance may be sufficient to resonate with the capacitor proper and cause the entire combination to have a high inductive reactance rather than the desired capacitive reactance.

Fig. 2 — Additional lead filtering for harmonics or other spurious frequencies in the high vhf TV band (174-216 MHz).
C1 — .001-μF disk ceramic.
C2 — 500- or 1000-pF fed-through bypass (Centralab FT-1000. Above 500 volts, substitute Centralab 858S-500.)
RFC — 14 inches No. 26 enamel close-wound on 3/16-inch dia. form or composition resistor body.

This effect is illustrated in the accompanying photographs. The response curve shown at A is for a 10-MHz low-pass filter arranged in a "pi" configuration. However, this particular circuit realization required some large-valued capacitors. Using ordinary capacitor types resulting in an unwanted resonance as evidenced by the sharp dip in the response curve at approximately 15 MHz. However, by going to the equivalent "T" configuration (see the section on filters in the chapter on electrical laws and circuits), a circuit realization for the desired response required much smaller capacitance values. The curve shown in the photo at B approximated this response quite closely and no effects of parasitic inductance were noticeable. When designing filters, it is advisable to compute the component values for as many configurations as possible in order to determine which one results in the most practical elements. If large capacitance values are unavoidable, either special low-inductance types should be used or a number of ordinary smaller-valued capacitors can be paralleled to reduce the effect of lead inductance.

A very desirable capacitor (C2) from an RFI point of view is shown in Fig. 2. Instead of having two or more plates arranged in a parallel fashion, the conductors are coaxial and are separated by the dielectric. Such feedthrough capacitors are highly recommended for conducting leads in and out of circuits where the radiation of harmonic energy is possible. In addition, the rfc illustrated in Fig. 2 could either consist of a small coil wound over a composition resistor as shown or it could be a ferrite bead on a straight piece of wire.

Decoupling from the AC Line

Direct feedback of rf energy into the ac power service is usually not a problem with modern transmitting equipment. However, currents induced on the antenna feed line may flow on the transmitter chassis and back into the ac line. A rig "hot" with rf or even the presence of "broadcast harmonics" while receiving may mean a problem of this sort. In the case where an antenna is being used that requires a ground (such as an end-fed wire), never use any part of the ac conduits, water systems, or other conductors in a building. It is always advisable to have a separate ground system for the antenna itself.

It is also good practice to use an antenna-matching network with no direct connection between the transmitter and antenna feed line. Any matching network that uses mutual-magnetic coupling exclusively will fulfill this requirement. Antenna pattern is another factor to consider and if possible, a type should be used that directs the minimum possible signal into other dwellings. For instance, ground-mounted vertical antennas have considerable low-angle radiation while a dipole directs energy at angles *below* the horizontal plane. A vertical ground plane or beam mounted on as high a tower as practical will generally be better from an RFI and TVI standpoint than antennas closer to the ground.

FILTERS AND INTERFERENCE

The judicious use of filters, along with other suppression measures such as shielding, has provided solutions to interference problems in widely varying applications. As a consequence, considerable attention has been given to the subject over the years that has resulted in some very esoteric designs. Perhaps the most modern approach is the optimization and/or realization for a particular application of a filter by means of a digital computer. However, there are a number of other types with component values cataloged in tabular form. Of these, the most important ones are the so-called Chebyshev and elliptic-function filters. (Butterworth filters are often considered as a special case of Chebyshev types only with a ripple factor of zero.)

Elliptic-function filters might be considered optimum in the sense that they provide the sharpest rolloff between the passband and stopband. Computed values for a low-pass filter with a 0.1-dB ripple in the passband and a cutoff frequency of 30.6 MHz are shown in Fig. 1. The filter is supposed to provide an attenuation of 35 dB above 40 MHz. An experimental model was built and the response is shown in Fig. 2. As can be seen, the filter came quite close to the design goals.

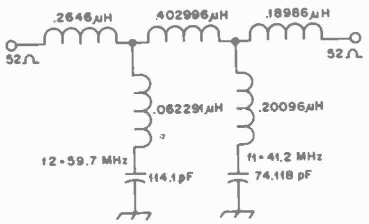


Fig. 1 — Schematic diagram showing component values of an experimental elliptic function filter.

Unfortunately, as with most of the designs in this section, alignment of the more complicated filters requires some sort of sweep-generator setup. This is the only practical way of "tweaking" a filter to the desired response. While building a sweep setup is not beyond the talents of an advanced experimenter, the lack of one is an obstacle in the home construction of filters.

Chebyshev Filters

Chebyshev low-pass filters (and Butterworth filters) have the same ladder-network circuit as the elliptic-function filter in Fig. 1 except that the inductors in the shunt arms are omitted. Tables for the element values are quite common and can be found in any number of references. However, how to determine the attenuation at a particular frequency is often not included in such tables and some explanation is in order. It will be recalled that a ripple factor was mentioned in conjunction with the elliptic-function filter in the previous section. This factor specifies the allowable amount of attenuation in the passband and represents a tradeoff from steepness of the attenuation between the passband and stopband. Larger ripple factors result in greater rolloff; however, the input impedance and consequently the VSWR of the filter become larger also. For moderate power-transmitting applications, a ripple factor of 0.1 dB is about the maximum permissible amount. This results in a VSWR of approximately 1.4:1. For low-level stages, VSWR is often not a problem and higher ripple factors can be used.

The attenuation (or insertion loss in the case of equal resistive terminations at the input and output of the filter) is given by:

$$L = 10 \log_{10} \frac{1}{1 + \epsilon^2 T_n^2(\omega)}$$

where $Tn(\omega)$ represents a Chebyshev polynomial of degree n, and n represents the number of inductors and capacitors in the filter (for instance, for an ordinary pi or T network, n would be 3). The term ω is just $2\pi f$ (f is the frequency in Hz and ϵ^2 will be discussed shortly). Chebyshev polynomials can be expressed in terms of ordinary trigonometric and hyperbolic functions by:

$$T_n(\omega) = \begin{Bmatrix} \cos(n \text{ arc } \cos\omega) & 0<\omega<1 \\ \cosh(n \text{ arc } \cosh\omega) & \omega>1 \end{Bmatrix}$$

For values of ω less than one, the polynomial oscillates between ± 1 while for greater values, it increases rapidly. Consequently, the value for ω equal to 1 represents the cutoff frequency of the filter. While the polynomial could be tabulated from tables of functions, the problem could be easily solved on many current calculator models. In fact, with programmable models such as the Hewlett Packard HP-25, finding the attenuation at any frequency only requires entry of the frequency, ripple factor and number of elements (number of poles). For those interested, a copy of such a program is available from ARRL for 25 cents and an s.a.s.e.

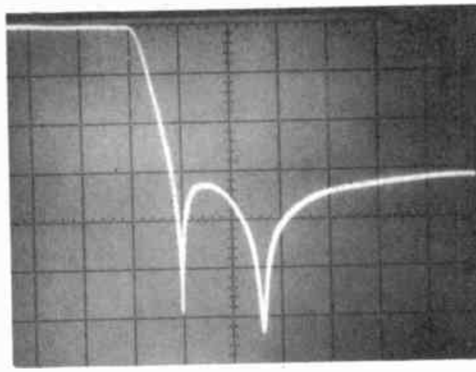


Fig. 2 — Response curve of the filter shown in Fig. 1. Vertical scale represents 10 dB/div. and horizontal scale is 10 MHz/div.

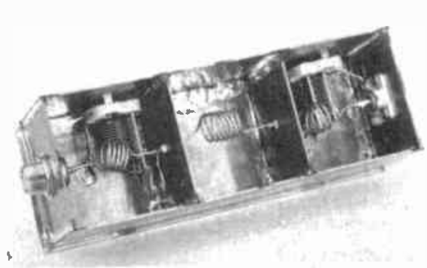
The term ϵ^2 is the ripple factor and is related to the ripple factor in decibels by the equation:

$$\epsilon^2 = 10^{\frac{\text{ripple (dB)}}{10}} - 1$$

This concept represents the most important aspect of current filter design. Limits or tolerances are set on the amount of ripple in either the passband or stopband (or both in the case of elliptic-function filters) and the filter is designed around these limits.

A Citation Eliminator

Quite often, some insight into the qualitative manner in which a filter works is useful. For example, consider the filter shown in Fig. 3. If the 2.26-μH coils were omitted, a high-pass configuration would result. By including the coils, the filter will possibly have a rolloff above the high-pass cutoff frequency and provide an unsymmetrical bandpass characteristic.



Pc board serves as an enclosure for this elliptic-function filter.

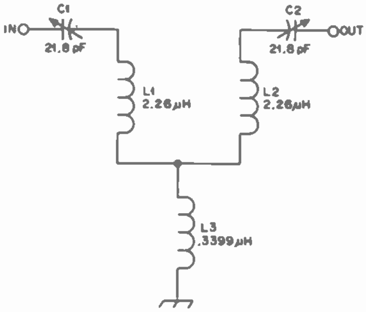


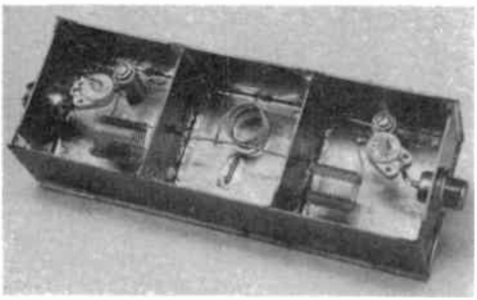
Fig. 3 – Schematic diagram of the "Citation Eliminator." Component values shown are theoretical computed inductances and capacitances for a bandpass filter resonant at 21.14 MHz with a bandwidth of 4.5 MHz.

C1, C2 – 4.5 to 25 pF, ceramic trimmer.

L1, L2 – 13 turns 3/4-inch dia., 16 tpi (B & W 3011).

L3 – 5 turns No. 16 solid wire 3/4-inch dia., approx. 1-inch long.

This is shown in Fig. 4 and an application of a filter of this type is as follows. Many older rigs suitable for cw work are often acquired by Novice operators because of their low cost. Unfortunately, operation on the higher bands such as 15 meters can be somewhat tricky and it is possible that the rig is tuned up on 20 meters instead. The aforementioned filter eliminates that possibility by providing rejection at 14 MHz and also at harmonics above the operating frequency of 21 MHz. It is relatively easy to align since all that is necessary is to grid-dip L1 to 21.14 MHz (with the input shorted and the output open) by means of C1. The process is repeated with L2 and C2, only the input is opened and the output terminal shorted. Further tweaking can be accomplished by adjusting the capacitors for minimum SWR with the output circuit connected to a dummy load. Some adjustment of L3 may be necessary which controls the coupling. Spreading or squeezing the coil turns farther apart or closer together, decreases or increases the inductance (and hence the coupling) accordingly.



A similar construction using pc board is shown for the "Citation Eliminator." Approximate dimensions are 2 X 2-3/4 X 2-3/4 (HWD). A top cover with hole for capacitor adjustment should be soldered on after filter is initially aligned. Power rating is suitable for transmitters in 75-watt input class.

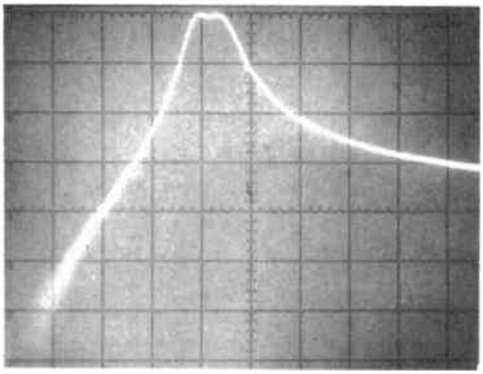


Fig. 4 – Response of the "Citation Eliminator." Attenuation and frequency scale 10 dB/div. (vertical) and 5 MHz/div. (horizontal).

An Absorptive Filter

The filter shown in Fig. 1 not only provides rejection by means of a low-pass section, it also includes circuitry that absorbs harmonic energy. A high-pass section consisting of L1, L2, C1 and C2 is terminated in a 50-ohm "idler load" and this combination performs the latter function. The advantages of this technique are that degradation of filter rejection caused by antenna mismatch at the harmonic frequency are not as severe (with a filter of this type) and the transmitter is terminated in a resistive load at the harmonic.

Construction and Test Techniques

If good performance above 100 MHz is not a necessity, this filter can be built using conventional fixed capacitors. Copper-clad Teflon board may not be readily available in small quantities from many supply houses. Regular fiberglass-insulated board is satisfactory for low power. One such filter has been used with an SB-100 transceiver running 100 watts. Although the *Q* of the fiberglass

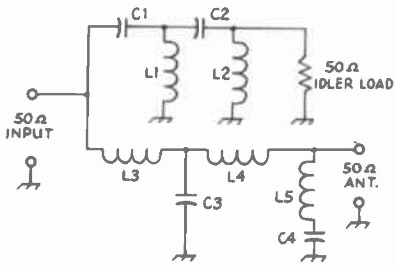


Fig. 1 – Schematic diagram of the absorptive filter. The pc board used is MIL-P-13949D, FL-GT-.062 in, C-2/2-11017, Class 1, Grade A. Polychem Bud Division. Capacitance between copper surfaces is 10-pF per square inch. Values are as follows for a design cutoff frequency of 40 MHz and rejection peak in Channel 2:

C1 – 52 pF	C4 – 21.6 pF	L3 – 0.3 µH
C2 – 73 pF	L1 – 0.125 µH	L4 – 0.212 µH
C3 – 126 pF	L2 – 0.52 µH	L5 – 0.24 µH

capacitors will be lower than that of Teflon-dielectric capacitors, this should not greatly affect the type of filter described here.

Test equipment needed to build this filter at home includes a reasonably accurate grid-dip oscillator, a *SWR* bridge, a reactance chart or the ARRL Lighting Calculator (for *L*, *C*, and *f*), a 50-ohm dummy load, and a transmitter.

Once the value of a given capacitor has been calculated, the next step is to determine the capacitance per square inch of the double-clad circuit board you have. This is done by connecting one end of a coil of known inductance to one side of the circuit board, and the other coil lead to the other side of the circuit board. Use the grid-dip oscillator, coupled lightly to the coil, to determine the resonant frequency of the coil and the circuit-board capacitor. When the frequency is known, the total capacitance can be determined by working the Lightning Calculator or by looking the capacitance up on a reactance chart. The total capacitance divided by the number of square inches on one side of the circuit board gives the capacitance per square inch. Once this figure is determined, capacitors of almost any value can be laid out with a ruler!

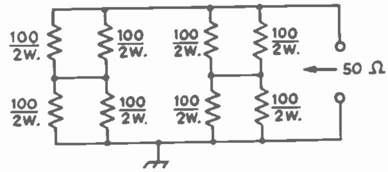


Fig. 3 — Dummy load for the high-pass section of the filter.

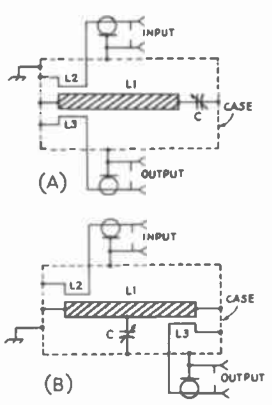
High voltages can be developed across capacitors in a series-tuned circuit, so the copper material should be trimmed back at least 1/8 inch from all edges of a board, except those that will be soldered to ground, to prevent arcing. This should not be accomplished by filing, since the copper filings would become imbedded in the board material and just compound the problem. The capacitor surfaces should be kept smooth and sharp corners should be avoided.

If the filter box is made of double-clad fiber glass board, both sides should be bonded together with copper stripped from another piece of board. Stripped copper foil may be cleaned with a razor blade before soldering. To remove copper foil from a board, use a straight edge and a sharp scribe to score the thin copper foil. When the copper foil has been cut, use a razor blade to lift a corner. Careful heating with a soldering iron will reduce the effort required to separate the copper from the board. This technique of bonding two pieces of board or two sides of a piece of board can also be used to interconnect two capacitors when construction in one plane would require too much area. Stray inductance must be minimized and sufficient clearance must be maintained for arc-over protection.

Capacitors with Teflon dielectric have been used in filters passing up to 2 kW PEP. One further word of caution: No low-pass filter will be fully effective until the transmitter with which it is used is properly shielded and all leads filtered.

Fig. 16-14 — Equivalent circuits for the strip-line filters. At A, the circuit for the 6- and 2-meter filters are shown. L2 and L3 are the input and output links. These filters are bilaterial, permitting interchanging of the input and output terminals.

At B, the representative circuit for the 220- and 432-MHz filters. These filters are also bilaterial.



The terminating loads for the high-pass section of the filter can be made from 2-watt, 10-percent tolerance composition resistors. Almost any dissipation rating can be obtained by suitable series-parallel combinations. For example, a 16-watt, 50-ohm load could be built as shown in Fig. 3. This load should handle the harmonic energy of a signal with peak fundamental power of 2 kilowatts. With this load, the harmonic energy will see a *SWR* under 2:1 up to 400 MHz. For low power (under 300 watts PEP), a pair of 2-watt 100-ohm resistors is adequate.

In the model shown the high-pass filter series capacitors are bonded and mounted on Teflon standoff insulators.

FILTERS FOR VHF TRANSMITTERS

High rejection of unwanted frequencies is possible with the tuned-line filters of Fig. 16-14. Examples are shown for each band from 50 through 450 MHz. Construction is relatively simple, and the cost is low. Standard boxes are used, for ease of duplication.

The filter of Fig. 16-15 is selective enough to pass 50-MHz energy and attenuate the 7th harmonic of an 8-MHz oscillator that falls in TV Channel 2. With an insertion loss at 50 MHz of about 1 dB, it can provide up to 40 dB of attenuation to energy at 57 MHz in the same line. This should be more than enough attenuation to take care of the worst situations, provided that the radiation is by way of the transmitter output coax only. The filter will not eliminate interfering energy that gets out from power cables, the ac line, or from the transmitter circuits themselves. It also will do nothing for TVI that results from deficiencies in the TV receiver.

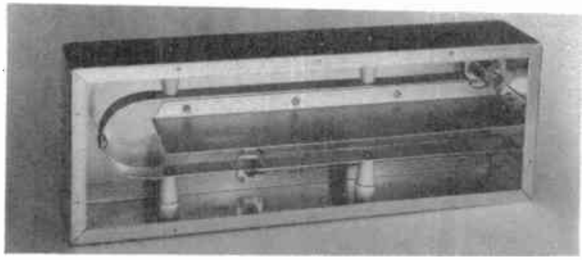
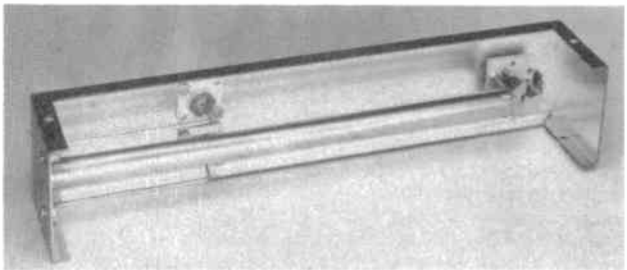


Fig. 16-15 – Interior of the 50-MHz strip line filter. Inner conductor of aluminum strip is bent into U shape, to fit inside a standard 17-inch chassis.

Fig. 16-16 – The 144-MHz filter has an inner conductor of 1/2-inch copper tubing 10 inches long, grounded to the left end of the case and supported at the right end by the tuning capacitor.



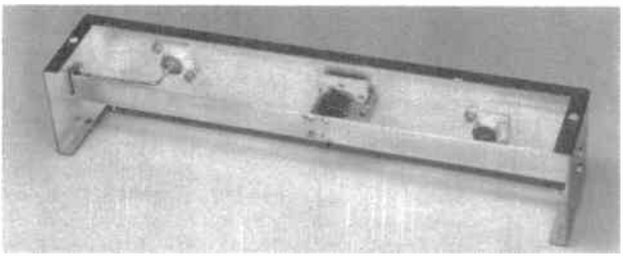


Fig. 16-17 – A half-wave strip line is used in the 220-MHz filter. It is grounded at both ends and tuned at the center.

The 50-MHz filter, Fig. 16-15, uses a folded line in order to keep it within the confines of a standard chassis. The case is a 6 X 17 X 3-inch chassis (Bud AC-433) with a cover plate that fastens in place with self-tapping screws. An aluminum partition down the middle of the assembly is 14 inches long, and the full height of the chassis, 3 inches.

The inner conductor of the line is 32 inches long and 13/16 inch wide, of 1/16-inch brass, copper or aluminum. This was made from two pieces of aluminum spliced together to provide the 32-inch length. Splicing seemed to have no ill effect on the circuit Q. The side of the "U" are 2 7/8 inches apart, with the partition at the center. The line is supported on ceramic standoffs. These were shimmed up with sections of hard wood or bakelite rod, to give the required 1 1/2-inch height.

The tuning capacitor is a double-spaced variable (Hammarlund HF-30-X) mounted 1 1/2 inches from the right end of the chassis. Input and output coupling loops are of No. 10 or 12 wire, 10 inches long. Spacing away from the line is adjusted to about 1/4 inch.

The 144-MHz model is housed in a 2 1/4 X 2 1/2 X 12-inch Minibox (Bud CU-2114-A).

One end of the tubing is slotted 1/4 inch deep with a hacksaw. This slot takes a brass angle bracket 1 1/2 inches wide, 1/4 inch high, with a 1/2-inch mounting lip. This 1/4-inch lip is soldered into the tubing slot, and the bracket is then bolted to the end of the box, so as to be centered on the end plate.

The tuning capacitor (Hammarlund HF-15-X) is mounted 1 1/4 inches from the other end of the box, in such a position that the inner conductor can be soldered to the two stator bars.

The two coaxial fittings (SO-239) are 11/16 inch in from each side of the box, 3 1/2 inches from the left end. The coupling loops are No. 12 wire, bent so that each is parallel to the center line of the inner conductor, and about 1/8 inch from its surface. Their cold ends are soldered to the brass mounting bracket.

The 220-MHz filter uses the same size box as the 144-MHz model. The inner conductor is 1/16-inch brass or copper, 5/8 inch wide, just long enough to fold over at each end for bolting to the box. It is positioned so that there will be 1/8 inch clearance between it and the rotor plates of the tuning capacitor. The latter is a Hammarlund HF-15-X, mounted slightly off-center in the box, so that its stator plates connect to the exact mid-point of the line. The 5/16-inch mounting hole in the case is 5 1/2 inches from one end. The SO-239 coaxial fittings are 1 inch in from opposite sides of the box, 2 inches from the ends. Their coupling links are No. 14 wire, 1/8 inch from the inner conductor of the line.

The 420-MHz filter is similar in design, using a 1 5/8 X 2 X 10-inch Minibox (Bud CU-2113-A). A half-wave line is used, with disk tuning at the center. The disks are 1/16-inch brass, 1 1/4-inch diameter. The fixed one is centered on the inner conductor, the other mounted on a No. 6 brass lead-screw. This passes through a threaded bushing, which can be taken from the end of a discarded slug-tuned form. An advantage of these is that usually a tension device is included. If there is none, use a lock nut.

Type N coaxial connectors were used on the 420-MHz model. They are 5/8 inch in from each side of the box, and 1 3/8 inches in from the ends. Their coupling links of No. 14 wire are 1/16 inch from the inner conductor.

Adjustment and Use

If you want the filter to work on both transmitting and receiving, connect the filter between antenna line and SWR indicator. With this arrangement you need merely adjust the filter for minimum reflected power reading on the SWR bridge. This should be zero, or close to it, if the antenna is well-matched. The bridge should be used, as there is no way to adjust the filter properly without it. If you insist on trying, adjust for best reception of signals on frequencies close to the ones you expect to transmit on. This works only if the antenna is well matched.

When the filter is properly adjusted (with the SWR bridge) you may find that reception can be improved by retuning the filter. Don't do it, if you want the filter to work best on the job it was intended to do: the rejection of unwanted energy, transmitting or receiving. If you want to improve reception with the filter in the circuit, work on the receiver input circuit. To get maximum power out of the transmitter and into the line, adjust the transmitter output coupling, not the filter. If the effect of the filter on reception bothers you, connect it in the line from the antenna relay to the transmitter only.

SUMMARY

The methods of harmonic elimination outlined here have been proved beyond doubt to be effective even under highly unfavorable conditions. It must be emphasized once more, however, that the problem must be solved one step at a time, and the procedure must be in logical order. It cannot be done properly without two items of simple equipment: a grid-dip meter and wavemeter covering the TV bands, and a dummy antenna.

To summarize:

1) Take a critical look at the transmitter on the basis of the design considerations outlined under "Reducing Harmonic Generation."

2) Check all circuits, particularly those connected with the final amplifier, with the grid-dip meter to determine whether there are any resonances in the TV bands. If so, rearrange the circuits so the resonances are moved out of the critical frequency region.

3) Connect the transmitter to the dummy antenna and check with the wavemeter for the presence of harmonics on leads and around the transmitter enclosure. Seal off the weak spots in the shielding and filter the leads until the wavemeter shows no indication at any harmonic frequency.

4) At this stage, check for interference with a TV receiver. If there is interference, determine the cause by the methods described previously and apply the recommended remedies until the interference disappears.

5) When the transmitter is completely clean on the dummy antenna, connect it to the regular antenna and check for interference on the TV receiver. If the interference is not bad, a Transmatch or matching circuit installed as previously described should clear it up. Alternatively, a low-pass filter may be used. If neither the Transmatch nor filter makes any difference in the interference, the evidence is strong that the interference, at least in part, is being caused by receiver overloading because of the strong fundamental-frequency field about the TV antenna and receiver. A Transmatch and/or filter, installed as described above, will invariably make a difference in the intensity of the interference if the interference is caused by transmitter harmonics alone.

6) If there is still interference after installing the Transmatch and/or filter, and the evidence shows that it is probably caused by a harmonic, more attenuation is needed. A more elaborate filter may be necessary. However, it is well at this stage to assume that part of the interference may be caused by receiver overloading, and take steps to alleviate such a condition before trying highly-elaborate filters and traps on the transmitter.

HARMONICS BY RECTIFICATION

Even though the transmitter is completely free from harmonic output it is still possible for interference to occur because of harmonics

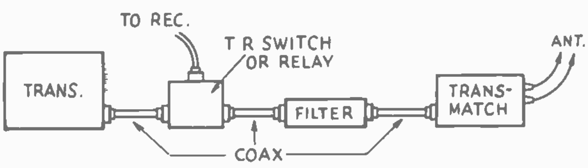


Fig. 16-18 – The proper method of installing a low-pass filter between the transmitter and a Transmatch. If the antenna is fed through coax, the Transmatch can be eliminated, but the transmitter and filter must be completely shielded. If a TR switch is used, it should be installed between the transmitter and low-pass filter. TR switches can generate harmonics themselves, so the low-pass filter should follow the TR switch.

generated outside the transmitter. These result from rectification of fundamental-frequency currents induced in conductors in the vicinity of the transmitting antenna. Rectification can take place at any point where two conductors are in poor electrical contact, a condition that frequently exists in plumbing, downspouting, BX cables crossing each other, and numerous other places in the ordinary residence. It also can occur at any exposed vacuum tubes in the station, in power supplies, speech equipment, etc., that may not be enclosed in the shielding about the rf circuits. Poor joints anywhere in the antenna system are especially bad, and rectification also may take place in the contacts of antenna changeover relays. Another common cause is overloading the front end of the communications receiver when it is used with a separate antenna (which will radiate the harmonics generated in the first tube) for break-in.

Rectification of this sort will not only cause harmonic interference but also is frequently responsible for cross-modulation effects. It can be detected in greater or less degree in most locations, but fortunately the harmonics thus generated are not usually of high amplitude. However, they can cause considerable interference in the immediate vicinity in fringe areas, especially when operation is in the 28-MHz band. The amplitude decreases rapidly with the order of the harmonic, the second and third being the worst. It is ordinarily found that even in cases where destructive interference results from 28-MHz operation the interference is comparatively mild from 14 MHz, and is negligible at still lower frequencies.

Nothing can be done at either the transmitter or receiver when rectification occurs. The remedy is to find the source and eliminate the poor contact either by separating the conductors or bonding them together. A crystal wavemeter (tuned to the fundamental frequency) is useful for hunting the source, by showing which conductors are carrying rf and, comparatively, how much.

Interference of this kind is frequently intermittent since the rectification efficiency will vary with vibration, the weather, and so on. The possibility of corroded contacts in the TV receiving antenna should not be overlooked, especially if it has been up a year or more.

TV RECEIVER DEFICIENCIES

When a television receiver is quite close to the transmitter, the intense rf signal from the transmitter's fundamental may overload one or more of the receiver circuits to produce spurious responses that cause interference.

If the overload is moderate, the interference is of the same nature as harmonic interference; it is caused by harmonics generated in the early stages of the receiver and, since it occurs only on channels harmonically related to the transmitting frequency, it is difficult to distinguish from harmonics actually radiated by the transmitter. In such cases additional harmonic suppression at the transmitter will do no good, but any means taken at the receiver to reduce the strength of the amateur signal reaching the first tube will effect an improvement. With very severe overloading, interference also will occur on channels *not* harmonically related to the transmitting frequency, so such cases are easily identified.

Cross-Modulation

Upon some circumstances overloading will result in cross-modulation or mixing of the amateur signal with that from a local fm or TV station. For example, a 14-MHz signal can mix with a 92-MHz fm station to produce a beat at 78 MHz and cause interference in Channel 5, or with a TV station on Channel 5 to cause interference in Channel 3. Neither of the channels interfered with is in harmonic relationship to 14 MHz. Both signals have to be on the air for the interference to occur, and eliminating either at the TV receiver will eliminate the interference.

There are many combinations of this type, depending on the band in use and the local frequency assignments to fm and TV stations. The interfering frequency is equal to the amateur fundamental frequency either added to or subtracted from the frequency of some local station, and when interference occurs in a TV channel that is not harmonically related to the amateur transmitting frequency the possibilities in such frequency combinations should be investigated.

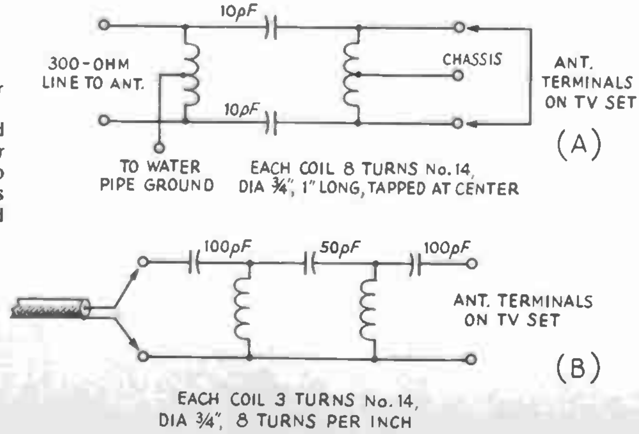


Fig. 16-19 — High-pass filters for installation at the TV receiver antenna terminals. A — balanced filter for 300-ohm line. B — for 75-ohm coaxial line. *Important:* Do not use a direct ground on the chassis of a transformerless receiver. Ground through a .001-μF mica capacitor.

I-f Interference

Some TV receivers do not have sufficient selectivity to prevent strong signals in the intermediate-frequency range from forcing their way through the front end and getting into the i-f amplifier. The once-standard intermediate frequency of, roughly, 21 to 27 MHz, is subject to interference from the fundamental-frequency output of transmitters operating in the 21-MHz band. Transmitters on 28 MHz sometimes will cause this type of interference as well.

A form of i-f interference peculiar to 50-MHz operation near the low edge of the band occurs with some receivers having the standard "41-MHz" i-f, which has the sound carrier at 41.25 MHz and the picture carrier at 45.75 MHz. A 50-MHz signal that forces its way into the i-f system of the receiver will beat with the i-f picture carrier to give a spurious signal on or near the i-f sound carrier, even though the interfering signal is not actually in the nominal passband of the i-f amplifier.

There is a type of i-f interference unique to the 144-MHz band in localities where certain uhf TV channels are in operation, affecting only those TV receivers in which double-conversion type plug-in uhf tuning strips are used. The design of these strips involves a first intermediate frequency that varies with the TV channel to be received and, depending on the particular strip design, this first i-f may be in or close to the 144-MHz amateur band. Since there is comparatively little selectivity in the TV signal-frequency circuits ahead of the first i-f, a signal from a 144-MHz transmitter will "ride into" the i-f, even when the receiver is at a considerable distance from the transmitter. The channels that can be affected by this type of i-f interference are:

Receivers with 21-MHz second i-f	*Receivers with 41-MHz second i-f*
Channels 14-18, incl.	Channels 20-25, incl.
Channels 41-48, incl.	Channels 51-58, incl.
Channels 69-77, incl.	Channels 82 and 83.

If the receiver is not close to the transmitter, a trap of the type shown in Fig. 16-21 will be effective. However, if the separation is small the 144-MHz signal will be picked up directly on the receiver circuits and the best solution is to readjust the strip oscillator so that the first i-f is moved to a frequency not in the vicinity of the 144-MHz band. This has to be done by a competent technician.

I-f interference is easily identified since it occurs on all channels – although sometimes the intensity varies from channel to channel – and the cross-hatch pattern it causes will rotate when the receiver's fine-tuning control is varied. When the interference is caused by a harmonic, overloading, or cross modulation, the structure of the interference pattern does not change (its intensity may change) as the fine-tuning control is varied.

High-Pass Filters

In all of the above cases the interference can be eliminated if the fundamental signal strength can be reduced to a level that the receiver can handle. To accomplish this with signals on bands below 30 MHz, the most satisfactory device is a high-pass filter having a cutoff frequency between 30 and 54 MHz, installed at the tuner input terminals of the receiver. Circuits that have proved effective are shown in Figs. 16-18 and 16-19. Fig. 16-18 has one more section than the filters of Fig. 16-19 and as a consequence has somewhat better cutoff characteristics. All the circuits given are designed to have little or no effect on the TV signals but will attenuate all signals lower in frequency than about 40 MHz. These filters preferably should be constructed in some sort of shielding container, although shielding is not always necessary. The dashed lines in Fig. 16-20 show how individual filter coils can be shielded from each other. The capacitors can be tubular ceramic units centered in holes in the partitions that separate the coils.

Simple high-pass filters cannot always be applied successfully in the case of 50-MHz transmissions, because they do not have sufficiently-sharp cutoff characteristics to give both good attenuation at 50-54 MHz and no attenuation above 54 MHz. A more elaborate design capable of giving the required sharp cutoff has been described (Ladd, "50-MHz TVI – Its Causes and Cures," *QST*, June and July, 1954). This article also contains other information useful in coping with the TVI problems peculiar to 50-MHz operation. As an alternative to such a filter, a high-Q wave trap tuned to the transmitting frequency may be used, suffering only the disadvantage that it is quite selective and therefore will protect a receiver from overloading over only a small range of transmitting frequencies in the 50-MHz band. A trap of this type is shown in Fig. 16-21. These "suck-out" traps, while absorbing energy at the frequency to which they are tuned, do not affect the receiver operation otherwise. The assembly should be mounted near the input terminals of the TV tuner and its case should be grounded to the TV set chassis. The traps should be tuned for minimum TVI at the transmitter operating frequency. An insulated tuning tool should be used for adjustment of the trimmer capacitors, since

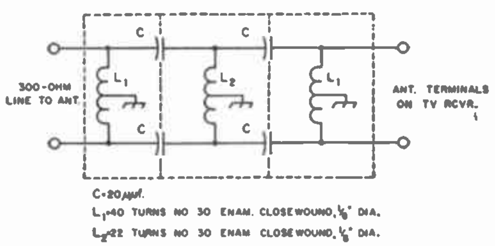


Fig. 16-20 – Another type of high-pass filter for 300-ohm line. The coils may be wound on 1/8-inch diameter plastic knitting needles. *Important:* Do not use a direct ground on the chassis of a transformerless receiver. Ground through a .001-μF

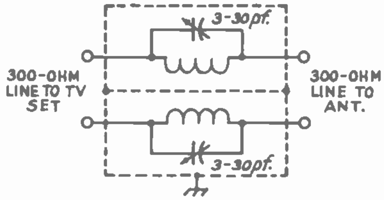


Fig. 16-21 – Parallel-tuned traps for installation in the 300-ohm line to the TV set. The traps should be mounted in an aluminum Minibox with a shield partition between them, as shown. For 50 MHz, the coils should have 9 turns of No. 16 enamel wire, close wound to a diameter of 1/2 inch. The 144-MHz traps should contain coils with a total of 6 turns of the same type wire, close-wound to a diameter of 1/4 inch. Traps of this type can be used to combat fundamental-overload TVI on the lower-frequency bands as well.

they are at a "hot" point and will show considerable body-capacitance effect.

High-pass filters are available commercially at moderate prices. In this connection, it should be understood by all parties concerned that while an amateur is responsible for *harmonic* radiation from his transmitter, it is no part of his responsibility to pay for or install filters, wave traps, etc. that may be required at the receiver to prevent interference caused by his *fundamental* frequency. Proper installation usually requires that the filter be installed right at the input terminals of the rf tuner of the TV set and not merely at the external antenna terminals, which may be at a considerable distance from the tuner. The question of cost is one to be settled between the set owner and the organization with which he deals. Don't overlook the possibility that the manufacturer of the TV receiver will supply a high-pass filter free of charge.

If the fundamental signal is getting into the receiver by way of the line cord a line filter such as those shown in Fig. 16-22 may help. To be most effective it should be installed inside the receiver chassis at the point where the cord enters, making the ground connections directly to the chassis at this point. It may not be so helpful if placed between the line plug and the wall socket unless the rf is actually picked up on the house wiring rather than on the line cord itself.

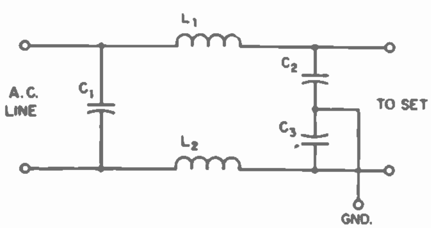


Fig. 16-22 – "Brute-force" ac line filter for receivers. The values of C1, C2 and C3 are not generally critical; capacitances from .001 to .01 μF can be used. L1 and L2 can be a 2-inch winding of No. 18 enameled wire on a half-inch diameter form. In making up such a unit for use external to the receiver, make sure that there are no exposed conductors to offer a shock hazard.

Antenna Installation

Usually, the transmission line between the TV receiver and the actual TV antenna will pick up a great deal more energy from a nearby transmitter than the television receiving antenna itself. The currents induced on the TV transmission line in this case are of the "parallel" type, where the phase of the current is the same in both conductors. The line simply acts like two wires connected together to operate as one. If the receiver's antenna input circuit were perfectly balanced it would reject these "parallel" or "unbalance" signals and respond only to the true transmission-line ("push-pull") currents; that is, only signals picked up on the actual antenna would cause a receiver response. However, no receiver is perfect in this respect, and many TV receivers will respond strongly to such parallel currents. The result is that the signals from a nearby amateur transmitter are much more intense at the first stage in the TV receiver than they would be if the receiver response were confined entirely to energy picked up on the TV antenna alone. This situation can be improved by using shielded transmission line – coax or, in the balanced form, "twinax" – for the receiving installation. For best results the line should terminate in a coax fitting on the receiver chassis, but if this is not possible the shield should be grounded to the chassis right at the antenna terminals.

The use of shielded transmission line for the receiver also will be helpful in reducing response to harmonics actually being radiated from the transmitter or transmitting antenna. In most receiving installations the transmission line is very much longer than the antenna itself, and is consequently far more exposed to the harmonic fields from the transmitter. Much of the harmonic pickup, therefore, is on the receiving transmission line when the transmitter and receiver are quite close together. Shielded line, plus relocation of either the transmitting or receiving antenna to take advantage of directive effects, often will result in reducing overloading, as well as harmonic pickup, to a level that does not interfere with reception.

UHF TELEVISION

Harmonic TVI in the uhf TV band is far less troublesome than in the vhf band. Harmonics from transmitters operating below 30 MHz are of such high order that they would normally be expected to be quite weak; in addition, the components, circuit conditions and construction of low-frequency transmitters are such as to tend to prevent very strong harmonics from being generated in this region. However, this is not true of amateur vhf transmitters, particularly those working in the 144-MHz and higher bands. Here the problem is quite similar to that of the low vhf TV band with respect to transmitters operating below 30 MHz.

There is one highly favorable factor in uhf TV that does not exist in the most of the vhf TV band: If harmonics are radiated, it is possible to move the transmitter frequency sufficiently (within the amateur band being used) to avoid interfering with a channel that may be in use in the locality. By restricting operation to a portion of the amateur band that will not result in harmonic interference, it is possible to avoid the necessity for taking extraordinary precautions to prevent harmonic radiation.

The frequency assignment for uhf television consists of seventy 6-Megahertz channels (Nos. 14 to 83, inclusive) beginning at 470 MHz and ending at 890 MHz. The harmonics from amateur bands above 50-MHz span the uhf channels as shown in Table 16-1. Since the assignment plan calls for a minimum separation of six channels between any two stations in one locality, there is ample opportunity to choose a fundamental frequency that will move a harmonic out of range of a local TV frequency.

TABLE 16-1

Harmonic Relationship – Amateur VHF Bands and UHF TV Channels

Amateur Band	*Harmonic*	*Fundamental Freq. Range*	*Channel Affected*
144 MHz	4th	144.0–144.5	31
		144.5–146.0	32
		146.0–147.5	33
		147.5–148.0	34
	5th	144.0–144.4	55
		144.4–145.6	56
		145.6–146.8	57
		146.8–148.0	58
	6th	144.0–144.33	79
		144.33–145.33	80
		145.33–147.33	81
		147.33–148.0	82
220 MHz	3rd	220–220.67	45
		220.67–222.67	46
		222.67–224.67	47
		224.67–225	48
	4th	220–221	82
		221–222.5	83
420 MHz	2nd	420–421	75
		421–424	76
		424–427	77
		427–430	78
		430–433	79
		433–436	80

COLOR TELEVISION

The color TV signal includes a subcarrier spaced 3.58 MHz from the regular picture carrier (or 4.83 MHz from the low edge of the channel) for transmitting the color information. Harmonics which fall in the color subcarrier region can be epected to cause break-up of color in the received picture. This modifies the chart of Fig. 16-2 to introduce another "severe" region centering around 4.8 MHz measured from the low-frequency edge of the channel. Hence with color television reception there is less opportunity to avoid harmonic interference by choice of operating frequency. In other respects the problem of eliminating interference is the same as with black-and-white television.

INTERFERENCE FROM TV RECEIVERS

The TV picture tube is swept horizontally by the electron beam 15,750 times per second, using a wave shape that has very high harmonic content. The harmonics are of appreciable amplitude even at frequencies as high as 30 MHz, and when radiated from the receiver can cause considerable interference to reception in the amateur bands. While measures to suppress radiation of this nature are required by FCC in current receivers, many older sets have had no such treatment. The interference takes the form of rather unstable, ac-modulated signals spaced at intervals of 15.75 kHz.

Studies have shown that the radiation takes place principally in three ways, in order of their importance: (1) from the ac line, through stray coupling to sweep circuits; (2) from the antenna system, through similar coupling; (3) directly from the picture tube and sweep-circuit wiring. Line radiation often can be reduced by bypassing the ac line cord to the chassis at the point of entry, although this is not completely effective in all cases since the coupling may take place outside the chassis beyond the point where the bypassing is done. Radiation from the antenna is usually suppressed by installing a high-pass filter on the receiver. The direct radiation requires shielding of high-potential leads and, in some receivers, additional bypassing in the sweep circuit; in severe cases, it may be necessary to line the cabinet with screening or similar shielding material.

Incidental radiation of this type from TV and broadcast receivers, when of sufficient intensity to cause serious interference to other radio services (such as amateur), is covered by Part 15 of the FCC rules. When such interference is caused, the user of the receiver is obligated to take steps to eliminate it. The owner of an offending receiver should be advised to contact the source from which the receiver was purchased for appropriate modification of the receiving installation. TV receiver dealers can obtain the necessary information from the set manufacturer.

It is usually possible to reduce interference very considerably, without modifying the TV receiver, simply by having a good amateur-band receiving installation. The principles are the same as those used in reducing "hash" and other noise – use a good antenna, such as the transmitting antenna, for reception; install it as far as possible from ac circuits; use a good feeder system such as a properly balanced two-wire line or coax with the outer conductor grounded; use coax input to the receiver, with a matching circuit if necessary; and check the receiver to make sure that it does not pick up signals or noise with the antenna disconnected.

HI-FI INTERFERENCE

Since the introduction of stereo and high-fidelity receivers, interference to this type of home-entertainment device has become a severe problem for amateurs. Aside from placing the amateur antenna as far as possible from any hi-fi installation, there is little else that can be done at the amateur's ham shack. Most of the hi-fi gear now being sold has little or no filtering to prevent rf interference. In other words, corrective measures must be done at hi-fi installation.

Hi-Fi Gear

Hi-fi gear can consist of a simple amplifier, with record or tape inputs, and speakers. The more elaborate installations may have a tape deck, record player, fm and a-m tuners, an amplifier, and two or more speakers. These units are usually connected together by means of shielded leads, and in most cases the speakers are positioned some distance from the amplifier, via long leads. When such a setup is operated near an amateur station, say within a few hundred feet, there are two important paths through which rf energy can reach the hi-fi installation to cause interference.

Step number one is to try to determine how the interference is getting into the hi-fi unit. If the volume control has no effect on the level of interference or very slight effect, the audio rectification of the amateur signal is taking place past the volume control, or on the output end of the amplifier. This is by far the most common type. It usually means that the amateur signal is being picked up on the speaker leads, or possibly on the ac line, and is then being fed back into the amplifier.

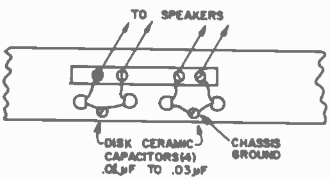


Fig. 16-23 — The disk capacitors should be mounted directly between the speaker terminals and chassis ground, keeping the leads as short as possible.

Experience has shown that most of the rf gets into the audio system via the speaker leads or the ac line, mostly the speaker leads. The amateur may find that on testing, the interference will only show up on one or two bands, or all of them. In hi-fi installations speakers are sometimes set up quite some distance from the amplifier. If the speaker leads happen to be resonant near an amateur band in use, there is likely to be an interference problem. The speaker lead will act as a resonant antenna and pick up the rf. One easy cure is to bypass the speaker terminals at the amplifier chassis. Use .01- to .03-μF disk capacitors from the speaker terminals directly to chassis ground; see Fig 16-23. Try .01 μF and see if that does the job. In some amplifiers .03 μF are required to eliminate the rf. Be sure to install bypasses on *all* the speaker terminals. In some instances, it may appear that one of each of the individual speaker terminals is grounded to the chassis. However, some amplifiers have the speaker leads above ground on the low side, for feedback purposes. If you have a circuit diagram of the amplifier you can check, but in the absence of a diagram, bypass all the terminals. If you can get into the amplifier, you can use the system shown in Fig. 16-24A.

In this system, two rf chokes are installed in series with the speaker leads from the output

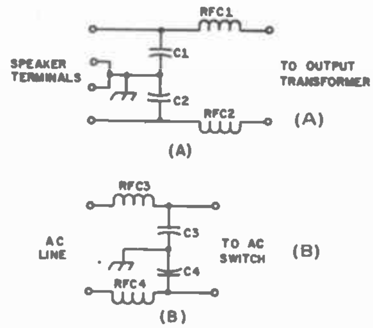


Fig. 16-24 — At A, the method for additional speaker filter, and at B, filtering the ac-line input. In both cases, these installations should be made directly inside the amplifier chassis, keeping the leads as short as possible.
C1,C2 — .01-to .03-μF disk ceramic.
C3, C4 — .01 disk ceramic, ac type.
RFC1 through RFC4 — 24 turns No. 18 enamel-covered wire, close-spaced and wound on a 1/4-inch diameter form (such as a pencil).

transformers, or amplifier output, to the speakers. These chokes are simple to make and help keep rf out of the amplifier. In particularly stubborn cases, shielded wire can be used for the speaker leads, grounding the shields at the amplifier chassis, and still using the bypasses on the terminals. When grounding, all chassis used in the hi-fi installation should be bonded together and connected to a good earth ground (such as a water pipe) if at all possible. It has been found that grounding sometimes eliminates the interference. On the other hand, don't be discouraged if grounding doesn't appear to help. Even with the bypassing and filtering grounding may make the difference.

Fig. 16-24B shows the method for filtering the ac line at the input of the amplifier chassis. The choke dimensions are the same as those given in Fig. 16-24A. Be sure that the bypasses are rated for ac because the dc types have been known to short out.

Antenna Pickup

If the hi-fi setup includes an fm installation, and many of them do, there is the possibility of rf getting into the audio equipment by way of the fm antenna. Chances for this method of entry are very good and precautions should be taken here to prevent the rf from getting to the equipment. A TV-type high-pass filter can prove effective in some cases.

Turntables and Tape Decks

In the more elaborate hi-fi setups, there may be several assemblies connected together by means of patch cords. It is a good idea when checking for RFI to disconnect the units, one at a time, observing any changes in the interference. Not only disconnect the patch cords connecting the pieces together, but also unplug the ac line cord for each item as you make the test. This will help you determine which section is the culprit.

Patch cords are usually, *but not always*, made of shielded cable. The lines *should* be shielded, which brings up another point. Many commercially available patch cords have poor shields. Some have wire spirally wrapped around the insulation, covering the main lead, rather than braid. This method provides poor shielding and could be the reason for RFI problems.

Record-player tone-arm connections to the cartridge are usually made with small clips. The existence of a loose clip, particularly if oxidation is present, offers an excellent invitation to RFI. Also, the leads from the cartridge and those to the amplifier are sometimes resonant at vhf, providing an excellent receiving antenna for rf. One cure for unwanted rf pickup is to install ferrite beads, one on each cartridge lead. Check all patch-cord connections for looseness or poor solder joints. Inferior connections can cause rectification and subsequent RFI.

Tape decks should be treated the same as turntables. Loose connections and bad solder joints all can cause trouble. Ferrite beads can be slipped over the leads to the recording and play-back pickup heads. Bypassing of the tone-arm or pickup-head leads is also effective, but sometimes it is difficult to install capacitors in the small area available. Disk capacitors (.001μF) should be used as close to the cartridge or pickup head as possible. Keep the capacitor leads as short as possible.

Preamplifiers

There are usually one or more preamplifiers used in a hi-fi amplifier. The inputs to these stages can be very susceptible to RFI. Fig. 16-24 illustrates a typical preamplifier circuit. In this case the leads to the bases of the transistors are treated for RFI with ferrite beads by the addition of RFC2 and RFC4. This is a very effective method for stopping RFI when vhf energy is the source of the trouble.

Within the circuit of a solid-state audio system, a common offender can be the emitter-base junction of a transistor. This junction operates as a forward-biased diode, with the bias set so that a change of base current with signal will produce a linear but amplified change in collector current. Should rf energy reach the junction, the bias could increase, causing nonlinear amplification and distortion as the result. If the rf level is high it can

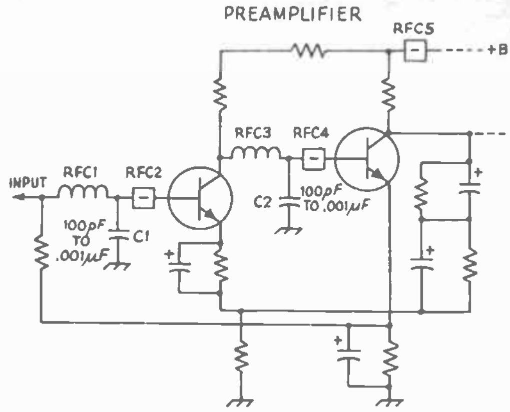


Fig. 16-25 — Typical circuit of a solid-state preamplifier.

completely block (saturate) a transistor, causing a complete loss of gain. Therefore, it may be necessary to reduce the transmitter power output in order to pinpoint the particular transistor stage that is affected.

In addition to adding ferrite beads it may be necessary to bypass the base of the transistor to chassis ground, C1 and C2, Fig. 16-25. A suitable value is 100 pF, and keep the leads short! As a general rule, the capacitor value should be as large as possible without degrading the high-frequency response of the amplifier. Values up to .001μF can be used. In severe cases, a series inductor (RFC1 and RFC3) may be required, Ohmite Z-50 or Z-144, or their equivalents (7 and 1.8 μH respectively). Fig. 16-25 shows the correct placement for an inductor, bypass capacitor, and ferrite bead. Also, it might help to use a ferrite bead in the plus-B lead to the preamplifier stages (RFC5 in Fig. 16-25). Keep in mind that Fig. 16-25 represents only one preamplifier of a stereo set. *Both* channels may require treatment.

FM Tuners

There is often an fm tuner used in a hi-fi installation. Much of the interference to tuners is caused by fundamental overloading of the first stage (or stages) of the tuner, effected by the amateur's signal. The cure is the installation of a high-pass filter, the same type used for TVI. The filter should be installed as close as possible to the antenna input of the tuner. The high-pass filter will attenuate the amateur *fundamental* signal, thus preventing overloading of the front end.

Shielding

Lack of shielding on the various components in a hi-fi installation can permit rf to get into the equipment. Many units have no bottom plates, or are installed in plastic cases. One easy method of providing shielding is to use aluminum foil. Make sure the foil doesn't short circuit the components, and connect it to chassis ground.

INTERFERENCE WITH STANDARD BROADCASTING

Interference with a-m broadcasting usually falls into one or more rather well-defined categories. An understanding of the general types of interference will avoid much cut-and-try in finding a cure.

Transmitter Defects

Out-of-band radiation is something that must be cured at the transmitter. Parasitic oscillations are a frequently unsuspected source of such radiations, and no transmitter can be considered satisfactory until it has been thoroughly checked for both low- and high-frequency parasitics. Very often parasitics show up only as transients, causing key clicks in cw transmitters and "splashes" or "burps" on modulation peaks in a-m transmitters. Methods for detecting and eliminating parasitics are discussed in the transmitter chapter.

In cw transmitters the sharp make and break that occurs with unfiltered keying causes transients that, in theory, contain frequency components through the entire radio spectrum. Practically, they are often strong enough in the immediate vicinity of the transmitter to cause serious interference to broadcast reception. Key clicks can be eliminated by the methods detailed in the chapter on keying.

BCI is frequently made worse by radiation from the power wiring or the rf transmission line. This is because the signal causing the interference, in such cases, is radiated from wiring that is nearer the broadcast receiver than the antenna itself. Much depends on the method used to couple the transmitter to the antenna, a subject that is discussed in the chapters on transmission lines and antennas. If it is at all possible the antenna itself should be placed so that it is not in close proximity to house wiring, telephone and power lines, and similar conductors.

The BC Set

Most present day receivers use solid-state active components, rather than tubes. A large number of the receivers in use are battery powered. This is to the amateur's advantage because much of the bc interference an amateur encounters is because of ac line pickup. In the case where the bc receiver is powered from the ac line, whether using tube or solid-stage components, the amount of rf pickup must be reduced or eliminated. A line filter such as is shown in Fig. 16-22 often will help accomplish this. The values used for the coils and capacitors are in general not critical. The effectiveness of the filter may depend considerably on the ground connection used, and it is advisable to use a short ground lead to a cold-water pipe if at all possible. The line cord from the set should be bunched up, to minimize the possibility of pick-up on the cord. It may be necessary to install the filter inside the receiver, so that the filter is connected between the line cord and the set wiring, in order to get satisfactory operation.

Cross-Modulation

With phone transmitters, there are occasionally cases where the voice is heard whenever the broadcast receiver is tuned to a bc station, but there is no interference when tuning between stations. This is cross-modulation, a result of rectification in one of the early stages of the receiver. Receivers that are susceptible to this trouble usually also get a similar type of interference from regular broadcasting if there is a strong local bc station and the receiver is tuned to some *other* station.

The remedy for cross modulation in the receiver is the same as for images and oscillator-harmonic response – reduce the strength of the amateur signal at the receiver by means of a line filter.

The trouble is not always in the receiver, since cross modulation can occur in any nearby rectifying circuit – such as a poor contact in water or steam piping, gutter pipes, and other conductors in the strong field of the transmitting antenna – external to both receiver and transmitter. Locating the cause may be difficult, and is best attempted with a battery-operated portable broadcast receiver used as a "probe" to find the spot where the interference is most intense. When such a spot is located, inspection of the metal structures in the vicinity should indicate the cause. The remedy is to make a good electrical bond between the two conductors having the poor contact.

Handling BCI Cases

Assuming that your transmitter has been checked and found to be free from spurious radiations, get another amateur to operate your station, if possible, while you make the actual check on the interference yourself. The following procedure should be used.

Tune the receiver through the broadcast band, to see whether the interference tunes like a regular bc station. If so, image or oscillator-harmonic response is the cause. If there is interference only when a bc station is tuned in, but not between stations, the cause is cross modulation. If the interference is heard at all settings of the tuning dial, the trouble is pickup in the audio circuits. In the latter case, the receiver's volume control may or may not affect the strength of the interference, depending on the means by which your signal is being rectified.

Having identified the cause, explain it to the set owner. It is a good idea to have a line filter with you, equipped with enough cord to replace the set's line cord, so it can be tried then and there. If it does not eliminate the interference, explain to the set owner that there is nothing further that can be done without modifying the receiver. Recommend that the work be done by a competent service technician, and offer to advise the service man on the cause and remedy. Don't offer to work on the set yourself, but if you are asked to do so use your own judgment about complying; set owners sometimes complain about the overall performance of the receiver afterward, often without justification. If you work on it, take it to your station so the effect of changes you make can be seen. Return the receiver promptly when you have finished.

MISCELLANEOUS TYPE OF INTERFERENCE

The operation of amateur phone transmitters occasionally results in interference on telephone lines and in audio amplifiers used in public-address work, plus other audio devices. The cause is rectification of the signal in an audio circuit.

Organs

An RFI problem area is the electronic organ. All of the techniques outlined for hi-fi gear hold true in getting rid of RFI in an organ. Two points should be checked – the speaker leads and the ac line. Many organ manufacturers have special servicemen's guides for taking care of RFI. However, to get this information you or the organ owner must contact the manufacturer, not the dealer or distributor. Don't accept the statement from a dealer or serviceman that there is nothing that can be done about the interference.

P-A Systems

The cure for RFI in p-a systems is almost the same as that for hi-fi gear. The one thing to watch for is rf on the leads that connect the various stations in a p-a system together. These leads should be treated the same as speaker *leads* and bypassing and filtering should be done at *both* ends of the lines. Also, watch for ac-line pickup of rf.

Telephone Interference

Telephone interference may be cured by connecting a bypass capacitor (about .001 μF) across the microphone unit in the telephone handset. The telephone companies have capacitors for this purpose. When such a case occurs, get in touch with the repair department of the phone company, giving the particulars. Section 500-150-100 of the Bell System Practices *Plant Series* gives detailed instructions. This section discusses causes and cures of telephone interference from radio signals. It points out that interference can come from corroded connections, unterminated loops, and other sources. It correctly points out that that rf can be picked up on the drop wire coming into the house, and also on the wiring within the house, but (usually) the detection of the rf occurs inside the phone. The detection usually takes place at the varistors in the compensation networks, and/or at the receiver noise suppressor and the carbon microphone. But interference suppression should be handled two ways: prevent the rf from getting to the phone, and prevent it from being rectified.

The telephone companies (Bell System) have two devices for this purpose. The first is a 40BA capacitor, which is installed at the service entrance protector, and the second is the 1542A inductor, which is installed at the connector block. According to the practices manual, the 40BA bypasses rf picked up on the drop wire coming into the house from the phone, and the 1542A suppresses rf picked up on the inside wiring. These are mentioned because in very stubborn cases they may be necessary. But first, it is suggested that the telephones be modified.

Since there are several different series of phones, they will be discussed separately:

500 series – These are the desk and wall phones most commonly in use. They come in several different configurations, but all use a 425-series compensation network. The letter designation can be A, B, C, D, E, F, G, or K, and all these networks contain varistors. The network should be replaced with a 425J, in which the varistors are replaced by resistors. Also, .01-μF disk-ceramic capacitors should be placed across the receiver suppressor. The suppressor is a diode across the receiver terminals. The carbon microphone in the handset should be bypassed with a .01-μF ceramic capacitor.

Series 1500, 1600, 1700 – These are the "Touch-Tone" phones, and the cure is similar to that for the 500 series, except that the network is a 4010B or D, and should be replaced with a 4010E.

Trimline series – These are the "Princess" series phones. The practice manual says that these should be modified by installing bypass capacitors across *all* components in the set that may act as demodulators. This statement is rather vague, but evidently a solution is known to the telephone company for these sets.

At the end of section 500-150-100 is an ordering guide for special components and sets, as follows:

Ordering Guide:
Capacitor, 40BA
Inductor, 1542A
-49 Gray, -50 Ivory
Set, Telephone, -rf Modified
Set, Telephone Hand, 220A, -rf Modified
Set, Telephone Hand, 2220B, -rf Modified
Set, Hand G, -rf Modified
Dial – (Touch-Tone dial only) -rf Modified.

The type "G" Handset is the one used with the 500 and Touch-Tone series phones. Also, Mountain Bell has put out an "Addendum 500-150-100MS, Issue A, January 1971" to the practices manual, which states that items for rf modified phones should be ordered on nonstock Form 3218, as follows:

(Telephone Set type)
Modified for BSP 500-150-100
for Radio Signal Suppression

The FCC

The Field Engineering Bureau of the FCC has a bulletin that will be of help to the amateur in cases involving RFI to audio devices. These bulletins are available from any of the field offices. The bulletin is addressed to the users of hi-fi, record players, public-address systems, and telephones. It clearly spells out the problem and the obligation of the owner of such gear.

It is suggested that the amateur obtain copies of this bulletin, which is listed as *Attachment III, Bulletin, Interference to Audio Devices.* When the amateur receives a complaint he can provide the complainer with a copy of the bulletin. This approach will help put the problem in correct perspective.

Test Equipment and Measurements

Measurement and testing seemingly go hand in hand, but it is useful to make a distinction between "measuring" and "test" equipment. The former is commonly considered to be capable of giving a meaningful quantitative result. For the latter a simple indication of "satisfactory" or "unsatisfactory" may suffice; in any event, the accurate calibration associated with real measuring equipment is seldom necessary, for simple test apparatus.

Certain items of measuring equipment that are useful to amateurs are readily available in kit form, at prices that represent a genuine saving over the cost of identical parts. Included are volt-ohm-milliammeter combinations, vacuum-tube and transistor voltmeters, oscilloscopes, and the like. The coordination of electrical and mechanical design, components, and appearance make it far preferable to purchase such equipment than to attempt to build one's own.

However, some test gear is either not available or can easily be built. This chapter considers the principles of the more useful types of measuring equipment and concludes with the descriptions of several pieces that not only can be built satisfactorily at home but which will facilitate the operation of the amateur station.

THE DIRECT-CURRENT INSTRUMENT

In measuring instruments and test equipment suitable for amateur purposes the ultimate "readout" is generally based on a measurement of direct current. A meter for measuring dc uses electromagnetic means to deflect a pointer over a calibrated scale in proportion to the current flowing through the instrument.

In the D'Arsonval type a coil of wire, to which the pointer is attached, is pivoted between the poles of a permanent magnet, and when current flows through the coil it sets up a magnetic field that interacts with the field of the magnet to cause the coil to turn. The design of the instrument is usually such as to make the pointer deflection directly proportional to the current.

A less expensive type of instrument is the moving-vane type, in which a pivoted soft-iron vane is pulled into a coil of wire by the magnetic field set up when current flows through the coil. The farther the vane extends into the coil the greater the magnetic pull on it, for a given change in current, so this type of instrument does not have "linear" deflection – the intervals of equal current are crowded together at the low-current end and spread out at the high-current end of the scale.

Current Ranges

The sensitivity of an instrument is usually expressed in terms of the current required for full-scale deflection of the pointer. Although a very wide variety of ranges is available, the meters of interest in amateur work have basic "movements" that will give maximum deflection with currents measured in microamperes or milliamperes. They are called microammeters and milliammeters, respectively.

Thanks to the relationships between current, voltage, and resistance expressed by Ohm's Law, it becomes possible to use a single low-range instrument – e.g., 1 milliampere or less full-scale pointer deflection – for a variety of direct-current measurements. Through its ability to measure current, the instrument can also be used indirectly to measure voltage. Likewise, a measurement of *both* current and voltage will obviously yield a value of resistance. These measurement functions are often combined in a single instrument – the volt-ohm-milliammeter or "VOM", a multirange meter that is one of the most useful pieces of measuring and test equipment an amateur can possess.

Accuracy

The accuracy of a dc meter of the D'Arsonval type is specified by the manufacturer. A common specification is "2 percent of full scale," meaning that a 0-100 microammeter, for example, will be correct to within 2 microamperes at any part of the scale. There are very few cases in amateur work where accuracy greater than this is needed. However, when the instrument is part of a more complex measuring circuit, the design and components of which *all* can cause error, the *overall* accuracy of the complete device is always less.

EXTENDING THE CURRENT RANGE

Because of the way current divides between two resistances in parallel, it is possible to increase the range (more specifically, to decrease the sensitivity) of a dc micro- or milliammeter to any desired extent. The meter itself has an inherent resistance – its internal resistance – which determines the full-scale current through it when its rated voltage is applied. (This rated voltage is of the order of a few millivolts.) By connecting an

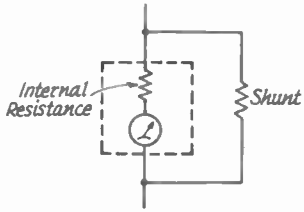


Fig. 17-1 — Use of a shunt to extend the calibration range of a current-reading instrument.

external resistance in parallel with the internal resistance, as in Fig. 17-1, the current will divide between the two, with the meter responding only to that part of the current which flows through the internal resistance of its movement. Thus it reads only part of the total current; the effect is to make more total current necessary for a full-scale meter reading. The added resistance is called a shunt.

It is necessary to know the meter's internal resistance before the required value for a shunt can be calculated. It may vary from a few ohms to a few hundred, with the higher resistance values associated with higher sensitivity. When known, it can be used in the formula below to determine the required shunt for a given current multiplication:

$$R = \frac{R_m}{n - 1}$$

where R is the shunt, R_m is the internal resistance of the meter, and n is the factor by which the original meter scale is to be multiplied.

Making Shunts

Homemade shunts can be constructed from any of various special kinds of resistance wire, or from ordinary copper wire if no resistance wire is available. The Copper Wire Table in this *Handbook* gives the resistance per 1000 feet for various sizes of copper wire. After computing the resistance required, determine the smallest wire size that will carry the full-scale current (250 circular mils per ampere is a satisfactory figure for this purpose). Measure off enough wire to provide the required resistance.

THE VOLTMETER

If a large resistance is connected in *series* with a current-reading meter, as in Fig. 17-2, the current

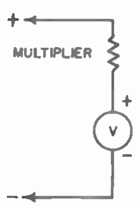


Fig. 17-2 — A voltmeter is a current-indicating instrument in series with a high resistance, the "multiplier."

multiplied by the resistance will be the voltage drop across the resistance, which is known as a multiplier. An instrument used in this way is calibrated in terms of the voltage drop across the multiplier resistor, and is called a voltmeter.

Sensitivity

Voltmeter sensitivity is usually expressed in ohms per volt, meaning that the meter's *full-scale* reading multiplied by the sensitivity will give the total resistance of the voltmeter. For example, the resistance of a 1000-ohms-per-volt voltmeter is 1000 times the full-scale calibration voltage, and by Ohm's Law the current required for full-scale deflection is 1 milliampere. A sensitivity of 20,000 ohms per volt, a commonly used value, means that the instrument is a 50-microampere meter.

The higher the resistance of the voltmeter the more accurate the measurements in high-resistance circuits. This is because in such a circuit the current flowing through the voltmeter will cause a change in the voltage between the points across which the meter is connected, compared with the voltage with the meter absent, as shown in Fig. 17-3.

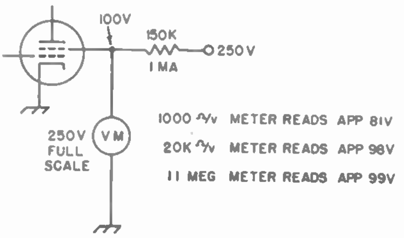


Fig. 17-3 — Effect of voltmeter resistance on accuracy of readings. It is assumed that the dc resistance of the screen circuit is constant at 100 kilohms. The actual current and voltage without the voltmeter connected are 1 mA and 100 volts. The voltmeter readings will differ because the different types of meters draw different amounts of current through the 150-kilohm resistor.

Multipliers

The required multiplier resistance is found by dividing the desired full-scale voltage by the current, in amperes, required for full-scale deflection of the meter alone. Strictly, the internal resistance of the meter should be subtracted from the value so found, but this is seldom necessary (except perhaps for very low ranges) because the meter resistance will be negligibly small compared with the multiplier resistance. An exception is when the instrument is already a voltmeter and is provided with an internal multiplier, in which case the multiplier resistance required to extend the range is

$$R = R_m(n - 1)$$

where R is the multiplier resistance, R_m is the total resistance of the instrument itself, and n is the factor by which the scale is to be multiplied. For example, if a 1000-ohms-per-volt voltmeter having a calibrated range of 0-10 volts is to be extended to 1000 volts, R_m is 1000 X 10 = 10,000 ohms, n is 1000/10 = 100, and R = 10,000 (100 – 1) = 990,000 ohms.

When extending the range of a voltmeter or converting a low-range meter into a voltmeter, the rated accuracy of the instrument is retained only when the multiplier resistance is precise. Precision wire-wound resistors are used in the multipliers of high-quality instruments. These are relatively expensive, but the home constructor can do quite well with 1-percent-tolerance composition resistors. They should be "derated" when used for this purpose – that is, the actual power dissipated in the resistor should not be more than 1/4 to 1/2 the rated dissipation – and care should be used to avoid overheating the body of the resistor when soldering to the leads. These precautions will help prevent permanent change in the resistance of the unit.

Ordinary composition resistors are generally furnished in 10- or 5-percent tolerance ratings. If possible errors of this order can be accepted, resistors of this type may be used as multipliers. They should be operated below the rated power dissipation figure, in the interests of long-time stability.

DC MEASUREMENT CIRCUITS

Current Measurement with a Voltmeter

A current-measuring instrument should have very low resistance compared with the resistance of the circuit being measured; otherwise, inserting the instrument will cause the current to differ from its value with the instrument out of the circuit. (This may not matter if the instrument is left permanently in the circuit.) However, the resistance of many circuits in radio equipment is quite high and the circuit operation is affected little, if at all, by adding as much as a few hundred ohms in series. In such cases the voltmeter method of measuring current, shown in Fig. 17-4, is frequently convenient. A voltmeter (or low-range milliammeter provided with a multiplier and operating as a voltmeter) having a full-scale voltage range of a few volts is used to measure the voltage drop across a suitable value of resistance acting as a shunt.

The value of shunt resistance must be calculated from the known or estimated maximum current expected in the circuit (allowing a safe margin) and the voltage required for full-scale deflection of the meter with its multiplier.

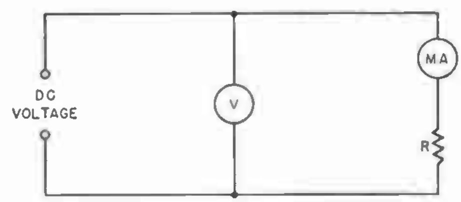


Fig. 17-5 – Measurement of power requires both current and voltage measurements; once these values are known the power is equal to the product – P = EI. The same circuit can be used for measurement of an unknown resistance.

Power

Power in direct-current circuits is determined by measuring the current and voltage. When these are known, the power is equal to the voltage in volts multiplied by the current in amperes. If the current is measured with a milliammeter, the reading of the instrument must be divided by 1000 to convert it to amperes.

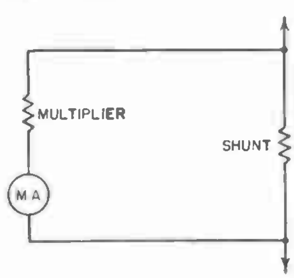


Fig. 17-4 – Voltmeter method of measuring current. This method permits using relatively large values of resistance in the shunt, standard values of fixed resistors frequently being usable. If the multiplier resistance is 20 (or more) times the shunt resistance, the error in assuming that all the current flows through the shunt will not be of consequence in most practical applications.

The setup for measuring power is shown in Fig. 17-5, where R is any dc "load," not necessarily an actual resistor.

Resistance

Obviously, if both voltage and current are measured in a circuit such as that in Fig. 17-5 the value of resistance R (in case it is unknown) can be calculated from Ohm's Law. For accurate results, the internal resistance of the ammeter or milliammeter, MA, should be very low compared with the resistance, R, being measured, since the voltage read by the voltmeter, V, is the voltage across MA and R in series. The instruments and the dc voltage should be chosen so that the readings are in the upper half of the scale, if possible, since the percentage error is less in this region.

THE OHMMETER

Although Fig. 17-5 suffices for occasional resistance measurements, it is inconvenient when frequent measurements over a wide range of resistance are to be made. The device generally used for this purpose is the ohmmeter. This consists fundamentally of a voltmeter (or milliammeter, depending on the circuit used) and a small dry battery, the meter being calibrated so the value of an unknown resistance can be read

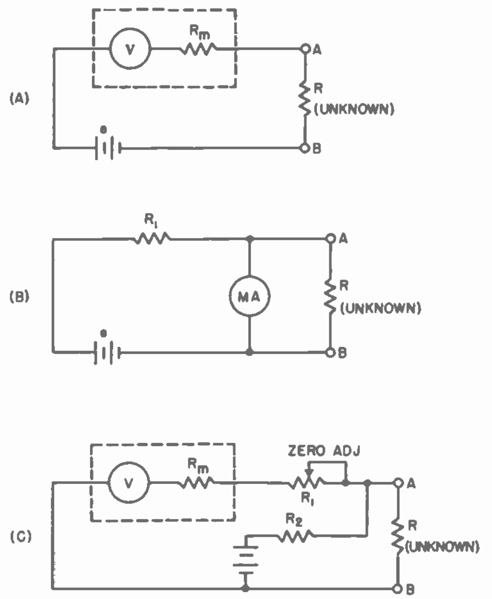


Fig. 17-6 – Ohmmeter circuits. Values are discussed in the text.

directly from the scale. Typical ohmmeter circuits are shown in Fig. 17-6. In the simplest type, shown in Fig. 17-6A, the meter and battery are connected in series with the unknown resistance. If a given deflection is obtained with terminals *A-B* shorted, inserting the resistance to be measured will cause the meter reading to decrease. When the resistance of the voltmeter is known, the following formula can be applied:

$$R = \frac{eR_m}{E} - R_m$$

where R is the resistance to be found,
e is the voltage applied (A-B shorted),
E is the voltmeter reading with R connected, and
R_m is the resistance of the voltmeter.

The circuit of Fig. 17-6A is not suited to measuring low values of resistance (below a hundred ohms or so) with a high-resistance voltmeter. For such measurements the circuit of Fig. 17-6B can be used. The unknown resistance is

$$R = \frac{I_2 R_m}{I_1 - I_2}$$

where R is the unknown,
R_m is the internal resistance of the milliammeter,
I_1 is the current with R disconnected from terminals A-B, and
I_2 is the current with R connected.

The formula is based on the assumption that the current in the complete circuit will be essentially constant whether or not the "unknown" terminals are short-circuited. This requires that R1 be very large compared with R_m – e.g., 3000 ohms for a 1-mA meter having an internal resistance of perhaps 50 ohms. A 3-volt battery would be necessary in this case in order to obtain a full-scale deflection with the "unknown" terminals open. R1 can be an adjustable resistor, to permit setting the open-terminals current to exact full scale.

A third circuit for measuring resistance is shown in Fig. 17-6C. In this case a high-resistance voltmeter is used to measure the voltage drop across a reference resistor, R2, when the unknown resistor is connected so that current flows through it, R2 and the battery in series. By suitable choice of R2 (low values for low-resistance, high values for high-resistance unknowns) this circuit will give equally good results on all resistance values in the range from one ohm to several megohms, provided that the voltmeter resistance, R_m, is always very high (50 times or more) compared with the resistance of R2. A 20,000-ohm-per-volt instrument (50-μA movement) is generally used. Assuming that the current through the voltmeter is negligible compared with the current through R2, the formula for the unknown is

$$R = \frac{eR2}{E} - R2$$

where R and $R2$ are as shown in Fig. 17-6C,
e is the voltmeter reading with A-B shorted, and
E is the voltmeter reading with R connected.

The "zero adjuster," R_1, is used to set the voltmeter reading exactly to full scale when the meter is calibrated in ohms. A 10,000-ohm variable resistor is suitable with a 20,000-ohms-per-volt meter. The battery voltage is usually 3 volts for ranges up to 100,000 ohms or so and 6 volts for higher ranges.

BRIDGE CIRCUITS

An important class of measurement circuits is the bridge, in which, essentially, a desired result is obtained by balancing the voltages at two different points in the circuit against each other so that there is zero potential difference between them. A voltmeter bridged between the two points will read zero (null) when this balance exists, but will indicate some definite value of voltage when the bridge is not balanced.

Bridge circuits are useful both on direct current and on ac of all frequencies. The majority of amateur applications is at radio frequencies, as shown later in this chapter. However, the principles of bridge operation are most easily introduced in terms of dc, where the bridge takes its simplest form.

The Wheatstone Bridge

The simple resistance bridge, known as the Wheatstone bridge, is shown in Fig. 17-7. All other bridge circuits – some of which are rather elaborate, especially those designed for ac – derive from this. The four resistors, R1, R2, R3, and R4 shown in A, are known as the bridge arms. For the

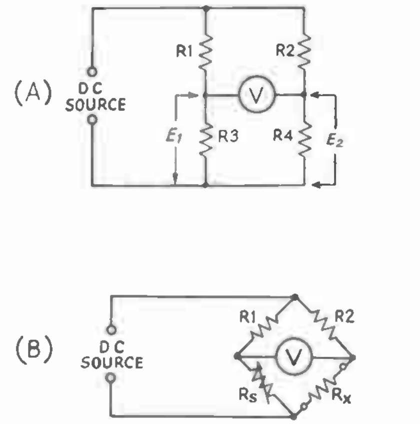


Fig. 17-7 — The Wheatstone bridge circuit. It is frequently drawn as at (B) for emphasizing its special function.

voltmeter reading to be zero, the voltages across R3 and R4 in series must add algebraically to zero; that is E1 must equal E2. R1R3 and R2R4 form voltage dividers across the dc source, so that if

$$\frac{R3}{R1 + R3} = \frac{R4}{R2 + R4}$$

E1 will equal E2.

The circuit is customarily drawn as shown at 17-7B when used for resistance measurement. The equation above can be rewritten

$$R_x = R_s \frac{R2}{R1}$$

to find R_x, the unknown resistance. R1 and R2 are frequently made equal; then the calibrated adjustable resistance (the standard), R_s, will have the same value as R_x when R_s is set to show a null on the voltmeter.

Note that the resistance *ratios*, rather than the actual resistance values, determine the voltage balance. However, the values do have important practical effects on the sensitivity and power consumption. The bridge sensitivity is the readiness with which the meter responds to small amounts of unbalance about the null point; the "sharper" the null the more accurate the setting of R_s at balance.

The Wheatstone bridge is rarely used by amateurs for resistance measurement, the ohmmeter being the favorite instrument for that purpose. However, it is worthwhile to understand its operation because it is the prototype of more complex bridges.

ELECTRONIC VOLTMETERS

It has been pointed out (Fig. 17-3) that for many purposes the resistance of a voltmeter must be extremely high in order to avoid "loading" errors caused by the current that necessarily flows through the meter. This tends to cause difficulty in measuring relatively low voltages (under perhaps 1000 volts) because a meter movement of given sensitivity takes a progressively smaller multiplier resistance as the voltage range is lowered.

The voltmeter resistance can be made independent of the voltage range by using vacuum tubes or field-effect transistors as electronic dc amplifiers between the circuit being measured and the actual indicator, which is usually a conventional meter movement. As the input resistance of the

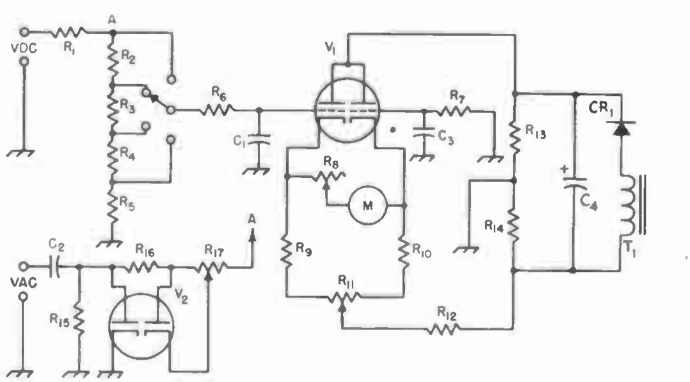


Fig. 17-8 — Vacuum-tube voltmeter circuit.

C1, C3 — .002- to .005-μF mica.
C2 — .01 μF, 1000 to 2000 volts, paper or mica.
C4 — 16 μF electrolytic, 150 volts.
CR1 — 400 PRV rectifier.
M — 0-200 microammeter.
R1 — 1 megohm, 1/2 watt.
R2-R5, incl. — To give desired voltage ranges, totaling 10 megohms.
R6, R7 — 2 to 3 megohms.
R8 — 10,000-ohm variable (calibrate).
R9, R10 — 2000 to 3000 ohms.
R11 — 5000- to 10,000-ohm control (zero set).
R12 — 10,000 to 50,000 ohms.
R13, R14 — App. 25,000 ohms. A 50,000-ohm slider-type wire-wound can be used.
R15 — 10 megohms.
R16 — 3 megohms.
R17 — 10-megohm variable.
T1 — 130-volt 15-mA transformer (only secondary shown).
V1 — Dual triode, 12AU7A.
V2 — Dual diode, 6AL5.

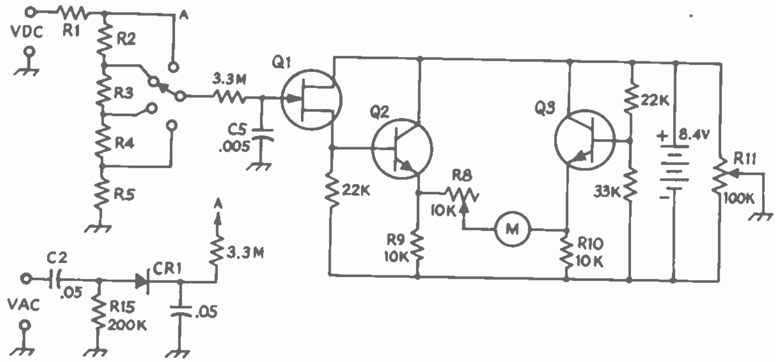


Fig. 17-9 – Electronic voltmeter using field-effect transistor for high input resistance. Components having the same functions as in the VTVM circuit of Fig. 17-8 carry the same designations. (Circuit is basic voltmeter circuit of the Heathkit IM-17.)
CR1 – Silicon diode.
Q1 – Field-effect transistor.
Q2, Q3 – Small-signal audio type.

electronic devices is extremely high – hundreds of megohms – they have essentially no loading effect on the circuit to which they are connected. They do, however, require a closed dc path in their input circuits (although this path can have very high resistance) and are limited in the amplitude of voltage that their input circuits can handle. Because of this, the device actually measures a small voltage across a portion of a high-resistance voltage divider connected to the circuit being measured. Various voltage ranges are obtained by appropriate taps on the voltage divider.

In the design of electronic voltmeters it has become practically standard to use a voltage divider having a resistance of 10 megohms, tapped as required, in series with a 1-megohm resistor incorporated in a probe that makes the actual contact with the "hot" side of the circuit under measurement. The total voltmeter resistance, including probe, is therefore 11 megohms. The 1-megohm probe resistor serves to isolate the voltmeter circuit from the "active" circuit.

The Vacuum-Tube Voltmeter

A typical vacuum-tube voltmeter (VTVM) circuit is given in Fig. 17-8. A dual triode, V1, is arranged so that, with no voltage applied to the left-hand grid, equal currents flow through both sections. Under this condition the two cathodes are at the same potential and no current flows through M. The currents can be adjusted to balance by potentiometer, R11, which takes care of variations in the tube sections and in the values of cathode resistors R9 and R10. When a positive dc voltage is applied to the left-hand grid the current through that tube section increases, so the current balance is upset and the meter indicates. The sensitivity of the meter is regulated by R8, which serves to adjust the calibration. R12, common to the cathodes of both tube sections, is a feedback resistor that stabilizes the system and makes the readings linear. R6 and C1 form a filter for any ac component that may be present, and R6 is balanced by R7 connected to the grid of the second tube section.

Values to be used in the circuit depend considerably on the supply voltage and the sensitivity of the meter, M. R12, and R13-R14, should be adjusted by trial so that the voltmeter circuit can be brought to balance, and to give full-scale deflection on M with about 3 volts applied to the left-hand grid (the voltage chosen for this determines the lowest voltage range of the instrument). The meter connections can be reversed to read voltages that are negative with respect to ground.

The small circuit associated with V2 is for ac measurements, as described in a later section.

As compared with conventional dc instruments, the VTVM has the disadvantages of requiring a source of power for its operation, and generally must have its "cold" terminal grounded in order to operate reliably. It is also somewhat susceptible to erratic readings from rf pickup when used in the vicinity of a transmitter, and in such cases may require shielding. However, its advantages outweigh these disadvantages in many applications.

The FET Voltmeter

The circuit of an electronic voltmeter using a field-effect transistor as an input device is shown in Fig. 17-9. Allowing for the differences between vacuum tubes and semiconductors, the operation of this circuit is analogous to that of Fig. 17-8. Transistors Q2 and Q3 correspond to the dual triode in the VTVM circuit, but since the input resistance of Q2 is fairly low, it is preceded by an FET, Q1, with source-coupled output. Note that in this circuit the "zero" or current-balance control, R11, varies the gate bias on Q1 by introducing an adjustable positive voltage in series with the source. This arrangement permits applying the adjustable bias to the gate through the voltmeter range divider, with no other provision needed for completing the dc gate-source path.

The small circuit associated with CR1 is for ac voltage measurement, to be discussed later.

As the power supply for the FET voltmeter is a self-contained battery, the grounding restrictions associated with a VTVM do not apply. The instrument can, however, be susceptible to rf fields if not shielded and grounded.

Electronic Ohmmeters

Most commercial electronic voltmeters include provision for measuring resistance and ac voltage, in addition to dc voltage. The basic ohmmeter circuit generally used is that of Fig. 17-6C. Since for practical purposes the input resistance of the vacuum tube or FET can be assumed to approach infinity, electronic ohmmeters are capable of measuring resistances in the hundreds of megohms – a much higher range than can be reached with an ordinary microammeter.

AC INSTRUMENTS AND CIRCUITS

Although purely electromagnetic instruments that operate directly from alternating current are available, they are seen infrequently in present-day amateur equipment. For one thing, their use is not feasible above power-line frequencies.

Practical instruments for audio and radio frequencies generally use a dc meter movement in conjunction with a rectifier. Voltage measurements suffice for nearly all test purposes. Current, as such, is seldom measured in the af range. When rf current is measured the instrument used is a thermocouple milliammeter or ammeter.

The Thermocouple Meter

In a thermocouple meter the alternating current flows through a low-resistance heating element. The power lost in the resistance generates heat which warms a "thermocouple," a junction of certain dissimilar metals which has the property of developing a small dc voltage when heated. This voltage is applied to a dc milliammeter calibrated in suitable ac units. The heater-thermocouple-dc meter combination is usually housed in a regular meter case.

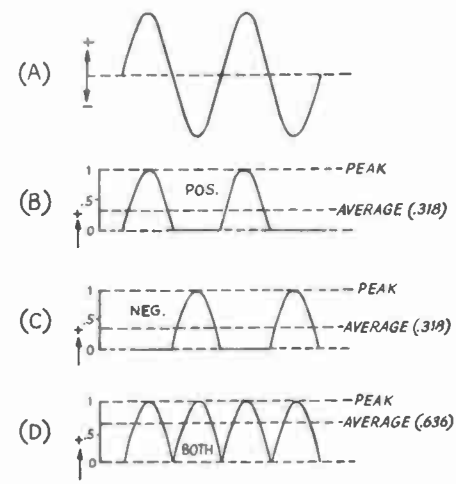


Fig. 17-11 – Sine-wave alternating current or voltage (A), with half-wave rectification of the positive half cycle (B) and negative half cycle (C). D – full-wave rectification. Average values are shown with relation to a peak value of 1.

Fig. 17-10 – Rf ammeter mounted in a Minibox, with connectors for placing the meter in series with a coaxial line. A bakelite-case meter should be used to minimize shunt capacitance (which introduces error) although a metal-case meter can be used if mounted on bakelite sheet with a large cut-out in the case around the rim. The meter can be used for rf power measurements ($P = I^2R$) when connected between a transmitter and a nonreactive load of known resistance.

Thermocouple meters can be obtained in ranges from about 100 mA to many amperes. Their useful upper frequency limit is in the neighborhood of 100 MHz. Their principal value in amateur work is in measuring current into a known load resistance for calculating the rf power delivered to the load. A suitable mounting for this is shown in Fig. 17-10, for use in coaxial lines.

RECTIFIER INSTRUMENTS

The response of a rectifier-type meter is proportional (depending on the design) to either the peak amplitude or average amplitude of the rectified ac wave, and never directly responsive to the rms value. The meter therefore cannot be calibrated in rms without preknowledge of the relationship that happens to exist between the "real" reading and the rms value. This relationship, in general, is not known, except in the case of single-frequency ac (a sine wave). Very many practical measurements involve nonsinusoidal wave forms, so it is necessary to know what kind of instrument you have, and what it is actually

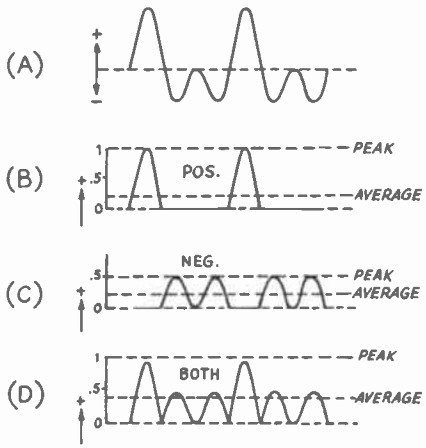


Fig. 17-12 – Same as Fig. 17-11 for an unsymmetrical waveform. The peak values are different with positive and negative half-cycle rectification.

reading, in order to make measurements intelligently.

Peak and Average with Sine-Wave Rectification

Fig. 17-11 shows the relative peak and average values in the outputs of half- and full-wave rectifiers (see power-supply chapter for further details). As the positive and negative half cycles of the sine wave have the same shape (A), half-wave rectification of either the positive half (B) or the negative half (C) gives exactly the same result. With full-wave rectification (D) the peak is still the same, but the average is doubled, since there are twice as many half cycles per unit of time.

Unsymmetrical Wave Forms

A nonsinusoidal waveform is shown in Fig. 17-12A. When the positive half cycles of this wave are rectified the peak and average values are as shown at B. If the polarity is reversed and the negative half cycles are rectified the peak value is different but the average value is unchanged. The fact that the average of the positive side is equal to the average of the negative side is true of *all* ac waveforms, but different waveforms have different averages. Full-wave rectification of such a "lop-sided" wave doubles the average value, but the peak reading is always the same as it is with the half cycle that produces the *highest* peak in half-wave rectification.

Effective-Value Calibration

The actual scale calibration of commercially-made rectifier-type voltmeters is very often (almost always, in fact) in terms of rms values. For sine waves this is satisfactory, and useful since rms is the standard measure at power-line frequency. It is also useful for many rf applications where the waveform is often closely sinusoidal. But in other cases, particularly in the af range, the error may be considerable when the waveform is not pure.

Turn-Over

From Fig. 17-12 it is apparent that the calibration of an average-reading meter will be the same whether the positive or negative sides are rectified. A half-wave *peak*-reading instrument, however, will indicate different values when its connections to the circuit are reversed (turn-over effect). Very often readings are taken both ways, in which case the sum of the two is the peak-to-peak value, a useful figure in much audio and video work.

Average- and Peak-Reading Circuits

The basic difference between average- and peak-reading rectifier circuits is that in the former the output is not filtered while in the latter a filter capacitor is charged up to the peak value of the output voltage. Fig. 17-13A shows typical average-reading circuits, one half-wave and the other full-wave. In the absence of dc filtering the meter responds to wave forms such as are shown at B, C and D in Figs. 17-11 and 17-12, and since the inertia of the pointer system makes it unable to follow the rapid variations in current, it averages them out mechanically.

In Fig. 17-13A CR1 actuates the meter; CR2 provides a low-resistance dc return in the meter circuit on the negative half cycles. R1 is the voltmeter multiplier resistance. R2 forms a voltage

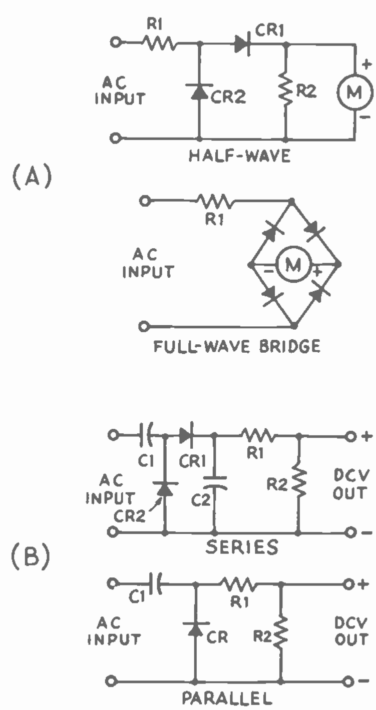


Fig. 17-13 – A – Half-wave and full-wave rectification for an instrument intended to operate on average values. B – half-wave circuits for a peak-reading meter.

divider with R1 (through CR1) which prevents more than a few ac volts from appearing across the rectifier-meter combination. A corresponding resistor can be used across the full-wave bridge circuit.

In these two circuits no provision is made for isolating the meter from any dc voltage that may be on the circuit under measurement. The error caused by this can be avoided by connecting a large capacitance in series with the "hot" lead. The reactance must be low compared with the meter impedance (see next section) in order for the full ac voltage to be applied to the meter circuit. As much as 1 μF may be required at line frequencies with some meters. The capacitor is not usually included in a VOM.

Series and shunt peak-reading circuits are shown in Fig. 17-13B. Capacitor C1 isolates the rectifier from dc voltage on the circuit under measurement. In the series circuit (which is seldom used) the time constant of the C2R1R2 combination must be very large compared with the period of the lowest ac frequency to be measured; similarly with C1R1R2 in the shunt circuit. The reason is that the capacitor is charged to the peak value of voltage when the ac wave reaches its maximum, and then must hold the charge (so it can register on a dc meter) until the next maximum of the same polarity. If the time constant is 20 times the ac period the charge will have decreased by about 5 percent by the time the next charge occurs. The *average* drop will be smaller, so the error is appreciably less. The error will decrease rapidly with increasing frequency, assuming no change in the circuit values, but will increase at lower frequencies.

In Fig. 17-13B R1 and R2 form a voltage divider which reduces the peak dc voltage to 71 percent of its actual value. This converts the peak reading to rms on sine-wave ac. Since the peak-reading circuits are incapable of delivering appreciable current without considerable error, R2 is usually the 11-megohm input resistance of an electronic voltmeter. R1 is therefore approximately 4.7 megohms, making the total resistance approach 16 megohms. A capacitance of .05 μF is sufficient for low audio frequencies under these conditions. Much smaller values of capacitance suffice for radio frequencies, obviously.

Voltmeter Impedance

The impedance of the voltmeter at the frequency being measured may have an effect on the accuracy similar to the error caused by the resistance of a dc voltmeter, as discussed earlier. The ac meter acts like a resistance in prallel with a capacitance, and since the capacitive reactance decreases with increasing frequency, the impedance also decreases with frequency. The resistance is subject to some variation with voltage level, particularly at very low voltages (of the order of 10 volts or less) depending upon the sensitivity of the meter movement and the kind of rectifier used.

The ac load resistance represented by a diode rectifier is approximately equal to one-half its dc load resistance. In Fig. 17-13A the dc load is essentially the meter resistance, which is generally quite low compared with the multiplier resistance R1, so the total resistance will be about the same as the multiplier resistance. The capacitance depends on the components and construction, test lead length and disposition, and such factors. In general, it has little or no effect at power-line and low audio frequencies, but the ordinary VOM loses accuracy at the higher audio frequencies and is of little use at rf. For radio frequencies it is necessary to use a rectifier having very low inherent capacitance.

Similar limitations apply to the peak-reading circuits. In the parallel circuit the resistive component of the impedance is smaller than in the series circuit, since the dc load resistance, R1R2, is directly across the circuit being measured, and is therefore in parallel with the diode ac load resistance. In both peak-reading circuits the effective capacitance may range from 1 or 2 to a few hundred pF. Values of the order of 100 pF are to be expected in electronic voltmeters of customary design and construction.

Linearity

Fig. 17-14, a typical current/voltage characteristic of a small semiconductor rectifier, indicates that the forward dynamic resistance of the diode is not constant, but rapidly decreases as the forward voltage is increased from zero. The transition from high to low resistance occurs at considerably less than 1 volt, but is in the range of voltage required by the associated dc meter. With an average-reading circuit the current tends to be proportional to the *square* of the applied voltage. This crowds the calibration points at the low end of the meter scale. For most measurement purposes, however, it is far more desirable for the output to be "linear;" that is, for the reading to be *directly* proportional to the applied voltage.

To achieve linearity it is necessary to use a relatively large load resistance for the diode – large enough so that this resistance, rather than the diode's own resistance, will govern the current flow. A linear or equally spaced scale is thus gained at the expense of sensitivity. The amount of resistance needed depends on the type of diode;

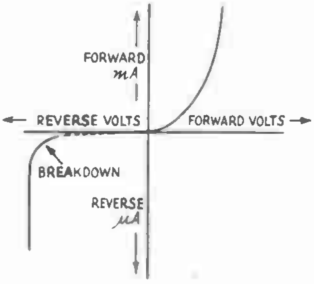


Fig. 17-14 – Typical semiconductor diode characteristic. Actual current and voltage values vary with the type of diode, but the forward-current curve would be in its steep part with only a volt or so applied. Note change in current scale for reverse current. Breakdown voltage, again depending on diode type, may range from 15 or 20 volts to several hundred.

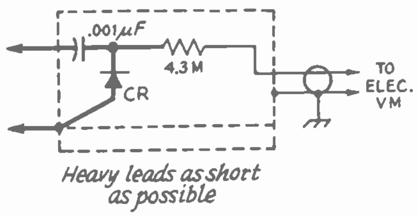


Fig. 17-15 – Rf probe circuit. CR is a small semiconductor rectifier, usually point-contact germanium. The resistor value, for exact voltage division to rms, should be 4.14 megohms, but standard values are generally used, including 4.7 megohms.

5000 to 50,000 ohms usually suffices for a germanium rectifier, depending on the dc meter sensitivity, but several times as much may be needed for silicon. The higher the resistance, the greater the meter sensitivity required; i.e., the basic meter must be a microammeter rather than a low-range milliammeter.

Reverse Current

When voltage is applied in the reverse direction there is a small leakage current in semiconductor diodes. This is equivalent to a resistance connected across the rectifier, allowing current to flow during the half cycle which should be completely nonconducting, and causing an error in the dc meter reading. This "back resistance" is so high as to be practically unimportant with silicon, but may be less than 100 kΩ with germanium.

The practical effect of back resistance is to limit the amount of resistance that can be used in the dc load resistance. This in turn affects the linearity of the meter scale.

The back resistance of vacuum-tube diodes is infinite, for practical purposes.

RF VOLTAGE

Special precautions must be taken to minimize the capacitive component of the voltmeter impedance at radio frequencies. If possible, the rectifier circuit should be installed permanently at the point where the rf voltage to be measured exists, using the shortest possible rf connections. The dc meter can be remotely located, however.

For general rf measurements an rf probe is used in conjunction with an electronic voltmeter, substituted for the dc probe mentioned earlier. The circuit of Fig. 17-15, essentially the peak-reading shunt circuit of Fig. 17-13B, is generally used. The series resistor, installed in the probe close to the rectifier, prevents rf from being fed through the probe cable to the electronic voltmeter, being helped in this by the cable capacitance. This resistor, in conjunction with the 10-megohm divider resistance of the electronic voltmeter, also reduces the peak rectified voltage to a dc value equivalent to the rms of the rf signal, to make the rf readings consistent with the regular ac calibration.

Of the diodes readily available to amateurs, the germanium point-contact type is preferred for rf applications. It has low capacitance (of the order of 1 pF) and in the high-back-resistance types the reverse current is not serious. The principal limitation is that its safe reverse voltage is only about 50-75 volts, which limits the rms applied voltage to 15 or 20 volts, approximately. Diodes can be connected in series to raise the overall rating.

Linearity at Radio Frequencies

The bypass or filter capacitance normally used in rf rectifier circuits is large enough, together with the resistance in the system, to have a time constant sufficient for peak readings. However, if the resistance is low (the load sometimes is just the microammeter or milliammeter alone) the *linearity* of the voltmeter will be affected as previously described, even if the time constant is fairly large. It is not safe to assume that the voltmeter is even approximately linear unless the load resistance is of the order of 10,000 ohms or greater.

Nonlinear voltmeters are useful as indicators, as where null indicators are called for, but should not be depended upon for actual measurement of voltage.

RF Power

Power at radio frequencies can be measured by means of an accurately-calibrated rf voltmeter connected across the load in which the power is being dissipated. If the load is a known pure resistance the power, by Ohm's Law, is equal to E^2/R, where E is the rms value of the voltage.

The method only indicates *apparent* power if the load is not a pure resistance. The load can be a terminated transmission line tuned, with the aid of bridge circuits such as are described in the next section, to act as a known resistance. An alternative load is a "dummy" antenna, a known pure resistance capable of dissipating the rf power safely.

AC BRIDGES

In its simplest form, the ac bridge is exactly the same as the Wheatstone bridge discussed earlier. However, complex impedances can be substituted for resistances, as suggested by Fig. 17-16A. The same bridge equation holds if Z is substituted for R in each arm. For the equation to be true, however, *the phase angles as well as the numerical values of the impedances must balance*; otherwise, a true null voltage is impossible to obtain. This means that a bridge with all "pure" arms (pure resistance or reactance) cannot measure complex impedances; a combination of R and X must be present in at least one arm besides the unknown.

The actual circuits of ac bridges take many forms, depending on the type of measurement intended and on the frequency range to be covered. As the frequency is raised stray effects (unwanted capacitances and inductances, principally) become more pronounced. At radio frequencies special attention must be paid to minimizing them.

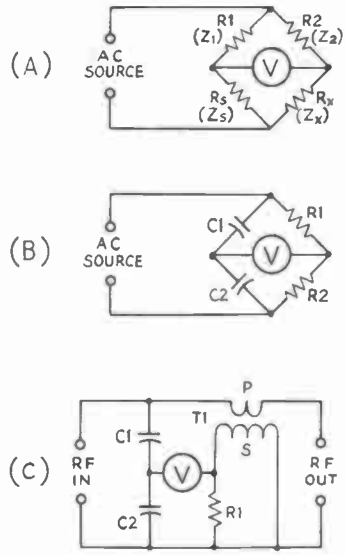


Fig. 17-16 – A – Generalized form of bridge circuit for either ac or dc. B – One form of ac bridge frequently used for rf measurements. C – SWR bridge for use in transmission lines. This circuit is often calibrated in power rather than voltage.

Most amateur-built bridges are used for rf measurements, especially SWR measurements on transmission lines. The circuits at B and C, Fig. 17-16, are favorites for this purpose. These basic forms are often modified considerably, as will be seen by the constructional examples later in the chapter.

Fig. 17-16B is useful for measuring both transmission lines and "lumped constant" components. Combinations of resistance and capacitance are often used in one or more arms; this may be required for eliminating the effects of stray capacitance.

Fig. 17-16C is used only on transmission lines, and only on those lines having the characteristic impedance for which the bridge is designed.

SWR Measurement – The Reflectometer

In measuring standing-wave ratio advantage is taken of the fact that the voltage on a transmission line consists of two components traveling in opposite directions. The power going from the transmitter to the load is represented by one voltage (designated "incident" or "forward") and the power reflected from the load is represented by the other. Because the relative amplitudes and phase relationships are definitely established by the line's characteristic impedance, its length and the load impedance in which it is terminated, a bridge circuit can separate the incident and reflected voltages for measurement. This is sufficient for determining the SWR. Bridges designed for this purpose are frequently called reflectometers.

Referring to Fig. 17-16A, if R1 and R2 are made equal, the bridge will be balanced when $R_X = R_S$. This is true whether R_X is an actual resistor or the input resistance of a perfectly matched transmission line, provided R_S is chosen to equal the characteristic impedance of the line. Even if the line is not properly matched, the bridge will still be balanced for power traveling *outward* on the line, since outward-going power sees only the Z_0 of the line until it reaches the load. However, power reflected back from the load does not "see" a bridge circuit, and the reflected voltage registers on the voltmeter. From the known relationship between the incident and reflected voltages the SWR is easily calculated:

$$SWR = \frac{V_0 + V_r}{V_0 - V_r}$$

where V_0 is the forward voltage and V_r is the reflected voltage. The forward voltage may be measured either by disconnecting R_X or shorting it.

The "Reflected Power Meter"

Fig. 17-16C makes use of mutual inductance between the primary and secondary of T1 to establish a balancing circuit. C1 and C2 form a voltage divider in which the voltage across C2 is in the same phase as the voltage at that point on the transmission line. The relative phase of the voltage across R1 is determined by the phase of the *current* in the line. If a pure resistance equal to the design impedance of the bridge is connected to the "RF Out" terminals, the voltages across R1 and C2 will be out of phase and the voltmeter reading will be minimum; if the *amplitudes* of the two voltages are also equal (they are made so by bridge adjustment) the voltmeter will read zero. Any other value of resistance or impedance connected to the "RF Out" terminals will result in a finite voltmeter reading. When used in a transmission line this reading is proportional to the reflected voltage. To measure the incident voltage the secondary terminals of T1 can be reversed. To function as described, the secondary leakage reactance of T1 must be very large compared to the resistance of R1.

Instruments of this type are usually designed for convenient switching between forward and reflected, and are often calibrated to read power in the specified characterisitic impedance. The net power transmission is equal to the incident power minus the reflected power.

Sensitivity vs. Frequency

In all of the circuits in Fig. 17-16 the sensitivity is independent of the applied frequency, within practical limits. Stray capacitances and couplings generally limit the performance of all three at the high-frequency end of the useful range. Fig. 17-16A will work right down to dc, but the low-frequency performance of Fig. 17-16B is degraded when the capacitive reactances become so large that voltmeter impedance becomes low in comparison (in all these bridge circuits, it is assumed that the voltmeter impedance is high compared with the impedance of the bridge arms). In Fig. 17-16C the performance is limited at low frequencies by the fact that the transformer reactance decreases with frequency, so that eventually the reactance is not very high in comparison with the resistance of R1.

The "Monimatch"

A type of bridge which is quite simple to make, but in which the sensitivity rises directly with frequency, is the Monimatch and its various offspring. The circuit cannot be described in terms of lumped constants, as it makes use of the distrbuted mutual inductance and capacitance between the center conductor of a transmission line and a wire placed parallel to it. The wire is terminated in a resistance approximating the characteristic impedance of the transmission line at one end and feeds a diode rectifier at the other. A practical example is shown later in this chapter.

FREQUENCY MEASUREMENT

The regulations governing amateur operation require that the transmitted signal be maintained inside the limits of certain bands of frequencies.* The exact frequency need not be known, so long as it is not outside the limits. On this last point there are no tolerances: It is up to the individual amateur to see that he stays safely "inside."

This is not difficult to do, but requires some simple apparatus and the exercise of some care. The apparatus commonly used is the frequency-marker generator, and the method involves use of the station receiver, as in Fig. 17-17.

THE FREQUENCY MARKER

The marker generator in its simplest form is a high-stability oscillator generating a series of signals which, when detected in the receiver, mark the exact edges of the amateur assignments. It does this by oscillating at a low frequency that has harmonics falling on the desired frequencies.

All U.S. amateur band limits are exact multiples of 25 kHz, whether at the extremes of a band or at points marking the subdivisions between types of emission, license privileges, and so on. A 25-kHz fundamental frequency therefore will produce the desired marker signals if its harmonics at the higher frequencies are strong enough. But since harmonics appear at 25-kHz intervals throughout the spectrum, along with the desired markers, the problem of identifying a *particular* marker arises. This is easily solved if the receiver has a reasonably good calibration. If not, most marker circuits provide for a choice of fundamental outputs of 100 and 50 kHz as well as 25 kHz, so the question can be narrowed down to initial identification of 100-kHz intervals. From these, the desired 25-kHz (or 50-kHz) points can easily be spotted. Coarser frequency intervals are rarely required; there are usually signals available from stations of known frequency, and the 100-kHz points can be counted off from them.

Transmitter Checking

In checking one's own transmitter frequency the signal from the transmitter is first tuned in on the receiver and the dial setting at which it is heard is noted. Then the *nearest* marker frequencies above and below the transmitter signal are turned in and identified. The transmitter frequency is obviously between these two known frequencies.

If the marker frequencies are accurate, this is all that needs to be known – except that the transmitter frequency must not be so close to a band (or subband) edge that sideband frequencies, especially in phone transmission, will extend over the edge.

If the transmitter signal is "inside" a marker at the edge of an assignment, to the extent that there is an audible beat note with the receiver's BFO turned off, normal cw sidebands are safely inside the edge. (This statement does not take into account *abnormal* sidebands such as are caused by clicks and chirps.) For phone the "safety" allowance is usually taken to be about 3 kHz, the nominal width of one sideband. A frequency difference of this order can be estimated by noting the receiver dial settings for the two 25-kHz markers which bracket the signal and dividing 25 by the number of dial divisions between them. This will give the number of kHz per dial division.

Transceivers

The method described above is applicable when the receiver and transmitter are separate pieces of equipment. When a transceiver is used and the

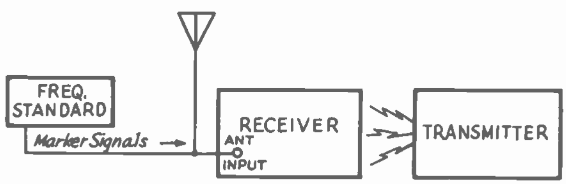


Fig. 17-17 – Setup for using a frequency standard. It is necessary that the transmitter signal be weak in the receiver – of the same order of strength as the marker signal from the standard. This requirement can usually be met by turning on just the transmitter oscillator, leaving all power off any succeeding stages. In some cases it may also be necessary to disconnect the antenna from the receiver.

* These limits depend on the type of emission and class of license held, as well as on international agreements. See the latest edition of *The Radio Amateur's License Manual* for current status.

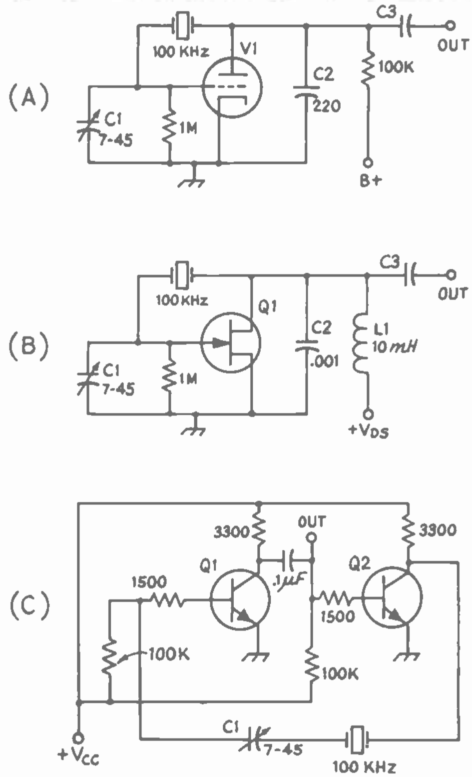


Fig. 17-18 — Three simple 100-kHz oscillator circuits. C is the most suitable of available transistor circuits (for marker generators) and is recommended where solid-state is to be used. In all three circuits C1 is for fine frequency adjustment. The output coupling capacitor, C3, is generally small — 20 to 50 pF — a compromise to avoid loading the oscillator by the receiver antenna input while maintaining adequate coupling for good harmonic strength.

transmitting frequency is automatically the same as that to which the receiver is tuned, setting the tuning dial to a spot between two known marker frequencies is all that is required.

