All New Projects for QRPers

W1FB's QRP NOTEBOOK

W1FB 75-M RECEIVER

ANT PEAK

USB

PWR LSB

PHONES AUDIO
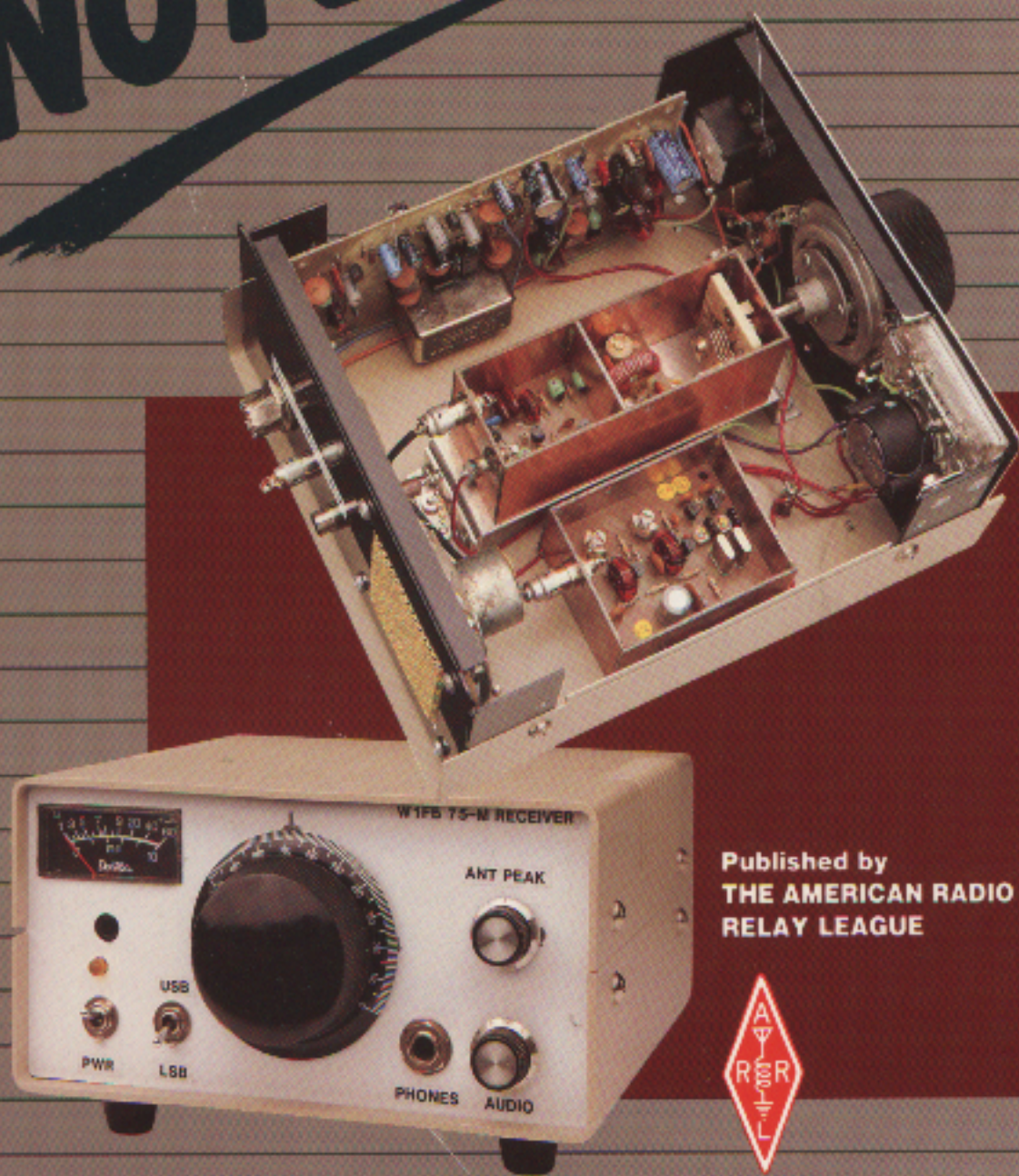
By Doug DeMaw, W1FB

W1FB's QRP NOTEBOOK

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Foreword

ARRL and Doug DeMaw have encouraged QRP operation and QRP operators for many years. Together, the League and DeMaw have published many articles and books of interest to QRP enthusiasts. In July of 1991, the Board of Directors of ARRL voted unanimously to recognize and congratulate Doug DeMaw, W1FB, for his many contributions to Amateur Radio. A brief history of Doug's career is printed on the back cover of this book.

Many of you, no doubt, purchased the first edition of the QRP Notebook. That book was the first in what was to become a series of Notebooks prepared by Doug at his home in Michigan. We are very pleased to publish this second, completely revised, edition.

Whether you are new to QRP or an old hand at it, you should find circuits and ideas of interest in this edition. In reading the foreword to the first edition, I've decided that the final two paragraphs, repeated below, are as true today as they were then.

Experimentation and low-power operating go hand in hand. Construction of a complete modern transceiver is a major undertaking, but some of the circuits in this book can be put together in an evening or a weekend from a few dollars' worth of parts. Once built, the equipment can be tested and improved as your understanding and skill grow. Many of the simpler circuits can be used later as parts of more complex projects.

We hope that this book will encourage you to pick up the soldering iron and give one of the circuits a try. Experience firsthand the thrill of a contact with equipment you built.

<div align="right">

David Sumner, K1ZZ
Executive Vice President

</div>

Newington, Connecticut
September 1991

Please note: The prices for parts and other items mentioned in this book were accurate when the book was published. Please check with the supplier before sending money, as all prices are subject to change.

Errata

Page	Fig No.	Description
55	3-15	L1 should be labeled L2
126	4-13	Q5 and Q6 are shown backward; the flat side should be to the right
		C11 should be labeled C13

Table of Contents

The Amateur's Code

The Radio Amateur is:

CONSIDERATE...never knowingly operates in such a way as to lessen the pleasure of others.

LOYAL...offers loyalty, encouragement and support to other amateurs, local clubs, and the American Radio Relay League, through which Amateur Radio in the United States is represented nationally and internationally.

PROGRESSIVE...with knowledge abreast of science, a well-built and efficient station and operation above reproach.

FRIENDLY...slow and patient operating when requested; friendly advice and counsel to the beginner; kindly assistance, cooperation and consideration for the interests of others. These are the hallmarks of the amateur spirit.

BALANCED...radio is an avocation, never interfering with duties owed to family, job, school, or community.

PATRIOTIC...station and skill always ready for service to country and community.

—The original Amateur's Code was written by Paul M. Segal, W9EEA, in 1928.—

US Customary—Metric Conversion Factors

International System of Units (SI)—Metric Units

Prefix	Symbol			Multiplication Factor
exa	E	10^{18}	=	1 000 000 000 000 000 000
peta	P	10^{15}	=	1 000 000 000 000 000
tera	T	10^{12}	=	1 000 000 000 000
giga	G	10^{9}	=	1 000 000 000
mega	M	10^{6}	=	1 000 000
kilo	k	10^{3}	=	1 000
hecto	h	10^{2}	=	100
deca	da	10^{1}	=	10
(unit)		10^{0}	=	1
deci	d	10^{-1}	=	0.1
centi	c	10^{-2}	=	0.01
milli	m	10^{-3}	=	0.001
micro	μ	10^{-6}	=	0.000001
nano	n	10^{-9}	=	0.000000001
pico	p	10^{-12}	=	0.000000000001
femto	f	10^{-15}	=	0.000000000000001
atto	a	10^{-18}	=	0.000000000000000001

Linear
1 metre (m) = 100 centimetres (cm) = 1000 millimetres (mm)

Area
$1 \ m^2 = 1 \times 10^4 \ cm^2 = 1 \times 10^6 \ mm^2$

Volume
$1 \ m^3 = 1 \times 10^6 \ cm^3 = 1 \times 10^9 \ mm^3$
1 litre (l) = 1000 cm^3 = 1 × 10^6 mm^3

Mass
1 kilogram (kg) = 1 000 grams (g)
(Approximately the mass of 1 litre of water)
1 metric ton (or tonne) = 1 000 kg

US Customary Units

Linear Units
12 inches (in) = 1 foot (ft)
36 inches = 3 feet = 1 yard (yd)
1 rod = 5½ yards = 16½ feet
1 statute mile = 1 760 yards = 5 280 feet
1 nautical mile = 6 076.11549 feet

Area
$1 \ ft^2 = 144 \ in^2$
$1 \ yd^2 = 9 \ ft^2 = 1 \ 296 \ in^2$
$1 \ rod^2 = 30¼ \ yd^2$
$1 \ acre = 4840 \ yd^2 = 43 \ 560 \ ft^2$
$1 \ acre = 160 \ rod^2$
$1 \ mile^2 = 640 \ acres$

Volume
$1 \ ft^3 = 1 \ 728 \ in^3$
$1 \ yd^3 = 27 \ ft^3$

Liquid Volume Measure
1 fluid ounce (fl oz) = 8 fluidrams = 1.804 in^3
1 pint (pt) = 16 fl oz
1 quart (qt) = 2 pt = 32 fl oz = 57¾ in^3
1 gallon (gal) = 4 qt = 231 in^3
1 barrel = 31½ gal

Dry Volume Measure
1 quart (qt) = 2 pints (pt) = 67.2 in^3
1 peck = 8 qt
1 bushel = 4 pecks = 2 150.42 in^3

Avoirdupois Weight
1 dram (dr) = 27.343 grains (gr) or (gr a)
1 ounce (oz) = 437.5 gr
1 pound (lb) = 16 oz = 7 000 gr
1 short ton = 2 000 lb, 1 long ton = 2 240 lb

Troy Weight
1 grain troy (gr t) = 1 grain avoirdupois
1 pennyweight (dwt) or (pwt) = 24 gr t
1 ounce troy (oz t) = 480 grains
1 lb t = 12 oz t = 5 760 grains

Apothecaries' Weight
1 grain apothecaries' (gr ap) = 1 gr t = 1 gr a
1 dram ap (dr ap) = 60 gr
1 oz ap = 1 oz t = 8 dr ap = 480 gr
1 lb ap = 1 lb t = 12 oz ap = 5 760 gr

Multiply ⟶

Metric Unit = Conversion Factor × US Customary Unit

⟵ Divide
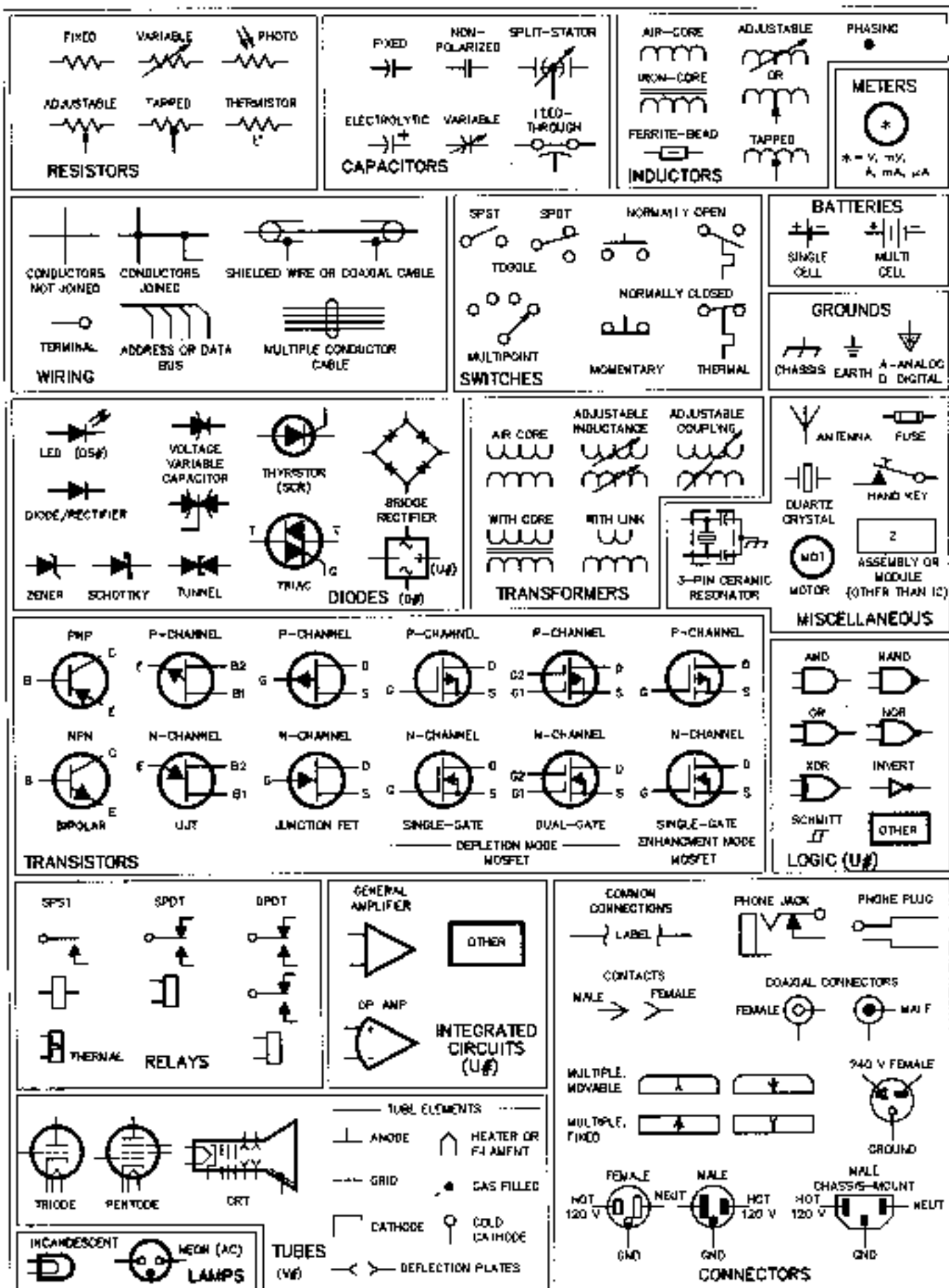
Metric Unit ÷ Conversion Factor = US Customary Unit

Metric Unit	Conversion Factor	× US Unit	Metric Unit	Conversion Factor	× US Unit
(Length)			(Volume)		
mm	25.4	inch	mm^3	16387.064	in^3
cm	2.54	inch	cm^3	16.387	in^3
cm	30.48	foot	m^3	0.028316	ft^3
m	0.3048	foot	m^3	0.764555	yd^3
m	0.9144	yard	ml	16.387	in^3
km	1.609	mile	ml	29.57	fl oz
km	1.852	nautical mile	ml	473	pint
(Area)			ml	946.333	quart
mm^2	645.16	$inch^2$	l	28.32	ft^3
cm^2	6.4516	in^2	l	0.9463	quart
cm^2	929.03	ft^2	l	3.785	gallon
m^2	0.0929	ft^2	l	1.101	dry quart
cm^2	8361.3	yd^2	l	8.809	peck
m^2	0.83613	yd^2	l	35.238	bushel
m^2	4047	acre	(Mass)	(Troy Weight)	
km^2	2.59	mi^2	g	31.103	oz t
(Mass)	(Avoirdupois Weight)		g	373.248	lb t
grams	0.0648	grains	(Mass)	(Apothecaries' Weight)	
g	28.349	oz	g	3.387	dr ap
g	453.59	lb	g	31.103	oz ap
kg	0.45359	lb	g	373.248	lb ap
tonne	0.907	short ton			
tonne	1.016	long ton			

Symbols Used in Schematic Diagrams

RESISTORS

FIXED | VARIABLE | PHOTO

ADJUSTABLE | TAPPED | THERMISTOR

CAPACITORS

FIXED | NON-POLARIZED | SPLIT-STATOR

ELECTROLYTIC | VARIABLE | FEED-THROUGH

INDUCTORS

AIR-CORE | ADJUSTABLE | PHASING

IRON-CORE | OR

FERRITE-BEAD | TAPPED

METERS

* = V, mV, A, mA, μA

WIRING

CONDUCTORS NOT JOINED | CONDUCTORS JOINED | SHIELDED WIRE OR COAXIAL CABLE

TERMINAL | ADDRESS OR DATA BUS | MULTIPLE CONDUCTOR CABLE

SWITCHES

SPST | SPDT | NORMALLY OPEN

TOGGLE

NORMALLY CLOSED

MULTIPOINT | MOMENTARY | THERMAL

BATTERIES

SINGLE CELL | MULTI CELL

GROUNDS

CHASSIS | EARTH | A - ANALOG D - DIGITAL

DIODES (D#)

LED (DS#) | VOLTAGE VARIABLE CAPACITOR | THYRISTOR (SCR) | BRIDGE RECTIFIER

DIODE/RECTIFIER

ZENER | SCHOTTKY | TUNNEL | TRIAC

TRANSFORMERS

AIR CORE | ADJUSTABLE INDUCTANCE | ADJUSTABLE COUPLING

WITH CORE | WITH LINK

MISCELLANEOUS

ANTENNA | FUSE

QUARTZ CRYSTAL | HAND KEY

3-PIN CERAMIC RESONATOR | MOTOR (M#) | ASSEMBLY OR MODULE (OTHER THAN IC)

2

TRANSISTORS

PNP | P-CHANNEL | P-CHANNEL | P-CHANNEL | P-CHANNEL | P-CHANNEL

NPN | N-CHANNEL | N-CHANNEL | N-CHANNEL | N-CHANNEL | N-CHANNEL

BIPOLAR | UJT | JUNCTION FET | SINGLE-GATE | DUAL-GATE | SINGLE-GATE ENHANCEMENT MODE MOSFET

DEFLETION MODE MOSFET

LOGIC (U#)

AND | NAND

OR | NOR

XOR | INVERT

SCHMITT | OTHER

RELAYS

SPST | SPDT | DPDT

THERMAL

INTEGRATED CIRCUITS (U#)

GENERAL AMPLIFIER | OTHER

OP AMP

CONNECTORS

COMMON CONNECTIONS | LABEL | PHONE JACK | PHONE PLUG

CONTACTS

MALE | FEMALE

COAXIAL CONNECTORS

FEMALE | MALE

MULTIPLE, MOVABLE

MULTIPLE, FIXED

240 V FEMALE

GROUND

FEMALE | MALE | MALE CHASSIS-MOUNT

HOT 120 V | NEUT | HOT 120 V | HOT 120 V | NEUT

GND | GND | GND

TUBES (V#)

TUBE ELEMENTS

ANODE | HEATER OR FILAMENT

GRID | GAS FILLED

CATHODE | COLD CATHODE

DEFLECTION PLATES

TRIODE | PENTODE | CRT

LAMPS

INCANDESCENT | NEON (AC)

AUTHOR MESSAGE AND ACKNOWLEDGEMENTS

As I prepare for my retirement from article and book writing I wish to enclose these thoughts with what may be my last effort at sharing with you the experiences and circuits that have made my QRP adventures so memorable. If I were asked to offer a single statement of encouragement to the body of amateurs, worldwide, it would be something like this: *Make an effort to avoid technical complacency. Don't rely on commercial equipment when assembling your amateur station. Experience the thrill of building transmitters and receivers, and gain valuable experience in the process. Half of the fun associated with our grand pastime is based on communicating by radio with equipment we have built.*

During my 41 years as a licensed amateur I have built hundreds of transmitters, receivers and allied equipment. I confess that I did not always understand some of the circuits I built, but along with the frustration of making them percolate came a learning experience that was more than worth the agony of struggling to obtain the expected performance. Later, after going to school and working as an electronics engineer, it became easy to design amateur gear that worked well. But, I shall always remember with joy those early days when most circuits seemed like magic, even though some of them were never honed to perfection.

It has been satisfying to present beginner types of articles, series and books in the pages of ARRL publications. Many letters have been received from young engineers who attribute their careers to the inspiration provided by League publications and my work. As a young person, you may have a career in electronics awaiting you, if you are willing to study the ARRL technical literature and construct equipment that can be used in your lab or ham shack. Amateur Radio is a resource that is yet to be tapped to its fullest measure by many licensed hams.

Certainly, my career would not have been successful without my long-term association with The ARRL and the many amateurs and engineers I have known. I appreciate especially the many long, wee-hour gab sessions I had with such people as Wes Hayward (W7ZOI), Byron Goodman (W1DX), Al Helfrick (K2BLA), Roy Lewallen (W7EL), Harold Johnson (W4ZCB) and Doug Blakeslee (N1RM), during which we discussed circuits and techniques without respite. My overall knowledge has been greatly enhanced by exchanging ideas with these men. I wish to thank the many applications engineers who offered sage advice over the years. Notable among them are Helge Granberg (K7ES), Roy Hejhall (K7QWR), Ed Oxner (KB6QJ) and Mike Metcalf (W7UDN). No small measure of credit for my success goes to my enduring wife Jean, W1CKK, who encouraged me as I did research and wrote technical papers and books.

Vy 73, Doug DeMaw, W1FB (ex W8HHS)

INTRODUCTION TO QRP

What motivates the low-power (QRP) amateur operator? This question does not have a simple answer. A vast number of QRPers are "turned on" by the relative simplicity of most home-made QRP equipment. Simple gear is not only easy to construct and operate, but it is fairly inexpensive to build. This has special appeal to those hams who lack technical backgrounds and have yet to develop their skills. Furthermore, many of published QRP circuits may be assembled on a PC (printed circuit) board that is available by mail from one or more PC-board vendors. The parts-placement guide for a given project is generally published in the related article. Guesswork is thus eliminated for the most part.

Other amateurs have the ability to design their own circuits. QRP equipment offers a short-term exercise in the workshop because many of the projects are simple. This enables an experimenter to try new circuits in an evening or within a couple of days. He can try new ideas and obtain fast results. He may continue to work with his new circuit until it is perfected, at which time a final model can be built, housed in a cabinet and used in his station.

Other QRPers are captured by the nostalgia that takes them back to the early days of Amateur Radio, when hams, through necessity, used only a few watts of RF power for communicating. In other words, they had to do things the hard way. Each successful QSO was logged as an achievement! Pride accompanied home-made gear and the ability to be heard at great distances.

I have spoken to a number of QRPers who expressed boredom from using store-bought, high-power (QRO) transmitters. Worldwide QSOs via brute-force methods were no longer stimulating to them. They found relief when giving up the mayhem of DX pileups, taught nerves from battling phone-band QRM and household power bills that were inflated from the use of high-power transmitters. QRP offered a new and exciting challenge as they worked for their WAS (worked all states) or DXCC (100 confirmed countries) with less than 5 watts of RF power. As past contesters they were able to satisfy their competitive urges by taking part in the many QRP contests that are offered each year.