The proper dial settings for the markers are those at which, with the BFO on, the signal is tuned to *zero* beat – the spot where the beat disappears as the tuning makes the beat tone progressively lower. Exact zero beat can be determined by a very slow rise and fall of background noise, caused by a beat of a cycle or less per second.

FREQUENCY-MARKER CIRCUITS

The basic frequency-determining element in most amateur frequency markers is a 100-kHz crystal. Although the marker generator should produce harmonics at 25-kHz and 50-kHz intervals, crystals (or other high-stability devices) for frequencies lower than 100 kHz are expensive and difficult to obtain. However, there is really no need for them, since it is easy to divide the basic frequency down to any figure one desires; 50 and 25 kHz require only two successive divisions, each by 2. In the division process, the harmonic output of the generator is greatly enhanced, making the generator useful at frequencies well into the vhf range.

Simple Crystal Oscillators

Fig. 17-18 illustrates a few of the simpler circuits. Fig. 17-18A is a long-time favorite where vacuum tubes are used and is often incorporated in receivers. C1 in this and the other circuits is used for exact adjustment of the oscillating frequency to 100 kHz, which is done by using the receiver for comparing one of the oscillator's harmonics with a standard frequency transmitted by WWV, WWVH, or a similar station.

Fig. 17-18B is a field-effect transistor analog of the vacuum-tube circuit. However, it requires a 10-mH coil to operate well, and since the harmonic output is not strong at the higher frequencies the circuit is given principally as an example of a simple transistor arrangement. A much better oscillator is shown at C. This is a cross-connected pair of transistors forming a multivibrator of the "free-running" or "astable" type, locked at 100 kHz by using the crystal as one of the coupling elements. While it can use two separate bipolar transistors as shown, it is much simpler to use an integrated-circuit dual gate, which will contain all the necessary parts except the crystal and capacitors and is considerably less expensive, as well as more compact, than the separate components. An example is shown later in the chapter.

Frequency Dividers

Electronic division is accomplished by a "bistable" flip-flop or cross-coupled circuit which produces one output change for every two impulses applied to its input circuit, thus dividing the applied frequency by 2. All division therefore must be in terms of some power of 2. In practice this is no handicap since with modern integrated-circuit flip-flops, circuit arrangements can be worked out for division by any desired number.

As flip-flops and gates in integrated circuits come in compatible series – meaning that they work at the same supply voltage and can be directly connected together – a combination of a dual-gate version of Fig. 17-18C and a dual flip-flop make an attractively simple combination for the marker generator.

There are several different basic types of flip-flops, the variations having to do with methods of driving (dc or pulse operation) and control of the counting function. Information on the operating principles and ratings of a specific type usually can be obtained from the manufacturer. The counting-control functions are not needed in using the flip-flop in a simple marker generator, although they come into play when dividing by some number other than a power of 2.

Frequency Standards

The difference between a marker generator and a frequency standard is that in the latter special pains are taken to make the oscillator frequency as stable as possible in the face of variations in temperature, humidity, line voltage, and other factors which could cause a small change in frequency.

While there are no definite criteria that distinguish the two in this respect, a circuit designated as a "standard" for amateur purposes should be capable of maintaining frequency within at least a few parts per million under normal variations in ambient conditions, without adjustment. A simple marker generator using a 100-kHz crystal can be expected to have frequency variations 10 times (or more) greater under similar conditions. It can of course be adjusted to exact frequency at any time the WWV (or equivalent) signal is available.

The design considerations of high-precision frequency standards are outside the scope of this chapter, but information is available from time to time in periodicals.

OTHER METHODS OF FREQUENCY CHECKING

The simplest possible frequency-measuring device is a parallel *LC* circuit, tunable over a desired frequency range and having its tuning dial calibrated in terms of frequency. It can be used only for checking circuits in which at least a small amount of rf power is present, because the energy required to give a detectable indication is not available in the *LC* circuit itself; it has to be extracted from the circuit being measured; hence the name absorption frequency meter. It will be observed that what is actually measured is the frequency of the rf *energy, not* the frequency to which the circuit in which the energy is present may be tuned.

The measurement accuracy of such an instrument is low, compared with the accuracy of a marker generator, because the *Q* of a practicable *LC* circuit is not high enough to make precise reading of the dial possible. Also, any two circuits coupled together react on each others' tuning. (This can be minimized by using the loosest coupling that will give an adequate indication.)

The absorption frequency meter has one useful advantage over the marker generator – it will respond *only* to the frequency to which it is tuned, or to a band of frequencies very close to it. Thus there is no harmonic ambiguity, as there sometimes is when using a marker generator.

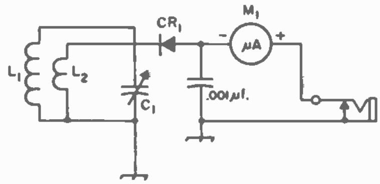


Fig. 17-19A – Absorption frequency-meter circuit. The closed-circuit phone jack may be omitted if listening is not wanted; in that case the positive terminal of M1 goes to common ground.

Absorption Circuit

A typical absorption frequency-meter circuit is shown in Fig. 17-19. In addition to the adjustable tuned circuit, L1C1, it includes a pickup coil, L2, wound over L1, a high-frequency semiconductor diode, CR1, and a microammeter or low-range (usually not more than 0-1 mA) milliammeter. A phone jack is included so the device can be used for listening to the signal.

The sensitivity of the frequency meter depends on the sensitivity of the dc meter movement and the size of L2 in relation to L1. There is an optimum size for this coil which has to be found by experiment. An alternative is to make the rectifier connection to an adjustable tap on L1, in which case there is an optimum tap point. In general, the rectifier coupling should be a little *below* (that is, less tight) the point that gives maximum response, since this will make the indications sharper.

Calibration

The absorption frequency meter must be calibrated by taking a series of readings on various frequencies from circuits carrying rf power, the frequency of the rf energy first being determined by some other means such as a marker generator and receiver. The setting of the dial that gives the highest meter indication is the calibration point for that frequency. This point should be determined by tuning through it with loose coupling to the circuit being measured.

OTHER INSTRUMENTS AND MEASUREMENTS

Many measurements require a source of ac power of adjustable frequency (and sometimes adjustable amplitude as well) in addition to what is already available from the transmitter or receiver. Rf and af test oscillators, for example, provide signals for purposes such as receiver alignment, testing of phone transmitters, and so on. Another valuable adjunct to the station is the oscilloscope, especially useful for checking phone modulation.

Rf Oscillators for Circuit Alignment

Receiver testing and alignment, covered in an earlier chapter, uses equipment common to ordinary radio service work. Inexpensive rf signal generators are available, both complete and in kit form. However, any source of signal that is weak enough to avoid overloading the receiver usually will serve for alignment work. The frequency marker generator is a satisfactory signal source. In addition, its frequencies, although not continuously adjustable, are known far more precisely, since the usual signal-generator calibration is not highly accurate. For rough work the dip meter described in the next section will serve.

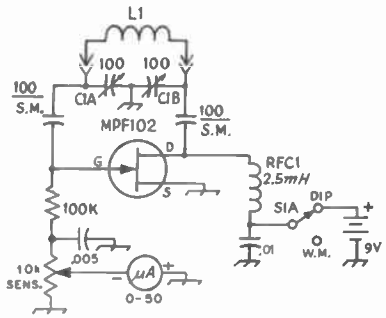


Fig. 17-19B – An FET source-dipper circuit suitable for use from 1.5 to 50 MHz. For operation at vhf and uhf the value of C1 should be made smaller, RFC1 would be a vhf type, and the bypass capacitors would be smaller in value. For uhf use Q1 would be changed to a uhf-type FET, a 2N4416 or similar.

THE DIP METER

The dip meter reverses the absorption-wavemeter procedure in that *it* supplies the rf power by incorporating a tunable oscillator from which the circuit being checked absorbs energy when this circuit and the oscillator are tuned to the same frequency and coupled together. In the vacuum-tube version the energy absorption causes a decrease or "dip" in the oscillator's rectified grid current, measured by a dc microammeter.

The same principle can be applied to solid-state oscillators. In some transistor versions the oscillator rf power is rectified by a diode to provide a meter indication. This technique can result in "dead spots" in the tuning range if the oscillator power is too low to enable the diode to conduct at all times. The circuit of Fig. 17-19B avoids the problem by measuring the changes in source current. In the W.M. (wavemeter) position of S1 the gate-source junction of Q1 serves as the detector diode.

Each tuning range of the dipper should overlap to provide sufficient coverage to check circuits of unknown resonant frequency. Plug-in coils are normally used to allow continuous coverage from 1.5 to at least 250 MHz.

Calibration

A dipper should have reasonably accurate calibration. Calibration of the dipper dial can be effected by monitoring the dipper output signal with a calibrated receiver. Make sure the *fundamental* frequency of the dipper is being used during calibration.

Operating the Dip Meter

The dip meter will check only resonant circuits, since nonresonant circuits or components will not absorb energy at a specific frequency. The circuit may be either lumped or linear (a transmission-line type circuit) provided only that it has enough Q to give sufficient coupling to the dip-meter coil for detectable absorption of rf energy. Generally the coupling is principally inductive, although at times there may be sufficient capacitive coupling between the meter and a circuit point that is at relatively high potential with respect to ground to permit a reading. For inductive coupling, maximum energy absorption will occur when the meter

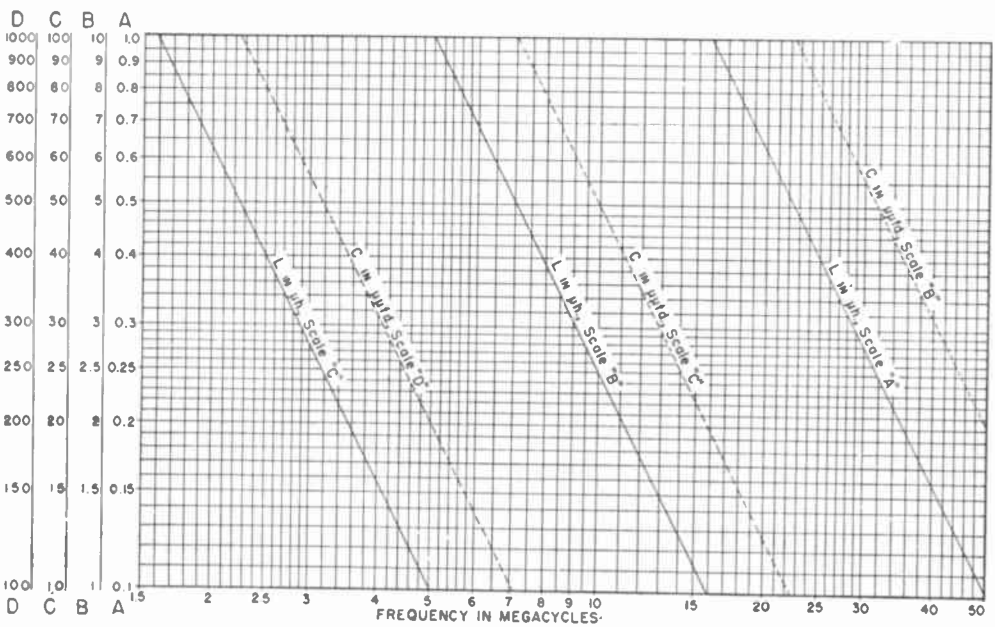


Fig. 17-20 – Chart for determining unknown values of L and C in the range of 0.1 to 100 μH and 2 to 1000 pF, using standards of 100 pF and 5 μH.

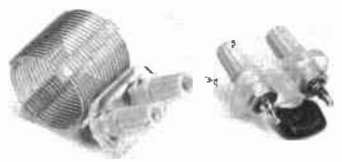


Fig. 17-21 – A convenient mounting, using binding-post plates, for L and C standards made from commercially available parts. The capacitor is a 100-pF silver mica unit, mounted so the lead length is as nearly zero as possible. The inductance standard, 5 μH, is 17 turns of coil stock, 1-inch diameter, 16 turns per inch.

is coupled to a coil (the same coupling rules that apply to any two coils are operative here) in the tuned circuit being checked, or to a high-current point in a linear circuit.

Because of distributed capacitance (and sometimes inductance) most circuits resonant at the lower amateur frequencies will show quasi-linear-type resonances at or close to the vhf region. A vhf dip meter will uncover these, often with beneficial results since such "parasitic" resonances can cause unwanted responses at harmonics of the intended frequency, or be responsible for parasitic oscillations in amplifiers. Caution must be used in checking transmission lines or antennas – and, especially, combinations of antenna and line – on this account, because these linear circuits have well-defined series of harmonic responses, based on the lowest resonant frequency, which may lead to false conclusions respecting the behavior of the system.

Measurements with the dip meter are essentially frequency measurements, and for best accuracy the coupling between the meter and circuit under checking must be as loose as will allow a perceptible dip. In this respect the dip meter is similar to the absorption wavemeter.

Measuring Inductance and Capacitance with the Dip Meter

With a carefully calibrated dip meter, properly operated, inductance and capacitance in the values ordinarily used for the 1.5-50 MHz range can be measured with ample accuracy for practical work. The method requires two accessories: an inductance "standard" of known value, and a capacitance standard also known with reasonable accuracy. Values of 100 pF for the capacitance and 5 μH for the inductance are convenient. The chart of Fig. 17-20 is based on these values.

The L and C standards can be quite ordinary components. A small silver-mica capacitor is satisfactory for the capacitance, since the customary tolerance is ±5 percent. The inductance standard can be cut from commercial machine-wound coil stock; if none is available, a homemade equivalent in diameter, turn spacing, and number of turns can be substituted. The inductance will be 5 μH within amply close tolerances if the specifications in Fig. 17-21 are followed closely. In any case, the inductance can easily be adjusted to the proper value; it should resonate with the 100-pF capacitor at 7100 kHz.

The setup for measuring an unknown is shown in Fig. 17-22. Inductance is measured with the unknown connected to the standard capacitance. Couple the dip meter to the coil and adjust the meter for the dip, using the loosest possible coupling that will give a usable indication. Similar procedure is followed for capacitance measurement, except that the unknown is connected to the standard inductance. Values are read off the chart for the frequency indicated by the dip meter.

Coefficient of Coupling

The same equipment can be used for measurement of the coefficient of coupling between two coils. This simply requires two measurements of inductance (of *one* of the coils) with the coupled coil first open-circuited and then short-circuited. Connect the 100-pF standard capacitor to one coil and measure the inductance with the terminals of the second coil open. Then short the terminals of the second coil and again measure the inductance of the first. The coefficient of coupling is given by

$$k = \sqrt{1 - \frac{L_2}{L_1}}$$

where k = coefficient of coupling
$L1$ = inductance of first coil with terminals of second coil open
$L2$ = inductance of first coil with terminals of second coil shorted.

AUDIO-FREQUENCY OSCILLATORS

Tests requiring an audio-frequency signal generally call for one that is a reasonably good sine wave, and the best oscillator circuits for this are RC-coupled, operating as nearly as possible as Class A amplifiers. Variable frequency covering the entire audio range is needed for determining frequency response of audio amplifiers, but this is

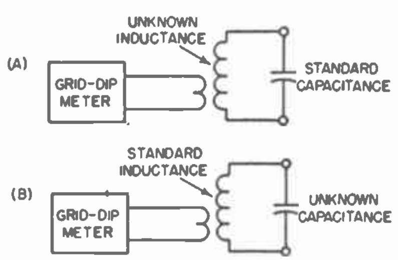


Fig. 17-22 – Setups for measuring inductance and capacitance with the dip meter.

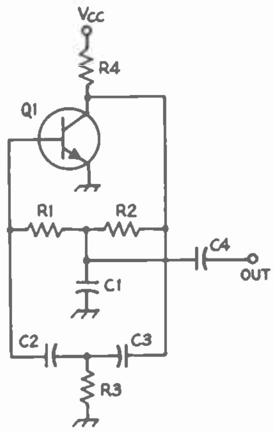


Fig. 17-23 – Twin-T audio oscillator circuit. Representative values for R1-R2 and C1 range from 18kΩ and .05 μF for 750 Hz to 15kΩ and .02 μF for 1800 Hz. For the same frequency range, R3 and C2-C3 vary from 1800 ohms and .02 μF to 1500 ohms and .01 μF. R4 should be approximately 3300 ohms. C4, the output coupling capacitor, can be .05 μF for high-impedance loads.

a relatively unimportant type of test in amateur equipment. The variable-frequency af signal generator is best purchased complete; kits are readily available at prices that compare very favorably with the cost of parts.

For most phone-transmitter testing, and for simple trouble shooting in af amplifiers, an oscillator generating one or two frequencies with good wave form is adequate. A "two-tone" (dual) oscillator is particularly useful for testing sideband transmitters, and a constructional example is found later in the chapter.

The circuit of a simple *RC* oscillator useful for general test purposes is given in Fig. 17-23. This "Twin-T" arrangement gives a wave form that is satisfactory for most purposes, and by choice of circuit constants the oscillator can be operated at any frequency in the usual audio range. R1, R2 and C1 form a low-pass type network, while C2C3R3 is high-pass. As the phase shifts are opposite, there is only one frequency at which the total phase shift from collector to base is 180 degrees, and oscillation will occur at this frequency. Optimum operation results when C1 is approximately twice the capacitance of C2 or C3, and R3 has a resistance about 0.1 that of R1 or R2 (C2 = C3 and R1 = R2). Output is taken across C1, where the harmonic distortion is least. A relatively high-impedance load should be used – 0.1 megohm or more.

A small-signal af transistor is suitable for Q1. Either npn or pnp types can be used, with due regard for supply polarity. R4, the collector load resistor, must be large enough for normal amplification, and may be varied somewhat to adjust the operating conditions for best waveform.

RESISTORS AT RADIO FREQUENCIES

Measuring equipment, in some part of its circuit, often requires essentially pure resistance – that is, resistance exhibiting only negligible reactive effects on the frequencies at which measurement is intended. Of the resistors available to amateurs, this requirement is met only by small composition (carbon) resistors. The inductance of wire-wound resistors makes them useless for amateur frequencies.

The reactances to be considered arise from the inherent inductance of the resistor itself and its leads, and from small stray capacitances from one part of the resistor to another and to surrounding conductors. Although both the inductance and capacitance are small, their reactances become increasingly important as the frequency is raised. Small composition resistors, properly mounted, show negligible capacitive reactance up to 100 MHz or so in resistance values up to a few hundred ohms; similarly, the inductive reactance is negligible in values *higher* than a few hundred ohms. The optimum resistance region in this respect is in the 50 to 200-ohm range, approximately.

Proper mounting includes reducing lead length as much as possible, and keeping the resistor separated from other resistors and conductors. Care must also be taken in some applications to ensure that the resistor, with its associated components, does not form a closed loop into which a voltage could be induced magnetically.

So installed, the resistance is essentially pure. In composition resistors the skin effect is very small, and the rf resistance up to vhf is very closely the same as the dc resistance.

Dummy Antennas

A dummy antenna is simply a resistor that, in impedance characteristics, can be substituted for an antenna or transmission line for test purposes. It permits leisurely transmitter testing without radiating a signal. (The amateur regulations strictly limit the amount of "on-the-air" testing that may be done.) It is also useful in testing receivers, in that electrically it resembles an antenna, but does not pick up external noise and signals, a desirable feature in some tests.

For transmitter tests the dummy antenna must be capable of dissipating safely the entire power output of the transmitter. Since for most testing it is desirable that the dummy simulate a perfectly-matched transmission line, it should be a pure resistance, usually of approximately 52 or 73

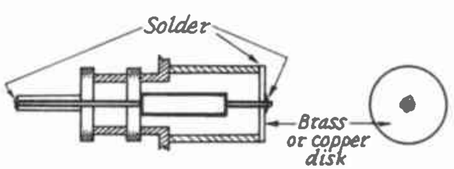


Fig. 17-24 – Dummy antenna made by mounting a composition resistor in a PL-259 coaxial plug. Only the inner portion of the plug is shown; the cap screws on after the assembly is completed.

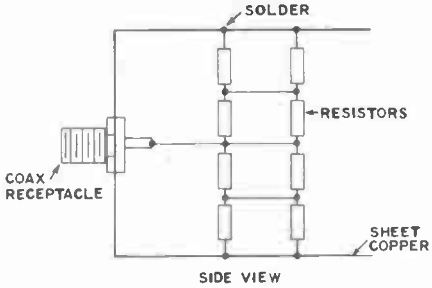


Fig. 17-25 – Using resistors in series-parallel to increase the power rating of a small dummy antenna. Mounted in this way on pieces of flat copper, inductance is reduced to a minimum. Eight 100-ohm 2-watt composition resistors in two groups, each four resistors in parallel, can be connected in series to form a 50-ohm dummy. The open construction shown permits free air circulation. Resistors drawn heavy are in one "deck"; light ones are in the other.

ohms. This is a severe limitation in home construction, because nonreactive resistors of more than a few watts rated safe dissipation are very difficult to obtain. (There are, however, dummy antenna kits available that can handle up to a kilowatt.)

For receiver and minipower transmitter testing an excellent dummy antenna can be made by installing a 51- or 75-ohm composition resistor in a PL-259 fitting as shown in Fig. 17-24. Sizes from one-half to two watts are satisfactory. The disk at the end helps reduce lead inductance and completes the shielding. Dummy antennas made in this way have good characteristics through the vhf bands as well as at all lower frequencies.

Increasing Power Ratings

More power can be handled by using a number of 2-watt resistors in parallel, or series-parallel, but at the expense of introducing some reactance. Nevertheless, if some departure from the ideal impedance characteristics can be tolerated this is a practical method for getting increased dissipations. The principal problem is stray inductance which can be minimized by mounting the resistors on flat copper strips or sheets, as suggested in Fig. 17-25.

The power rating on resistors is a *continuous* rating in free air. In practice, the maximum power dissipated can be increased in proportion to the reduction in duty cycle. Thus with keying, which has a duty cycle of about 1/2, the rating can be doubled. With sideband the duty cycle is usually not over about 1/3. The best way of judging is to feel the resistors occasionally; if too hot to touch, they may be dissipating more power than they are rated for.

THE OSCILLOSCOPE

The electrostatically deflected cathode-ray tube, with appropriate associated equipment, is capable of displaying both low- and radio-frequency signals on its fluorescent screen, in a form which lends itself to ready interpretation. (In contrast, the magnetically deflected television picture tube is not at all suitable for measurement purposes.) In the usual display presentation, the fluorescent spot moves across the screen horizontally at some known rate (horizontal deflection or horizontal sweep) and simultaneously is moved vertically by the signal voltage being examined (vertical deflection). Because of the retentivity of the screen and the eye, a rapidly deflected spot appears as a continuous line. Thus a varying signal voltage causes a pattern to appear on the screen.

Conventionally, oscilloscope circuits are designed so that in vertical deflection the spot moves upward as the signal voltage becomes more positive with respect to ground, and vice versa (there are exceptions, however). Also, the horizontal deflection is such that with an ac sweep voltage – the simplest form – positive is to the right; with a linear sweep – one which moves the spot at a uniform rate across the screen and then at the end of its travel snaps it back very quickly to the starting point – *time* progresses to the right.

Most cathode-ray tubes for oscilloscope work require a deflection amplitude of about 50 volts per inch. For displaying small signals, therefore, considerable amplification is needed. Also, special circuits have to be used for linear deflection. The design of amplifiers and linear deflection circuits is complicated, and extensive texts are available. For checking modulation of transmitters, a principal amateur use of the scope, quite simple circuits suffice. A 60-Hz voltage from the power line makes a satisfactory horizontal sweep, and the voltage required for vertical deflection can easily be obtained from transmitter rf circuits without amplification.

For general measurement purposes amplifiers and linear deflection circuits are needed. The most economical and satisfactory way to obtain a scope having these features is to assemble one of the many kits available.

Simple Oscilloscope Circuit

Fig. 17-26 is an oscilloscope circuit that has all the essentials for modulation monitoring: controls for centering, focusing, and adjusting the brightness of the fluorescent spot; voltage dividers to supply proper electrode potentials to the cathode-ray tube; and means for coupling the vertical and horizontal signals to the deflection plates.

The circuit can be used with electrostatic-deflection tubes from two to five inches in face diameter, with voltages up to 2500. Either set of deflecting electrodes (D1D2, or D3D4) may be used for either horizontal or vertical deflection, depending on how the tube is mounted.

In Fig. 17-26 the centering controls are not too high above electrical ground, so they do not need special insulation. However, the focusing and intensity controls are at a high voltage above ground and therefore should be carefully insulated. Insulated couplings or extension shafts should be used.

The tube should be protected from stray magnetic fields, either by enclosing it in an iron or steel box or by using one of the special CR tube

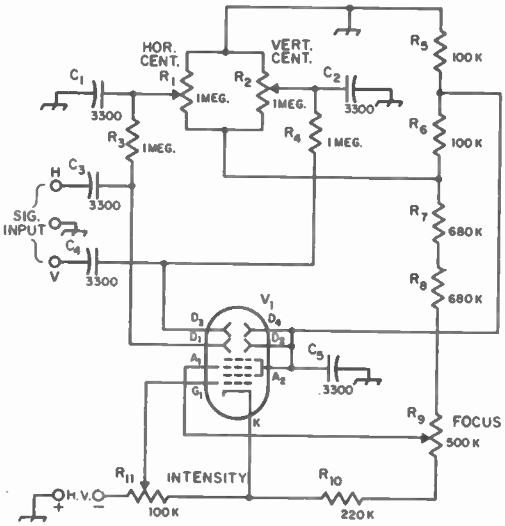


Fig. 17-26 – Oscilloscope circuit for modulation monitoring. Constants are for 1500- to 2500-volt high-voltage supply. For 1000 to 1500 volts, omit R8 and connect the bottom end of R7 to the top end of R9.
C1-C5, incl. – 1000-volt disk ceramic.
R1, R2, R9, R11 – Volume-control type, linear taper. R9 and R11 must be well insulated from chassis.
R3, R4, R5, R6, R10 – 1/2 watt.
R7, R8 – 1 watt.
V1 – Electrostatic-deflection cathode-ray tube, 2- to 5-inch. Base connections and heater ratings vary with type chosen.

shields available. If the heater transformer (or other transformer) is mounted in the same cabinet, care must be used to place it so the stray field around it does not deflect the spot. The spot cannot be focused to a fine point when influenced by a transformer field. The heater transformer must be well insulated, and one side of the heater should be connected to the cathode. The high-voltage dc can be taken from the transmitter plate supply; the current required is negligible.

Methods for connecting the oscilloscope to a transmitter for checking or monitoring modulation are given in earlier chapters.

Quasi-Linear Sweep

For wave-envelope patterns that require a fairly linear horizontal sweep, Fig. 17-27 shows a method of using the substantially linear portion of the 60-Hz sine wave – the "center" portion where the wave goes through zero and reverses polarity. A 60-Hz transformer with a center-tapped secondary winding is required. The voltage should be sufficient to deflect the spot well off the screen on both sides – 250 to 350 volts, usually. With such "over-deflection" the sweep is fairly linear, but it is as bright on retrace as on left-to-right. To blank it in one direction, it is necessary to couple the ac to the No. 1 grid of the CR tube as shown.

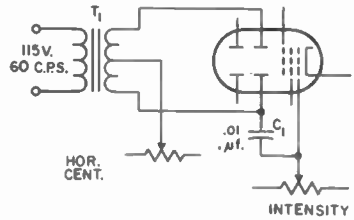


Fig. 17-27 – A quasi-linear time base for an oscilloscope can be obtained from the "center" portion of a sine wave. Coupling the ac to the grid gives intensity modulation that blanks the retrace.
C1 – Ceramic capacitor of adequate voltage rating.
T1 – 250- to 350-volt center-tapped secondary. If voltage is too high, use dropping resistor in primary side.

Lissajous Figures

When sinusoidal ac voltages are applied to both sets of deflecting plates in the oscilloscope the resultant pattern depends on the relative amplitudes, frequencies and phases of the two voltages. If the ratio between the two frequencies is constant and can be expressed in integers a stationary pattern will be produced.

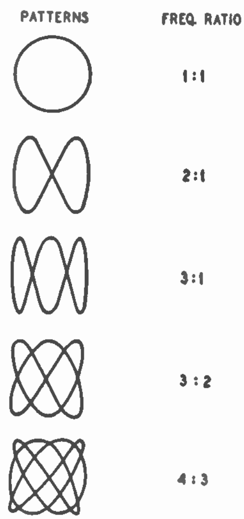


Fig. 17-28 – Lissajous figures and corresponding frequency ratios for a 90-degree phase relationship between the voltages applied to the two sets of deflecting plates.

The stationary patterns obtained in this way are called **Lissajous figures**. Examples of some of the simpler Lissajous figures are given in Fig. 17-28. The frequency ratio is found by counting the number of loops along two adjacent edges. Thus in the third figure from the top there are three loops along a horizontal edge and only one along the vertical, so the ratio of the vertical frequency to the horizontal frequency is 3 to 1. Similarly, in the fifth figure from the top there are four loops along the horizontal edge and three along the vertical edge, giving a ratio of 4 to 3. Assuming that the known frequency is applied to the horizontal plates, the unknown frequency is

$$f2 = \frac{n2}{n1} \; f1$$

where $f1$ = known frequency applied to horizontal plates,
$f2$ = unknown frequency applied to vertical plates,
$n1$ = number of loops along a vertical edge and,
$n2$ = number of loops along a horizontal edge

An important application of Lissajous figures is in the calibration of audio-frequency signal generators. For very low frequencies the 60-Hz power-line frequency is held accurately enough to be used as a standard in most localities. The medium audio-frequency range can be covered by comparison with the 440- and 600-Hz modulation on the WWV transmissions. It is possible to calibrate over a 10-to-1 range, both upwards and downwards, from each of the latter frequencies and thus cover the audio range useful for voice communication.

An oscilloscope having both horizontal and vertical amplifiers is desirable, since it is convenient to have a means for adjusting the voltages applied to the deflection plates to secure a suitable pattern size.

MARKER GENERATOR FOR 100, 50 AND 25 KHZ

The frequency generator in the accompanying illustrations will deliver marker signals of usable strength well into the vhf region when its output is connected to the antenna input terminals of a communications receiver. It uses a 100-kHz crystal in an integrated-circuit version of the solid-state multivibrator oscillator shown earlier. The oscillator is followed by a two-stage IC divider which produces 50- and 25-kHz marker intervals. Two inexpensive ICs are used, an MC-724P quad gate and an MC790P dual JK flip-flop. Two of the gates in the MC724P are used for the oscillator and a third serves as a following buffer amplifier and "squarer" for driving the first divide-by-2 circuit in the MC790P. This divider then drives the second divide-by-2 flip-flop. Outputs at the three frequencies are taken through a 3-position switch from taps as shown in the circuit diagram, Fig. 17-30.

Two of the three poles of the 4-position switch are used for controlling the collector voltage for the ICs. Voltage is on the MC724P in all active positions of the switch, but is applied to the MC790P only when 50- and 25-kHz markers are required. This saves battery power, since the MC790P takes considerably more current than the MC724P.

The outputs on all three frequencies are good square waves. To assure reasonably constant harmonic strength through the hf spectrum the output is coupled to the receiver through a small capacitance which tends to attenuate the lower-frequency harmonics. This capacitance, C3, is not critical as to value and may be varied to suit individual preferences. The value shown, 22 pF, is satisfactory for working into a receiver having an input impedance of 50 ohms.

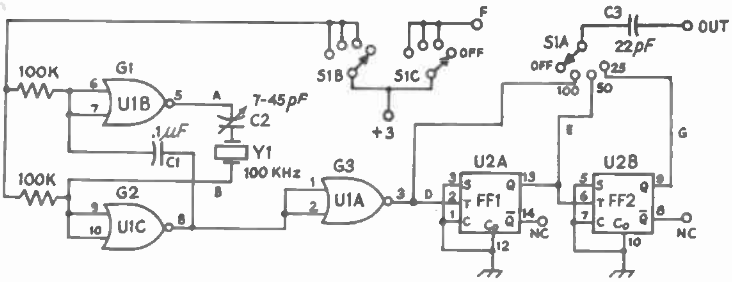
At 3 volts dc input the current taken in the 100-kHz position of S1 is 8 mA. In the 50- and 25-kHz positions the total current (both ICs) is 35 mA. The generator continues to work satisfactorily when the voltage drops as low as 1.5 volts. The oscillator frequency is subject to change as the voltage is lowered, the frequency shift amounting to approximately 30 Hz at 15 MHz on going from 3 to 2 volts. There is a slight frequency shift between the 100-kHz and 50/25-kHz positions, but this amounts to only 6 or 7 Hz at 15 MHz.

Fig. 17-29 – Frequency marker generating 100-, 50-, or 25-kHz intervals. Battery power supply (two "D" cells) is inside the cabinet, a 3 X 4 X 6-inch aluminum chassis with bottom plate. The trimmer capacitor for fine adjustment of frequency is available through the hole in the top near the left front.

Frequency changes resulting from temperature variations are larger; they may be as much as a few hundred Hz at 15 MHz in normal room-temperature variations. All such frequency changes can be compensated for by adjusting C2, and it is good practice to check the frequency occasionally against one of the WWV transmissions, readjusting C2 if necessary.

Layout and Construction

The physical layout of the circuit can be varied to suit the builder's tastes. The size of the box containing the generator shown in the photographs makes the batteries easily accessible for replacement. The method of mounting the crystal and C2 allows the latter to be reached through the top of the box for screwdriver adjustment, and makes possible the easy removal of the crystal since it plugs into a standard crystal socket. There is ample room for soldering the various wires that lead to the switch from the etched board on which the ICs, resistors, and C1 are mounted. The output



C1 – 0.1μF paper, low voltage.
C2 – 7-45-pF ceramic trimmer.
C3 – 22-pF dipped mica (ceramic also satisfactory).
S1 – 3-pole, 4-position rotary (Mallory 3134J).
U1 – Quad 2-input NOR gate, 1 section unused (Motorola MC724P).
U2 – Dual J-K flip-flop (Motorola MC790P).

Fig. 17-30 – Marker generator circuit. Pin 4 of both ICs is grounded. Connect pin 11 of U1 to point C, and pin 11 of U2 to point F.

jack is placed at the rear where it is convenient when the unit is alongside a receiver.

An etched board does not have to be used for wiring the ICs and associated parts, although it makes for neatness in construction. The wiring plan used in this one is shown in Fig. 17-32. Fig. 17-32 is not a conventional template, but is a scale drawing showing how the etched connections can run with a minimum number of cross-over points where jumpers are required (only one is needed in this layout). In following the wiring plan the resist can be put on as desired, so long as the separation between conductors is great enough to prevent short-circuits.

Fig. 17-32 shows the front or component side of the board. To get the reversed drawing that would be followed on the copper side, place a piece of paper under the figure, with a face-up piece of carbon paper under it. Then trace the wiring with a sharp pencil and the layout will be transferred to the back of the paper. The points where holes are to be drilled are shown by small dots and circles, the latter indicating the points at which external connections are to be made.

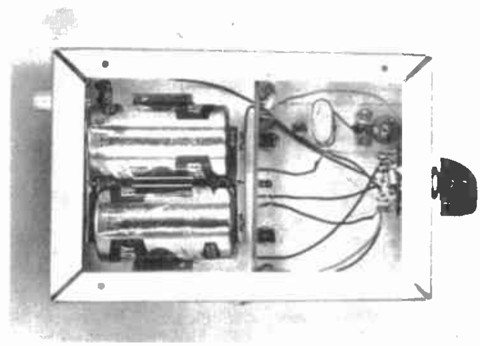


Fig. 17-31 – Integrated circuits and associated fixed capacitors and resistors are mounted on an etched broad measuring 3 3/4 X 2 1/2 inches, supported from one wall by an aluminum bracket. The 100-kHz crystal and trimmer capacitor are on a 1 X 2-inch plastic strip supported below the top on 1/2-inch spacers, with the capacitor facing upward so it can be adjusted from outside. The two dry cells are in a dual holder (available from electronics supply stores). The output connector is a phono jack, mounted on the rear wall (upper left in this view) with C3.

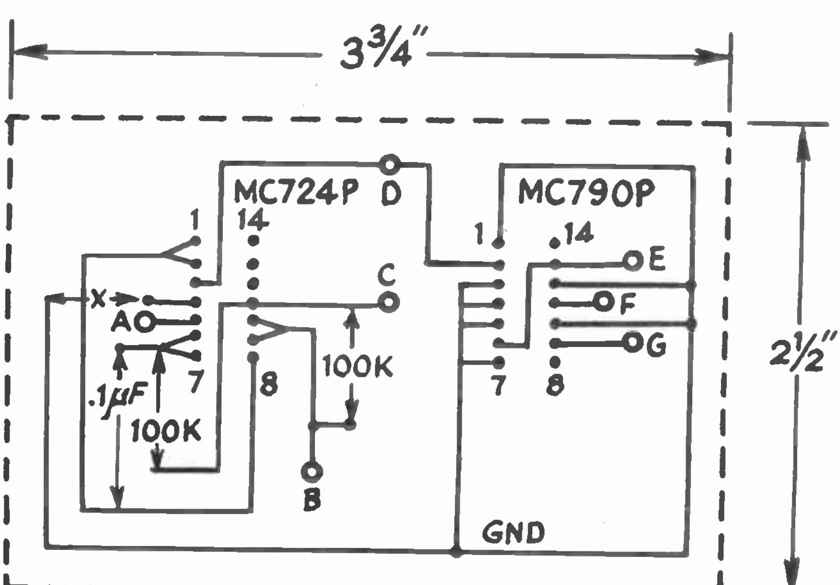
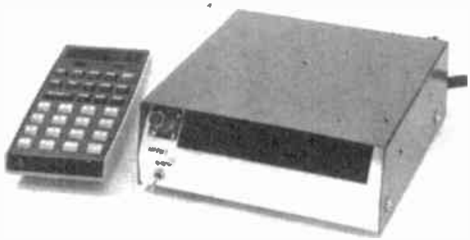


Fig. 17-32 – Wiring plan for the circuit board, component side. Dimensions for placement of parts are exact. X – jumper. Other letters indicate external connection points, corresponding to similarly lettered connections in Fig. 17-30.

50-MHZ FREQUENCY COUNTER



The counter is built into a homemade enclosure that measures 2 X 6 X 6 inches.

As the complexity of amateur radio equipment increases, the sophistication of the test equipment needed to effectively troubleshoot failures also increases. The following counter was developed to fulfill this need and will also certainly complement the station itself. This counter as a whole is very basic in design and some interesting features are its size and the displays. Each display (except for the least significant digit) contains four sections: The counter, latch, decoder and the display. The LSD does not have a counter included.

Circuit Description

The counter uses a crystal-controlled time base to generate the gating pulses. A 1-MHz oscillator is counted down to provide 1 kHz for the SN7493 (U6). U6 is connected to divide by twelve. The outputs of this divide-by-twelve counter are gated to provide the count, latch and reset gates. The timing relationships are shown in Fig. 2.

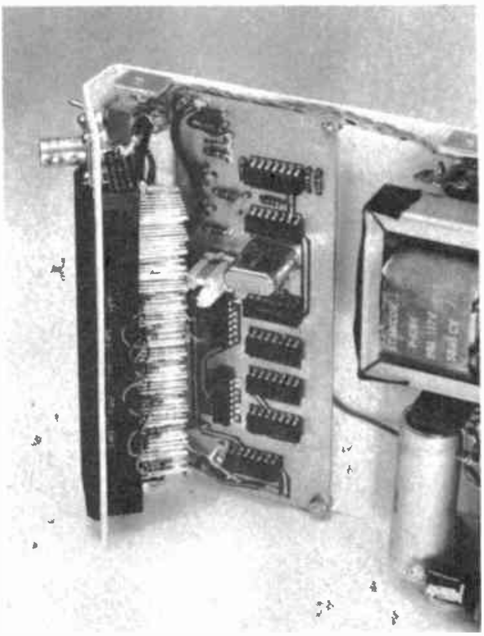
Displays are Texas Instruments TIL 306 devices and a single TIL 308. Each TIL 306 contains the four units necessary to display a counter frequency. The internal counter has an upper frequency limit of approximately 18 MHz. Following the counter is a latch, which is used to hold the data while the counter is operating. The frequency information, in BCD format, is decoded and it then drives the correct segments of the display. The maximum count output is used to drive the successive displays. Each display contains a feature called ripple blanking. If the number zero is detected in the latches and ripple blanking has been enabled, the display will be blanked. This function was incorporated to give leading zero blanking in the counter. Starting from left to right (MSD to LSD), if zero is detected that display will be blanked and the blanked data will be passed to the right. This means that 455.2 kHz will be displayed as 455.2 not 00455.2. The LSD is not connected for leading zero blanking. It was a bit disconcerting to turn the counter on and not have the display light up. Therefore, the LSD was not connected for leading zero blanking and zero will be displayed when the counter is on with no input.

The LSD, a TIL 308, does not have the internal counter. This allows a faster counter, greater than 18 MHz, to be used for the LSD. The N82S90, used as the counter for the LSD, is rated for 100-MHz operation.

Signal input is applied to a source follower, Q1. For the sake of simplicity, it was decided to use this form of input configuration instead of an amplifier for the input. An input signal of 0.25 V will be sufficient to trigger the counter, up to 50 MHz. Following the input network is an SN74S00 connected to act as a level translator. The signal from the input network is made TTL compatible by this circuit.

Construction

One of the features of this counter is that its construction is not critical. As can be seen in the photograph, the entire counter is built on a small



The use of a bezel and a single pc board allows very compact packaging in this counter. Leads that interconnect between the bezel and the circuit board have been made extra long. This excess length is tucked under the bezel and allows the pc board to be removed and worked on while still connected to the counter.

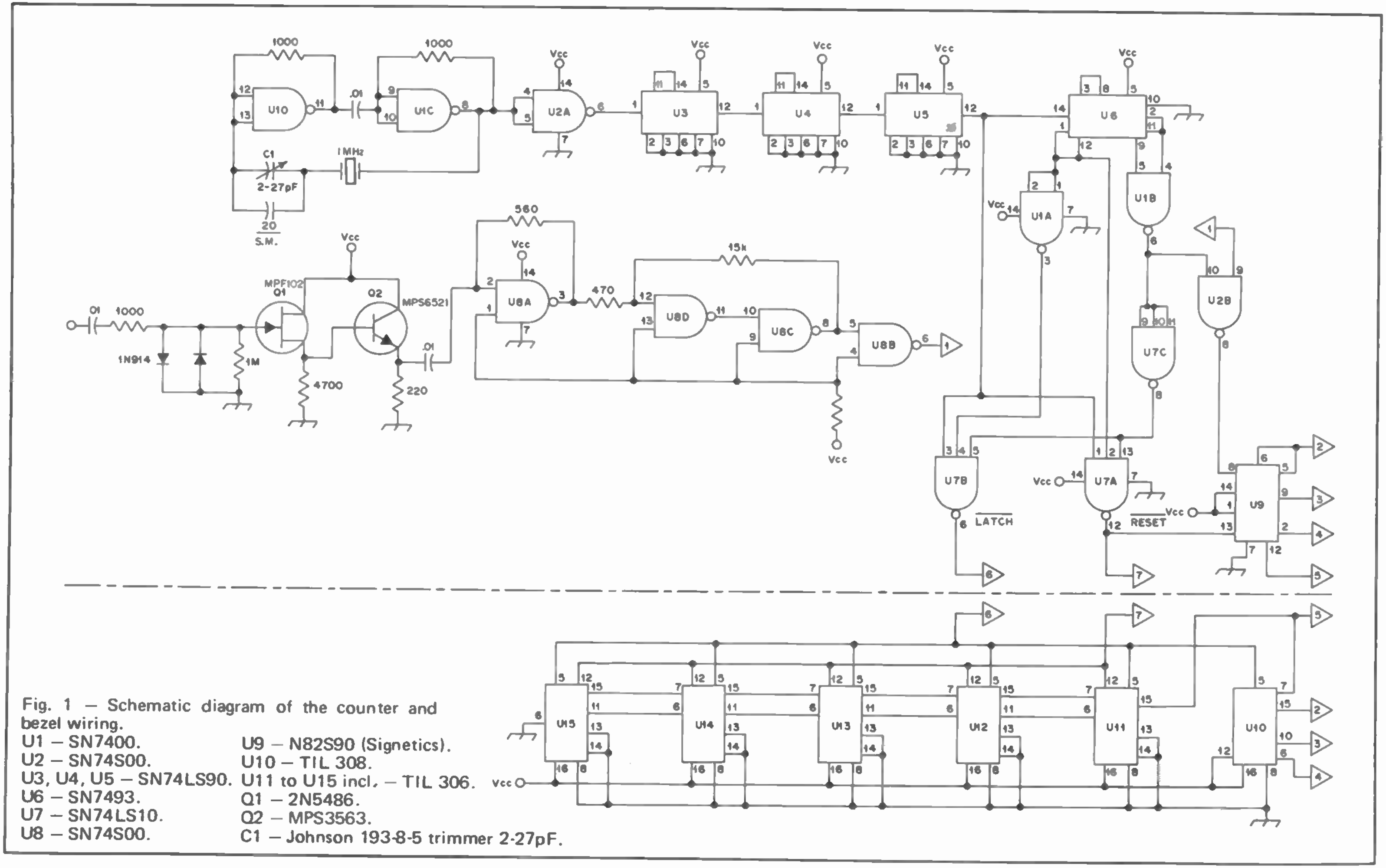


Fig. 1 — Schematic diagram of the counter and bezel wiring.

U1 — SN7400.
U2 — SN74S00.
U3, U4, U5 — SN74LS90.
U6 — SN7493.
U7 — SN74LS10.
U8 — SN74S00.
U9 — N82S90 (Signetics).
U10 — TIL 308.
U11 to U15 incl. — TIL 306.
Q1 — 2N5486.
Q2 — MPS3563.
C1 — Johnson 193-8-5 trimmer 2-27pF.

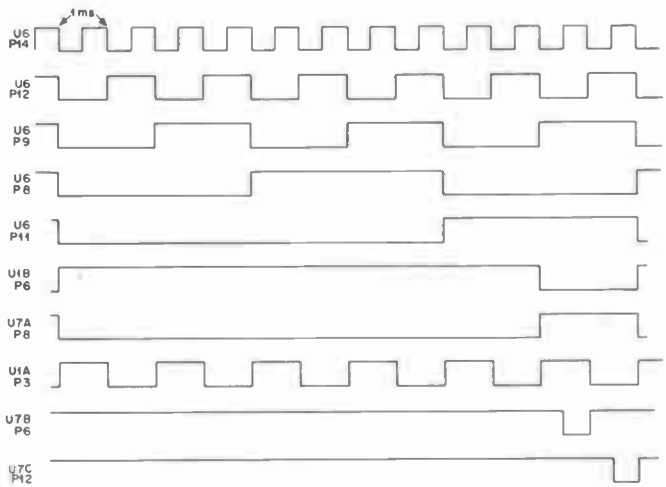


Fig. 2 — Timing diagram of the counter.

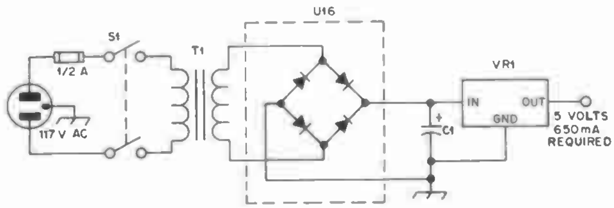


Fig. 3 — Schematic diagram of the power supply.
C1 — 2000 μF, 25 V.
T1 — 12.6 Vac, 1 A.
U16 — 50 PRV, 1-A encapsulated bridge rectifier.
VR1 — 5-V, 1-A voltage regulator LM309K.

circuit board and in the bezel. Instead of using a bezel, a circuit board could be etched and mounted behind the front panel. The foil on the top of the pc board is the ground interconnection for all the integrated circuits and the builder should liberally install bypass capacitors on the 5-volt line. This will prevent any transients on the line from showing up in the counted frequency. Construction techniques and added features are left entirely to individual preferences.

Operation and Adjustment

The only adjustment required is the crystal oscillator. This should be checked and set against a frequency standard. An error as small as 100 Hz in the crystal frequency can cause a frequency measurement to be off as much as 5 kHz at 50 MHz. This counter will function with signals as high as 65 MHz in its present configuration. The limiting factors are the selection of integrated circuits and the input network.

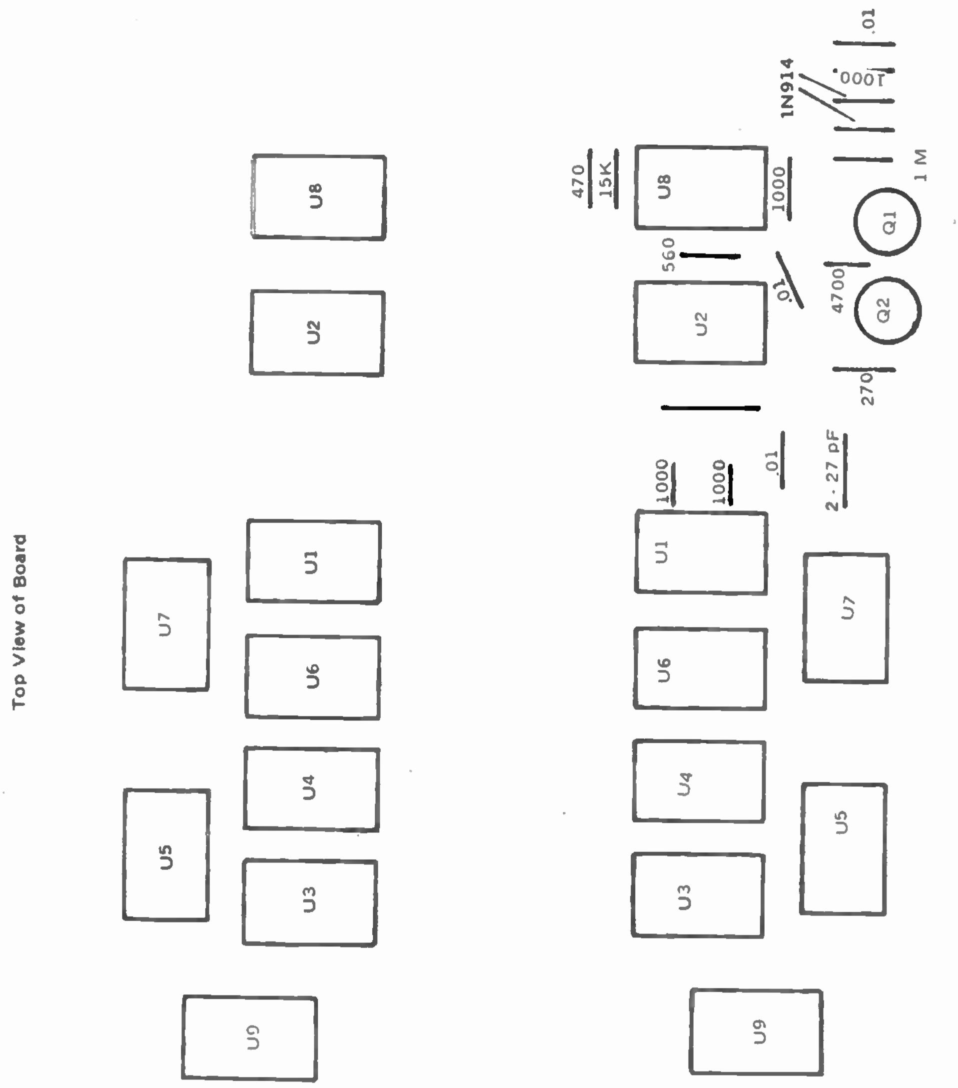


Fig. 4 – Template and parts layout for the counter (full scale). Note that double-sided pc board is required. Some of the holes for the discrete components will be surrounded by foil on the top of the board. Using a No. 33 drill slightly countersink the top of the board to remove the copper around the hole.

A CALIBRATED FIELD STRENGTH METER

There are many occasions when it is desirable to determine the relative performance of an antenna. While near-field pattern measurements are generally not accurate, they do show trends in terms of front-to-back ratio and may be used to determine what adjustments, if any, should be made to an existing system. The field-strength meter described here will detect large as well as small changes in radiated power from an antenna. For instance, the pattern of an hf-band Yagi may be checked by placing the meter and an associated sampling antenna several hundred feet from the beam. A watt or two of power is needed to make tests above 21 MHz, but for frequencies below this point, a grid-dip oscillator may serve as a "transmitter."

Fig. 1 gives the circuit diagram of the calibrated field-strength meter. L1 and L2 are resonated to the desired frequency with C1 to tune the hf bands. Adjustment is made to produce maximum meter deflection of the signal being sampled. Should the signal cause the meter to deflect off scale, the attenuator, R4, may be reset to reduce the level of the incoming energy.

Two operational amplifiers comprise a logarithmic circuit which produces a voltage output at pin 10 of U1B that is proportional to the logarithm (thus dB) of the input voltage. Forward bias is applied to CR1 via a 1-megohm resistor to improve conductivity at low signal input values. The output voltage from U1B is displayed by M1, a conventional milliammeter. Two scale ranges are available, 20 dB and 40 dB. With no signal applied, a small amount of quiescent current will appear on

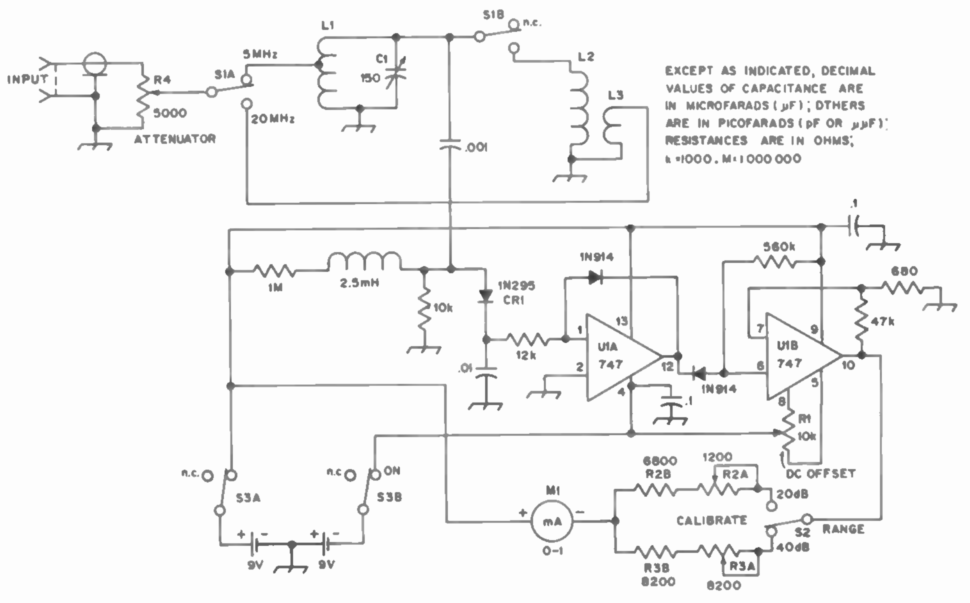
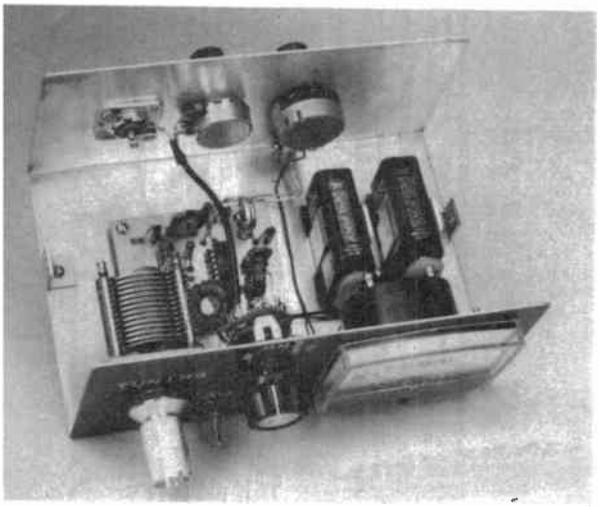


Fig. 1 — Circuit diagram for the calibrated field strength meter. Component designations not listed below are for text reference.

C1 — Variable capacitor, 140 pF maximum.
L1 — 44 turns of No. 24 enam. on a T-68-2 core tapped four turns from the ground end.
L2 — 15 turns of No. 24 enam on a T-68-2 core.
L3 — Two turns of No. 24 enam. wound over L2.
U1 — Dual 747 operational amplifier.
S1 — Dpdt rotary.
S2, S3 — Miniature toggle.



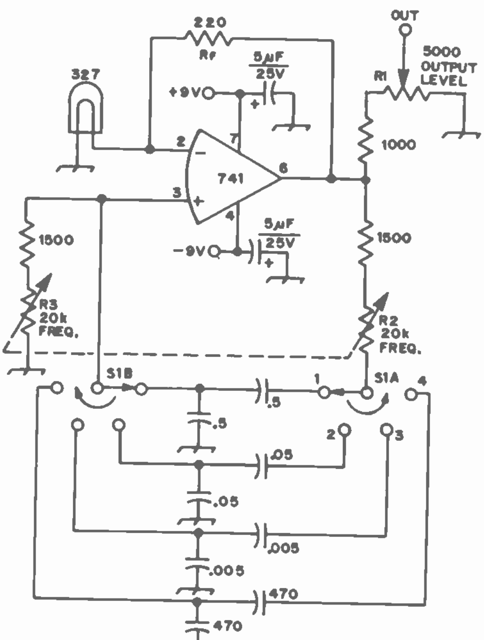
Inside view of the field strength meter. Most of the components are mounted on a circuit board.

M1. Readings made near this level will not be quite as accurate as those made in the upper portion of the scale. Accuracy is within one dB. About 1000 microvolts of signal is necessary to provide a meaningful movement of M1. R1 is the dc offset control and is mounted on the rear panel. It permits some variation of the absolute readings by shifting the dc levels at the output of U1B and may be used to set the meter to some convenient reference mark. The combined values of R2A and R2B should be 8000 ohms. R2A is a trim pot to allow proper adjustment to exactly that value. R3A and R3B serve a similar purpose and should be set for a total resistance of 16 kΩ.

AN AUDIO OSCILLATOR

A wide-range audio oscillator that will provide a moderate output level can be built from a single 741 operational amplifier (Fig. 1). Power is supplied by two nine-volt batteries, from which the circuit draws 4 mA. The frequency range is selectable from 15 Hz to 150 kHz, although a 1.5- to 15-Hz range can be included with the addition of two 5-µF nonpolarized capacitors and an extra switch position. Distortion is approximately one percent. The output level under a light load (10 kΩ) is 4 to 5 volts. This can be increased by using higher battery voltages, up to a maximum of plus and minus 18 volts, with a corresponding adjustment of R_f.

Pin connections shown are for the TO-5 case. If another package configuration is used, the pin connections may be different. R_f (220 Ω) is trimmed for an output level about five percent below clipping. This should be done for the temperature at which the oscillator will normally operate, as the lamp is sensitive to ambient temperature. Note that the output of this oscillator is direct coupled. If you are connecting this unit into circuits where dc voltage is present, use a coupling capacitor. As with any solid-state equipment, be cautious around plate circuits of tube-type equipment, as the voltage spike caused by charging a coupling capacitor may destroy the IC. This unit was originally described by Schultz in *QST* for November, 1974.



SWITCH 1	
POS.	FREQ.
1	15-150Hz
2	150-1500Hz
3	1500Hz-15kHz
4	15-150kHz

Fig. 1 — A simple audio oscillator that provides a selectable frequency range. R2 and R3 control the frequency and R1 varies the output level.

A TESTER FOR FET AND BIPOLAR TRANSISTORS

The circuit shown is intended solely as a tester for npn and pnp transistors, junction FETs, and dual-gate MOSETs. This equipment is not for use in checking audio or high-power rf transistors.

The circuit of Fig. 1 is an oscillator which is wired so that it will test various small-signal transistors by switching the battery polarity and bias voltage. A crystal for the upper range of the hf spectrum is wired into the circuit permanently, but could be installed in a crystal socket if the builder so desires. A 20-MHz crystal was chosen for this model. Any hf crystal cut for fundamental mode operation can be used.

When testing FETs the bias switch, S3, is placed in the FET position, thus removing R2 from the circuit. However, when testing bipolar transistors the switch position must be changed to BIPOL so that forward bias can be applied to the base of the bipolar transistor under test. R1 is always in the circuit, and serves as a gate-leak resistor for FETs being evaluated. It becomes part of the bias network when bipolars are under test. C1 is used for feedback in combination with the internal capacitances of the transistors being checked. Its value may have to be changed experimentally if crystals for lower frequencies are utilized in the circuit. Generally speaking, the lower the crystal frequency, the greater the amount of capacitance needed to assure oscillation. Use only that amount necessary to provide quick starting of the oscillator.

Components R3 and R4 are used as a voltage divider to provide bias for dual-gate MOSFETs. C2 is kept small in value to minimize loading of the oscillator by the low-impedance voltage doubler, CR1 and CR2. Rectified rf from the oscillator is monitored on M1. Meter deflection is regulated manually by means of control R5. S1 is used to select the desired supply voltage polarity – negative ground for testing n-channel FETs and npn bipolars, and a positive ground when working with p-channel and pnp devices.

When testing MOSFETs that are not gate protected (3N140 for one), make certain that the transistor leads are shorted together until the device is seated in the test socket. Static charges on one's hands can be sufficiently great to damage the insulation within the transistor. Use a single strand of wire from some No. 22 or 24 stranded hookup wire, wrapping it two or three times around the pigtails of the FET as close to the transistor body as possible. After the FET is plugged into the

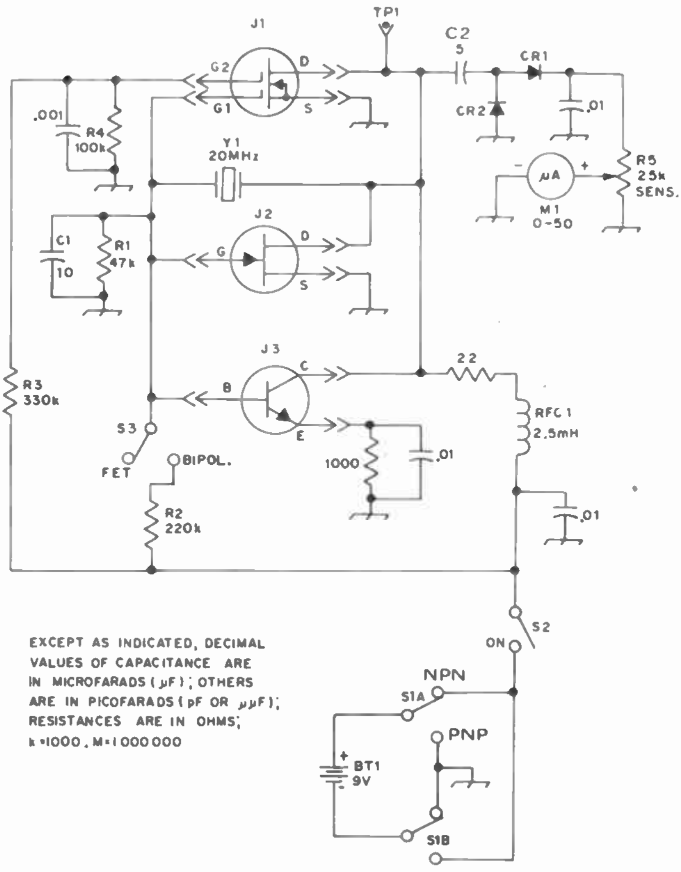


Fig. 1 – Schematic diagram of the transistor tester. Capacitors are disk ceramic or mica. Resistors are 1/2 or 1/4-watt composition execet for R5. Estimated cost for this tester (all parts new) is $15. Numbered components not appearing in parts list are so designated for text discussion.

- 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 (see tex).

socket, unwrap the wire and perform the tests. (It's not a bad idea to have an earth ground connected to the case of the tester when checking unprotected FETs.) Put the shorting wire back on the FET leads before removing the unit from the tester.

The meter indication is significant in checking any type of transistor. If the device is open, shorted, or extremely leaky, no oscillation will take place, and the meter will not deflect. The higher the meter reading, the greater the vigor of the transistor at the operating frequency. High meter readings suggest that the transistor is made for vhf or uhf service, and that its beta is medium to high. Lower readings may indicate that the transistor is designed for hf use, or that it has very low gain. Transistors that are known to be good but will not cause the circuit to oscillate are most likely made for low-frequency or audio applications.

A TESTER FOR CRYSTALS AND BIPOLAR TRANSISTORS

The circuit of Fig. 1 is intended primarily to test surplus crystals and bipolar transistors. It uses a Pierce oscillator. Battery polarity can be switched to allow testing of npn or pnp transistors. Crystal quality is indicated on M1. The greater the crystal activity, the higher the meter reading. A suitable transistor for use at Q1 (when testing crystals) is the 2N4124, MPS3563, or HEP53. All three have f_T ratings well into the vhf spectrum, and each has reasonably high beta. The two characteristics make the devices ideal as general-purpose oscillators.

This tester will work well from the upper hf range down to at least 455 kHz. S1 is used to change the value of feedback capacitance. The lower the frequency of operation, the greater the amount of capacitance required.

A transistor can be checked by plugging the unknown type into the panel socket while using a crystal of known frequency and condition. Both testers can be used as calibrators by inserting

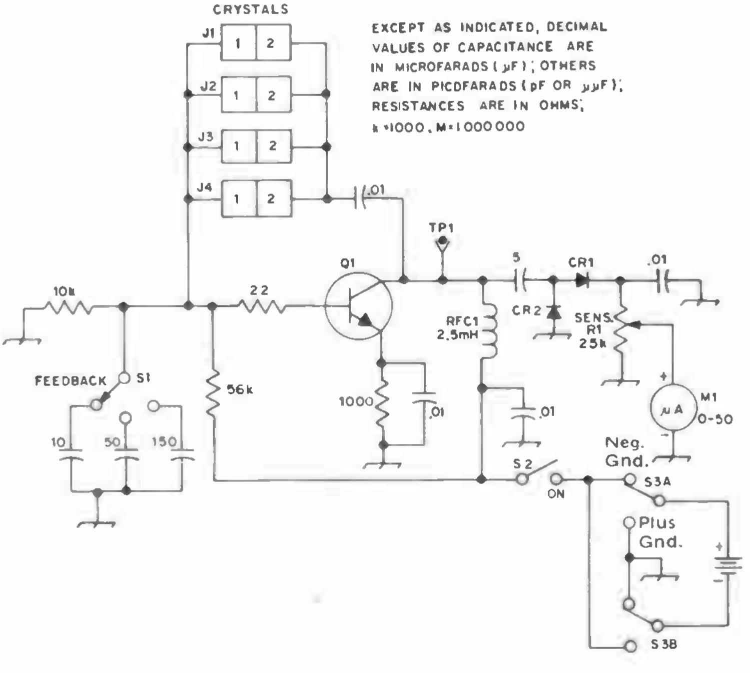


Fig. 1 — Schematic diagram of the No. 2 tester. Capacitors are disk ceramic. Fixed-value resistors are 1/2 or 1/4-watt composition. Estimated cost for this tester (all new parts) is $13.

BT1 — Small 9-V transistor-radio battery.
CR1, CR2 — 1N34A germanium diode or equiv.
J1-J4, incl. — Crystal socket of builder's choice.
M1 — Microampere meter. Calectro D1-910 used here.
R1 — 25,000-ohm linear-taper composition control with switch.
RFC1 — 2.5-mH rf choke.
S1 — Single-pole three-position phenolic rotary wafer type, miniature.
S2 — Part of R1.
S3 — Double-pole double-throw miniature toggle.
Q1 — Vhf npn bipolar, 2N4124, MPS3563, HEP53.

crystals for band-edge checking. The frequencies of unknown crystals can be checked by listening to the output from the test oscillators on a calibrated receiver or while using a frequency counter connected to the designated test point

Four crystal sockets are provided in the model shown here. J1 through J4 provide for testing of FT-243, HC-6/U, HC-17, and HC-25 crystals, the most popular holder styles in use today. Other types can be added by the builder if desired.

DIODE NOISE GENERATORS

A noise generator is a device for creating a controllable amount of rf noise ("hiss"-type noise) evenly distributed throughout the spectrum of interest. The simplest type of noise generator is a diode, either vacuum-tube or crystal, with dc flowing through it. The current is also made to flow through a load resistance which usually is chosen to equal the characteristic impedance of the transmission line to be connected to the receiver's input terminals. The resistance then substitutes for the line, and the amount of rf noise fed to receiver input is controlled by varying the dc through the diode.

The noise generator is useful for adjusting the "front-end" circuits of a receiver for best noise figure. A simple circuit using a crystal diode is shown in Fig. 17-51. The unit can be built into a small metal box; the main consideration is that the circuit from C1 through P1 be as compact as possible. A calibrated knob on R1 will permit resetting the generator to roughly the same spot each time, for making comparisons. If the leads are short, the generator can be used through the 144-MHz band for receiver comparisons.

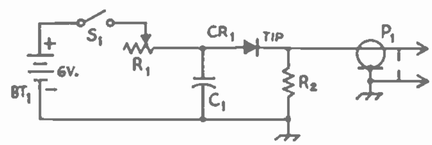


Fig. 17-51 — Circuit of a simple crystal-diode noise generator.
BT1 — Dry-cell battery, any convenient type.
C1 — 500-pF ceramic, disk or tubular.
CR1 — Silicon diode, 1N21 or 1N23. Diodes with "R" suffix have reversed polarity. (Do not use ordinary germanium diodes.)
P1 — Coaxial fitting, cable type.
R1 — 50,000-ohm control, ccw logarithmic taper.
R2 — 51 or 75 ohms, 1/2-watt composition.
S1 — Spst toggle (may be mounted on R1).

To use the generator, screw the coaxial plug onto the receiver's input fitting, open S1, and measure the noise output of the receiver by connecting an audio-frequency voltmeter to the receiver's af output terminals. An average-reading voltmeter is preferable to the peak-reading type, since on this type of noise the average-reading meter will give a fair approximation of rms, and the object is to measure noise *power*, not voltage.

In using the generator for adjusting the input circuit of a receiver for optimum noise figure, first make sure that the receiver's rf and af gain controls are set well within the linear range of response, and turn off the automatic gain control. With the noise generator connected but S1 open, adjust the receiver gain controls for an output reading that is far enough below the maximum obtainable to ensure that the receiver is operating linearly. This is your reference level of noise. Then close S1 and adjust R1 for a readily perceptible increase in output. Note the *ratio* of the two readings – i.e., the number of dB increase in noise when the generator is on. Then make experimental adjustments of the receiver input coupling, always with the object of obtaining the largest number of dB increase in output when the generator is switched on.

A simple crystal-diode noise generator is a useful device for the receiver adjustment, especially at vhf, and for comparing the performance of different receivers checked with the same instrument. It does not permit actual measurement of the noise figure, however, and therefore the results with one instrument cannot readily be compared with the readings obtained with another. In order to get a quantitative measure of noise figure it is necessary to use a temperature-saturated vacuum diode in place of the semiconductor diode. Suitable diodes are difficult to find.

RF PROBE FOR ELECTRONIC VOLTMETERS

The rf probe shown in Figs. 17-52 to 17-55, inclusive, uses the circuit discussed earlier in connection with Fig. 17-15.

The isolation capacitor, C1, crystal diode, and filter/divider resistor are mounted on a bakelite 5-lug terminal strip, as shown in Fig. 17-55. One end lug should be rotated 90 degrees so that it extends off the end of the strip. All other lugs should be cut off flush with the edge of the strip. Where the inner conductor connects to the terminal lug, unravel the shield three-quarters of an inch, slip a piece of spaghetti over it, and then solder the braid to the ground lug on the terminal strip. Remove the spring from the tube shield, slide it over the cable, and crimp it to the remaining quarter inch of shield braid. Solder both the spring and a 12-inch length of flexible copper braid to the shield.

Next, cut off the pins on a seven-pin miniature shield-base tube socket. Use a socket with a cylindrical center post. Crimp the terminal lug previously bent out at the end of the strip and insert it into the center post of the tube socket from the top. Insert the end of a phone tip or a

Fig. 17-52 — Rf probe for use with an electronic voltmeter. The case of the probe is constructed from a 7-pin ceramic tube socket and a 2 1/4-inch tube shield. A half-inch grommet at the top of the tube shield prevents the output lead from chafing. A flexible copper-braid grounding lead and alligator clip provide a low-inductance return path from the test circuit.

Fig. 17-54 — Inside the probe. The 1N34A diode, calibrating resistor, and input capacitor are mounted tight to the terminal strip with shortest leads possible. Spaghetti tubing is placed on the diode leads to prevent accidental short circuits. The tube-shield spring and flexible-copper grounding lead are soldered to the cable braid (the cable is RG-58/U coax). The tip can be either a phone tip or a short pointed piece of heavy wire.

pointed piece of heavy wire into the bottom of the tube socket center post, and solder the lug and tip to the center post. Insert a half-inch grommet at the top of the tube shield, and slide the shield over the cable and flexible braid down onto the tube socket. The spring should make good contact with the tube shield to insure that the tube shield (probe case) is grounded. Solder an alligator clip to the other end of the flexible braid and mount a phone plug on the free end of the shielded wire.

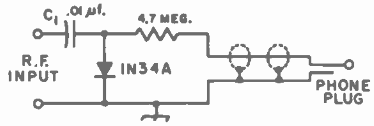


Fig. 17-53 — The rf probe circuit.

Mount components close to the terminal strip, to keep lead lengths as short as possible and minimize stray capacitance. Use spaghetti over all wires to prevent accidental shorts.

The phone plug on the probe cable plugs into the dc input jack of the electronic voltmeter and rms voltages are read on the voltmeter's negative dc scale.

The accuracy of the probe is within ±10 percent from 50 kHz to 250 MHz. The approximate input impedance is 6000 ohms shunted by 1.75 pF (at 200 MHz).

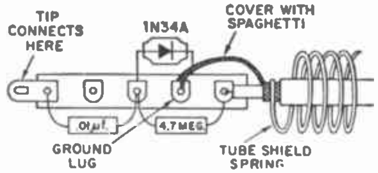


Fig. 17-55 — Component mounting details.

RF IMPEDANCE BRIDGE FOR COAX LINES

The bridge shown in Figs. 1 through 3 may be used to measure unknown complex impedances at frequencies below 30 MHz. Measured values are of equivalent series form, $R + jX$. The useful range of the instrument is from about 5 to 400 ohms if the unknown load is purely resistive, or 10 to 150 ohms resistive component in the presence of reactance. The reactance range is from 0 to approximately 100 ohms for either inductive or capacitive loads. Although the instrument cannot indicate impedances with the accuracy of a laboratory type of bridge, its readings are quite adequate for the measurement and adjustment of antenna systems for amateur use, including the taking of line lengths into account with a Smith chart or Smith transmission-line calculator.

The bridge incorporates a differential capacitor, C1, to obtain an adjustable ratio for measurement of the resistive component of the load. The capacitor consists of two identical sections on the same frame, arranged so that when the shaft is rotated to increase the capacitance of one section, the capacitance of the other section decreases. The capacitor is adjusted for a null reading on M1, and its settings are calibrated in terms of resistance at J3 so the unknown value can be read off the calibration. A coil-and-capacitor combination is used to determine the amount and type of reactance, inductive or capacitive. L1 and C2 in the bridge circuit are connected in series with the load. The instrument is initially balanced at the frequency of measurement with a purely resistive load connected at J3, so that the reactances of L1 and of C2 at its midsetting are equal. Thus, these reactances cancel each other in this arm of the bridge. With an unknown complex-impedance load then connected at J3, the setting of C2 is varied either to increase or decrease the capacitive reactance, as required, to cancel any reactance present in the load. If the load is inductive more capacitive reactance is required from C2 to obtain a balance, indicated by a null on M1, with less reactance needed from C2 if the load is capacitive. The settings of C2 are calibrated in terms of the value and type of reactance at J3. Because of the relationship of capacitive reactance to frequency, the calibration for the dial of C2 is valid at only one frequency. It is therefore convenient to calibrate this dial for equivalent reactances at 1 MHz, as shown in Fig. 4. Frequency corrections may then be made simply by dividing the reactance dial reading by the measurement frequency in megahertz.

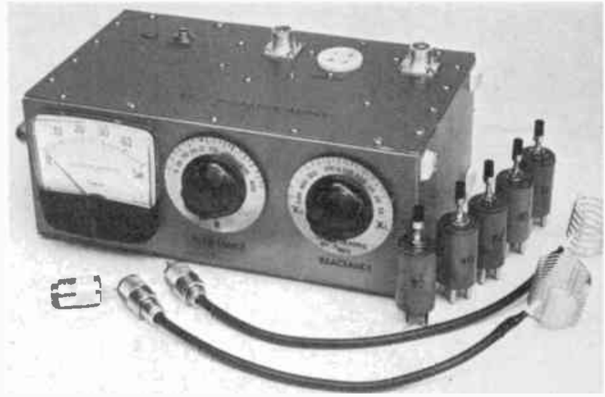


Fig. 1 — An *RCL* bridge for measuring unknown values of complex impedances. A plug-in coil is used for each frequency band. The bridge operates at an rf input level of about 5 volts; pickup-link assemblies for use with a grid-dip oscillator are shown. Before measurements are made, the bridge must be balanced with a nonreactive load connected at its measurement terminals. This load consists of a resistor mounted inside a coaxial plug, shown in front of the instrument at the left. The aluminum box measures 4 1/4 X 10 3/4 X 6 1/8 inches and is fitted with a carrying handle on the left end and self-sticking rubber feet on the right end and bottom. Dials are Millen No. 10009 with skirts reversed and calibrations added.

Construction

In any rf-bridge type of instrument, the leads must be kept as short as possible to reduce stray reactances. Placement of component parts, while not critical, must be such that lead lengths greater than about 1/2 inch (except in the dc metering circuit) are avoided. Shorter leads are desirable, especially for R1, the "standard" resistor for the bridge. In the unit photographed, the body of this resistor just fits between the terminals of C1 and J2 where it is connected. C1 should be enclosed in a shield and connections made with leads passing through holes drilled through the shield wall. The frames of both variable capacitors, C1 and C2, must be insulated from the chassis, with insulated couplings used on the shafts. The capacitor specified for C1 has provisions for insulated mounting. C2 is mounted on 1-inch ceramic insulating pillars.

Band-switching arrangements for L1 complicate the construction and contribute to stray reactances in the bridge circuit. For these reasons plug-in coils are used at L1, one coil for each band over which the instrument is used. The coils must be adjustable, to permit initial balancing of the bridge with C2 set at the zero-reactance calibration point. Coil data are given in Table I. Millen 45004 coil forms with the coils supported inside provide a convenient method of constructing these slug-tuned plug-in coils. A phenolic washer cut to the proper diameter is epoxied to the top or open end of each form, giving a rigid support for mounting of the coil by its bushing. Small knobs for 1/8-inch shafts, threaded with a No. 6-32 tap, are screwed onto the coil slug-tuning screws to permit ease of adjustment without a tuning tool. Knobs with setscrews should be used to prevent slipping. A ceramic socket to mate with the pins of the coil form is used for J2.

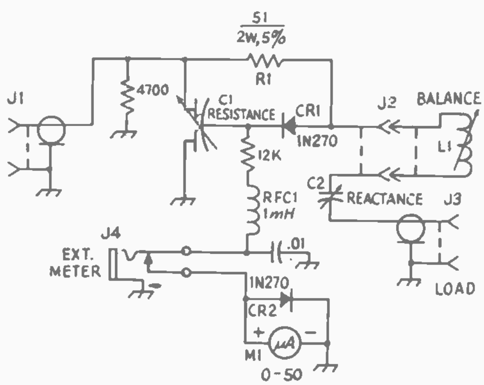


Fig. 2 — Schematic diagram of the impedance bridge. Capacitance is in microfarads; resistances are in ohms. Resistors are 1/2-W 10-percent tolerance unless otherwise indicated.

C1 — Differential capacitor, 11-161 pF per section (Millen 28801).
C2 — 17.5-327 pF with straight-line capacitance characteristic (Hammarlund RMC-325-S).
CR1, CR2 — Germanium diode, high back resistance.
J1, J3 — Coaxial connectors, chassis type.
J2 — To mate plug of L1, ceramic.
J4 — Phone jack, disconnecting type.
L1 — See text and Table I.
M1 — 0-50 μA dc (Simpson Model 1223 Bold-Vue, Cat. No. 15560 or equiv.).
R1 — For text reference.
RFC1 — Subminiature rf choke (Miller 70F103AI or equiv.).

Calibration

The resistance dial of the bridge may be calibrated by using a number of 1/2- or 1-watt 5-percent-tolerance composition resistors of different values in the 5- to 400-ohm range as loads. For this calibration, the appropriate frequency coil

TABLE 17-I

Coil Data for RF Impedance Bridge

Band	*Nominal Inductance Range, μH*	*Frequency Coverage, MHz*	*Coil Type or Data*
80	6.5–13.8	3.2–4.8	28 turns No. 30 enam. wire close-wound on Miller form 42A000CBI.
40	2.0–4.4	5.8–8.5	Miller 42A336CBI or 16 turns No. 22 enam. wire close-wound on Miller form 42A000CBI.
20	0.6–1.1	11.5–16.6	8 turns No. 18 enam. wire close-wound on Miller form 42A000CBI.
15	0.3–0.48	18.5–23.5	4 1/2 turns No. 18 enam. wire close-wound on Miller form 42A000CBI.
10	0.18–0.28	25.8–32.0	3 turns No. 16 or 18 enam. or tinned bus wire spaced over 1/4-inch winding length on Miller form 42A000CBI.

must be inserted at J2 and its inductance adjusted for the best null reading on the meter when C2 is set with its plates half meshed. For each test resistor, C1 is then adjusted for a null reading. Alternate adjustment of L1 and C1 should be made for a complete null. The leads between the test resistor and J3 should be as short as possible, and the calibration preferably should be done in the 3.5-MHz band where stray inductance and capacitance will have the least effect.

If the constructional layout of the bridge closely follows that shown in the photographs, the calibration scale of Fig. 4 may be used for the reactance dial. This calibration was obtained by connecting various reactances, measured on a laboratory bridge, in series with a 47-ohm 1-W resistor connected at J3. The scale is applied so that maximum capacitive reactance is indicated with C2 fully meshed. If it is desired to obtain an individual calibration for C2, known values of inductance and capacitance may be used in series with a fixed resistor of the same approximate value as R1. For this calibration it is *very important* to keep the leads to the test components as short as possible, and calibration should be performed in the 3.5-MHz range to minimize the effects of stray reactances. Begin the calibration by setting C2 at half mesh, marking this point as 0 ohms reactance. With a purely resistive load connected at J3, adjust L1 and C1 for the best null on M1. From this point on during calibration, do not adjust L1 except to rebalance the bridge for a new calibration frequency. The ohmic value of the known reactance for the frequency of calibration is multiplied by the frequency in MHz to obtain the calibration value for the dial.

Using the Impedance Bridge

This instrument is a low-input-power device, and is *not* of the type to be excited from a transmitter or left in the antenna line during station operation. Sufficient sensitivity for all measurements results when a 5-V rms rf signal is applied at J1. This amount of voltage can be delivered by most grid-dip oscillators. In no case should the power applied to J1 exceed 1 watt or calibration inaccuracy may result from a permanent change in the value of R1. The input impedance of the bridge at J1 is low, in the order of 50 to 100 ohms, so it is convenient to excite the bridge through a length of 52- or 75-ohm line such as RG-58/U or RG-59/U. If a grid-dip oscillator is used, a link coupling arrangement to the oscillator coil may be used. Fig. 1 shows two pick-up link assemblies. The larger coil, 10 turns of 1 1/4-inch-dia stock with turns spaced at 8 turns per inch, is used for the 80-, 40- and 20-meter bands. The smaller coil, 5 turns of 1-inch-dia stock with turns spaced at 4 turns per inch, is used for the 15- and 10-meter bands. Coupling to the oscillator should be as light as possible, while obtaining sufficient

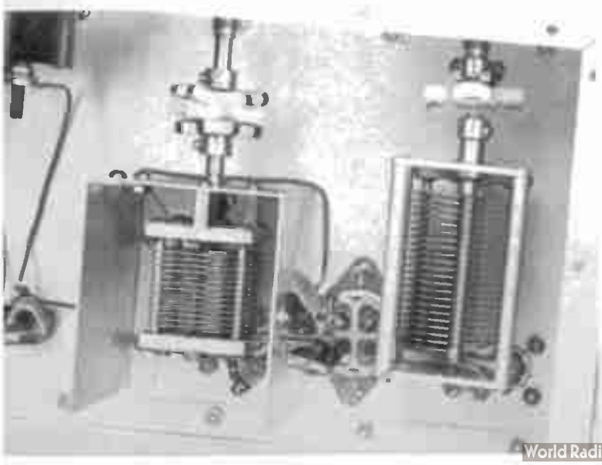
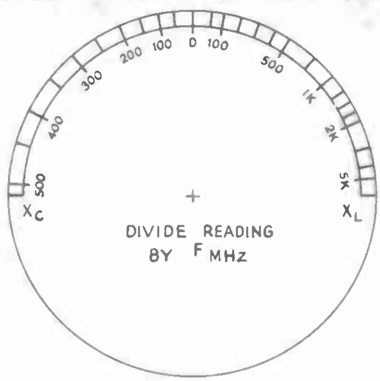


Fig. 3 — All components except the meter are mounted on the top of the box. C1 is visible inside the shield at the left, with C2 at the right and J2 mounted between them. J1 is hidden beneath C1 in this view; a part of J3 may be seen in the lower right corner of the box. Components for the dc metering circuit are mounted on a tie-point strip which is affixed to the shield wall for C1; all other components are interconnected with very short leads. The 4700-ohm input resistor is connected across J1.

Fig. 4 — Calibration scale for the reactance dial associated with C2. See text.



sensitivity, to prevent severe "pulling" of the oscillator frequency.

Before measurements are made, it is necessary to balance the bridge. Set the reactance dial at zero and adjust L1 and C1 for a null with a nonreactive load connected at J3. The bridge must be rebalanced after any appreciable change is made in the measurement frequency. A 51-ohm 1-W resistor mounted inside a PL-259 plug, as shown in Fig. 17-24, makes a load which is essentially nonreactive. After the bridge is balanced, connect the unknown load to J3, and alternately adjust C1 and C2 for the best null.

The calibration of the reactance dial is shown in Fig. 4. The measurement range for capacitive loads may be extended by "zeroing" the reactance dial at some value other than 0. For example, if the bridge is initially balanced with the reactance dial set at 500 in the X_L range, the 0 dial indication is now equivalent to an X_C reading of 500, and the total range of measurement for X_C has been extended to 1000.

A LOW-POWER RF WATTMETER

The wattmeter shown in Fig. 1 can be used with transmitters having power outputs from 1- to 25-watts within the frequency range of 1.8 to 30 MHz. For complete details, see *QST* for June, 1973. A bridge circuit based on a version of the one shown in Fig. 17-16C is used to measure the forward and reflected power on a transmission line.

The rf wattmeter.

It will be necessary to have a nonreactive 50-ohm dummy load for initial adjustment of the power meter. Connect the dummy load to one port of the instrument and apply rf power to the remaining port. S1 should now be thrown back and forth to determine which position gives the highest meter reading. This will be the FORWARD position. Adjust the sensitivity control for full-scale reading of the meter. Now, move the switch to the opposite (RELFECTED) position and adjust the trimmer nearest the transmitter input port for a null in the meter reading. The needle should drop to zero. It is recommended that these adjustments be made in the 10- or 15-meter band. Next, reverse the transmitter and load cables and repeat the nulling procedure while adjusting the trimmer on the opposite side of the pc board. Repeat these steps until a perfect null is obtained in both directions. The switch and the coax connectors can now be labeled, TRANSMITTER, LOAD, FORWARD, and REFLECTED, as appropriate.

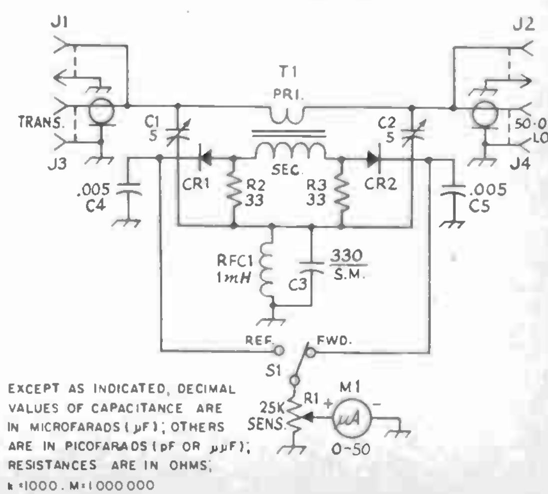
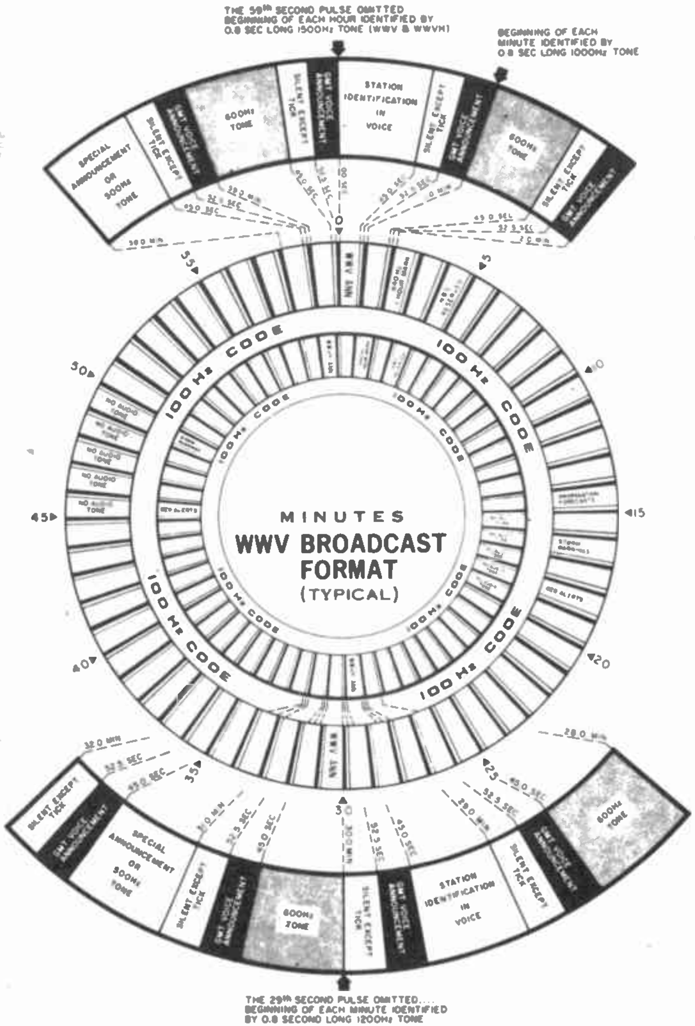


Fig. 1 — Schematic diagram of the wattmeter.
C1, C2 — 0.5- to 5-pF trimmer.
CR1, CR2 — 1N34A or equivalent.
M1 — 50-μA panel meter.
R1 — Linear-taper, 1/4 or 1/2 watt, 25,000 ohm.
R2, R3 — 33-ohm, 1/2-W composition resistor (matched pair recommended).
RFC1 — 1-mH rf choke.
S1 — Spdt toggle.
T1 — 60 turns 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.



STANDARD FREQUENCIES AND TIME SIGNALS

The National Bureau of Standards maintains two radio transmitting stations, WWV at Ft. Collins, Co., and WWVH near Kekaha, Kauai (Hawaii), for broadcasting standard radio frequencies of high accuracy. WWV broadcasts are on 2.5, 5, 10, 15, 20, and 25 MHz, and those from WWVH are on 2.5, 5, 10, 15, and 20 MHz. The broadcasts of both stations are continuous, night and day. Standard audio frequencies of 440, 500, and 600 Hz on each radio-carrier frequency by WWV and WWVH. The duration of each tone is approximately 45 seconds. A 600-Hz tone is broadcast during odd minutes by WWV, and during even minutes by WWVH. A 500-Hz tone is broadcast during alternate minutes unless voice announcements or silent periods are scheduled. A 440-Hz tone is broadcast beginning one minute after the hour by WWVH and two minutes after the hour by WWV. The 440-Hz tone period is omitted during the first hour of the UT day.

Transmitted frequencies from the two stations are accurate to ±2 parts in 10^{11}. Atomic frequency standards are used to maintain this accuracy.

Voice announcements of the time, in English, are given every minute. WWV utilizes a male voice, and WWVH features a female voice to distinguish between the two stations. WWV time and frequency broadcasts can be heard by telephone also. The number to call is (303) 499-7111, Boulder, CO.

All official announcements are made by voice. Time announcements are in GMT. One-second markers are transmitted throughout all programs except that the 29th and 59th markers of each minute are omitted. Detailed information on hourly broadcast schedules is given in the accompanying format chart. Complete information on the services can be found in NBS Special Publication 236, *NBS Frequency and Time Broadcast Services*, available for 25 cents from the Superintendent of Documents, U. S. Government Printing Office, Washington, D.C. 20402.

Geophysical Alerts

"Geoalerts" are broadcast in voice during the 19th minute of each hour from WWV and during the 46th minute of each hour from WWVH. The messages are changed each day at 0400 UT with provisions to schedule immediate alerts of outstanding occuring events. Geoalerts tell of geophysical events affecting radio propagation, stratospheric warming, etc.

Propagation Forecasts

Voice broadcasts of radio propagation conditions are given during part of every 15th minute of each hour from WWV. The announcements deal with short-term forecasts and refer to propagation along paths in the North Atlantic area, such as Washington, D.C. to London, or New York to Berlin.

CHU

CHU, the Canadian time-signal station, transmits on 3330.0, 7335.0 and 14,670.0 kHz. Voice announcements of the minute are made each minute; the 29th-second tick is omitted. Voice announcements are made in English and French.

A HETERODYNE DEVIATION METER

The instrument described here can be used to check the audio deviation of an fm transmitter, or to determine how far off frequency the transmitter carrier may be. It can also be used as a signal source to aid in setting a receiver on frequency, if a crystal of known accuracy is plugged into the oscillator.

The Circuit

As shown in Fig. 17-57 a transistor oscillator is used to feed energy to a mixer diode, CR1. A small pickup antenna is connected to the diode also, thereby coupling a signal from a transmitter to the mixer. The output from the diode, in the audio range, is amplified by U1, a 2747 operational amplifier. The 2747 amplifies and clips the audio, providing a square wave of nearly constant amplitude at the output. This square wave is applied to a rectifier circuit through variable coupling capacitors and a selector switch. A meter is connected to the rectifier circuit to read the average current. Since the amplitude of the input is constant, a change in frequency will produce a change of average current. Three ranges are selected by S1, with individual trimmers being placed in the circuit for calibration.

Fig. 17-56 – The deviation meter is constructed in a Calectro aluminum box. A four-position switch is at the lower right. The crystal plugs in on the left, with the frequency adjusting trimmer just below. A short whip or pickup wire can be plugged into the phono connector that is mounted on the back wall of the box.

Construction

An aluminum box is used for the enclosure, 6-1/4 × 3-1/2 × 2 inches. A meter switch, variable capacitor, and crystal socket are all mounted on the top panel. A small pc board is fastened to the

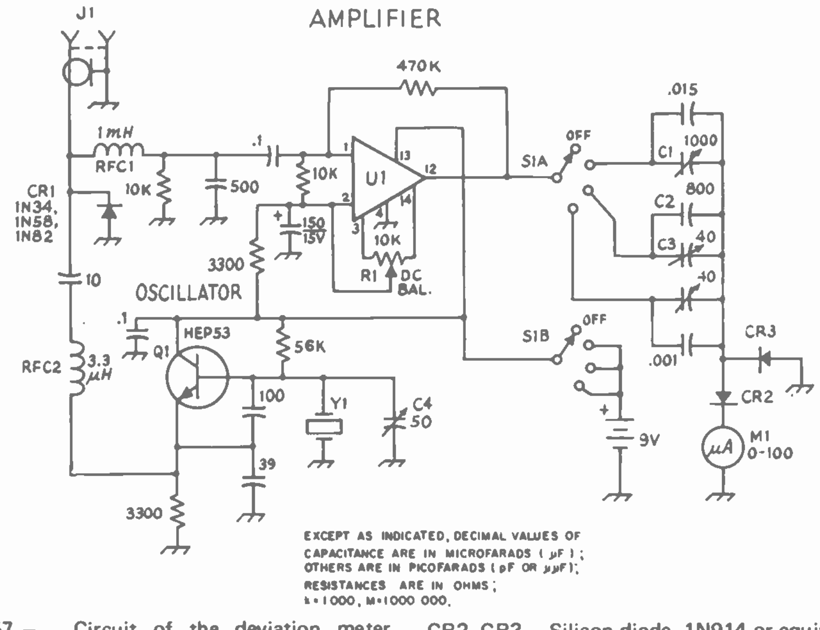


Fig. 17-57 – Circuit of the deviation meter. Connections shown are for a 2747 dual op amp. A 741 may be substituted with appropriate changes in pin numbers.

C1 – 360 to 1000 pF mica trimmer (J. W. Miller 160-A or equiv.).
C2, C3 – 3 to 30 pF mica trimmer (J. W. Miller 86 MA 2 or equiv.).
C4 – 50 pF miniature air variable (Hammarlund MAPC 50 or equiv.).
CR1 – Germanium diode, 1N34, 1N58, or 1N82 suitable.
CR2, CR3 – Silicon diode, 1N914 or equiv.
J1 – Coax connector, BNC or phono type suitable.
M1 – Microammeter, 0 to 1000 μA (Simpson Model 1212 Wide-Vue or equiv.).
Q1 – Motorola transistor.
R1 – 10,000-ohm miniature control, pc mount.
S1 – 2-pole, 4-position rotary switch, nonshorting.
U1 – Dual operational amplifier IC, Type 2747, one half not used.
Y1 – Crystal to produce harmonic on desired transmitter or receiver frequency. Fundamental range 6 to 20 MHz.

meter terminals as a convenient support. This board contains the IC and associated circuit components, as well as the rectifier diodes.

The oscillator is constructed on a separate pc board which mounts behind the crystal socket and variable capacitor. Metal spacers and 4-40 screws and nuts are used to fasten the oscillator board in place. A shield of pc board is placed between the oscillator and the amplifier to provide isolation. Power for the instrument is furnished by a 9-volt transistor radio battery that is held by a clip inside the box.

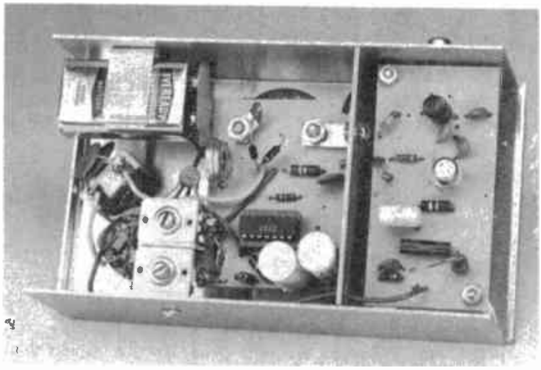


Fig. 17-58 – The dual op amp is located just below the center. Meter terminals are used as a convenient support for the amplifier pc board. The oscillator board is at the right, held in place by means of metal spacers.

Testing and Use.

Before calibrating the meter, the dc balance should be adjusted. A voltmeter should be connected to the output of U1, (pin 12) and R1 adjusted until the potential at this pin in one half of the supply voltage.

A low-level audio signal can be used to test the amplifier and meter circuit. As little as 10 mV, applied to pin 1, will produce a square wave at the output of the amplifier. Three ranges are provided in this meter; 0 - 1000Hz, 0 - 10 kHz, and 0 - 20 kHz. Each position can be calibrated by adjustment of the associated trimmer capacitor. The amount of capacitance needed may vary with different diodes, so fixed ceramic capacitors may be placed in parallel with the trimmers to bring the adjustment within range. As the frequency of the input to U1 is varied, the meter reading should correspond to that frequency over most of its range. On the upper frequency range, 0 - 20 kHz, a multiplication factor must be applied to the reading on the meter.

In use, a short whip or piece of wire is connected to J1, and the meter placed near a transmitter. A crystal that will produce a harmonic on the correct frequency is plugged into the socket. The selector switch should be in the first (0 - 1000 Hz) position. When the transmitter is turned on, the meter will indicate the difference in frequency between the transmitter and the harmonic from the oscillator. The trimmer, C4, should be adjusted for a minimum reading. Any hum, noise, or power-supply whine will cause a residual reading that could mask true zero beat. Modulation can be applied to the transmitter and the deviation control adjusted for the amount desired as indicated on the meter. Note that there is a difference between the indications obtained from a sine wave and those from voice. Readings will be lower with voice, the amount being dependent on the meter that is used and upon the individual voice.

Several transmitters can be netted to a system by setting the crystal in the device to the correct frequency at first, then adjust the frequency of each transmitter for an indication of zero beat.

Since there is some energy from the oscillator present at the input, J1, the same procedure can be used to align receivers to the correct frequency. When the deviation meter is acting as a signal source for checking either receivers or transmitters, the crystal should be checked for frequency drift several times during the test.

Construction Practices and Data Tables

TOOLS AND MATERIALS

While an easier, and perhaps a better, job can be done with a greater variety of tools available, by taking a little thought and care it is possible to turn out a fine piece of equipment with only a few of the common hand tools. A list of tools which will be indispensable in the construction of radio equipment will be found on this page. With these tools it should be possible to perform any of the required operations in preparing panels and metal chassis for assembly and wiring. It is an excellent idea for the amateur who does constructional work to add to his supply of tools from time to time as finances permit.

RECOMMENDED TOOLS

- Long-nose pliers, 6-inch and 4-inch
- Diagonal cutters, 6-inch and 4-inch
- Combination pliers, 6-inch
- Screwdriver, 6- to 7-inch, 1/4-inch blade
- Screwdriver, 4- to 5-inch, 1/8-inch blade
- Phillips screwdriver, 6- to 7-inch
- Phillips screwdriver, 3- to 4-inch
- Long-shank screwdriver with holding clip on blade
- Scratch awl or scriber for marking metal
- Combination square, 12-inch, for layout work
- Hand drill, 1/4-inch chuck or larger
- Soldering pencil, 30-watt, 1/8-inch tip
- Soldering iron, 200-watt, 5/8-inch tip
- Hacksaw and 12-inch blades
- Hand nibbling tool, for chassis-hole cutting
- Hammer, ball-peen, 1-lb head
- Heavy-duty jack knife
- File set, flat, round, half-round, and triangular. Large and miniature types recommended
- High-speed drill bits, No. 60 through 3/8-inch diameter
- Set of "Spintite" socket wrenches for hex nuts
- Crescent wrench, 6- and 10-inch
- Machine-screw taps, 4-40 through 10-32 thread
- Socket punches, 1/2", 5/8", 3/4", 1 1/8", 1 1/4", and 1 1/2"
- Tapered reamer, T-handle, 1/2-inch maximum pitch
- Bench vise, 4-inch jaws or larger
- Medium-weight machine oil
- Tin shears, 10-inch size
- Motor-driven emery wheel for grinding
- Solder, *rosin core only*
- Contact cleaner, liquid or spray can
- Duco cement or equivalent
- Electrical tape, vinyl plastic

Radio-supply houses, mail-order retail stores and most hardware stores carry the various tools required when building or servicing amateur radio equipment. While power tools (electric drill or drill press, grinding wheel, etc.) are very useful and will save a lot of time, they are not essential.

Twist Drills

Twist drills are made of either high-speed steel or carbon steel. The latter type is more common and will usually be supplied unless specific request is made for high-speed drills. The carbon drill will suffice for most ordinary equipment construction work and costs less than the high-speed type.

While twist drills are available in a number of sizes, those listed in bold-faced type in Table 18-1 will be most commonly used in construction of amateur equipment. It is usually desirable to purchase several of each of the commonly used sizes rather than a standard set, most of which will be used infrequently if at all.

Care of Tools

The proper care of tools is not alone a matter of pride to a good workman. He also realizes the energy which may be saved and the annoyance which may be avoided by the possession of a full kit of well-kept sharp-edged tools.

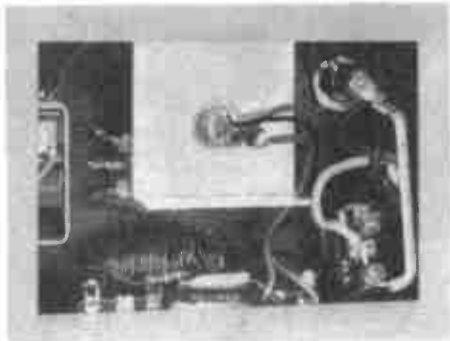
Drills should be sharpened at frequent intervals so that grinding is kept at a minimum each time. This makes it easier to maintain the rather critical surface angles required for best cutting with least wear. Occasional oilstoning of the cutting edges of a drill or reamer will extend the time between grindings.

The soldering iron can be kept in good condition by keeping the tip well tinned with solder and not allowing it to run at full voltage for long periods when it is not being used. After each period of use, the tip should be removed and cleaned of any scale which may have accumulated. An oxidized tip may be cleaned by dipping it in sal ammoniac while hot and then wiping it clean with a rag. If the tip becomes pitted it should be filed until smooth and bright, and then tinned immediately by dipping it in solder.

Useful Materials

Small stocks of various miscellaneous materials will be required in constructing radio apparatus, most of which are available from hardware or radio-supply stores. A representative list follows:

Fig. 1 — The SCR motor-speed control is housed in a small cabinet.



The working parts of the motor-speed control. The triac is centered on its aluminum heat sink, with the terminals of the speed-control resistor protruding from underneath. The rf-hash-suppression filter and components in the gate-triggering circuit are mounted on a tie-point strip, being visible at the bottom of the enclosure as shown in this view. The triac is barely discernable at the right end of the fixed resistor. Terminals of the strip which are associated with the mounting feet are unused, and are bent down to prevent accidental shorts to other parts of the circuit.

Sheet aluminum, solid and perforated, 16 or 18 gauge, for brackets and shielding.
1/2 X 1/2-inch aluminum angle stock.
1/ 4-inch diameter round brass or aluminum rod for shaft extensions. Machine screws: Round-head and flat-head, with nuts to fit. Most useful sizes: 4-40, 6-32 and 8-32, in lengths from 1/4 inch to 1 1/2 inches. (Nickel-plated iron will be found satisfactory except in strong rf fields, where brass should be used.)
Bakelite, lucite and polystyrene scraps.
Soldering lugs, panel bearings, rubber grommets, terminal-lug wiring strips, varnished-cambric insulating tubing.
Shielded and unshielded wire.
Tinned bare wire, Nos. 22, 14 and 12.

Machine screws, nuts, washers, soldering lugs, etc., are most reasonably purchased in quantities of a gross. Many of the radio-supply stores sell small quantities and assortments that come in handy.

TRIAC MOTOR-SPEED CONTROL

Most electric hand drills operate at a single high speed; however, from time to time, the need arises to utilize low or medium speeds. Low speeds are useful when drilling in tight spaces or on exposed surfaces where it is important that the drill bit doesn't slip, and when drilling bakelite, Plexiglas and similar materials. Medium speeds are useful for drilling non-ferrous metals such as aluminum and brass. One way to accomplish these ends with a single-speed electric drill is to use a silicon bidirectional thyristor (Triac) speed control.

The circuit for the Triac speed control is shown in Fig. 1. This type of circuit provides some degree of regulation with varying loads.

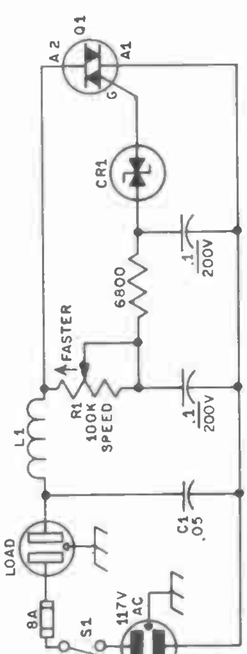


Fig. 1 — Schematic diagram of motor-speed control. Resistances are in ohms (k = 1000) and capacitances are in microfarads. **Important note**: The basing diagram for Q1 is correct as shown here. Some early literature accompanying the packaging of the HEP device appears to be in error.
C1 — .05-µF, 600-V 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.

Construction

Because of the small complement of parts, the Triac speed control can be constructed inside a very small container. The model described was built in a 2-3/4 × 2-1/8 × 1-5/8-inch Minibox. Since the mounting stud and main body of the Triac are common with the anode, care should be used to mount the Triac clear from surrounding objects. In the unit shown, two soldering lugs were soldered together and the narrow ends connected to one side of the female output connector; the large ends were used as a fastening point for the Triac anode stud.

Operation

Although the circuit described is intended to be used to reduce the speed of electric hand drills that draw six amperes or less, it has many other applications. It can be used to regulate the temperature of a soldering iron, which is being used to wire a delicate circuit, or it may be used for dimming lamps or for controlling the cooking speed of a small hot plate. Note, however, if the circuit is used with a device drawing from three to six amperes for a continuous period of over ten minutes, it will be necessary to provide a heat sink (insulated from the chassis) for the Triac anode case.

CHASSIS WORKING

With a few essential tools and proper procedure, it will be found that building radio gear on a metal chassis is a relatively simple matter. Aluminum is to be preferred to steel, not only because it is a superior shielding material, but because it is much easier to work and provides good chassis contacts.

The placing of components on the chassis is shown quite clearly in the photographs in this *Handbook*. Aside from certain essential dimensions, which usually are given in the text, exact duplication is not necessary.

Much trouble and energy can be saved by spending sufficient time in planning the job. When all details are worked out beforehand the actual construction is greatly simplified.

Cover the top of the chassis with a piece of wrapping paper, or, preferably, cross-section paper, folding the edges down over the sides of the chassis and fastening with adhesive tape. Then assemble the parts to be mounted on top of the chassis and move them about until a satisfactory arrangement has been found, keeping in mind any parts which are to be mounted underneath, so that interferences in mounting may be avoided. Place capacitors and other parts with shafts extending through the panel first, and arrange them so that

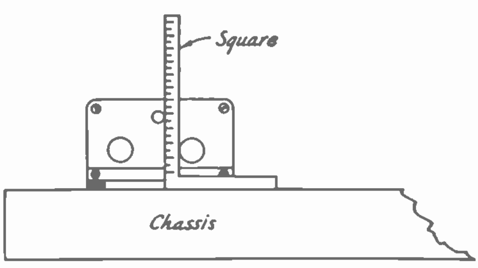


Fig. 18-3 — Method of measuring the heights of capacitor shafts. If the square is adjustable, the end of the scale should be set flush with the face of the head.

TABLE 18-I

Numbered Drill Sizes

Num.	*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.

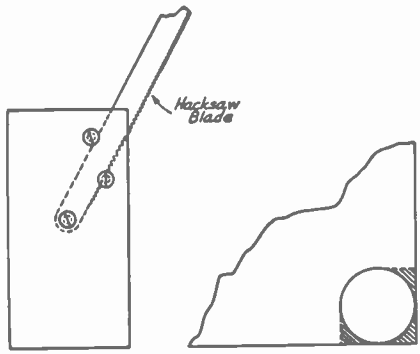


Fig. 18-4 – To cut rectangular holes in a chassis corner, holes may be filed out as shown in the shaded portion of B, making it possible to start the hack-saw blade along the cutting line. A shows how a single-ended handle may be constructed for a hack-saw blade.

the controls will form the desired pattern on the panel. Be sure to line up the shafts squarely with the chassis front. Locate any partition shields and panel brackets next, and then the tube sockets and any other parts, marking the mounting-hole centers of each accurately on the paper. Watch out for capacitors whose shafts are off center and do not line up with the mounting holes. Do not forget to mark the centers of socket holes and holes for leads under i-f transformers, etc., as well as holes for wiring leads. The small holes for socket-mounting screws are best located and center-punched, using the socket itself as a template, after the main center hole has been cut.

By means of the square, lines indicating accurately the centers of shafts should be extended to the front of the chassis and marked on the panel at the chassis line, the panel being fastened on temporarily. The hole centers may then be punched in the chassis with the center punch. After drilling, the parts which require mounting underneath may be located and the mounting holes drilled, marking sure by trial that no interferences exist with parts mounted on top. Mounting holes along the front edge of the chassis should be transferred to the panel, by once again fastening the panel to the chassis and marking it from the rear.