Still another advantage associated with QRP operation is the relative freedom from TVI and RFI. The fundamental overloading of TV receiver front-ends is seldom a factor at low levels of RF power. Also, the magnitude of the transmitter harmonics is very low when we use QRP transitters. This also minimizes potential interference to entertainment devices such as TV sets, FM receivers, VCRs and telephones.

In general, QRPers are a special breed of friendly operators. You have much to gain by getting involved in this growing movement.

Home-Made versus Commercial Equipment

The easy way to QRP operation is via store-bought equipment. Most 100-W or greater transceivers can be operated at 5 watts or less by simply reducing the drive to the final stage of the transmitter. This may be done by turning the DRIVE control to the left, or by detuning the PRESELECTOR circuit in some rigs. Of course, you will need an accurate RF wattmeter to ensure that the output power is less than 5 watts if you wish to use a high-power transceiver for QRP work. There can be no satisfaction if we cheat on the power amount! It would not be unlike going deer hunting, buying a freshly killed deer from someone else, then boasting to your friends about the deer you killed. Such things happen!

Commercial QRP equipment is available from Ten-Tec Corp. in TN. The Ten-Tec Argonaut is designed specifically for multiband QRP work. The Heath HW-9 transceiver falls into the same class, but the latter unit is for CW only, whereas the "Argo" enables you to use CW and SSB.

At least one mail-order vendor is dedicated to selling only QRP components and parts. The transmitters are for CW operation in the HF bands. The owner is KE8KL and his business name is Oak Hills Research (Big Rapids, MI). QRP-related kits are available also from A&A Engineering (W6UCM) in Anaheim, CA.

You will feel greater pride if you build your own QRP gear. Half the fun of QRPing comes from using homebrew transmitters, receivers, antennas and associated equipment. If you're a penny pincher like me, you will appreciate the monetary savings that come with parts scrounging and apparatus you built from scratch. But perhaps more importantly, you will "learn by doing." This new knowledge will help you to upgrade your license class and it may help lead you into a career in electronics.

QRP Communication Distance

If you have not tried your hand at QRPing you are probably curious about the communications distances that are possible with low power. This depends greatly upon the antenna you use, your station location and your operating technique. Let's compare a 100-watt signal to a 5-watt one. For example, if the other station gives you a signal report of 20 dB over S9 when you are using 100 watts, he should give you a report of 7 dB over S9 when you QRP to 5 watts. Not a tremendous loss at all! Your signal should remain Q5 at the 5-watt power level. Now, if you reduce your transmitter power to 1 watt, your report should be on the order of S9. You have lost 20 dB when reducing the power from 100 watts to 1 watt!

Many QRPers have earned the WAS award. Others have garnered the ARRL DXCC award. In fact, a favorite pastime of some QRP operators is to see how many miles they can communicate per watt or milli-watt. It is not unusual to find a QRP enthusiast using only 50 or 100 mW for contacts out to 500-1000 miles.

Antennas for QRP Work

We need to recognize at the beginning a need for effective antennas
when we are using low power. I have met QRPers who labored under
the misbelief that a simple antenna, such as an end-fed wire of
random length, close to the ground, is a proper accompanyment for
a QRP rig. Avoid this situation if you're unwilling to inflict
misery upon yourself! Communicating over great distances is tough
enough at the 5-watt or lower level, and a poor antenna worsens
the odds. On the other hand, I don't wish to suggest that your
kit of goodies contain a portable tower and a triband Yagi. A full-
size dipole, inverted-L, full-wave loop or quarter-wave vertical
(with at least four radial wires) will yield many solid QSOs. You
should strive to erect these simple antennas as high above ground
as practicable. They should installed away from nearby conductive
objects, such as power lines, telephone lines and metal buildings.
I have always had good luck when using inverted-V dipoles for QRP
work. They require but one support point, and offer good perform-
ance at modest heights above ground.

The ARRL publishes a number of good books that describe simple wire
antennas that work well. **The ARRL Antenna Book** heads the list. You
will also want a copy of **W1FB's Antenna Notebook**. Both books contain
useful information about antennas you can build easily and at modest
cost. The latter publication is written in plain language that
beginners can understand and there is no need for complex algebra.

You may want to construct portable antennas for QRP junkets, such
as camping or hiking trips. There is nothing wrong with using wire
of small gauge (no. 18 through 26) in order to reduce the weight
and bulk of the antenna. Tiny, lightweight insulators may be used
with this type of antenna to further minimize bulk. I frequently
use heavy-duty rubber bands or monofilament fish line as insulators
for temporary antennas. A spin casting fishing rod with a 1/4-ounce
lead weight is excellent for getting an antenna-support line over
a tree. Miniature coaxial cable (RG-174) is suitable for portable
HF antennas if the feeder is not more than 50 feet long at the
upper end of the HF spectrum (17 through 10 meters) or no longer
than 100 feet for MF and the low end of the HF spectrum, from 160
through 30 meters. RG-174 is a very lossy line, but it is small
and lightweight. This makes it worth considering for use with
a dipole that you wish to carry in your hip pocket or tote bag.
RG-58 coax cable is a better choice if you can manage the extra
weight and bulk.

The "bottom line" here is to use the best, most efficient antenna
system you can manage for QRP work. You need not be a masochist
to enjoy low-power operation.

Operating Technique

You need to approach QRP operation with a different philosophy
than when using higher power. Brute force success, such as we may

experience with full legal amateur power, must be forgotten at the 5-watt power level. Our signals do not overwhelm the other operator's receiver. In a like manner, our peanut-whistle signal does not stand out in a DX pileup or amid QRN or QRM. Patience is a virtue we must all develop in order to succeed. At least a modicum of operating skill is also a requirement for QRPers. To this end we need to develop timing for calling DX stations amid a pileup. Calling slightly off frequency (out at the fringes of a pileup) has been effective for me when chasing DX.

When calling CQ it is important to do it on a clear frequency, well away (when possible) from QSOs that are nearby in frequency. It may require several CQs before you obtain a response, but don't be discouraged. I have experienced this disappointment when using 500-600 watts of transmitter output power. It is a more common event when I operate with less than 5 watts.

I suggest that you avoid responding to the CQ of a station which has a weak signal. The operator may be using full legal power and an excellent antenna. But, band conditions at the time may be very poor, and hence his weak signal. In a reciprocal sense, your signal, under these bad conditions, may be inaudible at his end of the circuit. Try to answer stations that have loud or moderately loud signals. Your score will improve markedly if you adopt this technique.

QRP Signal Quality

A chirpy or buzz-laden CW signal is hard to copy. Be sure you have a clean note. This will aid your batting average. SSB signals that have distorted audio are similarly hard to copy. Strive to have high-quality audio. Needless to say, a transmitter that suffers from frequency drift is a force that works against the QRPer. The other station does not enjoy "tracking" a drifter. When we have drift in a QRP transmitter, the other operator may lose our signal quickly as if wanders into a mess of QRM on a nearby frequency.

Other Sources of QRP Information

Adrian Weise, WØRSP, has written two good books on the subject of QRP. You may want to contact him and obtain these books. The ARRL has published an anthology of QRP articles from past issues of QST (QRP Classics). I recommend all of the above books for the dedicated QRPer. Another ARRL publication, **Solid State Design for the Radio Amateur**, may be considered a bible for QRP builders. It has numerous QRP receiver, transmitter, transceiver and related circuits that you can build. The book treats theory in simple terms for the beginning technical experimenter.

QRP Socities

The following pages contain letters from key amateur organizations that represent QRPers. I strongly recommend that if you are not a member of The ARRL, you should join it to help support the cause of Amateur Radio. You will also receive the monthly journal

G-QRP CLUB
DEVOTED TO LOW POWER RADIO COMMUNICATION

G3RJV

Rev. George Dobbs
St. Aiden's Vicarage
498 Manchester Road
ROCHDALE,
Lancs.   OL11 3HE
ENGLAND
Phone: 0706 31812

Dear Fellow QRPers:

It is with great pleasure that I write a few words to go into the second edition of **The W1FB QRP Notebook**. I have commended the first edition of the book to many radio amateurs in the UK. For many years we have appreciated and enjoyed the work of Doug DeMaw, W1FB.

Traditionally, in the UK, QRP interest has been closely linked with the building of one's own equipment. Most G-QRPers have stations which are partly or wholly built by themselves. Listen for European QRP stations around 3560, 7030, 10,106, 14,060, 21,060 and 28,060 kHz on CW. Some SSB activity takes place on 3690, 7090, 14,285 and 28,885 kHz.

The G-QRP Club was founded in 1974 with some 30 members. In 1989 we enrolled our 5000th member. Although the club is UK based, it has members all over the world, including several hundred in the USA. The club exists to promote the interest and growth in low power (5 watts or less) radio communication and to encourage the building and operation of home-made equipment. The club has good links with other QRP groups worldwide, and especially with the QRP ARCI in the USA. Our chief vehicle is a quarterly magazine called **SPRAT**. **SPRAT** has always been a constructor's magazine and has a policy   that requires 2/3 of its pages be devoted to practical circuits the readers can build.

If you wish to know more about the G-QRP Club and its work, write to me and I will send you a sample copy of **SPRAT** and details of how to join our club.

On behalf of the QRPers in the UK, may I wish you pleasure in building circuits from this book, and success in all of your construction projects.

73 - RPE CU QRP,

George Dobbs, G3RJV

The ARRL, Inc.
225 Main Street
Newington, CT 0611

Phone: (203) 666-1541
Fax : (203) 665-7531

Dear QRP Enthusiast,

You belong to a group of folks that understands that dB over S9
are not necessary when noise and interference are absent. You
may very well build your own equipment. If not, you probably
make modifications to your gear. Chances are that you place a
high value on operating skill and finesse. You may enjoy
competitve activities like contests, chasing DX and earning
awards. You are involved in an interesting and satisfying aspect
of Amateur Radio, and it's great!

The American Radio Relay League encourages QRP activity. We do
that by publishing books (like this one), and articles in the
official journal, QST. The popular Worked All States award can
be endorsed for QRP. The IARU award, Worked All Continents, is
also endorsable for QRP. For more than 50 years ARRL has
encouraged QRP operation in the largest operating activity in the
world--ARRL Field Day. For several years the ARRL International
DX Contest has featured a separate category, complete with
certificates and plaques for QRP operators. The League's VHF
contests benefit greatly from QRP portable operations. Many
other ARRL contests feature QRP competition categories.

As long as hams enjoy and participate in QRP, ARRL will continue
to support and encourage that interest.

With very 73,

Charles L. Hutchinson, K8CH
Membership Services Manager

QRP ARCI
Paula Franke, President
P.O. Box 873
Beecher, IL 60401

*It has been said and proven many times that "Power is no substitute
for skill." This phrase struck home during a 1988 DXpedition. Larry
Maso, NU4B, and I operated KP2/QRP in a campground on the island
of St. John for one week in 1988. One morning a voice hailed me
through the tent. It belonged to a fellow camper who was lured
by the CW he was hearing. My visitor was a "big gun" DX type who
regaled me with a laundry list of his Amateur Radio accomplish-
ments, including DXCC Honor Roll. He commented wistfully that there
were no new frontiers to explore. He had lost the thrill for the
chase.*

*This won't happen to the average QRPer. We practice ham radio "limbo"
by experimenting with power to see "how low we can go!" Each QRP
operation includes the thrill of the chase. QRP ARCI supports its
members in this pursuit. Members need not operate QRP 100% of the
time. However, the club supports the internationally accepted power
level of 5 W output for CW and 10 W PEP for SSB, respective to
contests and awards.*

*Various contests, ranging from four-hour sprints to weekend marathons,
are sponsored nearly each month. Also, a QRP net can be found almost
daily. So, if you're in the mood for a QRP QSO, any time of the
day or night, check the QRP ARCI frequencies: CW -- 3560, 7040,
14,060, 21,060 and 28,060 kHz. SSB -- 3985, 7285, 14,285 and 28,-
885 kHz. Novices should try 3710, 7110, 21,110 and 28,110 kHz.*

*QRP ARCI has awards for WAS, WAC and DXCC, plus WAS-Milliwatt.
A popular award is our XM/W (1000 mi. per watt). The Great Circle
distance is divided by the QRP power and must exceed 1000 mi. per
watt.*

*Membership is $12 U.S. ($14 foreign). Members are assigned a QRP
ARCI number and receive four yearly issues (quarterly) of the club
journal (The QRP Quarterly). The journal contains a forum, circuits
and items of QRP interest.*

*QRP ARCI vigorously promotes and defends the interests of QRP oper-
ators worldwide. For more information send a large s.a.s.e. to
Publicity Chairman Mike Bryce, WB8VGE, 2225 Mayflower NW, Massillon,
OH 44647.*

Hope to meet you on the air soon!

Paula Franke, WB9TBU

of the League, **QST**. It frequently contains QRP projects that you can build, and it keeps you abreast of changes in FCC rules, new technical developments and operating news of interest.

SPRAT is the QRP journal of the G-QRP Club. This fine organization has members from many European countries and the USA. The magazine features all manner of simple QRP circuits, along with news of the growing QRP movement abroad. You will benefit from membership in this club.

QRP ARCI (QRP Amateur Radio Club International) publishes a journal called the **QRP Quarterly**. This interesting magazine features numerous circuits that are easy to construct, along with plenty of news about QRP activities worldwide. The QRP ARCI group meets yearly at the Dayton Hamvention in the QRP hospitality suite. The club also has a booth in the commercial exhibit area at HARA arena. Representatives from QRP groups abroad, such as from the G-QRP Club, are usually present at the Dayton Hamvention. I recommend that you join QRP ARCI to keep informed and to help support the movement.

Still another QRP organization is the Michigan QRP Club. You may obtain, as a member, the monthly journal entitled **The Five Watter**. Contact Tom Root, WB8UUJ, for membership information.

The foregoing QRP organizations sponsor QRP operating contests from time to time. You can sharpen your operating skills and meet QRPers from all corners of the world if you participate in the various contests. Information about the times and rules for these events are published in the journals of the QRP clubs.

Summary Remarks

The QRP facet of Amateur Radio offers one of the most rewarding challenges in our pastime. Owing to the relative simplicity and low cost of QRP gear, the opportunity exists for hams from all walks of life to grasp the fundamentals of electronics through experimenting. **W1FB's Design Notebook** (The ARRL, Inc.) is recommended reading if you wish to further your technical knowledge. It is another plain-language book that offers basic theory along with numerous practical circuits that appeal to QRPers.

The QRP workshop need not contain complicated laboratory test gear. A VOM (volt-ohmmeter), some common hand tools and soldering equipment will allow you to get started in the homebrewing craft. A 50-MHz oscilloscope and a signal generator can enhance your ability to test and troubleshoot circuits, but these items are by no means essential to your success as a builder.

It has been my observation that most QRPers are gentle souls that are willing to help beginners. Their on-the-air manners are a cut above what we find on some bands and frequencies in the HF spectrum. I find this low-key method of operating to be theraputic after the many years of stress and mayhem I endured while chasing DX and operating contests in "high power alley." I want to share these pleasures with you.

The first edition of this book was written as an experiment in 1986. I did not know how well a QRP book would be accepted, but it turned out to be a popular publication. Now, in 1990, it is time to update and improve the book. All of the text is new, and most of the projects are new. This second edition has been expanded greatly in order to provide you with more information about this fascinating hobby.

CHAPTER 2

QRP CONSTRUCTION METHODS

No piece of amateur equipment needs to have a commercial look. Many a would-be builder of ham gear has avoided contact with workshop tools, simply because he felt incapable of producing a classy unit that would elicit raves from his ham colleagues. It should go without saying that circuit performance is far more important than project appearance. Imparting the commercial look is generally beyond the craft of most hams, and the shop equipment needed for so lofty an objective would not be cost-effective. Certainly, we should endeavor to turn out the best looking project possible, but we need not be ashamed if it qualifies as an Ugly Duckling!

The word "ugly" brings to mind an expression that was coined by Wes Hayward, W7ZOI, in a QST article that described one of his QRP transmitters. He used point-to-point circuit wiring on a scrap of PC board. He referred to this method as "ugly construction." His finished work looked quite orderly to me. It was far from ugly, even though an etched PC pattern was not contained on the board stock.

You will hear about "dead bug" construction also. This entails the placement of transistors and ICs on a chassis or piece of circuit board with their leads or pins pointing upward (like a dead insect). Various components are soldered in mid air, so to speak, to the pins and leads of the transistors and ICs. Certainly, this is a quick way to test a circuit. An etched circuit board can be designed and made after the circuit is perfected. It is akin to putting the cart before the horse when we etch a PC board for an untried circuit! It represents a waste of time and money more often than not.

Early day amateurs built most of their equipment on wooden bases called "breadboards." This name became common because the wooden chassis resembled the boards that were used when slicing bread. The ends from orange crates provided many breadboards for my projects. Wooden panels were common also, as were panels made from Masonite when that product was developed. It was rather nice to be able to see the "works" of a receiver or transmitter after it was built in this manner. Cabinets were seldom used, which caused our projects to qualify as dust catchers! There is no reason why we can't build many of our solid-state projects on wooden bases. Wood is inexpensive and readily available.

Universal Breadboard

We can avoid the tedium of laying out and etching custom PC boards for many of our projects. If we adopt a universal type of PC breadboard it is possible to enjoy rapid completion of the circuit wiring, and the finished product will have a neat appearance. An example

of a workable board pattern is provided in Fig 2-1. There are many isolated pads available for point-to-point wiring. A site exists for a 16-pin IC as well. This part of the board may be used also for two 8-pin ICs or a 14-pin unit. Etched and drilled boards of this design, plus others with more IC sites, are available from FAR Circuits, 18N460 Field Court, Dundee, IL 60118.

You may choose to make your own universal breadboard by using the scale pattern in Fig 2-1. The foil that extends along the outer edge of the pattern, plus its branch foils, is used as the circuit ground. The long, single isolated conductor at the center of the pattern serves as the Vcc bus. I have built numerous QRP circuits on boards made from this pattern. They were neat and durable enough to qualify as permanent units after the circuits were refined and in proper working order.
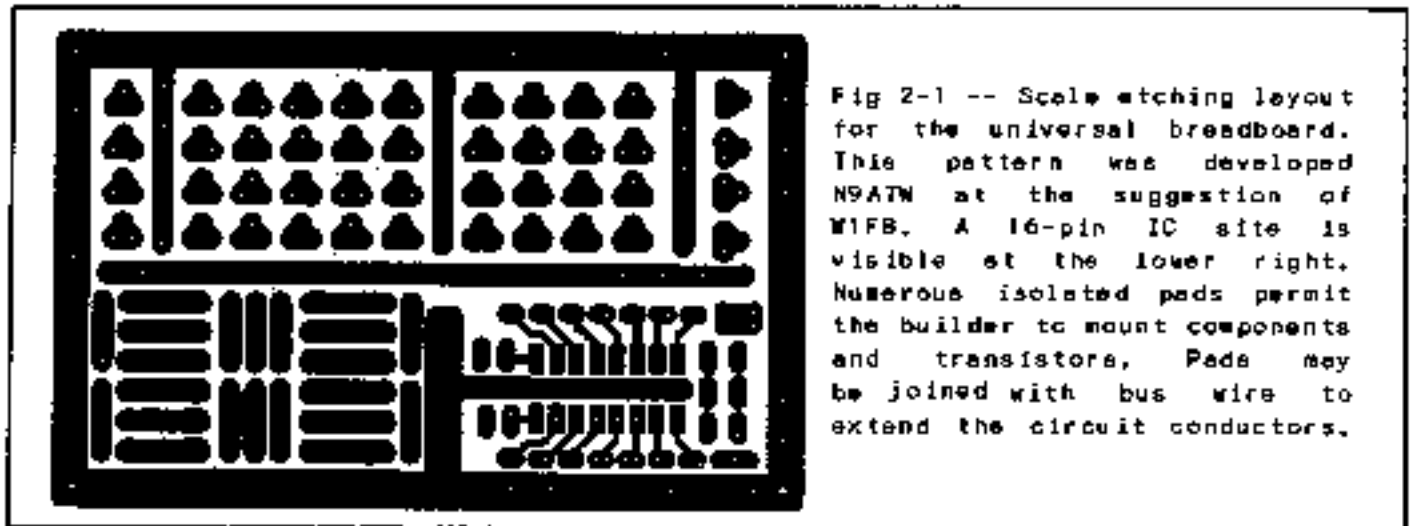
Fig 2-1 -- Scale etching layout for the universal breadboard. This pattern was developed N9ATW at the suggestion of W1FB. A 16-pin IC site is visible at the lower right. Numerous isolated pads permit the builder to mount components and transistors. Pads may be joined with bus wire to extend the circuit conductors.

Quick Circuit Boards

I have constructed one-shot circuit boards quickly by using a hobby motor with a metal routing bit or small cone-shaped abrasive bit to grind away the unwanted copper on a piece of unused PC board. A metal straight edge can be used as a guide for the cutting bit, thereby ensuring straight conductors. Various strips and isolated pads may be made in this manner if you have a steady hand.

You can construct a universal type of PC breadboard by using a hacksaw to cut the copper surface when forming numerous square isolated pads. The board material must be placed on a completely flat surface when making the cuts. Warped PC board stock does not lend itself to this process.

Another type of breadboard may be fashioned from an unused piece of PC-board stock by using epoxy cement or hot-melt glue to affix small squares of PC-board material to the main board. This creates isolated pads to which components can be soldered. I have found this variety of circuit board to be excellent for use in long-term gear.

What About Cabinets and Chassis?

Commercially manufactured project boxes and aluminum chassis have become too costly for many amateurs to consider. Cabinets often seem to exceed the cost of the circuit components by a huge margin. I tend to rebel at purchasing an $8 box for a $2 circuit. Most of my boxes are made from PC-board stock that I buy at hamfests where flea marketing takes place. Bargain prices generally prevail at these events. I create project boxes from this material by soldering the walls together with a 100-watt iron. I cut the stock with a heavy duty office paper cutter I bought at a used office equipment outlet. A saber saw may be used for cutting the board material, but it is difficult to obtain straight cuts when using a saw.

Once the chassis or cabinet is soldered, I abrade the outer surfaces with medium grade sandpaper to form tiny grooves in the copper. I scrub the finished work with hot soap and water, rinse it with clear water and dry it thoroughly. Next, I apply a spray coating of automotive primer paint. The finished coating of paint (your favorite color) may be sprayed on the work after the primer has dried. My labels are added to the panel after the finish coat of paint is dry. A final protective coating of clear polyurethane varnish may be added to protect the paint and the labels. Clear acrylic spray may be used in place of the varnish, but if will not offer the rugged protection made possible by clear varnish. Abrading the outer walls of the box helps the primer paint to adhere. Spray painted enclosures mar easily if paint is applied to a completely smooth surface.

Food Containers as Chassis

I seldom shop for groceries without inspecting the cookwares for potential chassis or project boxes. Cake pans, bread pans and other cookware items are relatively inexpensive, and they are quite suitable for use as foundation units when we build equipment. In a like manner, aluminum cookie sheets provide a good source for aluminum stock.

I tend to scrutinize the food containers for ideas that can be used in my shop. For example, rectangular sardine containers serve nicely as small chassis upon which to build QRP gear. You get to eat the contents as a bonus feature! A popular **QST** QRP transmitter, called "The Sardine Sender," was built on such a can. I once described a **QST** 40-meter QRP transmitter that was called "The Tuna Tin Two." It was built on a circular tuna-fish can. A circular PC board was mounted on the open end of the can. If there's a name for this game, it's called "saving dollars."

Other Sources for Project Boxes

Office supply stores contain a host of useful boxes that we may press into service as equipment enclosures. Recipe boxes -- plastic or metal -- make wonderful QRP project boxes. Bond boxes are ideal for those larger projects, and they come equipped with a carrying handle. I recall a 2-meter AM trans-receiver I built for **QST** publication in the 1960s. It was enclosed in a bond box. The article was entitled

"The Connecticut Bond Box." The completed unit was well suited to field applications because the bond box was rugged and had a durable finish (tan paint).

Metal project boxes may be constructed at low cost if you make them from galvanized iron furnace ducting. The material may be cut with tin shears or a saber saw. The box walls can be joined by means of solder seams along the mating edges. I have obtained free scraps of this metal from plumbing and heating shops.

More Innovation

No one seems to manufacture coil forms these days. Gone are those wonderful plug-in forms upon which we wound our coils. I have used a number unrelated items as coil forms, and with good results. For example, an inexpensive plug-in coil may be wound on the sleeve of a PL-55 phone plug. A three-circuit plug (such as one for stereo use) provides three electrical circuits for the coil. Once the coil is wound you can solder its leads to the terminals of the jack. The threads inside the plastic sleeve are filed away so that the sleeve forms a press-fit to the metal plug. A couple drops of glue will affix the sleeve to the metal plug after the coil wires are soldered in place. A coating of polystyrene Q Dope may be placed on the coil to protect it from damage. This will also keep the wire turns in place.

Small prescription pill vials may be purchased from the pharmacy at low cost. They serve nicely as small coil forms. You may also use PVC plumbing pipe for coil forms in low power RF circuits. PVC pipe works fine also for antenna insulators. Plastic coat hangers are inexpensive, and one hanger yields many antenna insulators if you saw the stock into 2-inch lengths. I have used coat-hanger pieces also as spreaders for open-wire feeders.

Have you considered using video cassette cases as enclosures for QRP rigs? They are inexpensive and weatherproof, which makes them ideal for housing portable QRP transceivers and Transmatches. I have purchased these boxes for as little as 75 cents each when they were on sale.

If I may return to the subject of coil forms, it is worth mentioning that expended shotgun shells are excellent for this purpose. I'm referring to the modern shells that have rigid plastic bodies. You may select the desired coil diameter in accordance with the gauge of the shell. The smallest size is 410 gauge and the largest one is 12 gauge. The primer in the brass base may be driven out by means of a hammer and metal punch. The resulting hole will allow the coil form to be affixed to a chassis by way of 6-32 hardware. Warning: *Do not attempt to unload a live cartridge except in a gun which is in good condition. Do not try to remove a primer that has not been fired in a gun.*

I have wound a number of VHF and UHF RF chokes and tuned-circuit inductors on plastic drinking straws. The material is rigid and may

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be cut easily with a razor blade or hobby knife. The plastic material
appears to have good dielectric quality as noted by checking various
straw-wound inductors with a laboratory Q meter.

In a similar vein, you can construct your own VHF toroid forms from
clear plastic or phenolic. There will be no core permeability with
these home-made toroids, but the self-shielding property of toroids
will prevail. I have made many toroid cores of this type from poly-
styrene sheeting. Plexiglass is suitable also.

Sources for Low-Cost Wire

Common hookup wire is relatively expensive today. I try to use scrap
wire for my projects in an effort to save money. Multiconductor tele-
phone cable yields a great many single conductors that are color-
coded. This small wire is ideal for tiny projects. I have obtained
scrap pieces of phone cable from installers, and the cable was free.

Be on the lookout for multiconductor cable that contains larger wires
inside the sheath. Some TV stations discard the cable from video
cameras when they become unreliable. This type of wire is also color-
coded and the individual wires contain stranded conductors. I once
bought a 10-foot length of 48-conductor cable at a ham radio flea
market for $2.50!

No. 12 or no. 14 stranded copper antenna wire is costly. I use a
cheap substitute that is strong, weatherproof and readily available.
I purchase no. 18 clear-plastic speaker wire (stranded conductor),
then pull the two conductors apart (easily done). In this manner
I can obtain 200 feet of antenna wire from a 100-foot roll of cable.
The cost per foot at this writing is 2-3 cents. I have found the
plastic insulation to be UV resistant over a four-year outdoor period
of use. The insulation does not crack or change color from air pollut-
ants. The bonus feature is that the plastic insulation adds strength
to the wire. It also prevents the copper conductor from oxidizing. I
have used this wire as low-impedance feed line (roughly 70 ohms)
on QRP DXpeditions in the West Indies. My dipoles and feeders were
made from a continuous piece of speaker cable. I pulled the wire
apart to form the dipole elements, tied a knot at the feed point,
then used the unseparated wire as a balanced feeder.

Don't overlook your farm-supply store as a source for no. 18 Copper-
weld wire. This material is sold in 1/4-mile rolls for as little
as $15. The steel core of the wire provides strength, but it tends
to kink and twist easily as you work with it, owing to its spring-
like characteristic. This wire is intended for electric-fence use.

Surplus Toroids, Rods and Slug-Tuned Coils

A few words of caution are in order at this time. Be cautious when
you purchase bargain-price toroids, rods and coil forms. The core
material may be entirely unsuitable for the application you have
in mind. Specifically, powdered-iron or ferrite cores are not all

the same. Various permeabilities are available for both types of core material. Each core "recipe" is designed for a particular range of operating frequencies. Generally speaking, the higher the permeability the lower the specified operating frequency in terms of resonant-circuit Q. The wrong core material can spoil the Q of a circuit, or even worse -- prevent the circuit from functioning. Broadband transformers that are wound on ferrite cores are a bit more tolerant of the core permeability, but an improper core can also degrade the transformer performance. An excellent text on the subject of broadband transformers was written by Jerry Sevick, W2FMI. This ARRL book is entitled **Transmission Line Transformers**, 2nd edition 1990. Another useful reference that is written in plain language is the W1FB book by Prentice-Hall, Inc., Englewood Cliffs, NJ. The title is **Ferromagnetic Core Design & Applications Handbook**. This book is sold also by Amidon Assoc., Inc., N. Hollywood, CA.

There is seldom a way to indentify the characteristics of an unmarked surplus toroid or slug-tuned coil when you examine these items at a flea market or see them listed in a surplus catalog. The vendors seldom know the core specifications. If the cores have standard color coding, such as Amidon and Micrometals Corp. powdered-iron cores, you will know what you're buying. Beware of other cores unless the seller can provide the specifications. You may end up with an audio-frequency core for that 20-meter project!

If you own a Q meter, you can test the toroids and coil forms at home. If the Q of the wound inductor is 100 or greater at the chosen operating frequency, the core will be suitable. If you lack a Q meter, relative tests for Q may be done by coupling a dip meter to the coil after resonating it with a suitable capacitance. If you can obtain a deep dip in the reading, the tuned-circuit Q is high. The higher the Q the farther you can move the dipper probe from the coil while still observing a dip.

Choosing the Right Fixed-Value Capacitor

Capacitors that are used for RF applications should have minimum unwanted inductive reactance (XL). The greater the stray inductance of the leads and internal structure the less effective they become as bypass and coupling capacitors. The situation worsens markedly as the operating frequency is increased.

Ceramic chip capacitors (leadless) exhibit the least XL, but they are suited only for direct soldering to PC boards (from foil to foil). Next best is the standard disc ceramic capacitor. The leads should be kept as short as practicable if they are to function effectively. NPO (zero temperature coefficient) disc ceramics are excellent for use in VFOs and other stability-dependent circuits. Beware of foreign-made NPO capacitors. I have had miserable results with some of them in VFO circuits. Drift was as bad as when I used standard non-NPO units!

Silver-mica capacitors ensure good tuned-circuit Q, and they exhibit low XL if the leads are kept short. The tolerance is good with these

capacitors. The marked value is generally very close to the actual value. Some amateurs use silver micas in VFO circuits, and they say that the frequency drift is minimal. I have had both good and bad luck with these units in VFOs. Some exhibit drift with heating while others are quite stable. I have found it necessary to hand pick the silver micas I have used in stable VFOs. It requires a cut-and-try process to find a stable group of capacitors. It is a time-consuming exercise because you must allow the PC board to cool down completely after soldering in a new capacitor. If the components are not allowed to cool, drift measurements are without meaning.

Polystyrene capacitors are somewhat more stable and predictable than silver micas. I have obtained good VFO stability with them up to 10 MHz. They are good capacitors for use also in RF filters. The marked values are very close to the actual capacitance values, much like silver micas. Polystyrene capacitors have a negative drift characteristic. This makes them useful in oscillator circuits that use powdered-iron cores in the VFO tuned circuit. Powdered iron has a positive drift characteristic, and the polystyrene caps compensate for the positive drift to enhance overall frequency stability. These capacitors are quite inexpensive compared to NPO ceramic ones.

Tantalum capacitors are better than electrolytic ones for bypassing in RF circuits. This is because the former units have much lower XL than electrolytic types exhibit. Tantalum capacitors are much smaller per uF than other types, and this makes them ideal for use in compact QRP projects.

Variable Capacitors

It is essential that you choose the proper type of variable capacitor for VFO and other frequency-critical tuned circuits. Not only is mechanical stability a vital consideration, but heat-related changes in capacitance must be taken into account also. Although I have had some acceptable results from NPO ceramic trimmers in VFOs, they can be mechanically unstable: vibration and heat can cause changes in capacitance (abrupt or gradual). I prefer to use small air trimmers if available space permits. Mica compression and plastic trimmers are entirely unsuitable in VFOs and similarly critical circuits.

Avoid using air variables that have aluminum plates. These capacitors are affected substantially by changes in heat, and this leads to VFO drift. Capacitors with plated brass vanes are the most stable of the available types. Wide plate spacing helps to minimize drift from heating. The VFO tuning capacitor should have a free-running bearing at each end of the rotor. Those with ball bearings are best, since they turn freely and stay put once set to the desired frequency. Tuning ease can be enhanced if you use a vernier drive which is free of backlash. The Jackson Brothers ball drives are my favorite units. Rim or friction-drive imported vernier mechanisms create backlash problems, especially if the variable capacitor does not turn freely.

Resistors in General

Resistors can be highly inductive, especially those that have a wire-wound structure. Most power resistors fit this description. They contain a coil of nichrome wire that establishes the resistance. These units are not suitable for use in RF circuits. Owing to the unwanted reactance of these components, a 50-ohm resistor might have an effective resistance (combined inductive reactance and resistivity) of, say, 100 ohms at a particular frequency. The situation will worsen as the operating frequency is increased.

The best resistor to use in an RF circuit is one that is composed of carbon with a composition outer insulation. Carbon-composition resistors are becoming scarce, and this is unfortunate. They have been replaced by carbon-film resistors. The resistive element is a layer of carbon that is formed into a spiral over the inner insulating form. The coiled carbon does introduce some reactance, but it need not be a concern at frequencies up to approximately 30 MHz. The effective XL of these resistors can be reduced by using two or more of them in parallel at critical circuit points. For example, we might elect to use two 100-ohm carbon-film resistors to obtain 50 ohms of resistance. Using these units in series worsens the XL problem.

Monolithic leadless resistors are now available for direct soldering to PC boards. These components exhibit the least unwanted inductance. They are sometimes called "chip resistors." I recommend this type of resistor in low-power VHF and UHF circuits where XL and XC must be minimized. They are excellent also in step attenuators.

You should be aware that resistors also exhibit capacitive reactance, and this has an effect also on the effective resistance of the unit. There is a certain amount of distributed capacitance across the body of a resistor. The longer the resistor body the lower the unwanted XC. In other words, a 1-watt resistor may be a better choice in a critical RF circuit than would be the case if we used a 1/8- or 1/4-watt resistor. I have seen a number of home-made 50-ohm dummy loads that measured 60, 70 or even 100 ohms at a specified frequency when I tested them with a laboratory reactance (RX) meter.

Transistor Selection and Use

You need not stock hundreds of transistors in order to have a working supply of general-purpose devices. There are thousands of transistor types available today, but most of them are so similar in performance that we would scarcely observe a difference in circuit quality were we to make substitutions with similar types of transistors. I like to keep a limited inventory of small-signal transistors on hand. I find that the common 2N3904 is a good device from dc to the lower end of the VHF spectrum. Likewise with the well known 2N2222A. I use 2N3906 transistors when a PNP device is required for low power applications, such as dc switches. The 2N4401 is a slightly more powerful version of the 2N2222A, but otherwise similar in characteristics. For VHF and UHF circuits I like to use the 2N5179 device.

There are many audio power transistors that are capable of HF band RF power amplification. These are generally much lower in cost than comparable transistors that are earmarked for RF service. Most hi-fi audio power transistors have an upper frequency limit (fT) that is in the VHF range. The Motorola MPS-U02 and MPS-U05 are examples of NPN audio transistors that perform well up to 30 MHz.

Here are some guidelines for selecting transistors for RF work.

(1) Obtain a data sheet for the chosen device or check its specs in the manufacturer's data book.

(2) Select a transistor that has at least twice the maximum VCE (collector to emitter) rating of the supply voltage (VCC) you plan to use in the circuit. This is necessary in order to have leeway for the sine-wave swing in ac or RF service. For example, a CW amplifier that uses a +12-V VCC will develop an ac voltage swing of roughly 24 V at the collector. In AM service the voltage swings to four times the VCC. Also, self-oscillation often causes high peak voltages at the transistor collector.

(3) Transistor gain is important. This is specified in dB. The greater the gain in dB, the lower the driving power needed to obtain the rated output power. For example, a transistor with 10 dB of gain, and a rated output power of 10 watts at a specified operating frequency, will require 1 watt of drive. The same transistor would require only 315 mW of drive to provide 10 W of output power if its gain were 15 dB. The greater the rated transistor gain the greater the potential for unwanted self-oscillation. High gain tends to set the stage for instability.

(4) The upper frequency limit of a transistor is an important factor. This is known as fT for a common (grounded) emitter circuit. The fT is specified for the frequency at which the device gain is unity or 1. For amplifier use the transistor should have an fT that is five to ten times the highest proposed operating frequency. This will ensure optimum stage gain at the chosen operating frequency. UHF transistors that are used at HF are quite touchy to work with and should be avoided. This is because the device gain increases substantially as the operating frequency in MHz is lowered. This phenomenon applies to all bipolar transistors, irrespective of the fT rating. In theory, the gain of a transistor increases 3 dB per octave lower. Thus, if a given tranistor has 10 dB of gain at 20 meters, the gain rises to 13 dB at 40 meters, and so on. The greater the transistor gain the more prone the amplifier becomes toward self-oscillation. Such a stage may be tamed by using resistive or capacitive loading at the input circuit, but this generally affects the input impedance and causes a loss of excitation power.

(5) Choose a transistor that has a PD (power dissipation) rating in milliwatts or watts somewhat greater than the proposed

dc input power for the stage. I prefer to use the rule of thumb that calls for a PD rating of approximately twice the stage dc input power. Thus, if my RF driver stage runs at 5 watts of dc input power (VCE X collector current) I use a device that is rated for a PD of 10 watts at 25 degrees C (case temperature). In a real-life situation we can use a transistor that has, say, an 8-W PD rating when operating with a dc input power of 5 W, but if this is done we need to pay close attention to keeping the device cool by means of an adequate heat sink.

Winding Toroidal Transformers

Numerous letters I receive from hams indicate a lack of knowledge concerning the winding of toroidal transformers. Some amateurs are especially perplexed about the meaning of the words bifilar and trifilar as they relate to multiple wires on the transformer core. The term "bifilar" means simply that two wires of identical length are wound on the core at the same time. In the case of a trifilar winding, we wind three wires on the toroid core at one time. The bifilar or trifilar windings may be laid on the core in parallel, or they may first be twisted together and placed on the core as a single wire. A suitable rule of thumb for twisting the wires is to have approximately eight twists per inch of wire. This may be done easily by clamping one end of the multifilar winding in the jaws of a vise and tightening the other end of the wires in the chuck of a hand drill. The drill is cranked slowly until the wires are twisted together.

Keeping Track of the Windings

It is helpful to use enamel wires that are of different colors. Enamel wire is made with red, green and tan/brown insulation. It is not an easy matter to find red and green wire in small-order quantity. In order to solve this problem I often use spray paint to color the wires before I wind them on a core. This makes indentification a simple matter when I connect the wires to the proper points in the circuit.

Transformer Polarity or Phasing

Some experimenters are confused by the black dots that appear over the transformer windings in some schematic diagrams. These important dots signify the phase of the windings, which is vital to correct circuit performance. The dots let you know which end of each winding connects to a key terminal on the PC board. All transformer windings have what is known as a "start" and "finish." The dots indicate the start or finish of each of the windings, but never are the dots used to signify both the start and finish of windings on a single core. A transformer that is connected to a circuit with the wrong phase can cancel the signal rather than pass it to the load.

Protecting the Transformer or Inductor Windings

Most powdered-iron toroids have been tumbled to provide smooth, rounded edges. Ferrite toroids are seldom made with smooth edges. In fact, many of them have sharp edges that can cut through the enamel insulation of magnet wire. The transformer performance can be spoiled when this happens because shorted turns often result. It is wise, therefore, to wrap a sharp-edge core with tape before adding the windings. I like to use Teflon pipe-thread tape (two layers) for this purpose. It is available from most plumbing supply stores. I have smoothed the sharp edges with a half-round file or sandpaper, but it can be tedious work when working with ferrite cores. They are very hard and brittle -- similar to ceramic.

I like to coat my completed transformer windings with a protective laquer, such as polyurethane varnish. A better agent is General Cement Corp. Polystyrene Q-Dope. This protective coating on the windings helps to prevent abrading of the wire insulation. It also keeps the turns in place and keeps out moisture and dirt.

I have used casting resin to encapsulate some of the transformers I have wound. This compound is available at hobby stores. It is suitable for use with broadband transformers, such as baluns. It is not a technique I recommend for use with high-Q, narrow-band transformers or inductors. Casting resin tends to lower the Q of the coil and it increases the distributed capacitance across the winding. This gives the effect of increased inductance.

Physical Relationship of Windings

Another frequent question is "Where do I place the secondary winding with respect to the primary winding?" For example, suppose we need to wind a toroidal transformer with a primary and a secondary winding. The primary has 20 turns of wire and the secondary has 5 turns. In this situation we will wind the larger winding first. The smaller winding is then wound over the larger one. Normally, I close wind the smaller winding over the low-Z (ground or VCC end) of the larger winding. This minimizes unwanted capacitive coupling between the windings and helps prevent harmonic currents from passing from one winding to the other by way of capacitive coupling. In some instances the smaller winding is spread over all of the larger winding area. It depends largely on how critical the circuit may be. In general terms, you may use either winding technique. Chances are that there will be no outward difference in circuit performance.

Calculating Transformer Wire Length

How much wire should you remove from the spool for a 30-turn winding? That's a common question, since nobody wants to waste expensive copper wire! Here's how I do it: I wrap a turn of wire or string through and around the core material to simulate one turn. I measure this wire, multiply it by the number of turns, and then add two inches to the wire length to allow for circuit connection.

The subject of winding toroids is treated in greater depth in **W1FB's Design Notebook** by The ARRL. The book is a useful adjunct to this volume because it treats similar subjects and it embodies the QRP topic.   I wish to recommend also the Amidon Associates, Inc. catalog, which has many pages of valuable information about toroids, balun cores, pot cores and rods.

Home-Made Heat Sinks

A variety of common materials are suitable for use as heat-sink stock. It is seldom necessary to purchase a commercial heat sink for transistors that operate at the QRP level. I often use sections of hardware store aluminum angle or channel for my heat sinks. Another inexpensive type of heat sink can be fashioned from 1/2- or 3/4-inch ID copper pipe caps -- the type used for household plumbing. A TO-220 style of transistor can be bolted to the flat end of the cap. The overall assembly is then affixed to a PC board by means of a 4-40 screw that is passed through the wall of the cap.

Larger heat sinks are made easily from aluminum channels that you can make from 16-gauge sheeting. The lips may be formed in a bench vise. Progressively smaller channels may be added inside the largest one to increase the effective area of this kind of heat sink. A thin layer of silicone heat-sink compound should be used between the mating surfaces of the channels to aid heat transfer. The channel pieces can be bolted together tightly with 6-32 screws and nuts. You may want to use dull black spray paint as a finish for the heat sink after all grease and oil is removed.

Most building-supply outlets stock a variety of copper and aluminum items that lend themselves nicely for use as heat sinks. Let your imagination operate freely as you browse in these stores!   Variety stores contain many baking items that are useful to hams. For example, I buy aluminum cookie sheets when I need sheet stock for heat sinks and small chassis or cabinets. Small cake tins and such are fine for use as chassis for QRP gear.

Most heat-sink compounds contain zinc oxide that is blended with oil and silicone grease. This material is messy and expensive. I like to use clear silicone grease as heat-sink compound. It is less costly and it is cleaner to use. The effectiveness of this grease can be enhanced by mixing it with zinc-oxide powder. You can often obtain the latter substance from your druggist.

Where to Get Electronics Parts

Your workshop activities are sometimes hampered by a shortage of parts for a particular project. This frustration can be minimized if you keep several of the more common items on hand. Bread-and-butter items such as disc-ceramic capacitors (0.001, 0.01 and 0.1 uF for example), 1/4-W resistors and popular small-signal transistors (2N2222, 2N3904, 2N3906, 2N4400 and 2N3553) are essential to have on hand for most QRP projects. Likewise with 741 op amps, an assort-

ment of enamel wire (gauges 20 through 30), small diameter stranded hookup wire, switches, jacks, screws, nuts and small variable caps. You may be wondering where these things may be found and how much money you need to spend in order to stock your shop. The answer is simple: Look for bargains at ham radio flea markets. If you see a quantity of a particular item at someone's swap table, ask the price for the entire lot. Eager vendors will often make you an offer you can't refuse! Don't be bashful about price haggling, because this is part of the ritual for some sellers.