Next, mount on the chassis the capacitors and any other parts with shafts extending to the panel, and measure accurately the height of the center of each shaft above the chassis, as illustrated in Fig. 18-3. The horizontal displacement of shafts having already been marked on the chassis line on the panel, the vertical displacement can be measured from this line. The shaft centers may now be marked on the back of the panel, and the holes drilled. Holes for any other panel equipment coming above the chassis line may then be marked and drilled, and the remainder of the apparatus mounted. Holes for terminals etc., in the rear edge of the chassis should be marked and drilled at the same time that they are done for the top.

Drilling and Cutting Holes

When drilling holes in metal with a hand drill it is important that the centers first be located with a center punch, so that the drill point will not "walk" away from the center when starting the hole. When the drill starts to break through, special care must be used. Often it is an advantage to shift a two-speed drill to low gear at this point. Holes more than 1/4-inch in diameter should be started with a smaller drill and reamed out with the larger drill.

The check on the usual type of hand drill is limited to 1/4-inch drills. Although it is rather tedious, the 1/4-inch hole may be filed out to larger diameters with round files. Another method possible with limited tools is to drill a series of small holes with the hand drill along the inside of the circumference of the large hole, placing the holes as close together as possible. The center may then be knocked out with a cold chisel and the edges smoothed up with a file. Taper reamers which fit into the carpenter's brace will make the job easier. A large rat-tail file clamped in the brace makes a very good reamer for holes up to the diameter of the file.

For socket holes and other large holes in an aluminum chassis, socket-hole punches should be used. They require first drilling a guide hole to pass the bolt that is turned to squeeze the punch through the chassis. The threads of the bolt should be oiled occasionally.

Large holes in steel panels or chassis are best cut with an adjustable circle cutter. Occasional application of machine oil in the cutting groove will help. The cutter first should be tried out on a block of wood, to make sure that it is set for the right diameter.

The burrs or rough edges which usually result after drilling or cutting holes may be removed with a file, or sometimes more conveniently with a sharp knife or chisel. It is a good idea to keep an old wood chisel sharpened and available for this purpose.

Rectangular Holes

Square or rectangular holes may be cut out by making a row of small holes as previously described, but is more easily done by drilling a 1/2-inch hole inside each corner, as illustrated in Fig. 18-4, and using these holes for starting and turning the hack saw. The socket-hole punch and the square punches which are now available also may be of considerable assistance in cutting out large rectangular openings.

SEMICONDUCTOR HEAT SINKS

Homemade heat sinks can be fashioned from brass, copper or aluminum stock by employing ordinary workshop tools. The dimensions of the heat sink will depend upon the type of transistor used, and the amount of heat that must be conducted away from the body of the semiconductor.

Fig. 18-5 shows the order of progression for forming a large heat sink from aluminum or brass

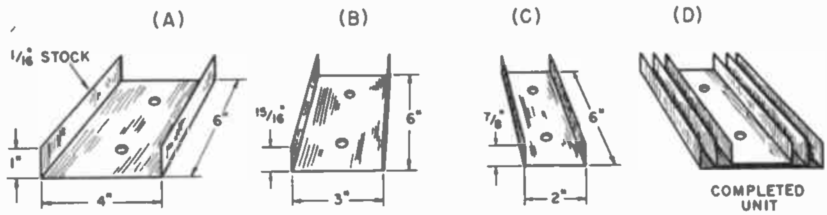


Fig. 18-5 — Details for forming channel type heat sinks.

channels of near-equal height and depth. The width is lessened in parts (B) and (C) so that each channel will fit into the preceding one as shown in the completed model at (D). The three pieces are bolted together with 8-32 screws and nuts. Dimensions given are for illustrative purposes only.

Heat sinks for smaller transistors can be fabricated as shown in Fig. 18-7. Select a drill bit that is one size smaller than the diameter of the transistor case and form the heat sink from 1/16 inch thick brass, copper or aluminum stock as shown in steps (A), (B), and (C). Form the stock around the drill bit by compressing it in a vise (A). The completed heat sink is press-fitted over the body of the semiconductor as illustrated at (D). The larger the area of the heat sink, the greater will be the amount of heat conducted away from the transistor body. In some applications, the heat sinks shown in Fig. 18-7 may be two or three inches in height (power transistor stages).

Another technique for making heat sinks for TO-5 type transistors (1) and larger models (1) is shown in Fig. 18-6. This style of heat sink will dissipate considerably more heat than will the type shown in Fig. 18-5. The main body of the sink is fashioned from a piece of 1/8-inch thick aluminum angle bracket – available from most hardware stores. A hole is bored in the angle stock to allow the transistor case to fit *snugly* into it. The transistor is held in place by a small metal plate whose center hole is slightly smaller in diameter than the case of the transistor. Details are given in Fig. 18-6.

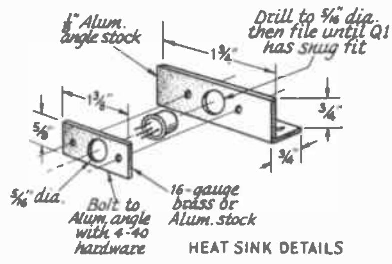
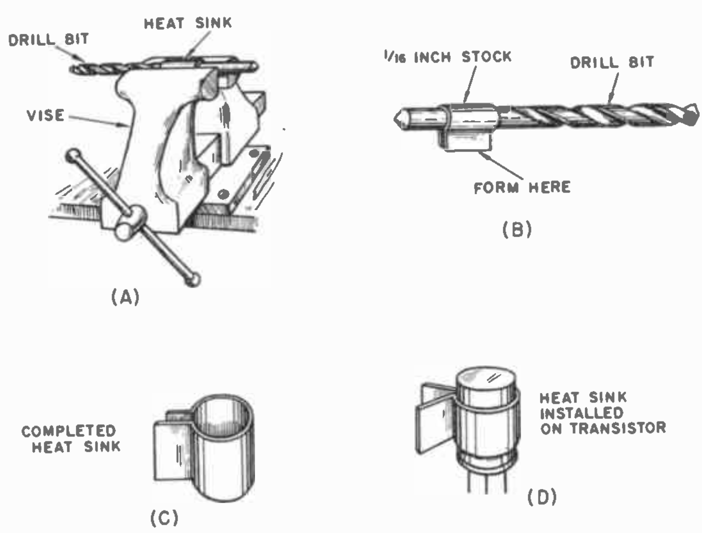


Fig. 18-6 — Layout and assembly details of another homemade heat sink. The completed assembly can be insulated from the main chassis of the transmitter by using insulating washers.

A thin coating of silicone grease, available from most electronics supply houses, can be applied between the case of the transistor and the part of the heat sink with which it comes in contact. The silicone grease will aid the transfer of heat from the transistor to the sink. This practice can be applied to all models shown here. In the example given in Fig. 18-5, the grease should be applied between the

Fig. 18-7 — Steps used in constructing heat sinks for small transistors.



three channels before they are bolted together, as well as between the transistor and the channel it contacts.

CONSTRUCTION NOTES

If a control shaft must be extended or insulated, a flexible shaft coupling with adequate insulation should be used. Satisfactory support for the shaft extension, as well as electrical contact for safety, can be provided by means of a metal panel bearing made for the purpose. These can be obtained singly for use with existing shafts, or they can be bought with a captive extension shaft included. In either case the panel bearing gives a "solid" feel to the control.

The use of fiber washers between ceramic insulation and metal brackets, screws or nuts will prevent the ceramic parts from breaking.

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.

Cutting and Bending Sheet Metal

If a sheet of metal is too large to be cut conveniently with a hack saw, it may be marked with scratches as deep as possible along the line of the cut on both sides of the sheet and then clamped in a vise and worked back and forth until the sheet breaks at the line. Do not carry the bending too far until the break begins to weaken; otherwise the edge of the sheet may become bent. A pair of iron bars or pieces of heavy angle stock, as long or longer than the width of the sheet, to hold it in the vise, will make the job easier. "C" clamps may be used to keep the bars from spreading at the ends. The rough edges may be smoothed with a file or by placing a large piece of emery cloth or sandpaper on a flat surface and running the edge of the metal back and forth over the sheet. Bends may be made similarly.

Finishing Aluminum

Aluminum chassis, panels and parts may be given a sheen finish by treating them in a caustic bath. An enamelled or plastic container, such as a dishpan or infant's bathtub, should be used for the solution. Dissolve ordinary household lye in cold water in a proportion of 1/4 to 1/2 can of lye per gallon of water. The stronger solution will do the job more rapidly. Stir the solution with a stick of wood until the lye crystals are completely dissolved. Be very careful to avoid any skin contact with the solution. It is also harmful to clothing. Sufficient solution should be prepared to cover the piece completely. When the aluminum is immersed, a very pronounced bubbling takes place and ventilation should be provided to disperse the escaping gas. A half hour to two hours in the solution should be sufficient, depending upon the strength of the solution and the desired surface.

Remove the aluminum from the solution with sticks and rinse thoroughly in cold water while swabbing with a rag to remove the black deposit. When dry, finish by spraying on a light coat of clear lacquer.

Soldering

The secret of good soldering is to use the right amount of heat. Too little heat will produce a "cold-soldered joint"; too much may injure a component. The iron and the solder should be applied simultaneously to the joint. Keep the iron clean by brushing the hot tip with a paper towel. Always use rosin-core solder, never acid-core. Solders have different melting points, depending upon the ratio of tin to lead. A 50-50 solder melts at 425 degrees F, while 60-40 melts at 371 degrees F. When it is desirable to protect from excessive heat the components being soldered, the 60-40 solder is preferable to the 50-50. (A less-common solder, 63-37, melts at 361 degrees F.)

When soldering transistors, crystal diodes or small resistors, the lead should be gripped with a pair of pliers up close to the unit so that the heat will be conducted away. Overheating of a transistor or diode while soldering can cause permanent damage. Also, mechanical stress will have a similar effect, so that a small unit should be mounted so that there is no appreciable mechanical strain on the leads.

Trouble is sometimes experienced in soldering to the pins of coil forms or male cable plugs. It helps if the pins are first cleaned on the inside with a suitable twist drill and then tinned by flowing rosin-core solder into them. Immediately clear the surplus solder from each hot pin by a whipping motion or by blowing through the pin from the inside of the form or plug. Before inserting the wire in the pin, file the nickel plate from the tip. After soldering, round the solder tip off with a file.

When soldering to the pins of polystyrene coil forms, hold the pin to be soldered with a pair of heavy pliers, to form a "heat sink" and insure that the pin does not heat enough in the coil form to loosen and become misaligned.

Wiring

The wire used in connecting amateur equipment should be selected considering both the maximum current it will be called upon to handle and the voltage its insulation must stand without breakdown. Also, from the consideration to TVI, the power wiring of all transmitters should be done with wire that has a braided shielding cover. Receiver and audio circuits may also require the use of shielded wire at some points for stability, or the elimination of hum.

No. 20 stranded wire is commonly used for most receiver wiring (except for the high-frequency circuits) where the current does not exceed 2 or 3 amperes. For higher-current heater circuits, No. 18 is available. Wire with cellulose acetate insulation is good for voltages up to about 500. For higher voltages, themoplastic-insulated wire should be used. Inexpensive wire strippers that make the removal of insulation from hookup wire an easy job are available on the market.

When power leads have several branches in the chassis, it is convenient to use fiber-insulated multiple tie points as anchorages or junction points. Strips of this type are also useful as insulated supports for resistors, rf chokes and capacitors. High-voltage wiring should have exposed points held to a minimum; those which cannot be avoided should be made as inaccessible as possible to accidental contact or short-circuit.

Where shielded wire is called for and capacitance to ground is not a factor, Belden type 8885 shielded grid wire may be used. If capacitance must be minimized, it may be necessary to use a piece of car-radio low-capacitance lead-in wire, or coaxial cable.

For wiring high-frequency circuits, rigid wire is often used. Bare soft-drawn tinned wire, size 22 to 12 (depending on mechanical requirements) is suitable. Kinks can be removed by stretching a piece of 10 or 15 feet long and then cutting into short lengths that can be handled conveniently. Rf wiring should be run directly from point to point with a minimum of sharp bends and the wire kept well spaced from the chassis or other grounded metal surfaces. Where the wiring must pass through the chassis or a partition, a clearance hole should be cut and lined with a rubber grommet. In case insulation becomes necessary, varnished cambric tubing (spaghetti) can be slipped over the wire.

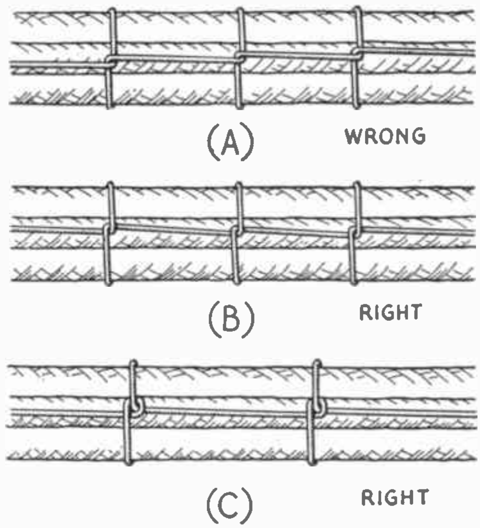


Fig. 18-8 – Methods of lacing cables. The method shown at C is more secure, but takes more time than the method of B. The latter is usually adequate for most amateur requirements.

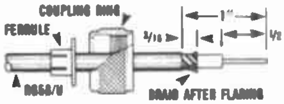
In transmitters where the peak voltage does not exceed 2500 volts, the shielded grid wire mentioned above should be satisfactory for power circuits. For higher voltages, Belden type 8656, Birnbach type 1820, or shielded ignition cable can be used. In the case of filament circuits carrying heavy current, it may be necessary to use No. 10 or 12 bare or enameled wire, slipped through spaghetti, and then covered with copper braid pulled tightly over the spaghetti. The chapter on TVI shows the manner in which shielded wire should be applied. If the shielding is simply slid back over the insulation and solder flowed into the end of the braid, the braid usually will stay in place without the necessity for cutting it back or binding it in place. The braid should be cleaned first so that solder will take with a minimum of heat.

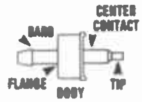
Rf wiring in transmitters usually follows the method described above for receivers with due respect to the voltages involved.

Where power or control leads run together for more than a few inches, they will present a better appearance when bound together in a single cable. The correct technique is illustrated in Fig. 18-8; both plastic and waxed-linen lacing cords are available. Plastic cable clamps are available to hold the laced cable.

To give a "commercial look" to the wiring of any unit, run any cabled leads along the edge of the chassis. If this isn't possible, the cabled leads should then run parallel to an edge of the chassis. Further, the generous use of tie points (mounted parallel to an edge of the chassis), for the support of one or both ends of a resistor or fixed capacitor,

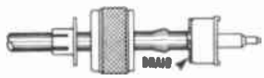
83-58FCP



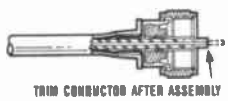


1. Strip cable – *don't nick braid, dielectric or conductor.* Slide ferrule, then coupling ring on cable. Flare braid slightly by rotating conductor and dielectric in circular motion.

2. Slide body on dielectric, barb going under braid until flange is against outer jacket. Braid will fan out against body flange.

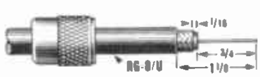


3. Slide nut over body. Grasp cable with hand and push ferrule over barb until braid is captured between ferrule and body flange. Squeeze crimp tip only of center contact with pliers; alternate-solder tip.



83-1SP PLUG (PL-259)

1. Strip cable, *don't nick braid, dielectric or conductor.* Tin exposed braid and conductor. Slide coupling ring on cable.



2. Screw body on cable. Solder braid through solder holes. Solder conductor to center contact

3. Screw coupling ring on body.

83-1SP PLUG WITH ADAPTERS

1. Strip jacket. *Don't nick braid.* Slide coupling ring and adapter on cable. Note – use 83-168 adapter for RG-58/U and 83-185 for RG-59/U.

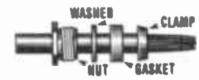
2. Fan braid slightly, fold back over adapter and trim to 3/8". Strip dielectric and tin exposed conductor. *Don't nick conductor.*

3. Screw body on adapter. Follow 2 and 3 under 83-1SP plug.

BNC CONNECTORS (STANDARD CLAMP)

1. Strip jacket. Fray braid and strip dielectric. *Don't nick braid or conductor.* Tin conductor.

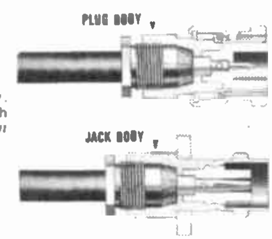
2. Taper braid. Slide nut, washer, gasket and clamp over braid. Clamp inner shoulder should fit squarely against end of jacket.



3. With clamp in place, comb out braid, fold back smooth as shown. Trim 3/32" from end.

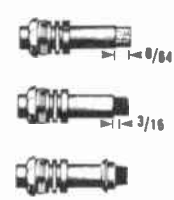
4. Solder contact on conductor through solder hole. Contact should butt against dielectric. Remove excess solder from outside of contact. Avoid excess heat to prevent swollen dielectric which would interfere with connector body.

5. Push assembly into body. Screw nut into body with wrench until tight. *Don't rotate body on cable to tighten.*



BNC CONNECTORS (IMPROVED CLAMP)

1. Follow 1, 2, 3 and 4 above except as noted. Strip cable as shown. Slide gasket on cable *with groove facing clamp.* Slide clamp on cable *with sharp edge facing gasket.* Clamp *should* cut gasket to seal properly.



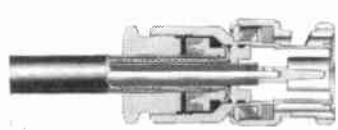


Fig. 18-9 – Cable stripping dimensions and assembly instructions for several popular coaxial cable connectors. This material courtesy of AMPHENOL® ELECTRONIC COMPONENTS, RF Division, Bunker Ramo Corp.

will add to the appearance of the finished unit. In a similar manner, "dress" the small components so that they are parallel to the panel or sides of the chassis.

Winding Coils

Close-wound coils are readily wound on the specified form by anchoring one end of a length of wire (in a vise or to a doorknob) and the other end to the coil form. Straighten any kinks in the wire and then pull to keep the wire under slight tension. Wind the coil to the required number of turns while walking toward the anchor, always maintaining a slight tension on the wire.

To space-wind the coil, wind the coil simultaneously with a suitable spacing medium (heavy thread, string or wire) in the manner described above. When the winding is complete, secure the end of the coil to the coil-form terminal and then carefully unwind the spacing material. If the coil is wound under suitable tension, the spacing material can be easily removed without disturbing the winding. Finish the space-wound coil by judicious applications of Duco cement, to hold the turns in place.

The "cold" end of a coil is the end at or close to chassis or ground potential. Coupling links should be wound on the cold end of a coil, to minimize capacitive coupling.

CIRCUIT-BOARD FABRICATION

Many modern-day builders prefer the neatness and miniaturization made possible by the use of etched or printed circuit boards. There are additional benefits to be realized from the use of circuit boards: Low lead inductances, excellent physical stability of the components and interconnecting leads, and good repeatability of the basic layout of a given project. The latter attribute makes the use of circuit boards ideal for group projects.

Methods

Perhaps the least complicated approach to circuit-board fabrication is the use of unclad perforated board into which a number of push-in terminals have been installed. The perforated board can be obtained with one of many hole patterns, dependent upon the needs of the builder. Perforated terminal boards are manufactured by such firms as Vector, Kepro, and Triad. Their products are available from the large mail-order houses.

Once the builder plots the layout of his circuit on paper, push-in terminals can be installed in the "perf" board to match the layout which was done on paper. The terminals serve as tie points and provide secure mounting-post anchors for the various components. Selected terminals can be wired together to provide ground and B-plus lines. Although this technique is the most basic of the methods, it is entirely practical.

An approach to etched-circuit board assembly can be realized by cutting strips of flashing copper, hobby copper, or brass shim stock into the desired shapes and lengths, then gluing them to a piece of unclad circuit board. Epoxy cement is useful for the latter. Alternatively, the strips can be held in place by means of brass eyelets which have been installed with a hand eyelet tool. If standard unclad circuit board is not handy, linoleum or Formica sheeting can be made to serve as a base for the circuit board. If this technique is used, the metal strips should be soldered together at each point where they join, assuring good electrical contact.

Etched-circuit boards provide the most professional end result of the three systems described here. They are the most stable, physically and electrically, and can be easily repeated from a single template. Etched-circuits can be formed on copper-clad perforated board, or on unpunched copper-clad board. There is no advantage in using the perforated board as a base unless push-in terminals are to be used.

Planning and Layout

The constructor should first plan the physical layout of the circuit by sketching a pictorial diagram on paper, drawing it to scale. Once this has been done, the interconnecting leads can be inked in to represent the copper strips that will remain on the etched board. The Vector Company sells layout paper for this purpose. It is marked with the same patterns that are used on their perforated boards.

After the basic etched-circuit design has been completed the designer should go over the proposed layout several times to insure against errors. When the foregoing has been done, the pattern can be painted on the copper surface of the board to be etched. Etch-resistant solutions are available from commercial suppliers and can be selected from their catalogs. Some builders prefer to use India ink for this purpose. Perhaps the most readily-available material for use in etch-resist applications is ordinary exterior enamel paint. The portions of the board to be retained are covered with a layer of paint, applied with an artist's brush, duplicating the pattern that was drawn on the layout paper. The job can be made a bit easier by tracing over the original layout with a ballpoint pen and carbon paper while the pattern is taped to the copper side of the unetched circuit board. The carbon paper is placed between the pattern and the circuit board. After the paint has been applied, it should be allowed to dry for at least 24 hours prior to the etching process. The Vector Company produces a rub-on transfer material that can also be used as etch-resist when laying out circuit-board patterns. Thin strips of ordinary masking tape, cut to size and firmly applied, serve nicely as etch-resist material too.

The Etching Process

Almost any strong acid bath will serve as an etchant, but the two chemical preparations recommended here are the safest to use. A bath can be prepared by mixing 1 part ammonium persulphate crystals with 2 parts clear water. A

Fig. 18-10 — A homemade stand for processing etched-circuit boards. The heat lamp maintains the etchant-bath temperature between 90 and 115 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.

normal quantity of working solution for most amateur radio applications is composed of 1 cup of crystals and 2 cups of water. To this mixture add 1/4 teaspoon of mercuric chloride crystals. The latter serves as an activator for the bath. Ready-made etchant kits which use these chemicals are available from Vector. A two-bag kit is sold as item 2594 and costs just over $1. Complete kits which contain circuit boards, etchant powders, etch-resist transfers, layout paper, and plastic etchant bags are also available from Vector at moderate prices.

Another chemical bath that works satisfactorily for copper etching is made up from one part ferric chloride crystals and 2 parts water. No activator is required with this bath. Ready-made solutions (one-pint and one-gallon sizes) are available through some mail-order houses at low cost. They are manufactured by Kepro Co. and carry a stock number of E-1PT and E-1G, respectively. One pint costs less than a dollar.

Etchant solutions become exhausted after a certain amount of copper has been processed, therefore it is wise to keep a quantity of the bath on hand if frequent use is anticipated. With either chemical bath, the working solution should be maintained at a temperature between 90 and 115 degrees F. A heat lamp can be directed toward the bath during the etching period, its distance set to maintain the required temperature. A darkroom thermometer is handy for monitoring the temperature of the bath.

While the circuit board is immersed in the solution, it should be agitated continuously to permit uniform reaction to the chemicals. This action will also speed up the etching process somewhat. Normally, the circuit board should be placed in the bath with the copper side facing down, toward the bottom of the tray. The tray should be non-metallic, preferably a Pyrex dish or a photographic darkroom tray.

The photograph, Fig. 18-10, shows a homemade etching stand made up from a heat lamp, some lumber, and an 8 rpm motor. An eccentric disk has been mounted on the motor shaft and butts against the bottom of the etchant tray. As the motor turns, the eccentric disk raises and lowers one end of the try, thus providing continuous agitation of the solution. The heat lamp is mounted on an adjustable, slotted wooden arm. Its height above the solution tray is adjusted to provide the desired bath temperature. Because the etching process takes between 15 minutes and one hour —dependent upon the strength and temperature of the bath — such an accessory is convenient.

After the etching process is completed, the board is removed from the tray and washed thoroughly with fresh, clear water. The etch-resist material can then be rubbed off by applying a few brisk strokes with medium-grade steel wool. WARNING: *Always use rubber gloves when working with etchant powders and solutions. Should the acid bath come in contact with the body, immediately wash the affected area with clear water. Protect the eyes when using acid baths.*

COMPONENT VALUES

Values of composition resistors and small capacitors (mica and ceramic) are specified throughout this *Handbook* in terms of "preferred values." In the preferred-number system, all values represent (approximately) a constant-percentage increase over the next lower value. The base of the system is the number 10. Only two significant figures are used.

"Tolerance" means that a variation of plus or minus the percentage given is considered satisfactory. For example, the actual resistance of a "4700-ohm" 20-percent resistor can lie anywhere between 3700 and 5600 ohms, approximately. The permissible variation in the same resistance value with 5-percent tolerance would be in the range from 4500 to 4900 ohms, approximately.

In the component specifications in this *Handbook*, it is to be understood that when no tolerance is specified the *largest* tolerance available in that value will be satisfactory.

Values that do not fit into the preferred-number system (such as 500, 25,000) easily can be substituted. It is obvious, for example, that a 5000-ohm resistor falls well within the tolerance range of the 4700-ohm 20-percent resistor used in the example above. It would not, however, be usable if the tolerance were specified as 5 percent.

TABLE 18-II

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

COLOR CODES

Standardized color codes are used to mark values on small components such as composition resistors and mica capacitors, and to identify leads from transformers, etc. The resistor-capacitor number color code is given in Table 18-III.

Fixed Capacitors

The methods of marking "postage-stamp" mica capacitors, molded paper capacitors and tubular ceramic capacitors are shown in Fig. 18-11.

Capacitors made to American War Standards or Joint Army-Navy specifications are marked with the 6-dot code shown at the top. Practically all surplus capacitors are in this category.

The 3-dot EIA code is used for capacitors having a rating of 500 volts and ±20 percent tolerance only; other ratings and tolerances are covered by the 6-dot EIA code.

Example: A capacitor with a 6-dot code has the following markings: Top row, left to right, black, yellow, violet; bottom row, right to left, brown, silver, red. Since the first color in the top row is black (significant figure zero) this is the AWS code and the capacitor has mica dielectric. The significant figures are 4 and 7, the decimal multiplier 10 (brown, at right of second row), so the capacitance is 470 pF. The tolerance is ±0%. The final color, the characteristic, deals with temperature coefficients and methods of testing (see Table 18-V).

A capacitor with a 3-dot code has the following colors, left to right: brown, black, red. The significant figures are 1, 0 (10) and the multiplier is 100. The capacitance is therefore 100 pF.

A capacitor with a 6-dot code has the following markings: Top row, left to right, brown, black, black; bottom row, right to left, black, gold, blue. Since the first color in the top row is neither black nor silver, this is the EIA code. The significant figures are 1, 0, 0 (100) and the decimal multiplier is 1 (black). The capacitance is therefore 100 pF. The gold dot shows that the tolerance is ±5% and the blue dot indicates 600-volt rating.

Ceramic Capacitors

Conventional markings for ceramic capacitors are shown in the lower drawing of Fig. 18-11. The colors have the meanings indicated in Table 18-III. In practice, dots may be used instead of the *narrow* bands indicated in Fig. 18-11.

Example: A ceramic capacitor has the following markings: Broad band, violet; narrow bands or dots, green, brown, black, green. The significant figures are 5, 1 (51) and the decimal multiplier is 1, so the capacitance is 51 pF. The temperature coefficient is −750 parts per million per degree C., as given by the broad band, the capacitance tolerance is ±5%.

Fixed Composition Resistors

Composition resistors (including small wire-wound units molded in cases identical with the composition type) are color-coded as shown in Fig. 18-12. Colored bands are used on resistors having axial leads; on radial-lead resistors the colors are placed as shown in the drawing. When bands are used for color coding the body color has no significance.

Examples: A resistor of the type shown in the lower drawing of Fig. 18-12 has the following color bands: A, red; B, red; C, orange; D, no color. The significant figures are 2, 2 (22) and the decimal multiplier is 1000. The value of resistance is therefore 22,000 ohms and the tolerance is ±20%.

A resistor of the type shown in the upper drawing has the following colors: body (A), blue; end (B), gray; dot, red; end (D), gold. The significant figures are 6, 8 (68) and the decimal multiplier is 100, so the resistance is 6800 ohms. The tolerance is ± 5%.

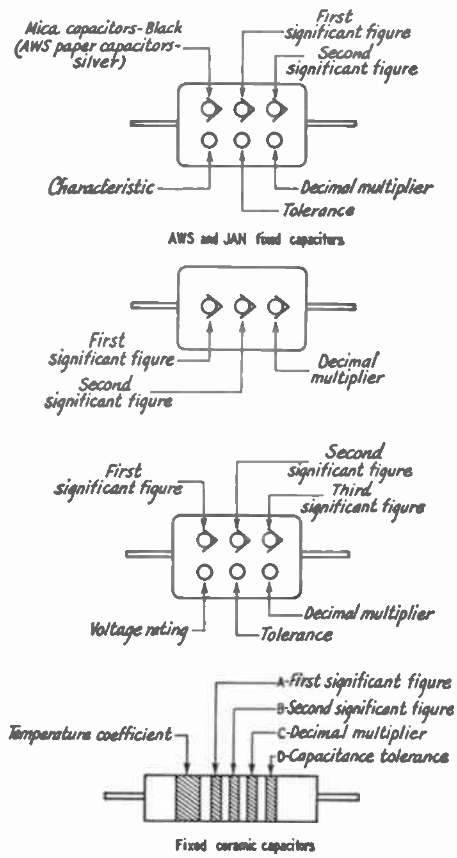


Fig. 18-11 — Color coding of fixed mica, molded paper and tubular ceramic capacitors. The color code for mica and molded paper capacitors is given in Table 18-III. Table 18-IV gives the color code for tubular ceramic capacitors.

TABLE 18-III

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.

TABLE 18-IV

Color Code for Ceramic Capacitors

Color	*Significant Figure*	*Decimal Multiplier*	Capacitance Tolerance *More than 10 pF (in %)*	Capacitance Tolerance *Less than 10 pF (in pF)*	*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

TABLE 18-V

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*

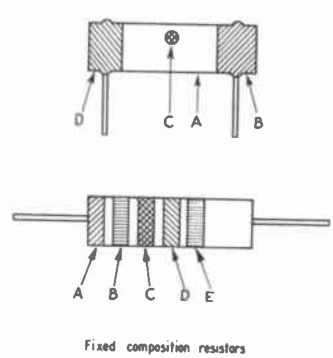


Fig. 18-12 – Color coding of fixed composition resistors. The color code is given in Table 18-III. The colored areas have the following significance:
A – First significant figure of resistance in ohms.
B – Second significant figure.
C – Decimal multiplier.
D – Resistance tolerance in percent. If no color 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.

TABLE 18-VI

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 1st color.

TABLE 18-VII

Metric Multiplier Prefixes

Multiples and submultiples of fundamental units (e.g., ampere, farad, gram, meter, watt)) may be indicated by the following prefixes.

Prefix	*Abbreviation*	*Multiplier*
tera	T	10^{12}
giga	G	10^{9}
mega	M	10^{6}
kilo	k	10^{3}
hecto	h	10^{2}
deci	d	10^{-1}
centi	c	10^{-2}
milli	m	10^{-3}
micro	μ	10^{-6}
nano	n	10^{-9}
pico	p	10^{-12}

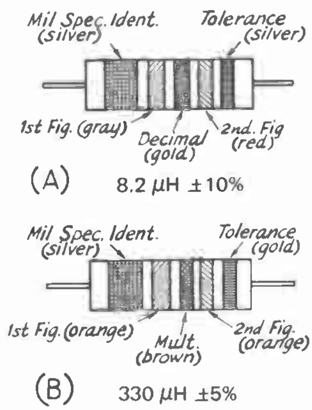


Fig. 18-13 — Color coding for tubular encapsulated rf chokes. At A, an example of the coding for an 8.2-µH choke is given. At B, the color bands for a 330-µH inductor are illustrated.

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.

PILOT-LAMP DATA

Lamp No.	*Bead Color*	*Base (Miniature)*	*Bulb Type*	RATING *Volts*	RATING *Amp.*
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[5]	White	Screw	T-3 1/4	2.9	0.17
292A[5]	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 Replace with No. 48.
5 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.

FINDING PARTS

No chapter on construction would be complete without information on where to buy parts. Amateurs, on a dwarfed scale, must function as purchasing agents in these perplexing times. A properly equipped buyer maintains as complete a catalog file as possible. Many of the companies listed in Chart I will provide free catalogs upon written request. Others may charge a small fee for catalogs. Mail ordering, especially for those distant from metropolitan areas, is today's means to the desired end when collecting component parts for an amateur project. Prices are, to some extent, competitive. A wise buyer will study the catalogs and select his merchandise accordingly.

Delays in shipment can be lessened by avoiding the use of personal checks when ordering. Bank or postal money orders are preferred by most distributors. Personal checks often take a week to clear, thereby causing frustrating delays in the order reaching you.

FREQUENCY-SPECTRUM REFERENCE CHART

Non-amateur Channel Assignments and Other Frequency Data

f(kHz)

15.734264 ±.000044 TV hor. scan freq.
17.8 (0.5)[a] NAA Cutler, ME.
18.6 (0.5)[a] NPG/NLK Jim Creek, WA.
21.4 (0.5)[a] NSS Annapolis, MD.
24.0 (0.5)[a] NBA Balboa, Panama, C.Z.
26.1 (0.5)[a] NPM, Hawaii.
60.0 (0.5)[a,b] WWVB Ft. Collins. CO.
85 Receiver i-f (command set or "Q5er").
100.0 (0.5)[a] Loran C (regional).
179 WGU-20 CD Station, East Coast. Bc of WX and time (a-m).
285 – 325 Marine RDF band. Two cw tones 1020 Hz apart.
285 – 405 Aero RDF; aero WX (a-m) 325 – 405.
415 – 490 Marine (cw).
455 Receiver i-f/mech. filters (Collins).
535 – 1605 Bc (a-m), 107 chans. every 10 kHz from 540 (carrier).

f(MHz)

1.8 – 2.0 Loran A (pulse xsm).
2.5 (0.5)[a,b] WWV, Ft. Collins, CO. WWVH Hawaii.
3.33 (50)[a,b] CHU Ottawa, Canada.
3.395 Transceiver i-f (Heath, Kenwood).
3.579545 ±10^{-5} TV chrominance subcarrier.
5.0 (0.5)[a,b] WWV, WWVH.
5.645 Receiver i-f (Drake).
7.335 (50)[a,b] CHU.
9.0 Xtal filters (KVG).
10.0 (0.5)[a,b] WWV, WWVH.
10.7 Receiver i-f (fm bc).
14.67 (50)[a,b]CHU.
15.0 (0.5)[a,b] WWV.
20.0 (0.5)[a,b] WWV.
25.0 (0.5)[a,b] WWV.
26.965 – 26.985 Citizens Band, chan. 1 – 3 (10-kHz sep.).
27.005 – 27.035 CB, chan. 4 – 7.
27.055 CB, chan. 8.
27.075 – 27.085 CB, chan. 10 – 11.
27.105 – 27.135 CB, chan. 12 – 15.
27.155 – 27.185 CB, chan. 16 – 19.
27.205 – 27.225 CB, chan. 20 – 22.
27.255 CB, chan. 23.
41.25 TV sound carrier (location in receiver i-f).
42.17 TV color subcarrier (location in receiver i-f).
45.75 TV picture carrier (location in receiver i-f).
54 – 72 TV chan. 2 – 4. (Three 6-MHz chans. starting from 54).
72, 75 RC chans.
76 – 88 TV chan. 5 – 6.
88.1 – 107.9 Bc (fm) 100 chan. from 88.1 (carrier) with 200-kHz sep.
120 – 130 Aero; RDF WX.
137.5, 137.62 WX Sat. (A4) ref. W1AW Bul. for orb. data.
162.4 Marine WX bc (fm, regional).
174 – 216 TV chan. 7 – 13.
470 – 890 TV chan. 14 – 83 (70 chan. 6-MHz wide).

[a]Standard-Frequency Transmission figure in brackets is error in parts 10^{10} (*Electronics Engineers' Handbook*, McGraw Hill, pp. 1-48).

[b]Standard time station. A3 xsms include time, WX, and propagation on WWV/WWVH. A3 time xsms on CHU (English/French). WWVB has no A3; info in BCD format generated by reducing carrier by 10 dB (binary 0).

Chart I

Code	Price	Supplier
E	* stamp ** $5	Aldelco P. O. Box 341 Lynbrook, NY 11563
L	* $1 ** $10	Allied Electronics 2400 W. Washington Blvd. Chicago, IL 60612
H		All Star Products, Inc. PO Box 487 Defiance, OH 43512
L	* free **	Amateur Electronic Supply 4828 W. Fond du Lac Ave. Milwaukee, WI 53216
B	* free ** none	Amidon Associates 12033 Otsego Street N. Hollywood, CA 91607
L	* free **	AM Tech PO Box 624 Marion, OH 52302
M, N	* free **	Andy Electronics 6427 Springer Houston, TX 77017
O	* free ** none	Atlantic Surplus Sales 580 Third Avenue Brooklyn, NY 11215
A, E, I	* free ** none	ATV Research 13th & Broadway Dakota City, NE 68731
B	* free ** none	Barken Electronics 274 Mt. Pleasant Ave. Livingston, NJ 07039
A	* free **	Barker & Williamson, Inc. Canal St. Bristol, PA 19007
L, M, N	** 50¢ * $5	Barry Electronics 512 Broadway New York, NY 10012
M, N	* free ** $10	Budget Electronics 2704 West North Avenue Chicago, IL 60647
C	* free ** none	Circuit Board Specialists P. O. Box 969 Pueblo, CO 81002
L,A	* free **	Cambridge Thermionic Corp. 445 Concord Ave. Cambridge, MA 02138
C, P	* free ** none	Circuit Board Specialists 3011 Norwich Ave. Pueblo, CO 81008
A, E	* free ** none	Circuit Specialists Co. PO Box 3047 Scottsdale, AZ 85257
J,M	* free **	Theodore E. Dames Co. 308 Hickory St. Arlington, NJ 07032
I, M, N	* free ** $3	Delta Electronics Co. PO Box 1 Lynn, MA 01903
L	* free ** none	Dominion Radio & Elect. Co. 535 Yonge St. Toronto, Ontario, CANADA
L	* free **	Electronics Distributors, Inc. 1960 Peck Street Muskegon, MI 49441
L	* free ** $5	Electro-Sonic Supply 543 Yonge St. Toronto, Ontario, CANADA
B, J	* free ** none	E. S. Electronic Labs Box 434 Excelsior Springs, MO 64024
M, N	* free **	Fair Radio Sales Box 1105 Lima, OH 45902
A	* free ** $5	Gerber 852 Providence Hwy. Dedham, MA 02026
P	* free **	Gregory Electronics Corp. 249 Rte. 46 Saddle Brook, NJ 07662
A,H,O	* free ** none	HAL Devices Box 365 Urbana, IL 61801
L	* free **	Ham Radio Center 8342 Olive Blvd. St. Louis, MO 63132
I K	* free ** $10	Hammond Transformer 394 Edinburgh Rd. N. Guelph, Ontario, CANADA
		U.S. Distributor for Hammond: Genesee Radio Co. 2550 Delaware Ave. Buffalo, NY 14216
L	* none **	Harrison Radio 20 Smith Street Farmingdale, L.I., NY 11735
A	* free **	Heath Co. Benton Harbor, MI 49022
L	* none **	Henry Radio 11240 W. Olympic Blvd. Los Angeles, CA 92801
L	* 25¢ ** $1.00	Hobby Industries Box 864 Council Bluffs, IA 51501
D	* free ** $5	International Crystal Co. 10 N. Lee Street Oklahoma City, OK 73102
D	* free **	JAN Crystals 2400 Crystal Drive Ft. Myers, FL 33901
A, M, N	* 25¢ ** $2	Jeff-Tronics 4252 Pearl Road Cleveland, OH 44109
C	* free ** none	Kepro Circuit Systems; 3630 Scarlet Oak St. St. Louis, MO 63122
F	* free ** $10	Kirk Electronics Division Electrotec Corp. 400 Town St. East Haddam, CT 06423
L	* free ** none	Lafayette Radio Elect 111 Jericho Tpk. Svosset, L.I., NY 11791 (See local phone directory)
M, N	* free ** $5	John Meshna, Jr. Box 62 E. Lynn, MA 01904
J	* free **	MFJ Enterprises PO Box 494 Mississippi State, MS 39762
A, G, H	* free ** $5	James Millen Mfg. Co. 150 Exchange Street Malden, MA 02148
A, B, H	* free **	J. W. Miller Company 19070 Reyes Avenue Compton, CA 90224
C		D.L. McClaren 19721 Maplewood Ave. Cleveland, OH 44135
A	* free ** $5	Nurmi Electronic Supply 1727 Donna Rd. West Palm Beach, FL 33401
B	* free ** none	Palomar Engineers Box 455 Escondido, CA 92025
C	* free ** none	PBI Electronics P. O. Box 5 Azuso, CA 91702
L	* none ** none	C. M. Peterson Co. Ltd. 220 Adelaide St., N. London, Ont. N6B3H4 CANADA
J	* free **	Piezo Technology, Inc. Box 7877 Orlando, FL 32804
E, M	* 15¢ ** none	Poly Paks Box 942 Lynnfield, MA 01940
M, N	* free ** $2	Precision Systems PO Box 6, Murray Hill, NJ 07974
D	* free ** none	Savoy Electronics, Inc Box 7127 Ft. Lauderdale, FL 33304
D	* free ** none	Sentry Mfg. Co. Crystal Park Chickasha, OK 73108
F	* free ** $10	Skylane Products 406 Bon Air Avenue Temple, Terrace, FL 33617
A, P	* free ** none	Spectronics, Inc. 1009 Garfield Street Oak Park, IL 60304
J	* free ** none	Spectrum International PO Box 1084 Concord, MA 01742
M, N	* free ** $4	Star Tronics Box 17127 Portland, OR 97217
C		Charles R. Sempirek Route 3, Box 1 Bellaire, OH 43906
A	* free ** none	Solid State Systems, Inc. 800 N. Providence Rd. Columbia, MO 65201
O		Teletype Corp., 5555 Touhy A. Skokie, IL 60076
K, A		Ten-Tec Inc. Highway 411, E. Sevierville, TN 37862
O	* sase ** none	Typetronics Box 8873 Ft. Lauderdale, FL 33310
A, B, E H	* free ** none	G. R. Whitehouse 11 Newberry Dr. Amherst, NH 03031
E, M	* free ** none	Weinschenker, K3DPJ Box 353 Irwin, PA 15642

Chart I Coding

A – New Components
B – Toroids and Ferrites
C – Etched-circuit board materials
D – Transmitting and receiving crystals
E – Solid-state devices
F – Antenna hardware
G – Dials and knobs
H – Variable capacitors
I – Transformers
J – I-f filters
K – Cabinet and boxes.
L – All of above, general distributor
M – Surplus parts
N – Surplus assemblies
O – RTTY equipment and parts
P – Surplus fm gear and parts

* Catalog price
** Minimum order

To the best of our knowledge, the suppliers shown in Chart I are willing to sell components to amateurs in small quantities by mail. This listing does not necessarily indicate that these firms have the approval of ARRL.

COPPER-WIRE TABLE

Wire Size A.W.G. (B&S)	Diam. in Mils [1]	Circular Mil Area	Turns per Linear Inch [2] Enamel	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.	Diam. in mm.	Nearest British S.W.G. No.
1	289.3	83690	—	—	—	—	—	3.947	.1264	119.6	7.348	1
2	257.6	66370	—	—	—	—	—	4.977	.1593	94.8	6.544	3
3	229.4	52640	—	—	—	—	—	6.276	.2009	75.2	5.827	4
4	204.3	41740	—	—	—	—	—	7.914	.2533	59.6	5.189	5
5	181.9	33100	—	—	—	—	—	9.980	.3195	47.3	4.621	7
6	162.0	26250	—	—	—	—	—	12.58	.4028	37.5	4.115	8
7	144.3	20820	—	—	—	—	—	15.87	.5080	29.7	3.665	9
8	128.5	16510	7.6	—	7.1	73	46	20.01	.6405	23.6	3.264	10
9	114.4	13090	8.6	—	7.8	—	—	25.23	.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	.912	20
20	32.0	1022	29.4	25.8	23.8	11	7.5	323.4	10.35	1.46	.812	21
21	28.5	810	33.1	—	26.0	—	—	407.8	13.05	1.16	.723	22
22	25.3	642	37.0	31.3	30.0	—	5	514.2	16.46	.918	.644	23
23	22.6	510	41.3	—	37.6	—	—	648.4	20.76	.728	.573	24
24	20.1	404	46.3	37.6	35.6	—	—	817.7	26.17	.577	.511	25
25	17.9	320	51.7	—	38.6	—	—	1031	33.00	.458	.455	26
26	15.9	254	58.0	46.1	41.8	—	—	1300	41.62	.363	.405	27
27	14.2	202	64.9	—	45.0	—	—	1639	52.48	.288	.361	29
28	12.6	160	72.7	54.6	48.5	—	—	2067	66.17	.228	.321	30
29	11.3	127	81.6	—	51.8	—	—	2607	83.44	.181	.286	31
30	10.0	101	90.5	64.1	55.5	—	—	3287	105.2	.144	.255	33
31	8.9	80	101	—	59.2	—	—	4145	132.7	.114	.227	34
32	8.0	63	113	74.1	62.6	—	—	5227	167.3	.090	.202	36
33	7.1	50	127	—	66.3	—	—	6591	211.0	.072	.180	37
34	6.3	40	143	86.2	70.0	—	—	8310	266.0	.057	.160	38
35	5.6	32	158	—	73.5	—	—	10480	335	.045	.143	38-39
36	5.0	25	175	103.1	77.0	—	—	13210	423	.036	.127	39-40
37	4.5	20	198	—	80.3	—	—	16660	533	.028	.113	41
38	4.0	16	224	116.3	83.6	—	—	21010	673	.022	.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 c.m. are commonly used.

SEMICONDUCTOR DIODE COLOR CODE

The "1N" prefix is omitted. A double-width band, which also identifies the cathode terminal end of the diode, is usually used as the first band. (An alternative method uses equal band widths with the set clearly grouped toward the cathode end.) The code is read starting at the cathode end.

Diodes having two-digit numbers are coded with a black band followed by second and third bands. A suffix letter is indicated by a fourth band.

Diodes with three-digit numbers are coded with the sequence numbers in the first, second and third bands. Any suffix letter is indicated by a fourth band.

Diodes with four-digit numbers are coded by four bands followed by a black band. A suffix letter is indicated by a fifth band replacing the black band.

The color code (numbers) is the same as the resistor-capacitor code. The suffix-letter code is A—brown, B—red, C—orange, D—yellow, E—green, and F—blue.

Wave Propagation

Though great advances have been made in recent years in understanding the many modes of propagation of radio waves, variables affecting communication over appreciable distances are very complex, and not entirely predictable. Amateur attempts to schedule operating time and frequencies for optimum results may not always succeed, but familiarity with the nature of radio propagation can certainly reduce the margin of failure and add greatly to one's enjoyment of the pursuit of any kind of DX.

The sun, ultimate source of life and energy on earth, dominates all radio communication beyond the local range. Conditions vary with such obvious sun-related earthly cycles as time of day and season of the year. Since these differ for appreciable changes in latitude and longitude, almost every communications circuit is unique in some respects. There are also short- and long-term solar cycles that influence propagation in less obvious ways. Furthermore, the state of the sun at a given moment is critical to long-distance communication, so it is understandable that propagation forecasting is still a rather inexact science.

With every part of the radio spectrum open to our use differing in its response to solar phenomena, amateurs have been, and still are, in a position to contribute to advancement of the art, both by accident and by careful investigation.

SOLAR PHENOMENA

Man's interest in the sun is older than recorded history. Sunspots were seen and discussed thousands of years ago, and they have been studied since Galileo observed them with the first telescope ever made. Records of sunspot observations translatable into modern terms go back nearly 300 years. Current observations are statistically "smoothed" to maintain a continuous record, in the form of the *Zurich Sunspot Number*, on which propagation predictions mentioned later are based.

A useful modern indication of overall solar activity is the *solar flux index*. Solar flux (noise) is measured on various frequencies in many places. A 2800-MHz measurement made several times daily in Ottawa is transmitted hourly by WWV. Because it is essentially current information, directly related to the sunspot number (see Fig. 19-1) and more immediately useful, it tends to displace the latter as a means of predicting propagation conditions.

SUNSPOT CYCLES

Even before their correlation with radio propagation variations was well-known, the periodic rise and fall of sunspot numbers had been studied for many years. These cycles average roughly 11 years in length, but have been as short as 9 and as long as 13 years. The highs and lows of the cycles also vary greatly. Cycle 19 peaked in 1958 with a sunspot number of over 200. Cycle 20, of nearer average intensity, reached 120 in 1969. By contrast, one of the lowest, Cycle 14, peaked at only 60 in 1907.[1] Several cycle lows have not reached zero levels on the Zurich scale for any appreciable period, while others have had several months of little or no activity.

Sunspot cycles should not be thought of as having sine-wave shape. There can be isolated highs during the normally low years. A remarkable example was a run of several days in October, 1974, only a few months from the approximate bottom of Cycle 20, when the solar flux reached 145, a level well above the highs of several cycles on record. Only 5 months later, several days of solar flux below 70 were recorded.

SOLAR RADIATION

Insofar as it affects most radio propagation, solar radiation is of two principal kinds: ultraviolet light and charged particles. The first travels at just under 300,000,000 meters (186,000 miles) per

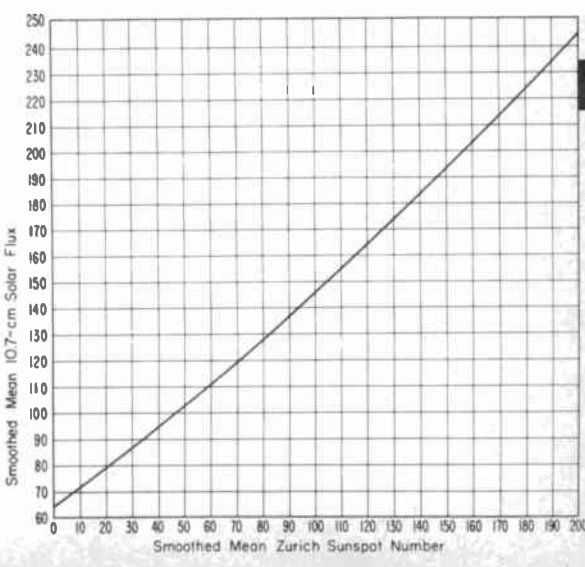


Fig. 19-1 — Relationship between smoothed mean Zurich sunspot number and the 2800-MHz solar flux. Highest solar flux recorded in 1974, Oct. 12, was 145, the equivalent of a sunspot number of 100. Lowest flux value in 1975 (early June) was 66, equating with a sunspot number very close to zero.

Fig. 19-2 — W1HDQ and W1SL look for sunspots with a simple projection system. The baffle at the top end of the small telescope provides a shaded area for viewing the sun's image (light circle) on the projection surface. Sunspots large enough to affect radio propagation are easily seen with this viewing system.

second, as does all electromagnetic radiation, so UV effects on wave propagation develop simultaneously with increases in observed solar noise, approximately 8 minutes after the actual solar event. Particle radiation moves more slowly, and by varying routes, so it may take up to 40 hours to affect radio propagation. Its principal effects are high absorption of radio energy and the production of auroras, both visual and the radio variety.

Variations in the level of solar radiation can be gradual, as with the passage of some sunspot groups and other long-lived activity centers across the solar disk, or sudden, as with solar flares. An important clew for anticipating variations in solar radiation levels and radio propagation changes resulting from them is the rotational period of the sun, approximately 27 days. Sudden events (flares) may be short-lived, but active areas capable of influencing radio propagation may recur at 4-week intervals for 4 or 5 solar rotations. Evidence of the "27-day cycle" is most marked during years of low solar activity.

Information on the condition of the sun, as it affects radio propagation, can be obtained in several ways. Projection of the sun's image as in Fig. 19-2 is particularly useful in the low years of the "11-year" cycle. At other times visible evidence of solar activity may be more difficult to sort out. Enough definition for our purposes is possible with the simplest telescopes. Low-cost instruments, 10 to 30-power, are adequate. A principal requirement is provision for mounting on a tripod having a pan-tilt head.[2]

Adjust the aiming to give a circular shadow of the scope body, then move the scope slowly until a bright spot appears on the projection surface. Put a baffle on the scope to enlarge the shaded area and adjust the focus to give a sharp-edged image of the solar disk. If there are any sunspots you will see them now. Draw a rough sketch of what you see, every time an observation is made, and keep it with your record of propagation observations.

Spots move across the image from left to right, as it is viewed with the sun at the observer's back. The line of movement is parallel to the solar equator. Not all activity capable of affecting propagation can be seen, but any spots seen have significance. Active areas may develop before spots are visible and may persist after spots associated with them are gone, but once identified by date they are likely to recur about 27 days later, emphasizing the worth of detailed records.

Variations in solar noise may be observed by aiming the antenna at the rising or setting sun. Sudden large increases may be heard regardless of the antenna position. Such bursts are often heard, but seldom recognized for what they are — warnings of imminent changes in propagation.

Vhf or uhf arrays capable of movement in elevation as well as azimuth are useful for solar noise monitoring. With a good system, the "quiet sun" can be "heard" at a low level.[3] Bursts that can be many dB higher indicate the start of a major event, such as a solar flare capable of producing an hf blackout and possibly vhf auroral propagation.

CHARACTERISTICS OF RADIO WAVES

All electromagnetic waves are moving fields of electric and magnetic force. Their lines of force are at right angles, and are mutually perpendicular to the direction of travel. They can have any position with respect to the earth. The plane containing the continuous lines of electric and magnetic force is called the *wave front*.

The medium in which electromagnetic waves travel has a marked influence on their speed of movement. In empty space the speed, as for light, is just under 300,000,000 meters per second. It is slightly less in air, and it varies with temperature and humidity to a degree, depending on the frequency. It is much less in dielectrics, where the speed is inversely proportional to the square root of the dielectric constant of the material.

Waves cannot penetrate a good conductor to any extent because the electric lines of force are practically short-circuited. Radio waves travel through dielectric materials with ease.

POLARIZATION

If the lines of force in the electric field are perpendicular to the surface of the earth the wave is said to be vertically polarized. If parallel with the earth, the polarization is said to be horizontal. It is possible to generate waves with rotating field lines. Known as circular polarization, this is useful in satellite communication, where polarization tends to be random. When the earth's surface is not available as a reference, polarization not of a rotating nature is described as linear or plane polarization, rather than vertical or horizontal, which become meaningless. Circular polarization is usable with plane-polarized antennas at the other end of the circuit, though with some small loss on most paths.

TYPES OF PROPAGATION

Depending on the means of propagation, radio waves can be classified as ionospheric, tropospheric, or ground waves. The ionospheric or sky wave is that main portion of the total radiation leaving the antenna at angles somewhat above the horizontal. Except for the reflecting qualities of the ionosphere, it would be lost in space. The tropospheric wave is that portion of the radiation kept close to the earth's surface as the result of bending in the lower atmosphere. The ground wave is that portion of the radiation directly affected by the surface of the earth. It has two components, an earth-guided surface wave, and the space wave, the latter itself being the resultant of two components, direct and ground-reflected. The terms "tropospheric wave" and "ground wave" are often used interchangeably, though this is not strictly correct.

THE IONOSPHERE

Long-distance communication and much over shorter distances, on frequencies below 30 MHz, is the result of bending of the wave in the ionosphere, a region between about 60 and 200 miles above the earth's surface where free ions and electrons exist in sufficient quantity to affect the direction of wave travel. Without the ionosphere, DX as we know it would be impossible.

Ionization of the upper atmosphere is attributed to ultraviolet radiation from the sun. The result is not a single region, but several layers of varying densities at various heights surrounding the earth. Each layer has a central region of relatively dense ionization that tapers off both above and below.

IONOSPHERIC LAYERS

The lowest useful region of the ionosphere is called the *E* layer. Its average height of maximum ionization is about 70 miles. The atmosphere here is still dense enough so that ions and electrons set free by solar radiation do not have to travel far before they meet and recombine to form neutral particles, so the layer can maintain its ability to bend radio waves only when continuously in sunlight. Ionization is thus greatest around local noon, and it practically disappears after sundown.

In the daylight hours there is a still lower area called the *D* region where ionization is proportional to the height of the sun. Wave energy in the two lowest frequency amateur bands, 1.8 and 3.5 MHz, is almost completely absorbed by this layer. Only the highest angle radiation passes through it and is reflected back to earth by the *E* layer. Communication on these bands in daylight is thus limited to short distances, as the lower angle radiation needed for longer distances travels farther in the *D* region and is absorbed.

The region of ionization mainly responsible for long-distance communication is called the *F* layer. At its altitude, about 175 miles at night, the air is so thin that recombination takes place very slowly. Ionization decreases slowly after sundown, reaching a minimum just before sunrise. The obvious effect of this change is the early disappearance of long-distance signals on the highest frequency that was usable that day, followed by loss of communication on progressively lower frequencies during the night. In the daytime the *F* layer splits into two parts, *F*1 and *F*2, having heights of about 140 and 200 miles, respectively. They merge again at sunset.

Scattered patches of relatively dense ionization develop seasonally at *E*-layer height. Such *sporadic E* is most prevalent in the equatorial regions, but it is common in the temperate latitudes in late spring and early summer, and to a lesser degree in early winter. Its effects become confused with those of other ionization on the lower amateur frequencies, but they stand out above 21 MHz, especially in the low-activity years of the solar cycle, when other forms of DX are not consistently available.

Duration of openings decreases and the length of skip increases with progressively higher frequencies. Skip distance is commonly a few hundred miles on 21 or 28 MHz, but multiple hop propagation can extend the range to 2500 miles or more. June and July are the peak months in the northern hemisphere. E_s propagation is most common in midmorning and early evening, but may extend almost around the clock at times. The highest frequency for E_s is not known, but the number of opportunities for using the mode drops off rapidly between the amateur 50- and 144-MHz bands, whereas 28 and 50 MHz are quite similar.

The greater the intensity of ionization in a layer, the more the wave path is bent. The bending also depends on wavelength; the longer the wave the more its path is modified for a given degree of ionization. Thus, for a given level of solar radiation, ionospheric communication is available for a longer period of time on the lower-frequency amateur bands than on those near the upper limit of hf spectrum. The intensity and character of solar radiation are subject to many short-term and long-term variables, the former still predictable with only partial success.

ABSORPTION

In traveling through the ionosphere, a radio wave gives up some of its energy by setting the ionized particles in motion. When moving particles collide with others, this energy is lost. Such *absorption* is greater at lower frequencies. It also increases with the intensity of ionization, and with the density of the atmosphere. This leads to a propagation factor often not fully appreciated: *signal levels and quality tend to be best when the operating frequency is near the maximum that is reflected back to earth at the time.*

VIRTUAL HEIGHT

An ionospheric layer is a region of considerable depth, but for practical purposes it is convenient to think of it as having finite height, from which a simple reflection would give the same effects (observed from the ground) as result from the gradual bending that actually takes place. It is given several names, such as *group height*, *equivalent height*, and *virtual height*.

The virtual height of an ionospheric layer for various frequencies and vertical incidence is determined with a variable-frequency sounding device that directs pulses of energy vertically and measures the time required for the round-trip path shown at the left in Fig. 19-3. As the frequency rises, a point is reached where no energy is returned vertically. This is known as the *critical frequency*, for the layer under consideration. A representation of a typical *ionogram* is shown in Fig. 19-4.[4] In this sounding the virtual height for 3.5 to 4 MHz was 400 km. Because the ionogram is a graphical presentation of wave travel time, double-hop propagation appears as an 800-km return for the same frequency. The critical frequency was just over 5 MHz on this occasion. Such a clear *F*-layer ionogram is possible only under magnetically quiet conditions, and at night, when little or no *E*- and *D*-layer ionization is present.

EFFECTS OF THE EARTH'S MAGNETIC FIELD

The ionosphere has been discussed thus far in terms of simple bending, or refraction, a concept useful for some explanatory purposes. But an understanding of long-distance propagation must take the earth's magnetic field into account. Because of it, the ionosphere is a birefringent medium (doubly refracting) which breaks up plane-polarized waves into what are known as the *ordinary* and *extraordinary waves*, f_oF2 and f_xF2 in the ionogram. This helps to explain the dispersal of plane polarization encountered in most ionospheric communication.[5]

Sudden marked increases in solar radiation, such as with solar flares, trigger instantaneous effects in the *F*, *E*, and *D* regions; slightly delayed effects, mainly in the polar areas; and geomagnetic effects, delayed up to 40 hours.

Onset of the *D*-region absorption is usually sudden, lasting a few minutes to several hours, leading to use of the term SID (sudden ionospheric disturbance). Shortwave fadeouts (SWFs) and SIDs exhibit wide variations in intensity, duration, and number of events, all tending to be greater in periods of high solar activity. Though their effects on radio propagation are of great importance, solar flares and associated disturbances are among the least predictable of solar-induced communications variables.

RADIATION ANGLE AND SKIP DISTANCE

The lower the angle above the horizon at which a wave leaves the antenna, the less refraction in the ionosphere or troposphere is required to bring it back, or to maintain useful signal levels in the case of tropospheric bending. This results in the emphasis on low radiation angles in the pursuit of DX, on the hf or vhf bands. It is rarely possible to radiate energy on a line tangent to the earth's surface, but even when this is done some bending is still required for communication over appreciable distances, because of earth curvature.

Some of the effects of radiation angle are illustrated in Fig. 19-3. The high-angle wave at the left is bent only slightly in the ionosphere, and so goes through it. The wave at the somewhat lower

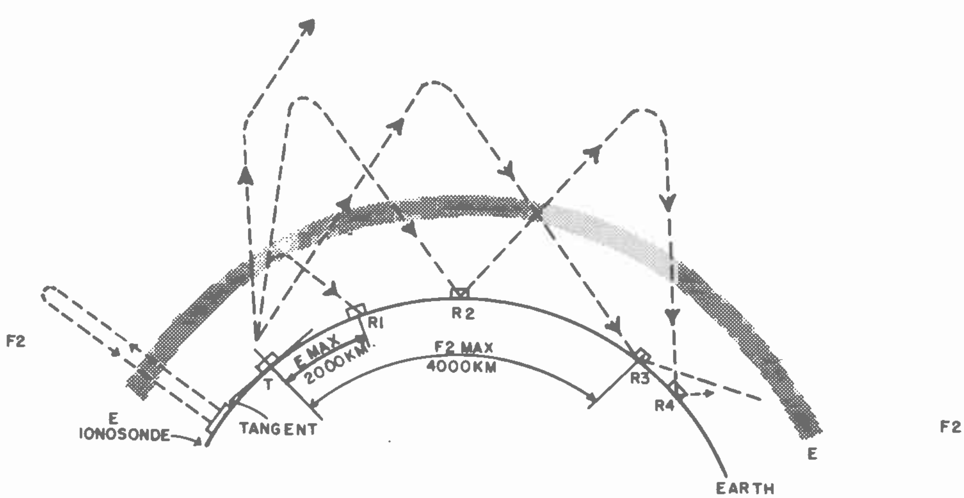


Fig. 19-3 — Three types of ionospheric propagation. Sounder, left, measures virtual-height and critical frequency of F_2 layer. Transmitter T is shown radiating at three different angles. Highest passes through the ionosphere after slight refraction. Lower-angle wave is returned to earth by the *E* layer, if frequency is low enough, at a maximum distance of 2000 kM. The *F*-layer reflection returns at a maximum distance of about 4000 kM, depending on the radiation angle. It is shown traversing a second path (double hop) from R2 to R4, the latter beyond single-hop range. The lowest-angle wave reaches the maximum practical single-hop distance at R3.

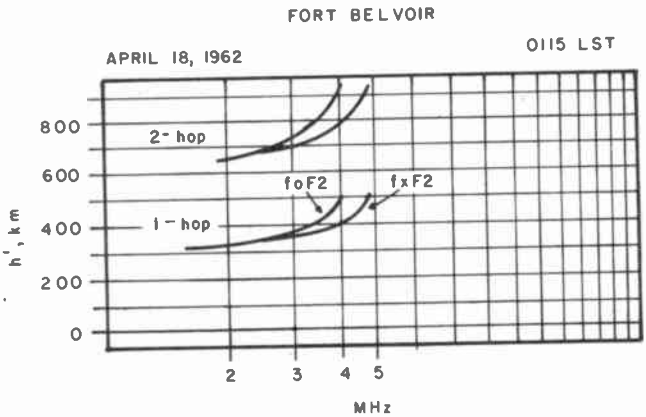


Fig. 19-4 — *F*-layer ionogram taken at night during magnetically quiet conditions. The traces show the breaking up into ordinary and extraordinary waves. Because it required twice the travel time, the double-hop return appears as having come from twice the height of the single-hop.

angle is just capable of being returned by the ionosphere. In daylight it might be returned via the *E* layer. Its area of return from the *F* layer, R2, is closer to the transmitting point, T, than is that of the lowest-angle wave. If R2 is at the shortest distance where returned energy is usable, the area between R1 and the outer reaches of the ground wave, near the transmitter, is called the *skip zone*. The distance between R2 and T is called the *skip distance*. The distances to both R1 and R2 depend on the ionization density, the radiation angle at T, and the frequency in use. The maximum distance for single-hop propagation via the *F* layer is about 2500 miles (4000 kilometers). The maximum *E*-layer single hop is about 1250 miles (2000 kilometers).

The maximum usable frequency (muf) for *F*-layer communication is about 3 times the critical frequency for vertical return, as at the left in Fig. 19-3. For *E*-layer propagation it is about 5 times.

MULTIPLE-HOP PROPAGATION

On its return to earth, the ionospherically propagated wave can be reflected back upward near R1 or R2, travel again to the ionosphere, and be refracted back to earth. This process can be repeated several times under ideal propagation conditions, leading even to communication over distances well beyond halfway around the world. Ordinarily ionospheric absorption and ground-reflection losses exact tolls in signal level and quality, so multiple-hop propagation usually yields lower signal levels and more distorted modulation than single-hop. This is not always the case, and under ideal conditions even long-way-around communication is possible with good signals. There is evidence to support the theory that signals for such communications, rather than hopping, may be ducted through the ionosphere for a good part of the distance.

FADING

Two or more parts of the wave may follow different paths, causing phase differences between wave components at the receiving end. Total field strength may be greater or smaller than that of one component. Fluctuating signal levels also result from the changing nature of the wave path, as in the case of moving air-mass boundaries, in tropospheric propagation on the higher frequencies. Changes in signal level, lumped under the term *fading*, arise from an almost infinite variety of phenomena; some natural, some man-made. Aircraft reflections are in the latter category.

Under some circumstances the wave path may vary with very small changes in frequency, so that modulation sidebands arrive at the receiver out of phase, causing distortion that may be mild or severe. Called *selective fading*, this problem increases with signal bandwidth. Double-sideband a-m signals suffer much more than single-sideband signals with suppressed carrier do.

THE SCATTER MODES

Much long-distance propagation can be described in terms of discrete reflection, through the analogy is never precise since true reflection would be possible only with perfect mirrors, and in a vacuum. All electromagnetic wave propagation is subject to scattering influences which alter idealized patterns to a great degree. The earth's atmosphere and ionospheric layers are scattering media, as are most objects that intervene in the wave path as it leaves the earth. Strong returns are thought of as reflections and weaker ones as scattering, but both influences prevail. Scatter modes have become useful tools in many kinds of communication.

FORWARD SCATTER

We describe a skip zone as if there were no signal heard between the end of useful ground-wave range and the points R1 or R2 of Fig. 19-3, but actually the transmitted signal can be detected over much of the skip zone, with sufficiently sensitive devices and methods. A small portion of the transmitted energy is scattered back to earth in several ways, depending on the frequency in use.

Tropospheric scatter extends the local communications range to an increasing degree with frequency, above about 20 MHz, becoming most useful in the vhf range. *Ionospheric scatter*, mostly from the height of the *E* region, is most marked at frequencies up to about 60 or 70 MHz. Vhf tropospheric scatter is usable within the limits of amateur power levels and antenna techniques, out to nearly 500 miles. Ionospheric forward scatter is discernible in the skip zone at distances up to 1200 miles or so.

A major component of ionospheric scatter is that contributed by short-lived columns of ionization formed around meteors entering the earth's atmosphere. This can be anything from very short bursts of little communications value to sustained periods of usable signal level, lasting up to a minute or more. Meteor scatter is most common in the early morning hours, and it can be an interesting adjunct to amateur communication at 21 MHz and higher, especially in periods of low solar activity. It is at its best during major meteor showers.[6]

BACKSCATTER

A complex form of scatter is readily observed when working near the maximum usable frequency for the *F* layer at the time. The transmitted wave is refracted back to earth at some distant point, which may be an ocean area or a land mass where there is no use of the frequency in question at the time. A small part of the energy is scattered back to the skip zone of the transmitter, via the ionospheric route.

Backscatter signals are generally rather weak, and subject to some distortion from multipath effects, but with optimum equipment they are usable at distances from just beyond the reliable local range out to several hundred miles. Under ideal conditions backscatter communication is possible over 3000 miles or more, though the term "sidescatter" is more descriptive of what probably happens on such long paths.

The scatter modes contribute to the usefulness of the higher parts of the DX spectrum, especially during periods of low solar activity when the normal ionospheric modes are less often available.

USING WWV BULLETINS

The National Bureau of Standards stations WWV and WWVH (see Chapter 17) transmit hourly propagation bulletins that are very useful for short-term communications planning. At 14 minutes after each hour (WWV only) information is given on expected propagation conditions, the current state of the geomagnetic field, and the solar flux index. At 18 minutes after each hour on WWV and at 46 after on WWVH, a summary for the previous day and a prediction for the current day are given. Detailed information on use of these bulletins appeared in *QST* for June, August, and September, 1975.[7]

PROPAGATION IN THE MF AND HF BANDS

The 1.8-MHz band offers reliable communication over distances up to about 25 miles during daylight. On winter nights ranges up to several thousand miles are possible.

The 3.5-MHz band is seldom usable beyond 200 miles in daylight, but long distances are not unusual at night, especially in years of low solar activity. Atmospheric noise tends to be high in the summer months on both 3.5 and 1.8 MHz.

The 7-MHz band has characteristics similar to 3.5 MHz, except that much greater distances are possible in daylight, and more often at night. In winter dawn and dusk periods it is possible to work the other side of the world, as signals follow the darkness path.

The 14-MHz band is the most widely used DX band. In the peak years of the solar cycle it is open to distant parts of the world almost continuously. During low solar activity it is open mainly in the daylight hours, and is especially good in the dawn and dusk periods. There is almost always a skip zone on this band.

The 21-MHz band shows highly variable propagation depending on the level of solar activity. During sunspot maxima it is useful for long-distance work almost around the clock. At intermediate levels it is mainly a daylight DX band. In the low years it is useful for transequatorial paths much of the year, but is open less often to the high latitudes. Sporadic-*E* skip is common in early summer and midwinter.

The 28-MHz band is excellent for DX communication in the peak solar-cycle years, but mostly in the daylight hours. The open time is shorter in the intermediate years, and is more confined to low-latitude and transequatorial paths as solar activity drops off. For about two years near the solar minimum, *F*-layer openings tend to be infrequent, and largely on north-south paths, with very long skip.

Sporadic-*E* propagation keeps things interesting in the period from late April through early August on this band, and on 21 MHz, providing single-hop communication out to 1300 miles or so, and multiple-hop to 2600 miles. Effects discussed in the following section on vhf propagation also show up in this band, though tropospheric bending is less than on 50 MHz.

THE WORLD ABOVE 50 MHZ

It was once thought that frequencies above 50 MHz would be useful only locally, but increased occupancy and improved techniques turned up many forms of long-distance vhf propagation. What follows supplements information given earlier in this chapter. First, let us consider the nature of our bands above 50 MHz.

50 to 54 MHz This borderline region has some of the characteristics of both higher and lower frequencies. Just about every form of wave propagation is found occasionally in the 50-MHz band, which has contributed greatly to its popularity. Its utility for service-area communication should not be overlooked. In the absence of any favorable condition, the well-equipped 50-MHz station should be able to work regularly over a radius of 75 to 100 miles or more, depending on terrain and antenna size and height.

Changing weather patterns extend coverage to 300 miles or more at times, mainly in the warmer months. Sporadic-E skip provides seasonal openings for work over 400 to 2500 miles, in seasons centered on the longest and shortest days of the year. Auroral effects afford vhf operators in the temperate latitudes an intriguing form of DX up to about 1300 miles. During the peak of "11-year" sunspot cycle 50-MHz DX of worldwide proportions may be workable by reflections of waves by the ionospheric F_2 layer. Various weak-signal scatter modes round out the 50-MHz propagation fare.

144 to 148 MHz Ionospheric effects are greatly reduced at 144 MHz. F-layer propagation is unknown. Sporadic-E skip is rare, and much more limited in duration and coverage than on 50 MHz. Auroral propagation is quite similar to that on 50 MHz, except that signals tend to be somewhat weaker and more distorted at 144. Tropospheric propagation improves with increasing frequency. It has been responsible for 144-MHz work over distances up to 2500 miles, and 500-mile contacts are fairly common in the warmer months. Reliable range on 144 is slightly less than on 50, under minimum conditions.

220 MHz and Higher Ionospheric propagation of the sorts discussed above is virtually unknown above about 200 MHz. Auroral communication is possible on 220 and 420 MHz, but probably not on higher frequencies, with amateur power levels. Tropospheric bending is very marked, and may be better on 432 than on 144 MHz, for example. Communication has been carried on over paths far beyond line of sight, on all amateur frequencies up through 10,000 MHz. Under minimum conditions, signal levels drop off slightly with each higher band.

PROPAGATION MODES

Known means by which vhf signals are propagated beyond the horizon are described below.

F_2-Layer Reflection Most communication on lower frequencies is by reflection of the wave in the F region, highest of the ionized layers. Its density varies with solar activity, the maximum usable frequency (muf) being highest in peak years of the sunspot cycle. Cycle 19 (in the recorded history of sunspot activity) hit an all-time high in the fall of 1958, which may never be equalled within the lifetime of some of us. Cycle 20 produced 50-MHz F_2 DX in 1968 to 1970, but less than Cycle 18 (1946 to 1949), and far less than Cycle 19.

The muf for F_2-layer propagation follows daily, monthly and seasonal cycles, all related to conditions on the sun, as with the hf bands. Frequent checks will show if the muf is rising or falling, and the times and directions for which it is highest. Two-way work has been done over about 1800 to 12,500 miles; even greater, if daylight routes around the earth the long way are included. The muf is believed to have reached about 70 MHz in 1958.

The TE Mode Also associated with high solar activity is a transequatorial mode, having an muf somewhat higher than the F_2. This is observed most often between points up to 2500 miles north and south of the *geomagnetic* equator, mainly in late afternoon or early evening.[8]

Sporadic-E Skip Patchy ionization of the E region of the ionosphere often propagates 28- and 50-MHz signals over 400 to 1300 miles or more. Often called "short skip," this is most common in May, June and July, with a shorter season around year end. Seasons are reversed in the southern hemisphere. E skip can occur at any time or season, but is most likely in mid-morning or early evening. Multiple-hop effects may extend the range to 2500 miles or more.

E_s propagation has been observed in the 144-MHz band, and on TV channels up to about 200 MHz. Minimum skip distance is greater, and duration of openings much shorter, on 144 MHz than on 50. Reception of strong E_s signals from under 300 miles on 50 MHz indicates some possibility of skip propagation on 144, probably to 800 miles or more.

Aurora Effect High-frequency communication may be wiped out or seriously impaired by absorption in the ionosphere, during disturbances associated with high solar activity and variations in the earth's magnetic field. If this occurs at night in clear weather, there may be a visible aurora, but the condition also develops in daylight, usually in late afternoon. Weak wavery signals in the 3.5-MHz and 7-MHz bands are good indicators.

Vhf waves can be returned to earth from the auroral region, but the varying intensity of the aurora and its porosity as a propagation medium impart a multipath distortion to the signal, which garbles or even destroys any modulation. Distortion increases with signal frequency and varies, often quite quickly, with the nature of the aurora. Single-sideband is preferred to modes requiring more bandwidth. The most effective mode is cw, which may be the only reliable communications method at 144 MHz and higher, during most auroras.

Propagation is generally from the north, but probing with a directional array is recommended. Maximum range is about 1300 miles, though 50-MHz signals are heard occasionally over greater distances, usually with little or no auroral distortion.

How often auroral communication is possible is related to the *geomagnetic* latitude of participating stations, auroras being most frequent in northeastern USA and adjacent areas of Canada. They are rare below about latitude 32 in the Southeast and about latitude 38 to 40 in the Southwest. The highest frequency for auroral returns depends on equipment and antennas, but auroral communication has been achieved up to at least 432 MHz.

Tropospheric Bending An easily-anticipated extension of normal vhf coverage results from abrupt changes in the refractive index of the atmosphere, at boundaries between air masses of differing temperature and humidity characteristics. Such warm-dry over cool-moist boundaries often lie along the southern and western edges of stable slow-moving areas of fair weather and high barometric pressure. Troposheric bending can increase signal levels from within the normal working range, or bring in more distant stations, not normally heard.

A condition known as *ducting* or *trapping* may simulate propagation within a waveguide, causing vhf waves to follow earth curvature for hundreds or even thousands of miles. Ducting incidence increases with frequency. It is rare on 50 MHz, fairly common on 144, and more so on higher frequencies. It occurs most often in temperate or low latitudes. It was the medium for the W6NLZ-KH6UK work on 144, 220 and 432 MHz, over a 2540-mile path. Gulf-Coast states see it often, the Atlantic Seaboard, Great Lakes and Mississippi Valley areas occasionally, usually in September and October.

Many local conditions contribute to tropospheric bending. Convection in coastal areas in warm weather; rapid cooling of the earth after a hot day, with upper air cooling more slowly; warming of air aloft with the summer sunrise; subsidence of cool moist air into valleys on calm summer evenings – these familiar situations create upper-air conditions which can extend normal vhf coverage.

The alert vhf enthusiast soon learns to correlate various weather signs and propagation patterns. Temperature and barometric-pressure trends, changing cloud formations, wind direction, visibility and other natural indicators can give him clues as to what is in store in the way of tropospheric propagation.

The 50-MHz band is more responsive to weather effects than 28, and 144 MHz is much more active than 50. This trend continues into the microwave region, as evidenced by tropospheric records on all our bands, up to and including work over a 275-mile path on 10,000 MHz.

The Scatter Modes Though they provide signal levels too low for routine communication, several scatter modes attract the advanced vhf operator.

Tropospheric scatter offers marginal communication up to 500 miles or so, almost regardless of conditions and frequency, when optimum equipment and methods are used.

Ionospheric scatter is useful mainly on 50 MHz, where it usually is a composite of meteor bursts and a weak residual scatter signal. The latter may be heard only when optimum conditions prevail. The best distances are 600 to 1200 miles.

Back scatter, common on lower frequencies, is observed on 50 MHz during ionospheric propagation, mainly of the F_2 variety. Conditions for 50-MHz backscatter are similar to those for the hf bands, detailed earlier in this chapter.

Scatter from meteor trails in the E region can cause signal enhancement, or isolated bursts of signal from a station not otherwise heard. Strength and duration of meteor bursts decrease with increasing signal frequency, but the mode is popular for marginal communication in the 50- and 144-MHz bands. It has been used on 220 MHz, and, more marginally, on 432 MHz.

Random meteor bursts can be heard by cooperating vhf stations at any time or season, but early-morning hours are preferred. Major meteor showers (August Perseids and December Geminids) provide frequent bursts. Some other showers have various periods, and may show phenomenal burst counts in peak years.[6] Distances are similar to other E-layer communication.

All scatter communication requires good equipment and optimum operating methods. The narrow-band modes are superior to wide-band systems.

Communication Via the Moon Though amateurs first bounced signals off the moon in the early 1950s, real communication via the earth-moon-earth (eme) route is a fairly recent accomplishment. Requirements are maximum legal power, optimum receiving equipment, very large high-gain antennas, and precise aiming. Sophisticated tracking systems, narrow bandwidth (with attendant requirements for receiver and transmitter stability) and visual signal-resolution methods are desirable. Lunar work has been done on all amateur frequencies from 50 to 2400 MHz, over distances limited only by the ability of the stations to "see" the moon simultaneously.

For more detailed vhf propagation information and references, see *The Radio Amateur's VHF Manual*, Chapter 2.

PROPAGATION PREDICTION

Information on the prediction of maximum usable frequencies (muf) and optimum working frequencies for F-layer propagation was formerly available from the U.S. Government Printing Office. The material took several forms, as methods developed for military communications use were adapted to worldwide civilian needs. Though the service was terminated in 1975, the basic methods are still of interest. A full description may be found in *QST* for March, 1972.[7] The government information is available in some technical libraries.

Other means are available to amateurs who wish to make their own predictions, both short- and long-term. An appreciable amount of observing and record-keeping time is involved at first, but the work can be streamlined with practice. Many amateurs who try it find the task almost as

TABLE 19-I

Some time and frequency stations useful for propagation monitoring.

Call	*Freq. (kHz)*	*Location*
WWV	2500, 5000, 10,000, 15,000, 20,000 25,000	Ft. Collins, Colorado
WWVH	Same as WWV	Kekaha, Kauai, Hawaii
CHU	3330, 7335, 14,760	Ottawa, Ontario, Canada
RAT, RWM	5000, 10,000, 15,000	Moscow, USSR*
RIM, RCH	2500, 5000, 10,000	Tashkent, USSR*
RID, RKM	5004, 10,004, 15,004	Irkutsk, USSR*
RTA	4996, 9996, 14,996	Novosibirsk, USSR*
ZUO	2500, 5000	Pretoria, South Africa
VNG	7500	Lyndhurst, Australia
BPV	5000, 10,000, 15,000	Shanghai, China
JJY	2500, 5000, 10,000, 15,000	Tokyo, Japan

*Call, from international table, may not check with actual reception. Locations and frequencies appear to be as given.

interesting as any operational success that may result from it. Properly organized, data collection and propagation prediction can become an ideal group project.

Getting Started

Because most factors have well-defined cyclical trends, the first step in propagation prediction is to become familiar with the rhythm of these trends for the geographical location and season under consideration. This job is made easier if we understand the causes of the ups and downs, so familiarity with basic information given earlier in this chapter is helpful.

What frequencies are "open," and where the cutoff in ionospheric propagation lies in the spectrum can be determined quite readily by tuning upward in frequency with a general-coverage receiver, until ionospherically propagated signals are no longer heard. The muf for the day and the times that a given frequency band opens or closes can be found in this way. A daily log will show if conditions are improving or deteriorating.

Listening in the amateur bands and on immediately adjacent frequencies may be the only way to do this, if the receiver is the amateur-bands-only variety. Most DX bands are narrower in other parts of the world than in the Americas, so there is no lack of round-the-clock occupancy by other services, ordinarily. Most receivers also cover somewhat more than the actual amateur assignments, at their widest, so some commercial and governmental signals can be found close by our band edges. A worldwide listing of stations, by frequency, is useful in identifying signals for propagation monitoring purpose.[8] Don't overlook W1AW; frequencies and schedule are listed in every *QST*.

Ability to tune to 5 MHz and multiples thereof, to receive the standard time-and-frequency stations now operating in many parts of the world, is a great aid. See Table 19-I. Most such stations operate continuously, with appreciable power and omnidirectional antennas. WWV and WWVH are excellent indicators, at any suitable distance from Colorado or Hawaii. Their signal behavior can tell the experienced observer at least as much about propagation – at the moment – as does the content of their propagation bulletins. Many receivers can be made to tune some of these frequencies by detuning their front-panel tracking controls. See *QST*, September, 1975, page 23, for suggestions. Simple crystal-controlled converters for the standard frequencies offer another possibility (*QST* for June, 1976, p. 25).

Recurring Phenomena

Because the sun is responsible for all radio-propagation variables, their rhythmic qualities are related to time, season and other sun-earth factors. Some are obvious. Others, particularly the rotational period of the sun, about 27.5 days, show best in long-term chart records kept on a monthly or four-week basis. Recurrence data are used in nearly all prediction work done presently, and the data can yield fair accuracy.

If the muf is high and conditions are generally good for several days, a similar condition is likely to prevail four weeks later, when the same area of the sun will be in view from the earth. Ionospheric disturbances also generally follow the 27-day cycle, though there may be marked differences in level from one period to the next.

Some solar-activity centers are short-lived, lasting less than a full rotation. Others go on and on, recognizable from their propagation effects for a year or more. Recurring phenomena are more apparent in the low-activity years of the solar cycle, most of them being far enough apart to be clearly identifiable. In April and May, 1976, for example, there were three well-separated areas affecting radio propagation. All were of "the old cycle." There were also three new-cycle areas seen briefly, but with no recognizable radio-propagation influence. The WWV propagation bulletins described will be seen to show recurring effects, if their content is charted for extended periods.

WWV Propagation Bulletins

Since the fall of 1974, WWV and WWVH have transmitted fairly detailed information hourly on the condition of the sun and the earth's magnetic field, and the radio spectrum. At 14 minutes after

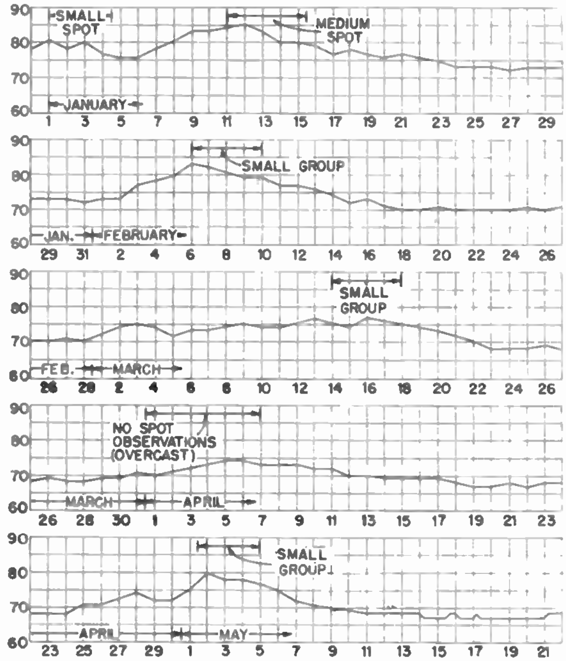


Fig. 19-5 — Graphs of 2800-MHz solar-flux information, as transmitted by WWV, for five consecutive four-week periods in early 1975. It will be seen that, even in this period of relatively low solar activity, the flux readings rise and fall with the passage of small spots. The rise in muf is mainly in the first half of the spot or group's passage across the solar disk. Observation in this period was with the system shown in Fig. 19-2.

the hour (WWV only), and changed four times daily, beginning at 0100 Universal Time, there is a series of statements. The first is a propagation quality forecast, "useless" to "excellent," in nine steps. Next is the condition of the earth's magnetic field, "quiet," "unsettled," or "disturbed." Then comes a letter-number coded forecast, W (disturbed), U (unsettled) or N (normal). The number reflects the 9-step wording given earlier. "Fair-to-good, quiet, November (for N) six" is commonly heard when conditions are above average. The forecast is for the North Atlantic path, but use of the information for other areas can be learned by experience.

The second section of the bulletin is the "K-index," given for the time of bulletin issuance, but relating to the hours just before then. It is a numerical figure for disturbance of the geomagnetic field, 0 to 9 in order of increasing severity. Essentially a current figure, and given with an expected trend, it is potentially quite valuable for short-term forecasting of hf (and occasionally vhf) propagation. A K-index of 0 or 1 indicates very low absorption in the entire hf range, and generally good *F*-layer propagation, up to the muf. Rising K-indices mean increasing absorption, affecting the lower bands first, and more severely on paths involving high latitudes. Up to 3 may show little effect on 21 or 28 MHz, or on low-latitude or transequatorial circuits. A K-index of 4 and expected to rise warns of probable more severe and general disturbance. Severe disturbance is associated with 6 or higher, and auroral propagation is likely on 28 MHz and higher frequencies above about latitude 40; perhaps farther south, east of the Mississippi.

Extremes of the numerical range provided are seldom used in the bulletins. A "W3" forecast is likely to be associated with severely disturbed conditions. "N7" is the usual limit on the good side. The bulletins' semantic nuances are not unlike those heard in weather forecasting, and for the same basic reason — the unpredictability of the sun at current levels of knowledge.

The final bulletin item is the solar flux, a reading taken on 2800 MHz at 1700 UTC in Ottawa. It is also given with an expected trend. Usually the solar-flux value is changed only with the 1900 bulletin, but the trend information may be changed at other times if marked variations are apparent. Especially in the transmissions immediately after the daily change, the solar flux may be the most revealing item of all, for reasons given in more detail later.

A second bulletin at 18 minutes after the hour (46 after on WWVH) is a brief statistical review of the previous day and a prediction for today. First given is the solar flux, the 1700 reading for the previous day (see above). Then comes the A-index, a 24-hour figure for geomagnetic disturbance reflecting the short-term K-index figures of the previous day. It is stated in a different way to give a more linear scale of geomagnetic variation, as a whole-day index for statistical purposes. When the A-index variation is plotted on a long-term basis, its significance as a recurrence warning becomes apparent. The "yesterday" portion of the bulletin concludes with the level of solar activity and the

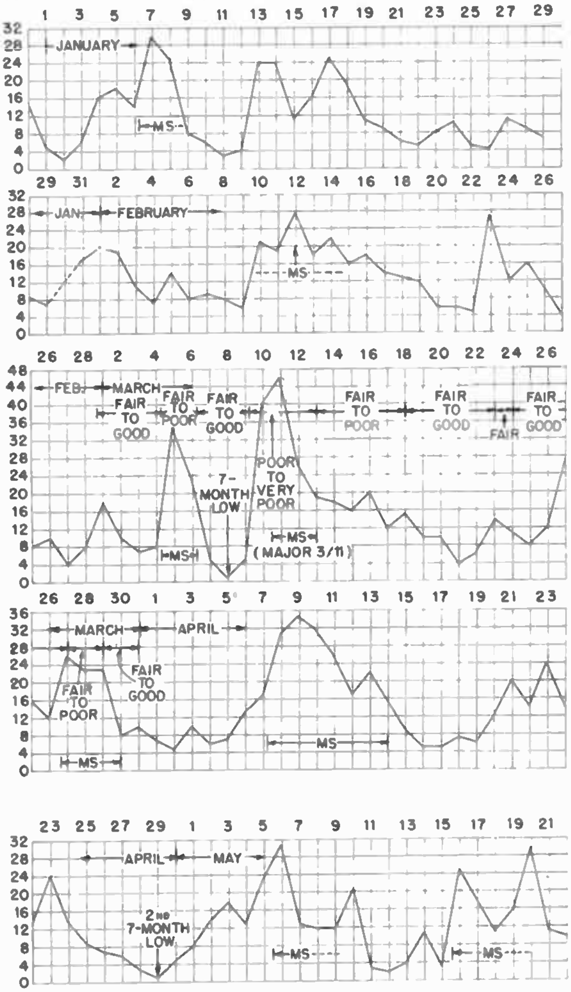


Fig. 19-6 — A-index information as transmitted by WWV, for the same four-week periods as Fig. 19-5. Propagation conditions as to level of absorption are indicated for the month of March, normally the most disturbed period of the year. Periods of geomagnetic storms are indicated by MS. It will be seen that disturbed periods show very marked 27-day recurrence effects, though the severity of disturbance is not necessarily consistent.

condition of the geomagnetic field.

The "today" portion is usually the expected levels of solar and geomagnetic activity. At times of exceptional conditions, present or expected, this final section may be modified to present information on many factors of interest to the serious observer.

Keeping Records

In a group project, or for the individual observer who has time for it, charting all the WWV information is both informative and rewarding. For a less time-consuming effort, concentrate on the 1900 bulletin (usually started at 1914 UTC but always given hourly thereafter, through 0014) and one running of the 18-after information. This will require only about 5 minutes of listening time, and the interval between the bulletins can be used for a quick propagation check, with practice.

Depending on the time available, charts of the WWV information can take many forms. Figures 19-5 and 19-6 show the solar flux and A-index plotted in four-week periods vertically to show recurrence effects clearly. The author keeps two charts. One on a monthly basis has the solar flux in the upper portion, the K-index (one square per unit) at the bottom, and the A-index (two squares per unit) superimposed on the K data. The K information is useful mainly if all bulletins are to be copied. It will then be important for short-term forecasting. The entire content of both WWV bulletins is put on this chart in a shorthand form developed through experience. Brief propagation notes and drawings of the sun are included, where significant. The previous month's solar-flux and A-index curves are inked in with a light or broken line, dates displaced to line up recurrences vertically.

The second chart on the same scale has only the solar flux and the A-index, kept with three consecutive four-week periods superimposed. For legibility these are in different colors. The longer term provides a good check on recurrences and shows clearly that there are occasional "surprises" that do not fit recurring dates. Some events disappear as activity centers die out. Others

Projection viewing of the sun's image with a 5-inch reflector-type telescope. White-paper viewing surface is cemented in the bottom of a black-sprayed cardboard box.

blossom "out of nowhere," at times, as new areas develop. Prediction of these seeming anomalies presents a challenge not yet met fully by anyone, including professionals. It is a wide-open field.

Solar Observation

Regular viewing of the sun should be a part of any major propagation-prediction effort. Even simple projection with a low-cost telescope, as shown in Fig. 19-2, is well worthwhile, if it is done consistently and if drawings are made. Improvements in technique need not be costly. A desirable first step is a light exclusion box that can cost practically nothing. A corrugated-paper box about 6 inches square (or round) and 12-inches deep is fitted with a cover of the same material, about 12 inches square. A hand hole is cut in the side of the box. The cover and interior are sprayed dull black, and a viewing surface of white paper is cemented inside the bottom.

The box is used in the same way as the shaded card in Fig. 19-2, but reduction in ambient light in the viewing area helpsgreatlyin making small detail more noticeable. It enables those with good eyesight to see spots and light variations on the solar surface that would have been missed before. For a complete black-box viewing system that can be assembled from simple optics and plywood, see reference 2.

The telescope used in the photograph, Fig. 19-2, was a low-priced zoom model, with a range of 8 to 25 power. Target scopes and the like in this general range work well, especially with light exclusion around the viewing surface. But many radio amateurs are also astronomers, and may have much better instruments that can be used for solar projection. Moderately priced, 2- or 3-inch refractors give beautiful projection detail with light-exclusion viewing.

Reducing ambient light allows use of a larger image, but this intensifies mechanical-stability problems with camera tripods and other portable supports. The need for sun tracking also increases. An equatorial mount and clock drive used in astronomical work take care of tracking, but only the best tripods give adequate stability.

Better definition with a given degree of optical quality is obtainable with direct viewing of the sun, but this requires a *safe solar filter* (neutral density 4 or more and designed for sun viewing) mounted over the scope aperture. Do not use eye-piece filters. Be sure the aperture filter is mounted so that it cannot come off accidentally. *Never* look at the sun without it.

Visual acuity is very important. Even people who think they have satisfactory vision with properly fitted glasses may find that keener eyes will see much fine detail and light gradation that they miss. Have a younger helper, if you are middle-aged or older.

Interpreting What You See

In viewing the sun with a celestial telescope equipped with a star diagonal and a vertical eyepiece, one sees the solar disk with the east limb on the right and the west limb on the left. This is the opposite of the view obtained with the setup of Fig. 19-2, but is more natural since it simulates a map. Visible solar activity moves across the disk from right to left, on a line parallel with the solar equator. The apparent position of the equator varies with the time of day and the position of the viewer, but it can be determined readily if drawings are made during each observation. Knowing the position of the equator is important in identifying activity as belonging to the old or new cycle, in times of transition. Old-cycle spots move near the equator. New-cycle activity appears some 30° above or below.

In good projection, or with properly safeguarded direct viewing, bright patches may be seen, especially near the east or west limbs. Known as plage, faculae, or flocculi, these patches identify active areas that may or may not include visible spots. When seen on the east limb, they may be advance notice of spots due in another day or two. They serve as warning of propagation changes several days away, and their appearance may coincide with the start of a steady rise in solar flux and in the muf as well. Faculae may identify new activity in which spots will appear four weeks later, or they may be the residue of declining activity that contained spots last time around. They can be a vital part of visual records, and their significance will increase as records accumulate.

In their first or last day on the east or west limb, respectively, sizeable spots or groups usually show as fine lines on or close to the edge of the image. Some detail will begin to show on the second day of new or recurring activity, and sketches should be made as accurately as possible. Note any changes in additional sketches, marked with date and time. Changes in appearance and growth or decay are significant indicators, becoming more so on consecutive rounds of long-lived activity centers.

Increasing size and number of spots will be reflected in a rise in solar flux on the WWV bulletins, particularly the one for 1900 UTC, and in rising *F*-layer muf. Sudden large growth, or a

Direct viewing of the sun should be done only with a telescope equipped with an aperture filter known to be safe for this purpose. Telescope is a Celestron 5 with the maker's solar filter, which passes 0.01 percent of impinging light. A brimmed hat shading the observer's eyes from direct rays of the sun helps to improve visual acuity.

major breakup of a large spot or group, may show radio effects at once – a rise in muf and perhaps a considerable increase in noise level. The latter is more obvious when using a directive array that can be aimed at the sun.

The noise burst and visible change will almost certainly be accompanied by particle radiation increase, the radio effects of which will be increased absorption of hf signal energy, and possibly auroral conditions on 28 MHz or the vhf bands, one to four days later. (Rising K-index on WWV, possibly without warning on previous bulletins.)

Slower growth, barely distinguishable from day to day, will be accompanied by rising solar-flux numbers, probably a point or two daily, and a gradual improvement in hf conditions that will last as long as the K-index remains low. A rise in muf will be apparent at such times, and propagation will remain good on all frequencies for several days, barring sudden solar change which is always a possibility.

If, on the first attempt at solar viewing, one sees sizeable spots or groups, it is well to remember that these may represent activity in a declining phase. If so, they may move across the disk with only minor apparent change. Keep watch though; the area could be brought back to active state again by forces not yet fully understood. This is why long-term predictions are doomed to occasional abject failure and why short-term prediction, using all the tools available, is such an exciting and useful pursuit.

Solar Flux vs. Sunspot Number

The question is often raised as to why we have both the 2800-MHz Solar-Flux Index and the Zurich Sunspot Number as indicators of solar activity. The answer lies in history and its continuity with the future.

Awareness of the infinite variation of the sun began almost with the invention of the telescope, but study of it was largely surreptitious and sporadic at first. There was a strong religious attachment to the sun and the heavens in Galileo's time, and a questioning attitude was actively discouraged. In addition, there is good evidence that the sun entered a period of some 70 years of relative calm in the middle 1600s. Though sunspots were seen frequently by Galileo and other early users of his invention, variations in the solar surface became so rare after about 1640 that new spots or groups were hailed as scientific curiosities.[9]

Record-keeping of sorts began about 100 years later, but up to the middle of the 18th century the accuracy of the surviving records is not high. Even thereafter, observing was done with many levels of sophistication, and it still is today. Thus, any long-term statistics must take the equipment used, the natural conditions of the site, and the skill and diligence of the observer into account. A Swiss astronomer devised a simple formula for what is called the "Wolf Number" in his honor:

$$R = K(10g + f)$$

where R is the sunspot number, K an observer rating factor, g the number of groups visible, and f the total number of spots seen, regardless of size. That the Wolf Number is still in use, with all its ambiguity, is recognition of its worth as a statistical link with history, but it has little in its favor otherwise. Today's radio amateur need only run off a few Wolf Numbers based on his daily observation of the sun and the concurrent behavior of the radio spectrum to realize that "sunspot number" has little relevance to the radio communications scene in this half of the 20th century.

Long-term statistics inevitably involve some "smoothing" of original data. Even the 2800-MHz solar flux given hourly on WWV is a smoothed value, but it tracks much better with observed variations in radio conditions and the appearance of the sun than any other available statistic.

The individual observer of the sun can establish his own Wolf-equation K factor by working out examples of his observed g and f, and finding the average for K that tracks well with the information given in Fig. 19-1. This is an interesting exercise, but it will show that, for our purposes, the solar flux is the only muf indicator we really need.

Monitoring Solar Noise

Radiation from the sun is recorded on many frequencies at many sites all over the world.

Combined charts of the results for the same time and date show marked differences from one frequency to another in the level of noise from the sun. Great equipment sophistication is not required for monitoring solar noise, and the curious amateur may find it an interesting project. In any location where man-made noise is low, a reasonably good receiver and a directional antenna that can be aimed at the sun will yield some solar noise. In fact, the "quiet sun" is a good noise generator with which to rate antenna and receiver system performance on most frequencies where directive antennas are used in amateur communication.

Increasing interest in amateur moonbounce and satellite communication have made antenna control in both azimuth and elevation much more common than in the past and have increased our sun-noise consciousness accordingly. Recording changes in received sun noise could add a useful dimension to any propagation-observation program: Hf, vhf, or uhf. The idea is not exactly new. See reference 3.

The most interesting projects in amateur radio tend to be those where we can start at the beginner level and get interesting results at moderate cost, then work up by stages, making worthwhile improvements for as long as we want to carry on the effort. Monitoring of the sun and the observation and recording of radio propagation variations, with a view to making better use of our on-the-air time, make a game of great fascination that can be played at ever-increasing levels of sophistication. You may never get enough of it.

Propagation References

[1] Tilton, "The DXer's Crystal Ball," *QST*, June, August, and September, 1975.

[2] Projection of the sun and interpretation of results are discussed in reference 1, and in *QST* for December, 1974, p. 83 and January, 1975, p. 84. A black-box viewing device (Tomcik, K4HYF) for sun projection is shown in July, 1964, *QST*. (Photocopy from ARRL, 75 cents and stamped envelope.)

[3] Bray and Kirchner, "Antenna Patterns from the Sun," *QST*, July, 1960. Wilson, "432-MHz Solar Patrol," *QST*, August, 1967.

[4] Davies, "Ionospheric Radio Propagation," NBS Monograph 80, out of print. Available in some technical libraries.

[5] See reference 4, p. 45.

[6] Bain, "VHF Propagation by Meteor Trail Ionization," *QST*, May, 1974. Table of major meteor showers, "Radio Amateurs VHF Manual," Ch. 2.

[7] Hall, "High-Frequency Propagation Estimations for the Radio Amateur," *QST*, March, 1972.

[8] *World Radio and TV Handbook*, Billboard Publications, London W1V1PG; 2160 Patterson Street, Cincinnati, Ohio 45124. Gilfer Associates, Box 239, Park Ridge, NJ 07656.

[9] Eddy, "The Maunder Minimum," digest of longer paper, *QST* for July, 1976.

WWV PROPAGATION BULLETIN SERVICE LOST

With portions of this *Handbook* already printing, changes in the WWV format were made, deleting the propagation bulletins formerly broadcast at 18 minutes after the hour. Termination of the 14-after information was also under consideration, though there was some prospect that this service might be saved. For the latest available information, see the propagation information that runs regularly in *QST*. Changes of any major nature will also be announced via W1AW bulletin.

Transmission Lines

Transmission lines, and the theory behind them, play an important role in many phases of radio communication. This is because the basic principles involved can be applied to a wide variety of problems. Types of transmission lines include simple two-conductor configurations such as the familiar coaxial cable and TV parallel-wire line. Such lines are useful from power frequencies to well up into the microwave region and form perhaps the most important class. The waveguide is representative of a second type. Here, the conductor configuration is rather complex and ordinary concepts such as voltage and current tend to become obscure. As a consequence, various parameters are expressed in terms of the electric and magnetic fields associated with the line. Finally, the propagation of electromagnetic energy through space itself is closely related to similar phenomena in wave guides and transmission lines. In fact, the only significant physical difference is that the power density in a wave propagated in space decreases with increasing distance while it is possible to transmit power over long distances with conventional lines with little attenuation. This is because power flow is essentially confined to one dimension in the latter case while the three-dimensional aspect of space does not permit such confinement.

Transmission Lines and Circuits

A transmission line differs from an ordinary circuit in one very important aspect. Delay effects associated with the finite propagation time of electromagnetic energy are often neglected in network design since the dimensions involved are normally small compared to the wavelength of any frequencies present in the circuit. This is not true in transmission-line considerations. The finite propagation time becomes a factor of paramount importance. This can be illustrated with the aid of Fig. 1. A transmission line separates a source at point g from a load at point a by a distance l. If the line is uniform (same conductor shape at any cross section along the line), only two parameters are required to express the line properties completely. These are the phase velocity, v_p, and the characteristic impedance, Z_o. If the line can be considered lossless as well, Z_o becomes a pure resistance, R_o.

Assume that a very short burst of power is emitted from the source. This is represented by the vertical line at the left of the series of lines in Fig. 2. As the pulse voltage appears across the load Z_a, all the energy may be absorbed or part of it may be reflected in much the same manner energy in a wave in water is reflected as the wave hits a steep breakwater or the end of a container. This reflected wave is represented by the second line in the series and the arrow above indicates the direction of travel. As the latter wave reaches the source, the process is again repeated with either all of the energy being absorbed or partially reflected.

The back-and-forth cycle is actually an infinite one but after a few reflections, the intensity of the wave becomes very small. If, instead of a short pulse, a continuous voltage is applied to the terminals of a transmission line, the voltage at any point along the line will consist of a sum of voltages of the composite of waves traveling toward the right and a composite of waves traveling toward the left. The total sum of the waves traveling toward the right is called the forward wave or incident wave while the one traveling toward the left is called the reflected wave. Provided certain conditions concerning Z_a are met, there will be a net flow of energy from the source to the load with a fraction of the energy being stored in the "standing" waves on the line. This phenomenon is identical to the case of a coupled resonator with ordinary circuit elements and sections of transmission line are often used for this purpose especially in the vhf/uhf region. The duplexer found in many vhf repeaters is a common example.

Line Factors and Equations

Since transmission lines are usually connected between lumped or discrete circuitry, it is con-

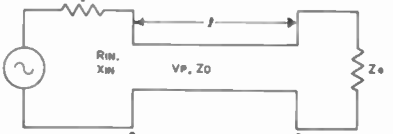


Fig. 1 — Source and load connected by means of a transmission line.

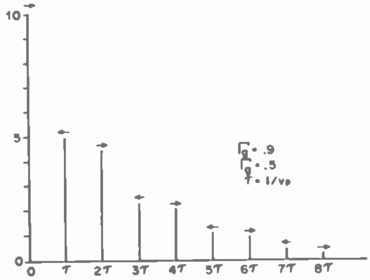


Fig. 2 — Magnitudes of components for forward and reverse traveling waves of a short pulse on a transmission line.

venient to be able to express the input impedance of a line in terms of the output or load impedance. A line treated this way is then similar to a filter or matching network with a given load impedance. One caution should be kept in mind in applying such relations and that is the manner in which the source and load are connected to the line can be important. There are always some "parasitic" effects arising from connectors and post-connector circuit configuration that may cause the line to "see" a different impedance than if measurements were made at the load terminals directly. This is indicated by the abrupt change in line dimensions at points *a* and *g* in Fig. 1. Even though the short line connecting the generator to the main transmission line (and the one connecting the load to the line) might have the same characteristic impedance, if the sizes are different, a mismatch will still occur. Normally, this effect can be neglected at hf but becomes important as the frequency of operation is extended into the vhf region and above.

In referring to the previous example shown in Fig. 2, the ratio of the voltage in the reflected wave to that of the voltage in the incident wave is defined as the voltage reflection coefficient designated by the Greek letter, Γ, or by ρ. The relation between the output resistance, R_a, the output reactance, X_a, the line impedance, Z_o, and the *magnitude* of the reflection coefficient is

$$\Gamma = \sqrt{\frac{(R_a - R_o)^2 + X_a^2}{(R_o + R_a)^2 + X_a^2}}$$

Note that if R_a is equal to R_o, and if x_a is 0 the reflection coefficient is 0 which represents "matched" conditions. All the energy in the incident wave is transferred to the load. In effect, it was as if there were an infinite line of characteristic impedance Z_o connected at *a*. On the other hand, if R_a is 0, regardless of the value of x_a the reflection coefficient is 1.0. This means all the power is reflected in much the same manner as radiant energy is reflected from a mirror.

If there are no reflections from the load, the voltage distribution along the line is constant or "flat" while if reflections exist, a standing-wave pattern will result. The ratio of the maximum voltage on the line to the minimum value (provided the line is longer than a quarter wavelength) is defined as the voltage standing-wave ratio (*VSWR*). The *VSWR* is related to the reflection coefficient by

$$VSWR = \frac{1 + \Gamma}{1 - \Gamma}$$

and this latter definition is a more general one valid for any line length. Quite often, the actual load impedance is unknown and an alternate way of expressing the reflection coefficient is

$$\Gamma = \sqrt{\frac{P_r}{P_f}}$$

where Pr is the power in the reflected wave and Pf is the power in the forward wave. The parameters are relatively easy to measure with power meters available commercially or with homemade designs. However, it is obvious there can be no other power sources at the load if the foregoing definition is to hold. For instance, the reflection coefficient of the generator in the example shown in Fig. 2 is 0.9. This value could have been obtained by substituting the generator resistance and reactance into a previous formula for reflection coefficient, but not by measurement if the source were activated.

Fortunately, it is possible to determine the input resistance and reactance of a terminated line if the load resistance and reactance are known, along with the line length and characteristic impedance. (With actual lines, the physical length must be divided by the velocity factor of the cable which gives the value of l in the following formula.) The equations are:

$$r_{in} = \frac{r_a(1 + \tan^2 \beta l)}{(1 - x_a \tan \beta l)^2 + (r_a \tan \beta l)^2}$$

$$x_{in} = \frac{x_a(1 - \tan^2 \beta l) + (1 - r_a{}^2 - x_a{}^2) \tan \beta l}{(1 - x_a \tan \beta l)^2 + (r_a \tan \beta l)^2}$$

for a 1-Ω line. Equations are often "normalized" this way in order to make universal tables or plots that cover a wide range of values. If characteristic impedances (Z_o) other than 1.0 are to be used, the following set of conversions apply where R_a and X_a are the load resistance and reactance and R_{in} and X_{in} represent the resistance and reactance at the input end of the line.

$$r_a = \frac{R_a}{Z_o}\ , \qquad R_{in} = Z_o r_{in}$$

$$x_a = \frac{X_a}{Z_o}\ , \qquad X_{in} = Z_o x_{in}$$

In order to determine the value of the tangent function, either the line length in meters or feet,

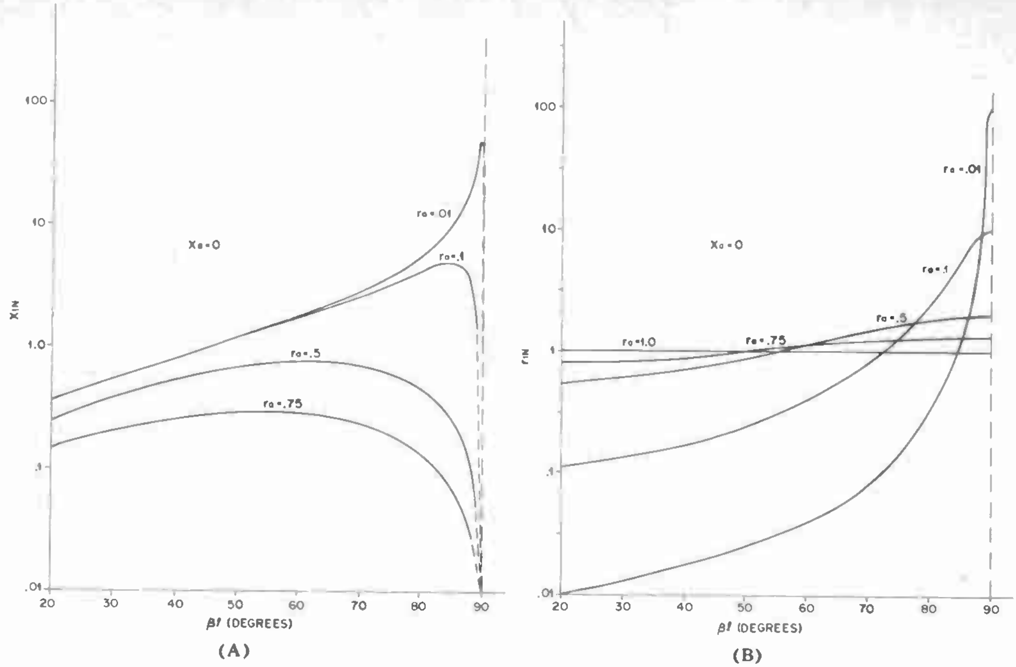


Fig. 3 — Normalized input reactance and resistance *vs.* line length for various values of r_a (x_a equal to 0).

along with the frequency in MHz can be substituted into the following expressions:

$$\beta l\,(\text{degrees}) = 1.2 f_{(\text{MHz})} \times l_{(\text{meters})}$$

$$\beta l\,(\text{degrees}) = 0.367 f_{(\text{MHz})} \times l_{(\text{feet})}$$

Since the foregoing transmission-line equations are somewhat awkward to work with, various plots have been devised that permit a graphical solution. However, with modern programmable calculators, even those in a moderate price class, it takes approximately 4 seconds to solve both equations. The plots shown in Fig. 3A and Fig. 3B were computed in this manner. The curves are for r_{in} and x_{in} for various values of r_a (x_a equal to 0) and line length in degrees. Note that 90 degrees appears to be a "critical" value and represents a line length of a quarter wavelength. As this value is approached, the transmission-line equations can be approximated by the formulas:

$$r_{in} \cong \frac{r_a}{r_a^2 + x_a^2}$$

$$x_{in} \cong \frac{-x_a}{r_a^2 + x_a^2}$$

If x_a is zero, the formula for a quarter-wavelength transformer is obtained which is

$$R_{in} = Z_o^2/R_a$$

Quite often, it is mistakenly assumed that power reflected from a load represents power "lost" in some way. This is only true if there is considerable loss in the line itself and the power is dissipated on the way back to the source. On the other hand, the quarter-wavelength transformer is an example where reflections on a lossless line can actually be used to advantage in matching a load impedance that is different from the source impedance.