Old radios and TV sets can be stripped of their components for use later on. I have purchased trade-in TV receivers from appliance. stores for as little as $5 each. Some dealers gave me two or three TV sets free of charge, just to get them out of their store rooms. If you have access to a landfill, check there for discarded radios and other electrical gadgets that contain useful parts.

No dedicated experimenter should be without a library of surplus electronics catalogs. Frequent short-term clearance sales by some of these merchants enable us to stock up on many day-to-day items we need for our projects. Be sure to check **The ARRL Handbook** for a list of electronics parts dealers that is keyed to the types of items they sell. A list of the mail-order houses I deal with follows.

All Electronics Corp.
P.O. Box 567
Van Nuys, CA 91408
1-800-826-5432

Digi-Key Corp.
701 Brooks Ave., S.
P.O. Box 677
Thief River Falls, MN 56701-0677
1-800-344-4539

Martin P. Jones & Assoc.
P.O. Box 12685
Lake Park, FL 33403-0685

Ocean State Electronics
P.O. Box 1458
6 Industrial Drive
Westerly, RI 02891
1-800-866-6626

Mouser Electronics
2401 Hwy. 287 N.
Mansfield, TX 76063
1-800-346-6873

Fair Radio Sales
P.O. Box 1105
Lima, OH 45802

Hoslett Electronics, Inc.
2700 Sunset Blvd.
Steubenville, OH 43952
1-800-524-6464

R & D Electronics
1224 Prospect Ave.
Cleveland, OH 44115
(216)621-1121

Jameco Electronics
1355 Shoreway Rd.
Belmont, CA 94002

Circuit Specialists
P.O. Box 3047
Scottsdale, AZ 85271
1-800-528-1417

Oak Hills Research
20879 Madison Ave,.
Big Rapids, MI 49307 .
1-800-842-3748
QRP kits & parts.

Never forget the sage expression "Caveat Emptor" when dealing with any surplus parts vendor. This "let the buyer beware" expression is, fortunately, seldom applicable. I have enjoyed integrity with all of the above vendors. If a wrong part was sent to me, I was always able to get the part exchanged or have my money refunded. I have not been as lucky with some other dealers in the mail-order business.

I wish to include in this chapter a list of dealers that sell special parts that we experimenters often need. Metal parts, plastics and

amateur radio kits are provided by the foregoing vendors.

ETCHED & DRILLED PC BOARDS & KITS

A & A Engineering
2521 W. LaPalma Ave.
Unit K
Anaheim, CA 92801
(714)952-2114

FAR Circuits (N9ATW)
18N640 Field Court
Dundee, IL 60118
Catalog avail.

TOROIDS, RODS & BALUN CORES

Amidon Assoc., Inc.
2216 E. Gladwick St.
Dominguez Hills, CA 90220
(310)763-5770

Palomar Engineers
Box 455
Escondido, CA 92025

METAL TUBING, GEARS & RODS

Small Parts, Inc.
6891 N.E. Third Ave.
P.O. Box 381966
Miami, FL 33238-1966
(305)751-0856

PLASTICS

United States Plastic Corp.
1390 Neubrecht Rd.
Lima, OH 45801
1-800-537-9724

Read the QST Ham Ads and display ads each month for information on new sources for small parts. The classified ads in CQ magazine and in various hobby electronics magazines also contain the names and addresses of surplus dealers.

Workshop Reference Library

In addition to the catalogs available from the foregoing vendors I wish to suggest the following publications as references that will be helpful during your experimental efforts.

ARRL Electronics Data Book

W1FB's Design Notebook

Solid State Design for the Radio Amateur

QRP Classics

The ARRL Handbook

W1FB's Antenna Notebook

Radio Communications Handbook
RSGB Publication

Motorola RF Device Data
Motorola Semiconductors, Inc.
P.O. Box 20912
Phoenix, AZ 85036

More Technical References

Numerous semiconductor manufacturers offer lists of their application notes and data books. Generally, the application notes are free, but there is a charge for books. You may send for literature lists at no charge. The following addresses may be helpful to you.

RCA Linear Integrated Circuits
RCA Solid State Division
Somerville, NJ 08876

Motorola Power MOSFET Transistor Data
Motorola Semiconductors
P.O. Box 20912
Phoenix, AZ 85036

Linear Applications Databook
National Semiconductor Corp.
2900 Semiconductor Drive
P.O. Box 58090
Santa Clara, CA 95052-8090

Handbook of Electronic Charts,
Graphs and Tables
by J. Lenk
Prentice-Hall, Inc.
Englewood Cliffs, NJ 07632

Linear Integrated Circuits
Signetics Corp.
811 E. Arques Ave.
Sunnyvale, CA 94086

MOSPOWER Design Catalog
Siliconix, Inc.
P.O. Box 4777
Santa Clara, CA 95054

IC Op-Amp Cookbook
Howard Sams & Co., Inc.
Indianapolis, IN 46268
by W. Jung

Handbook of Electronic Formulas,
Symbols & Definitions
by J. Brand
Van Nostrand Reinhold Co.
135 W. 50th St.
New York, NY 10020

The above list includes publications that are not produced by semiconductor manufacturers. I feel they are worthy of mention in the general context.

You may request a list of application notes and books that are available, but not listed here, from the foregoing manufacturers. Each of them publishes many books that you may want to include in your library.

Chapter Summary

I have attempted to provide in this chapter the basic data you need to get started in workshop procedures. Certainly, we have barely covered the subject here. The ARRL Handbook has an excellent chapter on workshop practices, along with a wealth of information that you will find useful as a QRPer. You may also want to equip yourself with the Radio Handbook by Bill Orr, W6SAI. It is available from Howard Sams & Co., Inc. (see above address). If you are to "learn by doing," it is essential that you read technical publications and establish an activity program in your workshop.

CHAPTER 3

RECEIVERS FOR QRP

The QRP enthusiast must decide which type of receiver is best suited to his needs, consistent with his technical ability and financial means. A variety of options are present: we may use the receiver portion of an existing transceiver for QRP signal reception. Likewise when it comes to employing a commercial receiver we may have on hand. Although receivers of this kind provide excellent weak-signal reception, they are large and seldom in keeping with the general theme of QRP operation. Perhaps of greater concern is the fact that when we use commercial QRP gear we are depriving ourselves of the complete thrill of QRP operation. A completely home-made station can provide a level of pride and accomplishment that is never experienced by those who start and finish their amateur careers with store-bought equipment. If you have never built a ham receiver, perhaps this is the time for you to gather some parts, heat your soldering iron and get with the program!

QRP Receivers in General

I have known QRPers who believed that a QRP receiver need not only be small, but it should have minimal circuitry and not cost much money. True, some very simple circuits offer acceptable performance, but most do not. Certainly, I recommend that the rank beginner to equipment construction start with a few simple circuits. This enables him to learn by doing, and to become familiar with how receivers operate. A good starting point is the DC (direct conversion) type of receiver. If you are a beginner I would like to suggest you build something like "The Neophyte Receiver" that was described in QST for February 1988, page 14. The circuit contains only two ICs and it performs on 40 or 80 meters. Circuit boards and parts kits for this receiver are available (see article footnote) for this WA3RNC project. Building a simple non-kit receiver can come later on.

DC receivers have the same limitations that are common to the older regenerative types of receiver. Single-signal reception is not available to the user. In other words, these receivers respond equally to signal energy that is above and below center frequency. This can make QRM seem much worse than it would be if we were using a super-heterodyne receiver with an IF filter, which rejects either the upper or lower sideband energy that may be present.

DC receivers can radiate a signal at the frequency to which they are tuned. This is because the VBFO (variable beat-frequency oscillator) operates at the incoming signal frequency. If an RF amplifier is used between the detector and the antenna, the problem is often solved because of the isolation afforded by the RF amplifier.

A DC receiver consists of a product detector. It replaces the first mixer of a superhet receiver. In effect, it functions as a mixer, but produces audio energy at its output rather than IF (intermediate frequency) energy. The VBFO provides an injection frequency for the product detector that is offset from the incoming signal by the amount necessary to cause a beat note for CW or a suitable offset for copying USB or LSB phone signals. In other words, if you desire, say, a 700-Hz CW beat note for a 3550-kHz signal, the VBFO is set for 3550.7 or 3549.3 kHz. The difference frequency appears in the headphones or speaker as a 700-Hz audio tone.

Since the product detector in a DC receiver may be only RF stage in the receiver (except for the VBFO), the overall receiver gain must be developed at audio frequency. The IF amplifiers in a superhet accomplish this task. Good weak-signal reception is possible when a DC receiver has an overall gain of 75 to 100 dB. This requires a high-gain audio system after the detector. An unfortunate byproduct of the high audio gain is receiver microphonics and hum. The minute movement of electrical parts (variable capacitors and such) in the receiver front end appear as audio energy and this noise is amplified many times. Bumping a DC receiver frequently results in a "clang" in the earphones.

A condition known as common-mode hum is found in DC receivers. It becomes more prevalent as the operating frequency is increased. It can be severe at 14 MHz and higher. This malady occurs when we use an ac-operated dc power supply with a DC receiver. RF energy from the VBFO enters the power supply and reaches the rectifier diodes. The diodes 120-Hz "hum modulate" the VBFO signal, and it enters the receiver at the listening frequency (via the antenna). This unwanted energy becomes manifest as a loud hum at the receiver output port. End-fed wire antennas make the problem worse, owing to the antenna being so close to the power supply leads, which radiate this energy. This ailment does not occur when we use a battery for the power source.

How might we eliminate the hum problem? The cure can be managed easily if we do a few things to the power supply. Fig 3-1 shows the circuit for a typical 12-V dc supply that has been suppressed for common-mode hum. Note that the rectifier diodes have 0.01-uF capacitors across them. This allows the stray RF energy to bypass the diodes rather than become rectified. The 120-V ac input line is also bypassed for RF energy. This keeps the line cord from radiating or picking up RF energy. Finally, RF chokes are used at the dc output lines of the power supply. These circuits may be added easily to existing power supplies. In addition to the foregoing preventive measures, a quality, noninductive earth ground should be connected to the case of the power supply and receiver. An antenna that is fed with coaxial cable also minimizes the potential for common-mode hum.

DC Receiver Selectivity

We have already learned that single-signal reception is not possible for a DC receiver. But, we can improve the overall selectivity by

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using a passive (coils and capacitors) or RC active (op amps, resist-
ors and capacitors) audio filter. Examples of this technique are
provided later in this book. The term "active" in electronics means
that the circuit requires an operating voltage. Passive circuits
do not.   RC active audio filters can be designed to yield a high-
pass, low-pass or band-pass response. There are elaborate versions
of these audio filters that enable the user to vary the response
frequency and the circuit Q. The greater the Q the sharper the filter
response. The common practical limit to the number of filter poles
(sections) is usually four. These filters may be designed to provide
audio gain, or the gain may be set for unity (1). Excessive Q causes
a condition known as "ringing," which can be annoying when listening
to weak signals. The best place to locate the audio filter is directly
after the first audio amplifier in the receiver. If the filter is
used between the receiver output and a speaker of headphones it can
exhibit impaired performance. This is caused by it being overloaded
by high levels of audio energy.
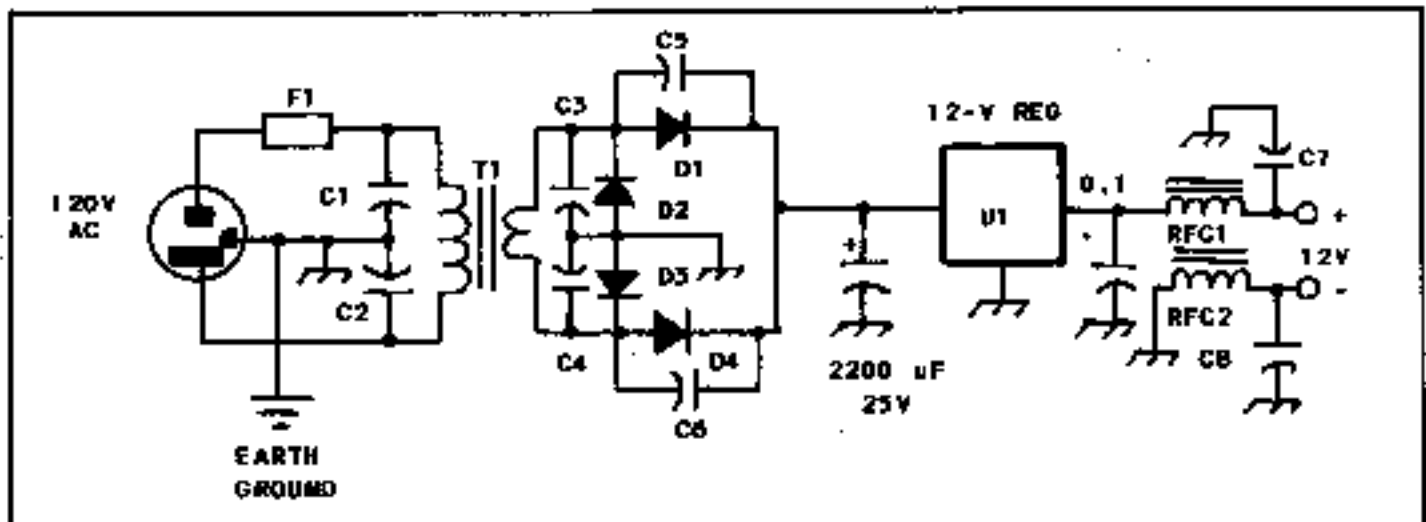
Fig 3-1 -- Example of how to treat a power supply for common-mode hum elim-
ination. C1 through C8, inclusive, are added to the circuit. They are 0.01-
uF disc ceramic units. C1 and C2 must have a 1000-V rating. The other units
need have only a 100-V rating. RFC1 and RFC2 for output currents up to 1 A
may consist of 20 turns of no. 20 enam. wire on an Amidon Assoc. FT-82-43
ferrite toroid core (850 mu). Larger wire diameter is needed for higher output
current. The on-off switch is not shown for this circuit example.

I am not attempting to villify DC receivers in this discussion. It
is important, however, that you understand their limitations and person-
ality traits. Certainly, they play an important role in the QRP pastime.
They are easy and inexpensive to construct. DC receivers are small
and lightweight. This makes them ideal for use in portable gear.   They
can be designed for low-current drain, and this is an important facet
of field operation when batteries supply the operating power.

I wish to mention finally that DC receivers often succumb to over-
loading from loud commercial AM broadcast stations that are nearby
in frequency. This results from the use of simple product detectors,

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such as those that employ a single dual-gate MOSFET (type 40673 or 3N212). Circuits of this type work all too well as AM detectors. The solution to the problem is to use a balanced product detector, such as a pair of JFETs, a CA3028A IC or a diode-ring doubly balanced type of detector. An RF attenuator (variable) at the receiver antenna input point is helpful also. It can reduce the strength of the offending signal and minimize the overloading effect. It will also reduce the strength of the desired signal. This is seldom a problem unless we are trying to copy a very weak signal. Fig 3-2 is a block diagram of a typical DC receiver.

Fig 3-2 -- Simple block diagram of a DC receiver. Q3 operates at the same frequency (approximate, see text) as the incoming signal. The difference frequency at the output of U1 is at audio frequency. Q2 is a low-noise audio preamplifier. FL1 may be a passive or RC active audio filter (see text). U2 boosts the audio signal to speaker or headphone volume.
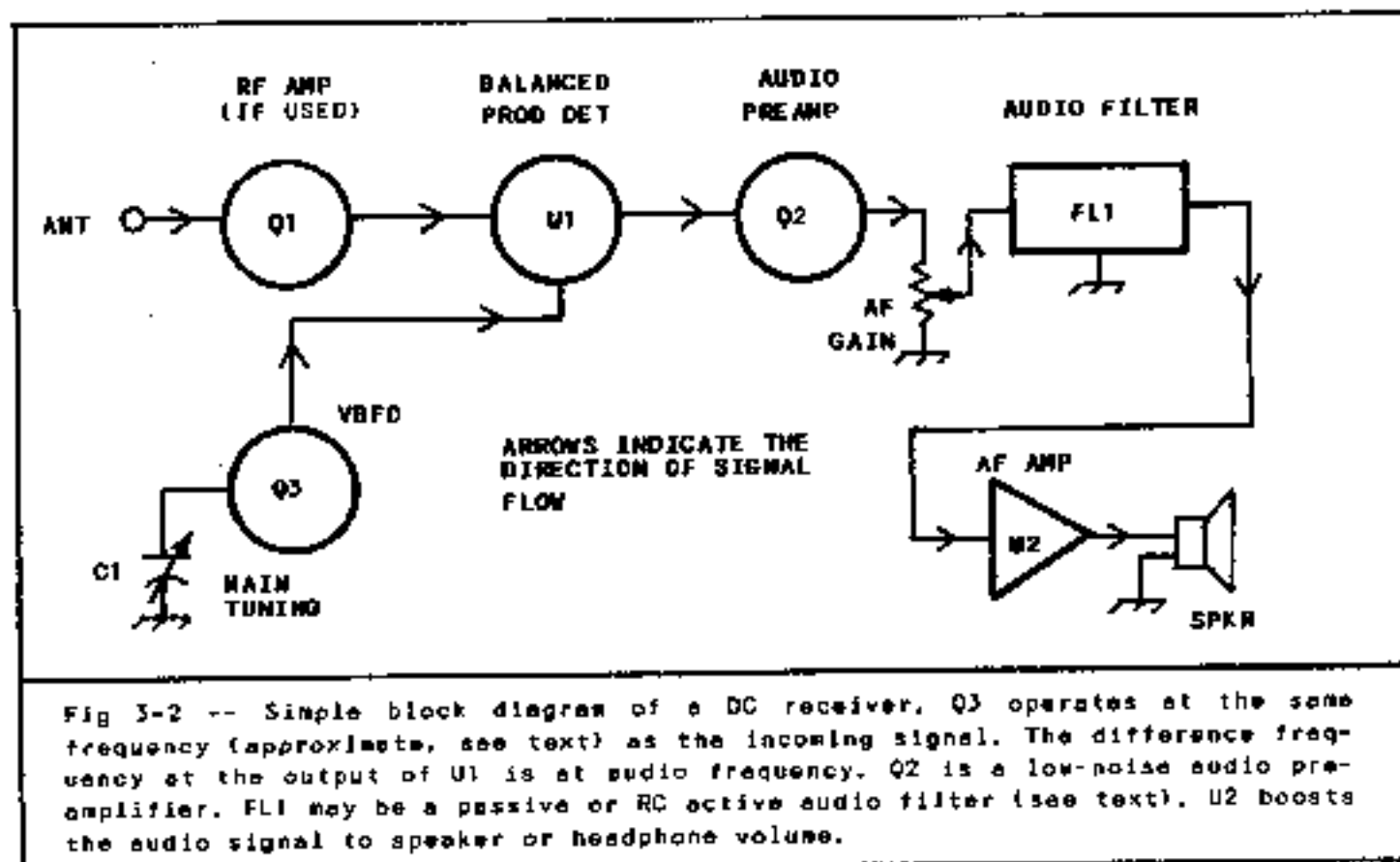
Superheterodyne Receivers

Reception can be improved greatly when we use a superheterodyne receiver instead of a DC one. The major benefit is improved selectivity, the elimination of receiver microphonics and single-signal reception. This style of receiver is practically immune to common-mode hum. Also, it becomes easier to include AGC (automatic gain control) and an S meter.

There are few negative features attendant to a well designed superhet. Notable among them is increased cost and complexity for a superhet that is designed for high performance. Also, the heterodyne scheme can cause "birdies" (spurious signals) to appear across the tuning range. This can become a major problem if the receiver is a double or triple conversion type: Many oscillator frequencies come into play. Their harmonics and resultant mixing products can appear as unmodulated

carriers as we tune through a particular amateur band. Even the best
of the commercial receivers have birdies somewhere in the tuning range.
These may be detected by disconnecting the antenna, advancing the audio
gain and tuning from band edge to band edge. Some of these unwanted
responses are quite weak, while others may register S6 or higher on
the S meter. I recall testing a specific high-priced commercial receiver
that exhibited 46 spurious responses from 1.8 to 29.7 MHz (ham band
only receiver). Fortunately, most of these spurs aren't noticeable
when atomspheric noise and signals are present.

Use caution if you design your own superheterodyne receiver. Try to
analyze the frequency relationships of the heterodyne oscillators,
the local oscillator and the BFO with respect to the receiving frequency
and the frequency converters of the system. Check to see where the
harmonics of the various oscillators fall. Even the BFO harmonics can
appear in the receiver tuning range.