If the terminating resistance is zero, the input resistance is also zero. In effect, the line and load act as a pure reactance which is given by the formula:

$$x_{in} = \frac{x_a + \tan\beta l}{1 - x_a \tan\beta l}$$

The special cases where the terminating reactance is either zero or infinity are given by the respective formulas:

$$x_{in} = \tan\beta l \quad , \qquad x_{in} = -\cot\beta l$$

A short length of line with a short circuit as a terminating load appears as an inductor while an open-circuited line appears as a capacitance.

MATCHING THE ANTENNA TO THE LINE

The load for a transmission line may be any device capable of dissipating rf power. When lines are used for transmitting applications the most common type of load is an antenna. When a transmission line is connected between an antenna and a receiver, the receiver input circuit (not the antenna) is the load, because the power taken from a passing wave is delivered to the receiver.

Whatever the application, the conditions existing at the load, and *only* the load, determine the standing-wave ratio on the line. If the load is purely resistive and equal in value to the characteristic impedance of the line, there will be no standing waves. In case the load is not purely resistive, and/or is not equal to the line Z_0, there will be standing waves. No adjustments that can be made at the input end of the line can change the SWR, nor is it affected by changing the line length.

Only in a few special cases is the load inherently of the proper value to match a practicable transmission line. In all other cases it is necessary either to operate with a mismatch and accept the SWR that results, or else to take steps to bring about a proper match between the line and load by means of transformers or similar devices. Impedance-matching transformers may take a variety of physical forms, depending on the circumstances.

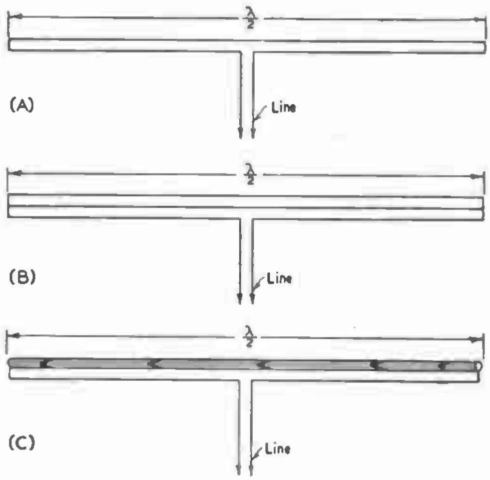


Fig. 20-10B — The folded dipole, a method for using the antenna element itself to provide an impedance transformation.

Note that it is essential, if the SWR is to be made as low as possible, that the load at the point of connection to the transmission line be purely resistive. In general, this requires that the load be tuned to resonance. If the load itself is not resonant at the operating frequency the tuning sometimes can be accomplished in the matching system.

THE ANTENNA AS A LOAD

Every antenna system, no matter what its physical form, will have a definite value of impedance at the point where the line is to be connected. The problem is to transform this antenna input impedance to the proper value to match the line. In this respect there is no one "best" type of line for a particular antenna system, because it is possible to transform impedances in any desired ratio. Consequently, any type of line may be used with any type of antenna. There are frequently reasons other than impedance matching that dictate the use of one type of line in preference to another, such as ease of installation, inherent loss in the line, and so on, but these are not considered in this section.

Although the input impedance of an antenna system is seldom known very accurately, it is often possible to make a reasonably close estimate of its value.

Matching circuits can be built using ordinary coils and capacitors, but are not used very extensively because they must be supported at the antenna and must be weatherproofed. The systems to be described use linear transformers.

The Quarter-Wave Transformer or "Q" Section

As mentioned previously, a quarter-wave transmission line may be used as an impedance transformer. Knowing the antenna impedance and the characteristic impedance of the transmission line to be matched, the required characteristic impedance of a matching section such as is shown in Fig. 20-10A is:

$$Z = \sqrt{Z1\,ZO} \qquad (20\text{-H})$$

Where Z1 is the antenna impedance and ZO is the characteristic impedance of the line to which it is to be matched.

Example: To match a 600-ohm line to an antenna presenting a 72-ohm load, the quarter-wave matching section would require a characteristic impedance of

$$\sqrt{72 \times 600} = \sqrt{43{,}200} = 208 \text{ ohms}$$

The spacings between conductors of various sizes of tubing and wire for different surge impedances

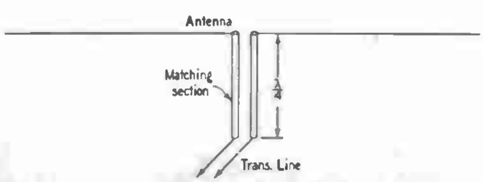


Fig. 20-10A — "Q" matching section, a quarter-wave impedance transformer.

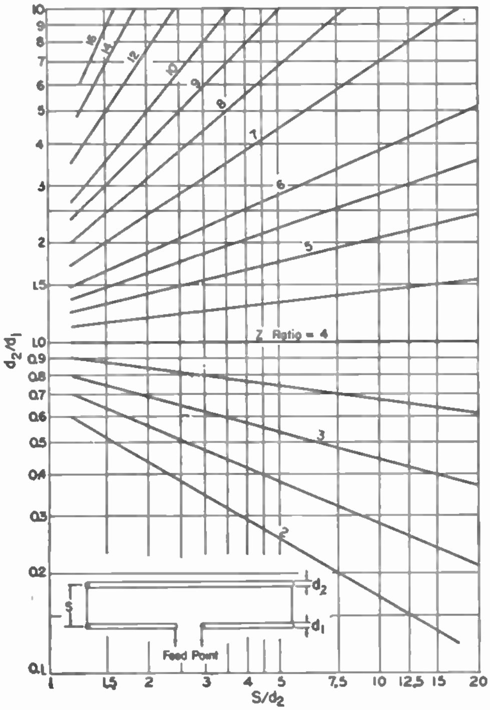


Fig. 20-11 – Impedance transformation ratio, two-conductor folded dipole. The dimensions d1, d2 and s are shown on the inset drawing. Curves show the ratio of the impedance (resistive) seen by the transmission line to the radiation resistance of the resonant antenna system.

are given in graphical form in the chapter on "Transmission Lines." (With 1/2-inch tubing, the spacing in the example above should be 1.5 inches for an impedance of 208 ohms.)

The length of the quarter-wave matching section may be calculated from

$$Length\ (\text{feet}) = \frac{246\,V}{f} \qquad (20\text{-}I)$$

where V = Velocity factor
f = Frequency in MHz

Example: A quarter-wave transformer of RG-11/U is to be used at 28.7 MHz. From the table 20-1, $V = 0.66$.

$$Length = \frac{246 \times 0.66}{28.7} = 5.65 \text{ feet}$$

$= 5$ feet 8 inches

The antenna must be resonant at the operating frequency. Setting the antenna length by formula is amply accurate with single-wire antennas, but in other systems, particularly close-spaced arrays, the antenna should be adjusted to resonance before the matching section is connected.

When the antenna input impedance is not known accurately, it is advisable to construct the matching section so that the spacing between conductors can be changed. The spacing then may be adjusted to give the lowest possible SWR on the transmission line.

Folded Dipoles

A half-wave antenna element can be made to match various line impedances if it is split into two or more parallel conductors with the transmission line attached at the center of only one of them. Various forms of such "folded dipoles" are shown in Fig. 20-10B. Currents in all conductors are in phase in a folded dipole, and since the conductor spacing is small the folded dipole is equivalent in radiating properties to an ordinary single-conductor dipole. However, the current flowing into the input terminals of the antenna from the line is the current in one conductor only, and the entire power from the line is delivered at this value of current. This is equivalent to saying that the input impedance of the antenna has been raised by splitting it up into two or more conductors.

The ratio by which the input impedance of the antenna is stepped up depends not only on the number of conductors in the folded dipole but also on their relative diameters, since the distribution of current between conductors is a function of their diameters. (When one conductor is larger than the other, as in Fig. 20-10B, the larger one carries the greater current.) The ratio also depends, in general, on the spacing between the conductors, as shown by the graphs of Figs. 20-11 and 20-12. An important special case is the 2-conductor dipole with conductors of equal diameter; as a simple antenna, not a part of a directive array, it has an

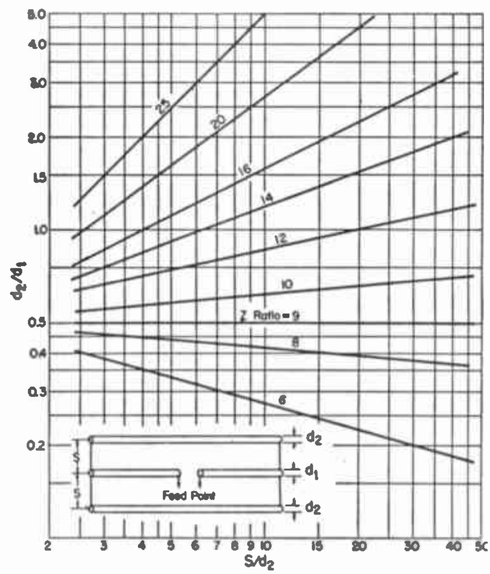


Fig. 20-12 – Impedance transformation ratio, three-conductor folded dipole. The dimensions d1, d2 and s are shown on the inset drawing. Curves show the ratio of the impedance (resistive) seen by the transmission line to the radiation resistance of the resonant antenna system.

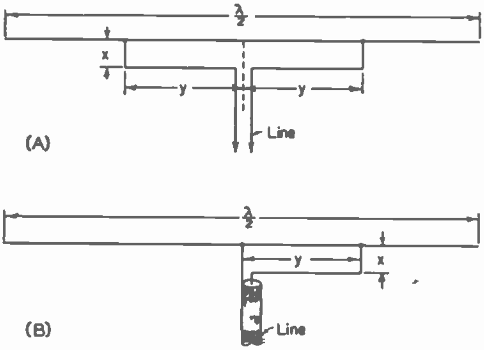


Fig. 20-13 – The "T" match and "gamma" match.

input impedance close enough to 300 ohms to afford a good match to 300-ohm Twin-Lead.

The required ratio of conductor diameters to give a desired impedance ratio using two conductors may be obtained from Fig. 20-11. Similar information for a 3-conductor dipole is given in Fig. 20-12. This graph applies where all three conductors are in the same plane. The two conductors not connected to the transmission line must be equally spaced from the fed conductor, and must have equal diameters. The fed conductor may have a different diameter, however. The unequal-conductor method has been found particularly useful in matching to low-impedance antennas such as directive arrays using close-spaced parasitic elements.

The length of the antenna element should be such as to be approximately self-resonant at the median operating frequency. The length is usually not highly critical, because a folded dipole tends to have the characteristics of a "thick" antenna and thus has a relatively broad frequency-response curve.

"T" and "Gamma" Matching Sections

The method of matching shown in Fig. 20-13A is based on the fact that the impedance between any two points along a resonant antenna is resistive, and has a value which depends on the spacing between the two points. It is therefore possible to choose a pair of points between which the impedance will have the right value to match a transmission line. In practice, the line cannot be connected directly at these points because the distance between them is much greater than the conductor spacing of a practicable transmission line. The "T" arrangement in Fig. 20-13A overcomes this difficulty by using a second conductor paralleling the antenna to form a matching section to which the line may be connected.

The "T" is particularly suited to use with a parallel-conductor line, in which case the two points along the antenna should be equidistant from the center so that electrical balance is maintained.

The operation of this system is somewhat complex. Each "T" conductor (*y* in the drawing) forms with the antenna conductor opposite it a short section of transmission line. Each of these transmission-line sections can be considered to be terminated in the impedance that exists at the point of connection to the antenna. Thus the part of the antenna between the two points carries a transmission-line current in addition to the normal antenna current. The two transmission-line matching sections are in series, as seen by the main transmission line.

If the antenna by itself is resonant at the operating frequency its impedance will be purely resistive, and in such case the matching-section lines are terminated in a resistive load. However, since these sections are shorter than a quarter wavelength their input impedance – i.e., the impedance seen by the main transmission line looking into the matching-section terminals – will be reactive as well as resistive. This prevents a perfect match to the main transmission line, since its load must be a pure resistance for perfect matching. The reactive component of the input impedance must be tuned out before a proper match can be secured.

One way to do this is to detune the antenna just enough, by changing its length, to cause reactance of the opposite kind to be reflected to the input terminals of the matching section, thus cancelling the reactance introduced by the latter. Another method, which is considerably easier to adjust, is to insert a variable capacitor in series with the matching section where it connects to the transmission line, as shown in Fig. 21-39. The capacitor must be protected from the weather.

The method of adjustment commonly used is to cut the antenna for approximate resonance and then make the spacing *x* some value that is convenient constructionally. The distance *y* is then adjusted, while maintaining symmetry with respect to the center, until the SWR on the transmission line is as low as possible. If the SWR is not below 2 to 1 after this adjusment, the antenna length should be changed slightly and the matching section taps adjusted again. This procedure may be continued until the SWR is as close to 1 to 1 as possible.

When the series-capacitor method of reactance compensation is used (Fig. 21-32), the antenna should be the proper length to be resonant at the operating frequency. Trial positions of the matching-section taps are then taken, each time adjusting the capacitor for minimum SWR, until the standing waves on the transmission line are brought down to the lowest possible value.

The unbalanced ("gamma") arrangement in Fig. 20-13B is similar in principle to the "T," but is adapted for use with single coax line. The method of adjustment is the same.

BALANCING DEVICES

An antenna with open ends, of which the half-wave type is an example, is inherently a balanced radiator. When opened at the center and fed with a parallel-conductor line this balance is maintained throughout the system, so long as the causes of unbalance discussed in the transmission-line chapter are avoided.

If the antenna is fed at the center through a coaxial line, as indicated in Fig. 20-14A, this balance is upset because one side of the radiator is connected to the shield while the other is connected to the inner conductor. On the side connected to the shield, a current can flow down over the *outside* of the coaxial line, and the fields thus set up cannot be canceled by the fields from the inner conductor because the fields *inside* the line cannot escape through the shielding afforded by the outer conductor. Hence these "antenna" currents flowing on the outside of the line will be responsible for radiation.

Linear Baluns

Line radiation can be prevented by a number of devices whose purpose is to detune or decouple the line for "antenna" currents and thus greatly reduce their amplitude. Such devices generally are known as baluns (a contraction for "balanced to unbalanced"). Fig. 20-14B shows one such arrangement, known as a bazooka, which uses a sleeve over the transmission line to form, with the outside of the outer line conductor, a shorted quarter-wave line section. As described earlier in this chapter, the impedance looking into the open end of such a section is very high, so that the end of the outer conductor of the coaxial line is effectively insulated from the part of the line below the sleeve. The length is an *electrical* quarter wave, and may be physically shorter if the insulation between the sleeve and the line is other than air. The bazooka has no effect on the impedance relationships between the antenna and the coaxial line.

Another method that gives an equivalent effect is shown at C. Since the voltages at the antenna terminals are equal and opposite (with reference to ground), equal and opposite currents flow on the surfaces of the line and second conductor. Beyond the shorting point, in the direction of the transmitter, these currents combine to cancel out. The balancing section "looks like" an open circuit to the antenna, since it is a quarter-wave parallel-conductor line shorted at the far end, and thus has no effect on the normal antenna operation. However, this is not essential to the line-balancing function of the device, and baluns of this type are sometimes made shorter than a quarter wavelength in order to provide the shunt inductive reactance required in certain types of matching systems.

Fig. 20-14D shows a third balun, in which equal and opposite voltages, balanced to ground, are taken from the inner conductors of the main transmission line and half-wave phasing section. Since the voltages at the balanced end are in series while the voltages at the unbalanced end are in parallel, there is a 4-to-1 step-down in impedance from the balanced to the unbalanced side. This arrangement is useful for coupling between a balanced 300-ohm line and a 75-ohm coaxial line, for example.

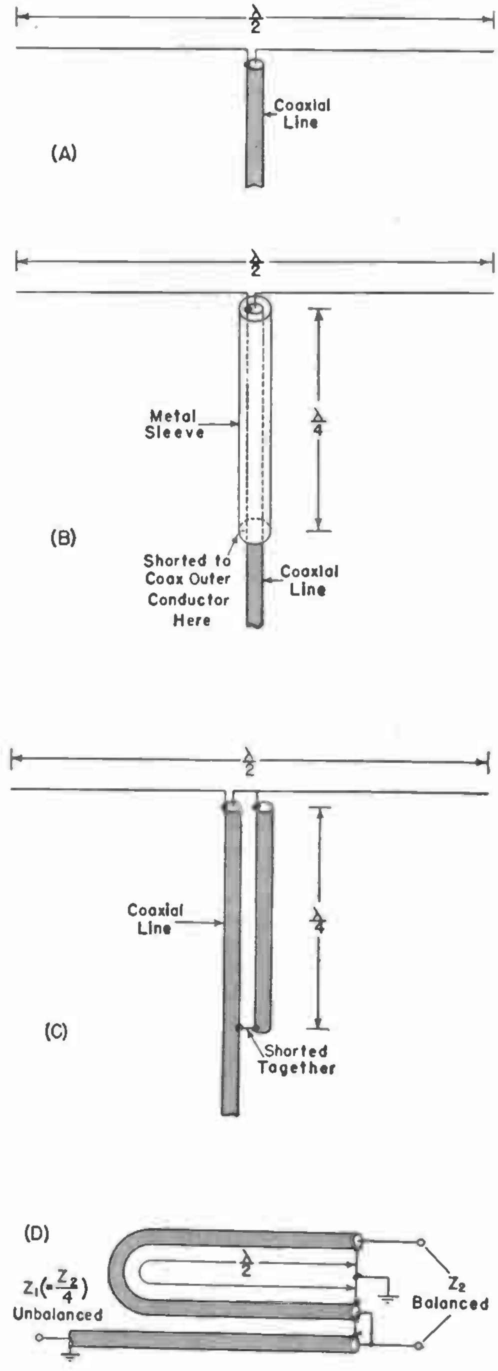


Fig. 20-14 – Radiator with coaxial feed (A) and methods of preventing unbalanced currents from flowing on the outside of the transmission line (B and C). The half-wave phasing section shown at D is used for coupling between an unbalanced and a balanced circuit when a 4-to-1 impedance ratio is desired or can be accepted.

OTHER LOADS AND BALANCING DEVICES

The most important practical load for a transmission line is an antenna which, in most cases, will be "balanced" – that is, symmetrically constructed with respect to the feed point. Aside from considerations of matching the actual impedance of the antenna at the feed point to the characteristic impedance of the line (if such matching is attempted) a balanced antenna should be fed through a balanced transmission line in order to preserve symmetry with respect to ground and thus avoid difficulties with unbalanced currents on the line and consequent undesirable radiation from the transmission line itself.

If, as is often the case, the antenna is to be fed through coaxial line (which is inherently unbalanced) some method should be used for connecting the line to the antenna without upsetting the symmetry of the antenna itself. This requires a circuit that will isolate the balanced load from the unbalanced line while providing effecient power transfer. Devices for doing this are called baluns. The types used between the antenna and transmission line are generally "linear," consisting of transmission-line sections.

The need for baluns also arises in coupling a transmitter to a balanced transmission line, since the output circuits of most transmitters have one side grounded. (This type of output circuit is desirable for a number of reasons, including TVI reduction.) The most flexible type of balun for this purpose is the inductively coupled matching network described in a subsequent section in this chapter. This combines impedance matching with balanced-to-unbalanced operation, but has the disadvantage that it uses resonant circuits and thus can work over only a limited band of frequencies without readjustment. However, if a fixed impedance ratio in the balun can be tolerated, the coil balun described below can be used without adjustment over a frequency range of about 10 to 1 – 3 to 30 MHz, for example.

Coil Baluns

The type of balun known as the "coil balun" is based on the principles of linear-transmission-line balun as shown in the upper drawing of Fig. 20-15. Two transmission lines of equal length having a characteristic impedance (Z_0) are connected in series at one end and in parallel at the other. At the series-connected end the lines are balanced to ground and will match an impedance equal to $2Z_0$. At the parallel-connected end the lines will be matched by an impedance equal to $Z_0/2$. One side may be connected to ground at the parallel-connected end, provided the two lines have a length such that, considering each line as a single wire, the balanced end is effectively decoupled from the parallel-connected end. This requires a length that is an odd multiple of 1/4 wavelength.

A definite line length is required only for decoupling purposes, and so long as there is adequate decoupling the system will act as a 4-to-1 impedance transformer regardless of line length. If each line is wound into a coil, as in the lower drawing, the inductances so formed will act as choke coils and will tend to isolate the series-connected end from any ground connection that may be placed on the parallel-connected end. Balun coils made in this way will operate over a wide frequency range, since the choke inductance is not critical. The lower frequency limit is where the coils are no longer effective in isolating one end from the other; the length of line in each coil should be about equal to a quarter wave-length at the lowest frequency to be used.

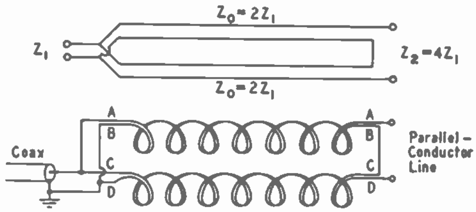


Fig. 20-15 – Baluns for matching between push-pull and single-ended circuits. The impedance ratio is 4 to 1 from the push-pull side to the unbalanced side. Coiling the lines (lower drawing) increases the frequency range over which satisfactory operation is obtained.

The principal application of such coils is in going from a 300-ohm balanced line to a 75-ohm coaxial line. This requires that the Z_0 of the lines forming the coils be 150 ohms.

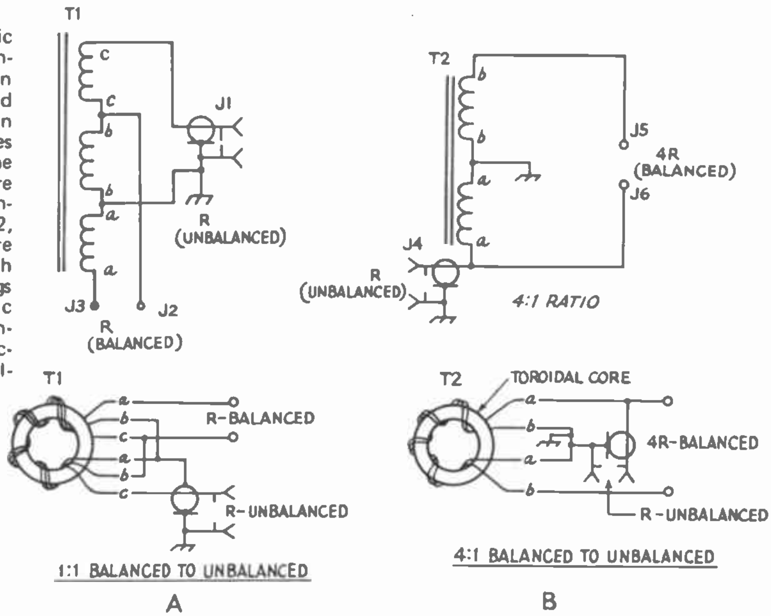
A balun of this type is simply a fixed-ratio transformer, when matched. It cannot compensate for inaccurate matching elsewhere in the system. With a "300-ohm" line on the balanced end, for example, a 75-ohm coax cable will not be matched unless the 300-ohm line actually is terminated in a 300-ohm load.

TWO BROAD-BAND TOROIDAL BALUNS

Air-wound balun transformers are somewhat bulky when designed for operation in the 1.8- to 30-MHz range. A more compact broad-band transformer can be realized by using toroidal ferrite core material as the foundation for bifilar-wound coil balun transformers. Two such baluns are described here.

In Fig. 20-16 at A, a 1:1 ratio balanced-to-unbalanced-line transformer is shown. This transformer is useful in converting a 50-ohm balanced line condition to one that is 50 ohms, unbalanced. Similarly, the transformer will work between balanced and unbalanced 75-ohm impedances. A 4:1 ratio transformer is illustrated in Fig. 20-16 at B. This balun is useful for converting a 200-ohm balanced condition to one that is 50 ohms, unbalanced. In a like manner, the transformer can be used between a balanced 300-ohm point and a 75-ohm unbalanced line. Both balun transformers will handle 1000 watts of rf power and are designed to operate from 1.8 through 60 MHz.

Fig. 20-16 – Schematic and pictorial representations of the balun transformers. T1 and T2 are wound on CF-123 toroid cores (see footnote 1, and the text). J1 and J4 are SO-239-type coax connectors, or similar. J2, J3, J5, and J6 are steatite feedthrough bushings. The windings are labeled a, b, and c to show the relationship between the pictorial and schematic illustrations.



Low-loss high-frequency ferrite core material is used for T1 and T2.[1,3] The cores are made from Q-2 material and cost approximately $5.50 in single-lot quantity. They are 0.5 inches thick, have an OD of 2.4 inches, and the ID is 1.4 inches. The permeability rating of the cores is 40. A packaged one-kilowatt balun kit, with winding instructions for 1:1 or 4:1 impedance transformation ratios, is available, but uses a core of slightly different dimensions.[2]

Winding Information

The transformer shown in Fig. 20-16 at A has a trifilar winding consisting of 10 turns of No. 14 formvar-insulated copper wire. A 10-turn bifilar winding of the same type of wire is used for the balun of Fig. 20-16 at B. If the cores have rough edges, they should be carefully sanded until smooth enough to prevent damage to the wire's formvar insulation. The windings should be spaced around the entire core as shown in Fig. 20-17. Insulation can be used between the core material and the windings to increase the power handling capabilities of the core.

Using the Baluns

For indoor applications, the transformers can be assembled open style, without benefit of a protective enclosure. For outdoor installations, such as at the antenna feed point, the balun should be encapsulated in epoxy resin or mounted in a suitable weather-proof enclosure. A Minibox, sealed against moisture, works nicely for the latter.

NONRADIATING LOADS

Typical examples of nonradiating loads for a transmission line are the grid circuit of a power amplifier (considered in the chapter on transmitters), the input circuit of a receiver, and another transmission line. This last case includes the "antenna tuner" – a misnomer because it is

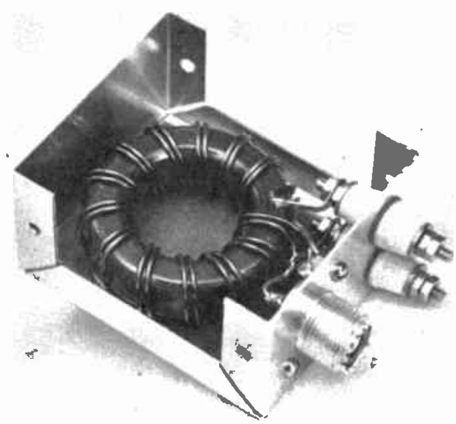


Fig. 20-17 – Layout of a kilowatt 4:1 toroidal balun transformer. Phenolic insulating board is mounted between the transformer and the Minibox wall to prevent short-circuiting. The board is held in place with epoxy cement. Cement is also used to secure the transformer to the board. For outdoor use, the Minibox cover can be installed, then sealed against the weather by applying epoxy cement along the seams of the box.

[1] Available in single-lot quantity from Permag Corp, 88-06 Van Wyck Expy, Jamaica, NY 11418.

[2] Amidon Associates, 12033 Otsego Street, North Hollywood, CA 91601.

[3] Toroid cores are also available from Ferroxcube Corp. of America, Saugerties, NY 12477.

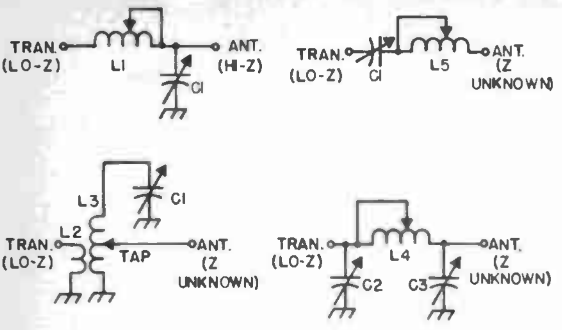


Fig. 20-18 – Networks for matching a low-*Z* transmitter output to random-length end-fed wire antennas.

actually a device for coupling a transmission line to the transmitter. Because of its importance in amateur installations, the antenna coupler is considered separately in a later part of this chapter.

Coupling to a Receiver

A good match between an antenna and its transmission line does not guarantee a low standing-wave ratio on the line when the antenna system is used for receiving. The SWR is determined wholly by what the line "sees" at the receiver's antenna-input terminals. For minimum SWR the receiver input circuit must be matched to the line. The rated input impedance of a receiver is a nominal value that varies over a considerable range with frequency. Most hf receivers are sensitive enough that exact matching is not necessary. The most desirable condition is that in which the receiver is matched to the line Z_0 and the line in turn is matched to the antenna. This transfers maximum power from the antenna to the receiver with the least loss in the transmission line.

COUPLING TO RANDOM-LENGTH ANTENNAS

Several impedance-matching schemes are shown in Fig. 20-18, permitting random-length wires to be matched to normal low-*Z* transmitter outputs. The circuit used will depend upon the length of the antenna wire and its impedance at the desired operating frequency. Ordinarily, one of the four methods shown will provide a suitable impedance match to an end-fed random wire, but the configuration will have to be determined experimentally. For operation between 3.5 and 30 MHz, C1 can be a 200-pF type with suitable plate spacing for the power level in use. C2 and C3 should be 500-pF units to allow for flexibility in matching. L1, L4, and L5 should be tapped or rotary inductors with sufficient *L* for the operating frequency. L3 can be a tapped Miniductor coil with ample turns for the band being used. An SWR bridge should be used as a match indicator.

COUPLING THE TRANSMITTER TO THE LINE

The type of coupling system that will be needed to transfer power adequately from the final rf amplifier to the transmission line depends almost entirely on the input impedance of the line. As shown earlier in this chapter, the input impedance is determined by the standing-wave ratio and the line length. The simplest case is that where the line is terminated in its characteristic impedance so that the SWR is 1 to 1 and the imput impedance is equal to the Z_0 of the line, regardless of line length.

Coupling systems that will deliver power into a flat line are readily designed. For all practical purposes the line can be considered to be flat if the SWR is no greater than about 1.5 to 1. That is, a coupling system designed to work into a pure resistance equal to the line Z_0 will have enough leeway to take care of the small variations in input impedance that will occur when the line length is changed, if the SWR is higher than 1 to 1 but no greater than 1.5 to 1.

Current practice in transmitter design is to

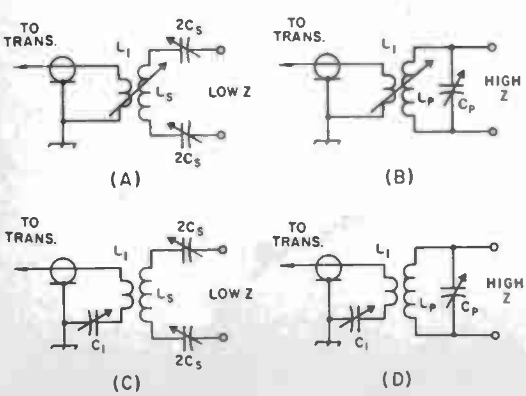
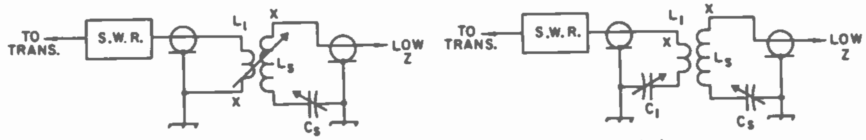


Fig. 20-19 – Simple circuits for coupling a transmitter to a balanced line that presents a load different than the transmitter output impedance. (A) and (B) are respectively series- and parallel-tuned circuits using variable inductive coupling between coils, and (C) and (D) are similar but use fixed inductive coupling and a variable series capacitor, C1. A series-tuned circuit works well with a low-impedance load; the parallel circuit is better with high-impedance loads (several hundred ohms or more).



(A) (B)

Fig. 20-20 — Coupling from a transmitter designed for 50- to 75-ohm output to a coaxial line with a 3- or 4-to-1 SWR is readily accomplished with these circuits. Essential difference between the circuits is (A) adjustable inductive coupling and (B) fixed inductive coupling with variable series capacitor.
In either case the circuit can be adjusted to give a 1-to-1 SWR on the meter in the line to the transmitter. The coil ends marked "x" should be adjacent, for minimum capacitive coupling.

provide an output circuit that will work into such a line, usually a coaxial line of 50 to 75 ohms characteristic impedance. The design of such output circuits is discussed in the chapter on high-frequency transmitters. If the input impedance of the transmission line that is to be connected to the transmitter differs appreciably from the value of impedance into which the transmitter output circuit is designed to operate, an impedance-matching network must be inserted between the transmitter and the line input terminals.

IMPEDANCE-MATCHING CIRCUITS FOR TRANSMISSION LINES

As shown earlier in this chapter, the input impedance of a line that is operating with a high standing-wave ratio can vary over quite wide limits. The simplest type of circuit that will match such a range of impedances to 50 to 75 ohms is a simple series- or parallel-tuned circuit, approximately resonant at the operating frequency. If the load presented by the line at the operating frequency is low (below a few hundred ohms), a series-tuned circuit should be used. When the load is higher than this, the parallel-tuned circuit is easier to use.

Typical simple circuits for coupling between the transmitter with 50- to 75-ohm coaxial-line output and a balanced transmission line are shown in Fig. 20-19. The inductor L1 should have a reactance of about 60 ohms when adjustable inductive coupling is used (Figs. 20-19A and 20-19B). When a variable series capacitor is used, L1 should have a reactance of about 120 ohms. The variable capacitor, C1, should have a reactance at maximum capacitance of about 100 ohms.

On the secondary side, L_s and C_s should be capable of being tuned to resonance at about 80 percent of the operating frequency. In the series-tuned circuits, for a given low-impedance load looser coupling can be used between L1 and L_s as the L_s-to-C_s ratio is increased. In the parallel-tuned circuits, for a given high-impedance load looser coupling can be used between L1 and L_p as the C_p-to-L_p ratio is increased. The constants are not critical; the rules of thumb are mentioned to assist in correcting a marginal condition where sufficient transmitter loading cannot be obtained.

Coupling to coaxial lines that have a high SWR, and consequently may present a transmitter with a load it cannot couple to, is done with an unbalanced version of the series-tuned circuit, as shown in Fig. 20-20. The rule given above for coupling ease and L_s-to-C_s ratio applies to these circuits as well.

The most satisfactory way to set up initially any of the circuits of Fig. 20-19 or 20-20 is to connect a coaxial SWR bridge in the line to the transmitter, as shown in Fig. 20-20. The "Monimatch" type of bridge, which can handle the full transmitter power and may be left in the line for continuous monitoring, is excellent for this purpose. However, a simple resistance bridge such as is described in the chapter on measurements is perfectly adequate, requiring only that the transmitter output be reduced to a very low value so that the bridge will not be overloaded. To adjust the circuit, make a trial setting of the coupling (coil spacing in Figs. 20-19A and B and 20-20A, C1 setting in others) and adjust C_s or C_p for minimum SWR as indicated by the bridge. If the SWR is not close to practically 1 to 1, readjust the coupling and return C_s or C_p, continuing this procedure until the SWR is practically 1 to 1. The settings may then be logged for future reference.

In the series-tuned circuits of Figs. 20-20A and 20-20C, the two capacitors should be set at similar settings. The "$2C_s$" indicates that a balanced series-tuned coupler requires twice the capacitance in each of two capacitors as does an unbalanced series-tuned circuit, all other things being equal.

It is possible to use circuits of this type without initially setting them up with an SWR bridge. In such a case it is a matter of cut-and-try until adequate power transfer between the amplifier and main transmission line is secured. However, this method frequently results in a high SWR in the link, with consequent power loss, "hot spots" in the coaxial cable, and tuning that is critical with frequency. The bridge method is simple and gives the optimum operating conditions quickly and with certainty.

A TRANSMATCH FOR BALANCED OR UNBALANCED LINES

Nearly all commercially made transmitters are designed to work into a 50- to 70-ohm load, and they are not usually equipped to handle loads that depart far from these values. However, many antenna systems (the antenna plus its feed line) have complex impedances that make it difficult, if not impossible, to load and tune a transmitter

Fig. 20-21 — The universal Transmatch shown here will couple a transmitter to almost any antenna system. If the amateur already has a matching indicator, the Monimatch section of the circuit can be eliminated. The counter dial and knobs are James Millen & Co. components.

properly. What is required is a coupling method to convert the reactive/resistive load to a non-reactive 50-ohm load. This task can be accomplished with a Transmatch, a device that consists of one or more *LC* circuits. It can be adjusted to tune out any load reactance plus, when necessary, transforming the load impedance to 50 or 70 ohms.

As has been discussed earlier in this chapter, losses in transmission lines depend on several factors: the size of the conductors, the spacing between conductors, the dielectric material used in the construction of the feed line, and the frequency at which the line is used. Coaxial lines can be classed as lossy lines when compared to a low-loss line such as open-wire feeders, at least below 100 MHz. Because losses increase as the SWR increases, the type of line used to feed an antenna should be chosen carefully. If the transmission line has very low-loss characteristics, high standing wave ratios can be tolerated with no practical loss of power in the line.

A wire antenna, fed at the center with open-wire line, is the most efficient multiband antenna devised to date. For all practical purposes, the feed line is lossless, so extremely high SWRs can be tolerated. This should not be construed to mean that coaxial feed lines cannot be used because of a high SWR, but only the very expensive types are really suitable in this application.

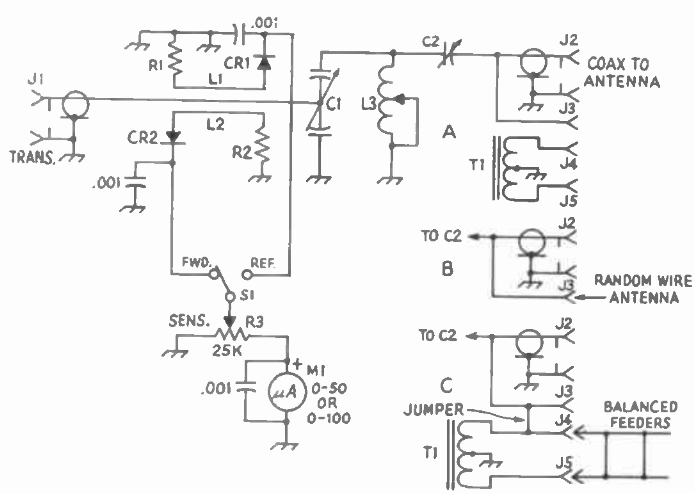


Fig. 20-22 — Circuit diagram of the Transmatch. The .001-μF capacitors used are disk ceramic.

C1 — Dual-section or air variable, 200 pF per section (E. F. Johnson 154-507 or Millen 16250).

C2 — Air variable 350 pF, (E. F. Johnson 154-10 or Millen 16520A).

CR1, CR2 — 1N34A germanium diode.

J1, J2 — Coax chassis connector, type SO-239.

J3, J4, J5 — Isolantite feedthrough insulators.

L1, L2 — See Fig. 20-25.

L3 — Roller inductor, 28 μH (E. F. Johnson 229-203).

M1 — 50 or 100 μA.

R1, R2 — 68-ohm, 1/2-watt carbon or composition.

R3 — 25,000-ohm control, linear taper.

S1 — Spst toggle.

T1 — Balun transformer, see text and Fig. 20-23.

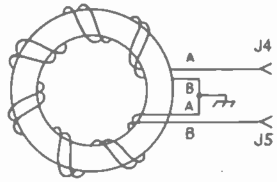


Fig. 20-23 – Details of the balun bifilar windings. The drawing shows the connections required. In the actual balun, the turns should be closed spaced on the inside of the core and spread evenly on the outside.

The Transmatch shown in Fig. 20-22 is designed to handle practically any mismatch that an amateur is likely to encounter. The unit can be used with either open-wire feeders, balanced lines, coaxial lines, or even an end-fed single wire. Frequency range of the unit is from 3 to 30 MHz, accomplished without the use of bandswitching. Basically, the circuit is designed for use with unbalanced lines, such as "coax." For balanced lines, a 1:4 (unbalanced-to-balanced) balun is connected to the output of the Transmatch.

The chassis used for the Transmatch is made of a 16 X 25-inch sheet of aluminum. When bent to form a U, the completed chassis measures 16 X 13 X 6 inches. When mounting the variable capacitors, the roller inductor and the balun, allow at least 1/2-inch clearance to the chassis and adjoining components. The capacitors should be mounted on insulated standoff insulators. The balun can be mounted on a cone insulator or piece of Plexiglas.

The balun requires three ferrite cores stacked for 2-kW or two cores for 1-kW power levels. Amidon type T-200-2 cores are used in making the balun.[1] Each core should be covered with two layers of 3M No. 27 glass-cloth insulating tape. Next, the cores are stacked and covered with another layer of the tape. The winding consists of 15 bifilar turns of No. 14, Teflon-covered wire. Approximately 20 feet of wire (two 10-foot lengths) are required.

A template for the etched-circuit Monimatch is shown in Fig. 20-25. Details for making etched circuits are given in the Construction Practices chapter. If the builder desires, a power-type bridge can be substituted. Such a unit is described in the Measurements chapter. In addition to providing standing-wave indications for Transmatch adjustment purposes, the power bridge will accurately measure transmitter output power.

For coax-to-coax feeder matching, the antenna feed line should be connected to J2 of Fig. 20-22. C1 and C2 should be set at maximum capacitance and power applied to the transmitter. The SWR indicator should be switched to read reflected power. Then, adjust L3 until there is a drop in the reflected reading. C1 and C2 should then be reset, along with L3, until a perfect match is obtained. It

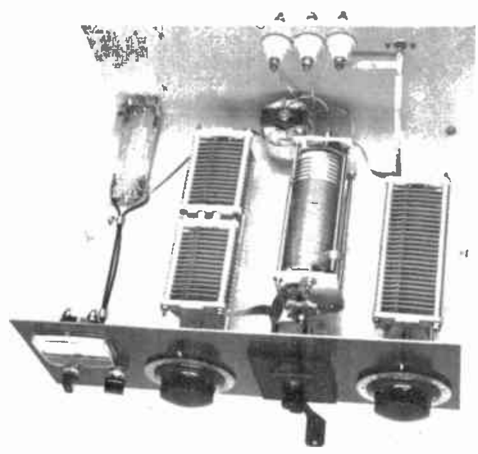


Fig. 20-24 – Interior view of the Transmatch. The etched-circuit Monimatch is mounted 1/2 inch above the chassis. Both C1 and C2 must be mounted on insulated stand-offs and insulated shaft couplers used between the capacitors and the panel knobs. Likewise, T1 should be installed on an insulated mounting. An isolantite cone is used in the unit shown (the balun could be mounted on a piece of Plexiglas). Feedthrough isolantite insulators, mounted through the rear deck, are used for the antenna connectors.

will be found that with many antenna systems, several different matching combinations can be obtained. Always use the matching setting that uses the most capacitance from C1 and C2, as maximum C provides the best harmonic attenuation.

End-fed wires should be connected to J3. Use the same adjustment procedures for setting up the Transmatch as outlined above. For balanced feeders, the feed line should be connected to J4 and J5, and a jumper must be connected between J3 and J4 (see Fig. 20-22 at C).

A slight modification will permit this Transmatch to be used on the 160-meter band. Fixed capacitors, 100 pF each (Centralab type 850S-100N), can be installed across each of the stator sections of C1, providing sufficient C to tune to 1.8 MHz. But, the fixed capacitors must be removed when using the Transmatch on the other hf bands.

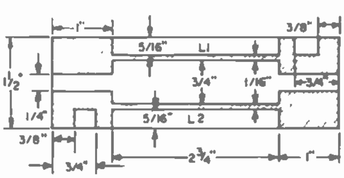


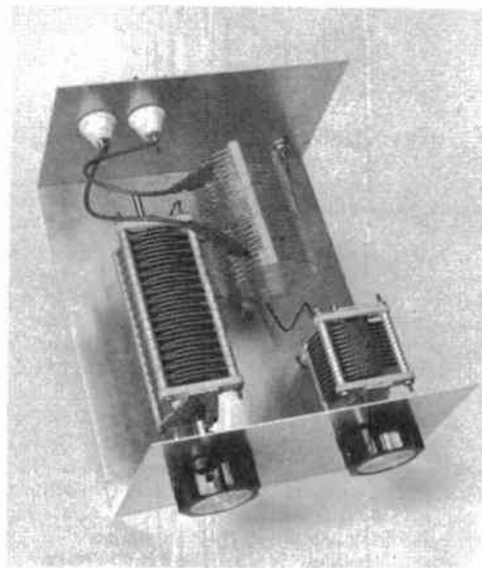
Fig. 20-25 – Template for the etched-circuit Monimatch, foil side shown, etched portion shaded.

[1] Amidon Associates, 12033 Otsego Street, North Hollywood, CA 91601.

A LINK-COUPLED MATCHING NETWORK

Link coupling offers many advantages over other types of systems where a direct connection between the transmitter and antenna is required. This is particularly true on 80 meters where commercial broadcast stations often induce sufficient voltage to either cause rectification or front-end overload. Transceivers and receivers that show this tendency can usually be cured by using only magnetic coupling between the transceiver and antenna system. There is no direct connection and better isolation results along with the inherent band-pass characteristics of magnetically coupled tuned circuits.

Although link coupling can be used with either single-ended or balanced antenna systems, its most common application is with balanced feed. The model shown here is designed for 80- and 40-meter operation since antennas for these bands are the most likely ones to require a matching network.



Construction

An additional bonus of this type of matching network is that it is very simple to build and requires only a few hours time. The layout is not critical and the general approach can be seen in the photograph. Also, components such as capacitors are not critical and just about any type with ratings similar to those specified in Fig. 1 can be substituted. A homemade hand-wound coil could also be used instead of the one specified for L1.

This matching network is capable of handling up to 2 kW PEP provided the voltage rating of C2 is sufficient. Also, instead of C1, a variable link might be employed. This would improve the matching capability of the network over one with a fixed coil. But, even a fixed-link coil might be used if one is willing to experiment with determining the correct amount of turns for the link.

TABLE 1

Band	*Center Section (No. turns)*	*Each End Section (No. turns)*
80 M	12	18
40 M	8	10

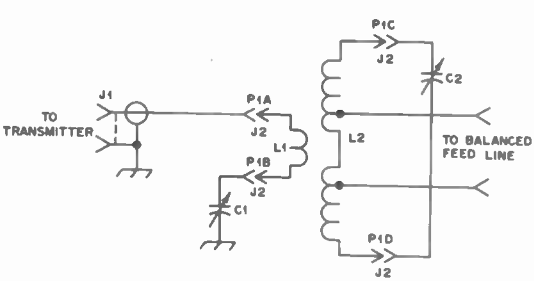


Fig. 1 — Schematic diagram of the 80- and 40-meter link-coupled matching network.
C1 — Variable capacitor, 244 pF, 2000 V Johnson 154-0001-001 or equiv.
C2 — Variable capacitor, 145 pF, 4500 V Johnson 154-0015-001 or equiv.
J1 — Coaxial fitting.
J2 — Millen 41305.
L1 — Barker & Williamson 3034. See Table 1 for details.
P1 — Millen 40305.

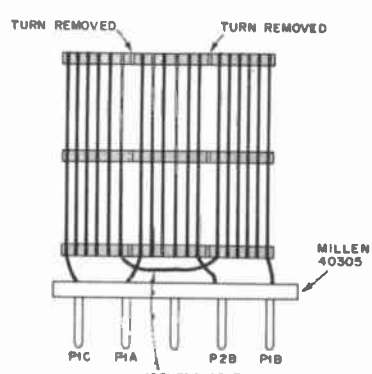


Fig. 2 — Pictorial view of L1 and L2 showing manner in which coils are connected to P1.

A VARIABLE-INDUCTOR ANTENNA COUPLER

Antenna couplers come in all sizes and circuit configurations. The design described here is different than most because it uses a fixed-value capacitor and an inductor which is adjustable. The system may be used on only 80 meters with a half-wavelength dipole, center fed with 300-ohm TV lead-in or 450-ohm open-wire line. Other antennas are unsuitable. The purpose for this description is to stimulate interest in variable-capacitor substitutes rather than in antenna tuning flexibility. No doubt, one could change the inductor and capacitor values for operation on any one of the hf bands but no information will be given here.

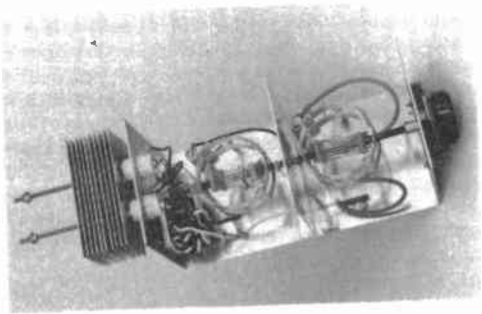
Electrical Design

The circuit shown in Fig. 1 is similar to ones shown earlier in the chapter. The two inductors are used to provide excellent circuit balance for open-wire fed antennas. A fixed-value capacitor is fabricated from aluminum plates and attached to the rear panel on ceramic pillars as shown in the photograph. The capacitance value is preset with the antenna connected and the variable inductors set to midrange. Capacitor adjustment is accomplished by adding or removing plates. Proper phase shift is given by a phase-reversal transformer at the input side of the tuner network. The transformer is attached to the inductors a few turns from the cold end at approximately the 100-ohm points. The two 100-ohm taps in parallel provide a 50-ohm input impedance for the coupler. The system shown here has a range of about 500 kHz when set properly at the center of the 80-meter band with all of the adjusting being accomplished by the variable inductors. The advantage of a one-knob Transmatch is obvious.

Mechanical Details

Most of the construction details may be viewed in the photograph. A U-shaped enclosure is used with a shield located at the center between the two coils. A long phenolic rod, coupled to the vernier dial mechanism, runs the entire length of the chassis and is held in place at the rear with a shaft bushing. These are smaller in diameter than the fixed portion to allow free movement through a complete 180-degree turn.

The rotating inductors are placed in the center of each coil section to provide maximum inductance change with rotation. Several turns are removed at the center of each fixed coil section to allow the phenolic shaft to pass through freely. High-voltage test-probe wire is used for the connections because it is flexible and will withstand the high voltages present in a coupler of this type. The unit shown here is able to handle output from a 2-kilowatt amplifier.



Inside view of the variable inductor antenna coupler. A shield is placed between the two inductors. Care must be taken to assure that moving wires attached to the inductors have a minimum of stress placed on them during dial rotation.

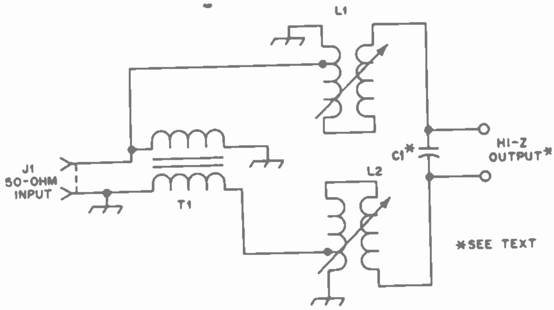


Fig. 1 — Circuit diagram for the antenna coupler.
L1, L2 — 12 turns, six each side of center, 2-1/2-inch dia., 12 tpi, with tap at five turns from cold end. Rotating section is six turns, 2-inch dia., 12 tpi.
T1 — 11 turns of twisted No. 14 PVC covered wire on a T-200 core from Amidon.

Chapter 21

Radiation and Antennas

ANTENNA THEORY

A fundamental principle of nature is that changes in magnetic, electric, and gravitational fields cannot propagate at infinite velocity. In the case of ordinary circuits, the effects of this finite propagation velocity can usually be neglected. This is because circuit dimensions are small enough so that the fields produced in one part of the circuit almost cancel those produced in other parts. On the other hand, radiation would be impossible if it were not for the fact that the speed of light is finite.

The manner in which energy is radiated from a circuit can be illustrated with the aid of Figs. 1 and 2. In Fig. 1, a dc voltage of 4 is applied to the upper electrode in the picture and –4 to the bottom one. Such a configuration is called a dipole although the electrodes do not have to be long thin conductors, necessarily. The actual shape is unimportant for this discussion except for the requirement of symmetry about the *xy* axis. The upper and lower members also have to be identical. Consequently, the field lines and equipotential contours will be symmetrical about *xy*. A three-dimensional view can be visualized by rotating the plot shown in Fig. 1 about this axis.

At distances far removed from the dipole, "distortion" produced by the conductor shape becomes negligible. That is, the conductors can be considered as two point charges. The fields of a configuration of this type are well known and the strength varies as the inverse of the distance cubed. While the analysis only holds for the dc case, it is also reasonably valid for slowly varying ac voltages.

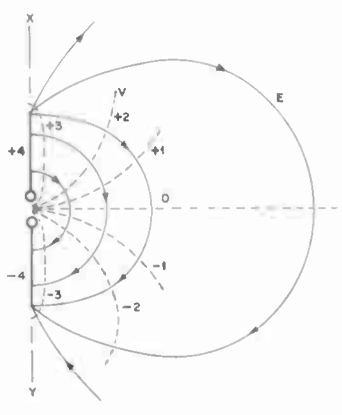


Fig. 1 — Static-field configuration of dipole. Solid lines indicate the direction of electric field and dashed lines show equipotential contours.

However, since the intensity diminishes so rapidly with distance, such fields would not prove very useful for communication purposes. In fact, fields of this type represent energy storage similar to that of ordinary capacitors used in radio work.

Radiation

If a rapidly varying ac voltage is applied to the dipole terminals, the plot illustrated in Fig. 1 is altered drastically because of the finite propagation velocity. It takes time for a change in field conditions at the dipole to reach a given point in space. The time interval is equal to the distance (between the dipole and the point) divided by the speed of light. Fig. 2 indicates how the static field is modified by this effect. Instead of building up instantaneously to the pattern of Fig. 1, the field follows the moving charges on the dipole.

During the first quarter cycle, as charge builds up on the dipole, the field lines tend to spread out toward the position they would occupy under static conditions (as shown in Fig. 2A). However, during the next quarter cycle, the dipole is being discharged and some of the lines break away to form closed loops (Fig. 2B). The process repeats itself during the next two quarter cycles except that the upper half of the dipole is charged negatively. An illustration of the field plot just before a complete cycle is shown in Fig. 2C.

On the other hand, if there were no delay effects, the field throughout space would follow the charge changes instantaneously. Once the source was turned off and the dipole discharged, the field would also be zero everywhere. This implies that all of the energy in the field would be returned to the dipole terminals. However, because of the time delay, "past history" is independent of present events. The effects of turning off the source could not propagate fast enough to affect variations in field that had occurred previously. Consequently, fields such as those of Fig. 2 will continue to propagate out into space in much the same manner as water waves on a pond are formed when an object that disturbs the surface is dropped in. The implication here is that energy is irretrievably lost from the dipole. This lost energy represents electromagnetic energy in the form of radiation such as radio or light waves.

Radiated Power

Along with the energy lost or radiated from the dipole, a certain fraction is also returned during each rf cycle. Consequently, the fields near the antenna represent both energy storage and radiation components. However, at points far enough away, the fields can be considered to belong only to the radiation component. Since it is assumed that the rf energy propagates at the same velocity in all directions, the power flowing through any imaginary sphere of a set concentric with an origin at the antenna must be the same. This is illustrated by the dashed circles in Fig. 2C. The circles represent contours of constant delay time.

An *average* power density can be defined as the ratio of total power flowing through any sphere and its surface area. Since the surface area of a sphere is given by

$$Area = 4\pi r^2$$

the average power density must be

$$S_{ave} = \frac{P}{4\pi r^2} \ (w/m^2)$$

where r is the distance from the origin to a point on the surface (and is also a radius of the sphere).

The power density can be related to the square of the electric and magnetic field strengths in a similar way that power is related to the square of the voltage and current in ordinary circuits.

$$S_{ave} \sim E_{ave}^2 \ ; \ S_{ave} \sim H_{ave}^2$$

Upon taking the square root of the power-density function it can be seen that the radiation fields must be proportional to the inverse of the distance from the source or

$$E_{ave} \sim \frac{1}{r} \ , H_{ave} \sim \frac{1}{r}$$

A derivation of the electric (E) and magnetic (H) fields around an ac-excited dipole is given in a number of texts including *Electromagnetics,* and *Antennas* by John Kraus (McGraw-Hill Book Co.). The only components that vary as the inverse of distance are an electric field that is at right angles to the direction of propagation and a magnetic component also perpendicular to the direction of wave travel. Both components are in time phase with each other but are at right angles in space. Such a wave front is called a transverse-electromagnetic wave (TEM) and forms the principal component in both ordinary transmission lines and waves in space. An illustration of this wave type is shown in Fig. 3.

Antenna Patterns and Gain

The "inverse-square law" discussed in the last section not only concerns the power density of radio waves but describes the variation with distance of other forms of radiation as well. Light waves are an example and the following analogies have counterparts in antenna theory.

For instance, average power density implies a source that radiates equally in all directions. While no such source exists in actuality, many examples approach this condition quite closely. The candle in Fig. 4A would appear as a point source of light of equal intensity when viewed from any direction with the obvious exception of points directly below the base. A bare light bulb would be another simple example. An ideal radiator with this charac-

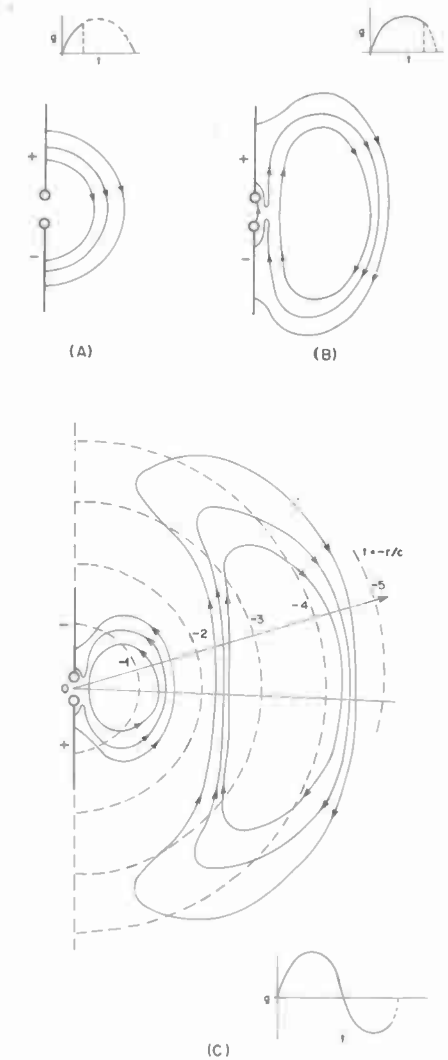


Fig. 2 — Modification of electric field when an ac voltage is applied to the antenna terminals. Inset shows charge on the dipole as a function of time.

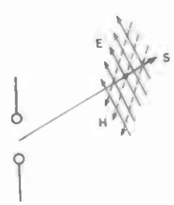


Fig. 3 — Transverse electromagnetic wave. Solid lines represent direction of electric field and dashed lines show direction of magnetic field. There are no field components in direction of power flow (arrow labeled *S*).

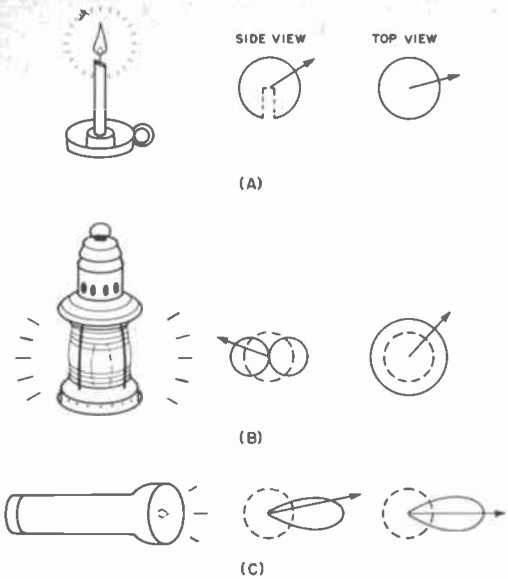


Fig. 4 — Optical analogies to radio antenna patterns. Dashed lines in Figs. 4B and 4C indicate pattern for same power into an isotropic source.

teristic is called an isotropic source and the radiation pattern when viewed from the top or side would just be a circle. That is, the power density directed along an arrow as shown in Fig. 4A would be the same regardless of the direction. In the case of an isotropic source, the peak or maximum power density is the same as the average power density discussed in the previous section.

A type of lantern often used for marine purposes shown at Fig. 4B typifies another source. Light rays that would normally radiate upwards are bent or refracted so that the radiation in the horizontal direction is increased. (The lantern lens illustrated here is called a Fresnel lens and a similar principle is used in certain microwave antennas.) As a consequence, the radiation in the horizontal direction exceeds the average power density by some factor. This effect occurs naturally with the dipole antenna and the radiation pattern of the dipole with the orientation shown in Fig. 3 is illustrated at the right in Fig. 4B. For a dipole that is short compared with the wavelength of operation, the maximum power density exceeds the average power density (the power density that would result for the same input power to an isotropic source) by a factor of 1.5. If the length is increased to a half wavelength, the factor is increased to 1.64.

On the other hand, the lantern radiates the same amount of light through a 360-degree arc in the horizontal direction. When viewed from above, the radiation pattern appears as a circle. Antenna patterns of this type are said to be omnidirectional. Vertical dipoles in free space and ground-plane antennas are examples.

Finally, the flashlight (Fig. 4C) is representative of a number of antennas very popular in amateur radio. Power that would normally be radiated in many directions is concentrated into a small arc or beamwidth. As in the flashlight, the energy can be concentrated by a reflecting surface or by a lens-like object called a director. A radiating element in the antenna corresponds to the bulb.

A convenient way of determining how effective an antenna is in concentrating power in a given direction is expressed in a parameter called gain. The gain of an antenna is the ratio of power density directed in some direction compared to a test antenna. Both antennas must have the same power input and usually, the *orientation of the most interest is along a line where the power density is the greatest* (for both antennas). Unless otherwise stated, gain is usually considered in the latter context.

If the test antenna is assumed to be an isotropic source the orientation wouldn't be important. However, since such antennas do not exist in nature, a practical model is required. The most common type of reference antenna is the half-wave dipole. At vhf and uhf, special antennas with reflectors and known characteristics are employed while in the microwave region, horns and other types are used.

Directivity is a parameter very similar to gain except that antenna losses are assumed to be zero. On the other hand, gain includes the effects of losses as well so it is important to know which characteristic is being discussed, especially in the consideration of antenna specifications. For instance, an antenna that is electrically small at the frequency of operation may have impressive directivity characteristics while being quite inefficient because of the losses present. Consequently, the gain will be quite low.

Radiation Resistance

Considerable confusion can result if such factors as antenna gain, directivity, efficiency, size, and aperture are not regarded in their proper contests. For instance, experienced amateurs might balk at the fact that the directivity of a half-wavelength dipole is only 1.64 compared with 1.5 for a short dipole giving a difference of 0.4 dB. This would imply that a four-foot dipole on 80 meters for instance, would be just as good as a half-wave version which would correspond to approximately 135 feet! The old amateur-radio adage, "the bigger the better," in regard to antenna performance would seem to be misleading.

However, another factor must be taken into account and that is the radiation resistance "seen" by the source or transmission line when connected to the antenna terminals. It will be recalled in the discussion concerning the radiation process shown in Fig. 2, energy was lost because some of the field lines broke away from the dipole and were propagated into space. In essence then, a "good" antenna configuration is one where the field extends far away from the conductors. Considered from a slightly different point of view, antenna dimensions and associated field shape should be large at the wavelength of operation so that delay

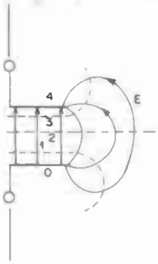


Fig. 5 — Electric-field pattern of parallel-plate capacitor. Complete pattern can be visualized by rotating two-dimensional field plot around axis.

effects in the fields caused by currents on different parts of the conductors will be maximum. For instance, the parallel-plate capacitor shown in Fig. 5 would be a relatively poor radiator. Most of the field is confined between the plates and only the "fringing" field at the edge of the plates and beyond would contribute to a radiation component. (An exception would be if the dimensions of the capacitor were so large the circumference approached an appreciable fraction of a wavelength. Then, the radiation would be from the fields across the slot and the entire system would be a "dual" of an antenna made from solid conductors. This effect is often of an undesirable nature in regard to designing effective shields out of sheets that are not completely bonded or soldered together. "Leakage" of rf energy will occur from the cracks.) A similar effect occurs with the so-called inverted-vee dipole. Instead of a straight dipole, the conductors are run off at an angle from the feed point. This permits the use of one high support with the ends of the antenna tied to lower supports near the ground. However, if the angle of the vee at the apex becomes too sharp, the fields tend to cancel rather than radiate.

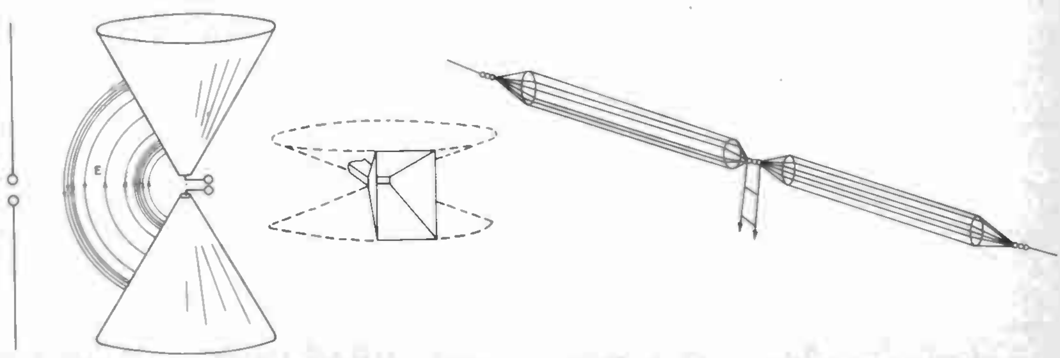
The result of any of these effects is that the radiation resistance of the antenna becomes very low in value. This means that a high current is required to produce the same radiated power in comparison with an antenna with a higher value of radiation resistance. As a consequence, the effect of losses in such devices as matching networks, ground systems, and similar areas where currents are required to produce the radiated field become significant. A point may be reached where more power is dissipated in the losses than radiated.

In cases where the losses can be neglected, the gain and directivity are the same. But while the directivity may remain the same, the antenna gain will decrease as the effect of antenna loss increases. As pointed out earlier, a small-sized antenna has the capability of being a good performer, but considerable care must be taken to insure losses do not offset any advantages. Also, since the ratio of energy stored to energy lost in the form of radiation (per rf cycle) is greater in a smaller antenna compared to a larger one (at the same frequency), the bandwidth becomes smaller. The ratio of energy stored to energy lost will be recognized as being proportional to the Q of ordinary circuit theory. Hence, a small antenna represents a high-Q system and less frequency variation is permitted before retuning will be required.

The Biconical Antenna

In the last section, the effect of the linear dimensions on antenna bandwidth was mentioned. However, the shape and size of the conductors plays a role as well. An important antenna from a theoretical point of view is the biconical-dipole configuration. This is shown in Fig. 6B. If the cone angle is made very small, the configuration approaches that of a wire dipole (Fig. 6A). On the other hand, a horn antenna used at microwave frequencies can be thought of as being a sector of a biconical dipole with a very large cone angle (Fig. 6C).

Fig. 6 — The biconical dipole can be thought of as a connection between "linear" dipoles and "aperture" types such as horn antennas. An approximation of the biconical dipole is the cage antenna consisting of single conductors which simulate a solid surface (Fig. 6D).



(A) (B) (C) (D)

An interesting feature of the biconical antenna is that the principal fields caused by currents along its surface are in the form of TEM waves. It will be recalled that such waves are the ones involved in radiation. They are also the principal modes on ordinary transmission lines such as coaxial cable and parallel-conductor line. From an intuitive reasoning, it might be expected that the biconical would provide a better "match" between the modes on the transmission line and those of the radiation field and consequently have a greater bandwidth. This is true and both the biconical dipole and the sectoral horn have large bandwidths while the linear dipole (Fig. 6A) has a relatively narrow bandwidth. Generally speaking, increasing the conductor width tends to result in greater bandwith. In cases where increasing the conductor surface may not be practical, simulating a larger surface by a series of wires is often employed. The "cage" dipole shown in Fig. 6D is a very old method of accomplishing this goal. Another advantage is that antenna currents are spread over a number of conductors with consequent reduction in ohmic losses. This is more of a problem in lf and vlf antennas where the antenna length is usually very short compared to the wavelength of operation. However, the effect of such loss is likely to be negligible with amateur antennas although the idea of the cage dipole offers attractive advantages in regard to antenna bandwidth.

Effective Aperture

Some concept of the part played by the physical configuration of the antenna in the process of radiation should be emerging from the previous sections. Power is transferred from the antenna feed line to the conductors where the currents generate electric and magnetic fields. Rather than contributing to the radiation directly, the conducting surfaces of the antenna act as guides for currents which in turn produce the fields. Consequently, it is the field and not the physical configuration that plays the predominant role in determining the manner in which energy is radiated. For instance, it is often assumed that increasing antenna size results in, "more power being radiated," or a similar platitude. Instead, rf energy is merely collimated toward some direction – hopefully, the desired one.

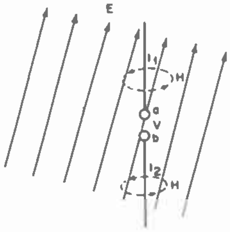


Fig. 7 – Fields from a passing radio wave will induce local currents on an antenna and a voltage across the terminals *ab*. Since a magnetic field will be produced by the currents on the conductors, some power will be reradiated or scattered.

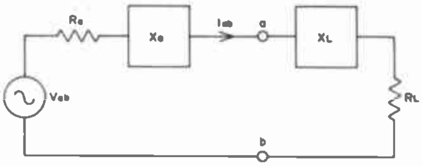


Fig. 8 – Equivalent circuit of antenna and load.

Gain and radiation pattern are the important characteristics of any antenna system. However, it is sometimes useful to introduce a third property called effective area or effective aperture. The effective area of an antenna is the ability to transfer energy from a passing radio wave to a terminating load such as the input of a receiver or a transmission line. It is defined as the ratio of power delivered to the load and power density in the passing TEM wave. For example, if the power delivered to the load was 10 watts for a power density of 1 watt/m^2 the effective area would be 10 square meters. Since aperture is related to the antenna when it is operated in the receive mode, the subject will be covered in more detail in the next section.

Receiving Antennas

A source connected to an antenna excites fields in the surrounding region which propagate outward in spherical fashion. At great distances, the nature of the radiation approaches that of a transverse-electromagnetic wave. If another antenna connected to a load is in the path of this wave, currents will be induced on the conductors and a voltage will be produced across the load. It is evident that the field configurations will differ considerably between identical antennas if one is being used for transmitting and the other one for receiving.

Although the fields surrounding the antenna are different for receiving and transmitting, there are several very important similarities *concerning conditions at the antenna terminals*. In most practical applications, the terminal characteristics are the ones of interest. First, consider the antenna shown in Fig. 7 that is immersed in the fields of a passing wave. It is also convenient to assume all conductor losses are negligible and the only resistive component is the desired load connected to the terminals. Consequently, if the terminals *ab* are open as shown in Fig. 7, the antenna will not absorb any power. However, "local" currents I_1 and I_2 will be induced on the conductors since a field component parallel to the wires will cause the charges on the surface to move. In turn, this current will produce a magnetic field at the same frequency as the passing wave. Since the field would be indistinguishable from one caused by a source connected to an antenna terminal, rf energy will be reradiated or "scattered." Although the antenna is not absorbing any power, it is still capable of reflecting or scattering rf energy.

Next, assume that a load is connected across the terminals *ab* as shown in Fig. 8. An equivalent

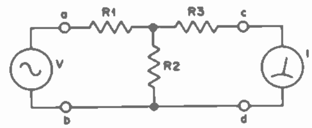


Fig. 9 — Show that the source and ammeter can be interchanged without changing the meter reading.

circuit of the antenna can be represented by an "open-circuit antenna voltage" shown as V_{ab} in series with a radiation resistance, R_a and an antenna reactance, X_a. The load is shown as a reactance, X_L in series with a resistance, R_L. If R_a is equal to R_L and if X_a and X_L are of opposite signs so that they cancel, maximum power will be transferred from the fields of the passing wave to the load. However, half of the power is "dissipated" in R_a. The significance of the latter phenomenon is as follows. Currents such as I_{ab} caused by induced fields are indistinguishable from ones generated by a source connected to the antenna terminals. Consequently, the current flowing in R_a represents some more reradiated power in addition to that scattered from the antenna conductors themselves.

Reciprocity

An important theorem concerning circuits is the principle of reciprocity. If a network consists of linear, bilateral (no one-way current flows), passive (no amplifiers) elements, the theorem states that a current produced in one part of the network by a source in another part will be the same if the source and ammeter are interchanged. The only requirement is that the source and ammeter do not upset the network configuration after the interchange is made. A source and ammeter with zero internal resistance would satisfy this requirement and the reader might verify the theorem for the resistive network shown in Fig. 9. Show that the current through the ammeter is the same if the source is connected at terminals *c* and *d* while the ammeter is connected across *a* and *b* (assume both have zero impedance).

The theorem can be extended to two antennas in space (as shown in Fig. 10) provided conditions similar to those of the circuit case exist (no changes caused by the ionosphere, for instance). A voltage applied to the terminals of antenna 1 produces a current in the ammeter connected to antenna 2. By the reciprocity theorem, the source and ammeter could be interchanged and the ammeter reading would be the same.

An interesting implication of the reciprocity theorem for antennas is that the radiation patterns for both receiving and transmitting *must be identical.* For instance, if the receiving pattern of antenna 1 had a null in the direction of antenna 2, but none while transmitting, the current produced in antenna 1 when the source and ammeter were interchanged would be zero. However, this would contradict the reciprocity theorem since a non-zero reading would occur with antenna 1 operating in the transmitting mode.

The fact that the receive and transmit patterns are identical has implications for the directivity and aperture of antennas. This is illustrated in Fig. 11. Assume that the antenna shown at Fig. 11A has the pattern with a directivity indicated by the dashed lines. If an incident wave of power density S passes by the antenna a power P_1 will be delivered to the load, R_L. Next, assume that the directivity has been increased as indicated by the sharper pattern shown in Fig. 11B. As a consequence, the received power P_2 will be larger than P_1. However, if a greater amount of power is transferred to the load, the aperture must have increased also. Thus, if the directivity increases, the effective aperture increases by the same amount. Since directivity is usually defined as the ratio of *maximum* radiation intensity to average intensity, it is the *maximum* effective aperture that increases in proportion to the directivity.

Received and Transmitted Power Relations

Using the fact that the directivity and maximum effective aperture increase in direct proportion along with the definition of an isotropic source, a simple relation between directivity and maximum effective aperture can be derived by

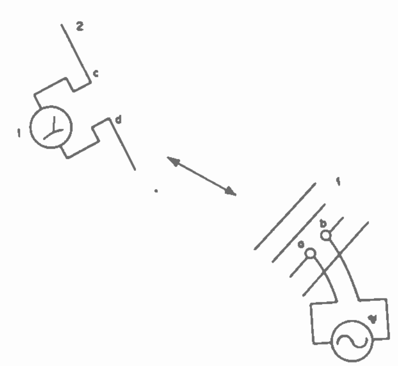


Fig. 10 — Reciprocity as applied to transmitting and receiving antennas.

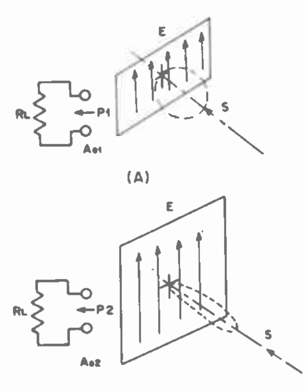


Fig. 11 — Increase in directivity results in a similar increase in aperture.

comparing these characteristics for actual antennas. This relation is given by

$$D = \frac{4\pi}{\lambda^2} A_{\text{max}}$$

It is now possible to derive a relation between the power delivered to the transmitting antenna and the power received by another antenna that is at a distance r from the source antenna.

A previous formula for the power density at a distance r away from an isotropic antenna was

$$S = \frac{P_t}{4\pi r^2}$$

Since the received power is related to the effective aperture (here assumed to be the maximum area) and power density by

$$P_r = A_{er}S$$

the ratio of received power to that delivered to the terminals of the isotropic source will be

$$\frac{P_r}{P_t} = \frac{A_{er}}{4\pi r^2}$$

If instead of an isotropic source, the antenna has a directivity D, the relation becomes

$$\frac{P_r}{P_t} = \frac{A_{er}D_t}{4\pi r^2}$$

Upon substitution of the relation between directivity and aperture, the formula can also be written as

$$\frac{P_r}{P_t} = \frac{A_{er}A_{et}}{\lambda^2 r^2}$$

which is known as the Friis Transmission Formula.* It can also be written in terms of the directivities as

$$\frac{P_r}{P_t} = D_t D_r \left(\frac{\lambda}{4\pi r}\right)^2$$

The transmission formula is a good approximation for vhf and uhf antennas with sufficient gain and are high enough so that ground reflections can be neglected. It is less useful at hf since the path is usually not very well defined and attenuation caused by ionospheric conditions will introduce additional errors. Similar formulas can be derived for purely ground-wave propagation although the curvature of the earth begins to affect the inverse-distance approximation at distances greater than 50 miles. Also, the conductivity of the earth over the propagation path of the ground wave is an important factor. The lowest loss occurs over sea water, but the attenuation increases rapidly over land areas where the conductivity is low. The subject of propagation is taken up in more detail in another chapter including the effects of the ionosphere.

Frequency Scaling

A point mentioned earlier was that the field configuration played the most important role in antenna operation. The transmission formulas and other relations in the last section dramatize this phenomenon although at first sight, the connection might not be so obvious. Fields are geometrical entities and like other mathematical relations they are independent of the dimensional system. The fact that a meter or foot happened to be picked at certain physical size does not alter geometrical shape or affect mathematical operations. Of course, there is a connection to the "outside world" through various constants.

The formula for the directivity in terms of the aperture would seem to contradict this philosophy since it would appear that the directivity varied with wavelength. However, the aperture of practical antennas is also proportional to the square of the wavelength and this squared term cancels. For instance, the maximum effective aperture of a dipole is

$$A_{em} = \frac{3}{8\pi} \lambda^2$$

provided the *electrical* length is small compared to a wavelength. Consequently, a short dipole has the same directivity at lf as it does in the uhf region.

Since the dimensions of the dipole change, the distance between two dipoles in a transmit/receive situation should also be changed in order to keep the same geometrical proportions. Examination of the transmission formulas in terms of the directivities indicates this is so. As the wavelength becomes shorter, the separation also must be made shorter if the same ratio of received to transmitted power is to result.

This dependence on geometry rather than frequency permits scaling of situations involving radiation such as the transmit-receive antenna problem. The principle is quite useful since an antenna can be modeled at some convenient frequency (such as at vhf) and then scaled appropriately to the desired frequency of operation. However, such scaling must include proper consideration of *all* parameters. This becomes a problem when factors such as loss must be taken into account as well.

While the object of such scaling might be to improve an antenna from the standpoint of bandwidth, the most common goal in amateur designs is an improvement of gain (perhaps in some desired direction if the system is immobile). The next section deals with some of the factors that affect gain.

Arrays and the Half-Wave Dipole

Horns, lens antennas, and the parabolic-reflector are antenna examples that closely resem-

*Friis, A Note on a Simple Transmission Formula, *Proc. I. R. E.*, Vol. 34, pp. 254-256, May, 1946.

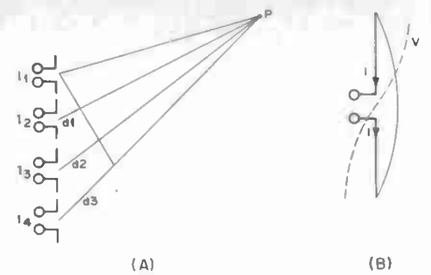


Fig. 12 — An antenna array is usually considered to consist of a series of separate elements. However, the concept can be extended somewhat to include configurations consisting of continuous conductors.

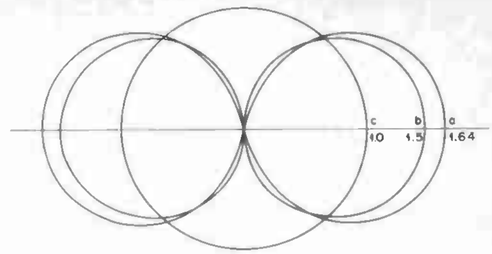


Fig. 13 — Radiation patterns and directivities of an isotropic source (*c*), a short dipole (*b*), and a half-wave dipole (*a*).

ble their counterparts in optics. The increase in gain is accomplished by reflection or refraction of the propagated wave front. There is also another very common antenna type that works on a principle analogous to the optical phenomenon of diffraction. Cancellation or reenforcement of the fields from a number of sources at a given point occurs because of phase shift caused by different time delays. This is illustrated in Fig. 12A. Assuming the currents in each short dipole (I_1 through I_4) are identical in magnitude and phase, the fields will reenforce at a point broadside to the array. (An array can consist of a number of separate antennas arranged in some configuration. The individual antennas that make up the array may or may not be identical.) However, at other angles, there will be a difference in distance from each antenna to a given point P indicated by d_1 through d_3. Some cancellation of the antenna fields will occur and the overall pattern of the array will be different from a single dipole taken alone.

One might wonder if instead of an array of separate dipoles, the configuration consisted of a single continuous conductor; would the same effect of cancellation and reenforcement occur because of different distances from a point to different parts of the conductor? For instance, the so-called half-wave dipole shown in Fig. 12B is *assumed* to have the sinusoidal current and voltage distribution illustrated. Since the dimensions of the antenna are no longer small compared to a wavelength, the fields from each "element" of the conductor will only be in phase at a point broadside to the antenna as was the case of Fig. 12A. At other angles, cancellation of the fields will exist and as a result, the half-wave dipole will exhibit some gain over its electrically short counterpart.

The pattern for the half-wave dipole is shown in Fig. 13 along with those of the electrically short dipole and an isotropic radiator for comparison. Maximum radiation is indicated for the three types by points a, b and c, respectively. However, a theoretical difficulty exists with the current and voltage distribution. A purely sinusoidal distribution would imply an infinite SWR on the "transmission line" formed by the dipole conductors. If this were true, there could be no radiation since all resistive components (including radiation resistance) would be zero. Therefore, the actual distribution departs somewhat from a pure sinusoid in order to account for the radiation resistance which is approximately 73 ohms.

This effect doesn't alter the radiation pattern of a thin-wire dipole and the sinusoidal approximation is a good one even for antennas longer than a half wavelength. However, extending the antenna to lengths much greater than a wavelength often doesn't yield any particular advantage since the radiation pattern breaks up into a number of minor lobes rather than increasing intensity in the broadside direction. Some of these patterns are shown in *The ARRL Antenna Book*.

MISCELLANEOUS ANTENNA TYPES

There are three simple antenna configurations consisting of thin conductors. These are the short dipole, small loop and the helix. The short dipole has been discussed already and it was pointed out that the half-wave dipole could be considered to be a continuous array of short dipoles. Small-loop antennas are seldom used for transmitting since the radiation resistance is very low. However, they are very practical for receiving at hf and below since atmospheric noise usually offsets any effects the low gain might have on signal-to-noise ratio. Unlike the short dipole, the small loop is relatively insensitive to local electric fields and can be used to advantage where such fields cause interference. Another property of the small loop is that very sharp nulls in the radiation pattern exist which make the loop a very useful type for direction finding purposes.

The helix antenna is capable of two entirely different modes of operation that depend upon the physical configuration. A long, thin helix where the circumference of each turn is small compared to a wavelength radiates in much the same manner as the linear dipole. Since radiation is greatest broadside to the longest dimension (as with the dipole), this type of operation is designated as the normal mode. However, unlike the dipole, there is a small component of electric field at right angles

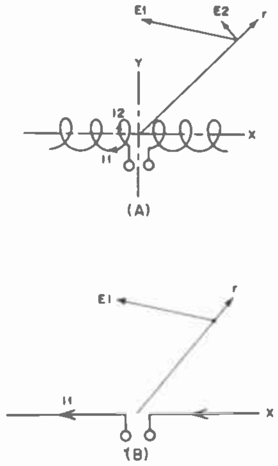


Fig. 14 — The helical antenna.

to the principal "dipole field" as illustrated in Fig. 14A. This is because there is a component of current in the y direction (I_2) along with ones in the x direction (I_1). These components generate fields E_2 and E_1, respectively.

Actually, the fields produced by I_2 are very similar to those generated by a small loop and the normal mode is sometimes analyzed by considering the helix to consist of a series of dipoles and loops connected in series. Since fields generated by loops and dipoles are in phase quadrature, E_1 and E_2 produce a resultant field that is elliptically polarized. Fields from a dipole are linearly polarized since the E_2 component is always zero (Fig. 14B).

Short-Loaded Dipoles

The normal helix mode is important theoretically in the design of short-loaded dipoles used in the hf region. Typical examples are those used for mobile purposes and in shortened loaded beams. The effects of the energy stored in the electric fields surrounding the dipole can be approximated by a series of capacitors as shown in Fig. 15A. The capacitance between the upper and lower halves will be greatest near the terminals (C_1) and decreases for points farther away (C_2 and C_3). As a consequence, the current caused by an applied voltage at the terminals will progressively diminish as the ends of the antenna are reached. The current distribution along the dipole is indicated by the dashed lines in Fig. 15A. Since the radiation resistance is proportional to the length (of an electrically-short dipole) where the current is the greatest, the sections near the ends of the antenna are not being used effectively.

Because a short dipole appears as a large capacitive reactance in series with a small radiation resistance, an inductive reactance is required to "tune out" the unwanted reactance and bring the system to "resonance." Ordinarily, the physical location of the inductors would not be important but there are advantages to be gained by employing the configuration shown in Fig. 15B. By placing such "loading coils" off center, and causing them to resonate with the outer sections (C_2 and C_3), the current distribution can be improved as shown in Fig. 15B. The sections of the dipole out to the coils now have higher currents and the radiation resistance increases. However, since the capacitive reactance out near the ends (C_2 and C_3) is greater than that near the terminals (C_1), more inductance is required. This effect is sometimes offset by increasing the capacitance of outer sections by means of "capacitive hats."

A question arises as to the most desirable configuration for the loading coils since the Q of any coil is a function of shape. One point that should be kept in mind, however, is that a resistive component that causes the Q to decrease is the result of radiation. Normally, coils are formed so that radiation losses are minimum, but in the case of loading-coil design, this imposes an unnecessary restriction for the configuration shown in Fig. 15B. Such coils should be designed for low loss and not high Q. Spacing between turns should be as large as practical and large-sized conductors should be used. Long, thick loading coils have been incorporated into many very successful antenna designs even though the circuit Q of such inductors may be lower than a value some designers would consider desirable.

The answer to such success can be seen with the aid of Fig. 14A. Since the field intensity of a small loop is proportional to the area of the coil, the component of field caused by the "loop-part" of the helix (E_2) is usually negligible if the diameter of the helix is not too great. This is ordinarily the case with the loading coils shown in Fig. 15B and the radiation will be almost linearly polarized. Such a system with a long loading coil might be thought of as a hybrid helix-dipole configuration. Occasionally, antennas are constructed that consist wholly of helical elements. However, the thinner conductors often employed in such designs to offset construction difficulties may negate any

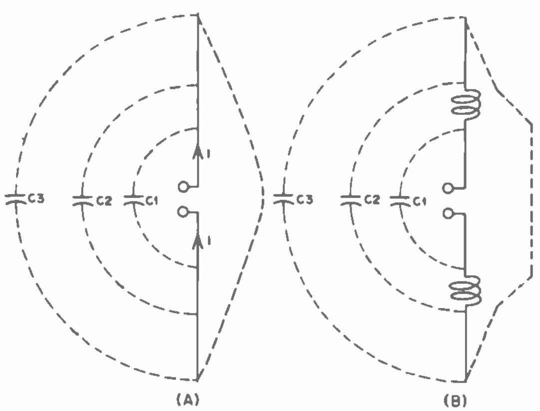


Fig. 15 — Current distributions on unloaded and loaded dipoles.

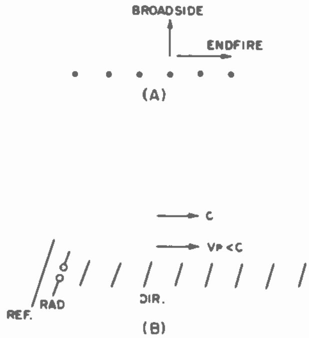


Fig. 16 — Maximum radiation may occur at right angles or parallel to the plane of an array. The Yagi-Uda array is an example of an "endfire" array.

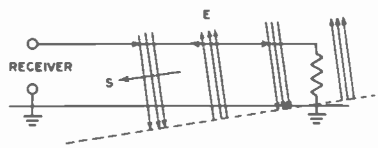


Fig. 17 — The Beverage antenna.

advantages over the hybrid combination, especially if capacitive loading is employed in the latter also.

Traveling-Wave Antennas

If the circumference of the helix is increased to approximately a wavelength, an entirely different mode of operation is possible than the so-called normal mode discussed in the previous section. Maximum radiation occurs along a direction that is parallel to the axis of the helix and consequently, is designated as the axial mode. However, the phenomenon is characteristic of a number of very important antenna types.

The antenna types covered so far have been mostly of the broadside variety with maximum radiation in a direction perpendicular to the long dimension of the antenna or array. It is also possible to adjust the phase of the individual elements that make up an array so that radiation is maximum in a direction parallel to the long dimension. An antenna with this type of pattern is called an end-fire array as illustrated in Fig. 16A.

There are a number of ways to excite the elements of the array illustrated in Fig. 16A. The most direct way would be to couple them with transmission lines whose lengths were such that the proper phasing was insured. However, this is not only cumbersome but becomes increasingly difficult to accomplish as the frequency of operation is raised.

Since current in one element will induce a current in another one in close proximity, it is possible to simulate the phasing lines by the configuration of Fig. 16B. Here, a number of linear elements are arranged so that the excitation is caused by coupling effects alone. Only one element is connected to the source (called the driven element or radiator) while the rest are coupled indirectly. The latter are called parasitic elements and may be either directors or reflectors depending upon their effect on antenna pattern.

Elements that redirect energy from the direction of travel are called reflectors while those that enhance radiation in the direction of wave travel are called directors. Only one or two reflectors are required to minimize the backlobes of the radiation pattern while many directors are required to obtain maximum gain in the forward direction. With hf antennas, the number of elements is limited, because of construction difficulties, to 5 or 6. However, vhf and uhf antennas may have as many as 13 elements and sometimes more. Regardless of the number, the configuration shown in Fig. 16B is called the Yagi-Uda array in honor of its discoverers.

It is also possible to think of an array with periodic (or almost periodic) spacings and elements such as the one shown in Fig. 16A in a slightly different manner. The driven element, reflector, and the first few directors perhaps, act as a launcher for a wave that travels toward the end of the antenna along the directors. The velocity of this wave is somewhat less than the speed of light so that the phase conditions for optimum end-fire radiation are satisfied. Maximum gain occurs if the so-called Hansen-Woodyard conditions are satisfied and such arrays are sometimes referred to as supergain arrays.

The array need not consist of linear elements such as the one shown in Fig. 16B. A helix operating in the axial mode, and a waveguide with holes or slots arranged periodically are other examples. Since rf energy "leaks out" as the wave progresses inside of the guide, such configurations are called leaky-wave arrays.

Two other antennas that work on a wave principle are shown in Figs. 17 and 18. A vertically polarized wave traveling over poorly conducting ground becomes attenuated. The effect of this attenuation was as though the wave was being tilted into the ground rather than traveling along the surface. Consequently, there is a component of electric field parallel to the surface of the earth which will induce a current in a horizontal wire as shown in Fig. 17. A wave traveling from right to

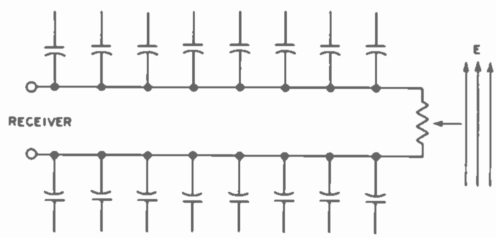


Fig. 18 — The Fishbone antenna.

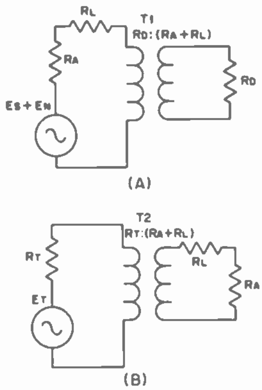


Fig. 19 — Equivalent circuits of a typical receiver input circuit (A) and a transmitter output stage (B) as related to antenna characteristics.

left induces a voltage that can be detected at the receiver while the power in a wave going from left to right gets absorbed in the resistor. As a result, the antenna displays a unidirectional property and is very useful in receiving applications where directional noise is a problem.

The antenna shown in Fig. 17 is called the Beverage antenna and a similar principle is employed in the fishbone antenna illustrated in Fig. 16. A number of radiators are coupled lightly to a parallel-wire transmission line. A passing wave induces a traveling wave on the transmission line which either terminates at the receiver or is dissipated in the resistor. So-called long-wire antennas such as the V and rhombic are sometimes terminated in order to obtain a unidirectional characteristic. While such termination improves the receiving characteristics, it is obvious that the unidirectional pattern doesn't result in higher gain in the transmitting mode since the power that would normally be radiated in the reverse direction is dissipated.

LOSSES IN ANTENNA SYSTEMS AND NOISE

Experienced radio operators might balk at the concept of reciprocity as applied to transmitting and receiving antennas. For instance, a particular antenna type might be very good for receiving purposes but be totally unsuitable for transmitting.

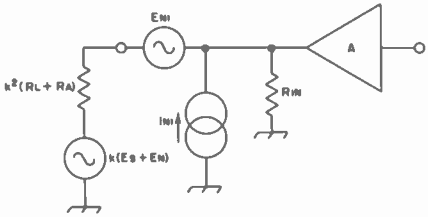


Fig. 20 — Sources of noise in amplifier input consist of both voltage and current generation.

An example would be the small loop. Small loops have been used in direction-finding receivers for years but because of the low radiation resistance in comparison to othet types they are seldom used for transmitting. The Beverage antenna would be another type used solely for receiving.

The answer to this apparent paradox can be resolved by considering the effects of noise and losses on the entire communications circuit. A typical receiver antenna and input stage is shown in Fig. 19A. R_D represents the input load required by the receiver, T_1 is a matching network that transforms the total resistance of the antenna circuit to R_D, R_a is the antenna radiation resistance, R_L is the loss resistance of the antenna circuit, E_s is the voltage induced in the antenna by the fields of the received signal, and E_n is the voltage induced in the antenna by unwanted interference or noise.

A word about T_1 and R_D is perhaps in order. There may be reactances associated with the circuits of the antenna and/or the first stage of the receiver. T_1 usually performs a double function of matching to $R_L + R_a$ to R_D along with tuning out this unwanted reactance. Normally, R_D would just be the input resistance of the first stage, however, the required source resistance might differ considerably from one which would result in maximum power gain. This is shown in Fig. 20. The input of a device such as a transistor can be represented by an input resistance, R_{in} along with an internal source of noise current (I_{ni}) and noise voltage (E_{ni}). These sources should not be confused with noise generated in the external circuitry including the antenna.

Assuming that T_1 is adjustable and that there are no other sources of power gain between the antenna and the circuit of Fig. 20, the conditions for optimum signal-to-noise ratio can be determined as follows. Varying the voltage transformation ratio by a factor of k means the impedance transformation ratio must change by a factor of k^2. An obvious choice of k might seem to be a high value that resulted in the greatest amount of signal being applied to the input circuit. However, the transformed resistance would be increased accordingly. Consequently, a greater percentage of the current I_{ni} would be diverted into R_{in} and the noise figure (which is a measure of merit with low numbers being the most desirable) would be degraded.

On the other hand, decreasing k would result in a lower source resistance and less internal noise current would flow through R_{in}. However, the applied voltage ($E_s + E_n$) would be correspondingly less in comparison to the internal noise voltage (E_{ni}) and the noise figure would again be degraded.

An optimum transformation ratio exists and is usually found by experiment. However, the transformed source resistance may not be equal to R_{in} for this ratio and conditions for maximum power gain do not occur. In order to distinguish a match for maximum power gain from one for best noise figure the term noise match is often employed.

Ordinarily, obtaining best noise figure is only important at vhf and higher since atmospheric noise is the limiting factor at lower frequencies. At hf then, considerable loss (R_L) can exist in the antenna system and as long as there is sufficient signal to override the internal noise of the first stage, the effects of such loss can be compensated by merely employing more gain in the receiver. Therefore, a rather inefficient antenna system can be used in many hf receiving applications without seriously degrading overall signal-to-noise ratio. In fact, some of the more inefficient types may have desirable characteristics in regard to signal-to-noise ratio.

The Small Loop Antenna

An example of such an antenna is the small loop shown in Fig. 21. Currents induced in the loop from a passing radio wave produce a voltage in the secondary of the coupling transformer, T_c, in a conventional fashion. However, if the loop is in the presence of local electric fields such as those from fluorescent lights, the induced currents from these fields tend to cancel. This is shown in Fig. 21 where the unwanted noise source is represented by E_n and a coupling capacitor, C_n. The currents coupled into the loop tend to split into two components, I_{n1} and I_{n2}, which travel in opposite directions. Since the currents are out of phase when they reach T_c, no voltage is induced in the secondary since the resultant difference between I_{n1} and I_{n2} is usually very low. An important factor here is to keep the loop balanced with respect to ground and surrounding objects. An alternate approach is to shield the loop but with a break in the shield so as not to form a completed turn. Otherwise, the loop would be "shielded" from the desired magnetic field also.

Although the efficiency of this system is quite low, the signal-to-noise ratio is usually very good and a loop may be the answer to a receiving problem where local fields cause serious interference. In addition, it is often possible to orient the loop so as to null out the offending noise field thus providing further rejection.

Transmitting-Antenna Efficiency

Since the ambient-noise level is the limiting factor at hf, antenna efficiency in receiving applications is usually secondary to suppressing unwanted external noise as much as possible. This is less true on the higher amateur bands such as 15 and 10 meters with the 20-meter band being more-or-less the crossover point. Antenna gain and efficiency requirements tend to a common limit for both transmitting and receiving starting at these frequencies and continuing up through vhf and above. That is, an inefficient transmitting antenna will likely be deficient for receiving also.

On the other hand, transmitting antenna efficiency is important at hf and below. The obvious reason being that the signal field strength at the receiving antenna is directly proportional to that at the transmitter. Since it is the signal-to-noise ratio that is important, little can be done in the way of efficiency at the receiving end to improve matters.

A circuit similar to the one in the discussion on

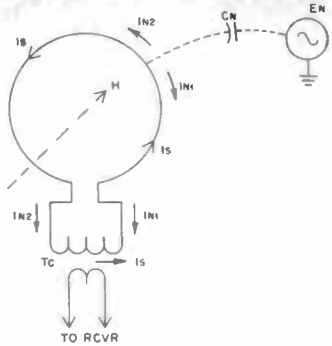


Fig. 21 — Loop antenna and electric-noise rejection.

receiving antennas is shown in Fig. 19B for the transmitting case. T_2 performs a function similar to T_1 in that it matches the antenna and loss resistance to the transmitter along with tuning out undesired reactive components. The attenuation resulting from the loss resistance, R_L, is given by

$$Attn = 10 \log_{10} (1 + R_L/R_a)$$

and it is obvious this resistance should be made as low as possible. The following section deals with some of the factors important in accomplishing this goal.

ANTENNA LOSSES

As pointed out in a previous section, it is the field configuration surrounding the antenna that plays the most important part in radiation rather than the actual conductor geometry. This concept is nowhere as clearly dramatized as in the case of losses associated with an antenna system. Merely considering conditions on the conductor surfaces is not enough. For instance, in discussing antenna designs with some amateurs, the subject of ground systems arises on occasion. A "good ground" is often assumed to be a rod driven five feet or so into the earth. While this might be suitable for some systems, it is entirely unsuitable from a loss standpoint in others. Radials are sometimes used to improve matters but unless they catch in the lawn mower, they are probably buried too deep!

That ground losses are primarily a surface

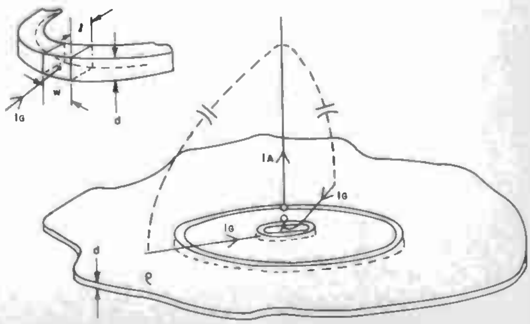


Fig. 22 — Effect on conductivity on ground loss.

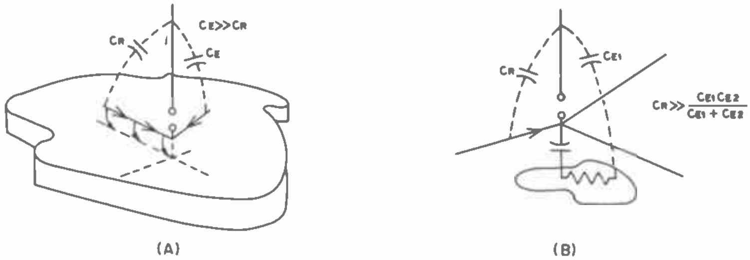


Fig. 23 — Buried radials (A) are less effective than elevated ones (B).

phenomena extending from the antenna can be seen with the aid of Fig. 22. A vertical antenna is placed over a conducting ground screen of thickness d and resistivity ρ. Currents flowing through the vertical wire return to the ground screen via the capacitance of the wire to the screen (shown in dashed lines) and back to the "ground" connection of the antenna.

For the time being, it is also assumed that the current distribution is independent of depth and that the total current flowing through any annular ring is constant regardless of the radius of the ring, at least in the near-field zone of the antenna. Since the current flows inward radially, some idea of the critical areas in regard to ground losses can be determined. The resistance of any sector of a particular ring is just the product of the resistivity, ρ, and length, l, divided by the cross-sectional area. However, it is evident the cross-sectional area must be the product of the width, W, and the thickness, d as can be seen in the inset of Fig. 22. Summarizing, the resistance of any slab or sector of the ring is just

$$R = \rho \frac{l}{A} = \rho \frac{l}{Wd}$$

Extending the idea of a sector to the entire circumference of the ring, the total "width" of the ring is then

$$W_{\text{cir}} = 2\pi r$$

Finally, the resistance of any particular ring is

$$R = \rho \frac{l}{2\pi r d}$$

where r is the average radius of the ring. The total resistance is just the sum of the resistances of the individual rings and it is evident that the ones with the smallest radii are going to contribute the most to this sum.

In order to lower the total resistance, it is evident ρ in the vicinity of the base of the antenna should be made as low as possible. This could be accomplished by using a wide radial metal plate in contact with the ground screen provided the metal had a much lower resistivity than the screen (which might be considered the earth for purposes of discussion here).

Radial Systems

In many instances, it would be impractical to use a solid metal plate so a radial grid of wires is used instead. A question arises as to the best place to install the wires (or plate for that matter) and the answer can be seen with the aid of Fig. 23. In Fig. 23A, the radials are buried beneath the earth. As a consequence, part of the current flows on the surface of the earth and just below it until the radials are reached. This only makes a bad situation worse since the capacitance of the antenna to parts of the earth where there are no radials is high to begin with. Therefore, a buried radial system consisting of a few deeply buried wires may not be very effective.

On the other hand, elevating the radial system decreases the capacitance of the vertical wire to earth and more of the current flows on the radial system. Such an elevated system is called a counterpoise and its effectiveness is even greater than might be expected. This is because the capacitance to earth diminishes very rapidly since there are really two components in series as shown in Fig. 23B. As few as three radial wires may be required in the ground-plane antenna (here the counterpoise consists of a radial system of quarter-wavelength wires) while over a hundred radials are needed in some broadcast installations where the radials are buried.

Dipole "Modes"

Ordinarily, the center-fed dipole is considered to be a balanced system. However, it can also be employed in an unbalanced configuration as shown in Fig. 24. Unfortunately, there are two possible

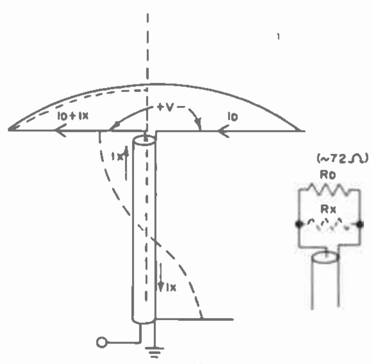


Fig. 24 — Even-mode (I_D) and odd-mode (I_x) excitation of unbalanced dipole.

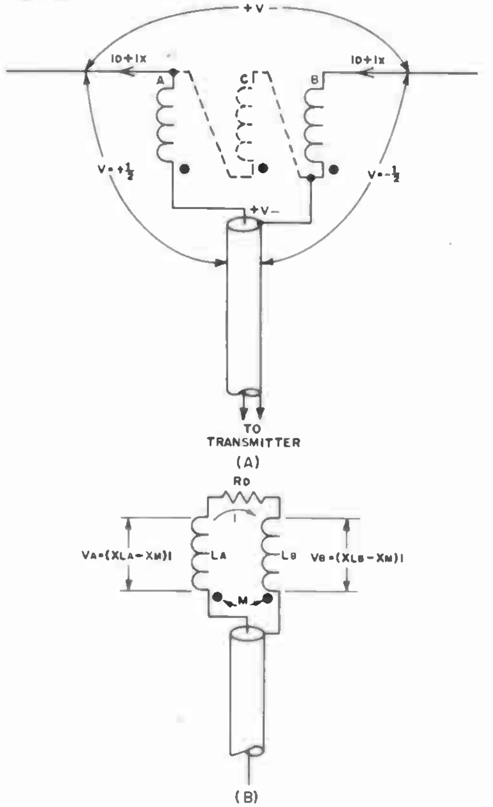


Fig. 25 — Unbalanced-to-balanced transformer. In Fig. 25A, voltages denoted by (1) are for a configuration with the coil *c* absent while voltages prefixed by (2) show the effect of adding coil *c*.

modes of excitation as indicated in the drawing. One mode which might be called the dipole mode (and the desired one) is identical with the mode for the balanced case. The current distribution for the latter is labeled I_D. A second mode which might be thought of as 3/4-wavelength excitation is also possible. The current distribution for this mode is labeled I_x.

Whether or not the latter mode can cause a problem depends mostly on the length of the coaxial line feeding the antenna. For line lengths that are not multiples of a half wavelength, the input impedance of this mode will be high and the load resistance presented to the feed line will be high also. This is represented by R_x in the inset with the resistance of the desired mode designated by R_D. In order to avoid difficulties with modes such as the one in Fig. 24, the antenna is excited in a manner so that the dipole is isolated from the feed line. Two methods for suppressing these modes are shown in Figs. 25 and 26.

In Fig. 25A, two tightly coupled coils are connected in series with the feed line as shown. Assuming that both coils are identical and that unity coupling exists between the windings, the voltage between one side of the dipole and ground is zero, while the voltage to ground on the other side is V. Consequently, the voltage applied across the dipole terminals is the same as that across the output of the feed line so the device is basically a 1:1 transformer. The explanation of this action can be seen with the aid of Fig. 25B.

A current flowing through the load would normally produce a voltage drop across L_a and L_b that would be equal to the product of the current and the respective reactances. However, if the coils are connected so that the mutual-reactance component is negative, the total reactance of each coil is just the difference between the coil and mutual reactance. For unity coupling and identical coils, the latter two reactances are the same and the difference is zero. Consequently, the voltage drop across either coil reactance caused by the current is just cancelled by the induced voltage from the mutual term and the total voltage across either coil is zero.

A third winding indicated by the dashed lines in Fig. 25A is sometimes used but does not enter into the principal operation of the transformer as outlined in Fig. 25B. Such transformers are called baluns and can be constructed with or without the third winding. Another approach is illustrated in Fig. 26. Rather than construct a transformer, a section of transmission line is coiled up near the junction of the feed line and the dipole terminals as shown in Fig. 26A. The effect is to place an inductance in series with the unwanted mode resistance while not affecting the desired mode (Fig. 26B and inset).

In many instances, a balun is not really required and unless symptoms of feed-line radiation exist, will not reduce losses. Also, the design of such transformers that will operate properly over a wide

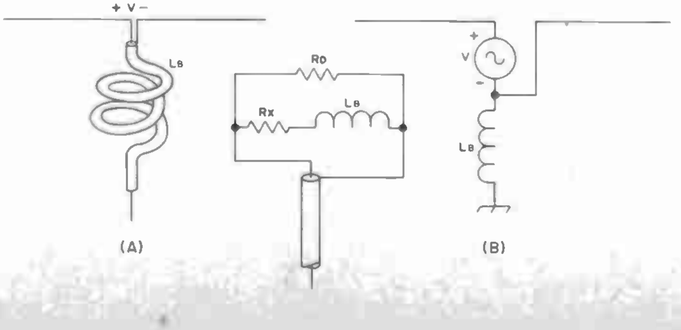


Fig. 26 — Coil balun.

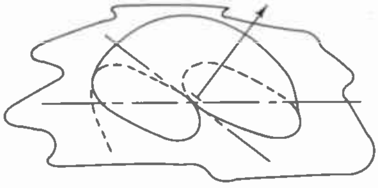


Fig. 27 — Radiation pattern of horizontal dipole over ideal absorber (heights much less than λ/4).

bandwidth is rather difficult. One problem is that the coupled coils also have capacitance which limit the high-frequency response unless the coils are wound so as to form a transmission line with the same characteristic impedance as the feed line and the load. The low-frequency response, particularly if a third winding is used, is limited by the self inductance of each coil. Although ferrite or powdered-iron cores are often used to improve this response, a poorly designed balun may actually make matters worse instead of improving them.

Normally, a balun wouldn't be required with the configuration shown in Fig. 23B since the radials tend to decouple the radiating element from the feed line. On the other hand, the dipole shown in Fig. 24 might be considered as having only one radial so that the possibility of such coupling is greater. This is why the dipole is ordinarily fed with a balanced feed line that is matched to the unbalanced output of the transmitter through a matching network using mutual magnetic coupling exclusively. Such a system eliminates the need for a balun while providing additional rejection above and below the operating frequency.