Superhets Need Not Be Complex

A typical noncomplex superheterodyne receiver is shown below in block
diagram form. The differences are many when you compare this system
to that shown in Fig 3-2.
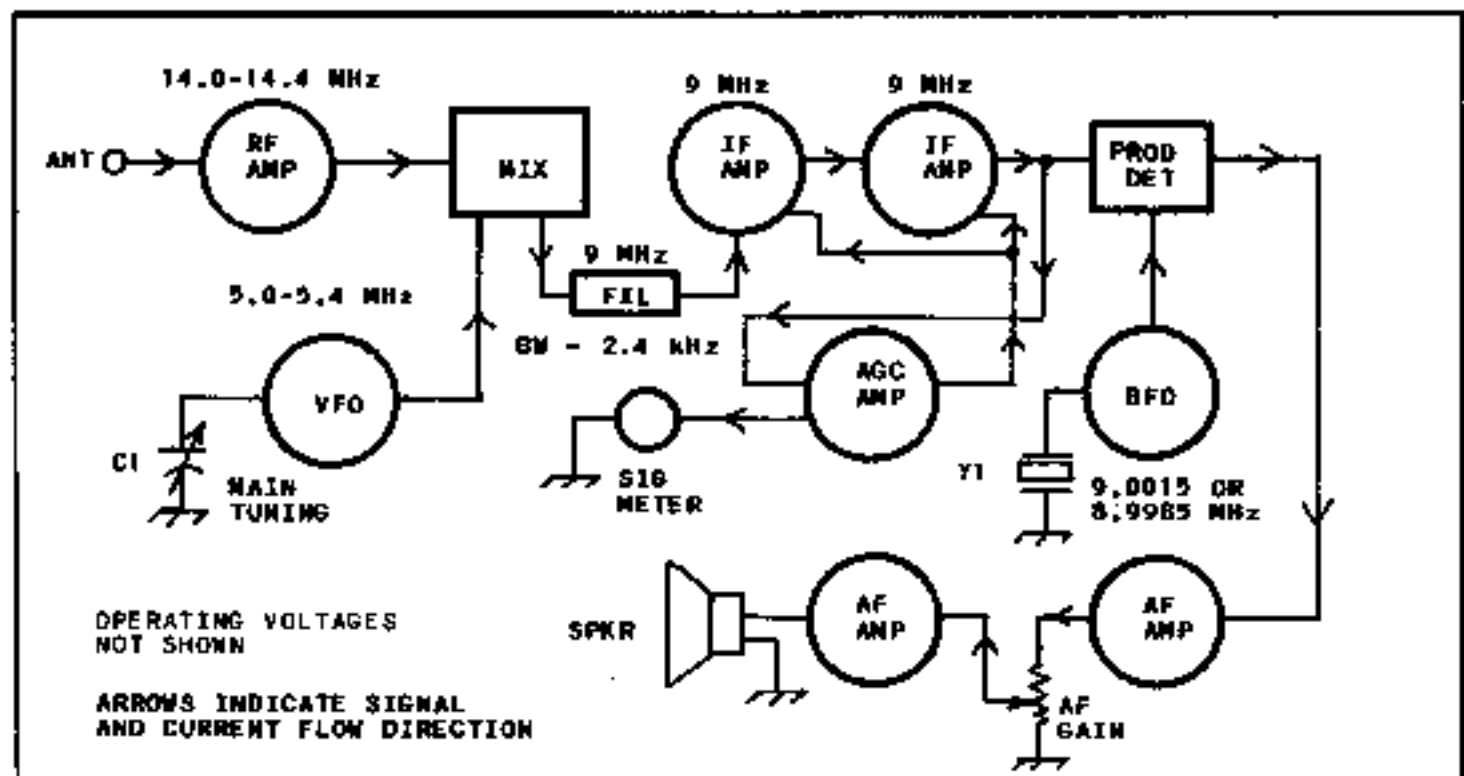
Fig 3-3 -- Block diagram of a single-conversion superheterodyne receiver. Passive
or active devices may be used for the mixer and product detector. The two freq-
uencies for Y1 permit USB and LSB reception. Frequencies listed are to show
just one possible arrangement for 20-meter reception. Multiple IF filters may
be used to provide various degrees of selectivity. A manual gain control may
be used for varying the output level of the RF amplifier.

The foregoing block diagram suggests that a superhet can contain a minimum number of stages and still offer good performance. Actually, we may remove some of the circuits in Fig 3-3 and still obtain good results. For example, no rule has been written that says we must use IF amplifiers, AGC, an RF amplifier or an S meter. It is important, however, to ensure sufficient overall receiver gain if we are to copy weak signals. For example, a bare-bones superhet may contain only a mixer, an IF filter, a product detector, BFO and a stage of audio amplification. Of course, we also need a VFO or LO (local oscillator). We may elect to eliminate the IF filter if we are willing to use an active audio filter after the first audio stage.

A practical QRP superhet might have an active mixer (to provide conversion gain), an active product detector (more gain) and one or two high-gain audio amplifiers. We still need our VFO and BFO with this so-called "gutless wonder." In the simplest example we may develop this type of receiver around only five or six transistors or two or three ICs. Design challenges of this kind have stimulated many hams toward getting the most for the least when building equipment. Needless to say, the absence of an AGC circuit requires the operator to "ride herd" on the receiver gain by means of the audio gain control. The annoyance of tuning across a very strong signal can become an ear-shattering event when there is no AGC. The same is true when we use a DC receiver. For the most part, S meters are usually "guess meters," so they serve no vital purpose in QRP work. Eliminating them simplifies the circuit. When we strip a superhet circuit to the bare essentials we need to keep in mind the need to have 75 to 100 dB of overall receiver gain if we are doing weak-signal work. If not, the audio gain may be fully advanced and the weak signal may still not be loud enough to copy.

If sufficient overall gain is available we need not use an RF amplifier ahead of the mixer for operation from 160 through 30 meters. This is because atmospheric and man-made noise from the antenna is of greater amplitude than the noise generated within a typical mixer. We do need an RF amplifier above, say, 10 MHz in order to provide a noise figure that is lower than that of the mixer alone. This calls for an RF amplifier that has sufficient gain to override the mixer noise. The RF amplifier must exhibit a low noise figure in order to reach this objective. In other words, the incoming noise from the antenna at 10 MHz and higher is usually less than the mixer noise. Current flow within an amplifier generates noise. Hence, it is worthwhile to experiment with the biasing (total current) of the RF amplifier in order to obtain a low noise figure. Amplifier instability can also generate noise.

Front-End Selectivity

In addition to introducing selectivity at the IF (IF filter) we need to provide front-end selectivity where the antenna connects to the receiver. The input tuned circuits for RF amplifiers and mixers should have reasonably high Q in order to achieve this goal. Double- or triple-tuned LC circuits may be used to narrow the response of the input tuned circuit. Front-end selectivity helps prevent strong, unwanted signals within or outside the frequency of interest from overloading the receiver.

Local Oscillator Considerations

The local oscillator must be frequency-stable, especially if we use narrow IF filters. A drifting VFO can cause the signal of interest to vanish quickly from the passband of the IF system. This is true especially when we use IF filters that have a 250- or 500-Hz bandwidth. More on this subject when we get into VFO design.

The mixer injection power or voltage that is supplied by the VFO is also vital to proper receiver performance. Each type of mixer we use requires a specific level of LO injection in order to ensure optimum mixer conversion gain and dynamic range (ability to withstand very strong signals without overloading). Therefore, I urge you to learn the LO requirements of the mixer you employ. Most diode mixers call for +7 dBm of injection power (approximately 8.5 mW). Dual-gate MOSFET mixers, such as the 40673, need roughly 5 volts P-P of LO injection on gate no. 2 for top performance. Most IC mixers or bipolar transistor mixers require 3 to 4 volts P-P from the LO. Bear in mind that these are rules of thumb for the beginner.

The output waveform from the LO should be reasonably free of harmonics or other spurious energy. If the waveform is unsanitary, so to speak, various unwanted injection frequencies reach the mixer, along with the desired frequency. This can cause the reception of signals from frequencies other than the intended one. I like to use a harmonic filter at the output of my LO chain to provide a clean sine wave at the injection frequency. A simple half-wave low-pass filter with a loaded Q of 1 is usually ample.

It is wise to isolate the LO circuitry from the rest of the receiver. A shield compartment and decoupling networks in the dc supply lines to the VFO will help isolate the LO system. Unprotected oscillator circuits can pick up stray RF energy from other parts of the circuit, and this can seriously affect the VFO stability. LO isolation is rather important when the VFO is used within the cabinet of a transceiver. Ideally, the VFO circuit would be contained in a separate box, external to the transceiver. This general rule applies to DC receivers also.

IF System Notes

You may use one or two IF amplifiers in your superhet, or none at all, as discussed earlier. Two IF amplifiers offer better AGC action than does a single amplifier. Also, the overall receiver gain will be much greater when we utilize two IF amplifiers. Contrary to a misconception that exists, high-gain IF amplifiers do not require tuned input and output transformers. For example, IC amplifiers such as the MC1350P or CA3028A work nicely when an RF choke or broadband transformer is used in place of a conventional IF transformer. Capacitive coupling is used between the stages when RF chokes are used. This method aids amplifier stability by providing low-Q input and output circuits. Transformers are useful, however, to provide an impedance match between the IF amplifiers and their input and output loads to ensure maximum gain. Most IC types of RF/IF amplifiers exhibit a 2000-ohm input and 8000-ohm output characteristic (base to base and collector-to-collector, respectively).

A well designed AGC system will yield up to 80 dB of gain control when two IC amplifiers are used in the IF system. Examples of high-performance AGC amplifiers and control circuits are described in **Solid State Design for the Radio Amateur.**   S meters are used with these systems.

IF Filtering

We must exercise care when using a mechanical or crystal-lattice IF filter. All filters require a specified termination impedance if they are to function as the design specifications are stated. An improper load impedance can cause the filter response to be asymmetrical, and the ripple (dips and peaks across the nose of the response curve) may become prohibitive. Be sure to check the manufacturer's data sheet when you use one of these filters. Also, some filters call for resonating capacitors at the filter input and output terminals. These are essential if we are to obtain correct filter performance.

Home-made IF filters are inexpensive and fairly easy to design. Ladder filter design, based on low-cost, surplus crystals, is treated in depth in **W1FB's Design Notebook** (ARRL). The book contains two articles on the subject by Wes Hayward, W7ZOI. The procedure is simple enough to follow for even the most rank beginner.

Suitable crystals for home-made ladder filters are available for as little as $1 each. Most of these crystals are surplus units for computer use. Check your electronics surplus catalogs for listings.   I have made some excellent 250-Hz ladder filters from TV color-burst crystals that I bought for 75 cents apiece.   Filters of this variety can often be built for less than $5, whereas a comparable commercial filter can cost as much as $75.

Post IF Filtering

A typical superhet receiver has the IF filter between the mixer and the first IF amplifier. Some amateurs have cascaded IF filters at this circuit point in order to improve the filter skirt selectivity. Others prefer to use a "tail end" filter. This method calls for using the second filter immediately after the last IF amplifier. A second filter used in this manner reduces receiver noise by rejecting wide-band noise that comes from within the IF amplifiers. The results can be dramatic when doing weak-signal work. It is important to make sure both filters have the same center frequency.   The tail-end filter is chosen for a wider bandwidth than the first filter. For example, a 250-Hz filter might be used just after the mixer, but a 500-Hz filter is used after the last IF amplifier. This helps assure that the passband of the first filter falls within the passband of the tail-end filter.

Receiver Audio System

It is not unusual to find receivers that have inferior audio, even though the remainder of the circuit performs nicely. This is true with some expensive commercial receivers as well. The cause may often be attributed to a miserly designer who tried to cut cost by using an

inadequate of low-cost output amplifier. It is true that we seldom need more than 0.5 W of audio power to excite a speaker to comfortable room level. But, if the output amplifier is rated for a maximum output power of 0.4 or 0.5 watt, it is producing distortion at that power level. Also, some audio-output ICs have inherent cross-over distortion. Weak signals take on a "fuzzy" sound in the presence of this distortion.

It is better to use an output audio amplifier than can produce 5 or more watts of power. You will never use all of those watts, but at normal listening levels you will not hear distortion. This philosophy can be equated to that which applies to HI-FI systems, where it is not unusual to find a "golden ears" person using a 100-W amplifier.

Distortion is also caused by a loudspeaker that is incapable of handling the available audio power. Manufacturers tend to use speakers that have small magnets. This limits the handling power of the speaker. It is absurd to have a 1-watt audio amplifier and a 400-mW speaker, but I have observed this shortcoming in a number of commercial products. Audio quality can be improved markedly if we use a 10-W speaker with audio systems that produce up to 5 watts of output power. When in doubt, select a speaker with a very large magnet. Although it may cost more than a bargain-price smaller unit, the rewards are worth the added expense.

Audio response shaping is worth considering when you design a receiver. We seldom need the frequencies below approximately 400 Hz, and those above 2500 Hz are not necessary for communications work. The audio response at the low frequency end of the spectrum can be rolled off by selecting coupling capacitors that pass the highs while rejecting the lows. To achieve this effect you may have to use 0.1-uF coupling capacitors instead of 2.2- or 10-uF ones. Response at the upper end of the communications audio range can be limited by using bypass capacitors from the output ports of the audio amplifiers to ground. I use capacitors with values from 0.002 to 0.1 uF for this purpose. The effective QRM can be reduced markedly if you adopt this technique for your receiver. Low-pitched rumble and noise tends to vanish, and so do those high-pitched heterodynes and SSB monkey chatter.

Single Conversion versus Double Conversion

Which is best? Single conversion or double conversion in a superhet receiver? It is said that "beauty is in the eyes of the beholder." This philosophy probably applies to the choice of receiver circuits. Simplicity is the keynote of a single-conversion receiver. This circuit is relatively inexpensive to build. A double-conversion receiver needs additional stages and a collection of heterodyne crystals. Also, the chance for birdies is greater when the double-conversion method is used. Careful gain distribution is mandatory in any receiver if we are to realize high dynamic range. It becomes more critical in a double-conversion setup, mainly because there is more potential gain to control. On the other hand, double conversion is beneficial when we want to minimize image responses. For example, the receiver may have a first IF of 40 MHz. This energy is well away from the receive frequency,

which discourages the development of image responses. The 40-MHz energy is then converted (second mixer and LO) to some lower frequency, such as 3.3 MHz or perhaps 455 kHz. Therefore, we are dealing with two IF systems, two or more IF filters, two mixers and two LO chains. The picture becomes even more complex when a designer elects to build a triple-conversion receiver. Personally, I prefer to work with single-conversion schemes. I encounter fewer performance problems when I follow this course!

Minimize the Frills

You will be happier and spend less money if you avoid the so-called bells and whistles we find on modern amateur receivers. The QRPer does not need AGC, an S meter, RF gain control, digital frequency display or passband tuning in order to have fun and success. Notch filters, IF shift, computer interface, noise blankers and the like are also unnecessary for routine communications. Good dynamic range and frequency stability are, however, essential. The subject of dynamic range and how to achieve it is covered in **Solid State Design for the Radio Amateur.**

Similarly, we do not need to use ICs, just because they are available. PC board layout can become a nightmare with some ICs. Discrete devices, such as bipolar transistors and FETs can provide good performance. Circuit-board layout is simple and there is plenty of opportunity to experiment with operating parameters when we use discrete devices. Many ICs prohibit external changes that change the performance. Unstable operation frequently plagues a high-gain IC. This is because there is so much gain in a small package. For the same reason it can be a chore to design a PC-board pattern that does not encourage instability. It is not always easy to locate the recommended IC bypass capacitors at the pins of the IC, and this is essential for ensuring stable oper-ation. Long PC-board foils are inductive, and this contributes also to instability. Although most ICs are of modest cost ($1 to $3), we can buy popular bipolar transistors for as little as 10 cents each. Many FETs are available for as little as 50 cents apiece.

PRACTICAL RECEIVER DESIGN

We will begin this section by reviewing various common circuits that you will use when you build receivers. Working parts values are assigned to these tested circuits. This will take the guess work out of your design exercises. Generally speaking, substitute transistors are suit-able for these circuits, provided the substituted transistor or diode has similar electrical traits. For example, types 2N3904, 2N2222 and 2N4400 or 2N4401 may be interchanged in many circuits with no great difference in performance. Try to select a substitute device that has approximately the same maximum voltage, current, gain and fT (upper frequency gain limit). The fT, where gain is unity, may rated for a higher frequency than the transistor for which a substitute is chosen, and performance will be okay. A workable rule of thumb for fT is to choose a transistor or IC that has an upper frequency limit that is five times or greater the desired operating frequency. This ensures

maximum available gain (hfe) at the circuit operating frequency. Oscillators, on the other hand, can be made to work at or even slightly above the transistor fT, but output power will be low. Oscillators are, in effect, amplifiers that use feedback to make them oscillate.

Variations in resistor values as great as 20 percent are acceptable for the practical circuits in this book. The same is true of bypass and coupling capacitors. You should adhere to the capacitor types and values called for in tuned circuits, VFOs, filters and crystal oscillators. If you do not have the exact value specified you may use combinations of series or parallel capacitors to obtain the value called out in the parts list. In this context, if the circuit requires a 50-pF NPO capacitor, you can use two 100-pF NPO units in series to obtain the specified value.

Air variable tuning capacitors may be modified easily to arrive at a specified maximum capacitance (plates fully meshed) value. You can use a pair of needle nose pliers to remove vanes from the rotor or stator to reduce the total capacitance. It is usually a simple matter convert, for example, a 100-pF air variable to a 50- or 75-pF unit through this procedure.

Crystal Oscillators

The practical circuits in this section may be used in receivers or transmitters. We will examine only the more popular oscillator types here. A plethora of oscillator circuits has been developed and named after the inventors. Each breed of oscillator has its salient features, but in all cases the key to proper performance is the feedback power used in the circuit. Feedback consists of a portion of the oscillator output power which is fed back to the input circuit. This is called positive feedback. Negative feedback is used in amplifier circuits to prevent them from oscillating. This is called neutralization.

Poor oscillator performance can result from the use of too little or too much feedback voltage or power. An oscillator with a low value of feedback may not oscillate at all, or it may be a sluggish starter when power is applied. Excessive feedback, conversely, can cause the oscillator to generate spurious frequencies, cause the transistor to draw excessive current and lead to frequency drift. In a worst case situation the crystal can be damaged by excessive feedback. This is most likely to occur when we use solid state power oscillators, or tube types of oscillators.

Positive feedback is of the same phase as the oscillator input signal. Negative feedback is 180 degrees out of phase with the input signal. This is important to remember when you use a link on the output tuned circuit to obtain feedback. If circuit fails to oscillate, reverse the feedback winding to obtain the proper phase.

A suitable ballpark value for the feedback is 0.25 the oscillator output power. This clearly illustrates that the efficiency of an oscillator is much lower than if the same transistor were used as

a class C amplifier. The feedback power is subtracted from the out-
put power. Typical crystal oscillator efficiency is on the order
of 30 percent. A solid state class C amplifier can provide an
efficiency as great as 75 percent if the operating conditions are
optimized.

XC1 = 230 ohms   XC4 = 450 ohms
XC2 = 450 ohms
XC3 = 230 ohms

XC1 = 22 ohms
XC2 = 150 ohms
at resonance

XC1,XC2 = 450 ohms
XC3 = 450 ohms

XC1 = 225
ohms

XC2 = 150
ohms
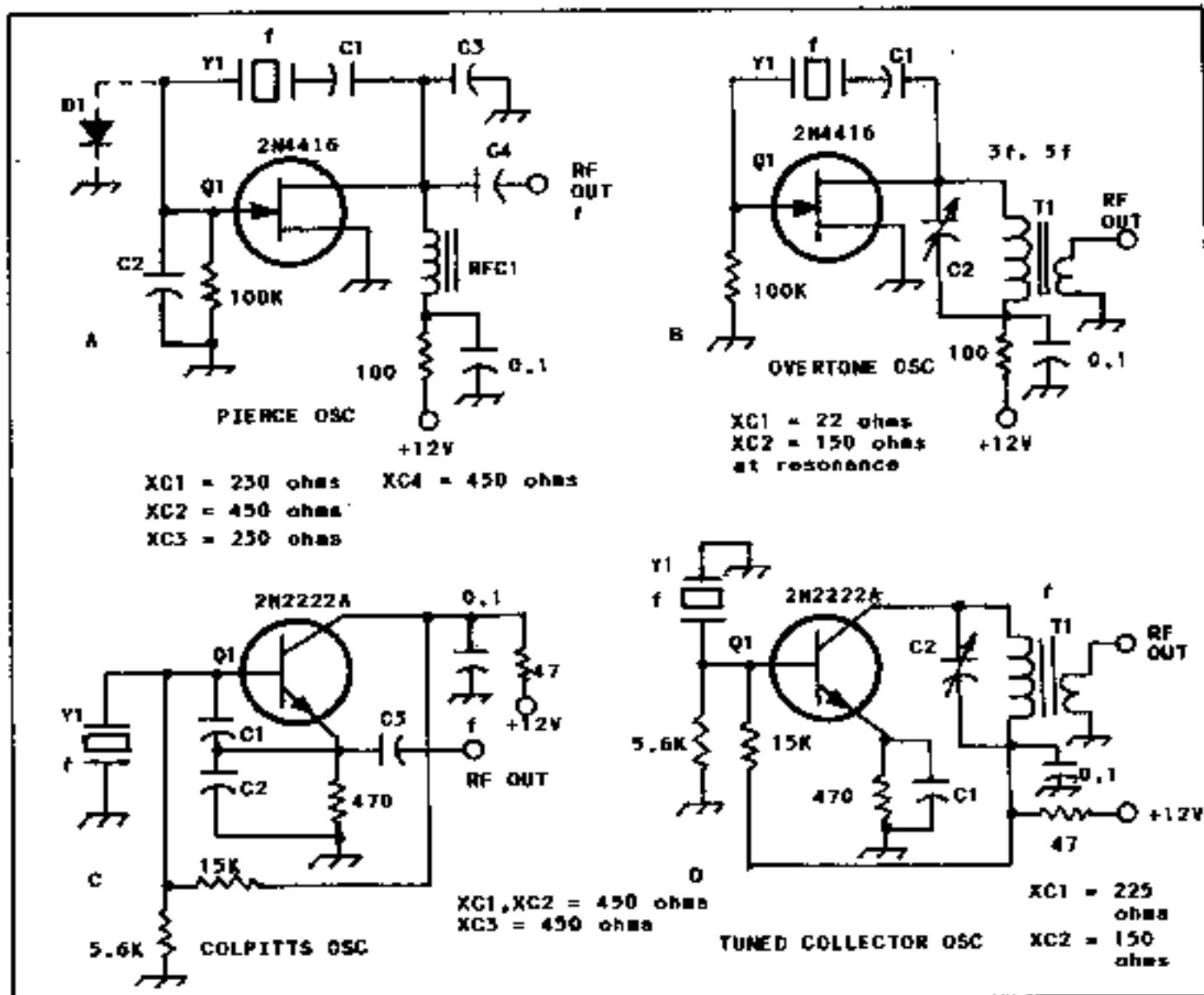
Fig 3-4 -- Practical examples of four crystal oscillators. Xc (capacitive
reactance in ohms) is provided above for the feedback, coupling and tuning
capacitors. These are approximate values. Some experimentation may be necess-
ary in order to obtain peak oscillator performance. See text for a thorough
treatment of these circuits. D1 may be added as a bias stabilization device.
A 1N914 is generally used for this application. D1 can be used with any FET
crystal or LC oscillator. It minimizes harmonic output currents.

The common oscillators in Fig 3-4 are shown with values for the XC of critical capacitors. You may obtain the capacitance value from these numbers by using 2 pi times XC times the frequency in MHz. Capacitance is in uF. Hence,

$$C_{uF} = \frac{1}{2\pi\ f\ XC} \qquad\qquad Eq\ 3-1$$

where XC is in ohms and 2 pi is 6.28.

Please bear in mind that the XC values expressed in Fig 3-4 are approximate. The absolute values depend upon the transistor gain and the activity of the crystal. Optimization of the feedback values may be accomplished by observing the oscillator output waveform with a scope and selecting the level of feedback that provides the least waveform distortion, consistent with reliable oscillator starting. Experimentation with the transistor bias value will also help to provide waveform improvement.

Fig 3-4A illustrates a Pierce oscillator. This untuned oscillator relies primarily on C2 and C3 to establish the feedback. RFC1 in combination with stray circuit capacitance (10 to 15 pF typically), must be resonant below the crystal frequency in order to obtain oscillation. The value of blocking capacitor C1 can be varied in the interest of the purest output waveform. Its value, in any event, should be the smallest one that permits reliable oscillation and acceptable power output. A bipolar transistor may be substituted for the JFET in Fig 3-4A. If this is done you may use the base and emitter resistance values shown at C and D of Fig 3-4.

An overtone oscillator is depicted at B of Fig 3-4. Note the similarity of this circuit to that at A. The feedback capacitors have been deleted and a tuned circuit (T1 and C2) has been substituted for RFC1 at A. This tuned circuit is resonant near the frequency of the desired crystal overtone (3rd, 5th, etc.). The tuned circuit must be resonant just slightly above the desired frequency in order to ensure oscillation. High-impedance output from this circuit may be taken from the FET drain through a small value coupling capacitor (XC = 450 ohms). The output link winding of T1 provides low-impedance coupling to the load -- 50 to 600 ohms, depending upon the number of turns used for the link.

A Colpitts oscillator is presented in Fig 3-4C. C1 and C2 are the feedback capacitors. C1 may sometimes be eliminated if the transistor used for Q1 has relatively high base-emitter (internal) capacitance. This may be determined through experimentation. A JFET or MOSFET can be used for Q1 at C of Fig 3-4. The 15K-ohm bias resistor is removed and the 5.6K-ohm resistor is replaced by a 100K-ohm unit if this is done. Output from the Colpitts oscillator is low, owing to the takeoff point being the emitter. An amplifier is needed after Q1 to develop output power that is similar to that at A, B and D of Fig 3-4.

A tuned collector oscillator is shown at D of Fig 3-4. Cl is the feedback capacitor for this fundamental oscillator. It works in combination with the internal base-emitter capacitance of Q1. This type of oscillator may be operated also at the harmonics of the crystal if C2 and T1 are tuned to the desired harmonic. If this is done it may be necessary to use a feedback divider of the kind shown at C in Fig 3-4. Harmonic operation of this circuit is not especially desirable because the output energy contains fundamental crystal energy, plus unwanted harmonics of the crystal frequency. An overtone oscillator (circuit B) is free of fundamental output energy.

All of the Fig 3-4 oscillators may be used as the heart of a QRP transmitter or as heterodyne oscillators in a superhet receiver or converter.   The circuit at C in Fig 3-4 may be used by itself as a milliwatt QRP transmitter. You may key the +12-V supply line or disconnect the 470-ohm emitter resistor from ground and key the circuit at that point.

Variable Crystal Oscillators

A variable crystal oscillator (VXO) offers relief from being locked to one crystal frequency. Inductive reactance (XL) and capacitive reactance (XC) are introduced in series with the crystal to shift its frequency of oscillation. The XC or XL can be made variable to allow shifting the operating frequency.

Fundamental crystals are used in VXO circuits, although you may use an overtone crystal at its fundamental frequency for VXO operation. Plated AT-cut quartz crystals are the best for VXO circuits. I prefer the ones that are contained in the larger HC-6/U metal holders. These crystals appear to provide the greatest frequency shift in a VXO. Surplus FT-243 crystals and other older crystals in nonmetallic holders do not shift frequency too well. In fact, they may be difficult to make oscillate in some oscillator circuits. These crystals are frequently referred to as being "sluggish." It may be helpful to ask your favorite crystal manufacturer if he can supply you with a crystal that is ground especially for VXO circuits. At one time, at least, International Crystal Mfg. Co. made a crystal of this kind. It was a bit more "rubbery" than standard AT-cut models.

You may be wondering how many kHz a crystal may be pulled or shifted in a VXO. This depends upon many factors. The nature of the crystal you use plays a role in this matter. The amount of stray circuit capacitance is part of the picture. The values for the XC and XL dictate to a large measure the maximum shift in frequency. Most importantly, however, is the crystal operating frequency. The lower the crystal frequency the smaller the frequency change. If you use, say, a 3.5-MHz crystal in a VXO you will do well to shift the frequency more than 1.5 kHz. Conversely, a 14-MHz crystal can usually be pulled some 15 kHz. A frequency change of 7 to 8 kHz is typical at 40 meters in a well designed VXO. I have successfully shifted a 20-MHz crystal 30 kHz.
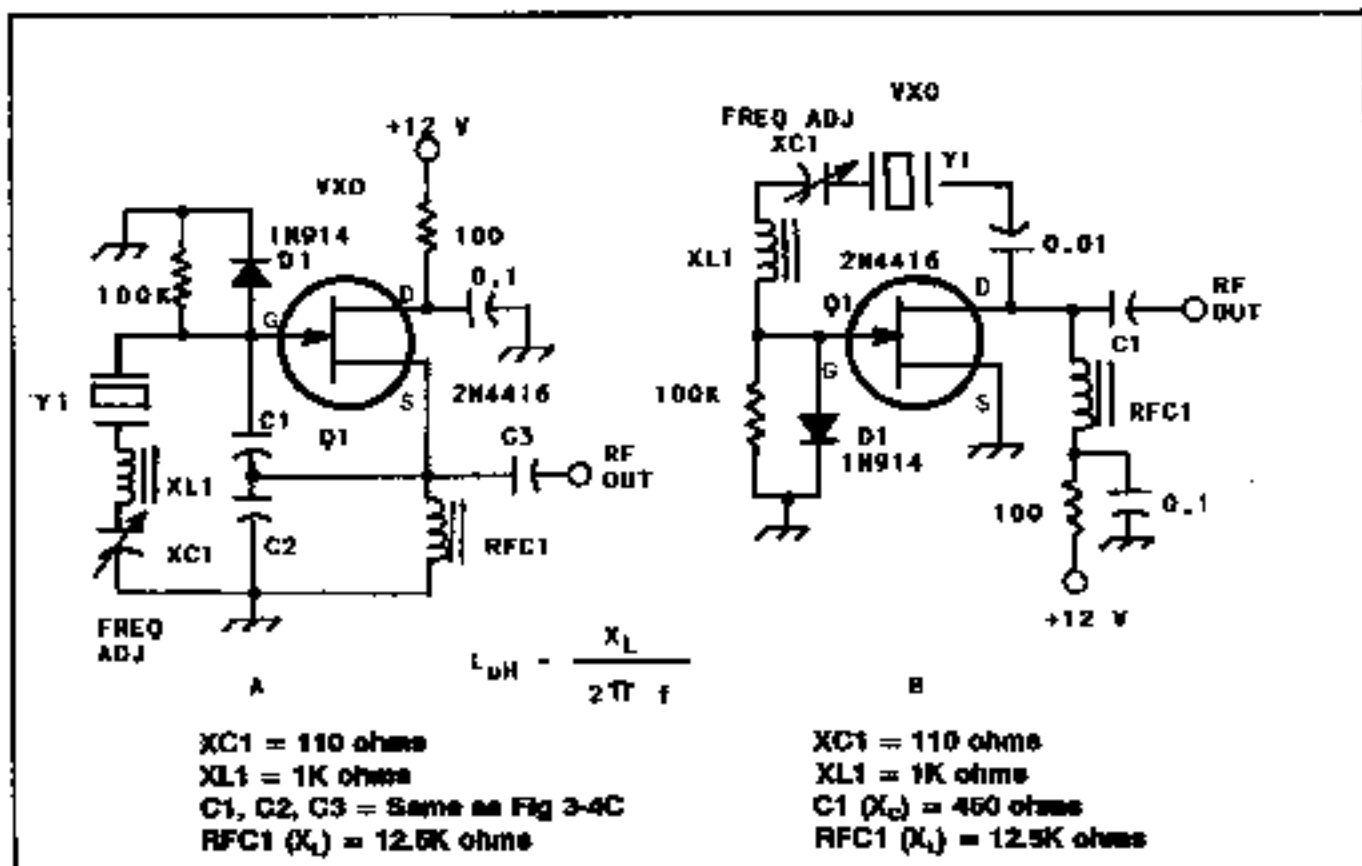
Fig 3-5 -- Two practical examples of VXOs. Circuit A follows the Colpitts
design of Fig 3-4C. The VXO at B is patterned after the Pierce oscillator
in Fig 3-4A. XL1 in both examples is approximately 15 uH at 14 MHz. At 1.8
MHz XL1 is approximately 100 uH. XC1 in both circuits is an air variable
capacitor that is panel mounted for easy adjustment. XC1 at B must have
its rotor and stator insulated from ground.

The values of the critical components are approximate. Again, there
is plenty of leeway for experimentation toward optimizing the osc-
illator performance. Be aware that all VXOs yield nonlinear response
when XC1 is adjusted. Maximum frequency change occurs toward the
minimum-capacitance end of the XC1 rotation. Increasing the maximum
capacitance of XC1 provides diminishing returns with regard to freq-
uency change.

The temptation usually exists to enhance the frequency range of a
VXO by increasing the inductance of XL1 in Fig 3-5. Again, a point
of diminishing returns is reached, respective to true VXO operat-
ion. Too great an inductance leads to regular VFO operation, and
the oscillator stability decreases. A VXO is normally as stable as
a straight crystal oscillator at a constant room temperature.

Too much stray circuit capacitance limits the upper range of VXO
adjustment. In fact, it can cause the highest operating frequency
to occur slightly below the marked crystal frequency. A good VXO
should enable you to shift the crystal frequency slightly above the
marked frequency (1.5 kHz at 14 MHz, for example). The circuit at
B of Fig 3-5 is capable of this function, whereas circuit A is not.
This is because C1 and C2 provide a set minimum capacitance. This

can be overcome by locating a small-value coupling capacitor between the Q1 gate and Y1. Caution: Too small a capacitor value here will prevent the circuit from oscillating.

XL1 in both circuits can be a miniature RF choke. The nearest standard value to that which you compute from the XL equation will be okay in these circuits. I use a 15-uH choke at 14 MHz and a 22 uH unit for 7 MHz.

If you employ the circuit of Fig 3-5B it will be necessary to keep the XC1 rotor and stator above ground. An insulated shaft coupler will prevent hand capacitance from shifting the oscillator frequency. You may use a trimmer capacitor for XC1 if you do not contemplate numerous frequency changes during a single QSO.

Best VXO stability is had by using a buffer-amplifier stage after the oscillator. This helps to isolate the VXO from load changes that can cause frequency shifts.

Transistors other than 2N4416 are suitable in these circuits. You may wish to use, for example, an MPF102 or a dual-gate MOSFET, such as a 40673 or 3N211. The gates of a dual-gate MOSFET can be tied together to allow the device to serve as a single-gate FET. This eliminates the need to apply forward bias to gate no. 2. There is no reason why you can't use bipolar transistors for the circuits in Fig 3-5. If you do, follow the biasing procedure set forth in Fig 3-4. The 2N4416 JFET has a better pinchoff characteristic than does an MPF102. This characteristic allows the transistor to produce greater output power than if a low-pinchoff device is used.

Standard LC VFOs

Frequency stability is the first consideration when we build VFOs. There are two kinds of drift to consider: (1) Short-term drift and (2) long-term drift. The former phenomenon takes place during the first two or three minutes after operating voltage is applied. Long-term drift may continue for many minutes, or even for hours.

Short-term drift is caused by the heating of the transistor junction, plus heating of the other VFO components by RF currents that flow through them. Long-term drift may be caused by an extension of short-term drift which is compounded by changes in ambient temperature within the VFO compartment. Changes in humidity contribute also to long-term drift.

Abrupt frequency changes can also occur in a VFO. This is usually caused by mechanical devices, such as trimmer capacitors, that cause changes in value from vibration. Changes in operating voltage may also cause frequency jumping.

Minimizing Drift

All LC types of VFOs exhibit some drift. Normally, in a well-designed VFO, this drift practically ceases after 30 minutes of operation.

I have seen VFOs that settle down within 10 minutes after turn-on. Typically, a stabilized VFO will ramp up and down in frequency by 10 or 15 Hz, even after a complete warmup. But, this does not cause a problem, since we don't notice any outward effects from so small a frequency change. In a worst case example, I have built VFOs that never stopped drifting. It was not uncommon to observe a 15-kHz drift at 7 MHz during a 1-hour period!   Here are some drift-preventive measures that I find helpful:

(1)   Use a regulated low operating voltage for the oscillator. The lost output power can be recouped by adding a class A amplifier after the oscillator.   Try an operating voltage of 6 to 8. Use a Zener diode to regulate this voltage.

(2)   Use very light coupling between the transistor and the tuned circuit. This helps to isolate the transistor from the tuned circuit. Minimize the coupling from the oscillator to the load by again using a small-value capacitor.

(3)   Utilize temperature-stable capacitors in the critical parts of the oscillator circuit. NPO (zero temperature coefficient) ceramic capacitors are excellent. Polystyrene capacitors are nearly as good as NPOs up to approximately 10 MHz. Silver micas are the worst of the lot. They may exhibit positive or negative drift traits with changes in temperature.

(4) Use a high-Q tuned circuit in your VFO. This minimizes wide-band noise output.

(5)   Install a small-signal silicon diode (Fig 3-5) between the FET oscillator gate and ground. The diode operates as a bias stabilizer, and this reduces potential drift. Harmonic output is lower when using this diode, and this is also a benefit.   This technique is not applicable to bipolar-transistor oscillators.

(6)   All trimmer capacitors should be small air variables or NPO -ceramic units. Avoid using mica trimmers or low cost plastic trimmers.

(7)   Air variables used for the main-tuning control in a VFO need to be the double-bearing type (bearing at each end of the rotor). Avoid using capacitors -with aluminum plates, since they expand and contract with changes in temperature. Plated brass vanes are best. The main tuning capacitor should turn easily with minimum torque.

(8)   Don't skimp on the number of buffer stages after the oscillator. The greater the number (up to three) the better the isolation from varying loads that cause frequency changes.

(9)   Enclose your VFO in its own shield compartment and decouple the dc leads that enter the box. This prevents stray RF energy from entering the VFO circuit and causing frequency shifts. Best results will be had if the VFO module is used outboard from the main piece of equipment (not always practical).

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(10)   Do not use double-sided PC boards for VFOs. The PC conductors form capacitors with the ground-plane side of the board, and the board insulating material serves as the dielectric. Capacitors thus formed are very unstable with changes in temperature. Also, stress on the PC board changes these unwanted capacitances. Use single-sided, glass-epoxy board material for VFOs.

(11)   Avoid using long PC-board traces (conductors) in critical parts of the VFO circuit, especially around the tuned circuit. Long conductors -- especially very narrow ones -- introduce stray induct-ance that becomes part of the tuned circuit. This can not only lower the tuned-circuit Q, but leads to frequency instability if the PC board expands and contracts from heat. Abrupt frequency changes may result if the PC board is stressed.

(12)   Use at least two coats of Polystyrene Q Dope (TM) on the tuned circuit coil after it has been wound in final form. This will keep the coil turns in a fixed position, which minimizes frequency shifts. The protective coating also keeps moisture and dirt from affecting the coil.

(13)   Avoid, if practicable, the use of any magnetic core inductor in the VFO tuned circuit. Rigid air-wound coils will yield the best stability traits. Powdered-iron and ferrite cores change permeability with temperature changes, and this causes drift. If you must use a core (toroid or coil slug), try to utilize no. 6 (yellow code) powdered iron. It is relatively stable with regard to heating. This material is carbonyl SF. When using slug-tuned coils try to have as little of the slug within the coil winding as possible. This lessens the heat-caused frequency drift. Also, the core should enter the high-impedance end of the inductor.

(14)   Whenever practicable, do not remove the operating voltage from your VFO, even during standby periods. This will eliminate the short-term drift that occurs when the oscillator is first turned on. If the VFO creates an interfering signal during receive, add a VFO offset circuit that will shift the VFO frequency during the receive period. This is illustrated in the universal VFO that appears later in this chapter.

(15)   Beware of some of the imported NPO capacitors. I have used a number of them that exhibit considerable capacitance change from heating. I have not experienced this problem with American NPO units.

The foregoing is a recipe book approach to VFO stability. You should have few problems if you adhere to these general "Band Aids" (TM).

Common VFO Circuits

Fig 3-6 shows various popular LC-oscillator circuits, along with methods for tuning them. It should be stressed that mechanical tuning (with air variable capacitors) is superior to VVC (voltage variable capacitance) diode tuning. These VVC or varactor diodes contribute

frequency instability. This is because additional semiconductor junctions are introduced to the circuit, and the junction capacitance changes with temperature. This is by no means an indictment of VVC diode tuning. It is merely an advisory. Certainly, VVC diodes have their place in amateur design, and are far less expensive (and more compact!) than most quality air variables. When these diodes are used for VFO tuning, make certain that you use a high-quality carbon composition potentiometer for varying the VVC operating voltage. I prefer a 2-watt Allen Bradley or equivalent control for this job. Most imported, low-cost controls are mechanically inferior, and the carbon element wears out quickly.

$$C_T = \left[\frac{1}{\frac{1}{C4} + \frac{1}{C6} + \frac{1}{C7}} \right] + \left[C1 + C2 + C3 \right]$$

SUGGESTED VALUES

XC1 = 1200 OHMS
XC2 = 3500 OHMS
XC3, XC4 = 880 OHMS
XC5 = 880 OHMS
XC6, XC7 = 90 OHMS
XL1 = 50 OHMS
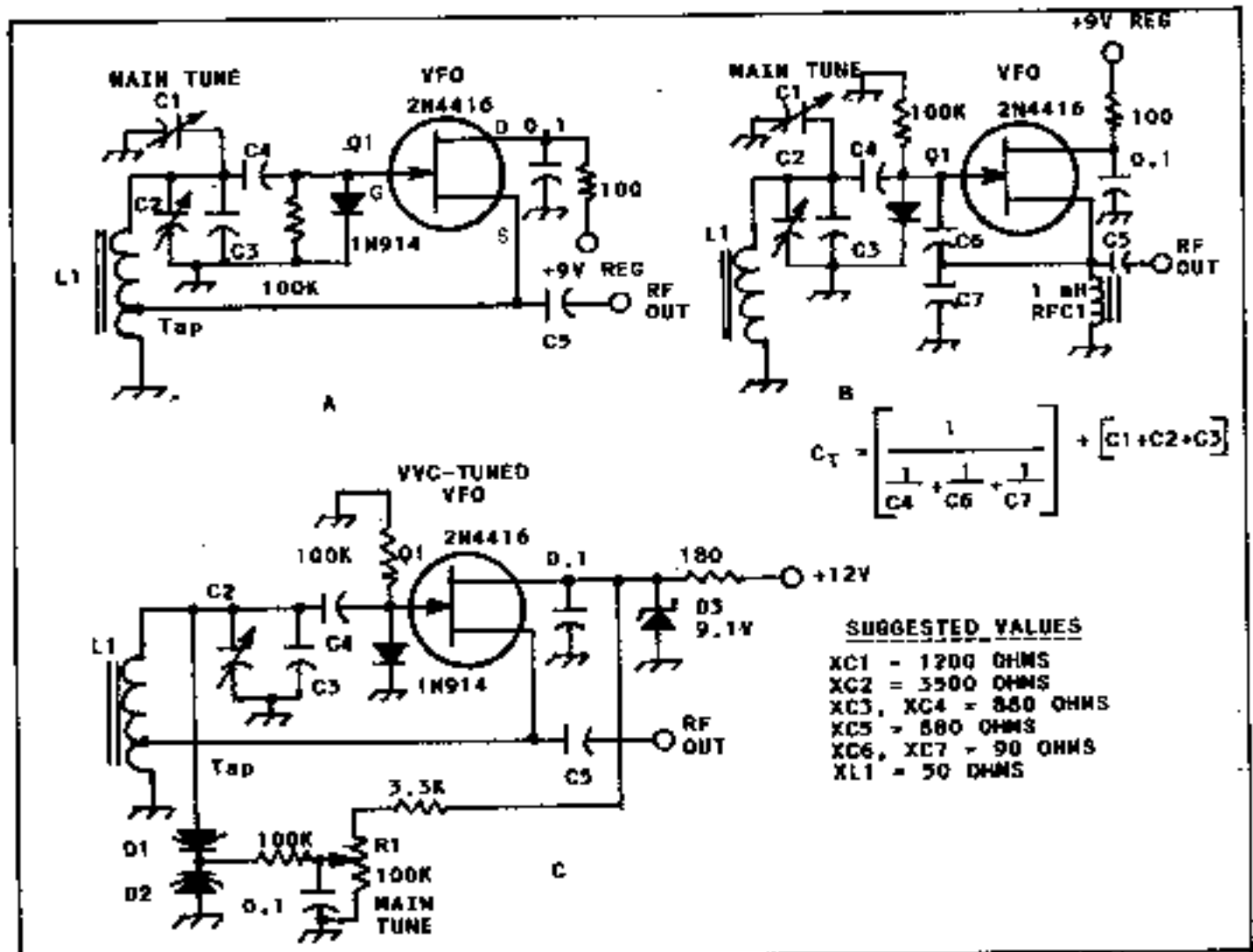
Fig 3-6 -- Examples of common VFOs. Circuits A and C use a tapped coil (L1) to provide feedback. The tap is 0.25 the total L1 turns above the grounded end of L1. Mechanical tuning (C1) is used for frequency control. Gate bias is stabilized by means of the 1N914 diode. In all examples above, C2, C3, C4 and C5 are NPO ceramic. Circuit B shows C6 and C7 rather than a coil tap for obtaining feedback. C6 and C7 are polystyrene. VVC tuning is used at C to change frequency. MV104 or similar VVC diodes may be used. R1 is an industrial grade, 2-W carbon-composition control. A vernier mechanism for R1 provides a slow tuning rate. C2 in all circuits is an NPO trimmer.

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The Hartley VFO circuit appears to be the most popular one among amateur builders. Therefore, I have featured this circuit in Fig 3-6. Each of the circuits needs to be followed by at least two buffer amplifiers in order to provide isolation from the overall VFO load. Power output from the source terminal of the oscillators is low. The buffer/amplifiers increase this power to usable levels. You may obtain the capacitor values in uF by using the formula in Eq 3-1. These values are given in XC on the diagram (Fig 3-6) and represent starting points for VFO design. The value of C1 at A and B must be chosen to provide the tuning range you desire. Depending upon your individual needs, C1 may need to have more or less total capacitance than the XC value indicates. In a like manner, the VVC diodes you select for D1 and D2 of Fig 3-6C will determine the overall tuning range of the VFO. These diodes are available from Motorola and others with various capacitance minimum and maximum values.