End-Fed Wire Antennas

In the discussion on ground losses, it was assumed that the total current passing through the perimeter of any annular ring was the same at any radius. This is not true with most antennas and depends upon the length. Vertical wires that are electrically short will have maximum current at the base with diminishing current at larger radii. It is evident the ground screen is of prime importance with such systems since the region near the base is the location where the resistance is greater also. A counterpoise is recommended and this is one reason why some mobile installations seem to perform adequately while the same antenna system would be inefficient if used with a poor ground. The fact that the automobile body is elevated provides some of the earth decoupling effect shown in Fig. 23B.

On the other hand, wire lengths approaching a half wavelength present a very high impedance at the base point and consequently, the input current is very small. This is true for wire antennas that are horizontal rather than vertical and a simple ground system for "cosmetic" purposes is usually sufficient. R_a in Fig. 19B is much greater than R_L so the loss in an end-fed half-wavelength system would be comparable to a center-fed one at the same height.

Radiation vs. "Antenna" Losses

The resistance, R_L, in Fig. 19 represents a loss in the antenna circuit itself but doesn't include any additional loss that might occur at distances great enough so that contributions to R_L would be neligible. The latter might be labeled radiation or "far field" losses while the former would be "near field" or antenna losses. It is interesting to compare the importance of the two losses and one limiting case is illustrated in Fig. 27.

A horizontal dipole is placed over a lossy surface but it is assumed the surface is far enough away so as not to influence R_L (Fig. 19). However, it must be close enough so that the antenna radiation pattern can be considered to be cut in half as shown. It is also assumed that reflections from the surface are negligible. While this wouldn't be true in a case where there was a well-defined boundary between the surface and space, it can be approximated by a surface where a gradual transition takes place. An analogous situation exists with sound propagation. For instance, reflections of propagated sound waves from a depth finder tend to be very weak over a soft oozy bottom but very strong if a hard sandy bottom exists. This is because the gradual nature of the soft bottom absorbs most of the energy while reflection takes place from the hard surface with a sharp boundary.

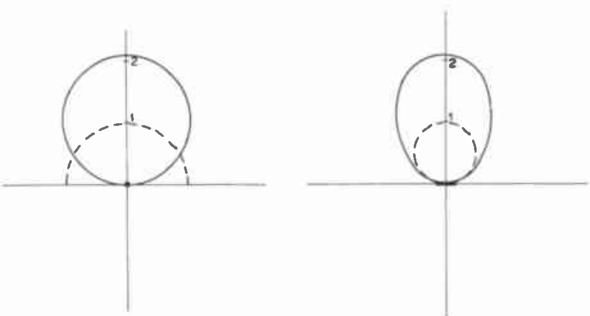


Fig. 28 — Radiation pattern of horizontal dipole over perfectly conducting surface (solid lines) at low height. Dashed curve is for same antenna over ideal absorber. (The view at the left is in the direction of the dipole while at the right, a side view is shown.)

Returning to the antenna shown in Fig. 27, all the radiated power directed below the horizon would be absorbed while power emitted at higher elevation angles is propagated into space. Consequently, exactly one-half the radiated power is absorbed. If, instead of placing the antenna over a lossy surface, it was over a perfectly reflecting plane the absorbed power would be reflected and propagated. However, it would be a mistake to assume that the power radiated in a given direction as shown in Fig. 27 by the arrow would merely be increased by 3 dB. Because of the reflections, little power would be propagated at the lower elevation angles and the actual pattern is illustrated in Fig. 28. The latter figure is for a horizontal antenna very close to the conducting plane. For radiation patterns of a dipole at other heights along with height multiplication patterns, see *The ARRL Antenna Book*. As the height of the antenna is increased, power radiated along the conducting surface is still zero but more power is propagated at the lower elevation heights.

An interesting feature of a "low" horizontal dipole is that it exhibits almost an omnidirectional pattern as shown in Fig. 29. Consequently, the orientation of such a system is not as critical as many amateurs might assume. Unless the dipole is at a height approaching a half wavelength, there is little advantage to be gained by orienting the antenna so that radiation is in some optimum direction. Other factors such as proximity to conducting objects would be more important in the installation.

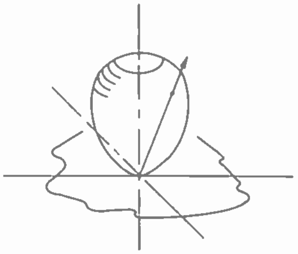


Fig. 29 — A "low" dipole exhibits an omnidirectional pattern and antenna orientation is not important normally.

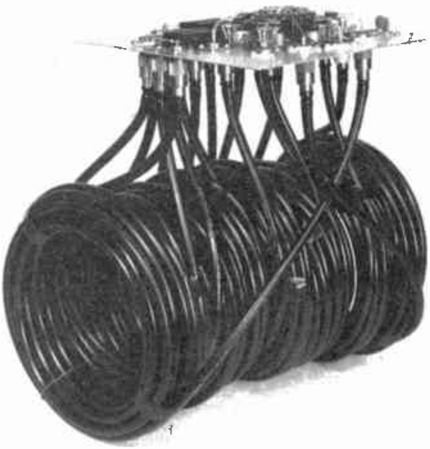
Since the earth is neither a perfect reflector nor absorber, the patterns of Figs. 27 through 29 represent theoretical limits. A problem often arising in amateur installations is that space limitations usually dictate a particular antenna configuration. For instance, it may be possible to erect a vertical antenna but with no provisions for radials. A horizontal dipole might also be possible but at a low height. If the primary objective was low-angle radiation, one might conclude that the vertical antenna would be the obvious choice. However, if the soil conductivity was low, the losses incurred because of the inadequate ground might offset the poor low-angle radiation pattern of a horizontal dipole. Consequently, some experimentation would be required in order to find the optimum system.

A 360° STEERABLE VERTICAL PHASE ARRAY FOR 7 MHz

The original design information for the array presented here appeared in *QST* for April, 1976 in an article written by Atchley, Stinehelfer and White. Featured in that article was a system designed for use in the 3.5 to 4.0 MHz range.

The configuration shown here makes use of four quarter-wavelength vertical elements in a square, with quarter-wave spacing between adjacent elements, as shown, with its predicted pattern, in Fig. 30. All elements are fed with equal amplitudes, the rear element at 0°, the two side elements at −90° and the lead element at −180°. The beam is transmitted along the diagonal from the rear to the lead element. Gain due to horizontal beam formation alone is approximately 5.3 dB over a single vertical element. Front-to-back ratio is on the order of 25 dB. Front-to-side ratio is 12 dB at 90° either side increasing to much higher levels at 135° either side. Since most of the vertical energy is concentrated at low angles by the array, as much as 4 dB of gain additional to the predicted 5.3 dB in the horizontal plane can be achieved in theory, with perfectly conducting ground. With a good radial system and less-than-perfect ground one can expect an approximate additional gain of 2 dB, or a total gain just over 7 dB for the system described.

A computer program calculation predicts the half-power beamwidth to be 97°. A suitable switching matrix is used to direct the beam pattern to four different quadrants, with only a slight loss



of forward gain at the cross-over points, and virtually no deterioration of the front-to-back and front-to-side suppression.

RF Power Dividers

Good power splitters are essential to the operation of phased arrays. The system described here

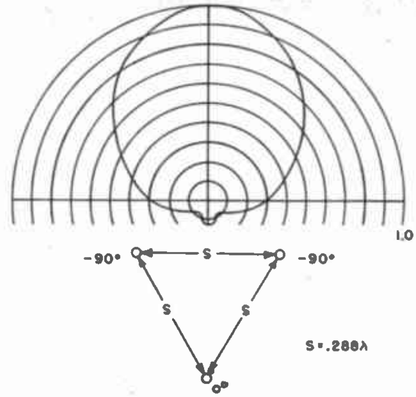


Fig. 30 — Polar plot of relative power, and planar view of the 4-element diamond array showing the pattern obtained with no dc voltage on the switching relays, as in Fig. 31. Minor lobes are too far down to show on this scale.

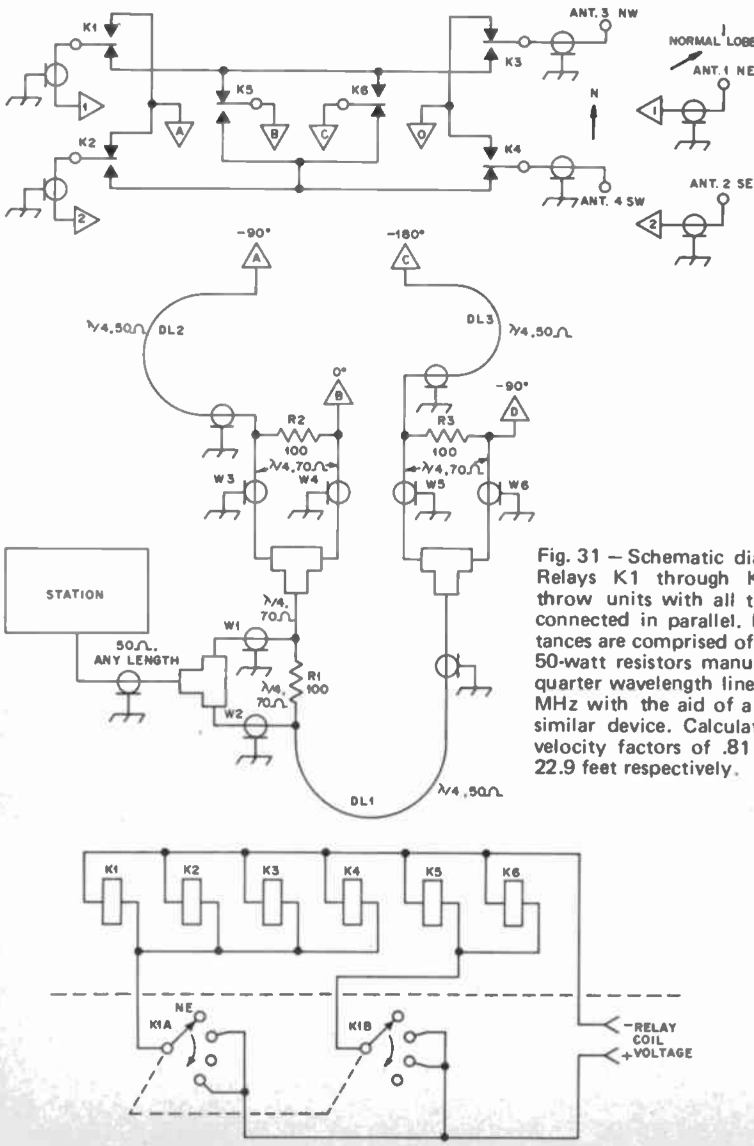
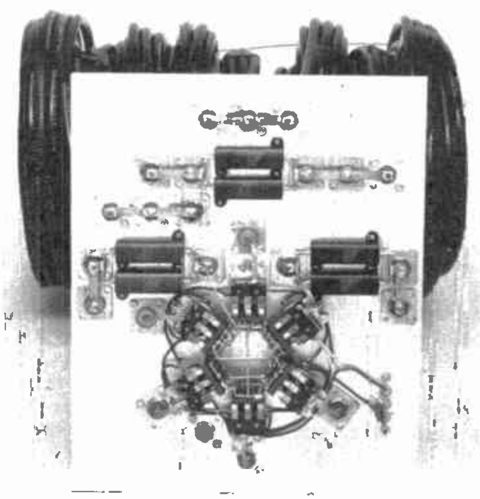


Fig. 31 — Schematic diagram of the phased array. Relays K1 through K6 are three-pole, double-throw units with all three sections of each relay connected in parallel. Each of the 100-ohm resistances are comprised of two 200-ohm noninductive 50-watt resistors manufactured by Nytronics. The quarter wavelength lines should be adjusted for 7.1 MHz with the aid of a noise bridge, RX bridge or similar device. Calculated lengths of line for the velocity factors of .81 and .66 are 28.1 feet and 22.9 feet respectively.



makes use of 2-way Wilkinson Power Dividers.[1] Those readers versed in microwave technology are no doubt familiar with these devices. As shown in Fig. 31, power from the transmitter is fed through 50-ohm line of any length to a T connection, feeding two quarter-wavelength lines, W1 and W2. The two inner conductors of the 70-ohm lines are connected through a 100-ohm noninductive resistance, R1. This type of divider gives an equal power split, matches to 50-ohm loads at the two outputs, and has the unique property that any energy returning to the two outputs out of phase, due to mismatches or mutuals, is absorbed in the 100-ohm resistor.

The 90-degree phase-delay cable (DL) used in one side of each of the three power-splitting hybrids serves to assure that equal power reflections from the antennas are absorbed. Theoretically the resistors absorb none of the forward power. This technique provides approximately 30 dB of isolation from one output terminal to the other, to unwanted energy. The Wilkinson Power Divider, when used to feed phased arrays, reduces the problems associated with element interaction due to mutual impedances between elements.

FEEDING, SWITCHING AND PHASING

The 4-element system shown here uses three Wilkinson 2-way power dividers, as shown in Fig. 31. Proper element phasing is accomplished with three 90-degree sections of RG-8/U cable, DL1, DL2, DL3. The four outputs are at 0°, -90°, -90° and -180° respectively in phase relationship, with power from the transmitter divided into four equal parts. Special attention should be paid to preserving symmetry throughout the system, in the hybrids, phase shifters, rf switching, feeds and antenna placement, in order to have the array perform uniformly as it is switched between the four headings.

The switching was done with six relays as shown in Fig. 31 and the accompanying photograph. With no voltage applied to the relay coils, the arms are in the positions shown in Fig. 31 giving northeasterly directivity – a convenient heading for operation from New England. The three Wilkinson power divider resistor networks along with the switching relays are mounted on a piece of sheet aluminum measuring 10 X 12 inches. Each of the three phasing lines and quarter-wavelength lines associated with the power dividers connect to coaxial fitting also mounted to the piece of sheet aluminum. The entire phasing/switching system can be transported or worked on as a single unit. Basic layout of this system is shown in the photograph and also in Fig. 33. Measured phase errors due to unequal lead lengths in the switching system are less than 2 degrees. The builder need not follow the layout shown here – generally speaking, if the leads are kept short and direct the system should function according to specifications listed earlier.

The relays used in this system were surplus 3 pole double throw units. Each of the three sections were connected in parallel to assure that the relays would handle the transmitter power. A three-wire control system is used to steer the array. Four different commands are required for four different directions as shown in the relay truth table in Fig. 31.

An aluminum chassis measuring 10 X 12 X 3 inches placed over the piece of aluminum on the resistor/relay side protects these components from the weather. Silicon seal is used to fill any holes or slots in the chassis.

Radiators and Radials

Relays K1 through K4 (Fig. 31) connect to their respective radiators through equal lengths of 50-ohm line. The radiators are constructed from 6061T6 aluminum tubing which is available in 12 foot lengths. Three pieces are required for each radiator (1 – 1-1/4, 1 – 1-1/8, 1 – 1 inch diameter). Constructional details are shown in Fig. 32. Base insulators are made from schedule 40 PVC pipe measuring 1-3/8 diameter. The 1-1/4 inch aluminum tubing does not fit securely inside the PVC pipe necessitating the use of shims between the aluminum tubing and PVC. One simple way of solving this problem is to cut thin strips (1/2 inch or so) of sheet aluminum and wind them over the aluminum tubing, sliding both the tubing and shim material into the PVC pipe. See Fig. 32 for details.

The radiators are supported at the base by 4-foot long pieces of 1-inch galvanized water pipe driven approximately 3 feet into the ground. Care should be taken to ensure that the pipe is kept true vertical when inserted into the ground. Although the radiators are self supporting, it is highly recommended that at least one set of guy wires be used for each radiator. Moderate winds have no difficulty bending the radiators if guys are not used. With the system described here, one set of guys, located at the 24 foot point are used. Heavy

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[1]Wilkinson, "An N-Way Hybrid Power Divider," IRE *Transactions on Microwave Theory and Techniques*, January, 1960.

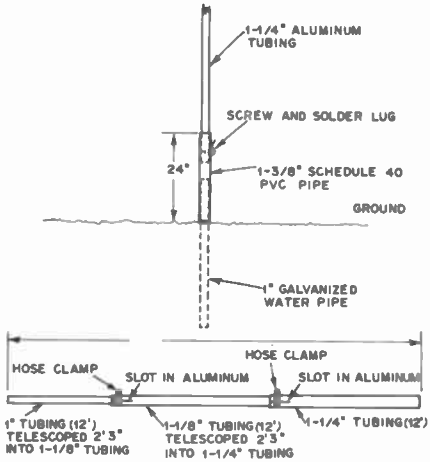

Fig. 32 – Details of element construction and mounting method. The water pipe should be driven into the ground so that approximately 1 foot of the pipe extends above ground level.

nylon cord was used for this purpose.

Though 120 radials are considered to be the optimum number, only 40 per element are used with this system. All are 35 feet long, No. 15 aluminum fence wire, lying directly on the ground. The radials at the center of the array, in addition to providing considerable symmetry of the mutuals, allows a higher packing density for the radials, reducing ground losses.

TESTING THE SYSTEM

Before power is applied to the array a quick resistance check should be made. One person should go to each element and short the input, while another person watches an ohmmeter placed across the main line at the station end. Make sure that a very low dc resistance is measured. Then, with the array in the normal position, that shown in the schematic, make sure that the resistance across the input is high. If it is low, check for moisture in the cables or connectors or for other leakage resistance. Silicone grease in the connectors is a good moisture preventive measure. It is recommended that these resistance checks be repeated periodically to be sure that all is well.

When the array is ready for use, go easy at first, as any reflected energy will be dissipated in the 100-ohm resistances. If they become hot, there is either a problem with the system or better matching of the elements will be necessary. Each 100-ohm resistance is made up of two Sage 200-ohm, 50-watt resistors connected in parallel.

USING THE SYSTEM

When receiving in the "search" mode, one hand tunes the receiver while the other operates the lobe selector switch, to see which position "listens" best. Large rotary arrays typically take 45 to 60 seconds to rotate 360°, which tends to discourage frequent directional checks. With the phased array a complete scan takes but a few seconds. The high front and side rejection eliminates most of the interference from signals in unwanted directions, and in transmitting the clean pattern helps to prevent interference to stations off the lobe of the beam that might be on the same frequency.

One useful by-product of this system has been the reduced atmospheric noise pickup from unwanted directions. In particular, when listening toward Europe atmospheric noise coming from electrical storms in the southwest is greatly reduced, improving the signal-to-noise ratio on signals arriving from the favored direction.

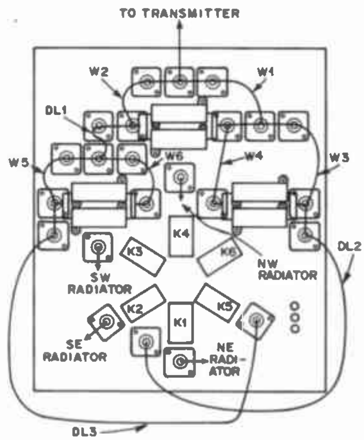

Fig. 33 – This drawing shows where each of the phasing and delay lines are connected to the divider/switching system.

LIMITED-SPACE ANTENNAS

Many amateurs with space restrictions often feel operation on the lower amateur bands is impractical for them. However, a number of techniques are available for employing radiators that are electrically short. In fact, the only technical disadvantages to small-sized antennas are that the bandwidth becomes narrow because of the increasing Q and losses impose limitations on efficiency since the radiation resistance is small. (See the discussion on radiation resistance in a

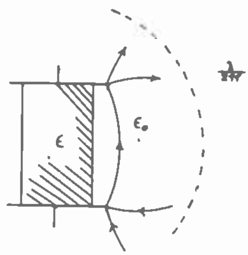

Fig. 34 – Increasing the spacing and plate size of the capacitor of Fig. 5 improves its radiating properties while adding a dielectric would increase the Q.

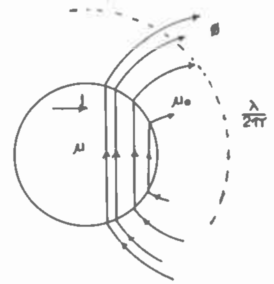

Fig. 35 – A coil wound on a high-permeability core with a shape that causes the flux lines to extend from the surface approaches the ideal electrically small radiator.

previous section in this chapter.)

On the other hand, making an antenna much longer than necessary often produces little if any advantages while actually introducing undesirable factors. For instance, a long antenna run near power lines or nearby dwellings is likely to be prone to noise on receiving and poor from an RFI standpoint. So even when space restrictions don't exist, there are occasions when electrically short antennas may still be superior to larger versions.

Bandwidth

Increasingly narrow bandwidth with decreasing electrical length is a physical reality that has to be dealt with in small-size antenna considerations. Considerable confusion sometimes exists in this area. If an electrically short antenna *doesn't* have a narrow bandwidth, something is wrong! There have been various attempts to "improve" the bandwidth of electrically short radiators but most are of a questionable nature. (One system actually used resistive loading and the bandwidth did *indeed* improve. But the limiting case of course is to connect the transmitter to a dummy load!)

Bandwidth is related to circuit Q and in order to increase the bandwidth, the Q of the antenna must be reduced. A general definition of Q is

$$Q = 2\pi \frac{\text{(Energy stored per rf cycle)}}{\text{(Energy lost per rf cycle)}}$$

and either the radiation resistance must be increased (which means the energy lost per cycle will increase) or the energy stored per cycle must be decreased. The problem then, is to find an optimum configuration where the field representing energy storage is a minimum while components that could contribute to radiation are at a maximum.

As with so many problems of a similar nature, the goal is simple to state but finding a solution is considerably more elusive. However, one way *not* to improve matters is illustrated in Fig. 34. A parallel-plate capacitor similar to the one shown in Fig. 5 is to be used as an antenna except that a dielectric with permittivity ϵ is placed between the plates. This would increase the total capacitance of the system which is desirable since the value of loading inductance required to resonate the circuit would be decreased. However, the dielectric would also increase the energy stored per rf cycle for the following reason.

The energy stored in the electric field of a capacitor is

$$W = \frac{1}{2} CV^2$$

It will be recalled from an earlier discussion that the fringing field was the component that contributed to radiation. However, the capacitance representing the fringing field is in parallel with the capacitance formed by the plates themselves. The applied voltage will then be the same for both. By placing the dielectric between the plates, this capacitance increases (which doesn't contribute to radiation) and the energy stored per rf cycle is increased. Since the fringing-field capacitance stays the same, so does the radiation resistance and antenna Q is increased.

The capacitor arrangement of Fig. 34 can be considered as an electric dipole of sorts. Dielectric loading increased the Q by concentrating stored energy in the interior regions of the antenna. The goal then, is to exclude as much energy as possible from such areas so that a greater percentage "couples" to the fringing field. One configuration that accomplishes this goal is illustrated in Fig. 35. A coil of wire is wound on a spherical core of permeability μ. Unlike a toroid, where most of the field is confined to the core itself, at least part of the path for each flux line is in air for the spherical form.

Energy stored in a magnetic field is given by

$$W = \frac{1}{2} \frac{B^2}{\mu}$$

where μ is the permeability and B is the magnetic flux density. The flux density is the ratio of total flux, ϕ, passing through a given area. Since ϕ

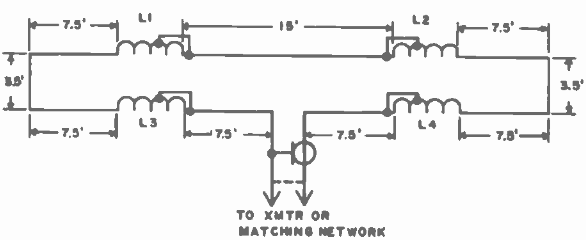

Fig. 36 — Dipole dimensions and coil data.
L1-L4, incl. — 82 μH for 3.86 MHz. Air wound preferable, 57 turns, 2-1/2-inch dia., 10 tpi of No. 16 solid wire (B&W 3031).

(denoted by the solid lines with arrows) is the same both inside and outside of the core, B must be the same also. Consequently, very little energy will be stored in the core if the ratio of μ/μ_o is large. This is because the ratio of B^2/μ becomes small as μ is increased. Therefore, the configuration of Fig. 35 is a very desirable one since most of the magnetic energy is stored in the fringing field.

The Radiansphere[1]

The magnetic and electric fields around an electrically small dipole at distances large compared to its dimensions are given by

$$H(\text{mag}) \sim \frac{1}{r^2}\left[j\left(\frac{2\pi}{\lambda}\right)r + 1\right]$$

$$E(\text{elec}) \sim \frac{1}{r^2}\left[j\left(\frac{2\pi}{\lambda}\right)r + \frac{1}{j\,\frac{2\pi}{\lambda}\,r} + 1\right]$$

Note that when the distance, r, is $\lambda/2\pi$, all the terms are equal. At much greater distances, the two proportional equations reduce to

$$H \sim \frac{1}{r}$$

$$E \sim \frac{1}{r}$$

$$r > \frac{\lambda}{2\pi}$$

which represent the "far" field or radiation field. The distance $\lambda/2\pi$ can be considered a transition zone between the exterior region where all terms represent power radiated and internal areas where terms representing energy storage are important also.

From a geometrical point of view, the distance $\lambda/2\pi$ would be a sphere surrounding the antenna as indicated by the dashed lines (Figs. 34 and 35). This sphere is defined as the radiansphere and is an important concept in the design of electrically small antennas. First, it provides a greater insight as to why attempting to store as much energy in the fringing fields is desirable. A greater portion of the field is then exterior to the radiansphere which in turn implies the radiation fields are increased.

Because of reciprocity, a similar process occurs on receiving. Perhaps the most familiar example of the application of this principle is the ferrite-loop antennas common in bc sets and direction finders. Like the sphere of Fig. 35, the ferrite core reduces the energy within the interior of the coil so that a greater fraction is contained in the fringing fields.

A Loaded Folded Dipole

Because of the low-valued radiation resistance and high Q of small antennas, formidable matching problems often exist. The reactive component must be tuned out and the radiation resistance matched to a feed line. In accordance with the principles of the preceding sections, an optimum system would be one that performs the matching and tuning functions while lowering the antenna Q at the same time.

One method that accomplishes these goals in part is called multiple tuning. A common example of multiple tuning is the folded dipole. Two half-wavelength radiators are paralleled and excited so that their far fields are in phase. Since the field produced by both radiators is doubled, the power density increases by a factor of 4. This means the power input to the antenna increases by a factor of 4 also. However, since only one radiator is being fed, the equivalent input resistance increases by a factor of 4 in order to satisfy the power increase. For instance, a folded dipole consisting of two 75-Ω radiators will have an input impedance of 300 Ω.

The principle can be applied to loaded antennas which is particularly advantageous because of the small radiation resistance. Instead of a matching transformer, the resistance is transformed in the antenna itself. It can then either be connected directly to the feed line or through some intermediate transformation if the value is still too low. Also, adding more elements will increase the resistance even more. For instance, a 3-radiator version would increase the resistance by a factor of 9 (or n^2 where n is the number of elements).

An experimental model is shown in Fig. 36. In spite of its size, the performance compared favorably with much larger antennas in on-the-air tests. However, bandwidth was quite narrow. The "window" centered around 3.86 MHz where the *VSWR* was less than 2:1 was approximately 20 kHz indicating a Q of 190. It is interesting to compute the radiansphere for 3.86 MHz which is approximately 41 feet. In this regard, the antenna of Fig. 36 is not that small. But considered in terms of a half-wavelength dipole, it is only 0.12 wavelengths long. In light of the latter fact, the narrow bandwidth is not surprising. However, it would also appear that a substantial reduction in Q might result if capacitive end loading was employed to extend the fringing field.

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[1] H.A. Wheeler, "The Radiansphere Around a Small Antenna," *Proceedings of the IRE* for August, 1959 (vol. 47 pp. 1325-1331).

DIRECTIVE ARRAYS WITH PARASITIC ELEMENTS

With few exceptions, the antennas described so far in Chapter 21 have unity gain or less, and are either omnidirectional or bidirectional. In order for antennas to have gain and take on directional characteristics they must employ additional elements. Antennas with these properties are commonly referred to as "beam" antennas. This section will deal with the design and characteristics of directional antennas with gain.

Parasitic Excitation

In most of these arrangements the additional elements receive power by induction or radiation from the driven element generally called the "antenna," and reradiate it in the proper phase relationship to achieve the desired effect. These elements are called *parasitic* elements, as contrasted to the driven elements which receive power directly from the transmitter through the transmission line.

The parasitic element is called a director when it reinforces radiation on a line pointing to it from the antenna, and a reflector when the reverse is the case. Whether the parasitic element is a director or reflector depends upon the parasitic-element tuning, which usually is adjusted by changing its length.

Gain vs. Spacing

The gain of an antenna with parasitic elements varies with the spacing and tuning of the elements and thus for any given spacing there is a tuning condition that will give maximum gain at this spacing. The maximum front-to-back ratio seldom, if ever, occurs at the same condition that gives maximum forward gain. The impedance of the driven element also varies with the tuning and spacing, and thus the antenna system must be tuned to its final condition before the match between the line and the antenna can be completed. However, the tuning and matching may interlock to some extent, and it is usually necessary to run through the adjustments several times to insure that the best possible tuning has been obtained.

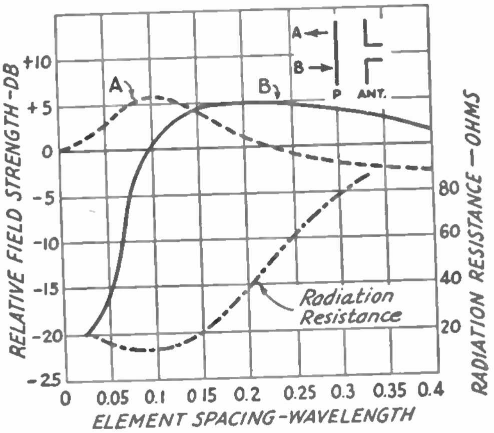


Fig. 21-36 -Gain vs. element spacing for an antenna and one parasitic element. The reference point, 0 dB, is the field strength from a half-wave antenna alone. The greatest gain is in the direction A at spacings of less than 0.14 wavelength, and in direction B at greater spacings. The front-to-back ratio is the difference in dB between curves A and B. Variation in radiation resistance of the driven element is also shown. These curves are for a self-resonant parasitic element. At most spacings the gain as a reflector can be increased by slight lengthening of the parasitic element; the gain as a director can be increased by shortening. This also improves the front-to-back ratio.

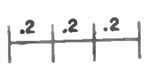
Two-Element Beams

A 2-element beam is useful where space or other considerations prevent the use of the larger structure required for a 3-element beam. The general practice is to tune the parasitic element as a reflector and space it about 0.15 wavelength from the driven element, although some successful antennas have been built with 0.1-wavelength spacing and director tuning. Gain *vs.* element spacing for a 2-element antenna is given in Fig. 21-36, for the special case where the parasitic element is resonant. It is indicative of the performance to be expected under maximum-gain tuning conditions.

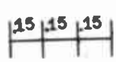
TABLE 21-II

Freq.	Driven Element A	Driven Element B	Reflector A	Reflector B	1st Director A	1st Director B	2nd Director A	2nd Director B
14050	33' 5 3/8"	33' 8"	35' 2 1/2"	35' 5 1/4"	31' 9 3/8"	31' 11 5/8"	31' 1 1/4"	31' 3 5/8"
14250	32' 11 3/4"	33' 2 1/4"	34' 8 1/2"	34' 11 1/4"	31' 4"	31' 6 3/8"	30' 8"	30' 10 1/2"
21050	22' 4"	22' 5 5/8"	23' 6"	23' 7 3/4"	21' 2 1/2"	21' 4"	20' 9 1/8"	20' 10 7/8"
21300	22' 3/4"	22' 2 3/8"	23' 2 5/8"	23' 4 1/2"	20' 11 1/2"	21' 1"	20' 6 1/4"	20' 7 3/4"
28050	16' 9"	16' 10 1/4"	17' 7 5/8"	17' 8 7/8"	15' 11"	16'	15' 7"	15' 9 1/2"
28600	16' 5 1/4"	16' 6 3/8"	17' 3 1/2"	17' 4 3/4"	15' 7 1/4"	15' 8 1/2"	15' 3 3/8"	15' 4 1/2"

A



B



Element lengths for 20, 15 and 10 meters, phone and cw. These lengths are for 0.2 or .15 wavelength element spacing.

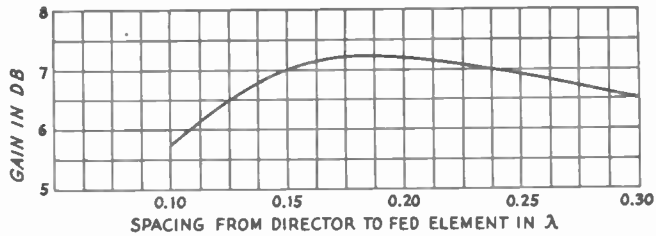


Fig. 21-37 – Gain of 3-element Yagi versus director spacing, the reflector spacing being fixed at 0.2 wavelength.

Three-Element Beams

A theoretical investigation of the 3-element case (director, driven element and reflector) has indicated a maximum gain of slightly more than 7 dB. A number of experimental investigations have shown that the optimum spacing between the driven element and reflector is in the region of 0.15 to 0.25 wavelength, with 0.2 wavelength representing probably the best overall choice. With 0.2-wavelength reflector spacing, Fig. 21-37 shows the gain variation with director spacing. It is obvious that the director spacing is not especially critical, and that the overall length of the array (boom length in the case of a rotatable antenna) can be anywhere between 0.35 and 0.45 wavelength with no appreciable difference in gain.

Wide spacing of both elements is desirable not only because it results in high gain but also because adjustment of tuning or element length is less critical and the input resistance of the driven element is higher than with close spacing. The latter feature improves the efficiency of the antenna and makes a greater bandwidth possible. However, a total antenna length, director to reflector, of more than 0.3 wavelength at frequencies of the order of 14 MHz introduces considerable difficulty from a constructional standpoint, so lengths of 0.25 to 0.3 wavelength are frequently used for this band, even though they are less than optimum.

In general, the gain of the antenna drops off less rapidly when the reflector length is increased beyond the optimum value than it does for a corresponding decrease below the optimum value. The opposite is true of a director. It is therefore advisable to err, if necessary, on the long side for a reflector and on the short side for a director. This also tends to make the antenna performance less dependent on the exact frequency at which it is operated, because an increase above the design frequency has the same effect as increasing the length of both parasitic elements, while a decrease in frequency has the same effect as shortening both elements. By making the director slightly short and the reflector slightly long, there will be a greater spread between the upper and lower frequencies at which the gain starts to show a rapid decrease.

When the over-all length has been decided upon, the element lengths can be found by referring to Fig. 21-38. The lengths determined by these charts will vary slightly in actual practice with the element diameter and the method of supporting the elements, and the tuning of a beam should always be checked after installation. However, the lengths obtained by the use of the charts will be close to correct in practically all cases, and they can be used without checking if the beam is difficult of access.

In order to make it even easier for the Yagi builder, Table 21-II can be used to determine the element lengths needed. Both cw and phone lengths are included for the three bands, 20, 15, and 10 meters. The 0.2 wavelength spacing will provide greater bandwidth than the 0.15 spacing. Antenna gain is essentially the same with either spacing. The element lengths given will be the same whether the beam has 2, 3 or 4 elements. It is recommended that "Plumber's Delight" type construction be used where all the elements are

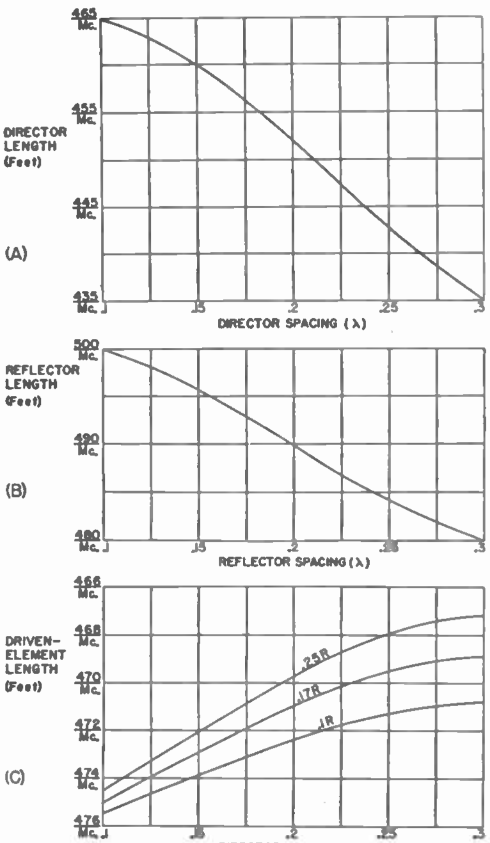


Fig. 21-38 – Element lengths for a 3-element beam. These lengths will hold closely for tubing elements supported at or near the center.

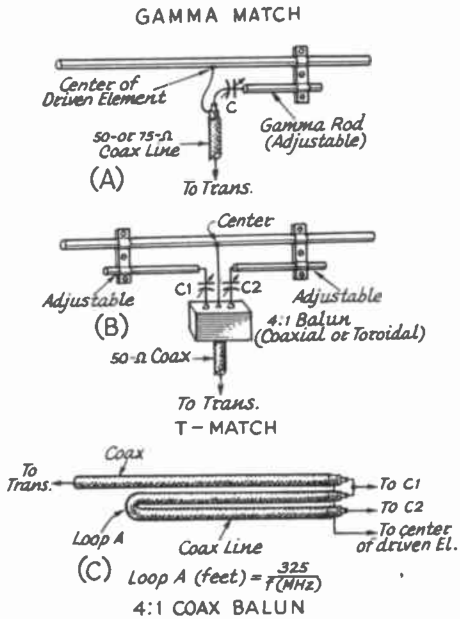


Fig. 21-39 — Illustrations of gamma and T-matching systems. At A, the gamma rod is adjusted along with C until the lowest possible SWR is obtained. A T-match is shown at B. It is the same as two gamma-match rods. The rods and C1 and C2 are alternately adjusted for a 1:1 SWR. A coaxial 4:1 balun transformer is shown at C. A toroidal balun can be used in place of the coax model shown. Details for the toroidal version are given in Chapter 20, and it has a broader frequency range than the coaxial version. The T-match is adjusted for 200 ohms and the balun steps this *balanced* value down to 50 ohms, *unbalanced*. Or, the T-match can be set for 300 ohms, and the balun used to step this down to 75 ohms, unbalanced. Dimensions for the gamma and T-match rods cannot be given by formula. Their lengths and spacing will depend upon the tubing size used, and the spacing of the parasitic elements of the beam. Capacitors C, C1 and C2 can be 140 pF for 14-MHz beams. Somewhat less capacitance will be needed at 21 and 28 MHz.

mounted directly on and grounded to the boom. This puts the entire array at dc ground potential, affording better lightning protection. A gamma section can be used for matching the feed line to the array.

Tuning Adjustments

The preferable method for checking the beam is by means of a field-strength meter or the S meter of a communications receiver, used in conjunction with a dipole antenna located at least 10 wavelengths away and as high as or higher than the beam that is being checked. A few watts of power fed into the antenna will give a useful signal at the observation point, and the power input to the transmitter (and hence the antenna) should be held constant for all of the readings.

Preliminary matching adjustments can be done on the ground. The beam should be set up so that the reflector element rests on earth with the remaining elements in a vertical configuration. In other words, the beam should be aimed straight up. The matching system is then adjusted for 1:1 SWR between the feed line and driven element. When the antenna is raised into its operating height, only slight touch-up of the matching network will be required.

A great deal has been printed about the need for tuning the elements of a Yagi-type beam. However, experience has shown that lengths given in Fig. 21-38 and Table II are close enough to the desired length that no further tuning should be required. This is true for Yagi arrays made from metal tubing. However, in the case of quad antennas, made from wire, the reflectors and directors *should* be tuned with the antenna in its operating location. The reason is that it is practically impossible to cut and install wire to the *exact* dimensions required for maximum gain or front-to-back.

Simple Systems: The Rotary Beam

Two- and three-element systems are popular for rotary-beam antennas, where the entire antenna system is rotated, to permit its gain and directivity to be utilized for any compass direction. They may be mounted either horizontally (with the plane containing the elements parallel to the earth) or vertically.

A four-element beam will give still more gain than a three-element one, provided the support is sufficient for about 0.2 wavelength spacing between elements. The tuning for maximum gain involves many variables, and complete gain and tuning data are not available.

The elements in close-spaced (less than one-quarter wavelength element spacing) arrays preferably should be made of tubing of one-half to one-inch diameter. A conductor of large diameter not only has less ohmic resistance but also has lower Q; both these factors are important in close-spaced arrays because the impedance of the driven element usually is quite low compared to that of a simple dipole antenna. With three- and four-element close-spaced arrays the radiation resistance of the driven element may be so low that ohmic losses in the conductor can consume an appreciable fraction of the power.

Feeding the Rotary Beam

Any of the usual methods of feed (described later under "Matching the Antenna to the Line") can be applied to the driven element of a rotary beam. The popular choices for feeding a beam are the gamma match with series capacitor and the T match with series capacitors and a half-wavelength phasing section, as shown in Fig. 21-39. These methods are preferred over any others because they permit adjustment of the matching and the use of coaxial line feed. The variable capacitors can be housed in small plastic cups for weatherproofing; receiving types with close spacing can be used at powers up to a few hundred watts. Maximum

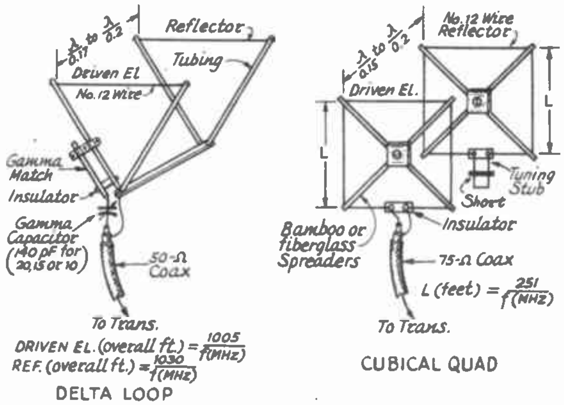


Fig. 21-40 – Information on building a quad or a Delta-Loop antenna. The antennas are electrically similar, but the Delta-Loop uses "plumber's delight" construction. Additional information is given in the text.

capacitance required is usually 140 pF at 14 MHz and proportionately less at the higher frequencies.

If physcially possible, it is better to adjust the matching device after the antenna has been installed at its ultimate height, since a match made with the antenna near the ground may not hold for the same antenna in the air.

Sharpness of Resonance

Peak performance of a multielement parasitic array depends upon proper phasing or tuning of the elements, which can be exact for one frequency only. In the case of close-spaced arrays, which because of the low radiation resistance usually are quite sharp-tuning, the frequency range over which optimum results can be secured is only of the order of 1 or 2 percent of the resonant frequency, or up to about 500 kHz at 28 MHz. However, the antenna can be made to work satisfactorily over a wider frequency range by adjusting the director or directors to give maximum gain at the *highest* frequency to be covered, and by adjusting the reflector to give optimum gain at the *lowest* frequency. This sacrifices some gain at all frequencies, but maintains more uniform gain over a wider frequency range.

The use of large-diameter conductors will broaden the response curve of an array because the larger diameter lowers the *Q*. This causes the reactances of the elements to change rather slowly with frequency, with the result that the tuning stays near the optimum over a considerably wider frequency range than is the case with wire conductors.

Combination Arrays

It is possible to combine parasitic elements with driven elements to form arrays composed of collinear driven and parasitic elements and combination broad-side-collinear-parasitic elements. Thus two or more collinear elements might be provided with a collinear reflector or director set, one parasitic element to each driven element. Or both directors and reflectors might be used. A broadside-collinear array can be treated in the same fashion.

DELTA LOOPS AND QUAD BEAMS

One of the more effective DX arrays is called the "cubical quad" or, simply, "quad" antenna. It consists of two or more square loops of wire supported by a bamboo or fiberglass cross-arm assembly. The loops are a quarter wavelength per side (full wavelength overall) one loop being driven, and the other serving as a parasitic element – usually a reflector. A variation of the quad is called the Delta Loop. The electrical properties of both antennas are the same, generally speaking, though some operators report better DX results with the Delta Loop. Both antennas are shown in Fig. 21-40. They differ mainly in their physical properties, one being of "Plumber's Delight" construction, while the other uses insulating support members. One or more directors can be added to either antenna if additional gain and directivity is desired, though most operators use the two-element arrangement.

It is possible to interlace quads or "deltas" for two or more bands, but if this is done the formulas given in Fig. 21-40 may have to be changed slightly to compensate for the proximity effect of the second antenna. For quads the length of the full-wave loop can be computed from

$$\textit{Full-wave loop}\ (\text{ft}) = \frac{1005}{f\,(\text{MHz})} \qquad \text{21-H}$$

If multiple arrays are used, each antenna should be tuned up separately for maximum forward gain as noted on a field-strength meter. The reflector stub on the quad should be adjusted for the foregoing condition. The Delta-Loop gamma match should be adjusted for a 1:1 SWR. No reflector tuning is needed. The Delta-Loop antenna has a broader frequency response than the quad, and holds at an SWR of 1.5:1 or better across the band it is cut for.

TABLE 21-III

Quantity	*Length (ft.)*	*Diameter (in.)*	*Reynolds No.*
2	8	1	9A
4	8	3/4	8A
1	8	1 1/4	10A
1	6	7/8	4231

2 U-bolts, TV antenna to mast type, 1 variable capacitor, 150 pF maximum, any type, 1 plastic freezer container, approximately 5 X 5 X 5 inches, to house gamma capacitor.
Gamma rod, 3/8- to 1/2-inch diameter aluminum tubing, 36 inches long. (Aluminum curtain rod or similar.)

The resonance of the quad antenna can be found by checking the frequency at which the lowest SWR occurs. The element length (driven element) can be adjusted for resonance in the most-used portion of the band by lengthening or shortening it.

It is believed that a two-element quad or Delta-Loop antenna compares favorably with a three-element Yagi array in terms of gain (see *QST*, May, 1963, and *QST*, January 1969 for additional information). The quad and Delta-Loop antennas perform very well at 50 and 144 MHz. A discussion of radiation patterns and gain, quads vs. Yagis, was presented by Lindsay in *QST*, May, 1968.

A SHORT 20-METER YAGI

Described here is a small, yet effective, three-element 20-meter Yagi that offers gain and good directivity. This system exhibits a front-to-back ratio in excess of 18 dB as measured with a good quality communications receiver.

Construction

The boom and all the elements are made from 1-1/4-inch diameter aluminum tubing available at most hardware stores. The two boom sections and the two pieces which make up the center portion of the driven element are coupled together using 15-inch sleeves of 1-3/8-inch OD aluminum tubing. Sheet metal screws should be used to secure the sections within the coupling sleeves.

The loading coils are wound on 1-1/8-inch diameter Plexiglas rod. Details are shown in Fig. 1. Be sure to slit the ends of the aluminum tubing where the compression clamps are placed. The coils are made from No. 14 enameled copper wire. The specified number of turns are equally spaced to cover the entire nine inches of Plexiglas.

The capacitance hats are constructed from 3/4-inch angle aluminum. Two pieces two feet in length are required for each hat. The model shown in the diagrams has the angle aluminum fastened to the element using aluminum strips however No. 8 sheet metal screws provide a suitable substitute. Solder lugs are fastened to the ends of the angle aluminum and No. 12 or 14 wire connects the ends of the aluminum resulting in a square loop. The wires should be soldered at each of the solder lugs.

All of the elements are secured to the boom with TV U-bolt hardware. Plated bolts are desirable to prevent rust from forming. An aluminum plate nine inches square by 1/4-inch thick was used as the boom-to-mast plate.

TABLE I

Complete parts list for the short beam.

QTY	*MATERIAL*
2	10-foot lengths of 1-1/4-inch dia. aluminum tubing (one for the reflector center section, one for the reflector end sections).
3	Eight-foot lengths of 1-1/4-inch dia. aluminum tubing (two lengths for the boom, one length for the director element center).
4	Six-foot lengths of 1-1/4-inch dia. aluminum tubing (two lengths for the driven element center, two lengths for the director and driven element ends).
2	15-inch lengths of 1-5/8-inch dia. aluminum tubing.
1	40-inch length of 3/8-inch dia. aluminum tubing.
4	Six-foot lengths of 3/4-inch angle aluminum.
6	12-inch lengths of 1-1/8-inch dia. Plexiglas rod.
1	Nine-inch square, 1/4-inch thick aluminum plate.
8	U-bolts.
12	Compression hose clamps.
8	Crutch caps.
38'	No. 12 enameled copper wire.
60'	No. 14 enameled copper wire.

A boom strut is recommended because the weight of the elements is sufficient to cause the boom to sag. A 1/8-inch diameter nylon line is plenty strong. A U-bolt clamp is placed on the mast several feet above the antenna and provides

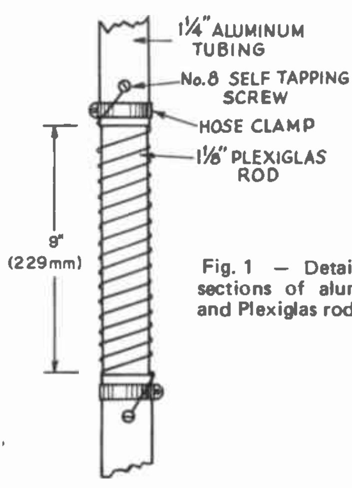
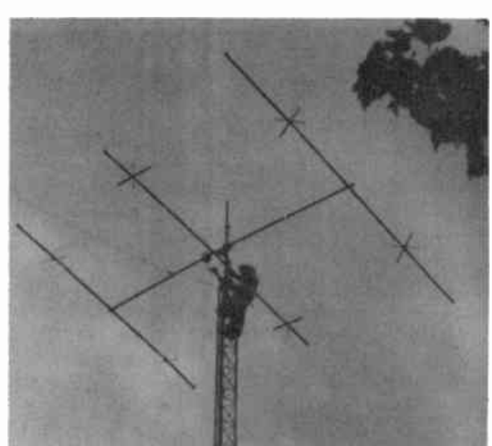
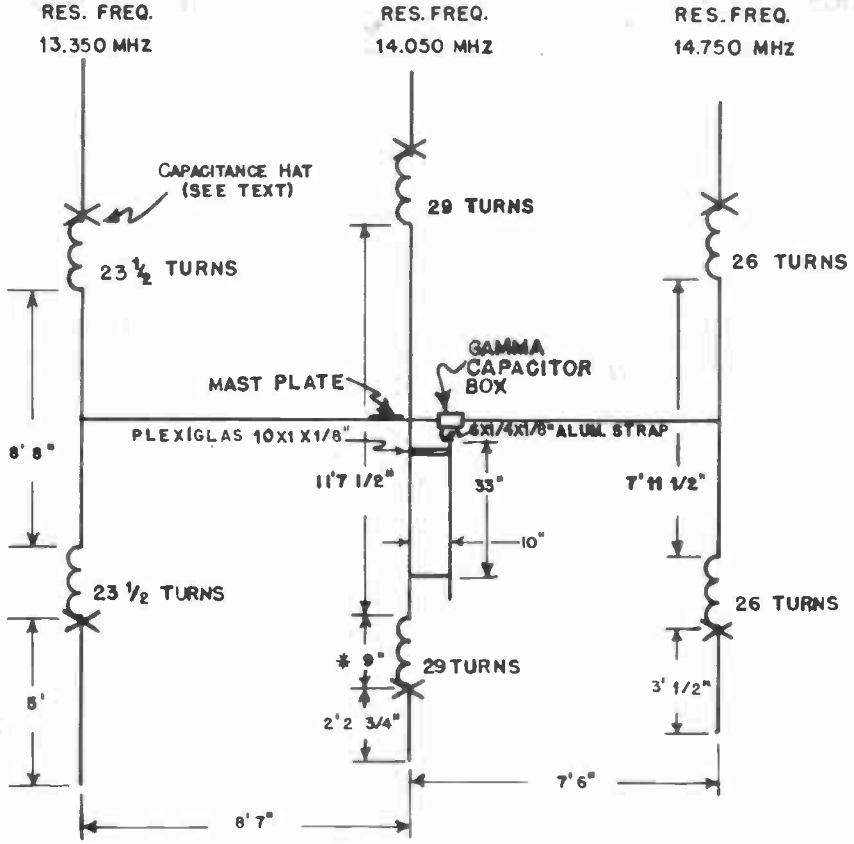


Fig. 1 — Details for joining sections of aluminum tubing and Plexiglas rod.

Shown here is WA1LNQ standing near the twenty-meter beam mounted atop the tower. Keep in mind the longest element is only 20 feet.





* ALL 6 PLEXIGLAS INSULATORS HAVE 9" (229MM) OF LENGTH EXPOSED

Fig. 2 — Constructional details for the 20-meter beam. The coils on each side of the element are identical. The gamma capacitor is a 140-pF variable unit manufactured by E. F. Johnson Co.

the attachment point for the center of the truss line. To reduce the possibility of water accumulating in the element tubing and subsequently freezing, crutch caps are placed over the ends. Rubber feet suitable for keeping furniture from scratching hardwood floors would serve the same purpose.

A piece of Plexiglas was mounted inside an aluminum Minibox to provide support and insulation for the gamma capacitor. A plastic refrigerator box would serve the purpose just as well. The capacitor housing is mounted to the boom by means of a U-bolt. The gamma rod is made of 3/8-inch aluminum 40-inches long and is connected to the gamma capacitor by a 6-inch length of strap aluminum.

Tune-Up and Operation

The builder is encouraged to follow the dimensions given in Fig. 1 as a starting point for the position of the gamma rod shorting strap. Connect the coaxial cable and install the antenna near or at the top of the tower. The gamma capacitor should be adjusted for minimum SWR at 14.100 MHz as indicated by an SWR meter (or power meter) connected in the feedline at the gamma capacitor box. If a perfect match cannot be obtained a slight repositioning of the gamma short might be required. The dimensions given favor the cw portion of the band. At 14.050 MHz the SWR is 1.1:1 and at 14.350 MHz the SWR is less than 2:1 making this antenna useful for phone as well as cw.

AN OPTIMUM-GAIN TWO-BAND ARRAY

If optimum performance is desired from a Yagi, the dual-4-element array shown in Fig. 21-43 will be of interest. This antenna consists of four elements on 15 meters interlaced with the same number for 10. Wide spacing is used, providing excellent gain and good bandwidth on both bands. Each driven element is fed separately with 50-ohm coax; gamma-matching systems are employed. If desired, a single feed line can be run to the array and then switched by a remotely controlled relay.

The element lengths shown in Fig. 21-44 are for the phone portions of the band, centered at 21,300 and 28,600 kHz. If desired, the element lengths can be changed for cw operation, using the dimensions given in Table 21-II. The spacing of the

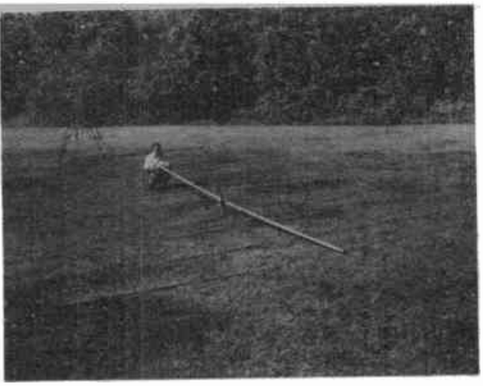


Fig. 21-43 – Ready for erection, this is the completed dual-band beam.

elements will remain the same for both phone and cw.

Construction Details

The elements are supported by commercially made U-bolt assemblies. Or, muffler clamps make excellent element supports. The boom-to-mast support is also a manufactured item that is designed to hold a 2-inch diameter boom and that can be used with mast sizes up to 2 1/2 inches in diameter. Another feature of this device is that it permits the beam to be tilted after it is mounted in place on the tower, providing access to the elements if they need to be adjusted once the beam has been mounted on the tower.

The elements are made from 6061-T6 aluminum tubing, which is available from metal suppliers. The tubing comes in 12-foot lengths and can be purchased in telescoping sizes. The center sections of the 15-meter beam elements are 1-inch outside diameter and the 10-meter sections are 3/4-inch. The ends of the tubing are slit with a

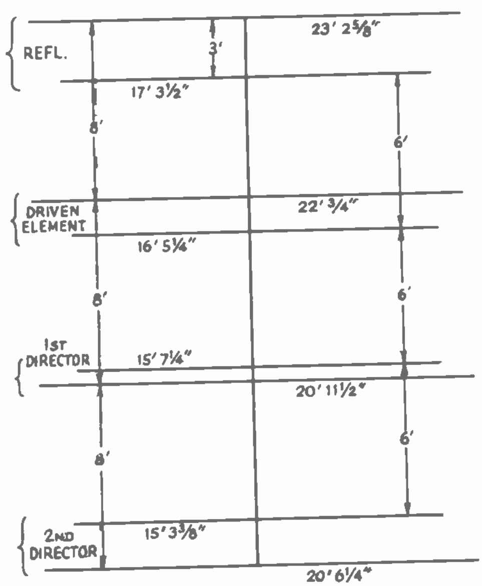


Fig. 21-44 – The element lengths shown are for the phone sections of the bands. Table 21-II provides the dimensions for cw frequencies.

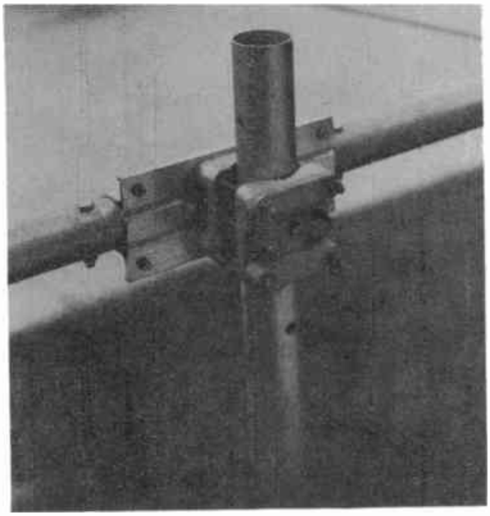
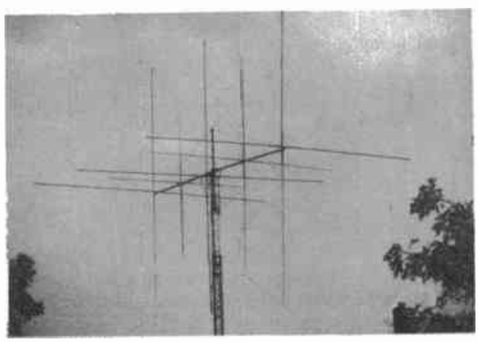


Fig. 21-45 – This is the boom-to-mast fixture that holds the two 12-foot boom sections together. The unit is made by Hy-Gain Electronics, P. O. Box 5407-HE, Lincoln, NE 68505.

hack saw, and hose clamps are used to hold the telescoping portions.

A THREE-BAND QUAD ANTENNA SYSTEM

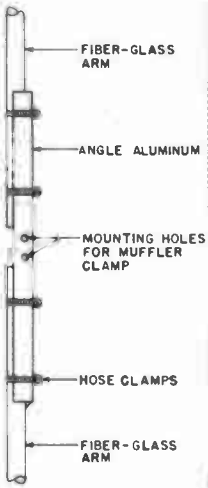
Quads have been popular with amateurs during the past few decades because of their light weight, relatively small turning radius, and their unique ability to provide good DX performance when mounted close to the earth. A two-element three-band quad, for instance, with the elements



The three-band quad antenna.

mounted only 35 feet above the ground, will give good performance in situations where a triband Yagi will not. Fig. 1 shows a large quad antenna which can be used as a basis for design for either smaller or larger arrays.

Five sets of element spreaders are used to support the three-element 20-meter, four-element 15-meter, and five-element 10-meter wire-loop system. The spacing between elements has been chosen to provide optimum performance con-



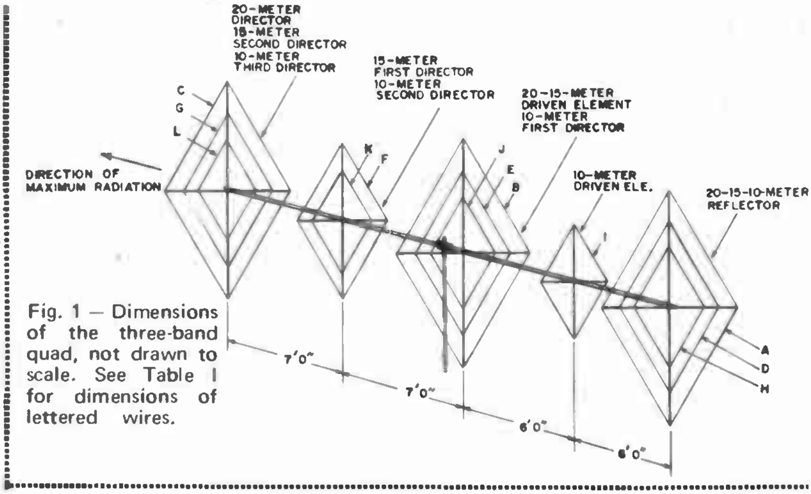


Fig. 1 — Dimensions of the three-band quad, not drawn to scale. See Table I for dimensions of lettered wires.

Fig. 2 — Details of one of two assemblies for a spreader frame. The two assemblies are jointed to form an X with a muffler clamp mounted at the position shown.

TABLE I

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' 6½"	(E) 47' 7½"	(F) 46' 5"	(G) 46' 5"	–
10 Meters	(H) 36' 2½"	(I) 35' 6"	(J) 34' 7"	(K) 34' 7"	(L) 34' 7"

Letters indicate loops identified in Fig. 1

sistent with boom length and mechanical construction. Each of the parasitic loops is closed (ends soldered together) and requires no tuning. All of the loop sizes are listed in Table I and are designed for a center frequency of 14.1, 21.1, and 28.3 MHz. Since quad antennas are rather broad-tuning devices excellent performance is achieved in both cw and ssb band segments of each band (with the possible exception of the very high end of 10 meters). Changing the dimensions to favor a frequency 200 kHz higher in each band to create a "phone" antenna is not necessary.

One question which comes up quite often is whether to mount the loops in a diamond or a square configuration. In other words, should one spreader be horizontal to the earth, or should the wire be horizontal to the ground (spreaders mounted in the fashion of an X)? From the electrical point of view, it is probably a trade-off. While the square configuration has its lowest point higher above ground than a diamond version (which may lower the angle of radiation slightly), the top is also lower than that of a diamond shaped array. Some authorities indicate that separation of the current points in the diamond system gives slightly more gain than is possible with a square layout. It should be pointed out, however, that there never has been any substantial proof in favor of one or the other, electrically.

Spreader supports (sometimes called spiders) are available from many different manufacturers. If the builder is keeping the cost at a minimum, he should consider building his own. The expense is about half that of a commercially manufactured equivalent and, according to some authorities, the homemade arm supports described below are less likely to rotate on the boom as a result of wind pressure.

A three-foot long section of one-inch-per-side steel angle stock is used to interconnect the pairs of spreader arms. The steel is drilled at the center to accept a muffler clamp of sufficient size to clamp the assembly to the boom. The fiber glass is attached to the steel angle stock with automotive hose clamps, two per pole. Each quad-loop spreader frame consists of two assemblies of the type shown in Fig. 2.

A 20-METER VERTICAL BEAM

An excellent parasitic array for 20 meters is a 3-element vertical beam originally described by W2FMI in June, 1972, *QST*. The antenna is actually one-half of a Yagi array using quarter-wave elements with spacing between elements of 0.2 wavelength (12-1/2 feet on 20 meters). This spacing results in a good compromise between gain and input impedance. Closer spacing would reduce the input impedance, and hence the efficiency, because of the inherent earth losses with vertical antennas. This vertical symmetrical Yagi allows for electrical beam switching (changing a director into a reflector by switching in a loading coil at the base) while maintaining a constant input impedance at the driven element. The dimensions of the three-element antenna, when used as a fixed or a switched array, are shown in Table 21-IV. The elements are constructed using 1/16-inch-wall aluminum tubing and consist of three telescoping sections with one-inch OD tubing used for the bottom portions. This results in a self-supporting structure. Actually, many choices are available, including No. 14 or 12 wire taped to bamboo poles.

The three-element array with the full image plane presents an input impedance of 15 ohms. Matching is accomplished with the step-down transformer, a 4:1 unbalanced-to-unbalenced toroidal balun. This transformer is also shown in Fig. 21-52 connected to the driven element.

Fig. 21-53 shows the geometry of the image plane. The inner square has a diagonal of 4/10 wavelength (25 feet). The outer wires of these sections are No. 14 wire and the inner wires are No. 18. All cross-connected wires were wire-wrapped and soldered. The pattern was chosen to give an easy path for the surface currents of a five-element array (parasitic elements at the four corners). The outer radials were all 0.4 wavelength long and also of No. 18 wire. Twenty-five wires emanated from each corner and nine from the sides.

TABLE 21-IV

Dimensions of 20-Meter Parasitic 3-Element Array

1) Fixed Array	
Director	15 ft 8 in.
Driven Element	16 ft
Reflector	17 ft 7 in.
Spacing Between Elements	12-1/2 ft
2) Switched Array	
Director and Reflector	15 ft
Driven Element	16 ft
Spacing Between Elements	12-1/2 ft
Loading Coil	2 ft No. 12 wire wound 3 turns with 3 in. dia. Length adjusted for max. F/B ratio

Fig. 21-52 — Base hardware of the driven element and the matching transformer.

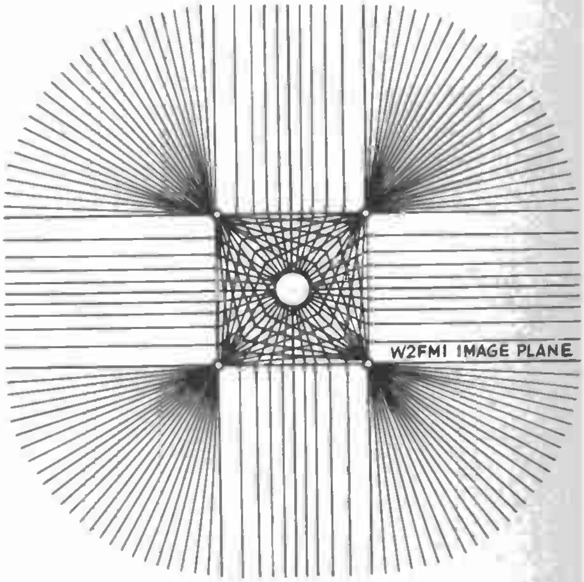


Fig. 21-53 — Geometry of the image plane used in this investigation. The pattern was chosen to approximate lines of current flow.

Fig. 21-54 — Base of one of the parasitic elements showing the relay enclosure, loading coil, and the indicator meter of the field-strength detector, which was located 2 wavelengths away.

STANDARD SIZES OF ALUMINUM TUBING

Many hams like to experiment with antennas but one problem in making antennas using aluminum tubing is knowing what sizes of tubing are available. If you want to build a beam, many questions about tubing sizes, weights, what size tubing fits into what other size, and so forth must be answered.

Table 21-V gives the standard sizes of aluminum tubing that are stocked by most aluminum suppliers or distributors in the United States and Canada. Note that all tubing comes in 12-foot lengths and also that any diameter tubing will fit into the next larger size, if the larger size has a 0.058-inch wall thickness. For example, 5/8-inch tubing has an outside diameter of 0.625 inches and will fit into 3/4-inch tubing with a 0.058-inch wall which has an inside diameter of 0.634 inches. Having used quite a bit of this type tubing it is possible to state that 0.009-inch clearance is just right for a slip fit or for slotting the tubing and then using hose clamps. To repeat, always get the next larger size and specify a 0.058-inch wall to obtain the 0.009-inch clearance.

With the chart, a little figuring will provide all the information needed to build a beam, including what the antenna will weigh. The 6061-T6 type of aluminum is a relatively high strength and has good workability, plus being highly resistant to corrosion and will bend without taking a "set."

Check the Yellow Pages for aluminum dealers.

TABLE 21-V

6061-T6 (61S-T6) ROUND ALUMINUM TUBE

In 12-Foot Lengths

O. D. Inches	Wall Thickness Inches	Wall Thickness 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.
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.

A SMALL YAGI FOR 40 METERS

Fig. 1 — The short 40-meter Yagi resembles a large 20-meter system.

A 7-MHz antenna for most amateur installations consists of a half-wave dipole attached between two convenient supports and fed power at the center with coaxial cable. When antenna gain is a requirement on this frequency, the dimensions of the system can become overwhelming. A full size three-element Yagi typically would have 68-foot elements and a 36-foot boom. Accordingly, half size elements present some distinct mechanical as well as economical advantages. Reducing the spacing between elements is not recommended since it would severely restrict the bandwidth of operation and make the tuning critical. Good directivity and reasonable gain are features of this array, yet the mechanical design allows the use of a "normal" heavy-duty rotator and a conventional tower support. Element loading is accomplished by lumped inductance and capacitance hats along the 38-foot elements.

Construction

The system described here is similar to the three-element antenna for 20 meters described earlier in this chapter. Some minor changes have been made to allow the use of standard sizes and lengths of aluminum tubing. All three elements are the same length; the tuning of the inductor is slightly different on each element, however. The two parasitic elements are grounded at the center with the associated boom-to-element hardware. A helical hairpin match is used to provide a proper match to the split and insulated driven element. Two sections of steel angle stock are used to reinforce the driven-element mounting plate since the Plexiglas center insulating material is not rigid and element sag might otherwise result. The parasitic element center sections are continuous sections of aluminum tubing and additional support is not needed here. Figs. 2 and 3 show the details clearly.

The inductors for each element are wound on 1-1/8-inch diameter solid Plexiglas cast rod. Each end of the coil is secured in place with a solder lug and the Plexiglas is held in position with an automotive compression clamp. The total number of turns needed to resonate the elements correctly is given in Fig. 5. The capacitance hats consist of 1/2-inch tubing three feet long (two pieces used) attached to the element directly next to the coil on each parasitic element and two inches away from

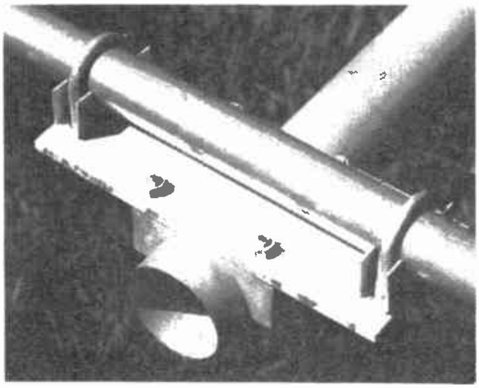


Fig. 2 — The parasitic elements are held in position with a small plate and four automotive muffler clamps.

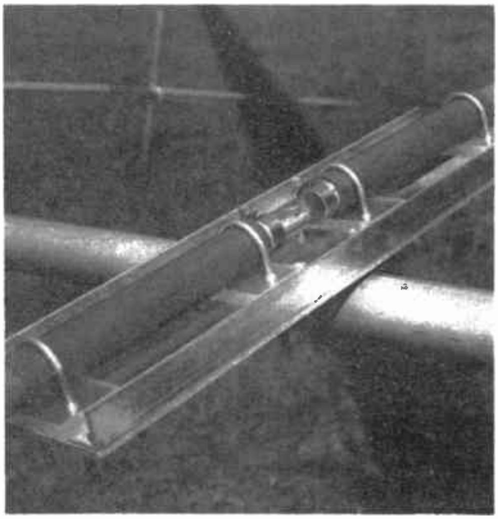


Fig. 3 — The driven element needs to be insulated from the boom. Insulation is provided by PVC tubing held in place on sheet plastic with U bolts.

the coil for the driven element. Complete details are given in Fig. 4.

The boom is constructed from three sections of aluminum tubing which measures 2-1/2 inches diameter and 12 feet long. These pieces are joined together with inner tubes made from 2-1/4-inch stock shimmed with aluminum flashing. Long strips, approximately one inch wide, are wound on the inner tubing before it is placed inside the boom sections. A pair of 3/8 × 3-1/2 inch steel bolts are placed at right angles to each other at every connection point to secure the boom. Caution: do not over tighten the bolts since this will distort the tubing making it impossible to pull apart sections, should the need arise. It is much better to install locking nuts over the original ones to assure mechanical security.

The helical hairpin details are given in Fig. 6. Quarter inch copper tubing is formed into seven turns approximately four inches long and 2-1/4 inches ID.

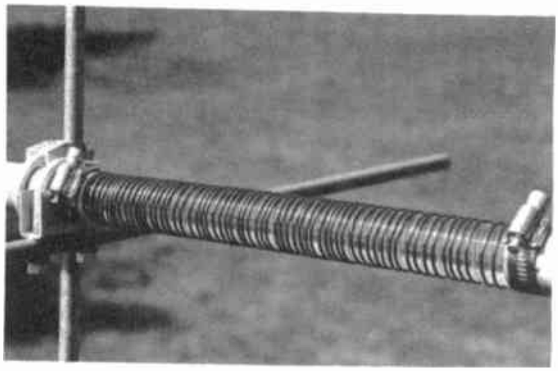


Fig. 4 – Each loading coil is wound on Plexiglas rod. The capacitance hats for the parasitic elements are mounted next to the coil, as shown here. The hose clamps compress the tubing against the Plexiglas rod. Each capacitance hat consists of two sections of tubing and associated muffler clamps.

Tuning and Matching

The builder is encouraged to carefully follow the dimensions given in Fig. 5. Tuning the elements with the aid of a grid-dip oscillator has proved to be somewhat unreliable and accordingly, no resonant frequencies will be given.

The hairpin matching system may not resemble the usual form but its operation and adjustment are essentially the same. For a detailed explanation of this network, see the Transmission Line chapter of *The ARRL Antenna Book*, thirteenth edition. The driven element resonant frequency required for the hairpin match is determined by the placement of the capacitance hats with respect to the ends of the coils. Sliding the capacitance hats away from the ends of the coils increases the resonant frequency (capacitive reactance) of the element to cancel the effect of the hairpin inductive reactance. The model shown here had capacitance hats mounted 2-1/2 inches out from the ends of the coils (on the driven element only). An SWR indicator or wattmeter should be installed in series with the feed line *at the antenna*. The hairpin coil may be spread or compressed with an insulated tool (or by hand if power is removed!) to provide minimum reflected power at 7.050 MHz. The builder should not necessarily strive for a perfect match by changing the position of the capacitance hats since this may reduce the bandwidth of the matching system. An SWR of less than 2 to 1 was achieved across the entire 40-meter band with the antenna mounted atop an 80-foot tower.

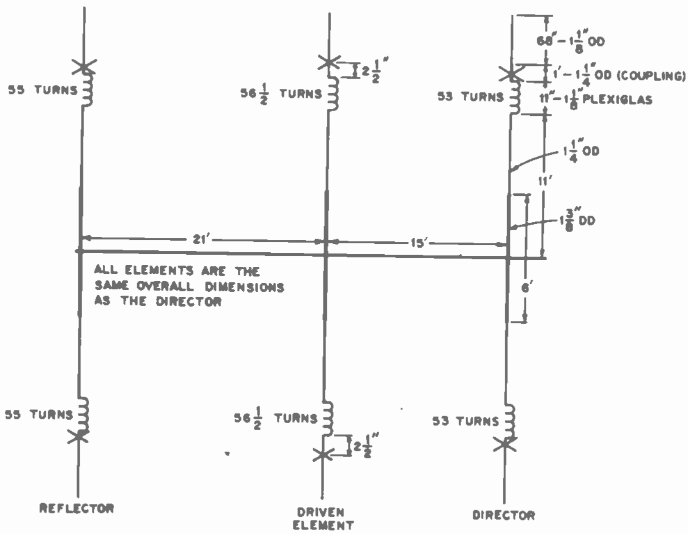


Fig. 5 – Mechanical details and dimensions for the 40-meter Yagi. Each of the elements uses the same dimensions; the difference is only the number of turns on the inductors and the placement of the capacitance hats. See the text for more details.

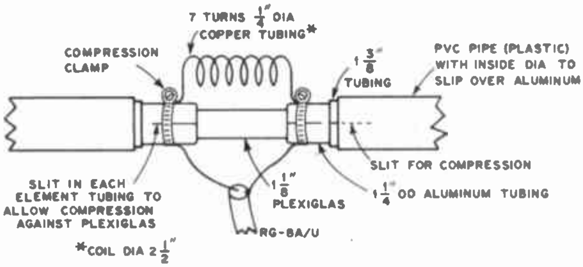


Fig. 6 — Driven-element hairpin matching details.

The tuning of the array can be checked by making front-to-back ratio measurements across the band. With the dimensions given here, the best figures of front-to-back (approximately 25 to 30 dB) should be noticed in the cw portion of the band. Should the builder suspect the tuning is incorrect or if the antenna is mounted at some height greatly different than 80 feet, retuning of the elements may be necessary.

ANTENNA SUPPORTS

"A"-FRAME MAST

The simple and inexpensive mast shown in Fig. 1 is satisfactory for heights up to 35 or 40 feet. Clear, sound lumber should be selected. The completed mast may be protected by two or three coats of house paint.

If the mast is to be erected on the ground, a couple of stakes should be driven to keep the bottom from slipping and it may then be "walked up" by a pair of helpers. If it is to go on a roof, first stand it up against the side of the building and then hoist it from the roof, keeping it vertical. The whole assembly is light enough for two men to perform the complete operation – lifting the mast, carrying it to its permanent berth, and fastening the guys – with the mast vertical all the while. It is entirely practicable, therefore, to errect this type of mast on any small, flat area of roof.

By using 2 X 3s or 2 X 4s, the height may be extended up to about 50 feet. The 2 X 2 is too flexible to be satisfactory at such heights.

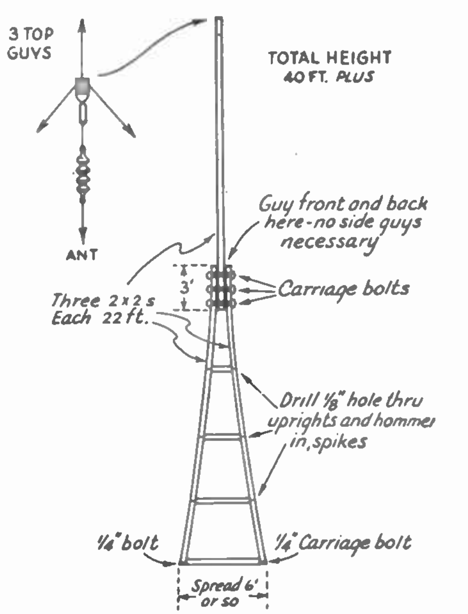


Fig. 1 — Details of a simple 40-foot "A"-frame mast suitable for erection in locations where space is limited.

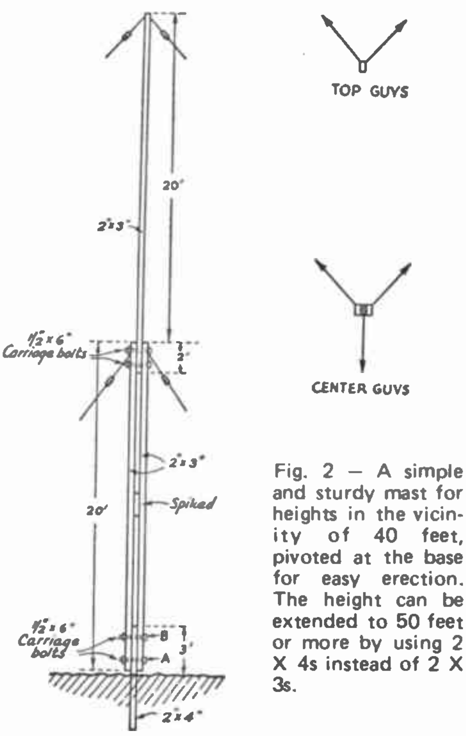


Fig. 2 — A simple and sturdy mast for heights in the vicinity of 40 feet, pivoted at the base for easy erection. The height can be extended to 50 feet or more by using 2 X 4s instead of 2 X 3s.

ANTENNA FORMULAS AND TABLES

The length in feet of a half-wavelength dipole is given by

$$L(\text{feet}) = \frac{492}{f(\text{MHz})}$$

An empirical formula that is more applicable to actual antennas is

$$L(\text{feet}) = \frac{468}{f(\text{MHz})}$$

At resonance, the input impedance of a half-wavelength dipole is 72 Ω. At lower frequencies, the input impedance appears as an increasing capacitive reactance with a decreasing radiation resistance (with decreasing frequency). At frequencies above resonance, the antenna appears as an increasing inductive reactance in series with an increasing radiation resistance. A peak in antenna impedance is reached when the electrical length is approximately one wavelength.

Similar considerations hold for vertical wires over conducting ground planes or random-length wire antennas. For thin conductors, the input impedance of a quarter-wavelength vertical wire is approximately 36 Ω. This impedance increases to over a 1000 Ω as the next resonance at approximately a half wavelength is approached. In the 5/8-wavelength mode, the impedance is very nearly 50 ohms which makes this length radiator useful for matching to 50-Ω transmission line. This effect is often used to advantage in vhf mobile vertical whips.

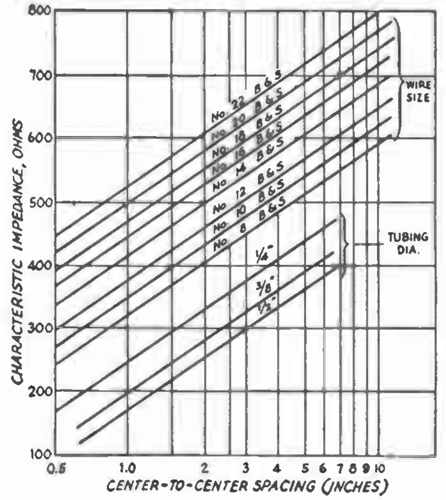


Fig. 1 — Chart showing the characteristic impedance of spaced-conductor parallel transmission lines with air dielectric. Tubing sizes given are for outside diameters.

TABLE I

Characteristics of Commonly-Used Transmission Lines

Type of Line	Z_0 *Ohms*	*Vel. %*	*pF per ft.*	*OD*	*Attenuation in dB per 100 feet* 3.5	7	14	21	28	50	144	420
RG58/A-AU	53	66	28.5	0.195	0.68	1.0	1.5	1.9	2.2	3.1	5.7	10.4
RG58 Foam Diel.	50	79	25.4	0.195	0.52	0.8	1.1	1.4	1.7	2.2	4.1	7.1
RG59/A-AU	73	66	21.0	0.242	0.64	0.90	1.3	1.6	1.8	2.4	4.2	7.2
RG59 Foam Diel.	75	79	16.9	0.242	0.48	0.70	1.0	1.2	1.4	2.0	3.4	6.1
RG8/A-AU	52	66	29.5	0.405	0.30	0.45	0.66	0.83	0.98	1.35	2.5	4.8
RG8 Foam Diel.	50	80	25.4	0.405	0.27	0.44	0.62	0.76	0.90	1.2	2.2	3.9
RG11/A-AU	75	66	20.5	0.405	0.38	0.55	0.80	0.98	1.15	1.55	2.8	4.9
Aluminum Jacket, Foam Diel.[1]												
3/8 inch	50	81	25.0	–	–	–	0.36	0.48	0.54	0.75	1.3	2.5
1/2 inch	50	81	25.0	–	–	–	0.27	0.35	0.40	0.55	1.0	1.8
3/8 inch	75	81	16.7	–	–	–	0.43	0.51	0.60	0.80	1.4	2.6
1/2 inch	75	81	16.7	–	–	–	0.34	0.40	0.48	0.60	1.2	1.9
Open-wire[2]	–	97	–		0.03	0.05	0.07	0.08	0.10	0.13	0.25	–
300-ohm Twin-lead	300	82	5.8		0.18	0.28	0.41	0.52	0.60	0.85	1.55	2.8
300-ohm tubular	300	80	4.6		0.07	0.25	0.39	0.48	0.53	0.75	1.3	1.9
Open-wire, TV type												
1/2 inch	400	95			0.028	0.05	0.09	0.13	0.17	0.30	0.75	–
1 inch	450	95			0.028	0.05	0.09	0.13	0.17	0.30	0.75	–

1 Polyfoam dielectric type line information courtesy of Times Wire and Cable Co.

2 Attenuation of open-wire line based on No. 12 conductors, neglecting radiation.

VHF and UHF Antennas

Improving his antenna system is one of the most productive moves open to the vhf enthusiast. It can increase transmitting range, improve reception, reduce interference problems, and bring other practical benefits. The work itself is by no means the least attractive part of the job. With even high-gain antennas, experimentation is greatly simplified, at vhf and uhf, because an array is a workable size, and much can be learned about the nature and adjustment of antennas. No large investment in test equipment is necessary.

Whether we buy or build our antennas, we soon find that there is no one "best" design for all purposes. Selecting the antenna best suited to our needs involves much more than scanning gain figures and prices in a manufacturer's catalog. The first step should be to establish priorities.

OBJECTIVES

Gain: Shaping the pattern of an antenna, to concentrate radiated energy, or received - signal pickup, in some directions at the expense of others is the only way to develop gain. This is best explained by starting with the hypothetical isotropic antenna, which would radiate equally in all directions. A point source of light illuminating the inside of a globe uniformly, from its center, is a visual analogy. No practical antenna can do this, so all antennas have "gain over isotropic" (dBi). A half-wave dipole in free space has 2.1 dBi. If we can plot the radiation pattern of antenna in all planes, we can compute its gain, so quoting it with respect to isotropic is a logical base for agreement and understanding. It is rarely possible to erect a half-wave antenna that has anything approaching a free-space pattern, and this fact is responsible for much of the confusion about true antenna gain.

Radiation patterns can be controlled in various ways. One is to use two or more driven elements, fed in phase. Such collinear arrays provide gain without markedly sharpening the frequency response, compared to that of a single element. More gain per element, but with a sacrifice in frequency coverage, is obtained by placing parasitic elements longer and shorter than the driven one, in the plane the first element, but not driven from the feedline. The reflector and directors of a Yagi array are highly frequency sensitive and such an antenna is at its best over frequency changes of less than one percent of the operating frequency.

Frequency Response: Ability to work over an entire vhf band may be important in some types of work. The response of an antenna element can be broadened somewhat by increasing the conductor diameter, and by tapering it to something approximating cigar shape, but this is done mainly with simple antennas. More practically, wide frequency coverage may be a reason to select a collinear array, rather than a Yagi. On the other hand, the growing tendency to channelize operations in small segments of our bands tends to place broad frequency coverage low on the priority list of most vhf stations.

Radiation Pattern: Antenna radiation can be made omnidirectional, bidirectional, practically unidirectional, or anything between these conditions. A vhf net operator may find an omnidirectional system almost a necessity, but it may be a poor choice otherwise. Noise pickup and other interference problems tend to be greater with such antennas, and those having some gain are especially bad in these respects. Maximum gain and low radiation angle are usually prime interests of the weak-signal DX aspirant. A clean pattern, with lowest possible pickup and radiation off the sides and back, may be important in high-activity areas, or where the noise level is high.

Height Gain: In general, the higher the better in vhf antenna installations. If raising the antenna clears its view over nearby obstructions, it may make dramatic improvements in coverage. Within reason greater height is almost always worth its cost, but height gain must be balanced against increased transmission-line loss. The latter is considerable, and it increases with frequency. The best available line may be none too good, if the run is long in terms of wavelength. Give line-loss information, shown in table form in Chapter 20, close scrutiny in any antenna planning.

Physical Size: A given antenna design for 432 MHz will have the same gain as one for 144 MHz, but being only one-third the size it will intercept only one-third as much energy in receiving. Thus, to be equal in communication effectiveness, the 432-MHz array should be at least equal in *size* to the 144-MHz one, which will require roughly three times as many elements. With all the extra difficulties involved in going higher in frequency, it is well to be on the big side, in building an antenna for the higher band.

DESIGN FACTORS

Having sorted out objectives in a general way, we face decisions on specifics, such as polarization, type of transmission line, matching methods and mechanical design.

Polarization: Whether to position the antenna elements vertical or horizontal has been a moot point since early vhf pioneering. Tests show little evidence on which to set up a uniform polarization policy. On long paths there is no consistent advantage, either way. Shorter paths tend to yield higher signal levels with horizontal in some kinds of terrain. Man-made noise, especially ignition interference, tends to be lower with horizontal. Verticals are markedly simpler to use in omnidirectional systems, and in mobile work.

Early vhf communication was largely vertical, but horizontal gained favor when directional arrays became widely used. The major trend to fm and repeaters, particularly in the 144-MHz band, has tipped the balance in favor of verticals in mobile work and for repeaters. Horizontal predominates in other communication, on 50 MHz and higher frequencies. It is well to check in advance in any new area in which you expect to operate, however, as some localities still use vertical almost exclusively. A circuit loss of 20 dB or more can be expected with cross-polarization.

Transmission Lines: There are two main categories of transmission lines: balanced and unbalanced. The former include open-wire lines separated by insulating spreaders, and Twin-Lead, in which the wires are embedded in solid or foamed insulation. Line losses result from ohmic resistance, radiation from the line, and deficiencies in the insulation. Large conductors, closely spaced in terms of wavelength, and using a minimum of insulation, make the best balanced lines. Impedances are mainly 300 to 500 ohms. Balanced lines are best in straight runs. If bends are unavoidable, the angles should be as obtuse as possible. Care should be taken to prevent one wire from coming closer to metal objects than the other. Wire spacing should be less than 1/20 wavelength.

Properly built, open-wire line can operate with very low loss in vhf and even uhf installations. A total line loss under 2 dB per hundred feet at 432 MHz is readily obtained. A line made of No. 12 wire, spaced 3/4 inch or less with Teflon spreaders, and running essentially straight from antenna to station, can be better than anything but the most expensive coax, at a fraction of the cost. This assumes use of baluns to match into and out of the line, with a short length of quality coax for the moving section from the top of the tower to the antenna. A similar 144-MHz setup could have a line loss under 1 dB.

Small coax such as RG-58 or 59 should never be used in vhf work if the run is more than a few feet. Half-inch lines (RG-8 or 11) work fairly well at 50 MHz, and are acceptable for 144-MHz runs of 50 feet or less. If these lines have foam rather than solid insulation they are about 30 percent better. Aluminum-jacket lines with large inner conductors and foam insulation are well worth their cost. They are readily water-proofed, and can last almost indefinitely. Beware of any "bargains" in coax for vhf or uhf uses. Lost transmitter power can be made up to some extent by increasing power, but once lost, a weak signal can never be recovered in the receiver.

Effects of weather should not be ignored. A well-constructed open-wire line works well in nearly any weather, and it stands up well. Twin-Lead is almost useless in heavy rain, wet snow or icing. The best grades of coax are impervious to weather. They can be run underground, fastened to metal towers without insulation, or bent into any convenient position, with no adverse effects on performance.

Impedance Matching

Theory and practice in impedance matching are given in detail in earlier chapters, and theory, at least, is the same for frequencies above 50 MHz. Practice may be similar, but physical size can be a major modifying factor in choice of methods. Only the matching devices used in practical construction examples later in this chapter will be discussed in detail here. This should not rule out consideration of other methods, however, and a reading of relevant portions of Chapters 20 and 21 is recommended.

Universal Stub: As its name implies, the double-adjustment stub of Fig. 22-1A is useful for many matching purposes. The stub length is varied to resonate the system, and the transmission line is tapped onto the stub at the point where line and stub impedances are equal. In practice this involves moving both the sliding short and the point of line connection for zero reflected power, as indicated on an SWR bridge connected in the line.

The universal stub allows for tuning out any small reactance present in the driven part of the

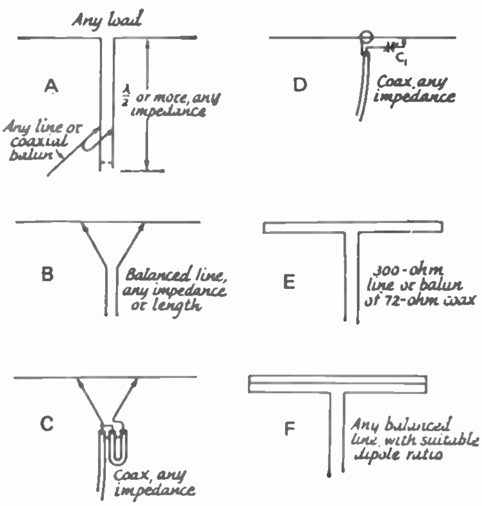


Fig. 22-1 – Matching methods commonly used in vhf antennas. The universal stub, A, combines tuning and matching. The adjustable short on the stub, and the points of connection of the transmission line, are adjusted for minimum reflected power in the line. In the delta match, B and C, the line is fanned out to tap on the dipole at the point of best impedance match. Impedances need not be known in A, B and C. The gamma-match, D, is for direct connection of coax. C1 tunes out inductance in the arm. Folded dipole of uniform conductor size, E, steps up antenna impedance by a factor of 4. Using a larger conductor in the unbroken portion of the folded dipole, E, gives higher orders of impedance transformation.

system. It permits matching antenna to line without knowledge of the actual impedances involved. The position of the short yielding the best match gives some indication of amount of reactance present. With little or no reactive component to be tuned out, the stub will be approximately a half-wavelength from load to short.

The stub should be stiff bare wire or rod, spaced no more than 1/20 wavelength. Preferably it should be mounted rigidly, on insulators. Once the position of the short is determined, the center of the short can be grounded, if desired, and the portion of the stub no longer needed can be removed.

It is not necessary that the stub be connected directly to the driven element. It can be made part of an open-wire line, as a device to match into or out of the line with coax. It can be connected to the lower end of a delta match, or placed at the feedpoint of a phased array. Examples of these uses are given later.

Delta Match: Probably the first impedance match was made when the ends of an open line were fanned out and tapped onto a half-wave antenna, at the point of most efficient power transfer, as in Fig. 22-1B. Both the side length and the points of connection either side of the center of the element must be adjusted for minimum reflected power in the line, but as with the universal stub, the impedances need not be known. The delta makes no provision for tuning out reactance, so the universal stub is often used as a termination for it, to this end.

Once thought to be inferior for vhf applications because of its tendency to radiate if improperly adjusted, the delta has come back to favor, now that we have good methods for measuring the effects of matching. It is very handy for phasing multiple-bay arrays with open lines, and its dimensions in this use are not particularly critical. It should be checked out carefully in applications like that of Fig. 22-1C, having no tuning device.

Gamma Match: An application of the same principle to direct connection of coax is the gamma match, Fig. 22-1D. There being no rf voltage at the center of a half-wave dipole, the outer conductor of the coax is connected to the element at this point, which may also be the junction with a metallic or wooden boom. The inner conductor, carrying the rf current, is tapped out on the element at the matching point. Inductance of the arm is tuned out by means of C1, resulting in electrical balance. Both the point of contact with the element and the setting of the capacitor are adjusted for zero reflected power, with a bridge connected in the coaxial line.

The capacitor can be made variable temporarily, then replaced with a suitable fixed unit when the required capacitance value is found, or C1 can be mounted in a waterproof box. Maximum should be about 100 pF for 50 MHz and 35 to 50 pF for 144. The capacitor and arm can be combined in one coaxial assembly, with the arm connecting to the driven element by means of a sliding clamp, and the inner end of the arm sliding inside a sleeve connected to the inner conductor of the coax. A commercially supplied assembly of this type is used in a 50-MHz array described later, or one can be constructed from concentric pieces of tubing, insulated by plastic sleeving. Rf voltage across the capacitor is low, once the match is adjusted properly, so with a good dielectric, insulation presents no great problem, if the initial adjustment is made with low power level. A clean, permanent high-conductivity bond between arm and element is important, as the rf current flow is high at this point.

Folded Dipole: The impedance of a half-wave antenna broken at its center is 72 ohms. If a single conductor of uniform size is folded to make a half-wave dipole as shown in Fig. 22-1E, the impedance is stepped up four times. Such a folded dipole can thus be fed directly with 300-ohm line with no appreciable mismatch. Coaxial line of 70 to 75 ohms impedance may also be used, if a 4:1 balun is added. (See balun information presented later in this chapter.) Higher impedance step up can be obtained if the unbroken portion is made larger in cross-section than the fed portion, as in 22-1F. For design information, see Chapter 20.

Baluns and Transmatches: Conversion from balanced loads to unbalanced lines, or vice versa, can be performed with electrical circuits, or their equivalents made of coaxial line. A balun made from flexible coax is shown in Fig. 22-2A. The looped portion is an electrical half-wavelength. The physical length depends on the propagation factor of the line used, so it is well to check its resonant frequency, as shown at B. The two ends are shorted, and the loop at one end is coupled to a dip-meter coil. This type of balun gives an impedance stepup of 4 to 1 in impedance, 50 to 200 ohms, or 75 to 300 ohms, typically.

Coaxial baluns giving 1-to-1 impedance transfer are shown in Fig. 22-3. The coaxial sleeve, open at the top and connected to the outer conductor of the line at the lower end (A) is the preferred type. A conductor of approximately the same size as the line is used with the outer conductor to form a quarter-wave stub, in B. Another piece of coax, using only the outer conductor, will serve this purpose. Both baluns are intended to present an infinite impedance to any rf current that might otherwise tend to flow on the outer conductor of the coax.

The functions of the balun and the impedance

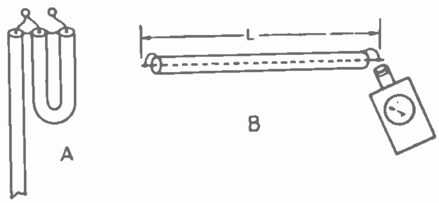


Fig. 22-2 – Conversion from unbalanced coax to a balanced load can be done with a half-wave coaxial balun, A. Electrical length of the looped section should be checked with a dip-meter, with ends shorted, B. The half-wave balun gives a 4:1 impedance step up.

transformer can be handled by various tuned circuits. Such a device, commonly called an antenna coupler or Transmatch, can provide a wide range of impedance transformations. A versatile example is described at the end of this chapter.

The *Q* Section: The impedance transforming property of a quarter-wave line is treated in Chapter 20. The parallel-bar Q section is not useful in low-impedance vhf matching situations, but Q sections of flexible coaxial line may be handy in phasing and matching vhf and uhf arrays. Such sections can be any odd multiple of a quarter-wave-length. An example of two 3/4-wave 75-ohm Q sections, used to phase and match a pair of Yagi bays, each of which has 50 ohms impedance, is given later in this chapter.

Mechanical Design

The small size of vhf and, especially, uhf arrays opens up a wide range of construction possibilities. Finding components is becoming difficult for home constructors of ham gear, but it should not hold back antenna work. Radio and TV distributors have many useful antenna parts and materials. Hardware stores, metals suppliers, lumber yards, welding-supply and plumbing-supply houses and even junkyards should not be overlooked. With a little imagination, the possibilities are endless.

Wood or Metal? Wood is very useful in antenna work, and it is almost universally available, in a great variety of shapes and sizes. Rug poles of wood or bamboo make fine booms. Round wood stock (dowelling) is found in many hardware stores in sizes suitable for small arrays. Square or rectangular boom and frame materials can be ripped to order in most lumber yards, if they are not available from the racks in suitable sizes.

There is no rf voltage at the center of a half-wave dipole or parasitic element, so no insulation is required in mounting elements that are centered in the support, whether the latter is wood or metal. Wood is good for the framework of multibay arrays for the higher bands, as it keeps down the amount of metal in the active area of the array.

Wood used for antenna construction should be well-seasoned and free of knots or damage. Available materials vary, depending on local sources. Your lumber dealer can help you better than anyone else in choosing suitable materials. Joining wood members at right angles is often done advantageously with gusset plates. These can be of thin outdoor-grade plywood or Masonite. Round materials can be handled in ways similar to those used with metal components, with U clamps and with other hardware.

Metal booms have a small "shorting effect" on elements that run through them. With materials sizes commonly employed, this is not more than one percent of the element length, and may not be noticeable in many applications. It is just perceptible with 1/2-inch tubing booms used on 432 MHz, for example. Formula lengths can be used as given, if the matching is adjusted in the frequency range one expects to use. The center frequency of an all-metal array will tend to be 0.5 to 1 percent higher than a similar system built of wooden supporting members.

Element Materials and Dimensions: Antennas for 50 MHz need not have elements larger than 1/2-inch diameter, though up to 1 inch is used occasionally. At 144 and 220 MHz the elements are usually 1/8 to 1/4 inch in diameter. For 420, elements as small as 1/16 inch in diameter work well, if made of stiff rod. Aluminum welding rod, 3/32 to 1/8 inch in diameter is fine for 420-MHz arrays, and 1/8 inch or larger is good for the 220 band. Aluminum rod or hard-drawn wire works well at 144 MHz. Very strong elements can be made with stiff-rod inserts in hollow tubing. If the latter is slotted, and tightened down with a small clamp, the element lengths can be adjusted experimentally with ease.

Sizes recommended above are usable with formula dimensions given in Table 22-I. Larger diameters broaden frequency response; smaller ones sharpen it. Much smaller diameters than those recommended will require longer elements, especially in 50-MHz arrays.

The driven element(s) of a vhf array may be cut from the formula

$$L \text{ (inches)} = \frac{5600}{\text{Freq. (MHz)}}$$

This is the basis for Table 22-I driven-element information. Reflectors are usually about 5 percent longer, and directors 5 percent shorter, though element spacing and desired antenna bandwidth affect parasitic-element lengths. The closer the reflector and director (especially the latter) are to the driven element the nearer they must be to the driven-element length to give optimum gain. This is another way of saying that close-spaced arrays tend to work effectively over narrower bandwidths than

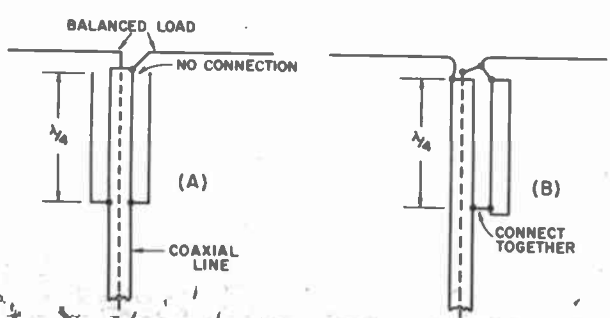


Fig. 22-3 — The balun conversion function, with no impedance change, is accomplished with quarter-wave lines, open at the top and connected to the coax outer conductor at the bottom. Coaxial sleeve, A, is the preferred type.

TABLE 22-1

Dimensions for VHF Arrays in Inches

*Freq. (MHz)**	*50**	*144**	*220**	*432**
Driven Element	111	38 5/8	25 7/16	13
Change per MHz	2	1/4	1/8	1/32
Reflector	116 1/2	40 1/2	26 3/4	13 1/2
1st Director	105 1/2	36 5/8	24 1/8	12 11/32
2nd Director	103 1/2	36 3/8	24	12 9/32
3rd Director	101 1/2	36 1/8	23 7/8	12 7/32
1.0 Wavelength	236	81 1/2	53 5/8	27 1/4
0.625 Wavelength	149	51	33 1/2	17
0.5 Wavelength	118	40 3/4	26 13/16	13 5/8
0.25 Wavelength	59	20 3/8	13 7/8	6 13/16
0.2 Wavelength	47 3/4	16 1/4	10 3/4	5 7/16
0.15 Wavelength	35 1/2	12 1/4	8	4

* Dimensions are for the *most-used* section of each band: 50 to 50.6 MHz, 144 to 145.5 MHz, 220 to 222 MHz, and 432 to 434 MHz. The element lengths should be adjusted for each megahertz difference in frequency by the amount given in the third line of the table. Example: If optimum performance is wanted much above 145 MHz, shorten all elements by about ¼ inch. For above 146 MHz, shorten by ½ inch. See text.

Element spacings are not critical, and table figures may be used, regardless of element lengths chosen. Parasitic element lengths are optimum for collinear arrays and small Yagis, having 0.2-wavelength spacing.

wide-spaced ones, though maximum gain may be possible with many different combinations of lengths and spacings.

Parasitic-element lengths of Table 22-1 are based on spacings of about 0.2 wavelength, common in relatively short Yagis and collinear arrays. Dimensions given later in the individual descriptions of antennas may be at variance with those of the table. Where this is evident, the length differences result from use of different element spacings, for the most part. Some designs are for maximum gain, without consideration of bandwidth. Still others have slightly modified spacings, to give optimum results with a particular boom length.

ANTENNAS FOR 50 MHz

Simple antennas such as dipoles, groundplanes, mobile whips and the like are covered adequately elsewhere in this *Handbook*. Adaptation of them to vhf work involves mainly reference to Table 22-1 for length information. We will be concerned here with arrays that give appreciable gain, or other properties needed in vhf communication.

Yagis, Short and Long: The Yagi array is practically standard for 50-MHz directive use. Usual sizes are three to six elements, though up to eight or nine in line are seen in ambitious installations. Director spacing, after the first three, must be very wide to be worthwhile, so boom lengths of 30 feet or more are needed for more than 6 elements. Though long Yagis certainly are desirable, it should be emphasized that the first two or three elements provide very high gain per unit of space. Even a 3-element Yagi, on as short a boom as 6 feet, is good for 7.5 dB over a dipole. To double the gain (add 3 dB) requires going to only 6 elements – but it takes a boom more than 20 feet long. If it is possible to put up a rotatable antenna at all, there is usually room for at least a 3-element structure, and the gain such an antenna provides is very helpful. Dimensions can follow those given for the first three elements of larger arrays described here.

Stacking Yagis: Where suitable provision can be made for supporting them, two Yagis mounted one above the other and fed in phase may be preferable to one long Yagi having the same theoretical or measured gain. The pair will require a much smaller turning space, for the same gain, and their lower radiation angle can provide interesting results. On long ionospheric paths a stacked pair occasionally may show an *apparent* gain much greater than the 2 to 3 dB that can be measured locally as the gain due to stacking.

Optimum spacing for Yagis of 5 elements or more is one wavelength, but this may be too much for many builders of 50-MHz antennas to handle. Worthwhile results can be obtained with as little as one half-wavelength (10 feet), and 5/8 wavelength (12 feet) is markedly better. The difference between 12 and 20 feet may not be worth the added structural problems involved in the wider spacing, at 50 MHz, at least. The closer spacings give lower measured gain, but the antenna patterns are cleaner than will be obtained with one-wavelength spacing. The extra gain with wider spacings

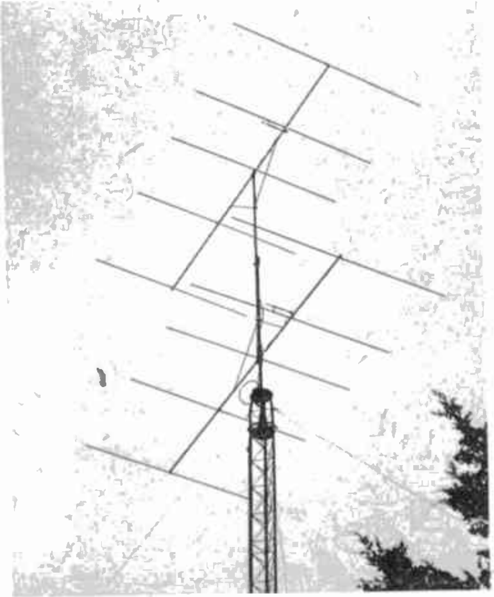


Fig. 22-4 – 5-over-5 stacked-Yagi array for 50 MHz, with all-coax feed.

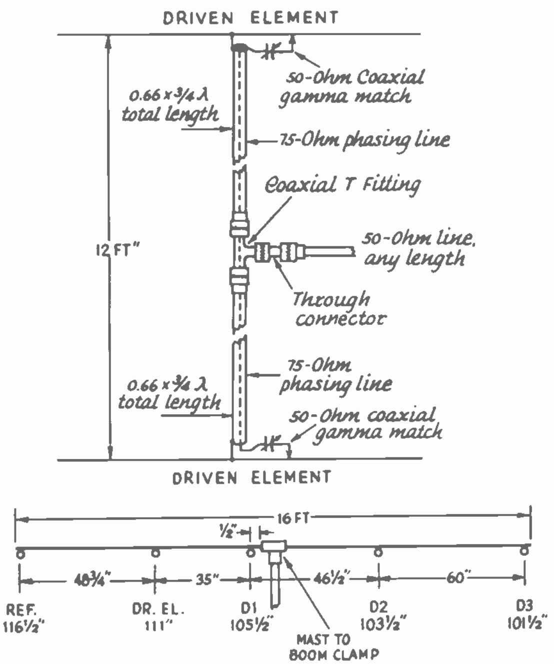


Fig. 22-5 — Principal dimensions of the 50-MHz 5-over-5, with details of the 3/4-wavelength *Q*-section matching system. The propagation factor of 0.66 applies only with solid-dielectric coax. Gamma-matching assemblies are coaxial-capacitor units (Kirk Electronics C6M).

is usually the objective on 144 MHz and higher bands, where the structural problems are not severe.

5-OVER-5 FOR 50 MHz

The information provided in Fig. 22-5 is useful for a single 5-element Yagi, or for the stacked pair of Fig. 22-4, either to be fed with a 50-ohm line. The phasing and matching arrangement may be used for any pair of Yagis designed for 50-ohm feed individually. With slight modification it will serve with Yagis designed for 200-ohm balanced feed.

Mechanical Details

Construction of the single Yagi bay or a stacked pair is simplified by use of components that should be available to most builders. Element-to-boom and boom-to-mast mounts are aluminum castings designed for these applications by Kirk Electronics, 134 Westpark Road, Dayton, Ohio 45459. The gamma matches shown schematically in Fig. 22-5 are of coaxial construction, waterproofed for long life, available from the same supplier.

Booms are made of two 8-foot lengths of 1 1/4-inch aluminum (Reynolds) found in many hardware stores. Reynolds makes a special fitting for joining sections of the tubing, but these are not widely available from the usual hardware-store stocks, so a handmade splice was substituted. A piece of the same-diameter tubing as the booms, 12 inches or more in length, is slotted with a hacksaw, and then compressed to fit inside the ends of the two 8-foot lengths, as seen in Fig. 22-6. If the splice is held in the compressed position with large pipe pliers or a hose clamp, the ends will slide inside the boom sections readily. When the splice is released from compression, the two tubes can be driven together. Self-tapping screws should be run through the tubes and the splice, to hold the assembly firm. Use at least two on each side of the splice.

Elements are 1/2-inch aluminum tubing, Alcoa alloy 6061-T6. Almost any aluminum should be suitable. Kirk Yagi clamps, one-piece aluminum castings designed for this job, are available for 3/8 as well as 1/2-inch elements, and 1 1/4-inch boom. The eyes through which the elements pass are drilled, but must be tapped for 10-32 setscrews to tighten the elements firmly in place, two screws per element. The portion of the clamp that surrounds the boom can be spread slightly to allow the clamp to slide along the boom to the desired point. The interior surface is slightly rough, so tightening the yoke with the screw provided with the clamps makes the element set firmly on the boom. The reflector, driven element and first director are all in back of the boom splice.

The vertical member of the stacked array is 1 1/4-inch thick-wall anodized steel tubing, commonly used in large antenna installations for home TV. Do not use thin-wall aluminum or light galvanized steel masting. The aluminum is not strong enough, and inexpensive steel masting rusts inside, weakening the structure and inviting failure.

Spacing between bays can be a half wavelength (10 feet), 5/8 wavelength (12 feet), or a full wavelength (20 feet), though the wide spacing imposes mechanical problems that may not be worth the effort for most builders. The 5/8-wave spacing is a good compromise between stacking gain and severe support problems, and is recommended with the materials used here.

The 10-foot lengths of steel masting could be used, with the bottom 8 feet running through the tower bearing to the rotator. A heavier main support is preferable, however, and it is "1-inch water pipe" in this installation. This is iron, about 1 3/8-inch outside diameter, extending about 8 feet out of the tower. The steel masting between the Yagi bays is fastened to the pipe with four TV-type U-clamps, spaced evenly in the overlapping area of the two supports.

The booms are braced to the mast fore and aft, using the longest pieces of element stock left over when the forward directors are cut from 12-foot lengths. Ends of the braces are flattened about one inch, and bent to the proper angle. Outer ends fasten to the booms with two self-tapping screws each. The mast ends are clamped to the support with one TV U-clamp for each pair. This bracing is good insurance against fluttering of the booms and elements, which can cause failures after long periods, even though a structure appears adequately strong.