Q1 in each example is a 2N4416 JFET. You may use other FETs, such as the MPF102, but output will be lower than with a 2N4416, owing to the superior pinchoff characteristic of the latter type. These circuits can be built around bipolar transistors as well. Forward-bias networks and emitter bias must be added when using bipolar transistors. I recommend a fairly "hot" bipolar transistor for VFO service. The 2N5179 and others with an fT in the UHF region are suitable. On the other hand, I have had good results with such transistors as the 2N3904, 2N2222 and 2N4400 in VFO circuits. Drift will be minimized if you use bipolar transistors with the least input and output capacitance, hence the recommendation for the 2N5179 device. FETs with high transconductance are best for use in VFOs. Certainly, dual-gate MOSFETs are excellent devices for VFO service. Among them are the RCA 40673, 3N211 and 3N212 transistors. A typical JFET or MOSFET has an input capacitance of approximately 6 pF. This must be added to the overall tuned circuit capacitance when calculating the inductance of L1 in Fig 3-6. Allow 10 pF (average) for stray circuit capacitance around the VFO tuned circuit.

The VFOs described in this chapter are suitable for receivers and transmitters. This information will not be repeated later.

Buffering the VFO

I mentioned earlier that buffering provides isolation between the oscillator and the load. Various circuits may be used, but narrow-band circuits (resonant tuned circuits) are best avoided. They may encourage VFO frequency "pulling" and tend to limit the overall bandwidth of the VFO frequency range. The exception to the foregoing philosophy is when we choose to add a harmonic filter at the output of the last buffer amplifier. A low-pass type of filter is most often used. It does not restrict the VFO overall bandwidth.

The buffer-amplifier system is used also to increase the output power of the oscillator. You must determine how much local-oscillator power is required for the mixer or transmitter input stage when designing your buffer-amplifiers. Additionally, knowledge of the input impedance of the mixer or transmitter stage is important. The output of

the last buffer, ideally, is tailored to match this impedance. Maximum power transfer occurs only when a matched condition exists. Most bipolar transistor class A transistor amplifiers present an input impedance of approximately 600 ohms. A diode balanced mixer, on the other hand, has a 50-ohm input impedance. The gates of FETs have a high impedance (megohms). The gate resistor establishes the effective gate impedance. Thus, if the gate resistor is 100,000 ohms, that is the effective gate impedance. Fig 3-7 illustrates how a buffer-amplifier chain may be structured.
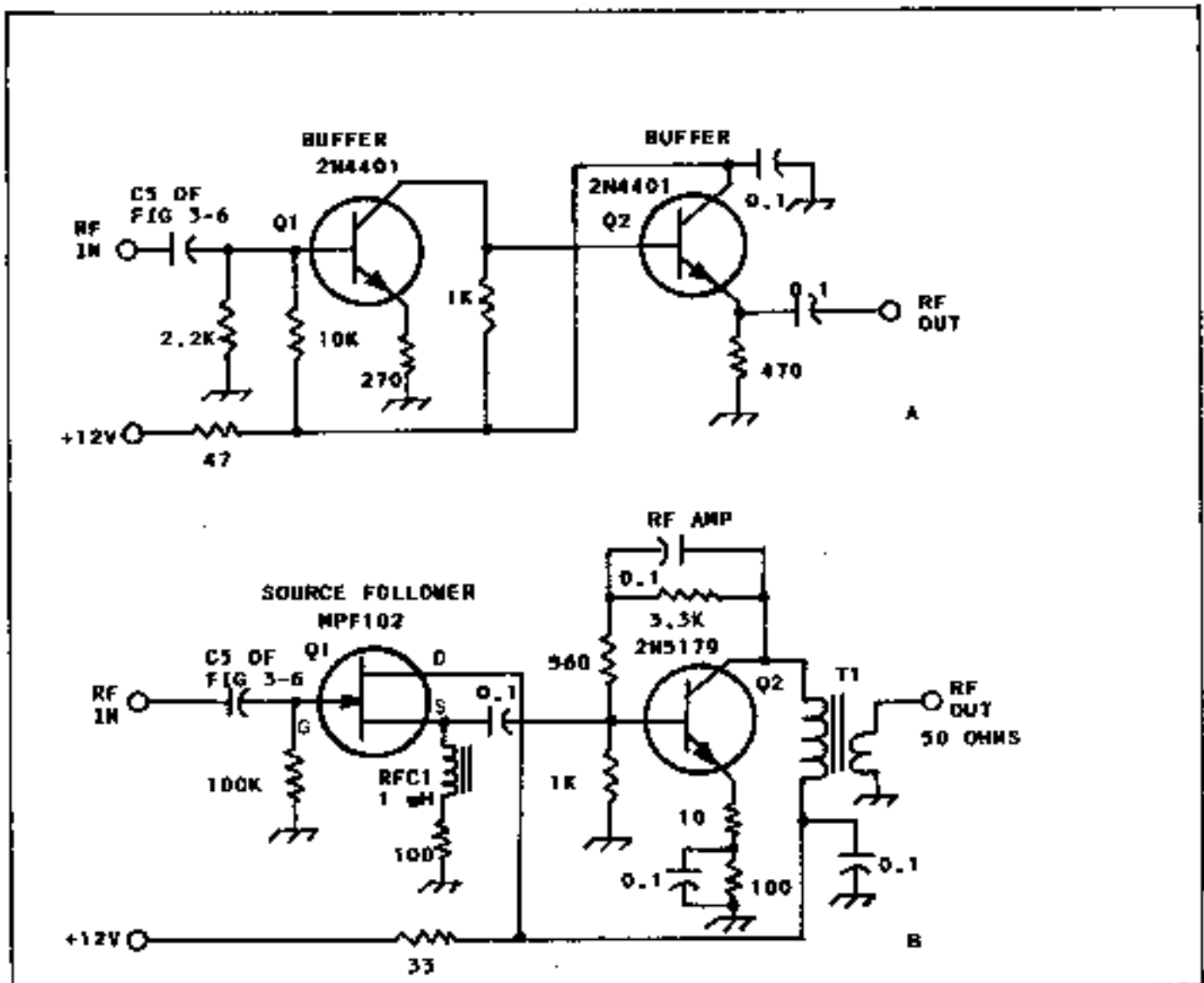
Fig 3-7 -- Two examples of VFO buffer/amplifier circuits. A direct-coupled system is shown at A. Transistors such as 2N2222A and 2N3904 may also be used for Q1 and Q2 at A. This system has a few dB of voltage gain and has an output impedance slightly less than 470 ohms. Circuit B has good isolation by virtue of FET Q1, which presents high Z to the VFO. Q1 has a gain of 0.9. Q2 has approximately 15 dB of gain. T1 has 16 turns of no. 28 enan. on an Amidon FT-37-43 ferrite toroid. Secondary has 8 turns.

A direct-coupled buffer section is illustrated at A of Fig 3-7. This simple, inexpensive approach to buffering a VFO is preferred by many builders, especially when a compact layout is required. It is broad-band and may be used from 500 kHz to 30 MHz with no circuit·changes.

Fig 3-7B shows the circuit I prefer for post-VFO use. FET Q1, because of its 100K-ohm input impedance presents light loading of the VFO output circuit. Output taken from the source of Q1 is routed to a broadband class-A amplifier (Q2) that exhibits a 50-ohm input imped-ance. This amplifier uses shunt and degenerative feedback to ensure stability and a flat frequency response from LF into the VHF spectrum. Harmonic output is low because Q2 operates linearly. A low-pass filter may be used at the output of T1 if further harmonic atten-uation is desired. Additional information about VFO buffer/amplifier systems is presented in **W1FB's Design Notebook** and **Solid State Design for the Radio Amateur** (both ARRL books).

Regulated dc voltage is sometimes used for Q1 at A and B in Fig 3-7. This helps to ensure a constant load at the VFO output: Even small variations in Q1 operating voltage can lead to phase shifts that are reflected to the VFO, and these changes cause shifts in frequency.

The Heterodyne VFO

Greater stability and higher cost are conditions that go with the more elaborate heterodyne type of VFO. These circuits require a VFO, mixer and heterodyne-oscillator system. Since two frequencies are involved in this style of VFO, we must use care to avoid passing spurious frequencies to the receiver or transmitter. Bandpass filter-ing is generally required after the mixer. We face greater expense because crystals are used in the heterodyne oscillator.

A practical circuit for a heterodyne VFO is shown in Fig 3-8. Design information for the tunable oscillator may be obtained elsewhere in this chapter. An RCA CA3028A IC is used as a singly balanced mixer. Q1 and Q2 are class A broadband linear amplifiers. Pierce oscillator Q3 provides the heterodyne frequency which, when mixed with the VFO frequency, yields U1 output from 7.0 to 7.3 MHz. Q1 amplifies this energy. It is filtered by bandpass filter FL1, then further amplified by Q2. A 7-MHz sine wave appears at the output port of T5.

You may modify this circuit for multiband use by having additional heterodyne oscillators and harmonic filters. These extra circuits may be band-switched by mechanical or electronic means. Fig 3-9 shows how the circuits can be switched with 1N914 diodes.

I mentioned earlier that this variety of VFO is more stable than the more common straight VFO. 1 need to qualify the statement by saying that any VFO is only as stable as its tunable oscillator. The improved stability comes from not using a simple VFO at freq-uencies above, say, 7 MHz. Most VFOs that operate below 7 MHz can be made to operate with minimum drift. A 14- or 21-MHz VFO, usually does, in contrast, drift appreciably unless great care is taken.
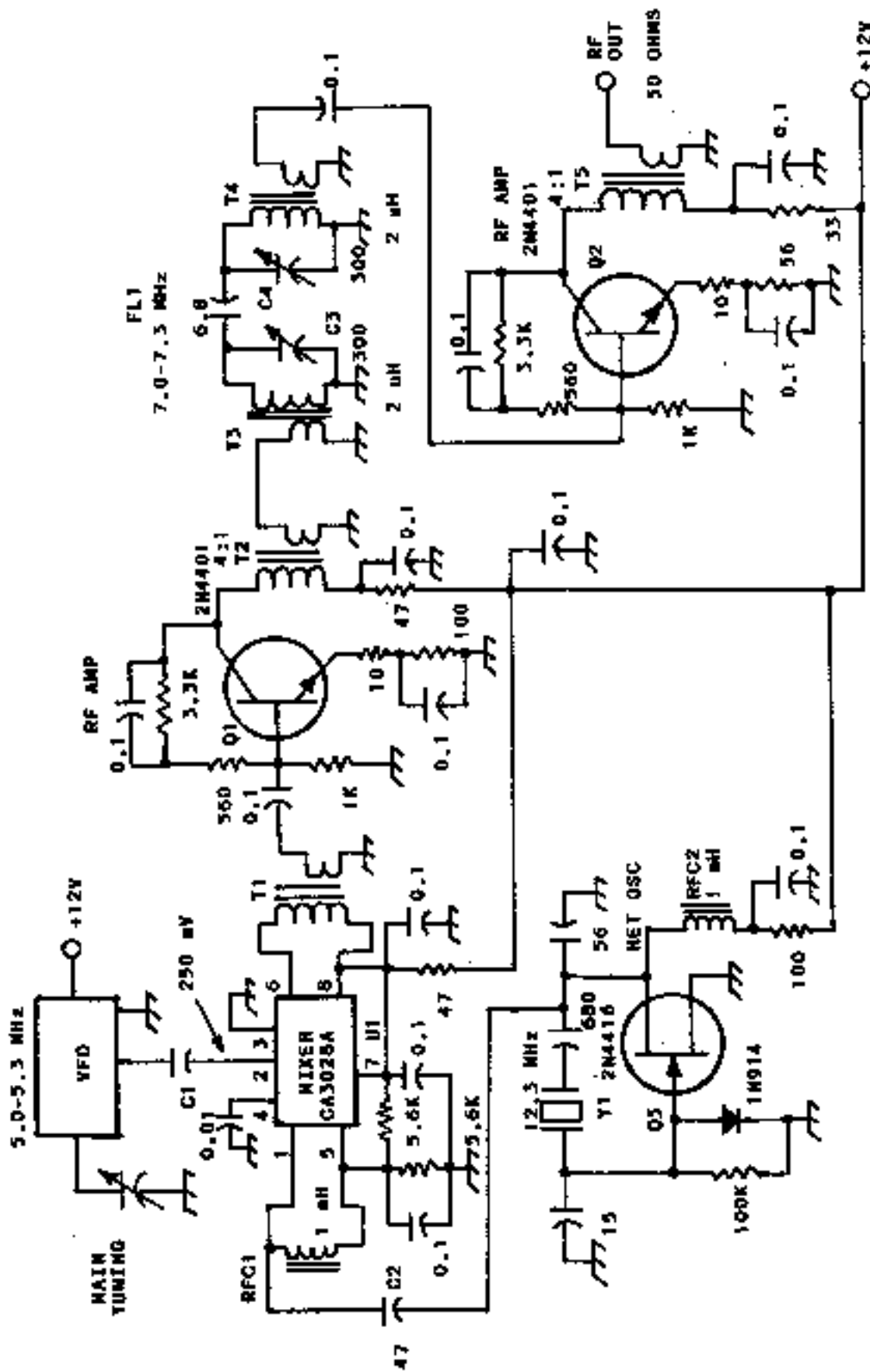
Fig 3-8 -- Practical circuit for a 40-meter heterodyne VFO. Fixed-value capacitors are disc ceramic. Resistors are 1/4-W carbon composition. C1 and C2 are chosen to provide 250 mV at pins 1 and 2 of U1. T1 has 18 turns of no. 28 enam. on an FT-37-43, with 6-turn sec. T2 and T5 have 16 turns of no. 28 enam. for primary and an 8-turn no. 28 sec. on an FT-37-43 toroid. T3 and T4 have 22 turns of no. 24 enam. on a T50-6 toroid. Links have 3 turns of no. 24 enam. wire. C3 and C4 are 500 pF mica trimmers. See text for additional information about this circuit.
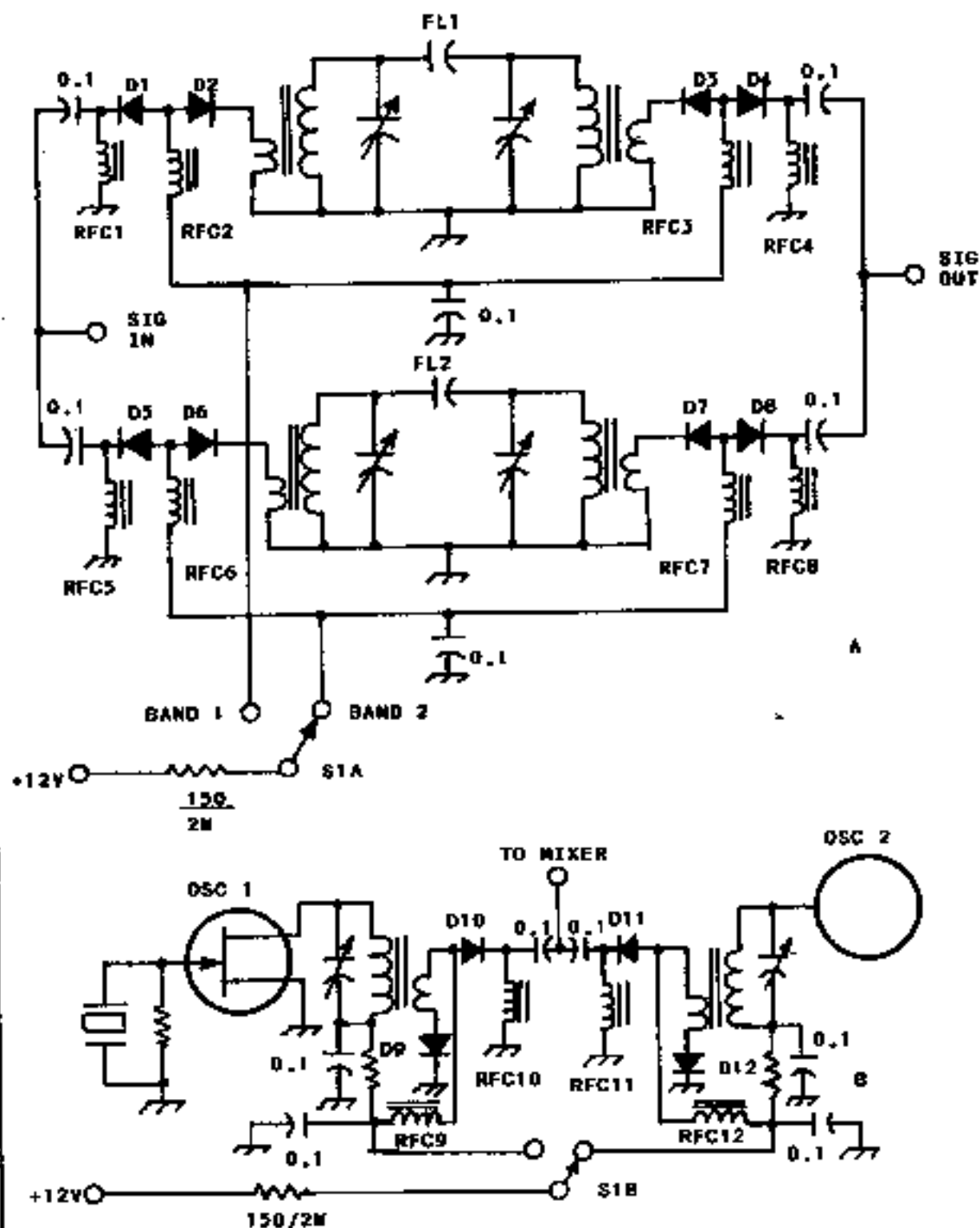
Fig 3-9 -- Two examples of diode switched RF circuits. Filter switching is shown at A. Oscillators may be switched as shown at B. All diodes are type 1N914. RF chokes are 1 mH. Additional filters and oscillators may be added by merely duplicating the above circuits and adding switch poles. All switching is done at low impedance.

It is important when using the circuits in Fig 3-9 to ensure ample
dc current through the switching diodes. If this is not done, the
diodes do not turn on completely. This results in a signal loss which
can be as great as 3-4 dB. By utilizing back-to-back diodes at each
switching point we obtain better filter isolation than if only one
diode were used for each switch.   The RF chokes specified at each
switch location are used in place of resistors in order to minimize
the dc resistance in the circuits. Miniature molded chokes are avail-
able from Mouser Electronics and Oak Hills Research.

You may include as many filters and oscillators necessary to achieve
multiband heterodyne oscillator service. The diode switching methods
shown in Fig 3-9 are, of course, suitable for many other types of
RF circuits. It is much easier to switch dc lines than it is to use
a mechanical switch for selecting the numerous RF circuits in a multi-
band system. When using diodes we can avoid long runs of RG-174 coax
line between the affected circuits and the switch terminals. It is
entirely practical to use diodes for switching multiple IF filters
into and out of a receiver IF circuit.

Frequency Synthesizers

Although synthesizers and digital frequency display represent the
modern way of life with expensive commercial ham gear, we QRPers
seldom use these techniques. They contribute to bulky equipment and
the associated circuitry consumes substantial power. Neither trait
is in keeping with the QRP theme. It is for these reasons that I
have chosen to exclude frequency synthesis from this book. You can
learn about synthesizers and frequency counters by consulting The
ARRL Handbook.

A Universal VFO

The VFO described in this section is stable and easy to build. It
may be used in any circuit that calls for a variable-frequency osc-
illator. It uses a tapped-coil Hartley arrangement. High- and low-
power output ports are available to accommodate a variety of circuit
requirements. The additional amplifier (for high power) can be dis-
abled by removing two jumper wires from the PC board. The circuit
is presented in Fig 3-10. Although a 2N4416 FET is specified for
Q1, you may use an MPF102 or many other JFETs that are earmarked
for operation at VHF.

Fig 3-10 shows a diode switch (D3 and D4) that serve as a frequency-
offset control. This is a necessary addition when we use the VFO
in a transceiver and need to establish a 700-Hz offset for CW during
transmit periods. C4 is adjusted for the desired frequency offset
when D3 and D4 are turned on. The operating frequency will then shift
lower by the required amount.