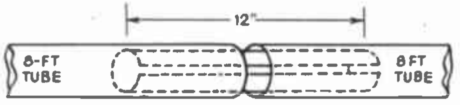


Fig. 22-6 — Details of the boom splices used in the 5-element 50-MHz Yagis. Two 8-foot lengths of 1 1/4-inch tubing are joined to make the 16-foot booms.

Phasing and Matching

A single 5-element Yagi can be fed directly with 50-ohm coax, through the Kirk coaxial gamma-match assembly (Type C6M). This has an adjustable coaxial capacitor, and an arm that connects to the driven element with a sliding clip. Both the capacitor and the point of connection should be adjusted for minimum reflected power, at the center of the frequency range most used. Doing this between 50.2 and 50.4 MHz is suitable for most operators, other than those using fm above 52.5 MHz. Each bay of the stacked pair should be set in this way. The pair can then be fed through a double *Q*-section of 75-ohm coax, as shown in Fig. 22-5.

The Kirk gamma-match assembly has an SO-239 coaxial fitting built in, so the phasing lines are fitted with PL-259 coaxial connectors at both ends. The inner ends attach to a matching coaxial T fitting. The main run of 50-ohm line connects to the center of the T, with a coaxial through-connector and a PL-259 fitting. When the antenna is installed all connectors should be wrapped tightly with plastic tape, and sprayed with Krylon or other protective spray. Dow-Corning Silastic RTV-732 sealant is also good for this use. If the coaxial phasing sections are wrapped around the booms and vertical support a few times, they will just reach the T-fitting, when 12-foot spacing is used.

The lines should be any odd multiple of a quarter-wavelength. If both are the same length the gamma arms should attach to the same side of the driven elements. If there is a half-wavelength difference in the lines, the arms should connect to opposite sides. The length given in Fig. 22-5 is nominal for solid-dielectric coax. If foam-dielectric line is used, the propagation factor given by the maker should be substituted for the 0.66 figure. It is best to grid-dip the line sections for resonant frequency, in any case. Cut the line three inches or more longer than the expected length. Solder a loop of wire between the center pin and the mounting flange of an SO-239 connector. Attach this to the PL-259 connector at one end of the line, and couple it to the dip-meter coil. Trim the line length until resonance at the midpoint of the intended frequency range is indicated. This will not change appreciably when the other coaxial connector is attached.

The line used in the model described is RG-59A/U, which is satisfactory for any amateur power level, so long as the SWR is kept low. Larger coax, such as RG-11A/U, is recommended for a greater margin of safety.

Adjustment and Testing

An individual Yagi can be tested and matched properly by mounting it a half-wavelength above ground, in a large area that is clear of obstructions for many wavelengths. The boom can also be tilted up, until the ground-reflected wave is not a factor in the field-strength meter reading. The SWR bridge should be connected at the gamma match, or an electrical half-wavelength therefrom. Apply low power (not over 10 watts) and adjust the gamma capacitor and the point of connection to the driven element for zero reflected power, at the desired frequency range. The model was flat from 50.2 to 50.4 with just perceptible reflected power showing at 50.1 to 50.5. Adjusted in this way the array should work well up to about 51 MHz.

The best way to check operation of the stacked pair is to support the array with the reflectors resting on the ground and the booms pointing straight up. A 6-foot step-ladder can be used for a temporary support. The bays can be fed separately with 50-ohm line, in this position, and the gamma settings should be the same as obtained in the first check, described above. Now connect the two 75-ohm phasing lines, and insert the SWR bridge in the 50-ohm line to the T fitting. The SWR should be the same as when the bays are fed separately through the 50-ohm line; close to 1:1. The array can be dismantled and reassembled atop the tower, and matching should remain correct.

The matching-phasing system described is useful for any two loads designed for 50-ohm feed. The 5/8-wave spacing is usable with up to at least 6-element bays, though wider bay spacing is needed for maximum gain with long Yagis. Individual antennas intended for 200-ohm balanced feed can be matched with 75-ohm coax in the phasing harness and baluns at each load.

Bay spacing is not critical. Close spacing gives somewhat lower gain, but a very clean pattern. The main lobe gets sharper and larger as spacing is increased, but minor lobes also increase. These take over from the main lobe if spacing of bays is carried too far. The effect of increasing bay spacing is shown graphically in Fig. 8-11 of *The Radio Amateur's VHF Manual*, and associated text.

144 OVER 50

Four phased 144-MHz Yagis are shown mounted above a 50-MHz 6-element Yagi in Fig. 22-7. The latter can be mechanically similar to the 5-element antennas of Fig. 22-4, though this two-band system was built almost entirely by hand. Element spacings are closer than in the 5-element 6-meter arrays, in order to fit 6 elements onto a 20-foot boom. The individual bays of the 2-meter array can be used singly, in pairs, or in the 4-bay system shown. Feed details are given for each application.

6-Element 50-MHz Yagi

The 6-meter elements were designed for light weight, with 1/2-inch tubing for half their length and thin-wall fuel-line tubing inserts for the outer

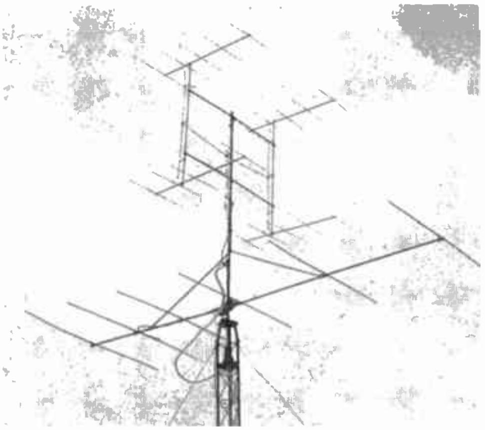


Fig. 22-7 – Antennas for two bands on a single support. Four 5-element Yagis for 144 MHz, top, have one-wavelength spacing each way. The 50-MHz Yagi is set up to make optimum use of 6 elements on a 20-foot boom.

portions. One-piece half-inch elements are equally good, though a bit bulkier. Elements can be run through the boom and held in place with clamps, as in Fig. 22-8, or mounted in Kirk castings. (See 5-element array description.) Lengths are 116, 110 1/2, 105 1/2, 104, 102 3/4, and 101 1/2 inches. Spacings, in the same order, are 36, 36, 42, 56 and 66 inches. The boom is made of two 10-foot aluminum mast sections, braced from above with 3/4-inch tubing. See Fig. 22-8.

The gamma matching was handled in two different ways. A coaxial capacitor and moving arm was hand-made, as shown in Fig. 22-9 using 1/2-inch and 1/4-inch tubes, insulated from one another by plastic sleeves that just fit inside the 1/2-inch fixed portion. The inner tubing can be wrapped with plastic tape to build up the needed thickness, to the same end. The arm is supported at two points with 1-inch ceramic pillars.

A second and simplier matching arrangement uses merely an extension of the main coaxial line, with a 100-pF fixed transmitting-type capacitor in series with the inner conductor and the sliding contact. The matching point was about 20 inches

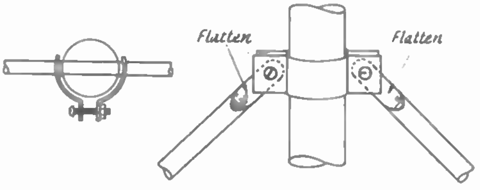


Fig. 22-8 – Elements may be run through a wood or metal boom, and held in place with simple aluminum clamps, left. At the right is a clamp for holding boom braces on the vertical support in the 50-MHz 6-element array.

out from the boom with a 100-pF capacitor. It is suggested that the matching be done first with a variable capacitor, substituting a fixed one when the desired value is found.

An element-mounting clamp no longer available appears in Fig. 22-9. The Kirk 1/2-to-1 1/4-inch element-mounting clamps (see 5-over-5 description) do this job nicely.

5-Element 144-MHz Yagis

An optimum design for 5-element 2-meter Yagis, to be used singly or combined in stacked systems, is shown in Fig. 22-10. Dimensions given work well from 144 to 146 MHz, if the matching is adjusted at 145. Lengths should be reduced 1/4 inch for each megahertz higher center frequency than 145 MHz. The original elements have center sections of 1/4-inch aluminum tubing, with 5/32-inch rod inserts that slide into the center members. One-piece elements of 1/8 to 1/4-inch tubing or rod will work equally well. The larger size will permit fastening in place with self-tapping screws bearing on the elements. For smaller sizes, use a clamp like that of Fig. 22-8. The booms are 3/4- or 1-inch diameter aluminum. Wood dowelling could be used equally well.

Feed Methods: A delta match is used in conjunction with a coaxial-line balun to feed a single 5-element Yagi. Some experimentation with delta dimensions may be required to achieve the best match. (See Fig. 22-1C and detailed description of the delta match earlier in this chapter.) This arrangement makes a fine small Yagi that can be dismantled readily, for carrying about in portable work.

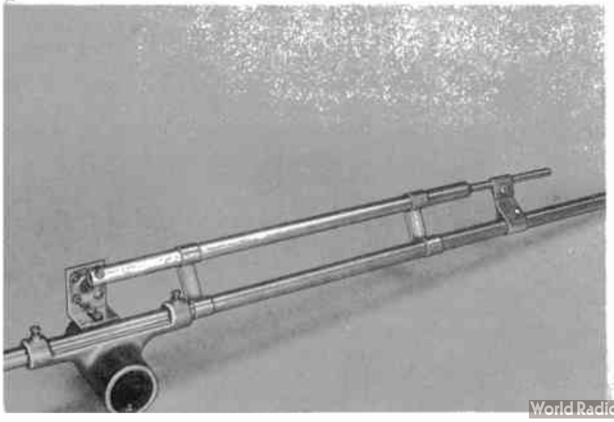


Fig. 22-9 – A hand-made coaxial gamma match for 50-MHz arrays. A 1/4-inch rod or tube 14 inches or longer slides inside a 1/2-inch sleeve that is connected to the coaxial fitting above the boom. The rod slides on plastic sleeves inside the larger section. Separation is maintained with two ceramic pillars mounted with wrap-around clips. Both the coaxial capacitor and the sliding clip between rod and element are adjusted for minimum reflected power in the coaxial line.

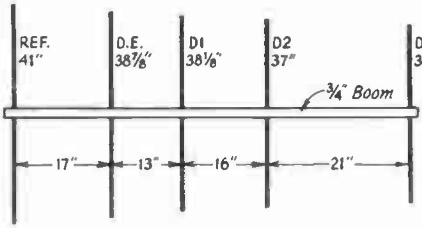


Fig. 22-10 – Optimum design for a 2-meter Yagi, using 5 elements on a 6-foot boom. When used singly, this antenna can be fed as shown in Fig. 22-1C, with 4-inch delta arms connected 3 inches either side of center. The balun loop would be about 27 inches long. With lengths shown, the antenna works well from 144 to above 146 MHz, but gain drops sharply above 147 MHz.

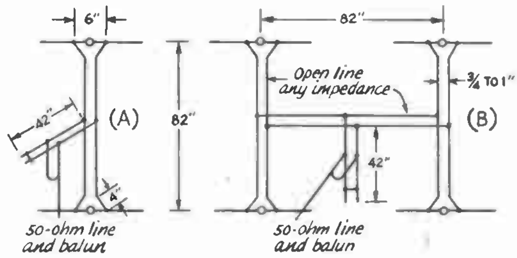


Fig. 22-11 – Stacking details for the 5-element Yagis of Fig. 22-7 and 22-10. The short on the universal stub, and the point of connection of the main transmission line, are adjusted for minimum reflected power in the latter. Balanced line could be connected similarly for the main turn.

Use of two 5-element Yagis with 1-wavelength spacing is shown in Fig. 22-11A. The phasing harness can be any open-wire line, preferably not spaced more than one inch. Delta dimensions are not critical in this application, as the matching is done with the universal stub at the center of the harness.

The 4-bay 20-element system in Fig. 22-7 and 22-11B uses two sets of 5-over-5, connected between centers with another 1-wavelength line. The universal stub is connected at the center of the horizontal section. In each case, the stub length and line-connection point are adjusted for minimum reflected power in the main line.

An interesting phasing method was used in the 4-bay array. Common electric zipcord, available in any hardware store, was split into its two parts. The insulation was left on, and spreaders made of ordinary 1/2-inch wood dowel were used to hold the wires one inch apart. Holes were drilled in these of such size that the zipcord could just be pulled through them. They are held in place with any good cement. If supported with TV-type screweyes that grip the spreaders, such a low-cost line is very durable. The array shown was taken down after two years of use in a very exposed location, and no deterioration was apparent. There was no breakage, even under several heavy ice loads each winter. Using several supports on each harness section is the key to this long life.

The transmission line was switched between the six- and two-meter arrays by means of a waterproofed antenna relay. To avoid the dangers of a 115-volt line run, 6.3-volt transformers were used at each end. This one-line hookup makes it possible to use a single rather expensive line to its fullest potential on two bands.

13-ELEMENT YAGI FOR 144 MHz

Many combinations of element lengths and spacings work well in long Yagis. The 13-element array detailed in Fig. 22-12 is the product of many months of joint experimental work by W2NLY and W6QKI. First described in *QST* for January, 1956, it has been a winner ever since. Elements are 1/8-inch hard-drawn aluminum wire, except for the folded-dipole driven element. This is the step up variety, intended to give a feed impedance of 200 ohms, for feeding with 50-ohm line and a coaxial balun.

The 24-foot boom carries a light load, and can

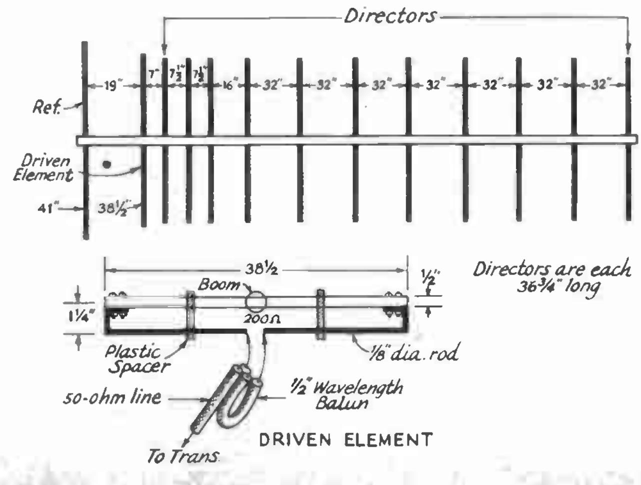
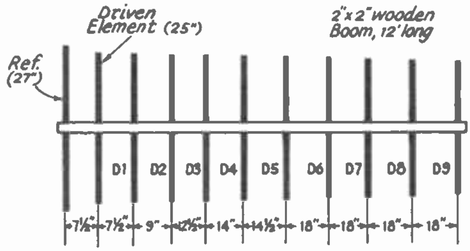


Fig. 22-12 – High-performance long Yagi for 144 MHz, from experimental work by W2NLY and W6QKI. Dimensions are for maximum gain between 144 and 145 MHz.



D1 = 23¼"
D2 = 23⅛"
D3 = 23"
D4 = 22⅞"
D5 = 22¾"
D6 = 22⅝"
D7 = 22½"
D8 = 22⅜"
D9 = 22¼"

Fig. 22-13 – 11-element Yagi for 220 MHz. Dimensions are for maximum gain in the lower 2 MHz of the band. Recommended feed method is a delta match, with universal stub and balun. Delta sides should be about 3 inches, tapped 2 inches either side of the element midpoint.

be made of thin-wall tubing if braced in the manner of the 50-MHz arrays previously described. Elements run through the boom and are held in place with clamps, as in Fig. 22-8. Lengths are for optimum gain between 144 and 145 MHz. Gain drops rapidly above 145.2 MHz. For a center frequency of 145 MHz, cut element lengths 1/8 inch. Broader frequency response can be obtained by tapering element lengths 1/8 inch per element, beginning with the second director.

Effective stacking of such long Yagis requires bay spacing of 1 1/2 to 2 wavelengths. Pairs or pairs of pairs can be fed in the manner of Fig. 22-15, using dimensions of Table 22-I.

11-ELEMENT YAGIS FOR 220 AND 432 MHz

High-gain antennas are almost a necessity for any serious work on 220 MHz and higher frequencies. The 11-element Yagis shown in Figs. 22-13 and 14 were worked out experimentally for maximum gain per element. They are intended primarily to be used in stacked pairs or sets of four, as shown (for 432 MHz) in Fig. 22-15.

Elements are stiff wire or welding rod, 1/8-inch diameter for 220, 3/32 or 1/8 inch for 432. Wood booms are shown, and are recommended for stacked arrays, particulary for 432. Metal booms should be 1/2-inch diameter for 432 and 3/4 to 1 inch for 220. Element lengths should be increased 0.5 to 1 percent if metal booms are used.

Frequency coverage without appreciable loss of gain, and no readjustment of matching, is about 1 percent of the operating frequency. Lengths of elements given are for 220 to 222 MHz and 432 to 434 MHz. Coverage can be extended somewhat higher by readjusting the matching for the desired higher frequency.

Recommended phasing is by open-wire line two wavelengths long each way. No. 12 wire spaced 1/2 to 3/4 inch with Teflon spreaders is ideal. If a metal supporting structure is used, it should preferably be entirely in back of the plane of the reflector elements.

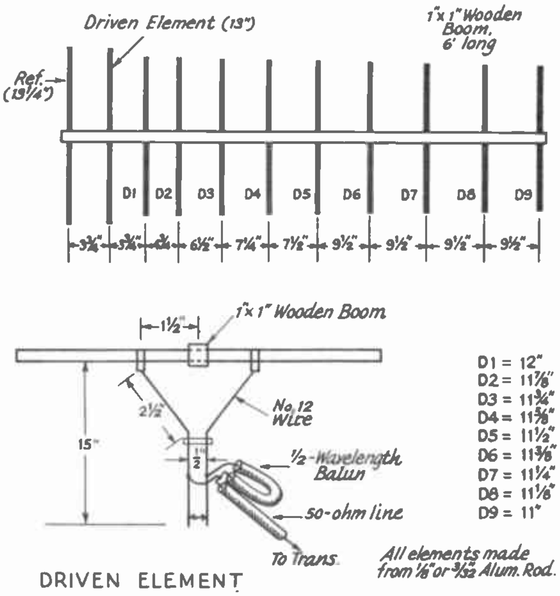


Fig. 22-14 – 11-element Yagi for 432 MHz, designed for optimum performance on a 6-foot boom. Operation should be uniform between 432 and 436 MHz, if the stub matching is adjusted when moving more than one megahertz in frequency.

COLLINEAR ANTENNAS

Information given thus far is mainly on parasitic arrays, but the collinear antenna has much to recommend it. Inherently broad in frequency response, it is a logical choice where coverage of an entire band is wanted. This tolerance also makes a collinear easy to build and adjust for any vhf application, and the use of many driven elements is popular in very large phased arrays, such as may be required for moonbounce (EME) communication.

Omnidirectional Verticals

Two or more half-wave elements mounted in a vertical line and fed in phase are often used to build up some gain, without directivity. A simple omnidirectional collinear of rugged construction is shown in Fig. 22-16. It is made entirely of copper pipe and matching elbow fittings, obtainable from plumbing supply houses and some hardware stores.

Initially the phasing stub was operated in the manner of Fig. 22-1A. When the optimum dimensions were found, the assembly was completed by making the angles with plumbing fittings, and the balun connections with bolts, nuts and star lugs.

Preferably the antenna should be mounted on a wooden support, though the center of the stub can be grounded for lightning protection. Dimensions given are for the upper half of the 2-meter band,

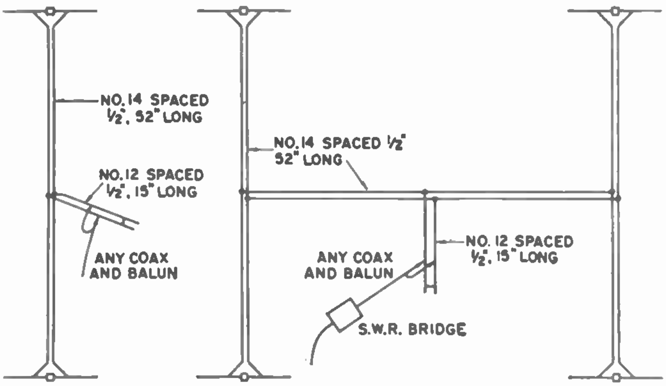


Fig. 22-15 — Phasing methods for using two or four 11-element Yagis for 432 MHz, with 2-wavelength spacing. Universal-stub match permits use of any type of transmission line.

though it works well enough all the way down to 144 MHz.

Any number of radiators can be used, if quarter-wave phasing stubs are connected between them. Commonly an odd number is used, and the center radiator is broken at its midpoint and fed with a universal stub. This type of antenna can be made of wire and strung up in a horizontal position. The pattern is bidirectional when this type of collinear is mounted horizontally.

Large Collinear Arrays

Bidirectional curtain arrays of 4, 6 and 8 half-waves in phase are shown in Fig. 22-17. Usually reflector elements are added, normally at about 0.2 wavelength in back of each driven element, for more gain and a unidirectional pattern. Such parasitic elements are omitted from the sketch in the interest of clarity. Dimensions are not critical, and may be taken from Table 22-1.

When parasitic elements are added, the feed impedance is low enough for direct connection open line or Twin-Lead, connected at the points indicated by black dots. With coaxial line and a balun, it is suggested that the universal stub match, Fig. 22-1A, be used at the feedpoint. All elements should be mounted at their electrical centers, as indicated by open circles in Fig. 22-17. The framework can be metal or insulating material, with equally good results. A model showing the preferred method of assembling an all-metal antenna is pictured in Fig. 22-18. Note that the metal supporting structure is entirely in back of the plane of the reflector elements. Sheet-metal clamps can be cut from scraps of aluminum to make this kind of assembly, which is very light in weight and rugged as well. Collinear elements should always be mounted at their centers, where rf voltage is zero – never at their ends, where the voltage is high and insulation losses and detuning can be very harmful.

Collinear arrays of 32, 48, 64 and even 128 elements can be made to give outstanding performance. Any collinear should be fed at the center of the system, for balanced current distribution. This is very important in large arrays, which are treated as sets of 6 or 8 driven elements each, and fed through a balanced harness, each section of which is a resonant length, usually of open-wire line. A 48-element collinear array for 432 MHz, Fig. 22-19, illustrates this principle.

PLANE AND PARABOLIC REFLECTORS

A reflecting plane, which may be sheet metal, wire mesh, or even closely-spaced elements of tubing or wire, can be used in place of parasitic reflectors. To be effective, the plane reflector must extend on all sides to at least a quarter-wavelength beyond the area occupied by the driven elements. The plane reflector provides high front-to-back ratio, a clean pattern, and somewhat more gain than parasitic elements, but large physical size rules it out for amateur use below 420 MHz. An interesting space-saving possibility lies in using a

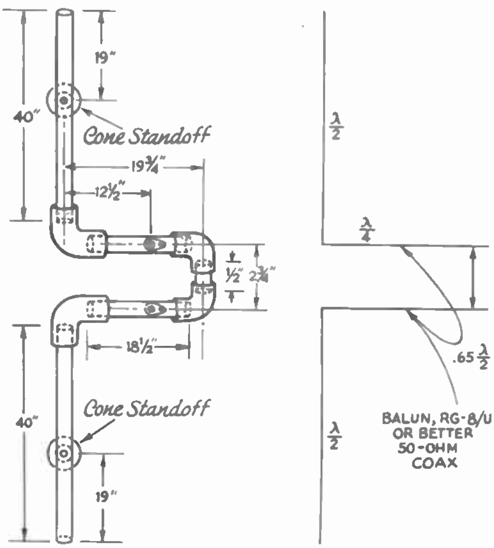


Fig. 22-16 — Rugged 2-meter omnidirectional vertical antenna made entirely of 1/2-inch copper pipe and elbows. The midpoint of the stub can be grounded, for lightning protection.

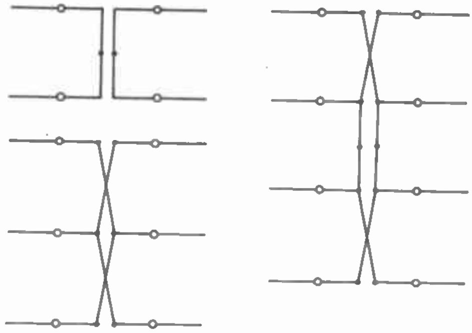


Fig. 22-17 – Element arrangements for 8, 12 and 16-element collinear arrays. Parasitic reflectors, omitted here for clarity, are 5 percent longer and 0.2 wavelength in back of the driven elements. Feed points are indicated by black dots. Open circles are recommended support points. The elements can run through wood or metal booms, without insulation, if supported at their centers in this way. Insulators at the element ends (points of high rf voltage) tend to detune and unbalance the system.

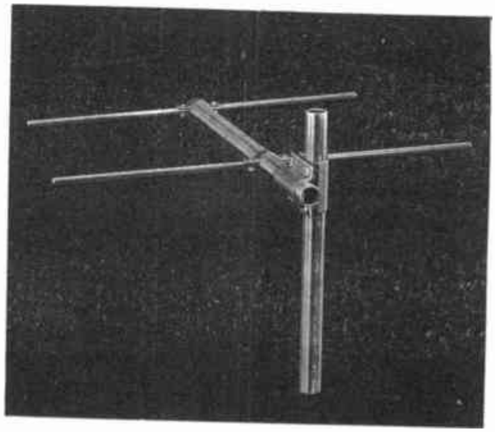


Fig. 22-18 – Model showing recommended method for assembling all-metal arrays. Suitable assembling clips can be cut and bent from sheet aluminum. Supporting structure should be in back of all active elements of the array.

single plane reflector with elements for two different bands mounted on opposite sides. Reflector spacing from the driven element is not critical. About 0.2 wavelength is common.

The reflector can be formed into parabolic shape for a focussing effect, similar to that in a searchlight. Parabolic reflectors must be very large in terms of wavelength. Principles involved in parabolic reflector design are discussed by WA9HUV in *QST* for June, 1971, page 100.

CIRCULAR POLARIZATION

Polarization is described as "horizontal" or "vertical," but these terms have no meaning once the reference of the earth's surface is lost. Many propagation factors can cause polarization change: reflection or refraction, passage through magnetic fields (Faraday rotation) and, satellite rolling, for examples. Polarization of vhf waves is often random, so an antenna capable of accepting any polarization is useful. Circular polarization, generated with helical antennas or with crossed elements fed 90 degrees out of phase, has this quality.

The circularly-polarized wave, in effect, threads its way through space, and it can be left- or right-hand polarized. These polarization "senses" are mutually exclusive, but either will respond to any plane polarization. A wave generated with right-hand polarization comes back with left-hand, when reflected from the moon, a fact to be borne in mind in setting up EME circuits. Stations communicating on direct paths should have the same polarization sense.

Both senses can be generated with crossed dipoles, with the aid of a switchable phasing harness. With helical arrays, both senses are provided with two antennas, wound in opposite directions.

Helical Antenna for 432 MHz

The 8-turn helix of Fig. 22-20 is designed for 432 MHz, with left-hand polarization. It is made

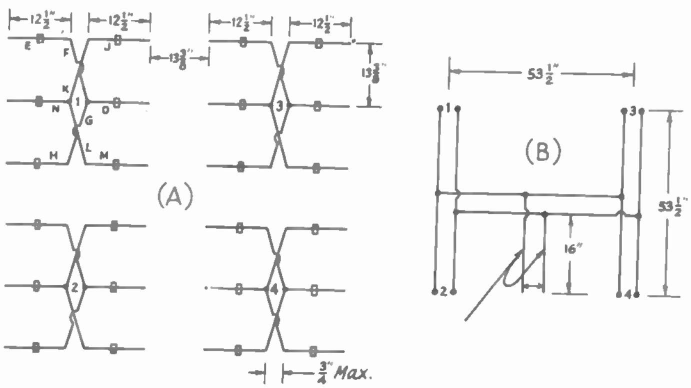


Fig. 22-19 – Large collinear arrays should be fed as sets of no more than 8 driven elements each, interconnected by phasing lines. This 48-element array for 432 MHz (A) is treated as if it were four 12-element collinears. Reflector elements are omitted for clarity. Phasing harness is shown at B.

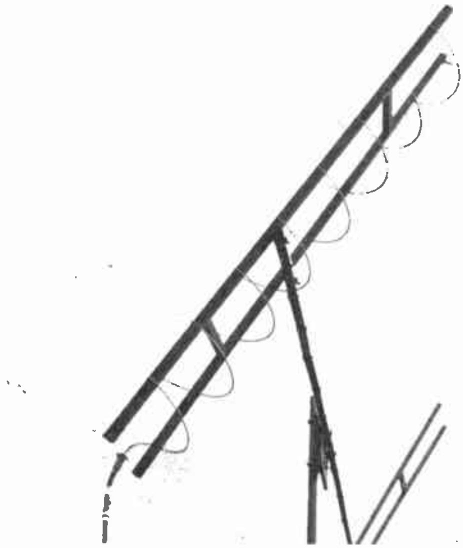


Fig. 22-20 – An 8-turn 432-MHz helical array, wound from aluminum clothesline wire. Left-hand polarization is shown. Each turn is one wavelength, with a pitch of 0.25 wavelength. Feed is with 50-ohm coax, through an 84-ohm Q section.

from 213 inches of aluminum clothesline wire, including 6 inches that are used for cutting back to adjust the feed impedance.

Each turn is one wavelength long, and the pitch is about 0.25 wavelength. Turns are stapled to the wooden supports, which should be water-proofed with liquid fiber glass or exterior varnish. The reflecting screen is one wavelength square, with a Type N coaxial fitting soldered at its center, for connection of the required coaxial Q section.

The nominal impedance of a helical antenna is 140 ohms, calling for an 84-ohm matching section to match to a 50-ohm line. This can be approximated with copper tubing of 0.4-inch inside diameter, with No. 10 inner conductor, both 6 1/2 inches long. With the antenna and transformer connected, apply power and trim the outer end of the helix until reflected power approaches zero.

The support arms are made from sections of 1 X 1 wood and are each 60 inches long. The spacing between them is 8.25 inches, outer dimension. The screen of the antenna in Fig. 22-20 is tacked to the support arms for temporary use. A wooden framework for the screen would provide a more rugged antenna structure. The theoretical gain of an 8-turn helical is approximately 14 decibels. Where both right- and left-hand circularity is desired, two antennas can be mounted on a common framework, a few wavelengths apart, and wound for opposite sense.

A TRANSMATCH FOR 50 AND 144 MHz

The antenna couplers as shown in Fig. 22-21 will permit unbalanced transmitter output lines (50-75 ohms) to be matched to balanced feeders in the 300 to 450-ohm impedance range. Also, "coax-to-coax" matching is possible with this circuit, permitting 50-ohm lines to be matched to 75-ohm lines, or vice versa. In situations where a high SWR condition exists where an antenna is being used in a part of the band to which it has not been tuned, this coupler will enable the transmitter to look into a flat load, thus permitting maximum loading for better efficiency.

Couplers of this type are beneficial in the reduction of harmonic energy from the transmitter, an aid to TVI reduction. It should be possible to realize a 30-dB or greater decrease in harmonic level by using this Transmatch between the transmitter and the feed line. When connected ahead of the receiver as well – a common arrangement – the added selectivity of the coupler's tuned circuits will help to reduce images and other undesired receiver responses from out-of-band signals. It is wise to remember that the use of devices of this kind will not correct for any mismatch that exists at the antenna end of the line. Although it assures a good match between the transmitter and the line, it can only disguise the fact that a mismatch exists at the antenna.

The Circuit

Balanced circuits are used for both bands, Fig. 22-22. Butterfly capacitors are employed to aid in securing good circuit symmetry. The links of each tuned circuit, L2 and L3, are series tuned by single-ended capacitors to help tune out reactance in the line.

Construction

A 4-1/2 X 4-1/2 X 2-inch homemade cabinet houses the 2-meter Transmatch; A Ten-Tec JW-5 is used as an enclosure for the 50-MHz unit. Other commercially made cabinets would be suitable, also. The two tuning controls are mounted in a line across the front of each cabinet. The main coil in each Transmatch is supported by a ceramic standoff insulator on one end and by the connection to the TUNING capacitor on the other. The links are self supporting. The coil taps are effected by bending standard No.6 solder lugs

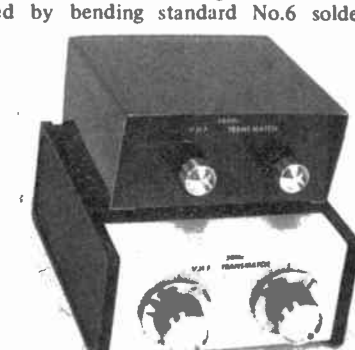


Fig. 22-21 – These 6- and 2-meter Transmatches may be used with powers up to 500 watts. They can be employed with either balanced or unbalanced feeders.

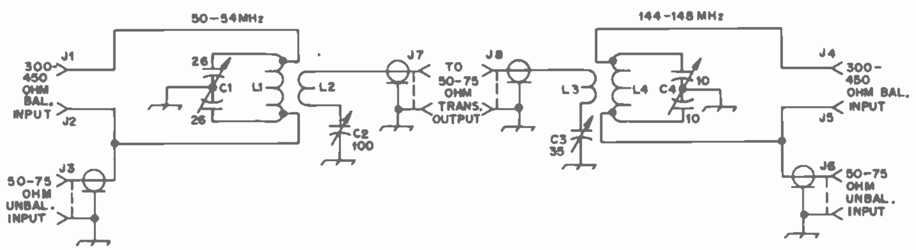


Fig. 22-22 — The schematic diagram of the vhf Transmatches. Capacitance is in pF unless otherwise noted. Resistance is in ohms, k = 1000.

C1 — 26-pF per section butterfly (E.F. Johnson 167-22).
C2 — 100-pF miniature variable (Millen 20100).
C3 — 35-pF miniature variable (Millen 20035).
C4 — 10-pF per section butterfly (E.F. Johnson 167-21).
J1-J4, incl. — Insulated binding post.
J5-J8, incl. — SO-239-style chassis connector.
L1 — 7 turns No. 10 copper wire, 1 1/2-inch dia, spaced one wire thickness between turns. Tap 2 1/2 turns from each end.
L2 — 2 turns No. 14 enam. or spaghetti-covered bare wire, 2-inch dia, over center of L1.
L3 — 2 turns No. 14 enam. or spaghetti-covered bare wire, 1 1/2-inch dia. over center of L4.
L4 — 5 turns No. 10 copper wire, 1-inch dia, spaced one wire thickness between turns. Tap 1 1/2 turns from each end.

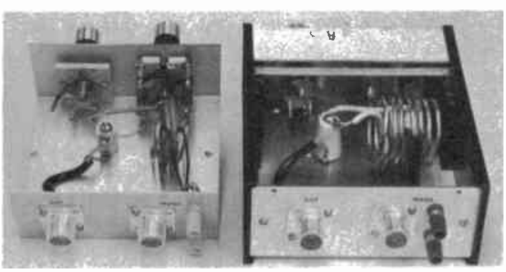


Fig. 22-23 — Inside view of the two Transmatches.

around the coil wire at the proper spots, then soldering the lugs in place. No. 20 bus wire is used to connect the taps of L1 to jacks J1 and J2. When operating coax-to-coax style, a short jumper wire connects J1 to its ground lug, or J4 to its ground lug, depending on the band being operated. The jumper must be removed for balanced-feeder operations.

Operation

Attach the vhf transmitter to J7 or J8 with a short length of coax cable. Connect a balanced feeder to J1 and J2 (for 50-MHz operation), or to J4 and J5 (for 144-MHz operation). A reflected-power meter or SWR bridge connected between the Transmatch and the transmitter will aid in the adjustment process. Adjust C1 and C2, alternately (for 50-MHz operation) for minimum meter reading on the SWR indicator. For 144-MHz operation, tune C3 and C4 in the same manner. Repeat the tuning until no further reduction in reflected power is possible. The meter should fall to zero, indicating a 1:1 match. No further adjustments will be needed until the transmitter frequency is moved 50 kHz or more. The tuning procedure is identical for matching coax to coax. In doing so, however, the antenna feed line (coax) is connected to either J3 or J6 and the shorting strap (discussed earlier) must be connected to J1 or J4. In some situations, it may be possible to get a better match by leaving the shorting strap off.

After the coupler is tuned up, the transmitter power can be increased to its normal level. These units will handle power levels up to 500 watts (transmitter output power) provided the coupler is tuned for a matched condition at all times. Reduced power (less than 50 watts) should be used during initial tune up, thus preventing parts from being damaged by heating or arcing. The coupler should never be operated without a load connected to its output terminals.

AN INEXPENSIVE DIRECTIONAL COUPLER

Precision in-line metering devices that are capable of reading forward and reflected power over a wide range of frequencies are very useful in amateur vhf and uhf work, but their rather high cost puts them out of the reach of many vhf enthusiasts. The device shown in Fig. 22-25 is an inexpensive adaptation of their basic principles. You can make it yourself for the cost of a meter, a few small parts, and bits of copper pipe and fittings that can be found in the plumbing stocks at many hardware stores.

Construction

The sampler consists of a short section of hand-made coaxial line, in this instance of 50 ohms impedance, with a reversible probe coupled to it. A small pickup loop built into the probe is terminated with a resistor at one end and a diode at the other. The resistor matches the impedance of the loop, not the impedance of the line section. Energy picked up by the loop is rectified by the diode, and the resultant current is fed to a meter equipped with a calibration control.

The principal metal parts of the device are a brass plumbing T, a pipe cap, short pieces of 3/4-inch ID and 5/16-inch OD copper pipe, and two coaxial fittings. Other available tubing combinations for 50-ohm line may be usable. The ratio of outer-conductor ID to inner-conductor OD should be 2.4/1. For a sampler to be used with other impedances of transmission line, see Chapter 20 for suitable ratios of conductor sizes. The photographs and Fig. 22-26 just about tell the rest of the story.

Soldering of the large parts can be done with a 300-watt iron or a small torch. A neat job can be done if the inside of the T and the outside of the pipe are tinned before assembling. When the pieces are reheated and pushed together, a good mechanical and electrical bond will result. If a torch is used, go easy with the heat, as an over-heated and discolored fitting will not accept solder well.

Coaxial connectors with Teflon or other heat-resistant insulation are recommended. Type N, with split-ring retainers for the center conductors, are preferred. Pry the split-ring washers out with a knife point or small screwdriver. Don't lose them, as they'll be needed in the final assembly.

The inner conductor is prepared by making eight radial cuts in one end, using a coping saw with a fine-toothed blade, to a depth of 1/2 inch. The fingers so made are then bent together, forming a tapered end, as seen in Fig. 22-26. Solder the center pin of a coaxial fitting into this, again being careful not to overheat the work.

In preparation for soldering the body of the coax connector to the copper pipe, it is convenient to use a similar fitting clamped into a vise as a holding fixture, with the T assembly resting on top, held in place by its own weight. Use the partially prepared center conductor to assure that the coax connector is concentric with the outer conductor. After being sure that the ends of the pipe are cut exactly perpendicular to the axis, apply heat to the coax fitting, using just enough so that a smooth fillet of solder can be formed where the flange and pipe meet.

Fig. 22-24 — Major components of the line sampler. The brass T and two end sections are at the back of the picture. A completed probe assembly is at the right. The N connectors have their center pins removed. The pins are shown with one inserted in the left end of the inner conductor and the other lying in the right foreground.

Before completing the center conductor, check its length. It should clear the inner surface of the connector by the thickness of the split ring on the center pin. File to length; if necessary, slot as with the other end, and solder the center pin in place. The fitting can now be soldered onto the pipe, to complete the 50-ohm line section.

The probe assembly is made from a 1-1/2-inch length of the copper pipe, with a pipe cap on the top to support the upper feedthrough capacitor, C2. The coupling loop is mounted by means of small Teflon standoffs on a copper disk, cut to fit inside the pipe. The disk has four small tabs around

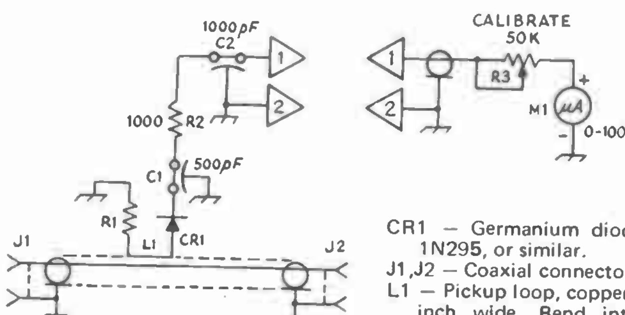


Fig. 22-25 — Circuit diagram for the line sampler.
C1 — 500-pF feedthrough capacitor, solder-in type.
C2 — 1000-pF feedthrough capacitor, threaded type.

CR1 — Germanium diode 1N34, 1N60, 1N270, 1N295, or similar.
J1,J2 — Coaxial connector, type N (UG-58A/U).
L1 — Pickup loop, copper strap 1 inch long × 3/16 inch wide. Bend into "C" shape with flat portion 5/8-inch long.
M1 — 0-100-μA meter.
R1 — Composition resistor, 82 to 100 ohms. See text.
R3 — 50,000-ohm composition control, linear taper.

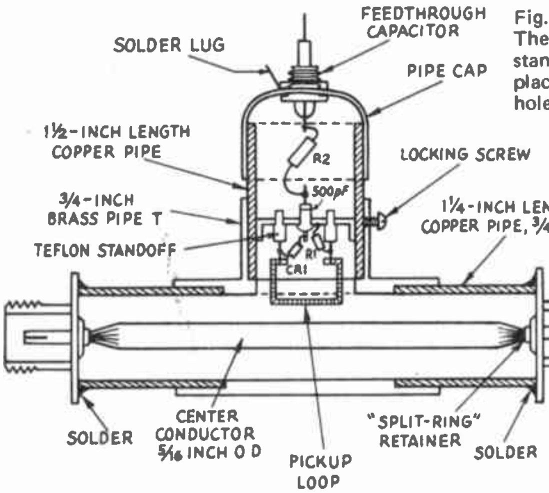


Fig. 22-26 – Cross-section view of the line sampler. The pickup loop is supported by two Teflon standoff insulators. The probe body is secured in place with one or more locking screws through holes in the brass T.

the edge for soldering inside the pipe. The diode, CR1, is connected between one end of the loop and a 500-pF feedthrough capacitor, C1, soldered into the disk. The terminating resistor, R1, is connected between the other end of the loop and ground, as directly as possible.

When the disk assembly is completed, insert it into the pipe, apply heat to the outside, and solder the tabs in place by melting solder into the assembly at the tabs. The position of the loop with respect to the end of the pipe will determine the sensitivity of a given probe. For power levels up to 200 watts the loop should extend beyond the face of the pipe about 5/32 inch. For use at higher power levels the loop should protrude only 3/32 inch. For operation with very low power levels the probe position can be determined by experiment.

The decoupling resistor, R2, and feedthrough capacitor, C2, can be connected, and the pipe cap put in place. The threaded portion of the capacitor extends through the cap. Put a solder lug over it before tightening its nut in place. Fasten the cap with two small screws that go into threaded holes in the pipe.

Calibration

The sampler is very useful for many jobs, even if it is not accurately calibrated, though it is desirable to calibrate it against a wattmeter of known accuracy. A good 50-ohm dummy load is a must.

The first step is to adjust the inductance of the loop or the value of the terminating resistor, for lowest reflected-power reading. The loop is the easier to change. Filing it to reduce its width will increase its impedance. Increasing the cross-section of the loop will lower it, and this can be done by coating it with solder. When the reflected-power reading is reduced as far as possible, reverse the probe and calibrate for forward power, by increasing the transmitter power output in steps and making a graph of the meter readings obtained. Use the calibration control, R3, to set the maximum reading.

Variations

Rather than use one sampler for monitoring both forward and reflected power by repeatedly reversing the probe, it is better to make two assemblies by mounting two T fittings end-to-end, using one for forward and one for reflected power. The meter can be switched between the probes, or two meters can be used.

The sampler described was calibrated at 146 MHz, as it was intended for 2-meter repeater use. On higher bands the meter reading will be higher for a given power level, and it will be lower for lower-frequency bands. Calibration for two or three adjacent bands can be achieved by making the probe depth adjustable, with stops or marks to aid in resetting for a given band. And, of course, more probes can be made, with each calibrated for a given band, as is done in some of the commercially available units.

Other sizes of pipe and fittings can be used, by making use of information given in Chapter 20 to select conductor sizes required for the desired impedances. (Since it is occasionally possible to pick up good bargains in 72-ohm line, you might like to make up a sampler for this impedance.)

Type N fittings were used because of their constant impedance, and their ease of assembly. Most have the split-ring retainer, which is simple to use in this application. Some have a crimping method, as do apparently all BNC connectors. If a fitting must be used that cannot be taken apart, drill a hole large enough to clear a soldering iron tip in the copper-pipe outer conductor. A hole of up to 3/8-inch diameter will have very little effect on the operation of the sampler.

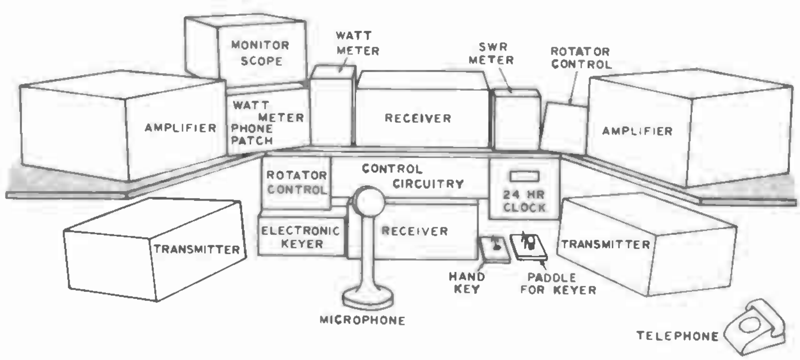
Assembling a Station

The actual location inside the house of the "shack" – the room where the transmitter and receiver are located – depends, of course, on the free space available for amateur activities. Fortunate indeed is the amateur with a separate room that he can reserve for his hobby, or the few who can have a special small building separate from the main house. However, most amateurs must share a room with other domestic activities, and amateur stations will be found tucked away in a corner of the living room, a bedroom, or even a large closet! A spot in the cellar or the attic can almost be classed as a separate room, although it may lack the "finish" of a normal room.

Regardless of the location of the station, however, it should be designed for maximum operating convenience and safety. It is foolish to have the station arranged so that the throwing of several switches is required to go from "receive" to "transmit," just as it is silly to have the equipment arranged so that the operator is in an uncomfortable and cramped position during his operating hours. The reason for building the station as safe as possible is obvious, if you are interested in spending a number of years with your hobby!

CONVENIENCE

The first consideration in any amateur station is the operating position, which includes the operator's table and chair and the pieces of equipment that are in constant use (the receiver, send-receive switch, and key or microphone). The table should be as large as possible, to allow sufficient room for the receiver or receivers, transmitter frequency control, frequency-measuring equipment, monitoring equipment, control switches, and keys and microphones, with enough space left over for the logbook, a pad and pencil. Suitable space should be included for radiogram blanks and a Callbook, if these accessories are in frequent use. If the table is small, or the number of pieces of equipment is large, it is often necessary to build a shelf or rack for the auxiliary equipment, or to mount it in some less convenient location in or under the table. If one has the facilities, a semicircular "console" can be built of wood, or a simpler solution is to use two small wooden cabinets to support a table top of wood or Masonite. A flush-type door will make an excellent table top. Homebuilt tables or consoles can be finished in any of the available oil stains, varnishes, paints or lacquers. Surplus com-



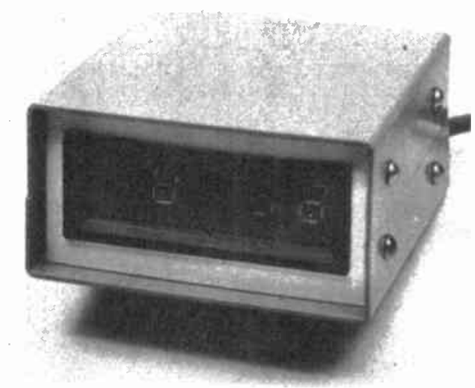
This neatly arranged station belongs to WØTDR in Missouri. The equipment is mounted in a homemade console placed on top of a desk. All controls are easily reachable. A telephone is conveniently located to the right of the operating position. Directly in front of the operator, above the lower receiver, is the control panel which handles antenna and station component switching. This layout is ideal for the right-handed operator.

puter furniture is readily available through various channels. Many of these consoles are ideal for an operating position. Many operators use a large piece of plate glass over part of their table, since it furnishes a good writing surface and can cover miscellaneous charts and tables, prefix lists, operating aids, calendar, and similar accessories.

TIME REFERENCES

Next to frequency, accurate measurement of time is an important part of a station's operating routine. While the matter of a minute or so may not seem like much, it *could* mean the difference between getting a coveted confirmation or waiting until a busy DX or contest operator has time to search for your contact in his log. As a consequence, the station clock should be both accurate and easy to read.

Digital clocks are ideal in both respects and just about any commercially manufactured model should do the job. However, a rather unique module is also available and should appeal to the ham who likes to build his own gear.



Digital Clock Module

Depending upon the module selected, it is possible for the builder to tailor this clock to his individual desires. The entire clock, with the exception of the power transformer and switches, is contained on one small pc board. The dimensions, approximately 1-3/8 X 3 X 1 inches (HWD), allow the builder to package the clock into almost any size container.

At present there are eight different modules available. The different combinations allow selection of: Line frequency (50 or 60 hertz), 12- or 24-hour display and clock/radio or alarm-tone output. The display contains four digits and also four LEDs as indicators. The normal display indicates the time in hours and minutes. Using the external switches it is possible to call up the seconds display, alarm time, and sleep time. The module used for this clock (MA1002A) is a 12-hour format that contains the clock/radio feature. There are two outputs available for controlling external devices. Each output (sleep or alarm) is a positive current source that can turn on an npn transistor for control purposes.

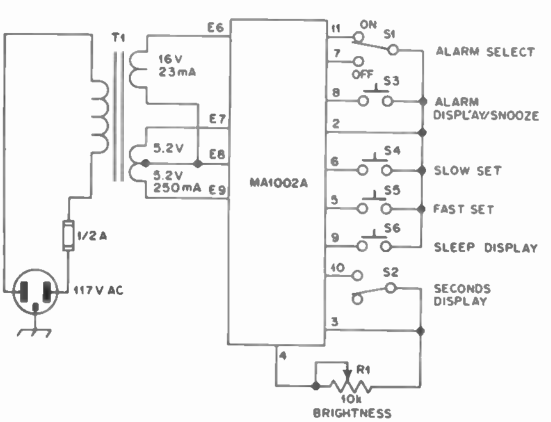
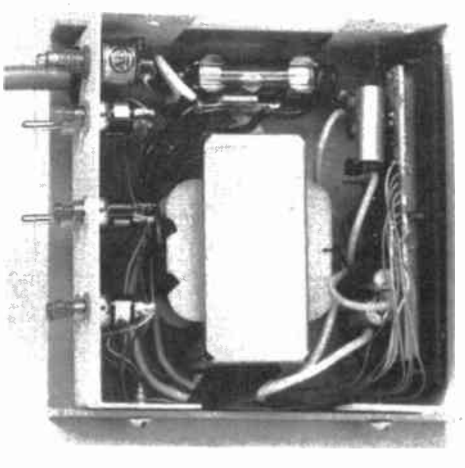


Fig. 1 — Schematic diagram of the clock. T1 is discussed in the text. S1 and S2 are spdt miniature toggle switches. S3 through S6 are momentary switches that are normally open. R1 is a miniature linear-taper potentiometer.



Interior view showing modified transformer.

TABLE I
MA1002 Display Modes

**Selected Display Modes*	*Digit No. 1*	*Digit No. 2*	*Digit No. 3*	*Digit No. 4*
Time Display	10's of Hours & AM/PM	Hours	10's of Minutes	Minutes
Seconds Display	Blanked	Minutes	10's of Seconds	Seconds
Alarm Display	10's of Hours & AM/PM	Hours	10's of Minutes	Minutes
Sleep Display	Blanked	Blanked	10's of Minutes	Minutes

*If more than one display mode input is applied, the display priorities are in the order of Sleep (overrides all others), Alarm, Seconds, Time (no other mode selected).

TABLE II
MA1002 Control Functions

Selected Display Mode	*Control Input*	*Control Function*
*Time	Slow	Minutes Advance at 2 Hz Rate
	Fast	Minutes Advance at 60 Hz Rate
	Both	Minutes Advance at 60 Hz Rate
Alarm/ Snooze	Slow	Alarm Minutes Advance at 2 Hz Rate
	Fast	Alarm Minutes Advance at 60 Hz Rate
	Both	Alarm Resets to 12:00 A.M. (12 hour format)
	Both	Alarm Resets to (0)0:00 (24 hour format)
Seconds	Slow	Input to Entire Time Counter is Inhibited (Hold)
	Fast	Seconds and 10's of Seconds Reset to Zero Without a Carry to Minutes
	Both	Time Resets to 12:00:00 A.M. (12 hour format)
	Both	Time Resets to (0):00:00 (24 hour format)
Sleep	Slow	Subtracts Count at 2 Hz
	Fast	Subtracts Count at 60 Hz
	Both	Subtracts Count at 60 Hz

*When setting time sleep minutes will decrement at rate of time counter, until the sleep counter reaches 00 minutes (sleep counter will not recycle).

Schematic Diagram and Construction

A look at the schematic diagram (Fig. 1) will show the simplicity of the entire clock. All that has to be provided is low-voltage ac and the controlling switches. After checking the catalogs of our parts suppliers, the transformers appeared to be a major stumbling block. It was not possible to find a manufacturer that produced transformers supplying the required voltages. At that point it was decided to rewind an available transformer. A Radio Shack 273-1480 was selected as the candidate. Its original secondary was rated 25.2 V at 1.2 amperes. The original secondary was removed and replaced with two new windings. The first, 106 turns of No. 30 enameled wire, produced 16.0 V ac under load. The second winding was 72 turns of No. 24 enameled wire. This winding measured 10.9 V ac under load. The entire job of rewinding the transformer can be done in less than two hours. A complete discussion of rewinding transformers can be found in the Bcginner and Novice cloumn of *QST*, February, 1970. This article was the basis for rewinding the transformer. The process is not long nor difficult and it produces a transformer that fulfills the requirements at a low cost.

Operation

The complete operation of the external switches and the display readouts are summarized in Tables I and II. This information was obtained from the MA1002 data sheet provided by National Semiconductor Corporation.

POWER CONNECTIONS AND CONTROL

Following a few simple rules in wiring your power outlets and control circuits will make it an easy job to change units within the station. If the station is planned in this way from the start, or if the rules are recalled when you are rebuilding, you will find it a simple matter to revise your station from time to time without a major rewiring job.

It is neater and safer to run a three-wire cable and box from a wall outlet over to the operating table or some central point, than to use a number of adapters and cube plugs at the wall outlet. If

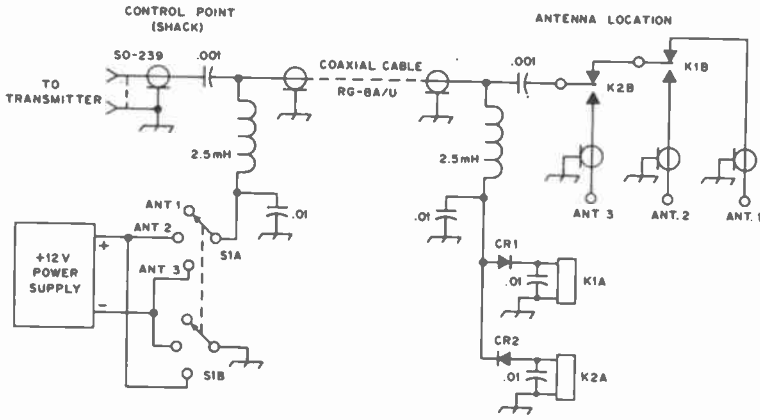


Fig. 23-1 – A remote antenna switching system using low voltage relays handles three different antennas. The coaxial cable is used as the control line. CR1 and CR2 can be any low voltage silicon diodes (Motorola 1N4001 or equiv.). K1 and K2 can be any 12-V dc relays with suitable contact ratings (Potter and Brumfield KA5DG or equiv.). The rotary switch should have at least two sections and three positions.

several outlets are located slightly above the table height, it will be convenient to reach the various plugs. Cable ties can be used for wrapping power cords to maintain a neat arrangement. The operating table should be positioned away from the wall slightly so it will be easier to reach the rear of equipment.

The power wiring should never be overloaded. Check the wire size to assure that the ratings are not exceeded. Consult an electrician for details on power handling capabilities of your house wiring. A 234-V ac line should be available with suitable current ratings. The outlet for this line should be different from the 117-V ac outlets to prevent confusion. A station which runs more than 500 watts input to the transmitter should have this higher voltage line to prevent lights from "blinking" with keying or modulation. It also provides better regulation. A single switch, either on the wall of the shack or at the operating position, should control all of the 117- and 234-volt outlets, except for lights and the line to which the clock is connected. This makes it a simple matter to turn the station to "standby" condition. In case of an emergency, a family member has one switch to shut off power but not the lights. The station equipment normally should be shut off with their own power switches before the main switch is turned off. With equipment left on, turning on the power with the main switch could cause a great surge on the line, which could trip a fuse or circuit breaker.

All power supplies should be fused. Pilot lights or other types of indicators always should be used to tell the operator when the unit is on. All switches for these power supplies should be clearly marked. Even though *you* may know the different functions of the control panel or power supply, a family member may not, and it is important that this vital information be available in case of trouble. In high voltage power supplies, it is recommended that an autotransformer be used in the primary circuit, aside from the power switch, to maintain better control of the high voltage. It also reduces the initial surge of current in the line caused by charging filter capacitors.

SWITCHES AND RELAYS

It is dangerous to use an overloaded switch in power circuits. After it has been used for some time, it may fail, leaving the power turned on even after the switch is set to the "OFF" position. For this reason, large switches or relays with adequate ratings should be used to control the plate power supply.

Any remote-control circuitry should be powered from low voltage. It is dangerous to have 117-V ac controlling remote antenna relays mounted atop a tower. It is recommended that low dc voltages be used for all control systems. One 12-V dc power supply of suitable current rating could be used to handle all control circuitry. As a back-up power source, an automobile battery could be tied in parallel with this supply in case of a power failure. Relay contacts used for antenna switching or rf switching should be rated at least 10 amperes, which will handle two kilowatts. A basic diagram of a remote antenna switching system is shown in Fig. 23-1. The coaxial cable is used to carry the control voltage. The two diodes provide proper operation of either relay.

The nature of the send-receive control circuitry depends almost entirely on the particular station equipment. It is impossible to list here anything but the broadest principles to follow. Commercially manufactured equipment usually has a section of the instruction book devoted to this point. In many cases the antenna-transfer relay is included in the transmitter so that the antenna is directly connected to the transmitter and a separate cable is connected from the transmitter to the receiver. When the transmitter is "on" the relay transfers the antenna to the transmitter output circuit.

EQUIPMENT INTERFACE

As the station grows in complexity, it is important to maintain a unique cabling system. The use of standard cable connectors makes the station components flexible. For low power rf or af, phono plugs and jacks are adequate. High power or voltage requires a higher quality component. A handy device that can simplify the ever present interfacing problem is a patch board with several different types of connectors. While experimenting or changing the station layout, this board can be quite helpful.

Audio patching is the most common situation the amateur encounters. The addition of a tape recorder or another aid to the station should be a simple process. Some tape recorder audio-output circuits are low impedance and could, without suitable coupling, undesirably load the circuit that is being interfaced. A coupling technique often used is that of a resistor (100KΩ) and blocking capacitor (.001 μF) in a series combination. Experimentation is necessary until the circuits are properly matched. The transmitted signal quality of the two units operating in unison should be checked thoroughly.

Often it is convenient to have another headphone jack for a visitor. An audio splitter is shown in Fig. 23-2 that will handle this function. The use of the two potentiometers allows each listener to set his own audio level. If the operator desires to listen to two receivers, at the same time or individually, the reverse of the described system and appropriate switch contacts are required.

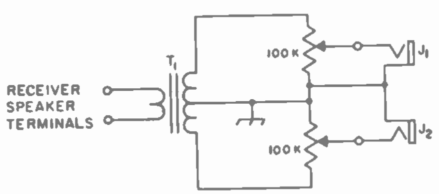


Fig. 23-2 — Diagram of the headphone splitter. The transformer, T1, is a universal output type. J1 and J2 are phone jacks. This circuit allows two sets of headphones to be operated from one receiver; each channel has its own volume control.

The amateur station can become quite sophisticated. As an aid to the operator and any one else within the family, a written record of all wiring is essential. Diagrams of the station wiring, ac voltage lines, rf and af cabling will reduce troubleshooting time or redesigning of the station. Documentation of all changes in antennas, transmitters, receivers, or amplifiers will keep the operator from going over the same road again.

SAFETY

Of prime importance in the layout of the station is the personal safety of the operator and of visitors, invited or otherwise, during normal operating practice. If there are small children in the house, every step must be taken to prevent their accidental contact with power leads of any voltage. A locked room is a fine idea, if it is possible; otherwise housing the transmitter and power supplies in metal cabinets is an excellent, although expensive solution. Lacking a metal cabinet, a wood cabinet or a wooden framework covered with wire screen is the next-best solution. Many stations have the power supplies housed in metal cabinets in the operating room or in a closet or basement, and this cabinet or entry is kept locked – with the key out of reach of everyone but the operator. The power leads are run through conduit to the transmitter, using ignition cable for the high-voltage leads. If the power supplies and transmitter are in the same cabinet, a lock-type main switch for the incoming power line is a good precaution.

An essential adjunct to any station is a *shorting stick* for discharging any high voltage to ground before any work is done in the transmitter. Even if interlocks and power-supply bleeders are used, the failure of one or more of these components may leave the transmitter in a dangerous condition. The shorting stick is made by mounting a small metal hook, of wire or rod, on one end of a dry stick or bakelite rod. A piece of ignition cable or other well-insulated wire is then run from the hook on the stick to the chassis or common ground of the transmitter, and the stick is hung alongside the transmitter. Whenever the power is turned off in the transmitter to permit work on the rig, the shorting stick is first used to touch the several high-voltage leads (plate rf choke, filter capacitor, tube plate connection) to insure that there is no high voltage at any of these points..

Some items which should be included in the station for safety reasons are a fire extinguisher and flashlight. Both should be convenient to reach. The fire extinguisher must be a carbon dioxide type to be effective in electrical fires. The flashlight batteries should be checked regularly. The extinguisher should likewise be inspected on a regular basis. A carbon dioxide type of extinguisher is recommended because it will cause the least amount of damage to equipment.

Family members should be instructed in the use of mouth-to-mouth resuscitation. A sign posted in

If space is available, a neat console can be constructed to house various types of station components. Surplus computer furniture can be used as well. Access to the equipment is through the back of the console. This station belongs to W7VRO.

ASSEMBLING A STATION

Voice operated control (VOX) used in conjunction with a microphone placed on a boom makes operating a nearly "hands-off" affair. This arrangement enables the operator, WB6DSV, to handle paperwork and watch meters and other important controls. This station is owned by W6OKK.

the station describing the necessary procedures to be followed in the event of an emergency should be pointed out to the family. Telephone numbers of the local police, fire department, and doctor should be included on this sign.

Fusing

A minor hazard in the amateur station is the possibility of fire through the failure of a component. If the failure is complete and the component is large, the house fuses will generally blow. However, it is unwise and inconvenient to depend upon the house fuses to protect the lines running to the radio equipment, and every power supply should have its primary circuit individually fused, at about 150 to 200 percent of the maximum rating of the supply. Circuit breakers can be used instead of fuses if desired.

Wiring

Control-circuit wires running between the operating position and a transmitter in another part of the room should be hidden, if possible. This can be done by running the wires under the floor or behind the base molding, bringing the wires out to terminal boxes or regular wall fixtures. Such construction, however, is generally only possible in elaborate installations, and the average amateur must content himself with trying to make the wires as inconspicuous as possible. If several pairs of leads must be run from the operating table to the transmitter, as is generally the case, a single piece of rubber- or vinyl-covered multiconductor cable will always look neater than several pieces of rubber-covered lamp cord, and it is much easier to sweep around or dust.

Solid or standard wire connected to a screw terminal (ac plug, antenna binding posts) should either be "hooked" around a *clockwise* direction, or, better yet, be terminated in a soldering lug. If the wire is hooked in a counter-clockwise position, it will tend to move out from under the screw head as the screw is tightened.

The antenna wires always present a problem, unless coaxial-line feed is used. Open-wire line from the point of entry of the antenna line should always be arranged neatly, and it is generally best to support it at several points. Many operators prefer to mount any antenna-tuning assemblies right at the point of entry of the feed line, together with an antenna changeover relay (if one is used), and then link from the tuning assembly to the transmitter can be made of inconspicuous coaxial line. If the transmitter is mounted near the point of entry of the line, it simplifies the problem of "What to do with the feeders?"

The station components which are located outside must be as safe as the arrangement in the shack. All antenna structures should be protected so that no one will be injured. There should be no low hanging wires or cables. A guard around a tower base is important to keep small children from climbing it. Several ways of protecting the tower base are possible. Cutting a sheet of 1/2-inch plywood lengthwise into three pieces and placing hinges on two edges and pad lock on the third edge will allow the entire structure to be stood up and wrapped around the tower base. The pad lock is essential. Other methods use hardware cloth (heavy mesh) with holes too small to get feet or hands through. Vertical antennas should be protected in a similar fashion, except use a wooden structure or fence.

Open-wire line should be insulated where it can be reached by someone. All control cables or other cables, if possible, should be buried underground or placed high enough so as not to be reached. If antenna work is planned, all cables leading to the tower should be disconnected and *power must be shut off.* Rotator controls should be unplugged. Any electrical wiring or contacts which are exposed to the outdoor environment must be protected from the weather. A water-tight box or a plastic bag will provide such protection. Corrosion to electrical contacts can cause TVI or RFI, poor connections, or losses in vital circuitry. Another

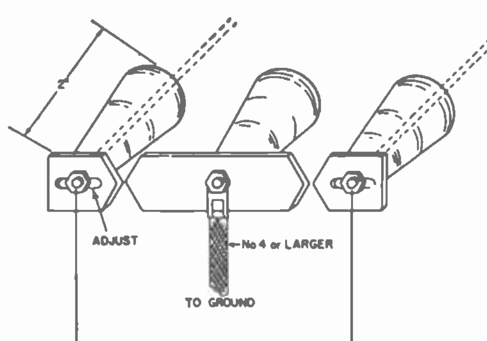


Fig. 23-3 — A simple lightning arrester made from three stand-off or feedthrough insulators and sections of a 1/8 × 1/2-inch brass or copper strap. It should be installed in the open-wire or Twin-Lead line at the point where it is nearest the ground outside the house. The heavy ground lead should be as short and direct as possible. Gap setting should be minimum for transmitter power.

consideration for control cables is rf bypassing. A strong rf field can cause a circuit to be actuated which could disrupt normal operation.

Where guyed towers are used, the guy wires should be arranged so as not to cause danger to someone walking through the area. If this is not possible, planting a shrub or tree near the guy anchor will tend to keep people clear of the vicinity.

LIGHTNING AND FIRE PROTECTION

The National Electrical Code (NFPA No. 70) adopted by the National Fire Protection Association, although purely advisory as far as the NFPA is concerned, is of interest because it is widely used in law and for legal regulatory purposes. Article 810 deals with radio and television equipment, and Section C treats specifically amateur transmitting and receiving stations. Some pertinent paragraphs are reprinted below.

810-13. Avoidance of Contacts with Conductors of Other Systems. Outdoor antenna and lead-in conductors from an antenna to a building shall not cross over electric light or power circuits and shall be kept well away from all such circuits so as to avoid the possibility of accidental contact. Where proximity to electric light and power service conductors of less than 250 volts between conductors cannot be avoided, the installation shall be such as to provide a clearance of at least two feet. It is recommended that antenna conductors be so installed as not to cross under electric light or power conductors.

810-15. Grounding. Masts and metal structures supporting antennas shall be permanently and effectively grounded, without intervening splice or connection.

810-56. Protection Against Accidental Contact. Lead-in conductors to radio transmitters shall be so located or installed as to make accidental contact with them difficult.

810-57.. Lightning Arrestors – Transmitting Stations. Each conductor of a lead-in for outdoor antenna shall be provided with a lightning arrestor or other suitable means which will drain static charges from the antenna system.

Exception No. 1. When protected by a continuous metallic shield which is permanently and effectively grounded.

Table 810-52

Size of Amateur-Station Outdoor Antenna Conductors

Material	Minimum Size of Conductors When Maximum Open Span Length Is Less than 150 feet	Over 150 feet
Hard-drawn copper	14	10
Copper-clad steel, bronze or other high-strength material	14	12

Exception No. 2. Where the antenna is permanently and effectively grounded.

In some areas the probability of lightning surges entering the home via the 117/230 volt-line may be high. A portion of the lightning surges originating on an overhead primary feeder can pass through the distribution transformer by electrostatic and electromagnetic coupling to the secondary circuit, even though the primary is protected by distribution-class lightning arresters. Radio equipment can be protected from these surges by the use of a "secondary service lightning arrester." A typical unit is the G.E. Model 9L15CCB007, marketed as the Home Lightning Protector. It is mounted at the weatherhead or in the service entrance box.

The best protection from lightning is that of completely disconnecting all equipment from antennas, and all ac receptacles. Eliminate the possible paths for any lightning stroke. Rotator cables or any other control cable from the antenna location should be disconnected during severe electrical storms.

Experiments have indicated that a high vertical conductor will generally divert to itself direct hits that might otherwise fall within a cone-shaped space of which the apex is the top of the conductor and the base a circle of radius approximately two times the height of the conductor. Thus a radio mast may afford some protection to low adjacent structures, but only when low-impedance grounds are provided.

This modern station belongs to JA1BRK which is set up to operate the hf and vhf bands. The equipment most often used is on the lower shelf, while the upper shelf holds auxiliary apparatus used for monitoring other frequencies. The large overhead lamp is especially useful.

Operating a Station

Good on-the-air operating practices are important to every amateur for at least three good reasons: to assure compliance with regulations, to permit a large volume of activity to be conducted as efficiently and as simply as possible, and as a matter of personal pride and competence. Good practices is a very bewildering subject at first to many new amateurs, but as in so many other fields, it soon becomes apparent that there is a sound basis of custom and tradition which has produced a body of standard practices. These have evolved over more than a half-century of experience. One of the League's important functions has been to formalize, to foster and to encourage good standard practices so that they have become universal and accepted. Some of our standard practices go back a long time; others have been developed to meet changing circumstances, requirements and technology.

It used to be that one standard was all that was required. Today, things are different. There are standard operating practices for cw, voice, RTTY and repeaters, with additional standards for ATV not too far away. Those for cw and voice are pretty firmly established, but RTTY is newer and repeater operation newer still. Your League will take a crack at all of them. If its recommendations don't "take hold," they will be changed until they become acceptable to a majority in a particular operating specialty. This has been the pattern on cw and phone and will be the pattern on RTTY, repeaters, satellites and whatever else comes along in the future. Operating is better than 50% of most amateurs' lives. Better learn to do it *right*.

Initially, we'll talk about phone and cw, because they can be covered together. RTTY and repeaters will be handled separately.

ESTABLISHING A CONTACT

The best way to do this, especially at first, is to *listen* until you hear someone calling CQ, and call *them*. This requires a little patience, but that's something else all amateurs must learn if we are to share our bands in harmony. Tune around near your own frequency. If you hear a CQ, put your vfo on that frequency (*without* putting a signal on the air), wait until he indicates he is listening, then call him, thus: "W6ZRJ, W6ZRJ, this is W7PGY, W7 Papa Golf Yankee calling, Over" On cw: W6ZRJ W6ZRJ DE W7PGY W7PGY $\overline{AR}$. If no answer (to anyone) this may be repeated; brief, repeated calls are preferred to long drawn out ones. Chances are, if he is to hear you at all, he will hear your first brief call; most amateurs seldom tune far from their transmitting frequency to listen after a CQ. Note the ending signals. These have a special significance of their own to indicate to a casual listener the "status of the contact."

In answer to your call (assuming you are heard), the called station will reply: "W7PGY from W6ZRJ, roger . . ." and then go into conversation. On cw, it would be W7PGY DE W6ZRJ R . . . . That "roger" (R) means that he has received your call correctly. That's *all* it means – RECEIVED. It does not mean correct, I agree, I will comply. It is not sent unless everything was received correctly. Note also that "roger" is the phonetic equivalent of the letter R only in this usage. The regular phonetic for R is "Romeo."

Perhaps W6ZRJ heard W7PGY but did not catch his call. In this case, he might come back with "The W7 station, please repeat your call, this is W6ZRJ, over." On cw: QRZ? W7? DE W6ZRJ $\overline{AR}$. The presence of interference (QRM) and atmospherics (QRN) in the amateur bands makes use of this procedure fairly frequent. The contact (QSO) can then continue. Please note the FCC requirements on identification (97.87).

CALLING CQ

If you hear no CQ, you may wish to make such a call yourself. Refrain from CQing unless you are willing to establish contact with whoever calls. CQ means "I wish to contact *any* amateur station." If this is not your desire, then don't CQ, or be specific in doing so. A CQ call can be somewhat longer than a call to a specific station, because you are trying to attract the attention of casual listeners, including those tuning around looking for someone to call. However, please avoid the common operating discrepancy of calling CQ endlessly; it clutters up the air and drives off potential "customers." The average call would go something like this: "Hello CQ, CQ, CQ, calling CQ, this is WØPAN, W zero Papa Alpha November, Bloomington, Minnesota, calling CQ and listening, go." On cw: CQ CQ CQ DE WØPAN WØPAN WØPAN K. After a brief standby for replies, if no

OPERATING ABBREVIATIONS AND PREFIXES

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 send faster? 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.

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 R-S-T 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 a.c. or less, very rough and broad.
2–Very rough a.c., very harsh and broad.
3–Rough a.c. tone, rectified but not filtered.
4–Rough note, some trace of filtering.
5–Filtered rectified a.c. 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., 469C). If it has clicks or noticeable other keying transients, add K (e.g., 469K). Of course a signal could have both chirps and clicks, in which case both C and K could be used (e.g., RST 469CK).

one answers and the frequency is still clear, you can try again. Short calls and frequent standbys are the best way to establish contact with the minimum QRM. This kind of procedure is easy to use when using VOX or keying through your VOX relay, or using cw break-in procedure.

THE QSO

During the contact, be sure to observe the FCC identification rules (see ARRL *License Manual*) Aside from that, there are no legal limits to what you can talk about, although it is recommended that controversial subjects connected with politics and morality be avoided. Keep everything on a friendly and cordial level, remembering that the conversation is not private and many others, including possibly members of the lay public, may be listening. Try to avoid the habitual utterances, procedures and inanities which so often make amateur radio contacts boring – things such as the drawn out 'ahhhhhh' to keep the VOX relay closed, or repeated "double dash" (dahdididididah) sign on cw, or hackneyed expressions such as "there" (referring to the other fellow) and "here" referring to yourself, or "we" when you mean "I." Both on cw and voice it is possible to be informal, friendly and conversational, and this is what makes an amateur radio QSO enjoyable. During the QSO, when you stand by the recommended signal is "go only" on voice, $\overline{\text{KN}}$ on cw, meaning that you want only the contacted station to come back to you. If you don't mind someone else breaking in, just "go" or K is sufficient. Of course, using VOX or break-in the conversation can proceed as it would face to face, without ending signals after each transmission; this is more normal in a voice contact than in a cw QSO.

ENDING THE QSO

When you decide to end the contact, *end* it. If the other fellow indicates a desire to end it, don't keep on talking, don't say "I won't hold you," then hold him. Express your pleasure at having contacted him and sign out, thus "W1QV from W6KW, clear." If you don't want further contacts, say "clear and leaving the air." On cw, it's $\overline{\text{SK}}$ W1QV DE W6KW, and, if leaving the air, CL.

All these things establish amateur radio as a cordial and fraternal hobby at the same time they foster orderliness and denote organization. Most of them have no legal standing; FCC regs say little about our internal procedures. The procedures we ourselves adopt are even more important than that, because they indicate that we are not just a bunch of hobbyists playing around in random fashion, but that we are an established communications *service* with distinct and distinctive procedures tailored to our special needs.

COURTESY

One thing that is considered the height of ill manners and "liddy" procedure in amateur radio is to tune up or make any transmission on a frequency which is already occupied. In some cases this is necessary, in others inadvertent; but it should always be avoided where possible. For example, if you are committed to a legal one-way transmission or schedule with a friend on a certain frequency at a certain time, it is sometimes unavoidable to cause temporary inconvenience to a going contact or even a net. In another situation, you may not hear another station on the frequency because of "skip," in which case an inquiry "Is the frequency in use?" or, on cw, the Morse letter C (didit dit) should bring a response if you are interfering with a station which you cannot hear. Use the same procedure in tuning up your antenna (use a dummy antenna for testing your rig) – don't *ever* fire up the rig and start tuning it without first turning on the receiver and checking the frequency. The amateur bands are crowded; consideration for the other guy will make things better for everybody.

RTTY PROCEDURES

On radioteletype, the methods of transmission and reception are somewhat different, so slightly different procedures are required. Voice is seldom a "written" mode and cw need not be, but RTTY always is. You type your transmission on a keyboard and it is received at the other end in printed form. Thus, most cw abbreviations can be used to good effect. In addition, such things as line feeds and carriage returns must be considered, as well as shifts for "letters" and "figures." These are nonprinting functions nevertheless essential for teleprinter operation.

Because of wide variations in RTTY machines, different mechanical procedures can often be used, but if you don't know the machine at the other end it is best to assume that it has none of the refinements.

As in other operating, the best thing to do is *listen*. The typical beadle-beadle of RTTY is familiar enough that it can be tuned in with an ordinary communications receiver, then put through the converter to copy on your printer. Some typical calls can be identified just by their sound, such as RY (the RTTY "test") and CQ and even your own call. The procedure is much the same as for cw – zero your vfo while copying and call your station on the same frequency. Even though he finishes his CQ with a carriage return (CR) and line feed (LF), it is a good idea to get into the habit of transmitting these functions, to "clear the machine." Thus: (2CR) (LF) K6DYX K6DYX K6DYX DE W1AW W1AW $\overline{\text{AR}}$ (2CR) (LF).

To initiate a CQ, find an unused point in the band, activate your carrier and transmit: (2CR) (LF) CQ CQ CQ DE K6DYX K6DYX K6DYX K (CR) (LF).

During the QSO, when you come to the end of a line (or the end-of-line indicator on tape equipment), send 2CR, LF, 2LTRS. That is, after your carriage return and line feed at the end of a line, the two nonprinting "letter" pulses serve to allow sluggish machines to get ready for the next line, and take less than a second to send. This is especially important with tape transmissions at the higher machine speeds – 75 and 100 wpm.

Most stations equipped for RTTY are also equipped with tape equipment. While RTTY can be sent manually from a keyboard, the use of tape for material which can be prepared ahead of time is much preferable, since it allows the machine to run at an even speed, faster than it could be typed by hand even by an expert typist. The tape is punched on a perforator and fed into a transmitter-distributor (TD) which is motor-driven. Thus, CQ calls or other prepared text (including message traffic) can be made up in advance. It is also fairly common practice to punch tape in ordinary QSOs, keeping ahead of the TD with the perforator. Many operators start punching their reply tape while they are still receiving from the operator at the other end, thus getting ahead far enough so that even if their typing speed is below the speed of the machine (usually 60 wpm) there is enough leeway to allow for the difference. Taped transmissions have no pauses, which can be irksome in manual transmissions.

RTTY equipment operates at different speeds and with different frequency shifts, depending on the sophistication of the equipment. Most amateurs, however, operate at a standard 60 wpm and 850-Hertz shift, and those with 100 wpm and 170-Hertz shift capability can usually switch to the standard. The considerate RTTY operator will be glad to do so whenever called upon, just as a considerate cw operator will slow down to the speed of his QSO.

REPEATER OPERATING

Although repeater operation is generally voice operation, it has some ramifications that are not present in the type of operation used in direct (i.e. not through a repeater) contact on phone. Most repeaters are of the "open" type where anyone with appropriate equipment operating on the repeater's input and output frequencies can participate. Such repeaters usually are accessed simply by depressing the mike button. Some "machines" have limited access, such as by means of a tone, a series of tones or pulses, or some other means to prevent their being triggered by a casual signal.

The primary purpose of repeaters is to extend the coverage for mobile and hand-held units. Fixed-station operation should be held to a minimum. Repeaters lend themselves very well to public service communications such as highway-accident reporting, and emergency-preparedness activities.

A repeater has to be built or purchased by somebody, installed by somebody, and maintained by somebody, usually at considerable expense and trouble. Sometimes this "somebody" is an individual but more often it is a group, either organized for the purpose or undertaking repeater operation as an additional club project. So a first point of repeater operating, not exactly an on-the-air concept, is to lend some kind of support to the group or individual that sponsors the repeater you use regularly.

Here are a few "dos" and "don'ts" put forward by repeater groups that may serve as useful guidelines for repeater operation:

1) Monitor the repeater you plan to use. Each system has its own peculiarities. Don't "key up" a repeater until you're familiar with its operation.

2) Identify properly. When operating mobile, you're required to indicate the call area you are in. Thus, "This is WA1RDX mobile one" would be proper. It is considered poor practice (indeed illegal) to key a repeater without identifying yourself.

3) When desiring to make a contact, all that is necessary is to indicate that you are on frequency. On some machines this may be accomplished by "This is WA1RDX mobile one monitoring." On others, standard practice calls for a single CQ followed by identification. Never send a long CQ; any respondent will be listening on frequency and hear the short call.

4) Keep transmissions short and thoughtful. Don't monopolize the repeater. Most repeaters go off automatically (time out) after a certain length of transmission (usually three minutes or less) and must be rekeyed. Remember, what you say may be monitored by many listeners using public-service band receivers. Don't give a bad impression of ham radio.

5) During a repeater contact, always pause a few seconds before transmitting to allow other stations access to the repeater. Someone may have an emergency to report or priority traffic.

6) Don't break into a contact unless you have something to add. Interrupting is no more polite on the air than it is in person.

7) Use simplex (i.e. direct contact, not through a repeater) operation whenever possible. This frees the repeater for use by stations unable to communicate directly.

8) Use the minimum power necessary to maintain communications. Not only is this an FCC requirement; it's also common courtesy.

9) Many repeaters have autopatch facilities, which is an interconnection between the repeater and the telephone system, to provide a public service. It is strictly forbidden to use the autopatch for anything that could be construed as business communications. Nor should the autopatch be used to simply avoid a toll call. Do not use an autopatch where regular telephone service is available. Abuses of autopatch privileges can lead to their loss.

The ARRL makes available an annually revised repeater directory listing all repeaters which have been registered. For details on how to obtain a copy, check recent issues of *QST*.

CW PROCEDURE

Cw operating procedure has been developing for over a century, for our present International (Continental) Code had its beginnings on the telegraph wire lines. There is more to talk about in cw procedures than any other mode for this reason, not because it is the most popular mode.

Phone many years ago outstripped cw as the most popular mode. But cw is far from dead. A listen to a rare DX pileup in the cw bands, or the cw section of any contest will demonstrate that conclusively. And it has many advantages over any other mode. Any amateur who avoids the use of cw because he is too lazy to become proficient enough in the code to realize its full benefits is missing almost half of amateur radio pleasure.

Good Sending

In many ways. cw can be compared with the spoken word. For the proficient cw man, it is indeed equivalent to this. But just as enunciation must be precise for best understanding in speaking, proper character formation and spacing is required in sending the code. And the learning processes are also similar. The beginning cw operator is subject to the same stresses and pressures as the child learning to talk, and can learn bad habits. He becomes subject to outside influences to his own possible detriment in everyday operating

Actually, it is far easier to learn code today than it was, say, forty years ago when nearly all amateur operation was by cw, because there are more helps available. On the other hand, there is less reason to learn it today than there was then. True, the licensing requirement still exists, but once you have your license if you prefer (and many amateurs do), you can spend 100 percent of your amateur operating time on voice or other modes that require no knowledge of the code. In the 1930s, you needed the code to communicate, not just to get your license. There are also, today, a great many gadgets on the market that, while seeming to make code easier only serve really to instill bad habits on the operator. Some teachers for example, would have you start out with an electronic keyer, but this weds you to such a device forever more. The best way to start is with an ordinary straight key, learning characters by their sound, and striving to imitate machine sending by learning to control the muscles used in manipulating this key. This makes "graduating" to a bug or an electronic key much easier at a later date.

In order to make your sending good, you have to know what good sending sounds like. The way to acquire this is to copy W1AW's bulletins and code practice, or other perfect sending, then strive to imitate it Sometimes you can get a copy of the practice text (it's listed in advance in *QST*), and try to send along with W1AW. Most amateur cw operators today have difficulty maintaining proper spacing, probably because so much equipment in use demands that we key through a VOX relay. On cw the control for this relay is usually set for minimum delay, so it will close quickly and open just as quickly; but on most equipment it still doesn't close quickly enough, so a part of the first dit or dah of the first character is cut off. This has a tendency to cause the operator to run his words together so the relay will stay closed while he is sending but open immediately when he stops, making his sending very difficult to copy.

Nobody's sending is perfect, and therefore *every* operator should continually strive for improvement. Watch out for the customary pitfalls as your cw proficiency develops. Do you ever send Q for MA, or P for AN? Do you have a "swing?" Yes, even on an electronic key you can develop personal idiosyncrasies. Be your own worst critic, and make sure your sending, at whatever speed, is beyond reproach.

Break-In

On cw you can have true break-in – the ability to hear the signal of the other station while you are keying your transmitter. Technical considerations are covered elsewhere in this manual. Once this part of it has been accomplished, the full advantages and benefits of break-in can be realized. Long calls are unnecessary, because you can hear immediately if the station being called comes back to someone else. Much QRM is thus eliminated. If both stations in a QSO are using break-in, no station transmits unnecessarily; if the transmitting station is not being received, the receiving station "breaks" him and transmission stops. If another signal comes on the frequency, it can be heard immediately and any appropriate action taken. If message or other recorded traffic is being transmitted, any material missed can be filled immediately because the transmission can be interrupted just by the tap of a key. You can even call a CQ using break-in, and stop the moment someone hears you and starts calling. The customary procedure is CQ CQ CQ DE WØPAN WØPAN BK (pause) CQ CQ CQ . . ., until someone breaks or until it seems obvious no one is going to. Alternatively, the Q signal QSK can be used, either in sending CQ or at the beginning of a QSO to indicate to the other station that you are equipped for break-in and invite him to use it. QSK

Voice-Operating Hints

1) Listen before calling.
2) Make short calls with breaks to listen. Avoid long CQs; do not answer over-long CQs.
3) Use push-to-talk or voice control. Give essential data concisely in first transmission.
4) Make reports honest. Use definitions of strength and readability for reference. Make your reports informative and useful. Honest reports and *full* word description of signals save amateur operators from FCC trouble.
5) Limit transmission length. Two minutes or less will convey much information. When three or more stations converse in round tables, brevity is essential.
6) Display sportsmanship and courtesy. Bands are congested . . . make transmissions meaningful . . . give others a break.
7) Check transmitter adjustment . . . avoid a-m overmodulation and splatter. On ssb check carrier balance carefully. Do not radiate when moving VFO frequency or checking nbfm swing. Use receiver BFO to check stability of signal. Complete testing before busy hours!

is the mark of a well-equipped and well-operated cw station.

VOICE OPERATING

The use of proper procedure to get best results is just as important as in using code. In telegraphy words must be spelled out letter by letter. It is therefore but natural that abbreviations and shortcuts have come into use. In voice work, however, abbreviations are not necessary, and have less importance in our operating procedure.

The letter "K" is used in telegraphic practice so that the operator will not have to pound out the separate letters. The voice operator can *say* the words "go" or "over."

One laughs on cw by sending HI. On phone, *laugh* when one is called for.

The matter of reporting *readability* and *strength* is as important to phone operators as to those using code. With telegraph nomenclature, it is necessary to spell out words to describe signals or use abbreviated signal reports. But on voice, we have the ability to "say it with words." "Readability four, strength eight" is the best way to give a quantitative report, but reporting can be done so much more meaningfully with ordinary words: "You are weak but I can understand you, so go ahead," or "Your signal is strong but you are buried under local interference."

Voice Equivalents to Code Procedure

Voice	*Code*	*Meaning*
Over	$\overline{AR}$	After call to specific station
End of message	$\overline{AR}$	End of transmission or record message
Wait; stand by	$\overline{AS}$	Self-explanatory
Roger	R	*All* received correctly
Go	K	Any station transmit
Go only	$\overline{KN}$	Addressed station only transmit
Clear	$\overline{SK}$	End of contact or communication
Closing station	CL	Going off the air

Phone-Operating Practice

Efficient voice communication, like good cw communication, demands good operating. Adherence to certain points "on getting results" will go a long way toward improving our phone-band operating conditions.

Use VOX or push-to-talk. If you use VOX (most home stations do), don't defeat its purpose by saying "aaahhh" to keep the relay closed. If you use push-to-talk (common on mobiles so traffic noises won't affect transmission), let go of the button every so often to make sure you aren't "doubling" with the other fellow. Don't be a monologuist – a guy who likes to hear himself talk.

Listen with care. It's natural enough to answer the loudest signal who calls, but do a little digging, if necessary, to answer the *best* signal instead, where there is a choice. Every amateur can't run a kilowatt, but there is no reason why every amateur cannot have a signal of the highest quality. Don't reward the guy who cranks up his gain and splatters by answering his call if another station is calling.

Interpose your call frequently. Say it often and distinctly, in measured tones. Too often, identification is muffled or slurred. The fastest voice communication doesn't come from the guy who talks fastest; it comes from the operator who speaks distinctly. Your call especially is important, you can be cited for improper identification if it cannot be understood.

Listen before transmitting. Make sure the frequency isn't being used before you come barging onto it. Our voice bands are pretty crowded and QRM is inevitable. But this is a reason for *more* courtesy, not less.

Keep modulation constant. By turning your gain "wide open" you are subjecting anyone listening to all kinds of extraneous noises that don't belong on the air. Speak as closely to the mike as you can without breath modulation, turn your gain down so that only your voice can be heard. A good stunt is to hold the mike at the corner of your mouth and talk across it, rather than into it. If you use a stationary mike, turn it so that your breath goes across it, not into it; otherwise, your "explosives" will distort your speech.

Have a pencil and paper always handy. Take notes on the other guy's conversation while he's talking, so you can answer him or comment on the things he has said; otherwise he might get the wrong impression that you are deliberately ignoring some of his remarks.

Avoid repetition. Don't repeat back what the other fellow has just said. Just *say* you received everything, don't try to prove it.

Avoid inanities. There are many of them in phone operation, and they are contagious. "That's a roger." "Yeeeaaah!" "By golly." The phoney laugh. The affected speech. If you must parrot, parrot the polished operator, not the affected or idiotic one.

Steer clear of such controversial or suggestive subjects as politics and sex, and of profanities, even those considered acceptable in today's permissive society.

Use phonetics only as required. When clarifying genuinely doubtful expressions and in getting your call identified positively we suggest use of the International Telecommunication Union list. However, don't overdo its use.

The speed of radiotelephone transmission (with perfect accuracy) depends almost entirely upon the skill of the two operators involved. One must learn to speak at a rate allowing perfect understanding as well as permitting the receiving operator to copy down the message text, if that is necessary. Because of the similarity of many English speech sounds, the use of word lists has been found necessary. All voice-operated stations should use a *standard* list as needed to identify call signals or unfamiliar expressions.

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

Round Tables. The round table has many advantages if run properly. It clears frequencies of interference, especially if all stations involved are on the same frequency, while the enjoyment value remains the same, if not greater. By use of push-to-talk, or vox, the conversation can be kept lively and interesting, giving each station operator ample opportunity to participate without waiting overlong for his turn.

Round tables can become very unpopular if they are not conducted properly. The monologuist, off on a long spiel about nothing in particular, cannot be interrupted; *make your transmissions short and to the point.* "Butting in" is discourteous and unsportsmanlike; *don't enter a round table, or any contact between two other amateurs, unless you are invited.* It is bad enough trying to copy through prevailing interference without the added difficulty of poor voice quality; *check your transmitter adjustments frequently.* In general, follow the precepts as hereinbefore outlined for the most enjoyment in round tables as well as any other form of radiotelephone communication.

DX OPERATING CODE

(For W/VE Amateurs)

Some amateurs interested in DX work have caused considerable confusion and QRM in their efforts to work DX stations. The points below, if observed by all W/VE amateurs, will go a long way toward making DX more enjoyable for everybody.

1. Call DX only after he calls CQ, QRZ?, signs $\overline{SK}$, or phone equivalent thereof.

2. Do *not* call a DX station:

a. On the frequency of the station he is working until you are *sure* the QSO is over. This is indicated by the ending signal $\overline{SK}$ on cw and any indication that the operator is listening, on phone.

b. Because you hear someone else calling him.

c. When he signs $\overline{KN}$, $\overline{AR}$, CL, or phone equivalents.

d. Exactly on his frequency.

e. After he calls a directional CQ, unless of course you are in the right direction or area.

3. Keep within frequency-band limits. Some DX stations operate outside. Perhaps they can get away with it, but you cannot.

4. Observe calling instructions of DX stations. "10U" means call ten kHz *up* from his frequency, "15D" means 15 kHz *down*, etc.

5. Give honest reports. Many foreign stations *depend* on W and VE reports for adjustment of station and equipment.

6. Keep your signal clean. Key clicks, chirps, hum or splatter give you a bad reputation and may get you a citation from FCC.

7. *Listen* for and *call* the station you want. Calling CQ DX is not the best assurance that the *rare* DX will reply.

8. When there are several W or VE stations waiting to work a DX station, avoid asking him to "listen for a friend." Let your friend take his chances with the rest. Also avoid engaging DX stations in rag-chews against their wishes.

WORKING DX

Most amateurs at one time or another make "working DX" a major aim. As in every other phase of amateur work, there are right and wrong ways to go about getting best results in working foreign stations, and it is the intention of this section to outline a few of them.

The ham who has trouble raising DX stations readily may find that poor transmitter efficiency is not the reason. He may find that his sending is poor, his calls ill timed, or his judgment in error. Working DX requires the know-how that comes with experience. If you just call CQ DX you may get a call from a foreign station, but it isn't likely to be a "rare one." On the other hand, unless you are experienced enough to know that conditions are right, your receiver is sensitive and selective enough and your transmitter and antenna properly tuned and oriented, you may get no calls at all, and succeed only in causing some unnecessary QRM.

The call CQ DX means slightly different things to amateurs on different bands:

a) On vhf, CQ DX is a general call ordinarily used only when the band is open, under favorable "skip" conditions. For vhf work, such a call is used for looking for new states and countries, also for distances beyond the customary "line-of-sight" range on most vhf bands.

b) CQ DX on our 7-, 14-, 21-, and 28-MHz bands may be taken to mean "General call to any foreign station." The term "foreign station" usually refers to any station on a different continent. If you do call CQ DX, remember that it implies you will answer any DX who calls. If you don't mean "general call to any DX station," then listen and call the station you do want.

Snagging the Rare Ones

Once in a while a CQ DX will result in snagging a rare DX contact, if you're lucky. This seldom happens, however; usually, what you have to do is listen – and listen – and then listen some more. *You gotta hear 'em before you can work 'em!* If everybody transmits, nobody is going to hear anything. Be a snooper. Usually, unless you are lucky enough to be among the first to hear him, a rare DX station will be found under a pileup, with stations swarming all over him like worker bees over a queen. The bedlam will subside when the DX station is transmitting (although some stations keep right on calling him), and you can hear him. Don't immediately join the pack, be a little cagey. Listen a while, get an idea of his habits, find out where he is listening (if not zero on himself), bide your time and wait your chance. Sometimes "tail-ending" works. This is the practice of waiting until the station your DX is working starts his sign-off, then just transmitting your own call. Be careful however; this could backfire. If your DX station doesn't respond to such tactics, best to avoid it. Some of them don't like it.

Make your calls short, snappy. No need to repeat his call (he knows it very well, all he needs to know is that you are calling him), but send your own call a couple of times. Try to find a time when few stations are calling him and he is not transmitting; then get in there! With experience, you'll learn all kinds of tricks, some of them clever some just plain dirty. You'll have no trouble discerning which is which. Learn to use the clever ones, and shun the dirty ones. More than you think depends on the impressions we make on our foreign friends!

Codes and Ethics

One of the most effective ways to work DX is to know the operating habits of the DX stations sought, and to abide by the procedures they use. Know when and where to call, and for how long, and when to remain silent waiting your chance. DXing has certain understood codes of ethics and procedures that will make this popular amateur pursuit more fun for everybody if everybody follows them. One of the sad things about DXing is to listen to some of the vituperation and abuse that goes on, mostly by stations on "this" side, as they trample on each other trying to raise their quarry. DX stations have been known to QRT in disgust at some of the tactics.

If W and VE stations will use the procedure in the "DX Operating Code" detailed elsewhere on these pages, we can all make a good impression on the air. ARRL has also recommended some operating procedures for DX stations aimed at controlling some of the thoughtless practices sometimes used by W/VE amateurs A copy of these recommendations (Op Aid No. 5) can be obtained free of charge from ARRL Headquarters .

Choosing Your Band

If it does nothing else in furthering your education, striving to work DX will certainly teach you a few things about propagation. You will find that four principal factors determine propagation characteristics. (1) The frequency of the band in which you do your operating. (2) The time of day or night. (3) The season of the year (4) The sunspot cycle. The proper choice of band depends pretty much on the other three factors. For example, the 3.5-4.0-MHz band at high noon in the summertime at the "node" part of the sunspot cycle is the poorest possible choice, while the same band at midnight during the wintertime at the 'null' part of the cycle might produce some very exciting DX. Similarly, you will learn by experience when to operate on which band for the best DX by juggling the above factors using both long-range and other indications of band conditions. WWV transmissions can also be helpful in indicating both current and immediate-forecast band conditions.

Conditions in the transmission medium often make it possible for the signals from low-powered transmitters to be received at great distances. In general, the higher the frequency band the less important power considerations become, for occasional DX work. This accounts in part for the relative popularity of the 14-, 21- and 28-MHz bands among amateurs who like to work DX.

QSL CARDS AND BUREAUS

Most amateurs who work another station for the first time, especially a foreign station, will later send the station a postcard confirming the contact. These cards are known as QSLs, taken from the international signal meaning, "I acknowledge receipt."

In 1976, in compliance with a directive from the ARRL Board of Directors, the ARRL-Membership Overseas QSL Service came into being. Members of the League may use this service 12 times a year. Any number of cards may be sent in at a time. They must, however, be presorted by country prefix, include a mailing label from a *QST* wrapper and be accompanied by a $1 check. Mailings to over 130 countries takes place weekly. Full information on operation of the bureau is available from ARRL hq.

Since it is rather expensive, for a foreign station especially, to send a QSL separately to each U.S. or Canadian station he's worked, ARRL has also set up a system of "incoming" QSL Bureaus, manned by amateur volunteers in each call area. The bureaus get packages of cards from overseas, which are sorted by call. Individual amateurs may claim their cards by sending a supply of stamped, self-addressed envelopes to the QSL manager in their area. QST carries the addresses of these bureaus nearly every issue. Or write to ARRL hq. for information.

KEEPING AN AMATEUR STATION LOG

Although recent FCC rulings have eliminated the legal necessity for detailed logging, you'll still want to maintain a log to preserve a record of your own activity within amateur radio, to be able to send QSLs, and to protect yourself. You'll be confident of meeting all of these by recording: (1) the date and time of each transmission, (2) all calls and transmissions made, whether contacts resulted, or not, (3) the input power to the last stage of the transmitter, (4) the frequency band used, (5) the time of ending each contact (QSO), and (6) the signature of the licensed operator. Written messages handled in standard form must be included in the log or kept on file for a period of at

DATE TIME (GMT)	STATION CALLED	CALLED BY	HIS SIGNAL RST	MY SIGNAL RST	FREQ	EMISSION TYPE	POWER INPUT WATTS	TIME OF ENDING QSO	OTHER DATA	NAME	QSL S	QSL R
16MAR.												
2300	W3CML	x	589	579	3.5	A1	250	2309	RCD S	BILL	—	
2315	CQ	x			7							
16	x	W5[illegible]	469	479	"	"	"	2330	CNDA [illegible]	[illegible]	✓	
35	K2SJC	x	57	57	3.8	A3A	150	2359	SKED			
17MAR												
0005	UA9CH	x	579	589	14	A1	25	0012	NOVOSIBIRSK	VLADIS	✓	
0013	x	UW9PT	579	579	"	"	"	0020		VIK	✓	
1200	CQ	x			"	"	"					
02	x	VK3NR	579	579	"		"	1217	MELBOURNE	NOEL	✓	

KEEP AN ACCURATE AND COMPLETE STATION LOG AT ALL TIMES. FCC REQUIRES IT.

A page from the official ARRL log is shown above, answering every FCC requirement in respect to station records. Bound logs made up in accord with the above form can be obtained from Headquarters for a nominal sum or you can prepare your own, in which case we offer this form as a suggestion. The ARRL log has a special wire binding and lies perfectly flat on the table.

least one year.

But a log can be more than just a legal record of station operation. It can be a "diary" of your amateur experience. Make it a habit to enter thoughts and comments, changes in equipment, operating experiences and reactions, anything that might make enjoyable reminiscences in years to come. Your log is a reflection of your personal experience in amateur radio. Make it both neat and complete.

ARRL headquarters stocks log books and message blanks for the convenience of amateurs.

PUBLIC SERVICE OPERATING

Amateurs interested in rendering public service in operating under ARRL sponsorship have formed the Amateur Radio Public Service Corps (ARPSC). This organization has two principal divisions. One is the Amateur Radio Emergency Service (ARES), an emergency-preparedness group of approximately 40,000 amateur operators signed up voluntarily to keep amateur radio in the forefront along preparedness lines. The other is the National Traffic System (NTS), a message-handling facility which operates daily (including weekends and holidays) for systematic handling of third-party traffic.

Also recognized by ARRL as a part of the organized amateur radio public service effort are the Radio Amateur Civil Emergency Service (RACES), a part of the amateur service serving civil defense under a separate sub-part of the amateur regulations; the Military Affiliate Radio Service, sponsored by the armed services to provide military training for amateurs; and numerous amateur groups organized into nets by individuals, clubs or other amateur entities for public service and registered with the League. The detailed workings of ARPSC and RACES are covered briefly herein and explained in somewhat more detail in *Public Service Communications, Operating an Amateur Radio Station*, available to interested amateurs without charge.

Here is an example of a plain-language message as it would be prepared for delivery. If the message were for relay instead of delivery, the information at the bottom would be filled in instead of that in the box.

MESSAGE HANDLING

Amateur operators in the United States and a few other countries enjoy a privilege not available to amateurs in most countries – that of handling third-party message traffic. In the early history of amateur radio in this country, some amateurs who were among the first to take advantage of this privilege formed an extensive relay organization which became the ARRL.

Thus, amateur message-handling has had a long and honorable history, and like most services, has gone through many periods of development and change. Those amateurs who handled traffic in 1914 would hardly recognize it the way some of us do it today, just as equipment in those days was far different from that in use now. Progress has been made and new methods have been developed in step with advancement in communication techniques of all kinds. Amateurs who handled a lot of traffic found that organized operating schedules were more effective than random relays, and as techniques advanced and messages increased in number, trunk lines were organized, spot frequencies began to be used, and there came into existence a number of traffic nets in which many stations operated on the same frequency to effect wider coverage in less time with fewer relays; but the old methods are still available to the amateur who handles only an occasional message.

Although message handling is as old an art as is amateur radio itself, there are many amateurs who do not know how to handle a formal message and have never done so. As each amateur grows older and gains experience in the amateur service, there is bound to come a time when he will be called upon to handle a written message, during a communications emergency, in casual contact with one of his many acquaintances on the air, or as a result of a request from a non-amateur friend. Regardless of the occasion, if it comes to you, you will want to rise to it! Considerable embarrassment is likely to be experienced by the amateur who finds he not only does not know the form in which the message should be prepared, but does not know how to go about putting it on the air.

Traffic work need not be a complicated or time-consuming activity for the casual or occasional message-handler. Amateurs may participate in traffic work to whatever extent they wish, from an occasional message now and then to becoming a part of organized traffic systems. This chapter explains some principles so the reader may know where to find out more about the subject and may exercise the message-handling privilege to best effect as the spirit and opportunity arise.

Responsibility

Amateurs who originate messages for transmission or who receive messages for relay or delivery should first consider that in doing so they are accepting the responsibility of clearing the message from their station on its way to its destination in the shortest possible time. Forty-eight hours after filing or receipt is the generally-accepted rule among traffic-handling amateurs, but it is obvious that if every amateur who relayed the message allowed it to remain in his station this long it might be a long time reaching its destination. Traffic should be relayed or delivered as quickly as possible.

Message Form

Once this responsibility is realized and accepted, handling the message becomes a matter of following generally-accepted standards of form and transmission. For this purpose, each message is divided into four parts: the preamble, the address, the text and the signature. Some of these parts themselves are subdivided. It is necessary in preparing the message for transmission and in actually transmitting it to know not only what each part is and what it is for, but to know in what *order* it should be transmitted, and to know the various procedure signals used with it when sent by cw. If you are going to send a message, you may as well send it right.

Standardization is important! There is a great deal of room for expressing originality and individuality in amateur radio, but there are also times and places where such expression can only cause confusion and inefficiency. Recognizing the need for standardization in message form and message transmitting procedures, ARRL has long since recommended such standards, and most traffic-interested amateurs have followed them. In general, these recommendations, and the various changes they have undergone from year to year, have been at the request of amateurs participating in this activity, and they are completely outlined and explained in *Operating an Amateur Radio Station*, a copy of which is available upon request or by use of the coupon at the end of this chapter.

Clearing a Message

The best way to clear a message is to put it into one of the many organized traffic networks, or to give it to a station that can do so. There are many amateurs who make the handling of traffic their principal operating activity, and many more still who participate in this activity to a greater or lesser extent. The result is a traffic system which spreads to all corners of the United States and covers most U.S. possessions and Canada. Once a message gets into an organized net, regardless of the net's size or coverage, it is systematically routed toward its destination in the shortest possible time.

Amateurs not experienced in message handling should depend on the experienced message-handler to get a message through, if it is important; but the average amateur can enjoy operating with a message to be handled either through a local traffic net or by free-lancing. The latter may be accomplished by careful listening for an amateur station at desired points, directional CQs, use of recognized calling and net frequencies, or by making and keeping a schedule with another amateur for regular work between specified points. He may well aim at learning and enjoying through doing. The joy and accomplishment in thus developing one's operating skill to the peak of perfection has a reward all its own.

If you decide to "take the bull by the horns" and put the message into a traffic net yourself (and more power to you if you do!), you will need to know something about how nets operate, and if on cw, the special Q signals and procedure they use to dispatch all traffic with a maximum of efficiency. The frequency and operating time of the net in your section, or of other nets into which your message can go, is given in ARRL's Net Directory. This annually-revised publication is available on request. Listening for a few minutes at the time and frequency indicated should acquaint you with enough fundamentals to enable you to report into the net and report your traffic. From that time on you follow the instructions of the net control station, who will tell you when and to whom (and on what frequency, if different from the net frequency) to send your message. Cw nets use the special "QN" signals, so it may be helpful to have a list of these before you (available from ARRL Hq., Operating Aid No. 9).

Network Operation

About this time, you may find that you are enjoying this type of operating activity and want to know more about it and increase your proficiency. Many amateurs are happily "addicted" to traffic handled after only one or two brief exposures to it. Much traffic is at present being conducted by cw, since this mode of communication seems to be popular for record purposes – but this does not mean that high code speed is a necessary prerequisite to working in traffic networks. There are many nets organized specifically for the slow-speed amateur, and most of the so-called "fast" nets are usually glad to slow down to accommodate slower operators.

It is a significant operating fact that code speed alone does *not* make for efficiency – sometimes the contrary! A high-speed operator who does not know procedure can "foul up" a net much more completely and more quickly than can a slow operator. Cw net operation provides an excellent opportunity to increase code speed. Given a little time your speed will reach the point where you can easily hold your own. Concentrate first on learning the net procedures.

Voice modes are also very popular for traffic work. Procedure is of paramount importance on phone, just as it is on cw. Procedure differs in that standard phonetics are an important ingredient in phone operation and *Q* and *QN* signals are not used. However, nets on all modes share the need for concise operation.

Teamwork is the theme of all net operation. The net which functions most efficiently is the net in which all participants are thoroughly familiar with the procedure used, and in which operators refrain from transmitting except at the direction of the net control station, and do not occupy time with extraneous comments, even the exchange of pleasantries. There is a time and place for everything. When a net is in session it should concentrate on handling traffic until all traffic is cleared. Before or after the net is the time for rag-chewing and discussion. Some further details of net operation are included in *Operating an Amateur Radio Station*, mentioned earlier, but there is no substitute for actual participation.

The National Traffic System

To facilitate and speed the movement of message traffic, there is in existence an integrated national system by means of which originated traffic can normally reach its destination area the same day the message is originated. This system uses the state or section net as a basis. Each section net sends a representative to a "region" net (normally covering a call area) and each "region" net sends a representative to an "area" net (normally covering a time zone). After the area net has cleared all its traffic, its members then go back to their respective region nets, where they clear traffic to the various section net representatives. By means of connecting schedules between the area nets, traffic can flow both ways so that traffic originated on the West Coast reaches the East Coast with a maximum of dispatch, and vice versa. In general, evening section nets function at 1900, evening region nets at 1945, evening area nets at 2030 and the same or different regional personnel again at 2130. Some section nets conduct a late session at 2200 to effect traffic delivery the same night. Local standard time is referred to in each case.

There also exists a segment of NTS that meets during the daytime hours. NTS(D) follows the same general sequence as NTS(E), but the times are less standard from region to region and area to area. Traffic from area to area is handled in a Continental Net, rather than by use of a Transcontinental Corps, as in NTS(E). *QST* covers details of NTS(D) as they unfold.

The NTS plan somewhat spreads traffic opportunity so that casual traffic may be reported into nets for efficient handling one or two days or nights per week; or the ardent traffic man can operate in both daytime and evening segments to roll up impressive totals and speed traffic reliably to its destination. Old-time traffic men who prefer a high degree of organization and teamwork have returned to the traffic game as a result of the new system. Beginners have shown more interest in becoming part of a system nationwide in scope, in which *anyone* can participate. The National Traffic System has vast and intriguing possibilities as an amateur service. It is open to any amateur who wishes to participate.

The above is but the briefest resume of what is of necessity a rather complicated arrangement of nets and schedules. Complete details of the System and its operation are included in the ARRL *Public Service Communications* Manual.

EMERGENCY COMMUNICATION

One of the most important ways in which the amateur serves the public, thus making his existence a national asset, is by his preparation for and his participation in communications emergencies. Every amateur, regardless of the extent of his normal operating activities, should give some thought to the possibility of his being the only means of communication should his community be cut off from the outside world. It has happened many times, often in the most unlikely places; it has happened without warning, finding some amateurs totally unprepared; it can happen to *you*. Are you ready?

There are two principal ways in which any amateur can prepare himself for such an eventuality. One is to provide himself with equipment capable of operating on any type of emergency power (i.e., either ac or dc), and equipment which can readily be transported to the scene of disaster. Mobile and hand-held equipment is especially desirable in most emergency situations.

Such equipment, regardless of how elaborate or how modern, is of little use, however, if it is not used properly and at the right times; and so another way for an amateur to prepare himself for emergencies, by no means less important than the first, is to *learn to operate efficiently*. There are many amateurs who feel that they know how to operate efficiently but who find themselves considerably handicapped at the crucial time by not knowing proper procedure, by being unable, due to years of casual amateur operation, to adapt themselves to snappy, abbreviated transmissions, and by being unfamiliar with message form and

Before Emergency

PREPARE yourself by providing emergency power for your station.

TEST your emergency equipment and operating ability in the annual Simulated Emergency Test and Field Day.

REGISTER with your ARRL Emergency Coordinator. If none, offer your services to local and civic relief agencies and explain what amateur radio can do during disasters.

In Emergency

LISTEN before you transmit, *always*!

REPORT to your Emergency Coordinator so he will have latest data on your facilities. Offer local civic and relief agencies your services directly in the absence of an EC.

RESTRICT all on-the-air work in accordance with FCC regulations, Sec. 97.107.

SOS is the International Distress Call for a dire emergency. The phone equivalent is MAYDAY. Use these calls for *emergency only*. False distress calls are unlawful.

RESPECT the fact that success in emergency depends on circuit discipline. The net control station is the supreme authority.

COOPERATE with those we serve. Be ready to help, but stay off the air unless there is a specific job to be done that you can handle more efficiently than any other station.

COPY bulletins from W1AW. During emergencies, special bulletins are transmitted as follows: *phone*, on the hour; *RTTY*, 15 minutes past the hour; *cw*, on the half hour.

After Emergency

REPORT to ARRL Headquarters promptly and fully so that the Amateur Service can receive full credit.

procedures. It is dangerous to overrate your ability in this; it is better to assume you have things to learn.

In general it can be said that there is more emergency equipment available than there are operators who know properly how to operate during emergency conditions, for such conditions require clipped, terse procedure with complete break-in on cw and fast push-to-talk or VOX on phone. The casual rag-chewing aspect of amateur radio, however enjoyable and worthwhile in its place, must be forgotten at such times in favor of the business at hand. There is only one way to gain experience in this type of operation, and that is by practice. During an emergency is no time for practice; it should be done beforehand, as often as possible, on a regular basis.

This leads up to the necessity for emergency organization and preparedness. ARRL has long recognized this necessity and has provided for it. The Section Communications Manager (whose address appears on page 8 of every issue of *QST*) is empowered to appoint certain qualified amateurs in his section for the purpose of coordinating emergency communication organization and preparedness in specified areas or communities. This appointee is known as an Emergency Coordinator for the city or town. One should be specified for each community. For coordination and promotion at section level a Section Emergency Coordinator arranges for and recommends the appointments of various Emergency Coordinators at activity points throughout the section. Emergency Coordinators organize amateurs in their communities according to local needs for emergency communication facilities.

The community amateurs taking part in the local organization are members of the Amateur Radio Emergency Service (ARES). *All* amateurs are invited to register in the ARES, whether they are able to play an active part in their local organization or only a supporting role. Application blanks are available from your EC, SEC, SCM or direct from ARRL Headquarters. In the event that inquiry reveals no Emergency Coordinator appointed for your community, your SCM would welcome a recommendation either from yourself or from a radio club of which you are a member. By holding an amateur operator license, you have the responsibility to your community and to amateur radio to uphold the traditions of the service.

Among the League's publications is a booklet entitled *Public Service Communications*. This booklet, while small in size, contains a wealth of information on ARES organization and functions and is invaluable to any amateur participating in emergency or civil defense work. It is free to ARES members and should be in every amateur's shack. Drop a line to the ARRL Communications Department if you want a copy, or use the coupon at the end of this chapter.

Every EC and SEC receives an *EC Workbook* from ARRL to aid in formulating almost every detail necessary for both planning and operating within the section.

The Radio Amateur Civil Emergency Service

Following World War II there was established within our government the Federal Civil Defense Administration (FCDA), which, at the behest of ARRL and other amateurs, considered the role of the amateur in civil defense communication should the U.S. become embroiled in another war. This resulted, in 1951, in the establishment of the Radio Amateur Civil Emergency Service (RACES) with rules promulgated by FCC as a part of the Amateur Radio Service. FCDA has evolved into the present Defense Civil Preparedness Agency, part of the Department of Defense, and the RACES rules have undergone several changes.

RACES is intended solely for civil defense communication during civil emergencies, through the medium of amateur radio, and is designed to continue operation during any extreme national emergency such as war. It shares certain segments of frequencies with the regular (i.e., normal) Amateur Service on a nonexclusive basis. Its regulations are a subpart of the familiar amateur regulations (Part 97) and are included in full in the ARRL *License Manual*.

A drastic change in RACES rules effected by FCC in February of 1976 greatly simplifies the procedures for amateur participation in RACES. The average amateur now simply makes himself available to his local civil defense (or whatever name) organization and becomes a part of that organization. He then uses his amateur call and observes regular amateur licensing privileges.

There is no longer a requirement for such things as a communications plan, a radio officer or personnel clearances. Local regulations may provide for some of these, but FCC regulations do not. Only licensed amateurs may act as control operators, and then only within their licensed privileges.

RACES stations are licensed only to civil defense organizations, not to individuals. They are given a WC prefix, a number corresponding to the call area in which they are located, and three letters. Application is made on FCC Form 610-B by the civil defense director or equivalent.

In the event of war, civil defense will place great reliance on RACES for back-up radio communication. Even in peacetime, RACES can be of great value in natural disaster communications. As a part of our total amateur public service effort, it deserves our whole-hearted and enthusiastic support and will permit us to continue to function in the public service, as amateurs, in RACES in wartime as we function in AREC and NTS during peacetime. If interested, inquire of your local civil defense agency and get sighed up with your radio officer.

ARRL OPERATING ORGANIZATION

Amateur operation must have point and constructive purpose to win public respect. Each individual amateur is the ambassador of the entire fraternity in his public relations and attitude toward his hobby. ARRL field organization adds point and purpose to amateur operating.

The Communications Department of the League is concerned with the practical operation of stations in all branches of amateur activity. Appointments or awards are available for rag-chewer, traffic enthusiast, phone operator, DX man and experimenter.

There are seventy-four ARRL Sections in the League's field organization, which embraces the United States, Canada, and certain other territory. Operating affairs in each Section are supervised by a Section Communications Manager (SCM) elected by members in that section for a two-year term of office. Organization appointments are made by the SCMs, elected as provided in the Rules and Regulations of the Communications Department, which accompany the League's By-Laws and Articles of Association. SCM addresses for all sections are given in full in each issue of *QST*. SCMs welcome monthly activity reports from all amateurs in their sections, regardless of status.

Whether your activity embraces phone or telegraphy, or both, there is a place for you in the League organization.

LEADERSHIP POSTS

To advance each type of station work and group interest in amateur radio, and to develop practical communications plans with the greatest success, appointments of ARRL members holding Conditional Class licenses or better to serve as leaders and organizers in particular single-interest fields are made by the SCM. Each leadership post is important. Each provides activities and assistance for appointee groups and individual members along the lines of natural interest. Some posts further the general ability of amateurs to communicate efficiently at all times by pointing activity toward networks and round tables; others are aimed specifically at establishment of provisions for organizing the amateur service as a standby communications group to serve the public in disaster, civil defense need or emergency of any sort. The SCM appoints the following in accordance with section needs and individual qualifications:

PAM Phone Activities Manager. Organizes activities for voice operators in his section. Promotes phone nets and recruits Official Phone Station appointees. The appointment of VHF-PAM is open to Technician licensees.

RM Route Manager. Organizes and coordinates cw traffic activities. Supervises and promotes nets and recruits Official Relay Station appointees.

SEC Section Emergency Coordinator. Promotes and administers section emergency radio organization.

EC Emergency Coordinator. Organizes amateurs of a community or other local areas for emergency radio service; maintains liaison with officials and agencies served, also with other local communication facilities. Sponsors tests, recruits for AREC and encourages alignment with RACES. A Technician Class licensee may receive this appointment if a qualified higher class licensee is not available.

STATION APPOINTMENTS

ARRL's field organization has a place for every active amateur who has a station. The Communications Department organization exists to increase individual enjoyment and station effectiveness in amateur radio work, and we extend a cordial invitation to every amateur to participate fully in the activities, to report results monthly, and to apply to the SCM for one of the following station appointments. ARRL membership and the conditional class or higher license or VE equivalent is prerequisite to all appointments, except where otherwise indicated.

OPS Official Phone Station. Sets high voice operating standards and procedures, furthers phone nets and traffic.

ORS Official Relay Station. Traffic service, operates cw nets; noted for 15 wpm and procedure ability. Open to RTTY traffickers.

ORS II Same as ORS, for the Novice and Technician operators, code speed minimum of 10 wpm.

OBS Official Bulletin Station. Transmits ARRL and FCC bulletin information to amateurs. Open to Technician licensees.

OVS Official VHF Station. Collects and reports vhf-uhf-shf propagation data, may engage in facsimile, TT, TV, work on 50 MHz and/or above. Takes part as feasible in vhf traffic work, reports same, supports vhf nets, observes procedure standards. Open to both Novice and Technician licensees.

OO Official Observer. Sends cooperative notices to amateurs to assist in frequency observance, insures high-quality signals, and prevents FCC trouble.

Emblem Colors

Members may wear the ARRL emblem with black-enamel background. A red background will indicate that the wearer is or has been SCM. SECs, ECs, RMs and PAMs may wear the emblem with green background. Observers and all *station* appointees are entitled to wear blue background emblems. Advisory Committee members wear yellow background pins and QSL Bureau personnel may wear orange pins.

RADIO CLUB AFFILIATION

ARRL affiliation is available to any amateur society in one of three categories: Category 1, all "local" radio clubs having at least 51% licensed amateurs and at least 51% ARRL membership; Category 2, radio club "councils," and similar organizations of large geographic area, same requirements as category 1. Category 3, high school, college and youth-group clubs having at least one officer or trustee who is a licensed amateur and an ARRL member.

A "Club Kit" is available upon request from the Club and Training Department; this kit contains all papers necessary for affiliation application plus other materials of interest to clubs. Once the completed affiliation package is returned the affiliation process begins.

ARRL affiliated clubs receive a quarterly bulletin from Headquarters and special information at intervals for posting on club bulletin boards or for relay to club members. A travel plan providing communications, technical, and legal/regulatorial contact from the Headquarters is worked out seasonally to give maximum benefits to as many as possible of the active affiliated radio clubs.

Material aimed at training and entertainment of club members is available, plus advice on club problems such as organization, conducting meetings and attracting new members. Training services for clubs include films, slide collections, and complete lesson plans, available upon request. Watch *QST* and *Radio Club News* for details on these items, or write the ARRL for the special benefits to affiliated clubs.

W1AW

The Maxim Memorial Station, W1AW, is dedicated to fraternity and service. Operated by the League headquarters, W1AW is located adjacent to the Headquarters offices on a seven-acre site. The station is on the air daily, except holidays, and available time is divided between the different bands and modes. Facilities for all commonly used amateur modes are provided for all bands from 1.8 to 144 MHz.

Operation is roughly proportional to amateur interest in different bands and modes with maximum legal power on most bands. W1AW's daily bulletins and code practice aim to give operational help to the largest number.

W1AW was established as a living memorial to Hiram Percy Maxim, to carry on the work and traditions of amateur radio. The station is on the air daily and is open to visitors at all times it is in operation. The W1AW schedule of operation and visiting hours is printed each month in the *Operating News* section of *QST*.

W1AW Code Practice

Approximate frequencies: 1.835 3.58 7.08 14.08 21.08 28.08 50.08 and 147.555 MHz. For practice purposes the order of words in each line may be reversed during the 5-15 wpm transmissions. Each tape carries checking references. Details on Qualifying Runs appear monthly in *QST* Operating Events.

Speeds	*EST/EDST*	*PST/PDST*
5-7-1/2-10-13-15		
	9 A.M./7 P.M. MWF	6 A.M./4 P.M. MWF
	4 P.M./10 P.M. TThSSu	1 P.M./7 P.M. TThSSu
35-30-25-20-15-13-10		
	9 A.M. TTh	6 A.M. TTh
	4 P.M./10 P.M. MWF	1 P.M./7 P.M. MWF
	7 P.M. TThSSu	4 P.M. TThSSu

OPERATING ACTIVITIES

Within the ARRL field organization there are many special activities. For all appointees and officials, regular CD (Communications Department) Parties are scheduled to develop operating ability and a spirit of fraternalism.

In addition, ARRL sponsors various other activities open to all amateurs. The DX-minded amateur may participate in the Annual ARRL International DX Competition during February and March. This popular contest may bring you the thrill of working new countries and building up your DXCC totals; certificate awards are offered to top scorers in each country and ARRL section (see page 8 of any *QST*) and to club leaders. Then there is the very-popular Sweepstakes in November. Of domestic scope, the SS affords the opportunity to work new states for that WAS award. A Novice activity is planned annually. Both a 10- and 160-Meter Contest are scheduled for early December. The interests of vhf enthusiasts are also provided for in contests held in January, June and September of each year. Where enough logs (three) are received to constitute minimum "competition" a certificate in spot activities, such as the "SS" and vhf party, is awarded the leading newcomer for his work considered only in competition with other newcomers.

As in all our operating, the idea of having a good time is combined in the annual June Field Day with the more serious thought of preparing ourselves to render public service in times of emergency. A premium is placed on the use of equipment without connection to commercial power sources. Clubs and individual groups always enjoy themselves in the "FD" and learn much about the requirements for operating under knockabout conditions afield.

ARRL contest activities are diversified to appeal to all operating interest, and will be found announced in detail in issues of *QST* preceding the different events.

AWARDS

The League-sponsored operating activities, heretofore mentioned, have useful objectives and provide much enjoyment for members of the fraternity. Achievement in amateur radio is also recognized by various certificates offered through the League and detailed below. Basic rules require that sufficient funds be included with submission of cards to insure their safe return. A W/VE must be an ARRL member to participate in the WAS and DXCC programs. (Novices and DX stations are exempted from this requirement.)

WAS Award

WAS means "Worked All States." An amateur, anywhere in the world, who succeeds in getting confirmed contacts with all fifty U.S. states and sends them in for examination, may receive this award from the League.

You can make the contacts over any period of time and on any or all amateur bands. If you wish, you may have your WAS award issued for some special way in which you made it, such as all cw, all phone, all on one band, all with lower power, etc. – only providing all cards submitted plainly show that a contact took place under the special circumstances for which you wish the award issued.

Before you send your cards, drop the ARRL Communications Department a line requesting a copy of the rules and an application blank.

5BWAS

The Five Band Worked All States Award became effective January 1, 1970. Only contacts made after that date count. Contacts must be confirmed with all 50 states on each of five amateur bands. Rules require applicants in the U.S. and possessions, Puerto Rico and Canada, to be a full member of ARRL. WAS rules apply.

DX Century Club Award

The DXCC is one of the most popular and sought-after awards in all of amateur radio, and among the more difficult to acquire. Its issuance is carefully supervised at ARRL headquarters by three staff members.

To obtain DXCC, an amateur must make two-way contact with 100 "countries" on the ARRL DXCC List. Written confirmations are required for proof of contact. Such confirmations must be sent to ARRL headquarters, where each one is carefully scrutinized to make sure it actually confirms a contact with the applying amateur, that it was not altered or tampered with, and that the "country" claimed is actually on the ARRL list. Further safeguards are applied to maintain the high standards of this award. A handsome king-size certificate is sent to each amateur qualifying.

The term "country" is an arbitrary one not necessarily agreeing with the dictionary definition of such. For DXCC purposes, many bodies of land not having independent status politically are classified as countries. For example, Alaska and Hawaii, while states of the U.S., are considered separate "countries" because of their distance from the mainland. There are over 300 such designations on the ARRL list. Once a basic DXCC is issued, the certificate can be endorsed, by sticker, for additional countries by sending the additional cards in to headquarters for checking.

Separate DXCC Awards are available for mixed modes, all phone, all cw, RTTY and 160 Meters.

Before applying, familiarize yourself with full information. Application forms (CD164) and the ARRL Countries List (detailing rules) may be obtained from Headquarters for a stamped addressed envelope.

Five-Band DXCC

Entirely separate from DXCC, ARRL also offers a Five-Band DXCC (5BDXCC) Award for those amateurs who submit written proof of having made two-way contact with 100 or more countries on each of five amateur bands since January 1, 1969.

For a copy of the complete rules, drop a line to ARRL Headquarters, 225 Main St., Newington, CT 06111.

WAC Awards

The WAC award, Worked All Continents, is issued by the International Amateur Radio Union (IARU) upon proof of contact with each of the six continents. Amateurs in the U.S.A., Possessions and Canada should apply for the award through ARRL, headquarters society of the IARU. Those elsewhere must submit direct to their own IARU member-society. Residents of countries not represented in the Union may apply directly to ARRL for the award. Two basic types of WAC certificates are issued. One contains no endorsements and is awarded for cw, or a combination of cw and phone contacts; the other is awarded when all work is done on phone. There is a special endorsement to the phone WAC when all the confirmations submitted clearly indicate that the work was done on two-way ssb. Special endorsements are also available for RTTY and SSTV. The *only* special band endorsements are for 1.8, 3.5, and 50 MHz.

Five- and Six-Band WAC Awards are based on contacts made on or after January 1, 1974. Write ARRL Headquarters for details.

Satellite "1000" Award

Contacts made on or after December 15, 1972, via the Oscar communications satellites count for this unique "DX Achievement" award. Only one contact per station, regardless of mode. To earn the award you must amass 1000 points. Each contact with a new station counts 10 points, with a new country 50 points, with a new continent 250 points. The fee for W/VE members and DX stations is $2 which includes return of the cards by registered mail. W/VE non-members' fee is $3.

6-Meter "600 Club" Award

This award uses contacts made on or after January 1, 1977. There are three variables built into the scoring system: QSOs, ARRL sections, and countries. Each new contact counts 2 points, new sections earn 6, new "countries" tally 25. A total of 600 or more points earns the award. The application furnishes full details regarding the ARRL section structure. Be sure to send for the form before you apply! Applicants in the U.S., Puerto Rico and U.S. possessions, and Canada, must be ARRL members and sufficient funds must be included for safe return of the QSLs. (Novices and DX applicants are exempt from fees.)

Code Proficiency Award

Many hams can follow the general idea of a contact "by ear" but when pressed to "write it down" they "muff" the copy. The Code Proficiency Award permits each amateur to prove himself as a proficient operator, and sets up a system of awards for step-by-step gains in copying proficiency. It enables every amateur to check his code proficiency, to better that proficiency, and to receive a certification of his receiving speed.

This program is a whale of a lot of fun. The League will give a certificate to any interested individual, who demonstrates that he can copy perfectly, for at least one minute, plain-language Continental code at 10, 15, 20, 25, 30, 35 or 40 words per minute, as transmitted monthly from W1AW and W6OWP.

As part of the ARRL Code Proficiency program W1AW transmits plain-language practice material frequently at speeds from 5 to 35 wpm, occasionally in reverse order. All amateurs are invited to use these transmissions to increase their code-copying ability. Non-amateurs are invited to utilize the lower speeds, 5, 7-1/2 and 10 wpm, which are transmitted for the benefit of persons studying the code in preparation for the amateur license examination. Refer to any issue of *QST* (Operating Events) for details.

Rag Chewers Club

The Rag Chewers Club is designed to encourage friendly contacts and discourage the "hello-good-bye" type of QSO. It furthers fraternalism through amateur radio.

Membership certificates are awarded to amateurs who report a fraternal-type contact with another amateur lasting a half hour or longer. This does not mean a half hour spent trying to get a message through or in trying to work a rare DX station, but a solid half hour of pleasant "visiting" with another amateur discussing subjects of mutual interest and getting to know each other. If nominating someone for RCC, please send the information to the nominee who will (in turn) apply to Headquarters for RCC.

Members sign "RCC" after their calls to indicate that they are interested in a chat, not just a contact. There is no fee for W/VE members and DX, a 25¢ fee for others.

Operating Aids

The following Operating Aids are available free, upon request: Emergency Operating, DX Operating Code, Contest Duplicate Contact Record, DXCC Countries List, WAS Record, ARRL Message Form, Ready Reference Information, A composite aid – Ending Signals, Time Conversion, Phonetic Alphabets, RST System and Steps in an Emergency, Emergency Reference Information.

A-1 Operator Club

The A-1 Operator Club should include in its ranks every good operator. To become a member, one must be nominated by at least two operators who already belong. General keying or voice technique, procedure, copying ability, judgment and courtesy all count in rating candidates under the club rules detailed at length in *Operating an Amateur Radio Station*. Aim to make yourself a fine operator, and one of these days you may be pleasantly surprised by an invitation to belong to the A-1 Operator Club, which carries a worthwhile certificate in its own right.

Brass Pounders League

Every individual reporting more than a specified minimum of official monthly traffic totals is given an honor place in the *QST* listing known as the Brass Pounders League and a certificate to recognize his performance is furnished by the SCM. In addition, a *BPL Traffic Award* (medallion) is given to individual amateurs working at their own stations after the third time they "make BPL" provided it is duly reported to the SCM and recorded in *QST*.

Public Service Honor Roll

A new listing, supplementing the BPL, was started in 1970. It takes into account the many public service functions of amateurs in addition to the handling of record messages. Points can be claimed for checking into and participating in nets, for serving as net control stations, as liaison between nets, for handling phone patches, for making BPL, for handling real emergency traffic and for serving as a net manager. Each such function has a maximum number of points per month so that nobody can make the PSHR by performing a single type of function, except handling emergency traffic. Versatility in public service is encouraged and rewarded. See *QST* for details.

Old Timers Club

The Old Timers Club is open to anyone who holds an amateur call at the present time, and who held an amateur license (operator or station) 20-or-more years ago. Lapses in activity during the intervening years are permitted.

If you can qualify as an "Old Timer," send an outline of your ham career. Indicate the date of your first amateur license and your present call. If eligible for the OTC, you will be added to the roster and will receive a membership certificate.

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 Kingdom
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
SSN-STZ	Sudan
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 Republics
URA-UTZ	Ukrainian Soviet Socialist Rep.
UUA-UZZ	Union of Soviet Socialist Republics
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	Viet Nam
XWA-XWZ	Laos
XXA-XXZ	Portuguese Overseas Provinces
XYA-XZZ	Burma
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
3EA-3FZ	Panama
3GA-3GZ	Chile
3HA-3UZ	China
3VA-3VZ	Tunisia
3WA-3WZ	Viet Nam
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
4TA-4TZ	Peru
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
5XA-5XZ	Uganda
5YA-5ZZ	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

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
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
DX	Distance, foreign countries
ES	And, &
FB	Fine business; excellent
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
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
SASE	Self-addressed, stamped envelope
SED	Said
SIG	Signature; signal
SINE	Operator's personal initials or nickname
SKED	Schedule
SRI	Sorry
SVC	Service; prefix to service message
TFC	Traffic
TMW	Tomorrow
TNX-TKS	Thanks
TT	That
TU	Thank you
TVI	Television interference
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
XMTR (TX)	Transmitter
XTAL	Crystal
XYL (YF)	Wife
YL	Young lady
73	Best regards
88	Love and kisses

▲ *Operating an Amateur Radio Station* covers the details of practical amateur operating. In it you will find information on Operating Practices, Emergency Communication, ARRL Operating Activities and Awards, the ARRL Field Organization, Handling Messages, Network Organization, "Q" Signals and Abbreviations used in amateur operating, and other helpful material. It's a handy reference that will serve to answer many of the questions concerning operating that arise during your activities on the air.

▲ *Public Service Communications* is the "bible" of the Amateur Radio Public Service Corps. Within its pages are contained the fundamentals of operation of the Amateur Radio Emergency Corps (AREC), the National Traffic System (NTS), and the Radio Amateur Civil Emergency Service (RACES), including diagrams of how each is organized and how it operates. The role of the American Red Cross and FCC's regulations concerning amateur operation in emergencies also come in for some special attention.

The two publications described above may be obtained without charge by any *Handbook* reader. Either or both will be sent upon request.

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Vacuum Tubes and Semiconductors

For the convenience of the designer, the receiving-type tubes listed in this chapter are grouped by filament voltages and construction types (glass, metal, miniature, etc.). For example, all miniature tubes are listed in Table I, all metal tubes are in Table II, and so on.

Transmitting tubes are divided into triodes and tetrodes-pentodes, then listed according to rated plate dissipation. This permits direct comparison of ratings of tubes in the same power classification.

For quick reference, all tubes are listed in numerical-alphabetical order in the index. Types having no table reference are either obsolete or of little use in amateur equipment. Base diagrams for these tubes are listed.

Tube Ratings

Vacuum tubes are designed to be operated within definite maximum (and minimum) ratings. These ratings are the maximum safe operating voltages and currents for the electrodes, based on inherent limiting factors such as permissible cathode temperature, emission, and power dissipation in electrodes.

In the transmitting-tube tables, maximum ratings for electrode voltage, current and dissipation are given separately from the typical operating conditions for the recommended classes of operation. In the receiving-tube tables, ratings and operating data are combined. Where only one set of operating conditions appears, the positive electrode voltages shown (plate, screen, etc.) are, in general, also the maximum rated voltages.

For certain air-cooled transmitting tubes, there are two sets of maximum values, one designated as CCS (Continuous Commercial Service) ratings, the other ICAS (Intermittent Commercial and Amateur Service) ratings. Continuous Commercial Service is defined as that type of service in which long tube life and reliability of performance under continuous operating conditions are the prime consideration. Intermittent Commercial and Amateur Service is defined to include the many applications where the transmitter design factors of minimum size, light weight, and maximum power output are more important than long tube life. ICAS ratings are considerably higher than CCS ratings. They permit the handling of greater power, and although such use involves some sacrifice in tube life, the period over which tubes give satisfactory performance in intermittent service can be extremely long.

The plate dissipation values given for transmitting tubes should not be exceeded during normal operation. In plate modulated amplifier applications, the maximum allowable carrier-condition plate dissipation is approximately 66 percent of the value listed and will rise to the maximum value under 100 percent sinusoidal modulation.

Typical Operating Conditions

The typical operating conditions given for transmitting tubes represent, in general, maximum ICAS ratings where such ratings have been given by the manufacturer. They do not represent the *only* possible method of operation of a particular tube type. Other values of plate voltage, plate current, etc., may be used so long as the maximum ratings for a particular voltage or current are not exceeded.

Detailed information and characteristic curves are available from tube and semiconductor manufacutrers, in books sold through radio dealers or direct from the factory.

Semiconductors

The semiconductor tablulation in this chapter is restricted to some of the more common diodes and transistors. The units listed were selected to represent those types that are useful for most amateur radio experimental applications. These diodes and transistors were chosen for their low cost and availability. Most of them can be obtained from the large mail-order houses or from the local manufacturer's distributor. Because there are thousands of diode and transistor types on today's market, this list is by no means complete.

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II — 6.3-Volt Metal Receiving Tubes V18
III — 6.3-Volt Glass Tubes, Octal Bases. V19
IV — Control and Regulator Tubes V19
V — Rectifiers V19
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01-A	—	4D
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0A5	—	Fig. 19
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0B3	V19	4AJ
0C2	V19	5BO
0C3A	V19	4AJ
0D3A	V19	4AJ
0G3	—	5BO
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0Z4	—	4R
0Z4A	—	4R
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1A6	—	6L
1A7GT	—	7Z
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1AB6	—	7DH
1AC6	—	7DH
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1AF5	—	6AU
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1B4	—	4M
1B5	—	6M
1B7GT	—	7Z
1B8GT	—	8AW
1C3	—	5CF
1C5GT	—	6X
1C6	—	6L
1C7G	—	7Z
1C21	—	4V
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1D5GT	—	5R
1D7G	—	7Z
1D8GT	—	8AJ
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1H5GT	—	5Z
1H6G	—	7AA
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1J5G	—	6X
1J6GT	—	7AB
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1K3	V19	3C
1L4	—	6AR
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1LA4	—	5AD
1LA6	—	7AK
1LB4	—	5AD
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2C37	—	Fig. 21
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2C43	V20	Fig. 11
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2C52	—	8BD
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2E25	V22	5BJ
2E26	V22	7CK
2E30	—	7CQ

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2V3G	—	4Y
2W3	—	4X
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2Y2	V19	4AB
2Z2	V19	4B
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3A3	—	8EZ
3A4	—	7BB
3A5	—	7BC
3A8GT	—	8AS
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3AQP1	—	12E
3B4	—	7CY
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3B7	—	7BE
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3B26	—	Fig. 18
3B27	—	4P
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3BP1A	—	14G
3C4	—	6BX
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3C23	—	3G
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3E22	—	8BY
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3EA5	—	7EW
3EP1	—	11N
3FP7	—	14B
3FP7A	—	14J
3GP1-4-5-11	—	11A
3GP1A	—	11N
3GP4A	—	11N
3JP1—12	—	14J
3JP1A—11A	—	14J
3KP1-4-11	—	11M
3LE4	—	6BA
3LF4	—	6BB
3MP1	—	12F
3Q4	—	7BA
3Q5GT	—	7AP
3RP1-4	—	12E
3RP1A	—	12E
3S4	—	7BA
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3-50A4	—	3G
3-50D4	—	2D
3-50G2	—	2D
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5A6	—	9L
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5ADP1-7-11	—	14J
5AMP1	—	14U
5AP1-4	—	11A
5AQP1	—	14G
5AS4A	V20	5T
5AT4	V20	5L
5ATP1-11	—	14V
5AU4	V20	5T
5AW4	V20	5T
5AX4GT	—	5T
5AZ4	—	5T
5BC3	—	9NT
5BP1	—	11A
5BP1A	—	11N
5BP7A	—	11N
5CP1-11	—	14B
5CP1A	—	14J
5CP1B-11B	—	14J
5CP7A	—	14J
5CP11A	—	14J
5CP12	—	14J
5D22	V23	5BK
5DJ4	—	8KS
5EA8	—	9AE
5FV8	—	9FA
5GP1	—	11A
5HP1-4	—	11A
5HP1A	—	11N
5JP1A-4A	—	11S
5LP1A-4A	—	11T
5MP1-11	—	7AN
5NP1-4	—	11A
5R4GY	V20	5T
5R4GYA	V20	5T

Type	Page	Base
5RP1A-4A	—	14P
5SP1-4	—	14K
5T4	—	5T
5U4G	V20	5T
5U4GA-GB	V20	5T
5UP1-11	—	12E
5V3A	V20	5T
5V4GA	V20	5L
5VF7	—	11N
5W4GT	—	5T
5X3	—	4C
5X4G	—	5Q
5XP1	—	14P
5XP1A-11A	—	14P
5Y3-G-GT	V20	5T
5Y4-G-GT	—	5Q
5Z3	V20	4C
5Z4	V20	5L
5-125B	V23	7BM
5-500A	V23	—
6A3	—	4D
6A4	—	5B
6A5GT	—	6T
6A6	—	7B
6A7	—	7C
6A8	V18	8A
6AB4	—	5CE
6AB5	—	6R
6AB6G	—	7AU
6AB7	—	8N
6AB8	—	9AT
6AC5GT	—	6Q
6AC6G	—	7AU
6AC7	V18	8N
6AD5G	—	6Q
6AD6G	—	7AG
6AD7G	—	8AY
6AD8	—	9T
6AE5G	—	6Q
6AE6G	—	7AH
6AE7GT	—	7AX
6AE8	V24	8DU
6AF3	—	9CB
6AF4A	V16	7DK
6AF5G	—	6Q
6AF6G	—	7AG
6AF7G	—	8AG
6AG5	V16	7BD
6AG6G	—	7S
6AG7	V18	8Y
6AH4GT	—	8EL
6AH5G	—	6AP
6AH6	V16	7BK
6AH7GT	—	8BE
6AJ4	V16	9BX
6AJ5	—	7BD
6AJ7	—	8N
6AJ8	—	9CA
6AK5	V16	7BD
6AK6	V16	7BK
6AK7	—	8Y
6AK8	—	9E
6AL3	—	9CB
6AL5	V16	6BT
6AL6G	—	6AM
6AL7GT	V19	8CH
6AM4	V16	9BX
6AM5	—	6CH
6AM6	—	7DB
6AM8A	—	9CY
6AN4	—	7DK
6AN5	V16	7BD
6AN6	—	7BJ
6AN7	—	9Q
6AN8A	V16	9DA
6AQ4	—	7DT
6AQ5A	V16	7BZ
6AQ6	V16	7BT
6AQ7GT	V19	8CK
6AQ8	—	9AJ
6AR5	V16	6CC
6AR6	V19	6BQ
6AR7GT	V19	7DE
6AR8	—	9DP
6AS5	—	7CV
6AS6	V16	7CM
6AS7GA	V19	8BD
6AS8	—	9DS
6AT6	V16	7BT
6AT8A	—	9DW
6AU4GT	V19	4CG
6AU5GT	V16	6CK
6AU6A	V16	7BK
6AU7	—	9A
6AU8A	—	9DX
6AV4	V20	5BS
6AV5GA	—	6CK
6AV5GT	—	6CK
6AV6	V16	7BT
6AV11	—	12BY
6AW7GT	—	8CQ
6AW8A	—	9DX
6AX4GT	V20	4CG
6AX5GT	V20	6S
6AX6G	—	7Q
6AX7	—	9A
6AX8	—	9AE
6AZ8	V16	9ED
6B4G	—	5S
6B5	—	6AS
6B6G	—	7V
6B7	—	7D
6B8	V18	8E
6B10	—	12BF
6BA6	V16	7BK
6BA7	V16	8CT
6BA8A	—	9DX
6BC4	V16	9DR
6BC5	—	7BD
6BC7	—	9AX
6BC8	—	9AJ
6BD4A	—	Fig. 80
6BD5GT	—	6CK
6BD6	—	7BK
6BD7	—	9Z
6BE6	V16	7CH
6BE7	—	9AA
6BE8A	V16	9EG
6BF5	V16	7BZ
6BF6	V16	7BT
6BG6GA	—	5BT
6BH5	—	9AZ
6BH6	V16	7CM
6BH8	V16	9DX
6BJ5	—	9CH
6BJ6A	V16	7CM
6BJ7	V16	9AX
6BJ8	V16	9ER

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6BK6	V16	7BT
6BK7A	—	9AJ
6BK7B	V16	9AJ
6BL7GTA	V19	8BD
6BL8	V16	9DC
6BM5	—	7BZ
6BN4A	V16	7EG
6BN6	V16	7DF
6BN7	—	9AJ
6BN8	V16	9ER
6BQ5	V16	9CV
6BQ6GTB/6CU6	V19	6AM
6BQ7A	V16	9AJ
6BR7	—	9BC
6BR8A	V16	9FA
6BS5	—	9BK
6BS7	—	9BB
6BS8	V16	9AJ
6BT6	—	7BT
6BT8	—	9FE
6BU5	—	8FP
6BU6	—	7BT
6BU8	—	9FG
6BV7	—	9BU
6BV8	—	9FJ
6BW4	V20	9DJ
6BW6	—	9AM
6BW7	—	9AQ
6BW8	—	9HK
6BW11	—	12HD
6BX4	V20	5BS
6BX6	—	9AQ
6BX7GT	V19	8BD
6BX8	V16	9AJ
6BY5G	V20	6CN
6BY6	—	7CH
6BY7	—	9AQ
6BY8	—	9FN
6BZ6	V16	7CM
6BZ7	V16	9AJ
6BZ8	V16	9AJ
6C4	V16	6BG
6C4	V20	6BG
6C5	—	6Q
6C6	—	6F
6C7	—	7G
6C8G	—	8G
6C10	—	12BQ
6CA4	V20	9M
6CA5	—	7CV
6CB5A	V19	8GD
6CB6A	V16	7CM
6CD6GA	V19	5BT
6CE5	V16	7BD
6CF6	—	7CM
6CG6	V16	7BK
6CG7	V16	9AJ
6CG8A	—	9GF
6CH6	—	9BA
6CH7	—	9EW
6CH8	—	9F
6CJ6	—	9AS
6CK4	V19	8JB
6CK6	—	9AR
6CL5	V19	8GD
6CL6	V16	9BV
6CL8A	—	9FX
6CM6	—	9CK
6CM7	—	9ES
6CM8	—	9FZ
6CN7	—	9EN
6CQ6	—	7DB
6CQ8	—	9GE
6CR6	—	7EA
6CR8	—	9GJ
6CS5	—	9CK
6CS6	—	7CH
6CS7	—	F9E
6CS8	—	9FZ
6CU5	—	7CV
6CU6	V19	6AM
6CU8	—	9GM
6CW4	V16	12AQ
6CW5	—	9CV
6CX7	—	9FC
6CX8	V16	9DX
6CY5	V16	7EW
6CY7	—	9EF
6CZ5	—	9HN
6D4	—	5AY
6D6	—	6F
6D7	—	7H
6D8G	—	8A
6D10	—	12BQ
6DA4A	—	4CQ
6DB5	—	9GR
6DB6	—	7CM
6DC6	—	7CM
6DE4	V20	4CG
6DE6	—	7CM
6DE7	—	9HF
6DG6GT	—	7S
6DJ8	V16	9AJ
6DK6	V16	7CM
6DN6	V19	5BT
6DN7	V19	8BD
6DQ5	V19	8JC
6DQ6B	V19	6AM
6DR7	—	9HF
6DS4	V16	12AQ
6DS5	—	7BZ
6DT5	—	9HN
6DT6	V16	7EN
6DT8	—	9DE
6DV4	—	12EA
6DW5	V16	9CK
6DX4	—	7DK
6DX8	—	9HX
6DZ4	—	7DK
6DZ7	V19	8JP
6E5	V19	6R
6E6	—	7B
6E7	—	7H
6E8G	—	8O
6EA5	—	7EW
6EA7	V19	8BD
6EA8	V16	9AE
6EB5	—	6BT
6EB8	V16	9DX
6EF6	V19	7S
6EH5	V16	7CV
6EH7	V16	9AQ
6EH8	V16	9JG
6EJ7	V16	9AQ
6ER5	V16	7FN
6ES5	—	7FP

Type	Page	Base
6ES8	V17	9DE
6EU7	V17	9LS
6EU8	V17	9JF
6EV5	—	7EW
6EW6	—	7CM
6EW7	—	9HF
6EX6	—	5BT
6EY6	V19	7AC
6EZ5	V19	7AC
6EZ8	V17	9KA
6F4	V20	7BR
6F4	V25	7BR
6F5	—	5M
6F6	V18	7S
6F7	—	7E
6F8G	—	8G
6FD7	—	9HF
6FG5	—	7GA
6FG7	—	9GF
6FH5	—	7FP
6FH6	V19	6AM
6FJ7	—	12BM
6FM7	—	12BJ
6FM8	—	9KR
6FQ5A	—	7FP
6FS5	—	7GA
6FV6	V17	7FQ
6FV8A	—	9FA
6FW8	—	9AJ
6FY5	—	7FN
6G5	—	6R
6G6G	—	7S
6GC5	V17	9EU
6GE5	—	12BJ
6GF5	—	12BJ
6GH8A	—	9AE
6GJ5	—	9NM
6GJ8	V17	9AE
6GK5	V17	7FP
6GK6	V17	9GK
6GM6	V17	7CM
6GM8	—	9DE
6GN8	V17	9DX
6GS8	—	9LW
6GT5	—	9NZ
6GU5	—	7GA
6GV8	—	9LY
6GW5	—	7GK
6GW6	V19	6AM
6GY8	—	9MB
6GZ5	—	7CV
6H4GT	—	5AF
6H5	—	6R
6H6	—	7Q
6H8G	—	8E
6HA5	—	76M
6HB5	—	12BJ
6HB6	V17	9PU
6HB7	—	9QA
6HF5	—	12FB
6HF8	—	9DX
6HG5	—	7BZ
6HJ8	—	9CY
6HK5	—	7GM
6HM5	—	76M
6HQ5	—	7GM
6HS6	—	7BK
6HS8	—	9LW
6HZ6	—	7EN
6HZ8	—	9DX
6J4	V17	7BQ
6J5	V18	6Q
6J6A	V17	7BF
6J6A	V20	7BF
6J7	V18	7R
6J8G	—	8H
6J11	—	12BW
6JB6	—	9QL
6JC6A	—	9PM
6JC8	—	9PA
6JD6	—	9PM
6JE6	—	9QL
6JE6A	—	9QL
6JF6	—	9QL
6JH6	—	7CM
6JH8	—	9DP
6JK8	—	9AJ
6JN8	—	9FA
6JT8	—	9DX
6JV8	—	9DX
6K5GT	—	5U
6K6GT	V19	7S
6K7	V18	7R
6K8	V18	8K
6K11	—	12BY
6KD6	—	12GW
6KD8	V17	9AE
6KE8	V17	9DC
6KM6	—	9QL
6KR8	V17	9DX
6KT6	V17	9PM
6KT8	V17	9QP
6KV8	—	9DX
6KY6	—	9GK
6KZ8	V17	9FZ
6L4	—	7BR
6L5G	—	6Q
6L6GA	—	7S
6L6GB	V18	7S
6L6GX	—	7S
6L7	V18	7T
6LF6	—	12GW
6LJ8	V17	9GF
6LM8	—	9AE
6LQ6	—	9QL
6LY8	V17	9DX
6M5	—	9N
6M6G	—	7S
6M7G	—	7R
6M8GT	—	8AU
6M11	—	12CA
6MQ8	—	9AE
6MU8	V17	9AE
6N4	—	7CA
6N5	—	6R
6N6G	—	7AU
6N7GT	V18	8B
6N7GT	V20	8B
6N8	—	9T
6P5GT	—	6Q
6P7G	—	7U
6P8G	—	8K
6Q4	—	9S
6Q5G	—	6Q
6Q6G	—	6Y
6Q7	V18	7V
6Q11	—	12BY
6R4	—	9R

Type	Page	Base
6R6G	—	6AW
6R7	V18	7V
6R8	—	9E
6S4	—	9AC
6S4A	—	9AC
6S6GT	—	5AK
6S7	—	7R
6S8GT	—	8CB
6SA7GT	V18	8R
6SB7Y	V18	8R
6SC7	V18	8S
6SD7GT	—	8N
6SE7GT	—	8N
6SF5	—	6AB
6SF7	V18	7AZ
6SG7	V18	8BK
6SH7	V18	8BK
6SH7L	—	8BK
6SJ7	V18	8N
6SJ7Y	V21	8N
6SK7	V18	8N
6SL7GT	V19	8BD
6SN7GTA	—	8BD
6SN7GTB	V19	8BD
6SQ7GT	V18	8Q
6SR7	V18	8Q
6SS7	—	8N
6ST7	—	8Q
6SU7GTY	—	8BD
6SV7	—	7AZ
6SZ7	—	8Q
6T4	V17	7DK
6T5	—	6R
6T6GM	—	6Z
6T7	—	7V
6T8	—	9E
6T8A	V17	9E
6T9	—	12FM
6U3	—	9BM
6U4GT	—	4CG
6U5	—	6R
6U6GT	—	7S
6U7G	—	7R
6U8	—	9AE
6U8A	V17	9AE
6V3	—	9BD
6V3A	—	9BD
6V4	V20	9M
6V5GT	—	6AO
6V6GTA	V18	7S
6V7G	—	7V
6V8	—	9AH
6W4GT	—	4CG
6W5G	—	6S
6W6GT	V19	7S
6W7G	—	7R
6X4/6063	V20	7CF
6X5GT	V20	6S
6X6G	—	7AL
6X8	—	9AK
6X8A	V17	9AK
6Y3G	—	4AC
6Y5	—	6J
6Y6G	—	7S
6Y6GA	V19	7S
6Y6GT	—	7S
6Y7G	—	8B
6Z3	V20	4G
6Z4	—	5D
6Z5	—	6K
6Z7G	—	8B
6ZY5G	—	6S
7A4	—	5AC
7A5	—	6AA
7A6	—	7AJ
7A7	—	8V
7A8	—	8U
7AB7	—	8BO
7AD7	—	8V
7AF7	—	8AC
7AG7	—	8V
7AH7	—	8V
7AJ7	—	8V
7AK7	—	8V
7B4	—	5AC
7B5	—	6AE
7B6	—	8W
7B7	—	8V
7B8	—	8X
7C4	—	4AH
7C5	—	6AA
7C6	—	8W
7C7	—	8V
7D7	—	8AR
7E5	—	8BN
7E6	—	8W
7E7	—	8AE
7EP4	—	11N
7EV6	—	7AC
7EY6	—	7AC
7F7	—	8AC
7F8	—	8BW
7G7	—	8V
7G8	—	8BV
7GP4	—	14G
7H7	—	8V
7J7	—	8BL
7JP1-4-7	—	14R
7K7	—	8BF
7L7	—	8V
7N7	—	8AC
7Q7	—	8AL
7R7	—	8AE
7S7	—	8BL
7T7	—	8V
7V7	—	8V
7VP1	—	14R
7W7	—	8BJ
7X6	—	7AJ
7X7	—	8BZ
7Y4	—	5AB
7Z4	—	5AB
8BP4	—	14G
9BM5	—	7BZ
9BW6	—	9AM
9NP1	—	6BN
10	—	4D
10EB8	—	9DX
10GP4	—	14G
10HP4	—	14G
10Y	—	4D
11/12	—	4F
12A4	—	9AG
12A5	—	7F
12A6	—	7S
12A7	—	7K
12A8GT	—	8A
12AB5	V17	9EU
12AB6	—	7CC
12AC6	—	7BK
12AD6	—	7CH
12AD7	—	9A
12AE6A	—	7BT
12AE7	—	9A

Type	Page	Base
12AF6	—	7BK
12AG6	—	7CH
12AH7GT	—	8BE
12AH8	—	9BP
12AJ6	—	7BT
12AL5	—	6BT
12AL8	—	9GS
12AQ5	V17	7BZ
12AT6	—	7BT
12AT7	V17	9A
12AU6	—	7BK
12AU7A	V20	9A
12AU7A	V17	9A
12AV5GA	—	6CK
12AV6	—	7BT
12AV7	V17	9A
12AW6	—	7CM
12AW7	—	7CM
12AX4GT	—	4CG
12AX4GTA	—	4CG
12AX7A	V17	9A
12AY7	V17	9A
12AZ7A	V17	9A
12B4	—	9AG
12B4A	—	9AG
12B6M	—	6Y
12B7	—	8V
12B7ML	—	8V
12B8GT	—	8T
12BA6	—	7BK
12BA7	—	8CT
12BD6	—	7BK
12BE6	—	7CH
12BF6	—	7BT
12BH7	—	9A
12BH7A	V17	9A
12BK5	—	9BQ
12BK6	—	7BT
12BL6	—	7BK
12BN6	—	7DF
12BQ6GA	—	6AM
12BQ6GT	—	6AM
12BQ6GTB	—	6AM
12BR7A	—	9CF
12BT6	—	7BT
12BU6	—	7BT
12BW4	—	9DJ
12BV7	—	9BF
12BX6	—	9AQ
12BY7	—	9BF
12BY7A	V17	9BF
12BZ6	—	7CM
12BZ7	—	9A
12C5	—	7CV
12C8	—	8E
12CA5	—	7CV
12CM6	—	9CK
12CN5	—	7CV
12CR6	—	7EA
12CS5	—	9CK
12CS6	—	7CH
12CT8	—	9DA
12CU5	—	7CV
12CU6	—	6AM
12CX6	—	7BK
12DB5	—	9GR
12DE8	—	Fig. 81
12DF5	—	9BS
12DF7	—	9A
12DK7	—	9HZ
12DL8	—	9HR
12DM7	—	9A
12DQ6A	—	6AM
12DQ7	—	9BF
12DS7	—	9JU
12DT5	—	9HN
12DT6	—	7EN
12DT7	—	9A
12DT8	—	9DE
12DU7	—	9JX
12DV7	—	9JY
12DV8	—	9HR
12DW5	—	9CK
12DW7	—	9A
12DW8	—	9JC
12DY8	—	9JD
12DZ6	—	7BK
12E5GT	—	6Q
12EA6	—	7BK
12EC8	—	9FA
12ED5	—	7CV
12EF6	—	7S
12EG6	—	7CH
12EK6	—	7BK
12EL6	—	7FB
12EM6	—	9HV
12EN6	—	7S
12F5GT	—	5M
12F8	—	9FH
12FK6	—	7BT
12FM6	—	7BT
12FP7	—	14E
12FQ8	—	9KT
12FR8	—	9KU
12FT6	—	7BT
12FX5	—	7CV
12FX8A	—	9KV
12G4	—	6BG
12G7G	—	7V
12G8	—	9CZ
12GA6	—	7CH
12GE5	—	12BJ
12GJ5	—	9NM
12GW6	—	6AM
12GN7	—	9BF
12GP7	—	14S
12H4	—	7DW
12H6	—	7Q
12HP7	—	11J
12J5GT	—	6Q
12J7GT	—	7R
12J8	—	9GC
12K5	—	7EK
12K7GT	—	7R
12K8	—	8K
12L6GT	—	7S
12L8GT	—	8BU
12Q7GT	—	7V
12R5	—	7CV
12S8GT	—	8CB
12SA7	—	8R
12SC7	—	8S
12SF5	—	6AB
12SF7	—	7AZ
12SG7	—	8BK
12SH7	—	8BK
12SJ7	—	8N
12SK7	—	8N
12SL7GT	—	8BD
12SN7GT	—	8BD
12SN7GTA	—	8BD
12SQ7	—	8Q
12SR7	—	8Q
12SW7	—	8Q

Type	Page	Base
12SX7	—	8BD
12SY7	—	8R
12U7	—	9A
12V6GT	—	7S
12W6GT	—	7S
12X4	V20	5BS
12Z3	—	4G
12Z5	—	7L
14A4	—	5AC
14A5	—	6AA
14A7	—	8V
14AF7	—	8AC
14AP1-4	—	12A
14B6	—	8W
14B8	—	8X
14C5	—	6AA
14C7	—	8V
14E6	—	8W
14E7	—	8AE
14F7	—	8AC
14F8	—	8BW
14H7	—	8V
14J7	—	8BL
14N7	—	8AC
14Q7	—	8AL
14R7	—	8AE
14S7	—	8BL
14V7	—	8V
14W7	—	8BJ
14X7	—	8BZ
14Y4	—	5AB
14Z3	—	4G
15	—	5F
15A6	—	9AR
15E	—	Fig. 51
16A5	—	9BL
17	—	3G
17Z3	—	9CB
18	—	6B
18FW6A	—	7CC
18FX6A	—	7CH
18FY6A	—	7BT
19	—	6C
19CL8A	—	9FX
19X3	—	9BM
19Y3	—	9BM
20	—	4D
20AP1-4	—	12A
20J8GT	—	8H
21A6	—	9AS
21A7	—	8AR
21EX6	—	5BT
22	—	4K
24-A	—	5E
24-G	—	2D
24XH	—	Fig. 1
25A6	—	7S
25A7GT	—	8F
25AC5GT	—	6Q
25AV5GA	—	6CK
25AV5GT	—	6CK
25AX4GT	—	4CG
25B5	—	6D
25B6G	—	7S
25B8GT	—	8T
25BK5	—	9BQ
25BQ6GA	—	6AM
25BQ6GT	—	6AM
25BQ6GTB	—	6AM
25C5	—	7CV
25C6G	—	7AC
25C6GA	—	7S
25CA5	—	7CV
25CD6G	—	5BT
25CD6GA	—	5BT
25CD6GB	—	5BT
25CU6	—	6AM
25D8GT	—	8AF
25DN6	—	5BT
25DQ6	—	6AM
25EC6	—	5BT
25EH5	—	7CV
25F5	—	7CV
25L6GT	—	7S
25N6G	—	7W
25S	—	6M
25SA7GT	—	8AD
25T	—	3G
25W4GT	—	4CG
25W6GT	—	7S
25X6GT	—	7Q
25Y4GT	—	5AA
25Y5	—	6E
25Z3	—	4G
25Z4	—	5AA
25Z5	V20	6E
25Z6	—	7Q
26	—	4D
26A6	—	7BK
26A7GT	—	8BU
26BK6	—	7BT
26C6	—	7BT
26CG6	—	7BK
26D6	—	7CH
26Z5W	—	9BS
27	—	5A
28Z5	—	5AB
30	—	4D
31	—	4D
32	—	4K
32ET5	—	7CV
32L7GT	—	8Z
33	—	5K
34	—	4M
34GD5	—	7CV
35/51	—	5E
35A5	—	6AA
35B5	V17	7BZ
35C5	—	7CV
35L6GT	—	7S
35T	—	3G
35TG	—	2D
35W4	V20	5BQ
35Y4	—	5AL
35Z3	—	4Z
35Z4GT	V20	5AA
35Z5GT	V20	6AD
35Z6G	—	7Q
36	—	5E
36AM3	V20	5BQ
37	—	5A
38	—	5F
39/44	—	5F
40	—	4D
40Z5GT	—	6AD
41	—	6B
42	—	6B
43	—	6B
45	—	4D
45Z3	—	5AM
45Z5GT	—	6AD
46	—	5C
47	—	5B
48	—	6A

Type	Page	Base
49	—	5C
50	—	4D
50A5	—	6AA
50AX6G	—	7Q
50B5	V17	7BZ
50BK5	—	9BQ
50C5	—	7CV
50C6G	—	7S
50C6GA	—	7S
50DC4	V20	5BQ
50FK5	V17	7CV
50L6GT	—	7S
50T	—	2D
50X6	—	7AJ
50Y6GT	V20	7Q
50Y7GT	—	8AN
50Z6G	—	7Q
50Z7G	—	8AN
51	—	5E
52	—	5C
53	—	7B
53A	—	Fig. 53
55	—	6G
56	—	5A
56AS	—	5A
57	—	6F
57AS	—	6F
58	—	6F
58AS	—	6F
59	—	7A
70A7GT	—	8AB
70L7GT	—	8AA
71-A	—	4D
72	—	4P
73	—	4Y
75	—	6G
75TH	—	2D
75TL	—	2D
76	—	5A
77	—	6F
78	—	6F
79	—	6H
80	V20	4C
81	—	4B
82	—	4C
83	V20	4C
83-V	V20	4AD
84/6Z4	—	5D
85	—	6G
85AS	—	6G
89	—	6F
90C1		5BO
99	—	4D
100TH	V21	2D
100TL	V21	2D
111H	—	2D
112-A	—	4D
117L7GT	—	8AO
117M7GT	—	8AO
117N7GT	V20	8AV
117P7GT	—	8AV
117Z3	V20	4CB
117Z4GT	—	5AA
117Z6GT	—	7Q
128AS	—	5A
150T	—	2N
152TH	—	4BC
152TL	—	4BC
175A	V23	Fig. 78
182-B	—	4D
183	—	4D
203-A	—	4E
203-H	—	3N
204-A	—	Fig. 39
205-D	—	4D
211	—	4E
212-E	—	Fig. 43
217-A	—	4AT
217-C	—	4AT
227-A	—	Fig. 53
241-B	—	Fig. 44
242-A	—	4E
242-B	—	4E
242-C	—	4E
249-B	—	Fig. 29
250TH	V21	2N
250TL	V21	2N
254	—	2N
254-A	—	Fig. 57
254-B	—	Fig. 57
261-A	—	4E
270-A	—	Fig. 39
276-A	—	4E
282-A	—	Fig. 57
284-B	—	3N
284-D	—	4E
295-A	—	4E
300T	—	2N
303-A	—	4E
304-A	—	Fig. 39
304-B	—	2D
304TH	V21	4BC
304TL	V21	4BC
305-A	—	Fig. 59
306-A	—	Fig. 63
307-A	—	Fig. 61
308-B	—	Fig. 43
310	—	4D
311	—	4E
311CH	—	Fig. 32
312-A	—	Fig. 68
312-E	—	Fig. 44
327-A	—	Fig. 50
327-B	—	Fig. 50
342-B	—	4E
356-A	—	Fig. 55
361-A	—	4E
376-A	—	4E
408A	—	7BD
417-A	V17	9V
482-B	—	4D
483	—	4D
485	—	5A
527	—	Fig. 53
559	—	Fig. 10
572B	V21	3G
575-A	—	4AT
592	—	Fig. 28
705-A	—	Fig. 45
717-A	—	8BK
756	—	4D
800	—	2D
801A/801	—	4D
802	—	6BM
803	V23	5J
804	—	Fig. 61
805	—	3N
806	—	2N
807	V22	5AW
807W	V22	5AW
808	—	2D
809	—	3G
810	V21	2N
811	—	3G

Type	Page	Base
811A	V21	3G
812	—	3G
812A	V21	3G
812H	—	3G
813	V23	5BA
814	V22	Fig. 64
815	V22	8BY
816	V20	4P
822	—	3N
822B	—	2N
826	—	7BO
828	—	5JP
829	—	7BP
829A	—	7BP
829B	V22	7BP
830	—	4D
830B	—	3G
831	—	Fig. 40
832	—	7BP
832A	V22	7BP
833A	V21	Fig. 41
834	—	2D
835	—	4E
836	V20	4P
837	—	6BM
838	—	4E
840	—	5J
841	—	4D
841A	—	3G
841SW	—	3G
843	—	5A
844	—	5AW
849	—	Fig. 39
850	—	Fig. 47
852	—	2D
860	—	Fig. 58
861	—	Fig. 42
864	—	4D
865	—	Fig. 57
866	—	4P
866A-AX	V20	4P
866B	V20	4P
866jr	V20	4B
871	—	4P
872A/872	V20	4AT
874	—	4S
878	—	4P
879	—	4AB
884	—	6Q
885	—	5A
902A	—	8CD
906P1-11	—	7AN
908A	—	7CE
909	—	5BP
910	—	7AN
911	—	7AN
914A	—	6BF
930B	—	3G
938	—	4E
950	—	5K
951	—	4M
954	—	5BB
955	—	5BC
955	—	5BC
956	—	5BB
957	—	5BD
958	—	5BD
958A	—	5BD
958A	—	5BD
959	—	5BE
967	—	3G
975A	—	4AT
1003	—	4R
1005	—	5AQ
1006	—	4C
1201	—	8BN
1203	—	4AH
1204	—	8BO
1206	—	8BV
1218A	—	7DK
1221	—	6F
1223	—	7R
1229	—	4K
1230	—	4D
1231	—	8V
1232	—	8V
1265	—	4AJ
1266	—	4AJ
1267	—	4V
1273	—	8V
1274	—	6S
1275	—	4C
1276	—	4D
1280	—	8V
1284	—	8V
1291	—	7BE
1293	—	4AA
1294	—	4AH
1299	—	6BB
1602	—	4D
1603	—	6F
1608	—	4D
1609	—	5B
1610	—	Fig. 62
1611	—	7S
1612	—	7T
1613	—	7S
1614	V22	7AC
1616	—	4P
1619	—	Fig. 74
1620	V18	7R
1621	—	7S
1622	—	7AC
1623	V20	3G
1624	—	Fig. 66
1625	V22	5AZ
1626	—	6Q
1627	—	2N
1628	—	Fig. 54
1629	—	6RA
1631	—	7AC
1632	—	7S
1633	—	8BD
1634	—	8S
1635	V19	8B
1641	—	Fig. 52
1642	—	7BH
1644	—	Fig. 4
1654	—	2Z
1802P1-11	—	11A
1805P1-4	—	11A
1806P1	—	11N
1851	—	7R
1852	V18	8N
1853	—	8N
2002	—	Fig. 1
2005	—	Fig. 1
2050	—	8BA
2051	—	8BA
2523N/128A	—	5A
4604	—	7CL
4652	—	Fig. 51
5514	—	4BO
5516	—	7CL

Type	Page	Base
5517	—	5BU
5556	—	4D
5562	—	Fig. 30
5590	—	7BD
5591	—	7BD
5608	—	7BD
5608A	—	7B
5610	—	6CG
5618	—	7CU
5651	V19	5BO
5654	—	7BD
5656	—	9F
5662	V19	Fig. 79
5663	—	6CE
5670	—	8CJ
5675	V20	Fig. 21
5679	—	7CX
5686	V17	9G
5687	V17	9H
5690	—	Fig. 38
5691	—	8BD
5692	—	8BD
5693	V18	8N
5694	—	8CS
5696	V19	7BN
5722	V17	5CB
5725	—	7CM
5726	—	6BT
5727	V19	7BN
5731	—	5BC
5749	—	7BK
5750	—	7CH
5751	—	9A
5755	—	9J
5763	V22	9K
5764	—	Fig. 21
5765	—	Fig. 21
5766	See 2C37	
5767	See 2C37	
5768	—	Fig. 21
5794	—	Fig. 21
5812	—	7CQ
5814	—	9A
5823	V19	4CK
5824	—	7S
5825	—	4P
5839	—	6S
5842	V17	9V
5844	—	7BF
5845	—	5CA
5847	—	9X
5852	—	6S
5857	—	9AB
5866	V21	Fig. 3
5867	—	Fig. 3
5871	—	7AC
5876	—	Fig. 21
5879	V17	9AD
5881	—	7AC
5890	—	12J
5893	V20	Fig. 21
5894A	V22	Fig. 7
5910	—	6AR
5915	—	7CH
5920	—	7BF
5933	V22	5AW
5961	—	8R
5962	V19	2AG
5963	—	9A
5964	—	7BF
5965	—	9A
5993	—	Fig. 35
5998	V19	8BD
6005	—	7BZ
6023	—	9CD
6026	—	Fig. 16
6028	—	7BD
6045	—	7BF
6046	—	7AC
6057	—	9A
6058	—	6BT
6059	—	9BC
6060	—	9A
6061	—	9AM
6062	—	9K
6063	V20	7CF
6064	—	7DB
6065	—	7DB
6066	—	7BT
6067	—	9A
6072	—	9A
6073	V19	5BO
6074	V19	5BO
6080	—	8BD
6082	—	8BD
6083	—	Fig. 5
6084	—	9BJ
6085	—	9A
6086	—	9BK
6087	—	5L
6101	—	7BF
6132	—	9BA
6135	—	6BG
6136	—	7BK
6137	—	8N
6140	—	9BY
6141	—	9BZ
6146	V22	7CK
6146A	V22	7CK
6146B	V22	7CK
6155	V23	5BK
6156	V23	5BK
6157	—	Fig. 36
6158	—	9A
6159B	V22	7CK
6173	—	Fig. 34
6186	—	7BD
6197	—	9BV
6201	—	9A
6211	—	9A
6216	—	Fig. 37
6218	—	9CG
6227	—	9BA
6252	V22	Fig. 7
6263	—	—

Type	Page	Base
6264	—	—
6265	—	7CM
6287	—	9CT
6308	—	8EX
6336A	—	8BD
6350	—	9CZ
6354	—	Fig. 12
6360	V22	Fig. 13
6374	—	9BW
6386	V17	8CJ
6417	V22	9K
6443	—	9BW
6485	—	7BK
6524	V22	Fig. 76
6550	V19	7S
6627	—	5BO
6660	—	7CC
6661	—	7CM
6662	—	7CM
6663	—	6BT
6664	—	5CE
6669	—	7BZ
6676	—	7CM
6677	—	9BV
6678	—	9AE
6679	—	9A
6680	—	9A
6681	—	9A
6816	V23	Fig. 77
6829	—	9A
6850	V22	Fig. 76
6883	V22	7CK
6884	V23	Fig. 77
6887	V17	6BT
6893	V22	7CK
6897	—	—
6907	—	Fig. 7
6939	V22	Fig. 13
6973	V17	9EU
7000	—	7R
7025	—	9A
7027A	V19	8HY
7034	V23	Fig. 75
7035	V23	Fig. 75
7054	—	9BF
7055	—	6BT
7056	—	6CM
7057	—	9AJ
7058	—	9A
7059	—	9AE
7060	—	9DX
7061	—	9EU
7077	—	—
7094	V23	Fig. 82
7137	—	7BQ
7167	—	7EW
7189A	V17	9CV
7247	—	9A
7258	V17	9DA
7270	V22	Fig. 84
7271	V22	Fig. 84
7308	—	9DE
7360	—	9KS
7408	—	7AC
7543	—	7BK
7551	V22	9LK
7558	V22	9LK
7581A	—	7AC
7586	V17	12AQ
7587	V17	12AS
7591	V19	8KQ
7695	—	9PX
7700	—	6F
7701	—	9MS
7717	—	7EW
7854	V22	Fig. 7
7868	—	9NZ
7895	V17	12AQ
7905	—	9PB
7984	—	12EU
8000	—	2N
8001	V22	7BM
8003	—	3N
8005	—	3G
8008	—	Fig. 8
8012	—	Fig. 54
8013-A	—	4P
8016	—	3C
8020	—	4P
8025	—	4AQ
8032	V22	7CK
8042	—	Fig. 51
8056	V17	12AQ
8058	V18	12CT
8072	V22	Fig. 85
8117	V22	Fig. 7
8121	V23	Fig. 86
8122	V23	Fig. 85
8163	V21	Fig. 3
8166/4-1000A	V23	—
8203	V22	12AQ
8295/172	V23	—
8298A	V22	7CK
8334	—	7DK
8393	V18	12AQ
8458	—	Fig. 13
8627	—	12CT
8628	V18	12AQ
8646	—	—
8677	V18	12CT
8808	—	Fig. 15
8873	V21	Fig. 87
8874	V21	—
8875	V21	—
8877	V21	—
9001	V18	7BD
9002	V18	7BS
9003	V18	7BD
9004	—	4BJ
9005	—	5BG
9006	V18	6BH
AT-340	—	5BK
AX9900	V21	Fig. 3
AX9901	—	Fig. 3

Type	Page	Base
AX9903	V22	Fig. 7
AX9905	—	Fig. 2
AX9909	—	Fig. 5
AX9910	V22	Fig. 7
BA	—	4J
BH	—	4J
BR	—	4H
CE220	—	4P
CK1005	—	5AQ
CK1006	—	4C
CK1007	—	Fig. 73
DR3B27	—	4P
DR123C	—	Fig. 15
DR200	—	2N
ECC81	—	9A
ECC82	—	9A
ECC83	—	9A
EF50	—	9C
F123A	—	Fig. 15
F127A	—	Fig. 15
G84	V19	4B
GL2C44	—	Fig. 9
GL5C24	—	Fig. 15
GL146	—	Fig. 56
GL152	—	Fig. 56
GL159	—	Fig. 56
GL169	—	Fig. 56
GL446A	—	Fig. 11
GL446B	—	Fig. 11
GL464A	—	Fig. 9
GL559	—	Fig. 10
GL6442	—	—
GL6463	—	9CZ
GL8012A	—	Fig. 54
HD203A	—	3N
HF60	—	2D
HF75	—	2D
HF100	—	2D
HF120	—	4F
HF140	—	4F
HF175	—	Fig. 46
HF200	—	2N
HF201A	—	Fig. 15
HF250	—	2N
HF300	—	2N
HK24	—	3G
HK54	—	2D
HK57	—	Fig. 33
HK154	—	2D
HK158	—	2D
HK252L	—	4BC
HK253	—	4AT
HK254	—	2N
HK257	—	7BM
HK257B	—	7BM
HK304L	—	4BC
HK354	—	2N
HK354C	—	2N
HK354D	—	2N
HK354E	—	2N
HK354F	—	2N
HK454H	—	2N
HK454L	—	2N
HK654	—	2N
HV12	—	3N
HV18	—	2N
HV27	—	3N
HY6J5GTX	—	6Q
HY6L5GTX	—	7AC
HY24	—	4D
HY25	—	3G
HY30Z	—	4BO
HY31Z	—	Fig. 60
HY40	—	3G
HY40Z	—	3G
HY51A	—	3G
HY51B	—	3G
HY51Z	—	4BO
HY57	—	3G
HY60	—	5AW
HY61	—	5AW
HY63	—	Fig. 72
HY65	—	Fig. 72
HY67	—	Fig. 65
HY69	—	Fig. 64
HY75	—	2T
HY75A	—	2T
HY114B	—	2T
HY615	—	Fig. 71
HY801A	—	4D
HY866Jr	—	4P
HY1231Z	—	Fig. 60
HY1269	—	Fig. 65
HYE1148	—	Fig. 71
KT66	—	7AC
KY21	—	—
NU2C35	—	Fig. 23
PE340	—	5BK
PL172/8295	V28	—
PL177A	V22	Fig. 14
PL6549	—	Fig. 14
PL6569	V21	Fig. 3
PL6580	V21	5BK
RK10	—	4D
RK11	—	3G
RK12	—	3G
RK15	—	4D
RK16	—	5A
RK17	—	5F
RK18	—	3G
RK19	—	4AT
RK20	—	Fig. 61
RK20A	—	Fig. 61
RK21	—	4P
RK22	—	Fig. 52
RK23	—	6BM
RK24	—	4D
RK25	—	6BM
RK25B	—	6BM
RK28	—	5J
RK28A	—	5J
RK30	—	2D
RK31	—	3G
RK32	—	2D
RK33	—	Fig. 69

Type	Page	Base
RK34	—	Fig. 70
RK35	—	2D
RK36	—	2D
RK37	—	2D
RK38	—	2D
RK39	—	5AW
RK41	—	5AW
RK42	—	4D
RK43	—	6C
RK44	—	6BM
RK46	—	Fig. 61
RK47	—	Fig. 64
RK48	—	Fig. 64
RK48A	—	Fig. 64
RK49	—	6A
RK51	—	3G
RK52	—	3G
RK56	—	5AW
RK57	—	3N
RK58	—	3N
RK59	—	Fig. 60
RK61	—	—
RK62	—	4D
RK63	—	2N
RK63A	—	2N
RK64	—	5AW
RK65	—	Fig. 48
RK66	—	Fig. 61
RK75	—	Fig. 61
RK100	—	Fig. 67
RK705A	—	Fig. 45
RK866	—	4P
T20	—	3G
T21	—	6A
T40	—	3G
T55	—	3G
T60	—	2D
T100	—	2D
T125	—	2N
T160L	V21	3G
T200	—	2N
T300	—	—
T814	—	3N
T822	—	3N
TB35	—	Fig. 30
TUF20	—	2T
TW75	—	2D
TW150	—	2N
TZ20	—	3G
TZ40	—	3G
UE100	—	2D
UE468	—	Fig. 32
UH35	—	3G
UH50	—	2D
UH51	—	2D
V70	—	3N
V70A	—	3N
V70B	—	3G
V70C	—	3G
V70D	—	3G
VR75	V19	4AJ
VR90	V19	4AJ
VR105	V19	4AJ
VR150	V19	4AJ
VT52	—	4D
VT127A	—	Fig. 53
WE304A	—	2D
X6030	—	Fig. 2
XXB	—	Fig. 6
XXD	—	8AC
XXL	—	5AC
XXFM	—	8BZ
ZB60	—	2D
ZB120	—	4E

SEMICONDUCTORS

Type	Page
1N21F	V24
1N34A	V24
1N35	V24
1N52A	V24
1N60	V24
1N64	V24
1N64A	V24
1N67A	V24
1N82A	V24
1N94	V24
1N270	V24
1N458A	V24
1N634	V24
1N645A	V24
1N914	V24
1N3754	V24
1N4001	V24
1N4002	V24
1N4004	V24
1N4719	V24
2N406	V24
2N441	V25
2N706A	V24
2N718A	V24
2N1179	V24
2N1302	V24
2N1306	V24
2N1491	V25
2N1970	V25
2N2102	V25
2N2157	V25
2N2222	V24
2N2270	V25
2N2631	V25
2N2869	V25
2N2876	V25
2N2925	V24
2N3053	V25
2N3055	V25
2N3119	V25

Type	Page
2N3391A	V24
2N3394	V24
2N3512	V25
2N3553	V25
2N3565	V24
2N3568	V24
2N3583	V25
2N3632	V25
2N3638	V24
2N3663	V24
2N3733	V25
2N3772	V25
2N3866	V24
2N3904	V24
2N3906	V24
2N3924	V25
2N3948	V25
2N4012	V25
2N4037	V25
2N4123	V24
2N4124	V24
2N4126	V24
2N4275	V24
2N4396	V25
2N4401	V24
2N4410	V24
2N4416	V25
2N4417	V25
2N4427	V25
2N4957	V24
2N4959	V24
2N5016	V25
2N5032	V24
2N5070	V25
2N5071	V25
2N5087	V24
2N5089	V24
2N5109	V24
2N5179	V24
2N5183	V24
2N5222	V24
2N5457	V26
2N5458	V26
2N5459	V26
2N5460	V25
2N5461	V26
2N5463	V26
2N5465	V26
2N5470	V25
2N5484	V26
2N5486	V26
2N5635	V25
2N5636	V25
2N5637	V25
2N5641	V25
2N5642	V25
2N5643	V25
2N5669	V26
2N5670	V26
2N5829	V24
2N5913	V25
2N5914	V25
2N5915	V25
2N5919	V25
2N5921	V25
2N5941	V25
2N5942	V25
2N5944	V25
2N5945	V25
2N5946	V25
2N5995	V25
2N5996	V25
2N6136	V25
2N40675	V25
3N128	V26
3N187	V26
3N200	V26
E300	V26
HEP51	V24
HEP53	V24
HEP56	V24
HEP801	V26
HEP802	V26
MBD101	V24
MJ480	V25
MMT3823	V26
MPF102	V26
MPF103	V26
MPF104	V26
MPF105	V26
MPF106	V26
MPF107	V26
MPF120	V26
MPF121	V26
MPF122	V26
MPS918	V24
MPS2926	V24
MPS3394	V24
MPS3563	V24
MPS3693	V24
MPS3694	V24
MPS3702	V24
MPS3706	V24
MPS6513	V24
MPS6514	V25
MPS6530	V25
MPS6534	V25
MPS6543	V25
MPS6569	V25
MPSA12	V25
MPSA55	V25
MPSU01	V25
MPSU51	V26
MSD7000	V24
TIS48	V25
TIS54	V26
TIXM10	V25
40231	V24
40235	V24
40600	V26
40601	V26
40602	V26
40603	V26
40604	V26
40673	V26

E.I.A. VACUUM-TUBE BASE DIAGRAMS

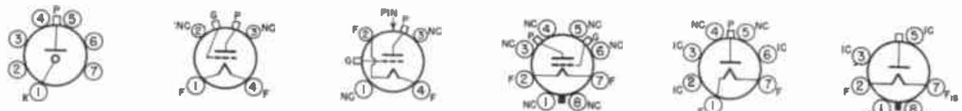
Socket connections correspond to the base designations given in the column headed "Base" in the classified tube-data tables. Bottom views are shown throughout. Terminal designations are as follows:

A = Anode	D = Deflecting Plate	IS = Internal Shield	RC = Ray-Control Eelectrode
B = Beam	F = Filament	K = Cathode	Ref = Reflector
BP = Bayonet Pin	FE = Focus Elect.	NC = No Connection	S = Shell
BS = Base Sleeve	G = Grid	P = Plate (Anode)	TA = Target
C = Ext. Coating	H = Heater	P_1 = Starter-Anode	U = Unit
CL = Collector	IC = Internal Con.	Pbr = Beam Plates	• = Gas-Type Tube

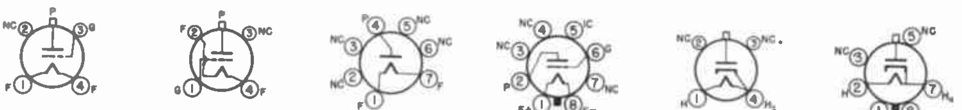
Alphabetical subscripts D, P, T and HX indicate, respectively, diode unit, pentode unit, triode unit or hexode unit in multi-unit types. Subscript CT indicates filament or heater tap.

Generally when the No. 1 pin of a metal-type tube in Table II, with the exception of all triodes, is shown connected to the shell, the No. 1 pin in the glass (G or GT) equivalent is connected to an internal shield.

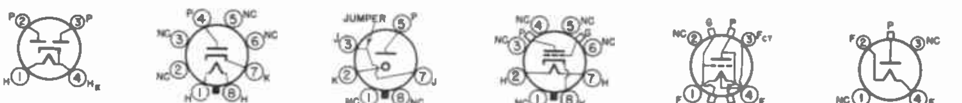
* On 12AQ, 12AS and 12CT: index = large lug; • = pin cut off



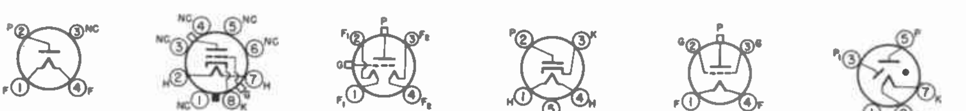
2AG 2D 2N 2T 2Z 3C



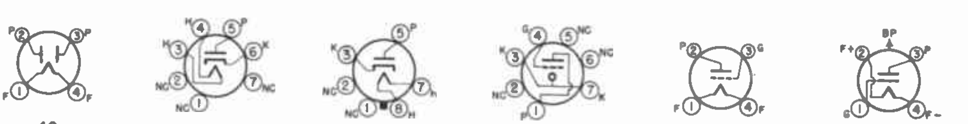
3G 3N 3T 4AA 4AB 4AC



4AD 4AH 4AJ 4AM 4AQ 4AT

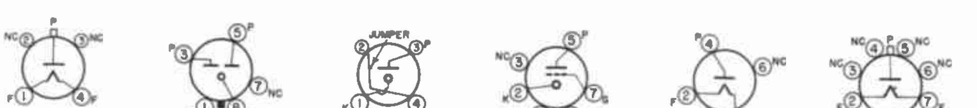


4B 4BB 4BC 4BJ 4BO 4BU

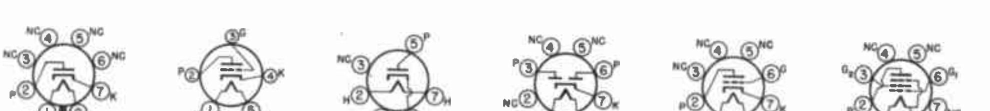


4C 4CB 4CG 4CK 4D 4E

4F 4G 4H 4J 4K 4M



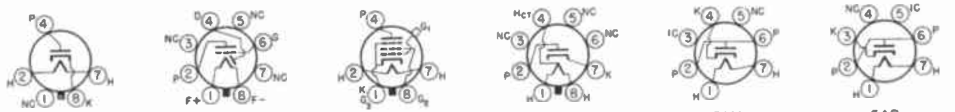
4P 4R 4S 4V 4X 4Y



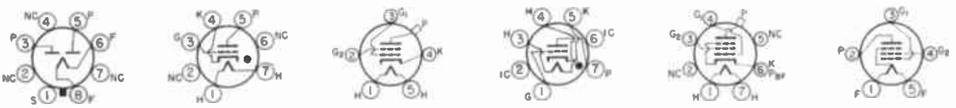
4Z 5A 5AA 5AB 5AC 5AD

TUBE BASE DIAGRAMS

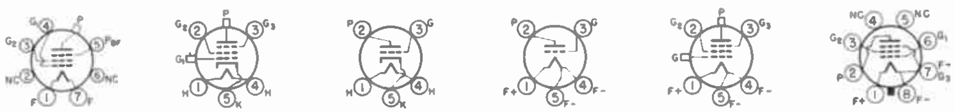
Bottom views are shown. Terminal designations on sockets are given on page V5.



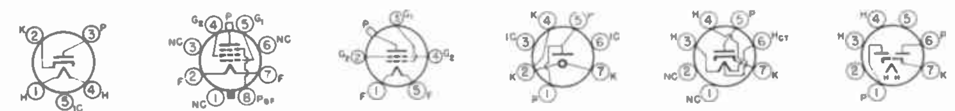
5AF 5AG 5AK 5AL 5AM 5AP



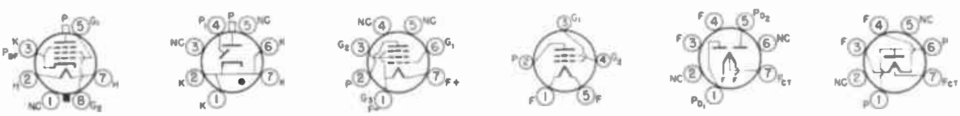
5AQ 5AS 5AW 5AY 5AZ 5B



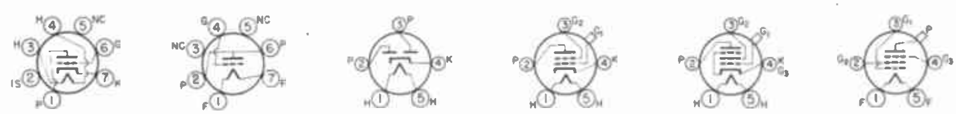
5BA 5BB 5BC 5BD 5BE 5BF



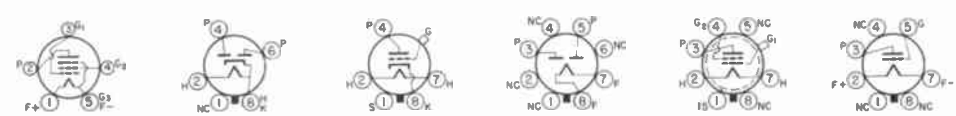
5BG 5BJ 5BK 5BO 5BQ 5BS



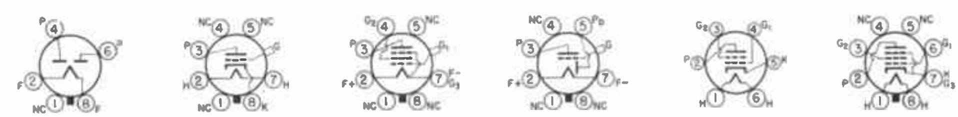
5BT 5BU 5BZ 5C 5CA 5CB



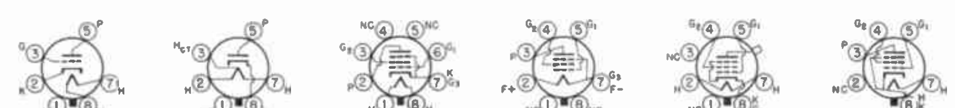
5CE 5CF 5D 5E 5F 5J



5K 5L 5M 5Q 5R 5S



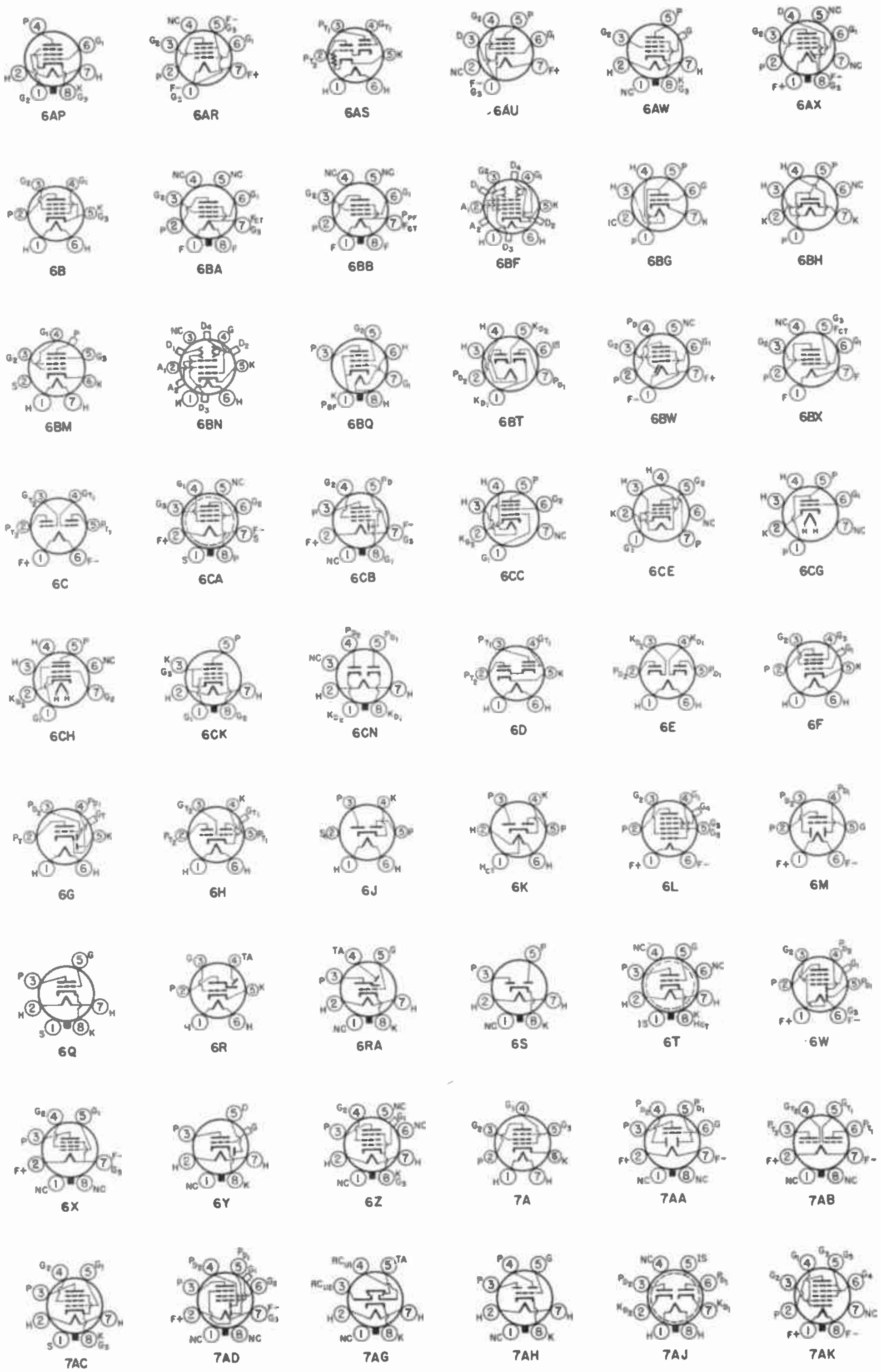
5T 5U 5Y 5Z 6A 6AA



6AB 6AD 6AE 6AF 6AM 6AO

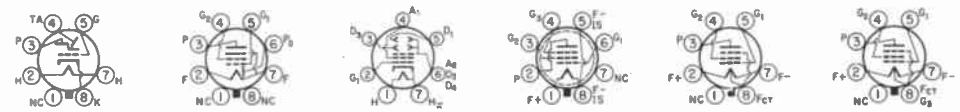
TUBE BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page V5.

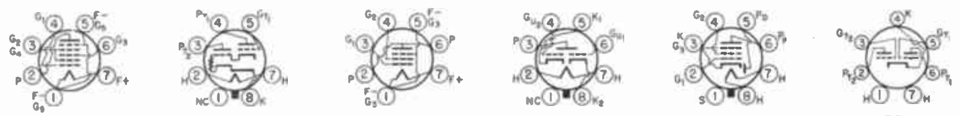


TUBE BASE DIAGRAMS

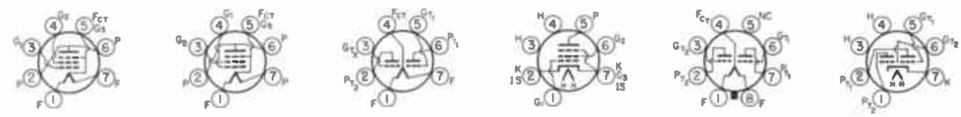
Bottom views are shown. Terminal designations on sockets are given on page V5.



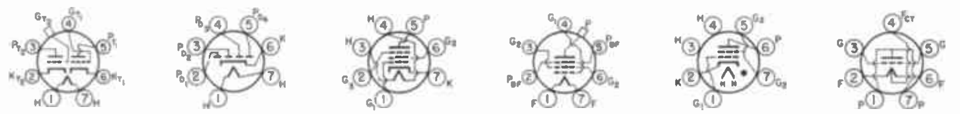
7AL 7AM 7AN 7AO 7AP 7AQ



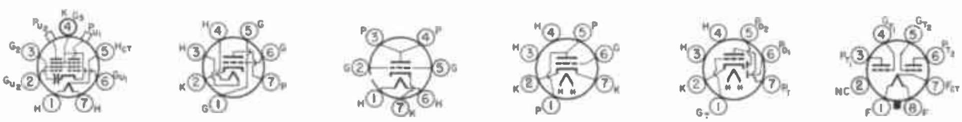
7AT 7AU 7AV 7AX 7AZ 7B



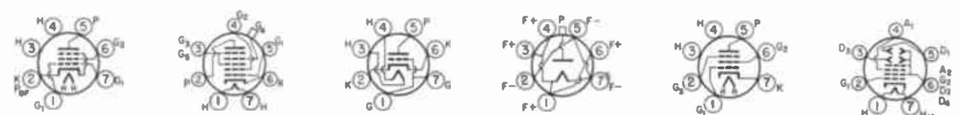
7BA 7BB 7BC 7BD 7BE 7BF



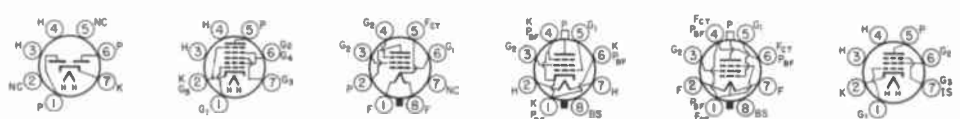
7BH 7BJ 7BK 7BM 7BN 7BO



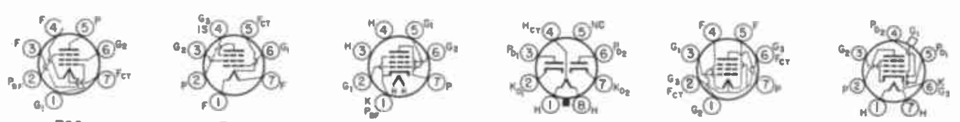
7BP 7BQ 7BR 7BS 7BT 7BW



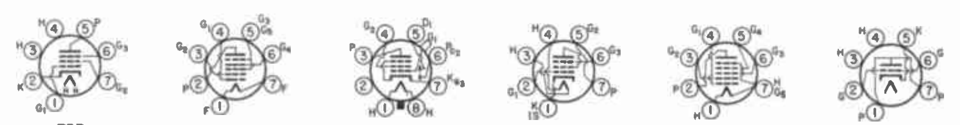
7BZ 7C 7CA 7CB 7CC 7CE



7CF 7CH 7CJ 7CK 7CL 7CM



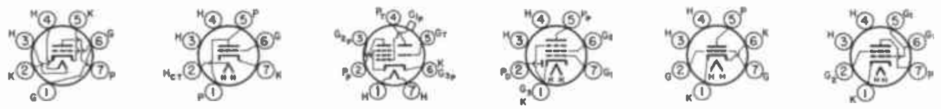
7CQ 7CU 7CV 7CX 7CY 7D



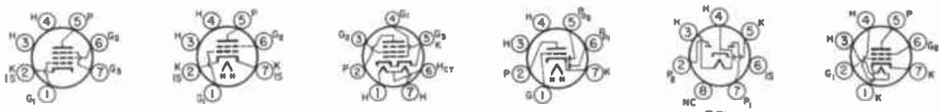
7DB 7DC 7DE 7DF 7DH 7DK

TUBE BASE DIAGRAMS

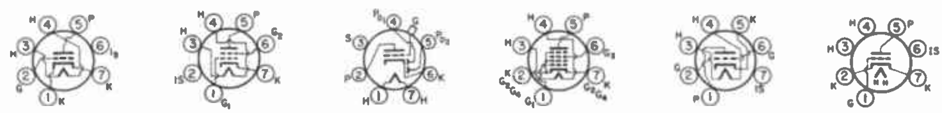
Bottom views are shown. Terminal designations on sockets are given on page V5.



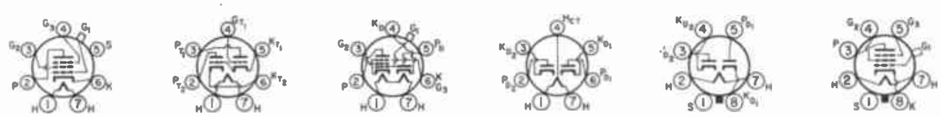
7DT | 7DW | 7E | 7EA | 7EG | 7EK



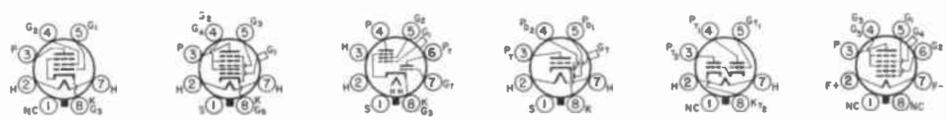
7EN | 7EW | 7F | 7FB | 7FL | 7FN



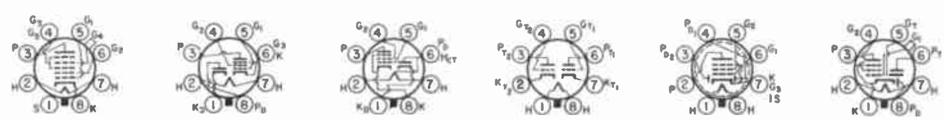
7FP | 7FQ | 7G | 7GA | 7GK | 7GM



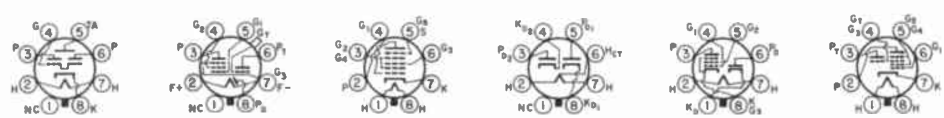
7H | 7J | 7K | 7L | 7Q | 7R



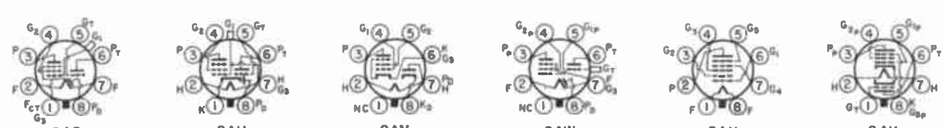
7S | 7T | 7U | 7V | 7W | 7Z



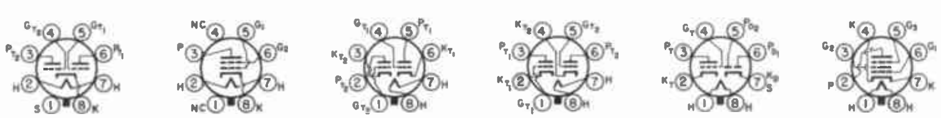
8A | 8AA | 8AB | 8AC | 8AE | 8AF



8AG | 8AJ | 8AL | 8AN | 8AO | 8AR



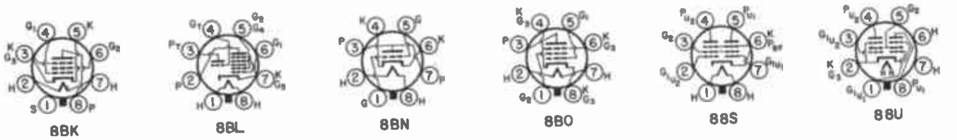
8AS | 8AU | 8AV | 8AW | 8AX | 8AY



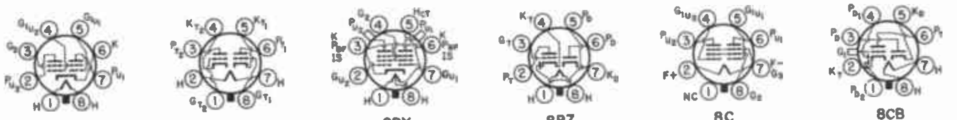
8B | 8BA | 8BD | 8BE | 8BF | 8BJ

TUBE BASE DIAGRAMS

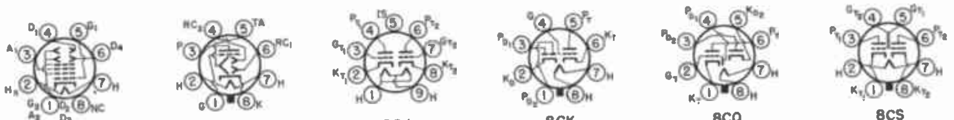
Bottom views are shown. Terminal designations on sockets are given on page V5.



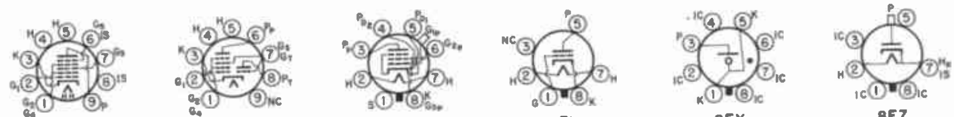
8BK 8BL 8BN 8BO 8BS 8BU



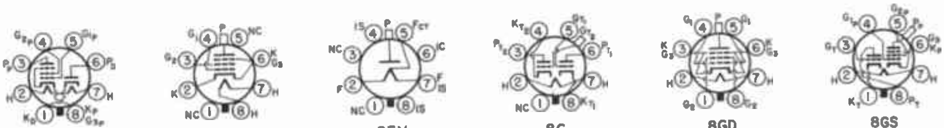
8BV 8BW 8BY 8BZ 8C 8CB



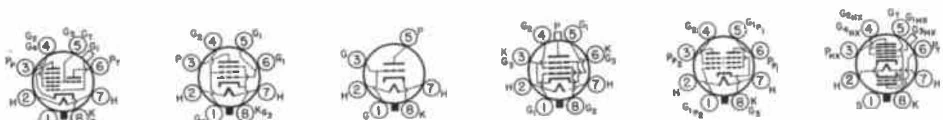
8CD 8CH 8CJ 8CK 8CQ 8CS



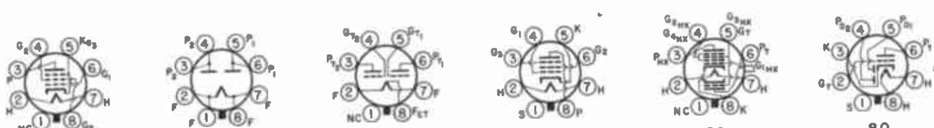
8CT 8DU 8E 8EL 8EX 8EZ



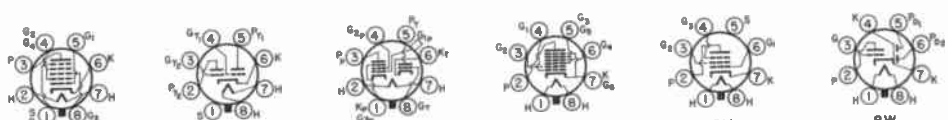
8F 8FP 8FV 8G 8GD 8GS



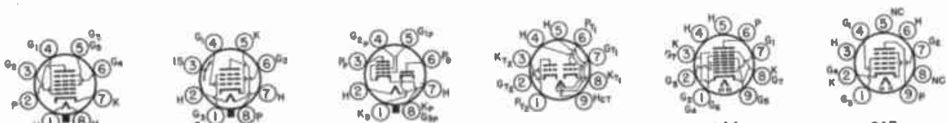
8H 8HY 8J8 8JC 8JP 8K



8KQ 8KS 8L 8N 8O 8Q



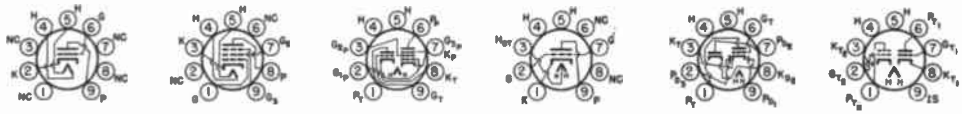
8R 8S 8T 8U 8V 8W



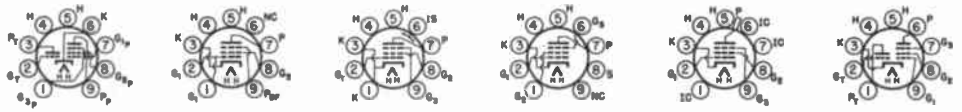
8X 8Y 8Z 9A 9AA 9AB

TUBE BASE DIAGRAMS

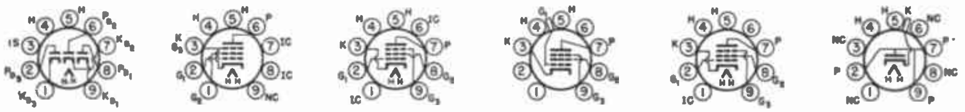
Bottom views are shown. Terminal designations on sockets are given on page V5.



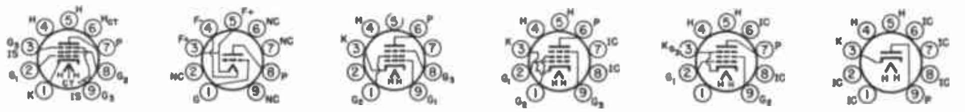
9AC 9AD 9AE 9AG 9AH 9AJ



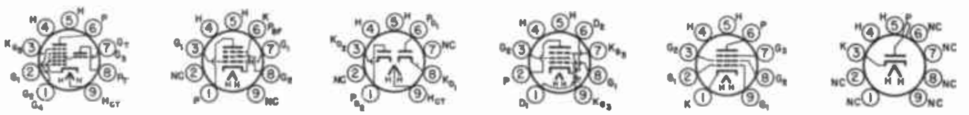
9AK 9AM 9AQ 9AR 9AS 9AT



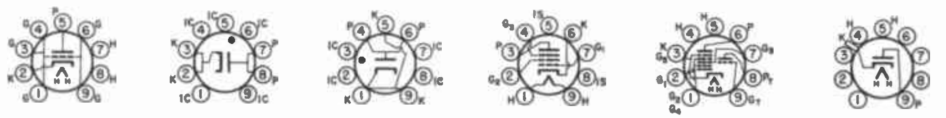
9AX 9AZ 9BA 9BB 9BC 9BD



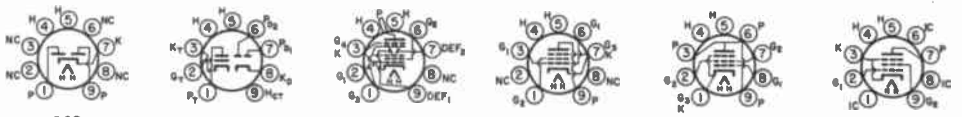
9BF 9BG 9BJ 9BK 9BL 9BM



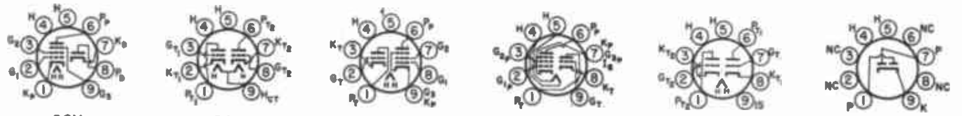
9BP 9BQ 9BS 9BU 9BV 9BW



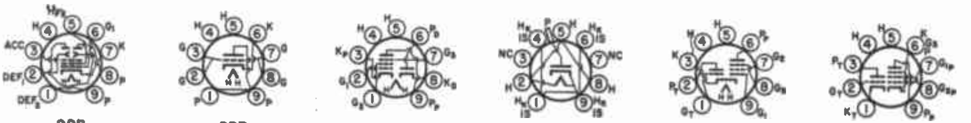
9BX 9BY 9BZ 9C 9CA 9CB



9CD 9CF 9CG 9CK 9CT 9CV



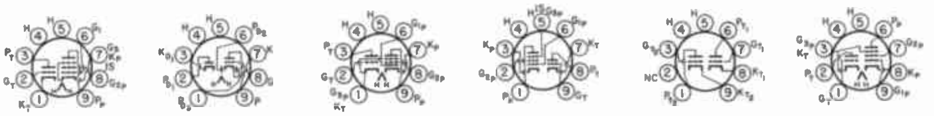
9CY 9CZ 9DA 9DC 9DE 9DJ



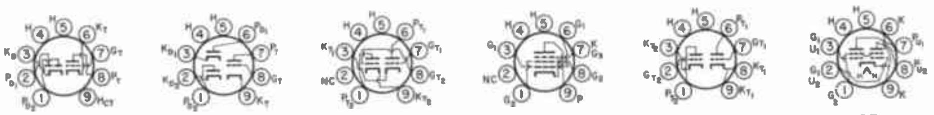
9DP 9DR 9DS 9DT 9DW 9DX

TUBE BASE DIAGRAMS

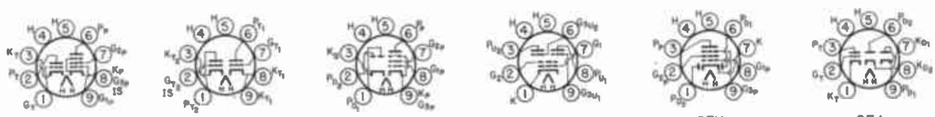
Bottom views are shown. Terminal designations on sockets are given on page V5.



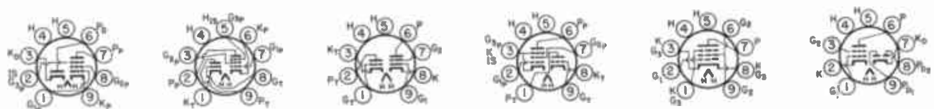
9DZ 9E 9EC 9ED 9EF 9FG



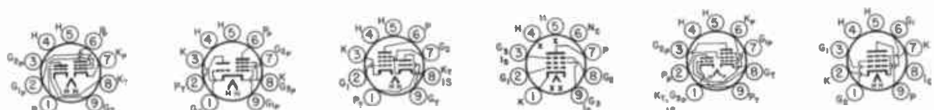
9EN 9ER 9ES 9EU 9EW 9F



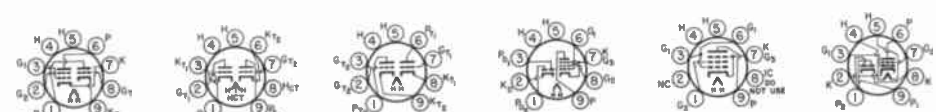
9FA 9FC 9FE 9FG 9FH 9FJ



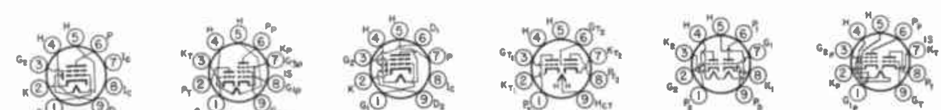
9FN 9FT 9FX 9FZ 9G 9GC



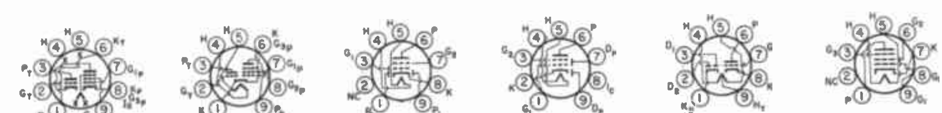
9GE 9GF 9GJ 9GK 9GM 9GR



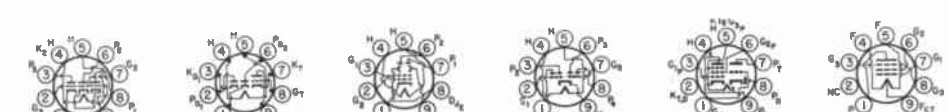
9GS 9H 9HF 9HK 9HN 9HR



9HV 9HX 9HZ 9J 9JC 9JD



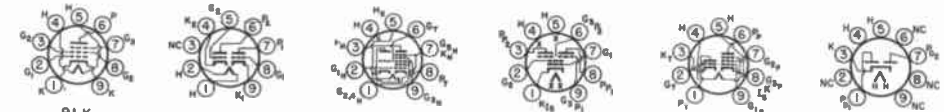
9JF 9JG 9JU 9JX 9JY 9K



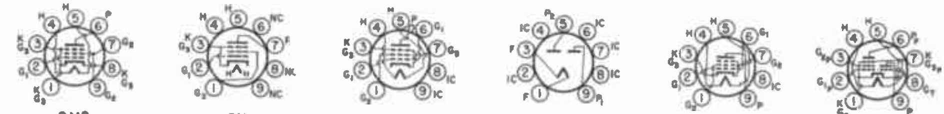
9KA 9KR 9KS 9KT 9KU 9L

TUBE BASE DIAGRAMS

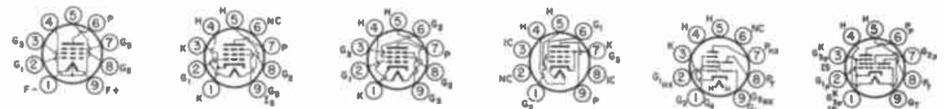
Bottom views are shown. Terminal designations on sockets and * meaning are given on page V5.



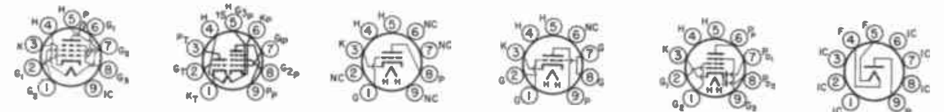
9LK 9LS 9KV 9LW 9LY 9M



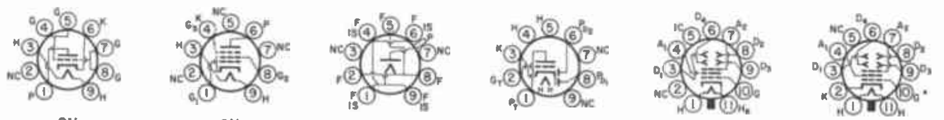
9MS 9N 9NM 9NT 9NZ 9PA



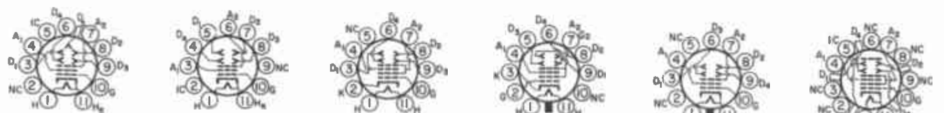
9PB 9PM 9PU 9PX 9Q 9QA



9QL 9QP 9R 9S 9T 9U

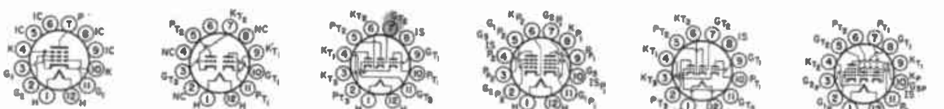


9V 9X 9Y 9Z 11A 11B

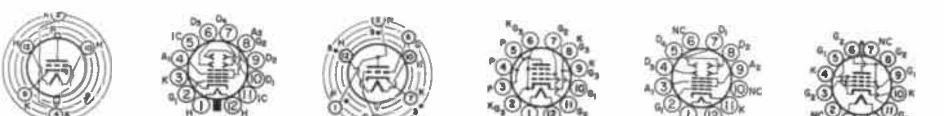


11C 11J 11L 11M 11N 11S

11T 11V 12A 12AQ * 12AS * 12BF



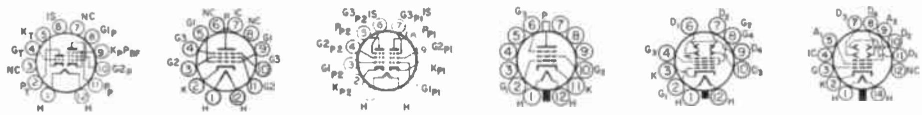
12BJ 12BM 12BQ 12BW 12BY 12CA



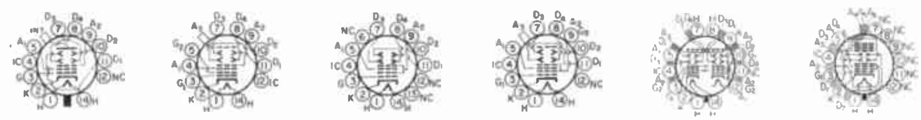
12CT * 12E 12EA * 12EU 12F 12FB

TUBE BASE DIAGRAMS

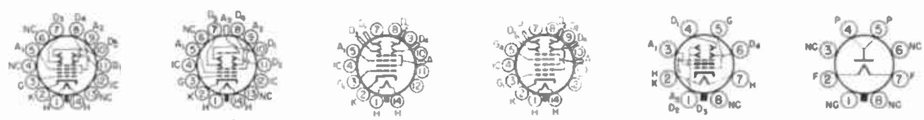
Bottom views are shown. Terminal designations on sockets are given on page V5.



12FM 12GW 12HD 12J 12T 14A



14B 14E 14G 14J 14K 14P



14R 14S 14U 14V FIG. 1 FIG. 2

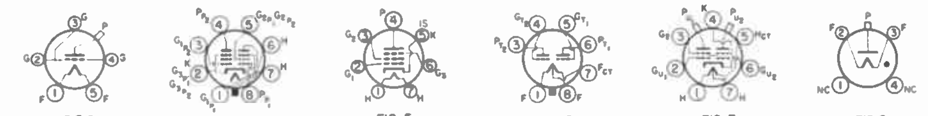


FIG. 3 FIG. 4 FIG. 5 FIG. 6 FIG. 7 FIG. 8

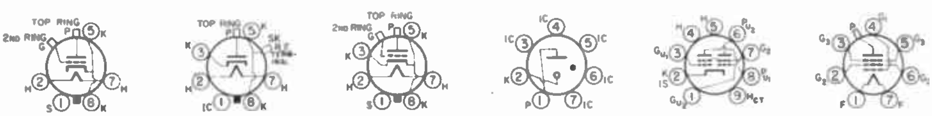


FIG. 9 FIG. 10 FIG. 11 FIG. 12 FIG. 13 FIG. 14

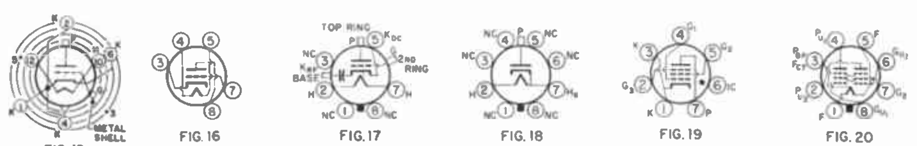


FIG. 15 FIG. 16 FIG. 17 FIG. 18 FIG. 19 FIG. 20

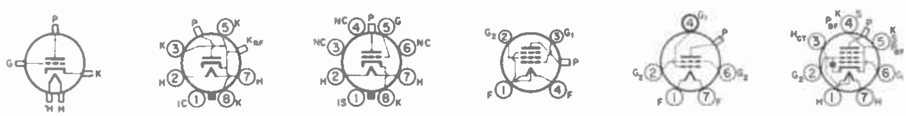


FIG. 21 FIG. 22 FIG. 23 FIG. 24 FIG. 25 FIG. 26

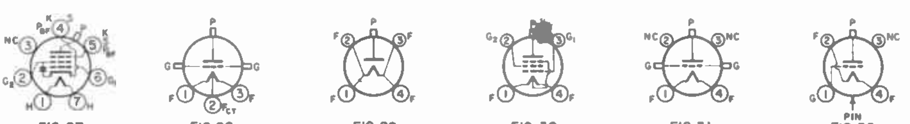


FIG. 27 FIG. 28 FIG. 29 FIG. 30 FIG. 31 FIG. 32

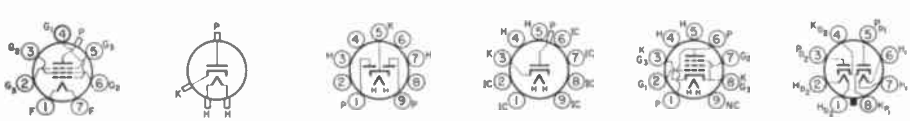


FIG. 33 FIG. 34 FIG. 35 FIG. 36 FIG. 37 FIG. 38

TUBE BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page V5.

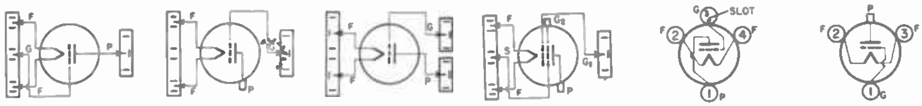


FIG. 39 FIG. 40 FIG. 41 FIG. 42 FIG. 43 FIG. 44

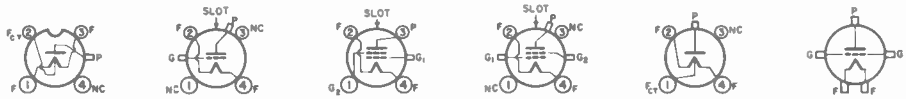


FIG. 45 FIG. 46 FIG. 47 FIG. 48 FIG. 49 FIG. 50

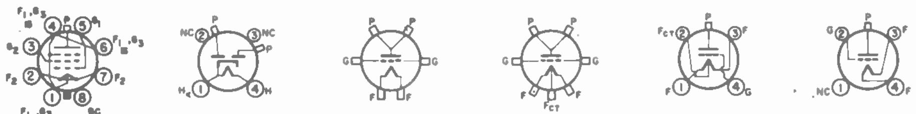


FIG. 51 FIG. 52 FIG. 53 FIG. 54 FIG. 55 FIG. 56

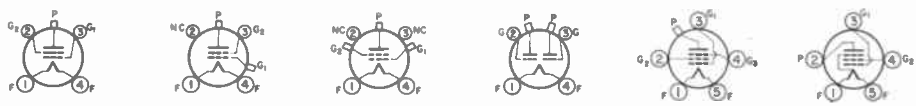


FIG. 57 FIG. 58 FIG. 59 FIG. 60 FIG. 61 FIG. 62

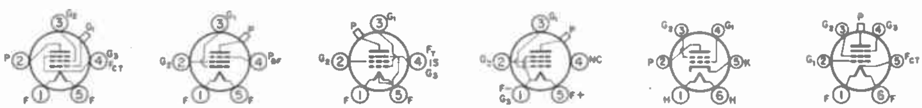


FIG. 63 FIG. 64 FIG. 65 FIG. 66 FIG. 67 FIG. 68

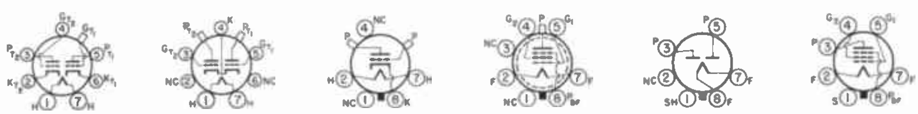


FIG. 69 FIG. 70 FIG. 71 FIG. 72 FIG. 73 FIG. 74

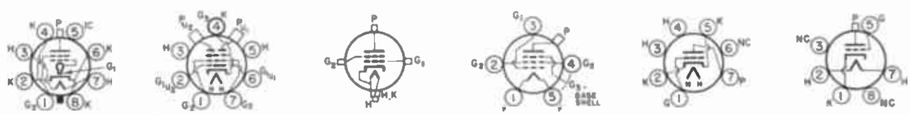


FIG. 75 FIG. 76 FIG. 77 FIG. 78 FIG. 79 FIG. 80

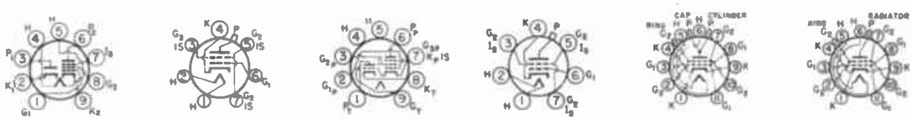


FIG. 81 FIG. 82 FIG. 83 FIG. 84 FIG. 85 FIG. 86

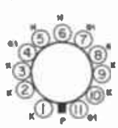


FIG. 87

Type	Name	Base	Fil. or Heater		Capacitances pF			Plate Supply V	Grid Bias	Screen Volts	Screen mA	Plate mA	Plate Res. Ohms	Transconductance[11]	Amp. Factor[4]	Load Res. Ohms	Watts Output
			V	Amp.	C_{in}	C_{out}	C_{gp}										
6AF4A	Uhf — Triode A1 Amp.	7DK	6.3	0.225	2.2	0.45	1.9	80	150*	—	—	16	2.27K	6600	15	—	—
	Osc. 950 MHz							100	10KΩ	—	0.4[9]	22	—	—	—	—	—
6AG5	Sharp Cut-off Pent.	7BD	6.3	0.3	6.5	1.8	0.03	250	180*	150	2.0	6.5	800K	5000	—	—	—
								100	180*	100	1.4	4.5	600K	4500	—	—	—
6AH6	Sharp Cut-off Pent. Pent. Amp.	7BK	6.3	0.45	10.0	2.0	0.03	300	160*	150	2.5	10	500K	9600	—	—	—
	Triode Amp.							150	160*	—	—	12.5	3.6K	11K	40	—	—
6AJ4	Uhf Triode	9BX	6.3	0.225	4.4	0.18	2.4	125	68*	—	—	16	4.2K	10K	42	—	—
6AK5	Sharp Cut-off Pent.	7BD	6.3	0.175	4.0	2.8	0.02	180	200*	120	2.4	7.7	690K	5100	—	—	—
								150	330*	140	2.2	7	420K	4300	—	—	—
								120	200*	120	2.5	7.5	340K	5000	—	—	—
6AK6	Pwr. Amp. Pent.	7BK	6.3	0.15	3.6	4.2	0.12	180	−9	180	2.5	15	200K	2300	—	10K	1.1
6AL5	Dual Diode[10]	6BT	6.3	0.3	—	—	—	Max. rms voltage −117. Max. dc output current −9 mA.[1]									
6AM4	Uhf Triode	9BX	6.3	0.225	4.4	0.16	2.4	150	100*	—	—	7.5	10K	9000	90	—	—
6AN5	Beam Pwr. Pent.	7BD	6.3	0.45	9.0	4.8	0.075	120	120*	120	12.0	35	12.5K	8000	—	2.5K	1.3
6AN8A‡	Medium-μ Triode	9DA	6.3	0.45	2.0	2.7	1.5	200	−6	—	—	13	5.75K	3300	—	—	—
	Sharp Cut-off Pent.				7.0	2.3	0.04	200	180*	150	2.8	9.5	30K	6200	—	—	—
6AQ5A‡	Beam Pwr. Pent.	7BZ	6.3	0.45	8.3	8.2	0.35	180	−8.5	180	3/4	30[2]	58K	3700	29[5]	5.5K	2.0
								250	−12.5	250	4.5/7	47[2]	52K	4100	45[5]	5K	4.5
6AQ6	Dual Diode — High-μ Triode	7BT	6.3	0.15	1.7	1.5	1.8	100	−1	—	—	0.8	61K	1150	70	—	—
								250	−3	—	—	1	58K	1200	70	—	—
6AR5	Pwr. Amp. Pent.	6CC	8.3	0.4	—	—	—	250	−16.5	250	5.7/10	35[2]	65K	2400	34[5]	7K	3.2
								250	−18	250	5.5/10	33[2]	68K	2300	32[5]	7.6K	3.4
6AS6	Sharp Cut-off Pent.	7CM	6.3	0.175	4	3	0.2	120	−2	120	3.5	5.2	110K	3200	—	—	—
6AT6	Duplex Diode — High-μ Triode	7BT	6.3	0.3	2.3	1.1	2.1	250	−3	—	—	1	58K	1200	70	—	—
6AU6A‡	Sharp Cut-off Pent.	7BK	6.3	0.3	5.5	5	0.0035	250	68*	150	4.3	10.6	1 meg.	5200	—	—	—
6AV6	Dual Diode — High-μ Triode	7BT	6.3	0.3	2.2	0.8	2.0	250	−2	—	—	1.2	62.5K	1600	100	—	—
6AZ8	Medium-μ Triode	9ED	6.3	0.45	2	1.7	1.7	200	−6	—	—	13	5.75K	3300	19	—	—
	Semiremote Cut-off Pent.				6.5	2.2	0.02	200	180*	150	3	9.5	300K	6000	—	—	—
6BA6	Remote Cut-off Pent.	7BK	6.3	0.3	5.5	5	0.0035	250	68*	100	4.2	11	1 meg.	4400	—	—	—
6BA7	Pentagrid Conv.	8CT	6.3	0.3	Osc. 20kΩ			250	−1	100	10	3.8	1 meg.	950	—	—	—
6BC4	Uhf Medium-μ Triode	9DR	6.3	0.225	2.9	0.26	1.6	150	100*	—	—	14.5	4.8K	10K	48	—	—
6BE6	Pentagrid Conv.	7CH	6.3	0.3	Osc. 20kΩ			250	−1.5	100	6.8	2.9	1 meg.	475	—	—	—
6BE8A‡	Medium-μ Triode	9EG	6.3	0.45	2.8	1.5	1.8	150	56*	—	—	18	5K	8500	40	—	—
	Sharp Cut-off Pent.				4.4	2.6	0.04	250	68*	110	3.5	10	400K	5200	—	—	—
6BF5	Beam Pwr. Amp.	7BZ	6.3	1.2	14	6	0.65	110	−7.5	110	4/10.5	39[2]	12K	7500	36[5]	2.5K	1.9
6BF6	Dual Diode — Medium-μ Triode	7BT	6.3	0.3	1.8	0.8	2	250	−9	—	—	9.5	8.5K	1900	16	10K	0.3
6BH6	Sharp Cut-off Pent.	7CM	6.3	0.15	5.4	4.4	0.0035	250	−1	150	2.9	7.4	1.4 meg.	4600	—	—	—
6BH8‡	Medium-μ Triode	9DX	6.3	0.6	2.6	0.38	2.4	150	−5	—	—	9.5	5.15K	3300	17	—	—
	Sharp Cut-off Pent.				7	2.4	0.046	200	82*	125	3.4	15	150K	7000	—	—	—
6BJ6A	Remote Cut-off Pent.	7CM	6.3	0.15	4.5	5.5	0.0035	250	−1	100	3.3	9.2	1.3 meg.	3800	—	—	—
6BJ7	Triple Diode	9AX	6.3	0.45	Max. peak inverse plate voltage = 330 V. Max. dc plate current each diode = 1.0 mA												
6BJ8‡	Dual Diode — Medium-μ Triode	9ER	6.3	0.6	2.8	0.38	2.6	250	−9	—	—	8	7.15K	2800	20	—	—
6BK6	Dual Diode — High-μ Triode	7BT	6.3	0.3	—	—	—	250	−2	—	—	1.2	62.5K	1600	100	—	—
6BK7B	Medium-μ Dual Triode[10]	9AJ	6.3	0.4	3	1	1.8	150	56*	—	—	18	4.6K	9300	43	—	—
6BL8	Triode	9DC	6.3	0.43	2.5	1.8	1.5	250	−1.3	—	—	14	—	5000	20	—	—
	Pentode				5.2	3.4	0.025	250	−1.3	175	2.8	10	400K	6200	47	—	—
6BN4A	Medium-μ Triode	7EG	6.3	0.2	3.2	1.4	1.2	150	220*	—	—	9	6.3K	6800	43	—	—
6BN6	Gated-Beam Pent.	7DF	6.3	0.3	4.2	3.3	0.004	80	−1.3	60	5	0.23	—	—	—	68K	—
6BN8‡	Dual Diode — High-μ Triode	9ER	6.3	0.6	3.6	0.25	2.5	250	−3	—	—	1.6	28K	2500	70	—	—
6BQ5	Pwr. Amp. Pent.	9CV	6.3	0.76	10.8	6.5	0.5	300	−7.3	200	10.8	49.5[2]	38K	—	—	5.2K	17[3]
6BQ7A	Medium-μ Dual Triode[10]	9AJ	6.3	0.4	2.85	1.35	1.15	150	220*	—	—	9	6.1K	6400	39	—	—
6BR8A‡	Medium-μ Triode	9FA	6.3	0.45	2.5	0.4	1.8	150	56*	—	—	18	5K	8500	40	—	—
	Sharp Cut-off Pent.				5	2.6	0.015	250	68*	110	3.5	10	400K	5200	—	—	—
6BS8	Low-Noise Dual Triode[10]	9AJ	6.3	0.4	2.6	1.35	1.15	150	220*	—	—	10	5K	7200	36	—	—
6BX8	Dual Triode[10]	9AJ	6.3	0.4	—	—	1.4	65	−1	—	—	9	—	6700	25	—	—
6BZ6	Semiremote Cut-off Pent.	7CM	6.3	0.3	7.5	1.8	0.02	200	180*	150	2.6	11	600K	6100	—	—	—
6BZ7	Medium-μ Dual Triode[10]	9AJ	6.3	0.4	2.5	1.35	1.15	150	220*	—	—	10	5.6K	6800	38	—	—
6BZ8	Dual Triode[10]	9AJ	6.3	0.4	—	—	—	125	100*	—	—	10[1]	5.6K	8000	45	—	—
6C4	Medium-μ Triode	6BG	6.3	0.15	1.8	1.3	1.6	250	−8.5	—	—	10.5	7.7K	2200	17	—	—
6CB6A‡	Sharp Cut-off Pent.	7CM	6.3	0.3	6.5	1.9	0.02	200	180*	150	2.8	9.5	600K	6200	—	—	—
6CE5‡	Rf Pent.	7BD	6.3	0.3	6.5	1.9	0.03	200	180*	150	2.8	9.5	600K	6200	—	—	—
6CG6	Semiremote Cut-off Pent.	7BK	6.3	0.3	5	5	0.008	250	−8	150	2.3	9	720K	2000	—	—	—
6CG7‡	Medium-μ Dual Triode[10]	9AJ	6.3	0.6	2.3	2.2	4	250	−8	—	—	9	7.7K	2600	20	—	—
6CL6	Pwr. Amp. Pent.	9BV	6.3	0.65	11	5.5	0.12	250	−3	150	7/7.2	31[2]	150K	11K	30[5]	7500	2.8
6CW4	Triode	12AQ	6.3	0.13	4.1	1.7	0.92	70	0	—	—	8	5.44K	12.5K	68	—	—
6CX8	Medium-μ Triode	9DX	6.3	0.75	2.2	0.38	4.4	150	150*	—	—	9.2	8.7K	4600	40	—	—
	Sharp Cut-off Pent.				9	4.4	0.06	200	68*	125	5.2	24	70K	10K	—	—	—
6CY5	Sharp Cut-off Tetrode	7EW	6.3	0.2	4.5	3	0.03	125	−7	80	1.5	10	100K	8000	—	—	—
6DJ8	Twin Triode	9AJ	6.3	0.365	3.3	1.8	1.4	90	−1.3	—	—	15	—	12.5K	33	—	—
6DK6	Sharp Cut-off Pent.	7CM	6.3	0.3	6.3	1.9	0.02	300	−6.5	150	3.8	12	—	9800	—	—	—
6DS4	High-μ Triode	12AQ	6.3	0.135	4.1	1.7	.92	70	0	—	—	8	5.44K	12.5K	68	—	—
6DT6	Sharp Cut-off Pent.	7EN	6.3	0.3	5.8	—	0.02	150	560*	100	2.1	1.1	150K	615	—	—	—
6DW5	Beam Pwr. Amp.	9CK	6.3	1.2	14	9	0.5	200	−22.5	150	2	55	15K	5500	—	—	—
6EA8‡	Triode	9AE	6.3	0.45	3	0.3	1.7	330	−12	—	—	18	5K	8500	40	—	—
	Sharp Cut-off Pent.				5	2.6	0.02	330	−9	330	4	12	80K	6400	—	—	—
6EB8	High-μ Triode	9DX	6.3	0.75	2.4	.36	4.4	330	−5	—	—	2	37K	2700	100	—	—
	Sharp Cut-off Pent.				11	4.2	0.1	330	−9	—	7	25	75K	12.5K	—	—	—
6EH5	Power Pentode	7CV	6.3	1.2	17	9	0.65	135	0	117	14.5	42	11K	14.6K	—	3K	1.4
6EH7	Remote Cut-off Pent.	9AQ	6.3	0.3	9	3	.005	200	−2	90	4.5	12	500K	12.5K	—	—	—
6EH8	Triode	9JG	6.3	0.45	2.8	1.7	1.8	125	−1	—	—	13.5	—	7500	40	—	—
	Pentagrid Conv.				4.8	2.4	0.02	125	−1	125	4	12	170K	6000	—	—	—
6EJ7	Sharp Cut-off Pent.	9AQ	6.3	0.3	10	3	.005	200	−2.5	200	4.7	10	350K	15K	—	—	—
6ER5	Tetrode	7FN	6.3	0.18	4.4	3.0	0.38	200	−1.2	0	0	10	8K	10.5K	80	—	—

Type	Name	Base	Fil. or Heater V	Fil. or Heater Amp.	Capacitances pF C_{in}	Capacitances pF C_{out}	Capacitances pF C_{gp}	Plate Supply V	Grid Bias	Screen Volts	Screen mA	Plate mA	Plate Res. Ohms	Transconductance[11]	Amp. Factor[4]	Load Res. Ohms	Watts Output
6ES8	Dual Triode	9DE	6.3	0.365	3.4	1.7	1.9	130	−1.2	—	—	15	—	12.5K	34	—	—
6EU7	Twin Triode	9LS	6.3	0.3	1.6	0.2	1.5	100	−1	—	—	0.5	80K	1250	100	—	—
6EU8	Triode	9JF	6.3	0.45	5.0	2.6	0.02	150	—	—	—	18	5K	8500	40	—	—
	Pentode				3.0	1.6	1.7	125	−1	125	4	12	80K	6400	—	—	—
6EZ8	Triple Triode — Triode No. 1	9KA	6.3	0.45	2.6	1.4	1.5	330	−4	—	—	4.2	13.6K	4200	57	—	—
	Triodes No. 2 & 3					1.2											
6FV6	Sharp Cut-off Tetrode	7FQ	6.3	0.2	4.5	3	0.03	125	−1	80	1.5	10	100K	8000	—	—	—
6GC5	Pwr. Pent.	9EU	6.3	1.2	18.0	7.0	0.9	110	−7.5	110	4	50	13K	8000	—	2K	2.1
6GJ8	Triode	9AE	6.3	0.6	3.4	1.6	2.6	125	−1	—	—	13.5	5K	8500	40	—	—
	Pentode				8	2.4	0.36	125	−1	125	4.5	12	150K	7500	—	—	—
6GK5	High-μ Triode	7FP	6.3	0.18	5	3.5	0.52	135	−1	—	—	11.5	5400	15K	78	—	—
6GK6	Power Pentode	9GK	6.3	0.76	10	7.0	0.14	250	−7.3	250	5.5	48	38K	11.3K	—	5.2K	5.7
6GM6	Pentode	7CM	6.3	0.4	10	2.4	0.036	125	—	125	3.4	14	200K	13K	—	—	—
6GN8	High-μ Triode	9DX	6.3	0.75	2.4	0.36	4.4	250	−2	—	—	2	37K	2700	100	—	—
	Sharp Cut-off Pent.				11	4.2	0.1	200	—	150	5.5	25	60K	11.5K	—	—	—
6HB6	Power Pentode	9PU	6.3	0.76	13	8.0	0.18	250	100*	250	6.2	40	24K	20K	—	—	—
6J4	Grounded-Grid Triode	7BQ	6.3	0.4	7.5	3.9	0.12	150	100*	—	—	15	4.5K	12K	55	—	—
6J6A‡	Medium-μ Dual Triode — A_1 Amp.[10]	7BF	6.3	0.45	2.2	0.4	1.6	100	50*	—	—	8.5	7.1K	5300	38	—	—
	Mixer							150	810*	—	—	4.8	10.2K	1900	Dsc. peak voltage = 3 V		
6KD8	Sharp Cut-off Pent.	9AE	6.3	0.4	5.0	2.6	0.015	125	−1	110	3.5	9.5	200K	5000	—	—	—
	Medium-μ Triode				1.5	2.8	1.8	125	−1	—	—	13.5	—	7500	40	—	—
6KE8	Medium-μ Triode	9DC	6.3	0.4	2.4	2.0	1.3	125	68*	—	—	13	5.0K	8000	40	—	—
	Sharp Cut-off Pent.				5.0	3.4	.015	125	33*	125	2.8	10	125K	12K	—	—	—
6KR8	Sharp Cut-off Pent.	9DX	6.3	0.75	13	4.4	0.075	200	82*	100	3.0	19.5	60K	20K	—	—	—
	Medium-μ Triode				4.2	3.0	2.6	125	68*	—	—	15	4400	10.4K	46	—	—
6KT6	Remote Cut-off Pent.	9PM	6.3	0.3	9.5	3	0.19	125	56*	125	4.2	17	—	18K	—	—	—
6KT8	High-μ Triode	9QP	6.3	0.6	32	1.6	3.0	250	−2	—	—	1.8	31.5K	3200	100	—	—
	Sharp Cut-off Pent.				7.5	2.2	0.046	125	−1	125	4.5	12	150K	10K	—	—	—
6KZ8	Sharp Cut-off Pent.	9FZ	6.3	0.45	5.5	3.4	0.01	125	−1	125	4	12	200K	7500	—	—	—
	Medium-μ Triode				3.2	1.8	1.6	125	−1	—	—	13.5	5400	8500	46	—	—
6LJ8	Sharp Cut-off Pent.	9GF	6.3	0.4	5.5	3.4	0.015	125	33*	125	3.5	12	125K	13K	—	—	—
	Medium-μ Triode				2.4	2.0	1.4	125	68*	—	—	13	5K	8000	40	—	—
6LY8	High-μ Triode	9DX	6.3	0.75	2.6	2.8	3.8	200	−2.0	—	—	1.0	59K	1700	100	—	—
	Sharp Cut-off Pent.				13.0	4.4	0.75	200	—	100	3	19.5	60K	20K	—	—	—
6MU8	Medium-μ Triode	9AE	6.3	0.6	3	2.2	2.2	330	0	—	—	11.5	5.8K	6000	35	—	—
	Sharp Cut-off Pent.							330	0	150	4.2	19	165K	9000	—	—	—
6T4	Uhf Triode	7DK	6.3	0.225	2.6	0.25	1.7	80	150*	—	—	18	1.86K	7000	13	—	—
6T8A‡	Triple Diode-High-μ Triode	9E	6.3	0.45	1.6	1	2.2	100	−1	—	—	0.8	54K	1300	70	—	—
								250	−3	—	—	1	58K	1200	70	—	—
6U8A‡	Medium-μ Triode	9AE	6.3	0.45	2.5	0.4	1.8	150	56*	—	—	18	5K	8500	40	—	—
	Sharp Cut-off Pent.				5	2.6	0.01	250	68*	110	3.5	10	400K	5200	—	—	—
6X8A‡	Medium-μ Triode	9AK	6.3	0.45	2.0	0.5	1.4	100	100*	—	—	8.5	6.9K	—	40	—	—
	Sharp Cut-off Pent.				4.3	0.7	0.09	250	200*	150	1.6	7.7	750K	—	—	—	—
12AB5	Beam Pwr. Amp. — A_1 Amp.	9EU	12.6	0.2	8	8.5	0.7	250	−12.5	250	4.5/7	47[2]	50K	4100	45[5]	5K	4.5
	AB_1 Amp.[3]							250	−15	250	5/13	79[2]	60K[1]	3750	70[5]	10K[6]	10
12AQ5	Beam Pwr. Amp. — A_1 Amp.	7BZ	12.6	0.225	8.3	8.2	0.35	250	−12.5	250	4.5/7	47[2]	52K	4100	45[5]	5K	4.5
	AB_1 Amp.[3]							250	−15	250	5/13	79[2]	60K[1]	3750[1]	70[5]	10K[6]	10
12AT7	High-μ Dual Triode[10]	9A	12.6	0.15	2.2[7]	0.5[7]	1.5[7]	100	270*	—	—	3.7	15K	4000	60	—	—
			6.3	0.3	2.2[8]	0.4[8]	1.5[8]	250	200*	—	—	10	10.9K	5500	60	—	—
12AU7A	Medium-μ Dual Triode[10]	9A	12.6	0.15	1.6[7]	0.5[7]	1.5[7]	100	0	—	—	11.8	6.25K	3100	19.5	—	—
			6.3	0.3	1.6[8]	0.35[8]	1.5[8]	250	−8.5	—	—	10.5	7.7K	2200	17	—	—
12AV7	Medium-μ Dual Triode[10]	9A	12.6	0.225	3.1[7]	0.5[7]	1.9[7]	100	120*	—	—	9	6.1K	6100	37	—	—
			6.3	0.45	3.1[8]	0.4[8]	1.9[8]	150	56*	—	—	18	4.8K	8500	41	—	—
12AX7A	High-μ Dual Triode — A_1 Amp.[10]	9A	12.6	0.15	1.6[7]	0.46[7]	1.7[7]	250	−2	—	—	1.2	62.5K	1600	100	—	—
	Class B		6.3	0.3	1.6[8]	0.34[8]	1.7[8]	300	0	—	—	40[2]	—	—	14[5]	16K[6]	7.5
12AY7	Medium-μ Dual Triode[10] — A_1 Amp.	9A	12.6	0.15	1.3	0.6	1.3	250	−4	—	—	3	—	1750	40	—	—
	Low-Level Amp.		6.3	0.3				150	2700*	Plate resistor = 20K. Grid resistor = 0.1 meg. V. G. = 12.5							
12AZ7A‡	High-μ Dual Triode[10]	9A	12.6	0.225	3.1[7]	0.5[7]	1.9[7]	100	270*	—	—	3.7	15K	4000	60	—	—
			6.3	0.45	3.1[8]	0.4[8]	1.9[8]	250	200*	—	—	10	10.9K	5500	60	—	—
12BH7A‡	Medium-μ Dual Triode[10]	9A	12.6	0.3	3.2[7]	0.5[7]	2.6[7]	250	−10.5	—	—	11.5	5.3K	3100	16.5	—	—
			6.3	0.6	3.2[8]	0.4[8]	2.6[8]										
12BY7A‡	Sharp Cut-off Pent.	9BF	12.6	0.3	11.1	3	0.055	250	68*	150	6	25	90K	12K	1200	—	—
			6.3	0.6													
35B5	Beam Pwr. Amp.	7BZ	35	0.15	11	6.5	0.4	110	−7.5	110	3/7	41[2]	—	5800	40[5]	2.5K	1.5
50B5	Beam Pwr. Amp.	7BZ	50	0.15	13	6.5	0.5	110	−7.5	110	4/8.5	50[2]	14K	7500	49[5]	2.5K	1.9
50FK5	Pwr. Pent.	7CV	50	0.1	17	9	0.65	110	62*	115	12	32	14K	12.8K	—	3K	1.2
5686	Beam Pwr. Pent.	9G	6.3	0.35	6.4	8.5	0.11	250	−12.5	250	3[5]	27[5]	45K	3100	—	9K	2.7
5687	Medium-μ Dual Triode[10]	9H	12.6	0.45	4[7]	0.5[7]	4[7]	120	−2	—	—	36	1.7K	11K	18.5	—	—
			6.3	0.9	4[8]	0.5[8]	4[8]	250	−12.5	—	—	12.5	3K	5500	16.5	—	—
5722	Noise Generating Diode	5CB	6.3	1.5	—	2.2	—	200	—	—	—	35	—	—	—	—	—
4A2 / 15874	High-μ Triode	9V	6.3	0.3	9.0	1.8	0.55	150	62*	—	—	26	1.8K	24K	43	—	—
5879	Sharp Cut-off Pent.	9AD	6.3	0.15	2.7	2.4	0.15	250	−3	100	0.4	1.8	2 meg.	1000	—	—	—
6386	Medium-μ Dual Triode[10]	8CJ	6.3	0.35	2	1.1	1.2	100	200*	—	—	9.6	4.25K	4000	17	—	—
6887	Dual Diode	6BT	6.3	0.2	Max. peak inverse plate voltage = 360 V. Max. dc plate current each diode = 10 mA.												
6973	Pwr. Pentode	9EU	6.3	0.45	6	6	0.4	440	−15	300	—	—	73K	4800	—	—	—
7189A	Pwr. Pentode	9CV	6.3	0.76	10.8	6.5	0.5	250	−7.3	250	5.5	48	40K	11.3K	—	—	—
7258	Sharp Cut-off	9DA	12.6	0.195	7	2.4	0.4	330	—	125	3.8	12	170K	7800	—	—	—
	Medium-μ Triode				2	0.26	1.5	330	−3	—	—	15	4.7K	4500	21	—	—
7586	Medium-μ Triode	12AQ	6.3	0.135	4.2	1.6	2.2	75	100*	—	—	10.5	3000	11.5K	35	—	—
7587	Sharp Cut-off Tet.	12AS	6.3	0.15	6.5	1.4	0.01	125	68*	50	2.7	10	200K	10.5K	—	—	—
7895	High-μ Triode	12AQ	6.3	0.135	4.2	1.7	0.9	110	0	—	—	7	6800	9400	64	—	—
8056	Medium-μ Triode	12AQ	6.3	0.135	4.0	1.7	2.1	12	0	—	—	5.8	1.6K	8000	12.5	—	—

TABLE I—MINIATURE RECEIVING TUBES—*Continued*

Type	Name	Base	Fil. or Heater V	Fil. or Heater Amp.	Capacitances pF C_{in}	Capacitances pF C_{out}	Capacitances pF C_{gp}	Plate Supply V	Grid Bias	Screen Volts	Screen mA	Plate mA	Plate Res. Ohms	Transconductance[11]	Amp. Factor[4]	Load Res. Ohms	Watts Output
8058	High-μ Triode	12CT	6.3	0.135	6.0	0.046	1.3	110	47*	—	—	10	—	10K	—	—	—
8393	Medium-μ Triode	12AQ	13.5	0.060	4.4	1.7	2.4	75	100*	—	—	10.5	3000	11.5K	35	—	—
8628	High-μ Triode	12AQ	6.3	0.10	10	3.4	1.7	150	3.3K*	—	—	0.3	41K	3100	127	7K	—
8677	Power Triode	12CT	6.3	0.15	6.0	1.2	—	180	1.2K*	—	—	20	3K	5400	70	—	1.4
9001	Sharp Cut-off Pent.	7BD	6.3	0.15	3.6	3	0.01	250	−3	100	0.7	2	1 meg.	1400	—	—	—
9002	Uhf Triode	7BS	6.3	0.15	1.2	1.1	1.4	250	−7	—	—	6.3	11.4K	2200	25	—	—
9003	Remote Cut-off Pent.	7BD	6.3	0.15	3.4	3	0.1	250	−3	100	2.7	6.7	700K	1800	—	—	—
9006	Uhf Diode	6BH	6.3	0.15	Max. ac voltage = 270. Max. dc output current = 5 mA.												

‡ Controlled heater warm-up characteristic.
Ω Oscillator gridleak or screen-dropping resistor ohms.
* Cathode resistor ohms.
** Space-charge grid.
[1] Per Plate.
[2] Maximum-signal current for full-power output.
[3] Values are for two tubes in push-pull.
[4] Unless otherwise noted.
[5] No signal plate mA.
[6] Effective plate-to-plate.
[7] Triode No. 1.
[8] Triode No. 2.
[9] Oscillator grid current mA.
[10] Values for each section.
[11] Micromhos.
[12] Through 33K.

TABLE II—METAL RECEIVING TUBES

Characteristics given in this table apply to all tubes having type numbers shown, including metal tubes, glass tubes with "G" suffix, and bantam tubes with "GT" suffix. For "G" and "GT"-tubes not listed (not having metal counterparts), see Tables III and V.

Type	Name	Base	Fil. or Heater V	Fil. or Heater Amp.	Capacitances pF C_{in}	Capacitances pF C_{out}	Capacitances pF C_{gp}	Plate Supply V	Grid Bias	Screen Volts	Screen mA	Plate mA	Plate Res. Ohms	Transconductance[12]	Amp. Factor[13]	Load Res. Ohms	Watts Output
6A8	Pentagrid Conv.	8A	6.3	0.3	—	—	—	250	−3	100	2.7	3.5	360K	550	—	—	—
								E_{bb} (Osc.) 250 V through 20K. Grid resistor (Osc.) 50K. I_b = 4 mA. I_{g1} = 0.4 mA.									
6AC7 1852	Sharp Cut-off Pent.	8N	6.3	0.45	11	5	0.15	300	160*	150	2.5	10	1 meg.	9000	—	—	—
								300	160*	60K[8]	2.5	10	1 meg.	3000	—	—	—
6AG7	Pwr. Amp. Pent.	8Y	6.3	0.65	13	7.5	0.06	300	3	150	7/9	30/31	130K	11K	—	10K	3
6B8	Dual-Diode — Pent.	8E	6.3	0.3	6	9	0.005	250	−3	125	2.3	10	600K	1325	—	—	—
6F6	Pwr. Amp. Pent. — A_1 Amp.[1,5]	7S	6.3	0.7	6.5	13	0.2	250	20	20[10]	—	31/34	2.6K	2600	6.8	4K	0.85
	AB_2 Amp.[1,6]							350	730*	132[11]	—	50/60	—	—	—	10K[7]	9
								350	−38	123[11]	—	48/92	—	—	—	6K[7]	13
	A_1 Amp.[5]							250	−16.5	250	6/11	34/36	80K	2500	—	7K	3.2
								285	−20	285	7/13	38/40	78K	2500	—	7K	4.8
	AB_2 Amp.[6]							375	26	250	5/20	34/82	—	—	82[11]	10K[7]	18.5
								375	340*	250	8/18	54/77	—	—	94[11]	10K[7]	19
6J5	Medium-μ Triode	6Q	6.3	0.3	3.4	3.6	3.4	250	−8	—	—	9	7.7K	2600	20	—	—
6J7	Sharp Cut-off Pent. — A_1 Amp.	7R	6.3	0.3	7	12	0.005	250	−3	100	0.5	2	1 meg.	1225	—	—	—
	Biased Detector							250	10K*	100	Zero signal cathode current = 0.43 mA.					0.5 meg.	
6K7	Variable-μ Pent. — R.f. Amp.	7R	6.3	0.3	7	12	0.005	250	−3	125	2.6	10.5	600K	1650	990	—	—
	Mixer							250	−10	100	Osc. peak volts = 7						
6K8	Triode — Hexode Conv. — Hexode	8K	6.3	0.3	—	—	—	250	−3	100	6	2.5	600K	350	—	—	—
	Triode							100	50K[8]	—	—	3.8	I_{g1} (Osc.) = 0.15 mA.				
6L6-GB[2]	Beam Pwr. Amp. — A_1 Amp.[1,5]	7AC	6.3	0.9	11.5	9.5	0.9	250	−20	20[10]	—	40/44	1.7K	4700	8	5K	1.4
	A_1 Amp.[5] Self Bias							250	167*	250	5.4/7.2	75/78	—	—	14[10]	2.5K	6.5
								300	218*	200	3/4.6	51/55	—	—	12.7[10]	4.5K	6.5
	A_1 Amp.[5] Fixed Bias							250	−14	250	5/7.3	72/79	22.5K	6000	14[10]	2.5K	6.5
								350	−18	250	2.5/7	54/66	33K	5200	18[10]	4.2K	10.8
	A_1 Amp.[6] Self Bias							250	125*	250	10/15	120/130	—	—	35.6[11]	5K[7]	13.8
								270	125*	270	11/17	134/145	—	—	28.2[11]	5K[7]	18.5
	A_1 Amp.[6] Fixed Bias							250	−16	250	10/16	120/140	24.5[5]	5500[5]	32[11]	5K[7]	14.5
								270	−17.5	270	11/17	134/155	23.5[5]	5700[5]	35[11]	5K[7]	17.5
	AB_1 Amp.[6] Self Bias							360	270*	270	5/17	88/100	—	—	40.6[11]	9K[7]	24.5
	AB_1 Amp.[6] Fixed Bias							360	−22.5	270	5/11	88/140	—	—	45[11]	3.8K[7]	18
								360	−22.5	270	5/15	88/132	—	—	45[11]	6.6K[7]	26.5
	AB_2 Amp.[6] Fixed Bias							360	−18	225	3.5/11	78/142	—	—	52[11]	6K[7]	31
								360	−22.5	270	5/16	88/205	—	—	72[11]	3.8K[7]	47
6L7	Pentagrid — Mixer Amp. — A_1 Amp.	7T	6.3	0.3	—	—	—	250	−3	100	6.5	5.3	600K	1100	−3[14]	—	—
	Mixer							250	−6	150	9.2	3.3	1 meg.	350	−15[14]	—	—
6N7GT	Class-B Twin Triode — B Amp.[9]	8B	6.3	0.8	—	—	—	300	0	—	—	35/70	—	—	82[11]	8K[7]	10
	A_1 Amp.[15]							250	−5	—	—	6	11.3K	3100	—	—	—
6Q7	Dual Diode — High-μ Triode	7V[2]	6.3	0.3	5	3.8	1.4	250	−3	—	—	1	58K	1200	70	—	—
6R7	Dual Diode — Triode	7V[2]	6.3	0.3	4.8	3.8	2.4	250	−9	—	—	9.5	8.5K	1900	16	10K	0.28
6SA7GT	Pentagrid Conv.	8R[2]	6.3	0.3	9.5	12	0.13	250	0[3]	100	8	3.4	800K	Grid No. 1 resistor 20K.			
6SB7Y	Pentagrid Conv.	8R	6.3	0.3	9.6	9.2	0.13	100	−1	100	10.2	3.6	50K	900	—	—	—
								250	−1	100	10	3.8	1 meg.	950	—	—	—
								250	22K[8]	12K[8]	12/13	6.8/6.5	Osc. Section in 88—108 MHz. Service.				
6SC7	High-μ Dual Triode[5]	8S	6.3	0.3	2	3	2	250	−2	—	—	2	53K	1325	70	—	—
6SF7	Diode — Variable-μ Pent.	7AZ	6.3	0.3	5.5	6	0.004	250	−1	100	3.3	12.4	700K	2050	—	—	—
6SG7	Hf Amp. Pent.	8BK	6.3	0.3	8.5	7	0.003	250	−2.5	150	3.4	9.2	1 meg.	4000	—	—	—
6SH7	Hf Amp. Pent.	8BK	6.3	0.3	8.5	7	0.003	250	−1	150	4.1	10.8	900K	4900	—	—	—
6SJ7[4]	Sharp Cut-off Pent.	8N	6.3	0.3	6	7	0.005	250	−3	100	0.8	3	1 meg.	1650	—	—	—
6SK7	Variable-μ Pent.	8N	6.3	0.3	6	7	0.003	250	−3	100	2.6	9.2	800K	2000	—	—	—
6SQ7GT	Dual Diode — High-μ Triode	8Q	6.3	0.3	3.2	3	1.6	250	−2	—	—	0.9	91K	1100	100	—	—
6SR7	Dual Diode — Triode	8Q	6.3	0.3	3.6	2.8	2.4	250	−9	—	—	9.5	8.5K	1900	16	—	—
6V6GTA	Beam Pwr. Amp. — A_1 Amp.[5]	7AC	6.3	0.45	10	11	0.3	180	−8.5	180	3/4	29/30	50K	3700	8.5[10]	5.5K	2
								250	−12.5	250	4.5/7	45/47	50K	4100	12.5[10]	5K	4.5
								315	−13	225	2.2/6	34/35	80K	3750	13[10]	8.5K	5.5
	AB_1 Amp.[6]							250	−15	250	5/13	70/79	60K	3750	30[11]	10K[7]	10
								285	−19	285	4/13.5	70/92	70K	3600	38[11]	8K[7]	14
1620	Sharp Cut-off Pent.	7R	6.3	0.3	7	12	0.005	250	−3	100	0.5	2	1 meg.	1225	—	—	—
5693	Sharp Cut-off Pent.	8N	6.3	0.3	5.3	6.2	0.005	250	−3	100	0.85	3	1 meg.	1650	—	—	—

* Cathode resistor-ohms.
[1] Screen tied to plate.
[2] No connection to Pin No. 1 for 6L6G, 6Q7G, 6RGT/G, 6S7G, 6SA7GT/G and 6SF5-GT.
[3] Grid bias = 2 volts if separate oscillator excitation is used.
[4] Also type 6SJ7Y.
[5] Values are for single tube or section.
[6] Values are for two tubes in push-pull.
[7] Plate-to-plate value.
[8] Osc. grid leak — Scrn. res.
[9] Values for two units.
[10] Peak af grid voltage.
[11] Peak af G-G voltage.
[12] Micromhos.
[13] Unless otherwise noted.
[14] G_3 voltage.
[15] Units connected in parallel.

TABLE III—6.3-VOLT GLASS TUBES WITH OCTAL BASES

(For "G" and "GT"-type tubes not listed here, see equivalent type in Tables II and V; characteristics and connections will be similar)

Type	Name	Plate Dissi-pation (Watts)	Base	Fil. or Heater V	Fil. or Heater Amp.	Capacitances pF C_{in}	Capacitances pF C_{out}	Capacitances pF C_{gp}	Plate Supply V	Grid Bias	Screen Volts	Screen mA	Plate mA	Plate Res. Ohms	Transcon-ductance[4]	Amp. Factor	Load Res. Ohms	Watts Output
6AL7GT	Electron-Ray Indicator	—	8CH	6.3	0.15	—	—	—	Outer edge of any of the three illuminated areas displaced 1/16 in. min. outward with +5 volts to its electrode. Similar inward disp. with −5 volts. No pattern with −6 volts grid.									
6AQ7GT	Dual Diode — High-μ Triode	—	8CK	6.3	0.3	2.8	3.2	3	250	−2	—	—	2.3	44K	1600	70	—	—
6AR6	Beam Pent.	—	6BQ	6.3	1.2	11	7	0.55	250	−22.5	250	5	77	21K	5400	—	—	—
6AR7GT	Dual Diode — Remote Pent.	—	7DE	6.3	0.3	5.5	7.5	0.003	250	−2	100	1.8	7	1.2 meg.	2500	—	—	—
6AS7GA	Low-μ Twin Triode — DC Amp.[1]	—	8BD	6.3	2.5	6.5	2.2	7.5	135	250*	—	—	125	0.28K	7000	2	—	—
6AU5GT	Beam Pwr. Amp.[3]	10	6CK	6.3	1.25	11.3	7	0.5	115	−20	175	6.8	60	6K	5600	—	—	—
6BL7GTA	Medium-μ Dual Triode[1]	—	8BD	6.3	1.5	4.4	0.9	6	250	−9	—	—	40	2.15K	7000	15	—	—
6BQ6GTB 6CU6	Beam Pwr. Amp.[3]	11	6AM	6.3	1.2	15	7	0.6	250	−22.5	150	2.1	57	14.5K	5900	—	—	—
6BX7GT	Dual Triode[1]	—	8BD	6.3	1.5	5	3.4	4.2	250	390*	—	—	42	1.3K	7600	10	—	—
6CB5A	Beam Pwr. Amp.[3]	26	8GD	6.3	2.5	22	10	0.4	175	−30	175	6	90	5K	8800	—	—	—
6CD6GA	Beam Pwr. Amp.[3]	20	5BT	6.3	2.5	24	9.5	0.8	175	−30	175	5.5	75	7.2K	7700	—	—	—
6CK4	Low-μ Triode	—	8JB	6.3	1.25	8	1.8	6.5	550	−26	—	—	55	1.0K	6500	6.7	—	—
6CL5	Beam Pwr. Amp.[3]	25	8GD	6.3	2.5	20	11.5	0.7	175	−40	175	7	90	6K	6500	—	—	—
6DN6	Beam Pwr. Pent.[3]	15	5BT	6.3	2.5	22	11.5	0.8	125	−18	125	6.3	70	4K	9000	—	—	—
6DN7	Dissimilar Dual Triode	—	8BD	6.3	0.9	2.2	0.7	4	350	−8	—	—	8	9K	2500	22	—	—
						4.6	1	5.5	550	−9.5	—	—	68	2K	7700	15	—	—
6DQ5	Beam Pwr. Amp.[3]	24	8JC	6.3	2.5	23	11	0.5	175	−25	125	5	110	5.5K	10.5K	—	—	—
6DQ6B	Beam Pwr. Amp.[3]	18	6AM	6.3	1.2	15	7	0.55	250	−22.5	150	2.4	75	20K	6600	—	—	—
6DZ7	Twin Pwr. Pent.[1]	13.2	8JP	6.3	1.52	11	5	0.6	300	120*	250	15	80	—	—	—	9K[2]	12
6E5	Electron Ray — Triode	—	6R	6.3	0.3	—	—	—	250	—	—	—	—	—	—	—	—	—
6EA7	Dissimilar — Dual Triode	—	8BD	6.3	1.05	2.2	0.6	4	350	−3	—	—	1.5	34K	1900	65	—	—
						6	1.3	8	550	−25	—	—	95	770	6500	5	—	—
6EF6	Beam Pwr. Amp.[5]	—	7S	6.3	0.9	11.5	9	0.8	250	−18	250	2	50	—	5000	—	—	—
6EY6	Beam Pwr. Pent.	—	7AC	6.3	0.68	8.5	7	0.7	350	−17.5	300	3	44	60K	4400	—	—	—
6EZ5	Beam Pwr. Pent.	—	7AC	6.3	0.8	9	7	0.6	350	−20	300	3.5	43	50K	4100	—	—	—
6FH6	Beam Pwr. Pent.	—	6AM	6.3	1.2	33	8	0.4	770	−22.5	220	1.7	75	12K	6000	—	—	—
6GW6	Beam Power Amp.[3]	17.5	6AM	6.3	1.2	17	7	0.5	250	−22.5	150	2.1	70	15K	7100	—	—	—
6K6GT	Pwr. Amp. Pent.	—	7S	6.3	0.4	5.5	6	0.5	315	−21	250	4/9	25/28	110K	2100	—	9K	4.5
6SL7GT	High-μ Dual Triode[1]	—	8BD	6.3	0.3	3.4	3.8	2.8	250	−2	—	—	2.3	44K	1600	70	—	—
6SN7GTB	Medium-μ Dual Triode[1]	—	8BD	6.3	0.6	3	1.2	4	250	−8	—	—	9	7.7K	2600	20	—	—
6W6GT	Beam Pwr. Amp.	—	7S	6.3	1.2	15	9	0.5	200	180*	125	2/8.5	46/47	28K	8000	—	4K	3.8
6Y6GA	Beam Pwr. Amp.	—	7S	6.3	1.25	15	1	0.7	200	−14	135	2.2/9	61/66	18.3K	7100	—	2.6K	6
1635	High-μ Dual Triode	—	8B	6.3	0.6	—	—	—	300	0	—	—	6.6/54	—	—	—	12K[2]	10.4
6550	Power Pentode	35	7S	6.3	1.6	14	12	0.85	400	−16.5	225	18	105	27K	9000	—	3K	20
7027A	Beam Pwr. Amp.	—	8HY	6.3	0.9	10	7.5	1.5	450	−30	350	19.2	194	—	6000	—	6K[2]	50
7591	Beam Pwr. Amp.	19	8KQ	6.3	0.8	10	5	0.25	450	200*	400	22	94	—	—	—	9K[2]	28

* Cathode resistor-ohms.
[1] Per section.
[2] Plate-to-plate value.
[3] Horz. Deflection Amp.
[4] Micromhos. A
[5] Vert. Deflection mp.

TABLE IV—CONTROL AND REGULATOR TUBES

Type	Name	Base	Cathode	Fil. or Heater Volts	Fil. or Heater Amp.	Peak Anode Voltage	Max. Anode mA	Minimum Supply Voltage	Oper-ating Voltage	Oper-ating mA	Grid Resistor	Tube Voltage Drop
0A2 6073	Voltage Regulator	5BO	Cold	—	—	—	—	185	150	5–30	—	—
0A3A/VR75	Voltage Regulator	4AJ	Cold	—	—	—	—	105	75	5–40	—	—
0B2 6074	Voltage Regulator	5BO	Cold	—	—	—	—	133	108	5–30	—	—
0B3/VR90	Voltage Regulator	4AJ	Cold	—	—	—	—	125	90	5–40	—	—
0C2	Voltage Regulator	5BO	Cold	—	—	—	—	105	75	5–30	—	—
0C3A/VR105	Voltage Regulator	4AJ	Cold	—	—	—	—	135	105	5–40	—	—
0D3A/VR150	Voltage Regulator	4AJ	Cold	—	—	—	—	185	150	5–40	—	—
5651	Voltage Regulator	5BO	Cold	—	—	115	—	115	87	1.5–3.5	—	—
5662	Thyratron — Fuse	Fig. 79	Htr.	6.3	1.5	200[1]	I_k to fuse — 150 Amp., 60 cycle, half-wave					50 V
5696	Relay Service	7BN	Htr.	6.3	0.15	500[1]	100 ma. peak current; 25-ma. average.					
5727	Gas Thyratron	7BN	Htr.	6.3	0.6	650	—	—	—	—	—	—
5823	Relay or Trigger	4CK	Cold	—	—	Max. peak inv. volts = 200; Peak mA = 100; Avg. mA = 25.						
5962	Voltage Regulator	2AG	Cold	—	—	—	—	730	700	5/55[2]	—	—
5998	Series Regulator	8BD	Htr.	6.3	2.4	250	125	—	110	100	350[4]	—

[1] Peak inverse voltage.
[2] Values in microamperes.

TABLE V—RECTIFIERS—RECEIVING AND TRANSMITTING

See Also Table IV—Controls and Regulator Tubes

Type	Name	Base	Cathode	Fil. or Heater Volts	Fil. or Heater Amp.	Max. AC Voltage Per Plate	DC Output Current mA	Max. Inverse Peak Voltage	Peak Plate Current mA	Type
0Z4-G	Full-Wave Rectifier	4R	Cold	—	—	300	75	1000	200	GAS
1G3-GT/ 1B3-GT	Half-Wave Rectifier	3C	Fil.	1.25	0.2	—	1.0	33000	30	HV
1K3/1J3	Half-Wave Rectifier	3C	Fil.	1.25	0.2	—	0.5	26000	50	HV
1V2	Half-Wave Rectifier	9U	Fil.	0.625	0.3	—	0.5	7500	10	HV
2X2-A	Half-Wave Rectifier	4AB	Htr.	2.5	1.75	4500	7.5	—	—	HV
2Y2	Half-Wave Rectifier	4AB	Fil.	2.5	1.75	4400	5.0	—	—	HV
2Z2/G84	Half-Wave Rectifier	4B	Fil.	2.5	1.5	350	50	—	—	HV

TABLE V — RECTIFIERS — RECEIVING AND TRANSMITTING — *Continued*

See Also Table IV — Controls and Regulator Tubes

Type	Name	Base	Cathode	Fil. or Heater Volts	Fil. or Heater Amp.	Max. AC Voltage Per Plate	D.C. Output Current mA	Max. Inverse Peak Voltage	Peak Plate Current mA	Type
3B24	Half-Wave Rectifier	Fig. 49	Fil.	5.0 2.5[5]	3.0 3.0	— —	60 30	20000 20000	300 150	HV
3B28	Half-Wave Rectifier	4P	Fil.	2.5	5.0	—	250	10000	1000	GAS
5AT4	Full-Wave Rectifier	5L	Htr.	5.0	2.25	550	800	1550	—	HV
5AU4	Full-Wave Rectifier	5T	Fil.	5.0	4.5	300[3] 400[3] 500[4]	350[3] 325[3] 325[4]	1400	1075	HV
5AW4	Full-Wave Rectifier	5T	Fil.	5.0	4.0	450[3] 550[4]	250[3] 250[4]	1550	750	HV
5R4GY 5R4GYA	Full-Wave Rectifier	5T	Fil.	5.0	2.0	900[3] 950[4]	150[3] 175[4]	2800	650	HV
5U4G	Full-Wave Rectifier	5T	Fil.	5.0	3.0	Same as Type 5Z3				HV
5U4GA	Full-Wave Rectifier	5T	Fil.	5.0	3.0	300[3] 450[3] 550[4]	275[3] 250[3] 250[4]	1550	900	HV
5U4GB 5AS4A	Full-Wave Rectifier	5T	Fil.	5.0	3.0	300[3] 450[3] 550[4]	300[3] 275[3] 275[4]	1550	1000	HV
5V3A	Full-Wave Rectifier	5T	Htr.	5.0	3.8	425[3] 500[4]	350	1400	1200	HV
5V4GA	Full-Wave Rectifier	5L	Htr.	5.0	2.0	375[3]	175	1400	525	HV
5Y3-G-GT	Full-Wave Rectifier	5T	Fil.	5.0	2.0	Same as Type 80				HV
5Z3	Full-Wave Rectifier	4C	Fil.	5.0	3.0	500	250	1400	—	HV
5Z4	Full-Wave Rectifier	5L	Htr.	5.0	2.0	400	125	1100	—	HV
6AV4	Full-Wave Rectifier	5BS	Htr.	6.3	0.95	—	90	1250	250	HV
6AX5GT	Full-Wave Rectifier	6S	Htr.	6.3	1.2	450	125	1250	375	HV
6BW4	Full-Wave Rectifier	9DJ	Htr.	6.3	0.9	450	100	1275	350	HV
6BX4	Full-Wave Rectifier	5BS	Htr.	6.3	0.6	—	90	1350	270	HV
6BY5G	Full-Wave Rectifier	6CN	Htr.	6.3	1.6	375[3]	175	1400	525	HV
6CA4	Full-Wave Rectifier	9M	Htr.	6.3	1.0	350[3]	150	1000	450	HV
6DE4	Half-Wave Rectifier	4CG	Fil.	6.3	1.6	—	175	5000	1100	HV
6V4	Full-Wave Rectifier	9M	Htr.	6.3	0.6	350	90	—	—	HV
6X4/6063 6X5GT	Full-Wave Rectifier	7CF 6S	Htr.	6.3	0.3	325[3] 450[4]	70	1250	210	HV
6Z3	Half-Wave Rectifier	4G	Fil.	6.3	0.3	350	50	—	—	HV
12X4	Full-Wave Rectifier	5BS	Htr.	12.6	0.3	650[3] 900[4]	70 70	1250 1250	210 210	HV
25Z5	Rectifier-Doubler	6E	Htr.	25	0.3	125	100	—	500	HV
35W4	Half-Wave Rectifier	5BQ	Htr.	35[1]	0.15	125	60	330	600	HV
35Z4GT	Half-Wave Rectifier	5AA	Htr.	35	0.15	250	100	700	600	HV
35Z5G	Half-Wave Rectifier	6AD	Htr.	35[1]	0.15	125	60	—	—	HV
36AM3	Half-Wave Rectifier	5BQ	Htr.	36	0.1	117	75	365	530	HV
50DC4	Half-Wave Rectifier	5BQ	Htr.	50	0.15	117	100	330	720	HV
50Y6GT	Full-Wave Rectifier	7Q	Htr.	50	0.15	125	85	—	—	HV
80	Full-Wave Rectifier	4C	Fil.	5.0	2.0	350[3] 500[4]	125 125	1400	375	HV
83	Full-Wave Rectifier	4C	Fil.	5.0	3.0	500	250	1400	800	MV
83-V	Full-Wave Rectifier	4AD	Htr.	5.0	2.0	400	200	1100	—	HV
117N7GT	Rectifier-Tetrode	8AV	Htr.	117	0.09	117	75	350	450	HV
117Z3	Half-Wave Rectifier	4CB	Htr.	117	0.04	117	90	300	—	HV
816	Half-Wave Rectifier	4P	Fil.	2.5	2.0	2200	125	7500	500	MV
836	Half-Wave Rectifier	4P	Htr.	2.5	5.0	—	—	5000	1000	HV
866-A-AX	Half-Wave Rectifier	4P	Fil.	2.5	5.0	3500	250	10000	1000	MV
866B	Half-Wave Rectifier	4P	Fil.	5.0	5.0	—	—	8500	1000	MV
866 Jr.	Half-Wave Rectifier	4B	Fil.	2.5	2.5	1250	250[2]	—	—	MV
872A/872	Half-Wave Rectifier	4AT	Fil.	5.0	7.5	—	1250	10000	5000	MV

[1] Tapped for pilot lamps.
[2] Per pair with choke input.
[3] Capacitor input.
[4] Choke input.
[5] Using only one-half of filament.

TABLE VI — TRIODE TRANSMITTING TUBES

Type	Maximum Ratings: Plate Dissipation Watts	Plate Voltage	Plate Current mA	DC Grid Current mA	Freq. MHz. Full Ratings	Amplification Factor	Cathode: Volts	Amperes	Capacitances: C_{in} pF	C_{gp} pF	C_{out} pF	Base	Typical Operation: Class of Service	Plate Voltage	Grid Voltage	Plate Current mA	DC Grid Current mA	Approx. Driving Power Watts	P-to-P Load Ohms	Approx. Output Power Watts
6J6A‡[2]	1.5	300	30	16	250	32	6.3	0.45	2.2	1.6	0.4	7BF	C-T	150	−10	30	1.6	0.035	—	3.5
6F4	2.0	150	20	8.0	500	17	6.3	0.225	2.0	1.9	0.6	7BR	C-T-O	150	−15 550[3] 2000[4]	20	7.5	0.2	—	1.8
12AU7A[7]	2.76[6]	350	12[6]	3.5[6]	54	18	6.3	0.3	1.5	1.5	0.5	9A	C-T-O	350	−100	24	7	—	—	6.0
6C4	5.0	350	25	8.0	54	18	6.3	0.15	1.8	1.6	1.3	6BG	C-T-O	300	−27	25	7.0	0.35	—	5.5
5675	5	165	30	8	3000	20	6.3	0.135	2.3	1.3	0.09	Fig. 21	G-G-O	120	−8	25	4	—	—	0.05
6N7GT[7]	5.5[6]	350	30[6]	5.0[6]	10	35	6.3	0.8	—	—	—	8B	C-T-O	350	−100	60	10	—	—	14.5
2C40	6.5	500	25	—	500	36	6.3	0.75	2.1	1.3	0.05	Fig. 11	C-T-O	250	−5	20	0.3	—	—	0.075
5893	8.0	400	40	13	1000	27	6.0	0.33	2.5	1.75	0.07	Fig. 21	C-T C-P	350 300	−33 −45	35 30	13 12	2.4 2.0	— —	6.5 6.5
2C43	12	500	40	—	1250	48	6.3	0.9	2.9	1.7	0.05	Fig. 11	C-T-O	470	—	38[7]	—	—	—	9[7]
3C24	25 17 25	2000 1600 2000	75 60 75	7[11]	60	24	6.3	3.0	1.7	1.6	0.2	2D	C-T C-P AB$_2$[7]	2000 1600 1250	−130 −170 −42	63 53 24/130	18 11 270[9]	4 3.1 3.4[8]	— — 21.4K	100 68 112
1623	30	1000	100	25	60	20	6.3	2.5	5.7	6.7	0.9	3G	C-T-O C-P B[7]	1000 750 1000	−90 −125 −40	100 100 30/200	20 20 230[9]	3.1 4.0 4.2[8]	— — 12K	75 55 145

	Maximum Ratings						Cathode		Capacitances				Typical Operation							
Type	Plate Dissipation Watts	Plate Voltage	Plate Current mA	DC Grid Current mA	Freq. MHz Full Ratings	Amplification Factor	Volts	Amperes	C_{in} pF	C_{gp} pF	C_{out} pF	Base	Class of Service	Plate Voltage	Grid Voltage	Plate Current mA	DC Grid Current mA	Approx. Driving Power Watts	P-to-P Load Ohms	Approx. Output Power Watts
811-A	65	1500	175	50	60	160	6.3	4.0	5.9	5.6	0.7	3G	C-T	1500	−70	173	40	7.1	—	200
													C-P	1250	−120	140	45	10.0	—	135
													G-G-B	1250	0	27/175	28	12	—	165
													AB_1	1250	0	27/175	13	3.0	—	155
812-A	65	1500	175	35	60	29	6.3	4.0	5.4	5.5	0.77	3G	C-T	1500	−120	173	30	6.5	—	190
													C-P	1250	−115	140	35	7.6	—	130
													B[7]	1500	−48	28/310	270[8]	5.0	13.2K	340
100TH	100	3000	225	60	40	40	5.0	6.3	2.9	2.0	0.4	2D	C-T C-P	3000	−200	165	51	18	—	400
													B[7]	3000	−65	40/215	335[8]	5.0[8]	31K	650
3-100A2 100TL	100	3000	225	50	40	14	5.0	6.3	2.3	2.0	0.4	2D	C-T C-P	3000	−400	165	30	20	—	400
													G-M-A	3000	−560	60	2.0	7.0	—	90
													B[7]	3000	−185	40/215	640[8]	6.0[8]	30K	450
3CX100A5[15]	100	1000	125[14]	50	2500	100	6.0	1.05	7.0	2.15	0.035	—	G-G-A	800	−20	80	30	6	—	27
	70	600	100[14]										C-P	600	−15	75	40	6	—	18
2C39	100	1000	60	40	500	100	6.3	1.1	6.5	1.95	0.03	—	G-I-C	600	−35	60	40	5.0	—	20
													C-T-O	900	−40	90	30	—	—	40
													C-P	600	−150	100[14]	50	—	—	—
AX9900/ 5866[15]	135	2500	200	40	150	25	6.3	5.4	5.8	5.5	0.1	Fig. 3	C-T	2500	−200	200	40	16	—	390
													C-P	2000	−225	127	40	16	—	204
													B[7]	2500	−90	80/330	350[8]	14[8]	15.68K	560
572B/T160L	160	2750	275	—	—	170	6.3	4.0	—	—	—	3G	C-T	1650	−70	165	32	6	—	205
													G-G-B[7]	2400	−2.0	90/500	—	100	—	600
810	175	2500	300	75	30	36	10	4.5	8.7	4.8	12	2N	C-T	2500	−180	300	60	19	—	575
													C-P	2000	−350	250	70	35	—	380
													G-M-A	2250	−140	100	2.0	4	—	75
													B[7]	2250	−60	70/450	380[8]	13[8]	11.6K	725
8873	200	2200	250	—	500	160	6.3	3.2	19.5	7.0	0.03	Fig. 87	AB_2	2000	—	22/500	98[8]	27[8]	—	505
250TH	250	4000	350	40[13]	40	37	5.0	10.5	4.6	2.9	0.5	2N	C-T-O	2000	−100	357	94	29	—	464
														3000	−150	333	90	32	—	750
													C-P	2000	−160	250	60	22	—	335
														2500	−180	225	45	17	—	400
														3000	−200	200	38	14	—	435
													AB_2[7]	1500	0	220/700	460[8]	46[8]	4.2K	630
250TL	250	4000	350	35[13]	40	14	5.0	10.5	3.7	3.0	0.7	2N	C-T-O	2000	−200	350	45	22	—	455
														3000	−350	335	45	29	—	750
													C-P	2000	−520	250	29	24	—	335
														2500	−520	225	20	16	—	400
														3000	−520	200	14	11	—	435
													AB_2[7]	1500	−40	200/700	780[8]	38[8]	3.8K	580
PL-6569	250	4000	300	120	30	45	5.0	14.5	7.6	3.7	0.1	Fig. 3	G-G-A	2500	−70	300	85	75[11]	—	555
														3000	−95	300	110	85[11]	—	710
														3500	−110	285	90	85[11]	—	805
														4000	−120	250	50	70[11]	—	820
8875	300	2200	250	—	500	160	6.3	3.2	19.5	7.0	0.03	—	AB_2	2000	—	22/500	98[8]	27[8]	—	505
304TH	300	3000	900	60[13]	40	20	5.0	25	13.5	10.2	0.7	4BC	C-T-O	1500	−125	665	115	25	—	700
														2000	−200	600	125	39	—	900
													C-P	1500	−200	420	55	18	—	500
							10	12.5						2000	−300	440	60	26	—	680
														2500	−350	400	60	29	—	800
													AB_2[7]	1500	−65	1065[8]	330[8]	25[8]	2.84K	1000
304TL	300	3000	900	50[13]	40	12	5.0	25	12.1	8.6	0.8	4BC	C-T-O	1500	−250	665	90	33	—	700
														2000	−300	600	85	36	—	900
													C-P	2000	−500	250	30	18	—	410
														2000	−500	500	75	52	—	810
														2500	−525	200	18	11	—	425
														2500	−550	400	50	36	—	830
							10	12.5					AB_1[7]	1500	−118	270/572	236[8]	0	2.54K	256
														2500	−230	160/483	460[8]	0	8.5K	610
													AB_2[7]	1500	−118	1140[8]	490[8]	39[8]	2.75K	1100
833A	350	3300	500	100	30	35	10	10	12.3	6.3	8.5	Fig. 41	C-T-O	2250	−125	445	85	23	—	780
														3000	−160	335	70	20	—	800
	450[15]	4000[15]			20[15]								C-P	2500	−300	335	75	30	—	635
														3000	−240	335	70	26	—	800
													B[7]	3000	−70	100/750	400[8]	20[8]	9.5K	1650
8874	400	2200	250	—	500	160	6.3	3.2	19.5	7.0	0.03	—	AB_2	2000	—	22/500	98[8]	27[8]	—	505
3-400Z	400	3000	400	—	110	200	5	14.5	7.4	4.1	0.07	Fig. 3	G-G-B	3000	0	100/333	120	32	—	655
PL-8580	400	4000[15]	350	120	—	45	5.0	14.5	7.6	3.9	0.1	5BK	G-G-A	4000	−110	350	92	105[11]	—	1080
														2500	−70	350	95	85	—	660
8163	400	3000	400	20[13]	30	350	5.0	14.1	8.0	5.0	0.3	Fig. 3	G-G-B	2500	0	72/400	140	35	—	640
3-500Z	500	4000	400	—	110	160	5	14.5	7.4	4.1	0.07	Fig. 3	G-G-B	3000	—	370	115	30	5K	750
													C-T	3500	−75	300	115	22	—	850
3-1000Z	1000	3000	800	—	110	200	7.5	21.3	17	6.9	0.12	Fig. 3	G-G-B	3000	0	180/670	300	65	—	1360
8077	1500	4000	1000	—	250	200	5.0	10	42	10	0.1	—	AB_2	2500	−8.2	1000	—	57	—	1520

* Cathode resistor in ohms.

[1] KEY TO CLASS-OF-SERVICE ABBREVIATIONS
- A_1 — Class-A_1 af modulator.
- AB_1 — Class-AB_1 push-pull af modulator.
- AB_2 — Class-AB_2 push-pull af modulator.
- B — Class-B push-pull af modulator.
- C-M — Frequency multiplier.
- C-P — Class-C plate-modulated telephone.
- C-T — Class-C telegraph.
- C-T-O — Class-C amplifier-osc.
- G-G-A — Grounded-grid class-C amp.
- G-G-B — Grounded-grid class-B amp. (Single Tone).
- G-G-O — Grounded-grid osc.
- G-I-C — Grid-isolation circuit.
- G-M-A — Grid-modulated amp.

[2] Twin triode. Values, except interelectrode capacitances, are for both sections in push-pull.
[3] Output at 112 MHz.
[4] Grid leak resistor in ohms.
[5] Peak values.
[6] Per section.
[7] Values are for two tubes in push-pull.
[8] Max. signal value.
[9] Peak af grid-to-grid volts.
[10] Plate-pulsed 1000-MHz. osc.
[11] Includes bias loss, grid dissipation, and feed-through power.
[12] 1000-MHz. cw osc.
[13] Max. grid dissipation in watts.
[14] Max. cathode current in mA.
[15] Forced-air cooling required.
[16] Plate-pulsed 3300-MHz. osc.
[17] 1900-MHz. cw osc.
[18] No Class-B data available.

TABLE VII—TETRODE AND PENTODE TRANSMITTING TUBES

Type	Maximum Ratings: Plate Dissipation Watts	Plate Voltage	Screen Dissipation Watts	Screen Voltage	Freq. MHz. Full Ratings	Cathode: Volts	Amperes	Capacitances: C_{in} pF	C_{gp} pF	C_{out} pF	Base	Typical Operation: Class of Service[14]	Plate Voltage	Screen Voltage	Suppressor Voltage	Grid Voltage	Plate Current mA	Screen Current mA	Grid Current mA	Approx. Driving Power Watts	P-to-P Load Ohms	Approx. Output Power Watts
8203	1.8	400	—	—	250	6.3	0.16	4.2	2.2	1.6	12AQ	C·P/C·T	155	—	—	14/2700[1]	21	—	5	0.4	—	1.55
6939[3]	7.5	275	3	200	500	6.3 / 12.6	0.75 / 0.375	6.6	0.15	1.55	Fig. 13	C·T	200	200	—	−20	60	13	2	1.0	—	7.5
												C·P	180	180	—	−20	55	11.5	1.7	1.0	—	6
												C·M	200	190	—	68K[1]	46	10	2.2	0.9	—	—
7551 / 7558	12	300	2	250	175	12.6 / 6.3	0.38 / 0.8	10	0.15	5.5	9LK	C·T	300	250	—	−55	80	5.1	1.6	1.5	—	10
												C·P	250	250	—	−75	70	3.0	2.3	1.0	—	7.5
5763 / 6417	13.5	350	2	250	50	6.3 / 12.6	0.75 / 0.375	9.5	0.3	4.5	9K	C·T	350	250	—	−28.5	48.5	6.2	1.6	0.1	—	12
												C·P	300	250	—	−42.5	50	6	2.4	0.15	—	10
												C·M[2]	300	250	—	−75	40	4	1	0.6	—	2.1
												C·M[4]	300	235	—	−100	35	5	1	0.6	—	1.3
2E26 / 6893	13.5	600	2.5	200	125	6.3 / 12.6	0.8 / 0.4	12.5	0.2	7	7CK	C·T	600	185	—	−45	66	10	3	0.17	—	27
												C·P	500	180	—	−50	54	9	2.5	0.15	—	18
												AB_1	500	200	—	−25	9/45	10[7]	0	0	—	15
6360[3]	14	300	2	200	200	6.3 / 12.6	0.82 / 0.41	6.2	0.1	2.6	Fig. 13	C·T	300	200	—	−45	100	3	3	0.2	—	18.5
												C·P	200	100	—	15K[1]	86	3.1	3.3	0.2	—	9.8
												C·M[11]	300	150	—	−100	65	3.5	3.8	0.45	—	4.8
												AB_2	300	200	—	−21.5	30/100	1/11.4	64[8]	0.04	6.5K	17.5
2E25	15	450	4	250	125	6	0.8	8.5	0.15	6.7	5BJ	C·T·O	450	250	—	−45	75	15	3	0.4	—	24
												C·P	400	200	—	−45	60	12	3	0.4	—	16
												AB_2[6]	450	250	—	−30	44/150	10/40	3	0.9[7]	6K	40
832A[3]	15	750	5	250	200	6.3 / 12.6	1.6 / 0.8	8	0.07	3.8	7BP	C·T	750	200	—	−65	48	15	2.8	0.19	—	26
												C·P	600	200	—	−65	36	16	2.6	0.16	—	17
6252/ AX9910[3]	20	750	4	300	300	6.3 / 12.6	1.3 / 0.65	6.5	—	2.5	Fig. 7	C·T	600	250	—	−60	140	14	4	2.0	—	—
												C·P	500	250	—	−80	100	12	3	4.0	—	—
												B	500	250	—	−26	25/73	0.7/16	52[8]	—	20K	23.5
1614	25	450	3.5	300	80	6.3	0.9	10	0.4	12.5	7AC	C·T	450	250	—	−45	100	8	2	0.15	—	31
												C·P	375	250	—	−50	93	7	2	0.15	—	24.5
												AB_1[6]	530	340	—	−36	60/160	20[7]	—	—	7.2K	50
815[3]	25	500	4	200	125	6.3 / 12.6	1.6 / 0.8	13.3	0.2	8.5	8BY	C·T·O	500	200	—	−45	150	17	2.5	0.13	—	56
												AB_2	500	125	—	−15	22/150	32[7]	—	0.36[7]	8K	54
6146 / 6146A	25	750	3	250	60	6.3	1.25	13	0.24	8.5	7CK	C·T	500	170	—	−66	135	9	2.5	0.2	—	48
													750	160	—	−62	120	11	3.1	0.2	—	70
8032 / 6883						12.6	0.585					C·T[12]	400	190	—	−54	150	10.4	2.2	3.0	—	35
												C·P	400	150	—	−87	112	7.8	3.4	0.4	—	32
													600	150	—	−87	112	7.8	3.4	0.4	—	52
6159B						26.5	0.3					AB_2[6]	600	190	—	−48	28/270	1.2/20	2[7]	0.3	5K	113
													750	165	—	−46	22/240	0.3/20	2.6[7]	0.4	7.4K	131
												AB_1[6]	750	195	—	−50	23/220	1/26	100[8]	0	8K	120
6524[3] / 6850	25	600	—	300	100	6.3 / 12.6	1.25 / 0.625	7	0.11	3.4	Fig. 76	C·T	600	200	—	−44	120	8	3.7	0.2	—	56
												C·P	500	200	—	−61	100	7	2.5	0.2	—	40
												AB_2	500	200	—	−26	20/116	0.1/10	2.6	0.1	11.1K	40
807 / 807W / 5933	30	750	3.5	300	60	6.3	0.9	12	0.2	7	5AW	C·T	750	250	—	−45	100	6	3.5	0.22	—	50
												C·P	600	275	—	−90	100	6.5	4	0.4	—	42.5
												AB_1	750	300	—	−35	15/70	3/8	75[8]	0	—	72
1625						12.6	0.45				5AZ	B[10]	750	—	—	0	15/240	—	555[8]	5.3[7]	6.65K	120
2E22	30	750	10	250	—	6.3	1.5	13	0.2	8	5J	C·T·O	750	250	22.5	−60	100	16	6	0.55	—	53
6146B/ 8298A	35	750	3	250	60	6.3	1.125	13	0.22	8.5	7CK	C·T	750	200	—	−77	160	10	2.7	0.3	—	85
												C·P	600	175	—	−92	140	9.5	3.4	0.5	—	62
												AB_1	750	200	—	−48	25/125	6.3	—	—	3.6K	61
AX-9903[3] / 5894A	40	600	7	250	250	6.3 / 12.6	1.8 / 0.9	6.7	0.08	2.1	Fig. 7	C·T	600	250	—	−80	200	16	2	0.2	—	80
829B[3] / 3E29[3]	40	750	7	240	200	6.3 / 12.6	2.25 / 1.125	14.5	0.12	7	7BP	C·T	500	200	—	−45	240	32	12	0.7	—	83
												C·P	425	200	—	−60	212	35	11	0.8	—	63
												B	500	200	—	−18	27/230	—	56[8]	0.39	4.8K	76
3D24	45	2000	10	400	125	6.3	3	6.5	0.2	2.4	Fig. 75	C·T·O	2000	375	—	−300	90	20	10	4.0	—	140
													1500	375	—	−300	90	22	10	4.0	—	105
4D22	50	750	14	350	60	12.6 / 25.2	1.6 / 0.8	28	0.27	13	Fig. 26	C·T	750	300	—	−100	240	26	12	1.5	—	135
													600	300	—	−100	215	30	10	1.25	—	100
4D32						6.3	3.75				Fig. 27	C·P	600	—	—	−100	220	28	10	1.25	—	100
													550	—	—	−100	175	17	6	0.6	—	70
												AB_2[6]	600	250	—	−25	100/365	26[7]	70[8]	0.45[7]	3K	125
8117[3]	60	750	7	300	175	6.3 / 12.6	1.8 / 0.9	11.8	3.7	0.09	Fig. 7	AB_1	600	250	—	−32.5	60/212	1.9/25	—	—	1410	76
814	65	1500	10	300	30	10	3.25	13.5	0.1	13.5	Fig. 64	C·T	1500	300	—	−90	150	24	10	1.5	—	160
												C·P	1250	300	—	−150	145	20	10	3.2	—	130
4-65A	65	3000	10	600	150	6	3.5	8	0.08	2.1	Fig. 25	C·T·O	1500	250	—	−85	150	40	18	3.2	—	165
													3000	250	—	−100	115	22	10	1.7	—	280
												C·P	1500	250	—	−125	120	40	16	3.5	—	140
													2500	250	—	−135	110	25	12	2.6	—	230
												AB_1	2500	400	—	−85	15/66	3[7]	—	—	—	100
7854[3]	68	1000	8	300	175	6.3 / 12.6	1.8 / 0.9	6.7	2.1	0.09	Fig. 7	C·T	750	260	—	−75	240	12.7	5.5	3.5	—	123
												C·P	600	225	—	−75	200	7.8	5.5	3.5	—	85
4E27/ 8001	75	4000	30	750	75	5	7.5	12	0.06	6.5	7BM	C·T	2000	500	60	−200	150	11	6	1.4	—	230
												C·P	1800	400	60	−130	135	11	8	1.7	—	178
PL-177A	75	2000	10	600	175	6	3.2	7.5	0.06	4.2	Fig. 14	C·T·C·P	2000	400	0	−125	150	12	5	0.8	—	220
													1000	400	0	−105	150	16	5	0.7	—	100
												AB_1	2000	600	—	−115	25/175	0/7	0	0	—	210
7270 / 7271	80	1350	—	425	175	6.3 / 13.5	3.1 / 1.25	8	0.4	0.14	Fig. 84	C·T	850	400	—	−100	275	15	8	10	—	135
												AB_1	665	400	—	−119	220	15	6	10	—	85
8072	100	2200	8	400	500	13.5	1.3	16	0.13	0.011	Fig. 85	C·T·O	700	200	—	−30	300	10	20	5	—	85

Type	Maximum Ratings					Cathode		Capacitances			Base	Typical Operation										
	Plate Dissipation Watts	Plate Voltage	Screen Dissipation Watts	Screen Voltage	Freq. MHz. Full Ratings	Volts	Amperes	C_{in} pF	C_{gp} pF	C_{out} pF		Class of Service[14]	Plate Voltage	Screen Voltage	Suppressor Voltage	Grid Voltage	Plate Current mA	Screen Current mA	Grid Current mA	Approx. Driving Power Watts	P-to-P Load Ohms	Approx. Output Power Watts
6816[9]	115	1000	4.5	300	400	6.3	2.1	14	0.085	0.015	Fig. 77	C-T-O	900	300	—	−30	170	1	10	3	—	80
6884						26.5	0.52					C-P	700	250	—	−50	130	10	10	3	—	45
												AB_1[6]	850	300	—	−15	80/200	0/20	30[8]	0	7K	80
												AB_2[6]	850	300	—	−15	80/335	0/25	46[6]	0.3	3.96K	140
813[13]	125	2500	20	800	30	10	5	16.3	0.25	14	5BA	C-T-O	1250	300	0	−75	180	35	12	1.7	—	170
													2250	400	0	−155	220	40	15	4	—	375
												AB_1	2500	750	0	−95	25/145	27[7]	0	0	—	245
												AB_2[6]	2000	750	0	−90	40/315	1.5/58	230[8]	0.1[7]	16K	455
													2500	750	0	−95	35/360	1.2/55	235[6]	0.35[7]	17K	650
4-125A 4D21 6155	125	3000	20	600	120	5	6.5	10.8	0.07	3.1	5BK	C-T-O	2000	350	—	−100	200	50	12	2.8	—	275
													3000	350	—	−150	167	30	9	2.5	—	375
												AB_2[6]	2500	350	—	−43	93/260	0/6	178[6]	1.0[7]	22K	400
												AB_1[6]	2500	600	—	−96	50/232	0.3/8.5	192[8]	0	20.3K	330
												GG	2000	0	—	0	10/105[17]	30[17]	55[17]	16[17]	10.5K	145
4E27A/ 5-125B	125	4000	20	750	75	5	7.5	10.5	0.08	4.7	7BM	C-	3000	500	60	−200	167	5	6	1.6	—	375
													1000	750	0	−170	160	21	3	0.6	—	115
803	125	2000	30	600	20	10	5	17.5	0.15	29	5J	C-T	2000	500	40	−90	160	45	12	2	—	210
												C-P	1600	400	100	−80	150	45	25	5	—	155
7094	125	2000	20	400	60	6.3	3.2	9.0	0.5	1.8	Fig. 82	C-T	1500	400	—	−100	330	20	5	4	—	340
												C-P	1200	400	—	−130	275	20	5	5	—	240
												AB_1	2000	400	—	−65	30/200	35[7]	60[8]	0	12K	250
4X150A	150[9]	2000	12	400	500	6	2.6	15.5	0.03	4.5	Fig. 75	C-T-O	1250	250	—	−90	200	20	10	0.8	—	195
												C-P	1000	250	—	−105	200	20	15	2	—	140
4X150G[15]						2.5	6.25	27	0.035	4.5	—	AB_2[6]	1250	300	—	−44	475[7]	0/65	100[8]	0.15[7]	5.6K	425
8121	150	2200	8	400	500	13.5	1.3	16	0.13	0.011	Fig. 5	C-T-O	1000	200	—	−30	300	10	30	5	—	165
4-250A 5D22 6156	250[9]	4000	35	600	110	5	14.5	12.7	0.12	4.5	5BK	C-T-O	2500	500	—	−150	300	60	9	1.7	—	575
													3000	500	—	−180	345	60	10	2.6	—	800
												C-P	2500	400	—	−200	200	30	9	2.2	—	375
													3000	400	—	−310	225	30	9	3.2	—	510
												AB_2[6]	2000	300	—	−48	510[7]	0/26	198[6]	5.5[7]	8K	650
												AB_1[6]	2500	600	—	−110	430[7]	0.3/13	180[8]	0	11.4K	625
4X250B	250[9]	2000	12	400	175	6	2.1	18.5	0.04	4.7	Fig. 75	C-T-O	2000	250	—	−90	250	25	27	2.8	—	410
												C-P	1500	250	—	−100	200	25	17	2.1	—	250
												AB_1[6]	2000	350	—	−50	500[7]	30[7]	100[8]	0	8.26K	650
7034/[9] 4X150A	250	2000	12	300	150	6	2.6	16	0.03	4.4	Fig. 75	C-T-O	2000	250	—	−88	250	24	8	2.5	—	370
												C-P	1600	250	—	−118	200	23	5	3	—	230
7035/ 4X150D	250	2000	12	400		26.5	0.58					AB_2[6]	2000	300	—	−50	100/500	0/36	106[6]	0.2	8.1K	630
												AB_1[6]	2000	300	—	−50	100/470	0/36	100[6]	0	8.76K	580
4CX-300A	300[9]	2000	12	400	500	6	2.75	29.5	0.04	4.8	—	C-T	2000	250	—	−90	250	25	27	2.8	—	410
												C-P	1500	250	—	−100	200	25	17	2.1	—	250
												AB_1[6]	2000	350	—	−50	500[7]	30[7]	100[6]	0	8.26K	650
175A	400	4000	25	600	—	5	14.5	15.1	0.06	9.8	Fig. 86	C-T-C-P	4000	600	0	−200	350	29	6	1.4	—	960
													2500	600	0	−180	350	40	7	1.6	—	600
												AB_1	2500	750	—	−143	100/350	1/35	0	0	—	570
4-400A	400[9]	4000	35	600	110	5	14.5	12.5	0.12	4.7	5BK	C-T-C-P	4000	300	—	−170	270	22.5	10	10	—	720
												GG	2500	0	—	0	80/270[17]	55[17]	100[17]	38[17]	4.0K	325
												AB_1	2500	750	—	−130	95/317	0/14	0	0	—	425
8122	400	2200	8	400	500	13.5	1.3	16	0.13	0.011	Fig. 86	C-T-O	2000	200	—	−30	300	5	30	5	—	300
5-500A	500	4000	35	600	30	10	10.2	19	0.10	12	—	C-T	3000	500	0	−220	432	65	35	12	—	805
												C-T	3100	470	0	−310	260	50	15	6	—	580
												AB_1	3000	750	0	−112	320	26	—	—	—	612
8166/ 4-1000A	1000	6000	75	1000	—	7.5	21	27.2	.24	7.6	—	C-T	3000	500	—	−150	700	146	38	11	—	1430
												C-P	3000	500	—	−200	600	145	36	12	—	1390
												AB_2	4000	500	—	−60	300/1200	0/95	—	11	7K	3000
												GG	3000	0	—	0	100/700[17]	105[17]	170[17]	130[17]	2.5K	1475
4CX1000A	1000	3000	12	400	400	6	12.5	35	.005	12	—	AB_1[6]	2000	325	—	−55	500/2000	−4/60	—	—	2.8K	2160
													2500	325	—	−55	500/2000	−4/60	—	—	3.1K	2920
													3000	325	—	−55	500/1800	−4/60	—	—	3.85K	3360
8295/ 172	1000	3000	30	600	—	6	8.2	38	.09	18	—	C-T	2000	500	35	−175	850	42	10	1.9	—	1155
													2500	500	35	−200	840	40	10	2.1	—	1440
													3000	500	35	−200	820	42	10	2.1	—	1770
												AB_1	2000	500	35	−110	200/800	12/43	110[8]	—	2.65K	1040
													2500	500	35	−110	200/800	11/40	115[8]	—	3.5K	1260
													3000	500	35	−115	220/800	11/39	115[6]	—	4.6K	1590

1. Grid-resistor.
2. Doubler to 175 MHz.
3. Dual tube. Values for both sections, in push-pull. Interelectrode capacitances, however, are for each section.
4. Tripler to 175 MHz.
5. Filament limited to intermittent operation.
6. Values are for two tubes
7. Max.-signal value.
8. Peak grid-to-grid volts.
9. Forced-air cooling required.
10. Two tubes triode connected, G_2 to G_1 through 20K Ω. Input to G_2.
11. Tripler to 200 MHz.
12. Typical Operation at 175 MHz.
13. ± 1.5 volts.
14. KEY TO CLASS-OF-SERVICE ABBREVIATIONS
    - AB_1 = Class-AB_1.
    - AB_2 = Class-AB_2.
    - B = Class-B push-pull at modulator.
    - C-M = Frequency multiplier.
    - C-P = Class-C plate-modulated telephone.
    - C-T = Class-C telegraph.
    - C-T-O = Class-C amplifier-osc.
    - GG = Grounded-grid (grid and screen connected together).
15. No Class B data available.
16. HK257B 120 MHz. full rating.
17. Single tone.

TABLE VIII — SEMICONDUCTOR DIODES[1]

This list contains but a small percentage of the available diode types. A complete listing would be impractical.

Small-Signal General-Purpose Diodes

Type	Material[2]	Use	Peak Reverse Volts	Max. Forward Voltage at Max. mA	Max. Forward mA at Max. V	Max. Reverse μ-amp
1N34A	G	General Purpose	75	50	5.0	30
1N35	G	General Purpose	50	—	—	2000
1N52A	G	General Purpose	85	50	5.0	100
1N60	G	Video Detector	25	50	5.0	40
1N67A	G	General Purpose	100	1	4.0	5
1N64A	G	General Purpose	100	1.0	4.0	5
1N94	S	High-Speed	75	1.0	10.0	5
1N270	G	General Purpose	100	90	—	100
1N458A	S	Signal Diode	125	1	100	25
1N634	G	60-Volt Very Low Z	120	—	50.0	45
1N645A	S	Signal Diode	225	1	400	.05
1N914	S	Fast Logic/HF Det.	75	1	75	.025
1N3754	S	Rectifier	100	1.2	150	300
1N4001	S	Rectifier	50	1.1	1000	30
1N4002	S	Rectifier	100	1.1	1000	30
1N4004	S	Rectifier	400	1.1	1000	30
1N4719	S	Rectifier	50	1.0	3000	1500
MSD7000	S	Dual Series Diode	100	1.1	200	100

Microwave Mixer and UHF Diodes

Type	Material[2]	Use	Average Freq.	Noise Figure
1N21F[3]	G	Mixer	3060 MHz	6dB
1N82A	S	Mixer	1000 MHz	14dB
MBD101	S	Mixer (V_R = 4v)	1000 MHz (C_T = 1pF @ 0V)	7dB

[1] A bar, plus sign, or color dot usually denotes the cathode end of crystal diodes.
Diode color code rings are grouped toward the cathode end.

[2] S = Silicon. G = Germanium.

[3] Polarity is such that the base is the anode and the tip is the cathode, R-types have opposite polarity.

TABLE IX — SEMICONDUCTORS

SMALL-SIGNAL TYPES

No.	Type	Maximum Ratings: Material[1]	Maximum Ratings: Diss. (Watts)	Maximum Ratings: V_{CEO} (Volts)	Characteristics: I_C (dc)	Characteristics: h_{FE} (Min.)	Characteristics: f_T (Typ.)	Characteristics: Noise Fig. (dB)	Other Data: Use (Typ.)	Other Data: Case Style	Other Data: Base Conn.	Other Data: Manufacturer[2]	Application
2N406	PNP	G	0.15	−18	—	34	0.65 MHz	—	Gen. Purpose	TO-1	7	R	Gen. Purpose
2N706A	NPN	S	0.3*	20	50 mA	20	400 MHz	—	rf	TO-18	8	M	rf, Switching
2N718A	NPN	S	0.5	50	150 mA	40	60 MHz	—	—	TO-18	8	R	Switching
2N1179	PNP	G	0.080*	−30	−10 mA	100	—	—	hf Amp.	TO-45	5	R	rf Mixer
2N1302	NPN	G	0.15	25	0.3A	20	—	—	Computer	TO-5	8	R	Osc., Amp.
2N1306	NPN	G	0.15	25	0.3A	60	—	—	Computer	TO-5	8	R	Osc., Amp.
2N2222	NPN	S	1.8	30	800 mA	35	250 MHz	—	Gen. Purpose	TO-18	8	M	vhf Amp., Osc.
2N2925	NPN	S	0.2*	25	100 mA	170	160 MHz	2.8	Gen. Purpose	—	1	GE	Osc., rf, i-f, af
2N3391A	NPN	S	0.2*	25	100 mA	250	160 MHz	1.9	Audio	—	1	GE	Low-noise Preamps.
2N3394	NPN	S	0.31	25	100 mA	55	—	—	Gen. Purpose	TO-92	2	M	Audio Amp.
2N3565	NPN	S	0.2	25	50 mA	150	—	—	—	TO-106	7	—	—
2N3568	NPN	S	0.3	60	500 mA	120	60 MHz	—	—	TO-105	—	—	—
2N3638	PNP	S	0.3	−25	−500 mA	100	150 MHz	—	—	TO-105	—	—	Switching
2N3663	NPN	S	0.12*	12	25 mA	20	900 MHz	4	rf	—	1	GE	vhf/uhf Osc., Amp., Mix.
2N3702	PNP	S	0.31	−25	−200 mA	60	100 MHz	—	Gen. Purpose	TO-92	2	M	vhf Osc., Amp.
2N3866	NPN	S	5	3	400 mA	5	800 MHz	—	Gen. Purpose	TO-39	8	M	uhf Amp., Osc.
2N3904	NPN	S	0.21	40	200 mA	40	300 MHz	—	Gen. Purpose	TO-92	2	M	vhf Amp., Osc.
2N3906	NPN	G	0.15	25	300 mA	60	—	—	Computer	TO-5	8	R	Osc., Amp.
2N4123	NPN	S	0.21	30	200 mA	50	250 MHz	—	Gen. Purpose	TO-92	2	M	vhf Amp., Osc.
2N4124	NPN	S	0.3	25	200 mA	120	250 MHz	5	Audio-rf	—	2	M	—
2N4126	PNP	S	0.3	−25	200 mA	120	250 MHz	4	Audio-rf	—	2	M	—
2N4275	NPN	S	0.28	15	—	18	—	—	—	—	—	—	Switching
2N4401	NPN	S	0.31*	40	600 mA	20	250 MHz	—	Gen. Purpose	TO-92	2	M	Osc., rf, i-f, af
2N4410	NPN	S	0.31*	80	250 mA	60	250 MHz	—	Gen. Purpose	TO-92	2	M	Osc., rf, i-f, af
2N4957	PNP	S	.2	30	30 mA	20	1600 MHz	2.6	rf Amp.	TO-72	9	M	rf Amp., Mix., Osc.
2N4959	PNP	S	.2	30	30 mA	20	1500 MHz	3.2	rf Amp.	TO-72	9	M	rf Amp., Mix., Osc.
2N5032	NPN	S	.2	10	20 mA	25	2000 MHz	3.0	rf Amp.	TO-72	9	M	Low-noise rf Amp.
2N5087	PNP	S	0.310*	−50	−50 mA	200	150 MHz	1	rf Amp.	TO-92	2	M	Low-noise rf Amp.
2N5089	PNP	S	0.310*	−25	−50 mA	450	175 MHz	2	rf Amp.	TO-92	2	M	Low-noise rf Amp.
2N5109	NPN	S	3.5*	40	0.4 A	70	—	3	vhf Amp.	TO-39	8	R	Wide-band Amp.
2N5179	NPN	S	0.200*	12	50 mA	25	900 MHz	4.5	rf Amp.	TO-72	9	M	uhf Amp., Osc., Mix.
2N5183	NPN	S	0.5	18	1 A	120	200 MHz	—	Gen. Purpose	TO-104	8	R	vhf Osc., Amp.
2N5222	PNP	S	0.310*	−15	−50 mA	20	450 MHz	—	rf Amp.	TO-92	18	M	rf Amp., Mix., Video i-f
2N5829	PNP	S	.2	30	30 mA	20	1600 MHz	2.3	rf Amp.	TO-72	9	M	rf Amp., Mix., Osc.
40231	NPN	S	0.5*	18	100 mA	55	60 MHz	2.8	Audio	TO-104	7	R	Preamps. and Drivers
40235	NPN	S	0.18*	35	50 mA	40	1200 MHz	3.3	rf	TO-104	9	R	vhf/uhf Amp., Osc., Mix.
HEP51	PNP	S	0.6	−25	−600 mA	80	150 MHz	—	—	TO-5	8	M	rf Amp.
HEP53	NPN	S	0.6	30	600 mA	85	200 MH	—	—	TO-5	8	M	rf Amp.
HEP56	NPN	S	0.31	20	100 mA	70	750 MHz	—	—	TO-92	18	M	uhf Osc.
MPS918	NPN	S	0.310*	15	—	20	200 MHz	6	Amp. Osc.	TO-92	2	M	uhf Amp., Osc.
MPS2926	NPN	S	0.31	18	100 mA	35	300 MHz	—	Gen. Purpose	TO-92	2	M	vhf Osc., Amp.
MPS3394	NPN	S	0.31	25	100 mA	55	—	—	Gen. Purpose	TO-92	2	M	Audio Amp.
MPS3563	NPN	S	0.310*	12	—	20	200 MHz	—	Amp. Osc.	TO-92	2	M	uhf Amp., Osc.
MPS3693	NPN	S	0.310*	45	—	40	200 MHz	4	rf Amp.	TO-92	2	M	50 MHz Amp.
MPS3694	NPN	S	0.310*	45	—	100	200 MHz	4	rf Amp.	TO-92	2	M	50 MHz Amp.
MPS3702	PNP	S	0.31	−25	−200 mA	60	100 MHz	—	Gen. Purpose	TO-92	2	M	vhf Osc., Amp.
MPS3706	NPN	S	0.310*	20	600 mA	600	100 MHz	—	af Amp.	TO-92	2	M	Audio Amp.

SMALL-SIGNAL TYPES — *Continued*

No.	Type	Maximum Ratings: Material[1]	Diss. (Watts)	V_{CEO} (Volts)	I_C (dc)	Characteristics: h_{FE} (Min.)	f_T (Typ.)	Noise Fig. (dB)	Use (Typ.)	Other Data: Case Style	Base Conn.	Manufacturer[2]	Application
MPS6514	NPN	S	.3	25	100 mA	150	480 MHz	2.0	Audio-rf	TO-92	2	M	af-rf Amp.
MPS6530	NPN	S	0.310*	40	600 mA	30	390 MHz	—	Amp.	TO-92	2	M	Complementary Amp
MPS6534	PNP	S	0.310*	−40	−600 mA	60	260 MHz	—	μhf Amp.	TO-92	2	M	Complementary Amp.
MPS6543	NPN	S	0.310*	25	—	25	750 MHz	—	Osc.	TO-92	2	M	uhf Osc.
MPS6569	NPN	S	0.310*	20	—	20	300 MHz	6	i-f Amp.	TO-92	18	M	vhf Amp., Video i-f
MPSA12	NPN	S	0.310*	20	—	35	—	—	Audio Amp.	TO-92	2	M	High-Z Pre-amp.
MPSA55	PNP	S	0.5	−60	−500 mA	50	50 MHz	—	Audio Amp.	TO-92	2	M	Audio Amp.
TIS48	NPN	—	1.2*	40	500 mA	40	500 MHz	—	rf	TO-92	3	TI	rf, Switching
TIS54	PNP	—	0.25*	−12	−80 mA	30	300 MHz	—	rf	TO-92	3	TI	rf, Switching
TIXM10	PNP	—	0.075*	−20	−30 mA	20	630 MHz	4	rf	TO-72	4	TI	rf, Preamp., vhf/uhf

LARGE-SIGNAL TYPES

No.	Type	Maximum Ratings: Material[1]	Diss. (Watts)	V_{CEO} (Volts)	I_C (dc)	Characteristics: h_{FE} (Min.)	f_T (Typ.)	Noise Fig. (dB)	Use (Typ.)	Other Data: Case Style	Base Conn.	Manufacturer[2]	Application
2N441	PNP	G	150	−40	−15 A	20	—	—	Gen. Purpose	TO-36	13	M	Switch, Amp.
2N1491	NPN	S	3.0*	30	100 mA	15	300 MHz	—	rf Amp.	TO-39	8	R	vhf Amp., Mix.
2N1970	PNP	G	170	−50	−15 A	17	—	—	Gen. Purpose	TO-36	13	M	Switch, Amp.
2N2102	NPN	S	5†	65	1 A	20	100 MHz	6	Gen. Purpose	TO-5	8	R	af, rf Amps (Linear)
2N2157	PNP	—	170†	−60	−30 A	40	100 kHz	—	af	TO-36	13	M	af, dc Amp., Switch.
2N2270	NPN	S	5†	—	1 A	10	—	6	Amp.	TO-5	9	R	Low-noise Amp.
2N2631	NPN	S	8.75†	60	1.5 A	—	200 MHz	—	rf	TO-39	8	R	Class C rf Amp., Osc.
2N2869	PNP	—	30†	−50	−10 A	50	200 kHz	—	Gen. Purpose	TO-3	11	R	af, Osc., Amp., Switch.
2N2876	NPN	S	17.5†	60	1.5 A	—	200 MHz	—	rf	TO-60	12	R	vhf Class-C Amp.
2N3053	NPN	S	5	60	700 mA	50	—	—	—	TO-5	8	R	Pwr. Switch
2N3055	NPN	S	115	60	15 A	20-70	—	—	Gen. Purpose	TO-3	11	R	Switch, Reg., Amp.
2N3119	NPN	S	4†	80	500 mA	50	450 MHz	—	Amp.	TO-5	8	R	Switch, Pulse Amp.
2N3512	NPN	—	4	35	500 mA	10	250 MHz	—	Audio-rf	—	7	R	—
2N3553	NPN	S	7†	40	1 A	10	500 MHz	—	rf	TO-39	8	M	Class A, B, C rf Mult., Amp., Osc.
2N3583	NPN	S	35†	175	2 A	10	15 MHz	—	hV Gen. Purp.	TO-66	11	R	rf, af Osc., Amp. dc Amp.
2N3632	NPN	S	23*	65	3 A	—	400 MHz	—	rf Amp.	TO-60	12	R	uhf Pwr. Amp., Osc.
2N3733	NPN	S	23*	65	3 A	—	400 MHz	—	rf Amp.	TO-60	12	R	uhf Pwr. Amp., Osc.
2N3772	NPN	S	150	60	20 A	15-60	800 kHz	—	Pwr. Amp.	TO-3	11	R	Pwr. Amp.
2N3866	NPN	S	5†	30	0.4 A	—	800 MHz	—	rf	TO-39	8	R	Class A, B, C rf Mult., Amp., Osc.
2N3924	NPN	S	7*	18	500 mA	—	350 MHz	—	rf Amp.	TO-39	8	M	uhf Pwr. Amp., Osc.
2N3948	NPN	S	1*	20	400 mA	15	700 MHz	—	rf Amp.	TO-39	8	M	uhf Pwr. Amp., Osc.
2N4012	NPN	S	11.6*	40	1.5 A	—	500 MHz	—	rf Amp.	TO-60	12	R	uhf Pwr. Amp., Osc.
2N4037	PNP	S	7†	−40	−1 A	50	60 MHz	—	Gen. Purpose	TO-5	8	R	Amp., Switching
2N4396	NPN	S	62†	60	5 A	60	4 MHz	—	Gen. Purpose	TO-3	11	R	rf, af Osc., Amp. dc Amp.
2N4427	NPN	S	3.5*	20	400 mA	—	500 MHz	—	rf Amp.	TO-39	8	R	uhf Amp.
2N5016	NPN	S	30*	65	4.5 A	—	600 MHz	—	rf Amp.	TO-60	12	R	uhf Pwr., rf Amp.
2N5070	NPN	S	70†	65	3.3 A	—	30 MHz	—	Amp.	TO-60	12	R	30 MHz Amp.
2N5071	NPN	S	70†	65	3.3 A	—	76 MHz	—	Amp.	TO-60	12	R	50 MHz Amp.
2N5470	NPN	S	3.5†	55	200 mA	—	2GHz	—	uhf Amp.	—	—	R	Microwave Osc., Amp.
2N5635	NPN	S	7.5*	35	1 A	5	—	—	rf Amp.	—	23	M	400 MHz, rf Amp.
2N5636	NPN	S	15*	35	1.5 A	5	—	—	rf Amp.	—	23	M	400 MHz, rf Amp.
2N5637	NPN	S	30*	35	3 A	5	—	—	rf Amp.	—	23	M	400 MHz rf Amp.
2N5641	NPN	S	15*	35	1 A	5	—	—	rf Amp.	—	23	M	400 MHz, rf Amp.
2N5642	NPN	S	30*	35	3 A	5	—	—	rf Amp.	—	23	M	400 MHz, rf Amp.
2N5643	NPN	S	60*	35	5 A	5	—	—	rf Amp.	—	23	M	400 MHz, rf Amp.
2N5913	NPN	S	3.5†	14	330 mA	—	900 MHz	—	uhf Amp.	TO-39	8	R	432 MHz Amp.
2N5914	NPN	S	10.7†	14	1.5 A	—	900 MHz	—	uhf Amp.	—	—	R	432 MHz Amp.
2N5915	NPN	S	10.7†	14	1.5 A	—	800 MHz	—	uhf Amp.	—	—	R	432 MHz Amp.
2N5919	NPN	S	25†	30	4.5 A	—	400 MHz	—	uhf Amp.	—	—	R	220 MHz Amp.
2N5921	NPN	S	8.3†	50	700 mA	—	2.3 GHz	—	uhf Amp.	—	—	R	Microwave Osc., Amp.
2N5941	NPN	S	80	35	6.0A	10	50 MHz	—	rf Amp.	—	25	M	30 MHz rf Amp.
2N5942	NPN	S	140	35	12 A	10	50 MHz	—	rf Amp.	—	26	M	30 MHz rf Amp.
2N5944	NPN	S	5	16	0.4A	20	470 MHz	—	uhf Amp.	—	27	M	432 MHz rf Amp.
2N5945	NPN	S	15	16	0.8 A	20	470 MHz	—	uhf Amp.	—	27	M	432 MHz rf Amp.
2N5946	NPN	S	37.5	16	2.0 A	20	470 MHz	—	uhf Amp.	—	27	M	432 MHz rf Amp.
2N5995	NPN	S	10.7	14	1.5 A	—	175 MHz	—	vhf Amp.	—	23	R	vhf rf Amp.
2N5996	NPN	S	35.7	18	5.0 A	—	175 MHz	—	vhf Amp.	—	23	R	vhf rf Amp.
2N6136	NPN	S	60	18	6.0 A	20	470 MHz	—	uhf Amp.	—	27	M	432 MHz rf Amp.
MJ480	NPN	S	87†	40	1 A	30	4 MHz	—	Gen. Purpose	TO-3	11	M	af, rf Amp., Osc.
MPS-U01	NPN	S	1.0*	30	1.5 A	70	50 MHz	—	af Amp.	—	20	M	Audio Amp.
MPS-U51	PNP	S	1.0*	−30	−1.5 A	70	50 MHz	—	Gen. Purpose	—	20	M	af Amp.

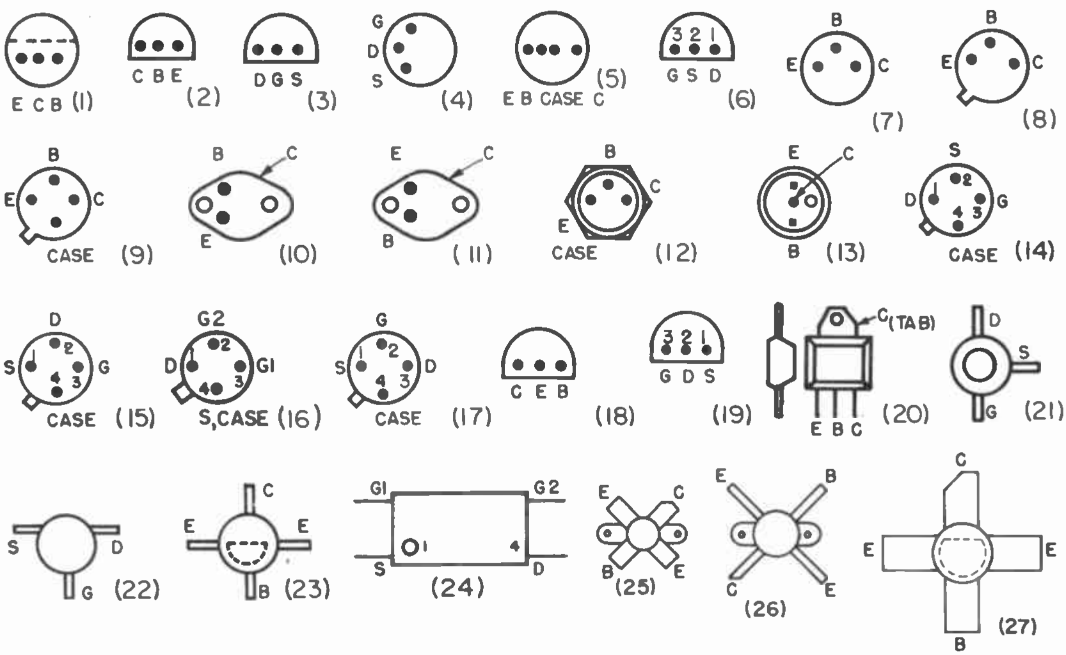
FIELD-EFFECT TRANSISTORS

No.	Type	Diss. (mW)	V_{DS}	V_{GS}	MIN. μ MHOS	C_{iss} (pF)	MAX. I_{DSS} (mA)	Top Freq. (MHz)	Case Style	Base Conn.	Manufacturer	Application
2N4416	N JFET	175	30	−6.0	4000	4	15	450	TO-72	15	M	vhf/uhf rf Amp., Mix., Osc.
2N4417	N FET	175	30	−30	4500	3.5	15	400	—	22	UC	vhf uhf Amp.
2N5460	P JFET	310	—	40	1000	5	5	—	TO-92	19	M	Gen. Purpose Audio

FIELD-EFFECT TRANSISTORS — *Continued*

No.	Type	Diss. (mW)	V_{DS}	V_{GS}	MIN. μMHOS	C_{iss} (pF)	MAX. I_{DSS} (mA)	Top Freq. (MHz)	Case Style	Base Conn.	Manu-facturer	Application
2N5461	P JFET	310	—	40	1500	5	9	—	TO-92	19	M	Gen. Purpose Audio
2N5463	P JFET	310	—	40	1000	5	5	—	TO-92	19	M	Gen. Purpose Audio
2N5465	P FET	310	—	60	2000	5	16	—	TO-92	19	M	Gen. Purpose Amp.
2N5669	N JFET	310	25	1.0	1600	4.7	4	—	TO-92	6	M	Amp. Switching
2N5670	N JFET	310	25	2.0	2500	4.7	8	—	TO-92	6	M	Amp. Switching
3N128	N IGFET	100	20	—	5000	5.8	—	200	TO-72	14	R	af, rf, Amp., Mix., Osc.
3N187	MOS n-channel Depletion type	330	20	−6+6	7000	4	8.5	300	TO-72	16	R	—
3N200	MOS n-channel Depletion type	330	20	−6+6	10,000	8.5	—	500	TO-72	16	R	uhf rf Amp.
40600	N Dual-Gate FET	400	20	−8	10,000	5.5	18	250	TO-72	16	R	uhf rf Amp.
40601	N Dual-Gate FET	400	20	−8	10,000	5.5	18	250	TO-72	16	R	vhf Mixer
40602	N Dual-Gate FET	400	20	−8	10,000	5.5	18	250	TO-72	16	R	vhf Amp.
40603	N Dual-Gate FET	400	20	−8	10,000	5.5	18	—	TO-72	16	R	rf Amp.
40604	N Dual-Gate FET	400	20	−8	10,000	5.5	18	—	TO-72	16	R	rf Mix.
40673	N Dual-Gate FET	330	20	−6	12,000	6	35	400	TO-72	16	R	rf Amp.
E300	N JFET	250	—	1	9000	5.5	30	100	—	4	SI	vhf Amp.
HEP801	N JFET	200	20	—	3000	—	9	—	TO-72	14	M	af Amp.
HEP802	N JFET	200	25	—	2000	—	20	—	TO-92	6	M	rf Amp.
MMT3823	N JFET	225	30	−30	3000	4.0	20	—	—	21	M	rf Amp., Mix.
MPF102	N JFET	200	25	−2.5	2000	4.5	20	200	TO-92	6	M	af, rf Amp., Mix., Osc.
MPF103/ 2N5457	N JFET	310	25	−25	1000	4.5	5	—	TO-92	6	M	Gen. Purpose Audio
MPF104/ 2N5458	N JFET	310	25	−25	1500	4.5	9	—	TO-92	6	M	Gen Purpose Audio
MPF105/ 2N5459	N JFET	200	25	−4.5	2000	4.5	16	100	TO-92	6	M	af, rf Amp., Mix., Osc.
MPF106/ 2N5484	N JFET	200	25	−25	2500	5	30	432	TO-92	6	M	af, rf Amp., Mix., Osc.
MPF107/ 2N5486	N JFET	310	—	−25	1000	5	20	400	TO-92	6	M	vhf-uhf rf Amp.
MPF120	N Dual-Gate MOS FET	500	25	±20	8000	4.5	18	105	—	24	M	rf Amp.
MPF121	N Dual-Gate MOS FET	500	25	±20	10,000	4.5	30	200	—	24	M	rf Amp.
MPF122	N Dual-Gate MOS FET	500	25	±20	8000	4.5	20	200	—	24	M	rf Mix.

* = Ambient Temp. of 25°C (No heat sink). † = Case Temp. of 25°C (with heat sink).
[1] S = Silicon. G = Germanium. [2] GE = General Electric. M = Motorola. R = RCA. SI = Siliconix. TI = Texas Instuments. UC = Union Carbide.



The leads are marked C - collector, B - base, E - emitter, G - gate, D - drain, and S - source.

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 – millihenry
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})

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