Q2 provides isolation between Q1 and class-A linear amplifier Q3.
Output may be taken directly from Q3 for many applications. C12 is
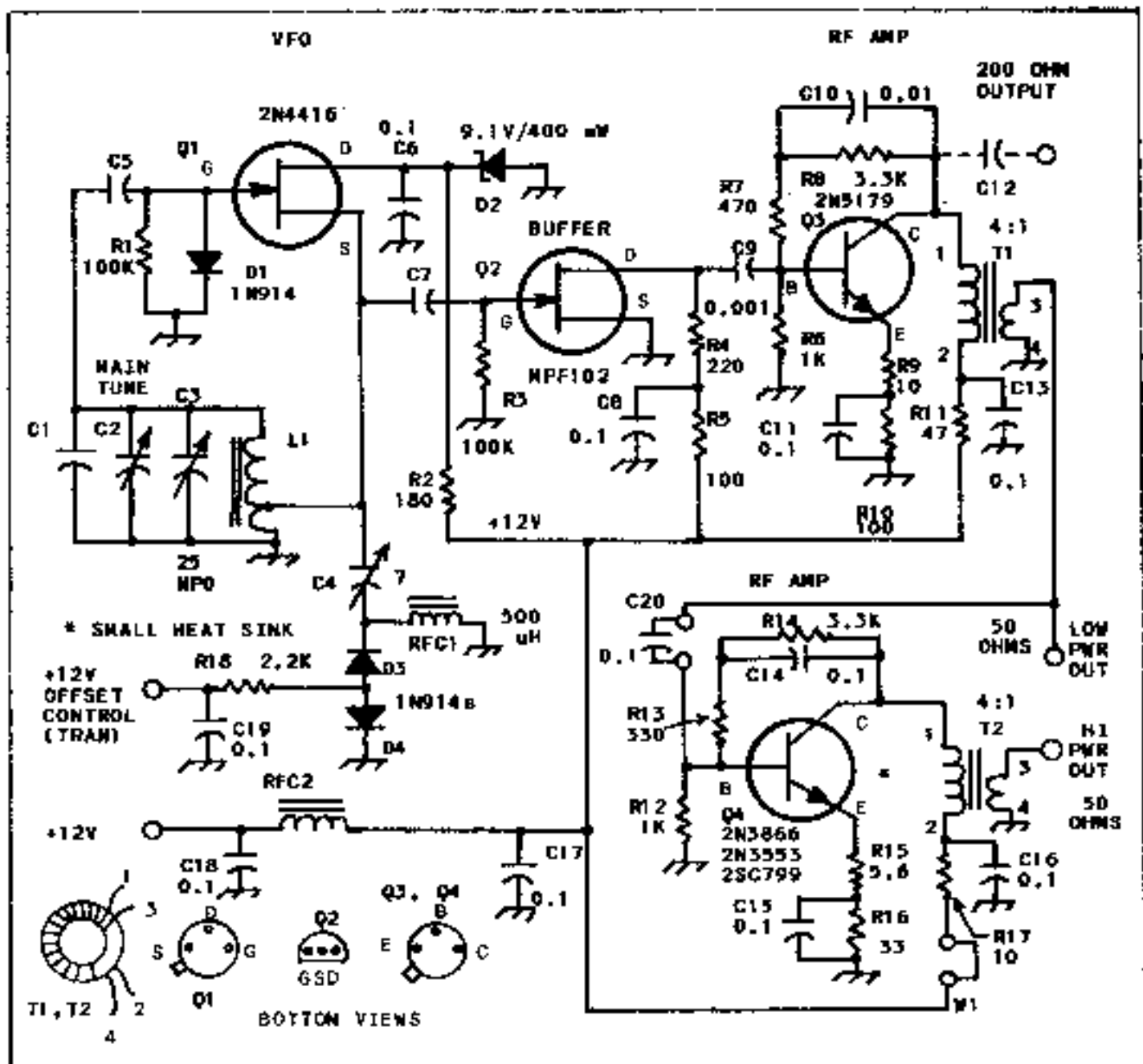
Fig 3-10 -- Schematic diagram of the universal VFO. Fixed-value capacitors are disc ceramic. Resistors are 1/4-W carbon film or carbon composition. See Table 3-1 for frequency-determining component values versus operating frequency. C2 is an air variable. C3 and C4 should be miniature air variable trimmers or NPO ceramic trimmers. RFC1 is a miniature 500-uH choke. For RFC2 wind 10 turns of no. 28 enam. wire on an Amidon FT-25-43 ferrite toroid. T1 and T2 have 16 primary turns of no. 28 enam. wire on Amidon FT-37-43 ferrite toroids (mu = 850). The secondary windings have 8 turns of no. 28 enam. wire. C20 and M1 are additional components for use during high-power operation only. A small press-on heat sink is used on the case of Q4.
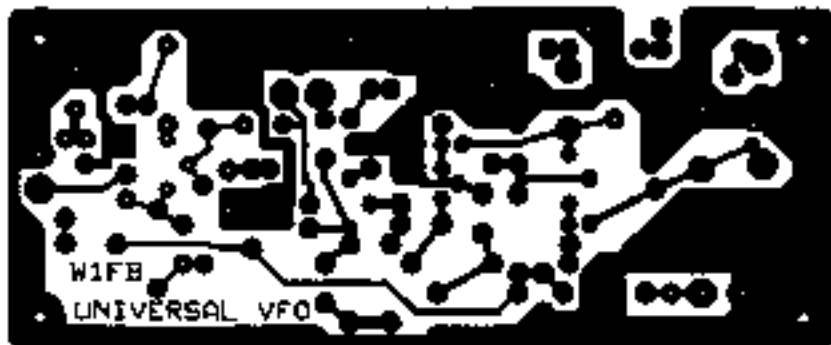
Fig 3-11 -- Etched side of the Universal VFO PC board.
The pattern is shown to scale. Drilled boards are
available from FAR Circuits (see text).
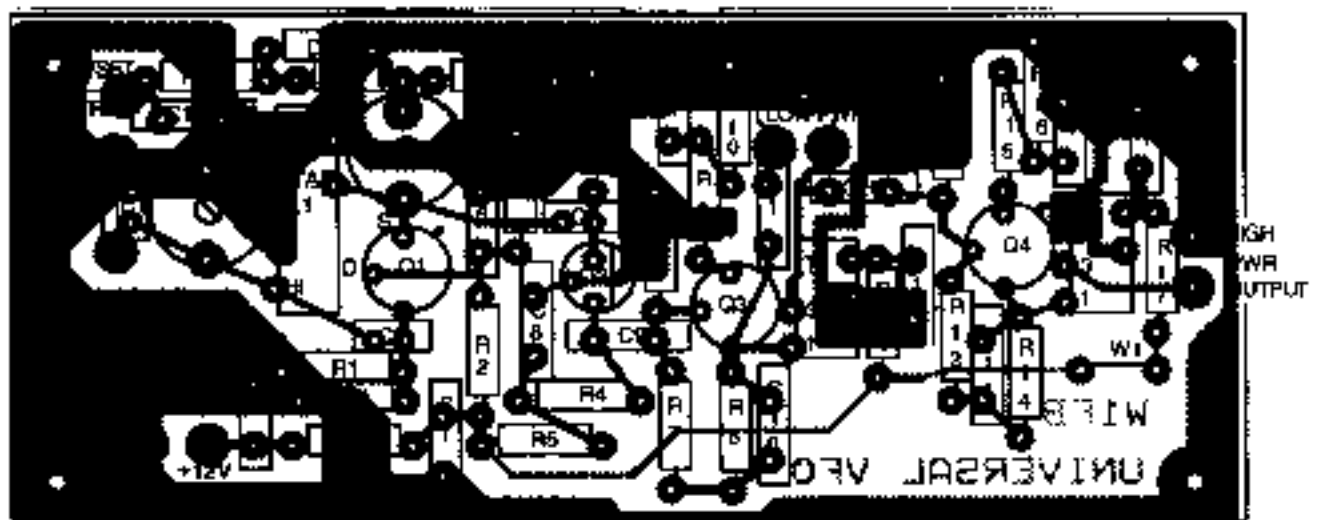
Fig 3-12 -- Expanded X-ray view of the VFO PC board as viewed from the
component side of the board. This pattern is shown at 1.5 times scale.

chosen for the required value of LO injection that is supplied to a mixer or other type of circuit. C12 is used only when the mating circuit has an impedance greater than 50 ohms. The collector of Q3 has an impedance of approximately 200 ohms.

When greater VFO output power is needed you can activate Q4 by means of components C20 and W1. The additional output power may be required when the VFO feeds a diode-ring mixer, for example. Harmonic filtering can be used after Q3 or Q4, depending upon which stage is used as the output one. Drilled and plated circuit boards for this project are available from FAR Circuits, 18N640 Field Court, Dundee, IL 60118. Price: $3.85 plus $1.50 shipping fee.

Construction Notes

This VFO should be contained in a shielded box. C2 is also in the enclosure. A suitable box for the assembly can be fashioned from sections of double-sided PC board. The mating walls of the enclosure are soldered together inside the box. A vernier drive is attached to the shaft of C2 in the interest of smooth tuning. C2 should be the double-bearing type, and it must turn easily in order to prevent backlash.

NPO capacitors are best for use at C1, C3, C4, C5 and C7. These units enhance the VFO stability. Try to avoid using silver micas at these circuit points. Likewise with polystyrene capacitors.

After you finish testing the VFO and establish the desired tuning range, apply a coating of polystyrene Q Dope (TM) or similar low-loss cement to L1. Add a second coat of cement 24 hours later. This will aid VFO stability by affixing the turns to the core. The sealant also prevents moisture from causing frequency changes. Finally, use a generous blob of epoxy cement to affix L1 to the PC board. Use care to avoid getting the epoxy glue on the coil turns. Affix the toroid at the gap between the ends of the winding.

When you tap L1 make certain that you do not end up with shorted coil turns. This can be accomplished by slipping a small piece of paper under the tapped turn to prevent the bare wire from resting against the adjacent coil turns. A physical description of how to tap coils is given in **W1FB's Design Notebook**.

A scale etching template for this project is shown in Fig 3-11, along with a parts-placement guide. Do not use double-sided PC board for this circuit. Glass epoxy board material should be used in preference to phenolic board.

VFO Summary

In the event you have a pet circuit for use in place of Q1 in Fig 3-10, you may still want to utilize the Q2, Q3 and Q4 circuits. They provide excellent VFO buffering and provide two power-output options.

Table 3-1

f(MHz)	C1(pF)	C2(pF)	C5, C7(pF)	L1
1.8-2	220	100	100	24 uH. 71 turns of no. 30 enam. on a T68-6 toroid core. Tap at 18 turns.
3.5-4	150	50	68	9.5 uH. 44 turns of no. 26 enam. on a T68-6 toroid core. Tap at 11 turns.
5-5.5	130	50	47	5 uH. 33 turns of no. 24 enam. on a T68-6 toroid core. Tap at 8 turns.
7-7.3	110	25	47	3.6 uH. 27 turns of no. 24 enam. on a T68-6 toroid core. Tap at 7 turns.

Component information for the universal VFO in Fig 3-10. C3 is a 25-pF NPO ceramic trimmer. C4 is a 7-pF NPO trimmer. The total capacitance of C2 may be reduced to restrict the VFO tuning range. C1, C5 and C7 are NPO ceramic.

A Series-Tuned Stable VFO

The circuit in Fig 3-13 is one I developed in the 1970s. I have used it many times in circuits that require exceptional stability with regard to short- and long-term drift. It takes advantage of the once popular series-tuned Colpitts oscillator that provided excellent stability in vacuum-tube circuits during the 1940s and 1950s.

The high capacitance used in the feedback circuit (C4 and C5) helps to minimize short-term drift caused by changes in junction capacitance when Q1 is turned on. C4 and C5 are part of the overall capacitance for the tuned circuit. Polystyrene capacitors are used in the feedback circuit to help compensate for the positive drift introduced by the powdered-iron core in L1. NPO ceramic capacitors are used for C2 and C3 to enhance the stability.

L1 is a J. W. Miller Co. high-Q 43-series slug-tuned coil. Although this inductor is a bit expensive, it is a good choice in this type of circuit. A solenoidal-wound coil of equivalent inductance, if wound on a ceramic form, would work equally well, but it would be substantially larger than the Miller coil. L1 has considerably greater inductance than would be required when using a parallel-tuned Colpitts oscillator. This offers the advantage of masking the stray

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inductance that exists in the circuit-board conductors and leads
that connect to L1 and C1. This parasitic inductance can reduce the
effective Q of the coil and may cause frequency changes when the
PC board is flexed or expands from heating. The parasitic inductance
is a very small part of the L1 inductance in this circuit.

Coupling between Q1 and source follower Q2 is light (C8) to minimize
oscillator loading. Q3 amplifies the oscillator energy, which is
filtered by the Q3 collector tuned circuit. This circuit is designed
to transform the 1000-ohm collector impedance to a 500-ohm output
impedance for interfacing a bipolar-transistor amplifier or the LO
port of an IC mixer. The network has a loaded Q of 4. You may wish
to change the tuned circuit to provide a 50-ohm output impedance.
Output voltage at 500 ohms is 9 volts P-P (3.2 V RMS). The output
energy is a sine wave.
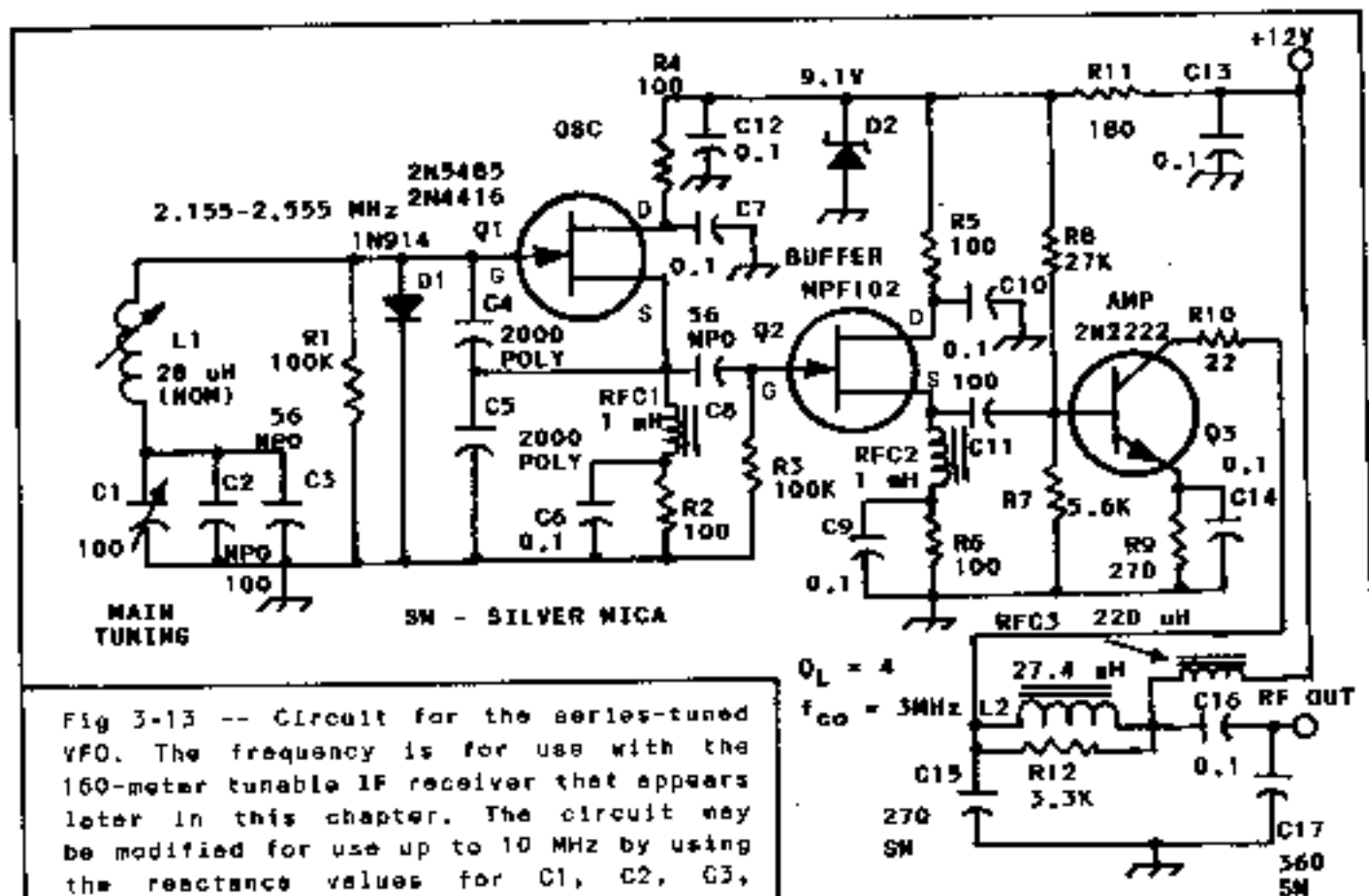
Fig 3-13 -- Circuit for the series-tuned
VFO. The frequency is for use with the
160-meter tunable IF receiver that appears
later in this chapter. The circuit may
be modified for use up to 10 MHz by using
the reactance values for C1, C2, C3,
C4, C5, C8, C15, C17, L1 and L2 to learn
the new component values (approximate).
C1 is a double-bearing air variable. L1 is a J.W. Miller no. 43A335CBI or
equivalent. L2 has 21 turns of no. 26 enam. wire on an Amidon Assoc. FT-
50-61 ferrite toroid (125 mu). RFC1, RFC2 and RFC3 are miniature Mouser
Electronics chokes. Fixed-value capacitors are 100-V disc ceramic unless
otherwise indicated. Resistors are 1/4-W carbon film or carbon composition.
The VFO output impedance is 500 ohms. Output voltage is 8 P-P. Total circuit
current is 35 mA. A PC board is available from FAR Circuits. 18N640 Field
Court, Dundee, IL 60118.
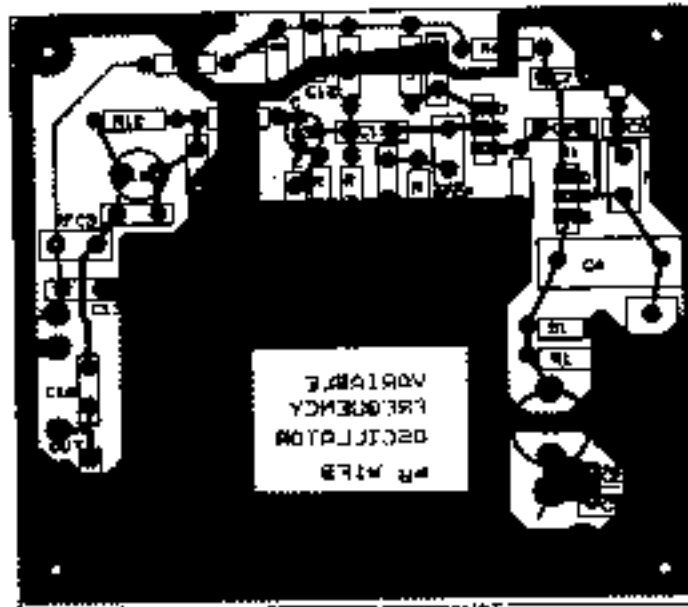
Fig 3-15 -- Parts placement guide for the VFO as viewed from the component side of the board.
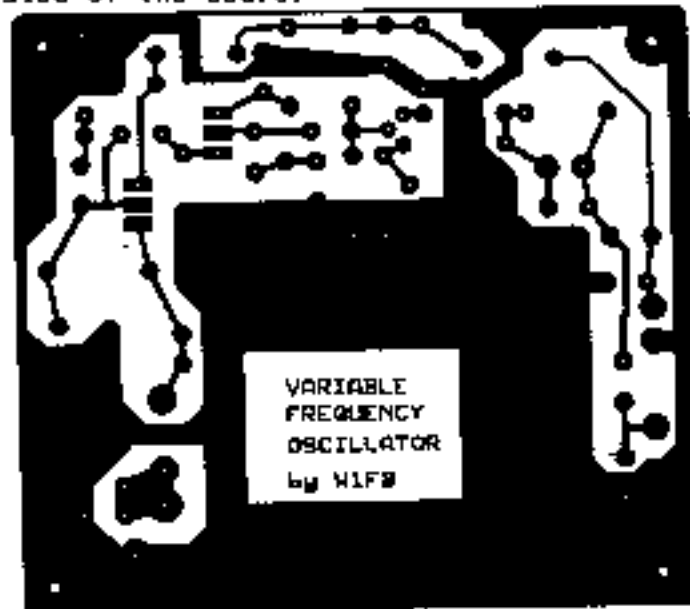
Fig 3-14 -- Scale etching pattern for the VFO as seen from the etched side. Drilled and plated boards for this VFO are available from FAR Circuits for $5.25 plus $1.50 shipping.

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The Fig 3-13 circuit is arranged for operation from 2.155 to 2.555 MHz. This range is required for the 160-meter tunable IF receiver that we shall discuss later in the chapter. The IF for the receiver is 455 kHz and the tuning range is 1.7-2.1 MHz.

The output filter (C15, C17 and L2) is based on pi-network equations that are published in **Motorola RF Device Data**. This book is available from Motorola Semiconductor Prod., Inc., Box 20912, Phoenix, AZ 85036. The equations appear on page 6-29 (network B). The same data is available in Motorola Application Note AN-267. This RF data book is a valuable tool for those of us who design our own QRP transmitters.

Q1 in Fig 3-13 should be a high transconductance JFET with a 400-MHz upper frequency rating. Although an MPF102 FET can be used in this circuit, output will be somewhat below that of the hotter FETs. This is because the 2N4416 or 2N5485 devices have a better pinchoff characteristic, which allows higher output before the FET channel pinches off.

Construction Information

The Fig 3-13 VFO will perform best if it is enclosed in a shield box. I used double-sided PC board material to form my enclosure. C1 is mounted on the front panel of the box. L1 is mounted on one of the side walls (near C2 and C3). The leads from the coil to the PC board should be short and direct. A phono jack is used for the RF output connector. A 0.001-uF feedthrough capacitor serves as the +12-V terminal. A U-shaped aluminum cover is attached to the box walls by means of four no. 4 sheet-metal screws. The VFO board is soldered directly to the inner walls of the box and is elevated above the bottom of the box approximately 3/8 inch. The assembled module may be attached to the chassis by means of four no. 6 spade bolts, or a pair of metal L brackets can be attached to the lower edges of two of the box walls. These brackets are then screwed to the main chassis.

A vernier drive mechanism should be used with C1 to ensure smooth tuning at a reduced rate. The larger of the imported vernier drives is suitable for this application. Japanese vernier mechanisms that have a 0-100 calibration scale are available from Mouser Electronics.

Fig 3-14 contains a scale etching template for this VFO. A parts overlay is offered in **Fig 3-15**.

VFO Wrap-Up

I mentioned earlier that this VFO can be tailored for output impedances other than 500 ohms. The desired output Z is merely cranked into the equations referenced above. Be sure to select a cutoff frequency that is somewhat above the operating frequency. R12 in Fig 3-13 is used to broaden the response of the output network and to prevent Q3 from self-oscillating. VFO output may be decreased by using a larger resistance value at R9.

If you observe a spurious (buzzy) signal in the receiver when you
tune the Fig 3-13 VFO, it is likely that Q3 is self-oscillating
near the VFO operating frequency. It will depend upon how close
the VFO load is to 500 ohms. A quick cure is to bridge a 470- or
560-ohm 1/4-W resistor from the VFO output (C17) to ground. This
will provide a constant load for the Q3 amplifier with minimum loss
of output signal.

As is the situation with the Universal VFO in Fig 3-10, this VFO
can be modified for use on other frequencies up to 14 MHz. Simply
determine the XC and XL values for Q1 and apply them to the desired
operating frequency. The new values will get you close to the range
of interest. Plates may be removed from C1 in Fig 3-13 to reduce
the tuning range of the VFO.   The output filter (Q3 LC network)
will need to be altered for the new operating frequency (see page
56, second paragraph).

IF Amplifiers and Such

Various IF amplifier systems, AGC circuits and audio amplifiers
are described in detail in **W1FB's Design Notebook** and in **Solid State
Design for the Radio Amateur.** These books are part of the QRPers
technical library and provide information that is not repeated here.
In this book we will concentrate more on practical circuits you
can build and use, such as complete receivers and modules that may
be used with them.

An Audio-Derived S Meter

An S meter may be added to a DC or superheterodyne receiver that
has no AGC circuit. This is done by sampling audio energy after
the product detector, amplifying and rectifying it, then applying
the resultant dc voltage to a meter. A practical circuit for an
audio-derived S meter is presented in Fig 3-16.
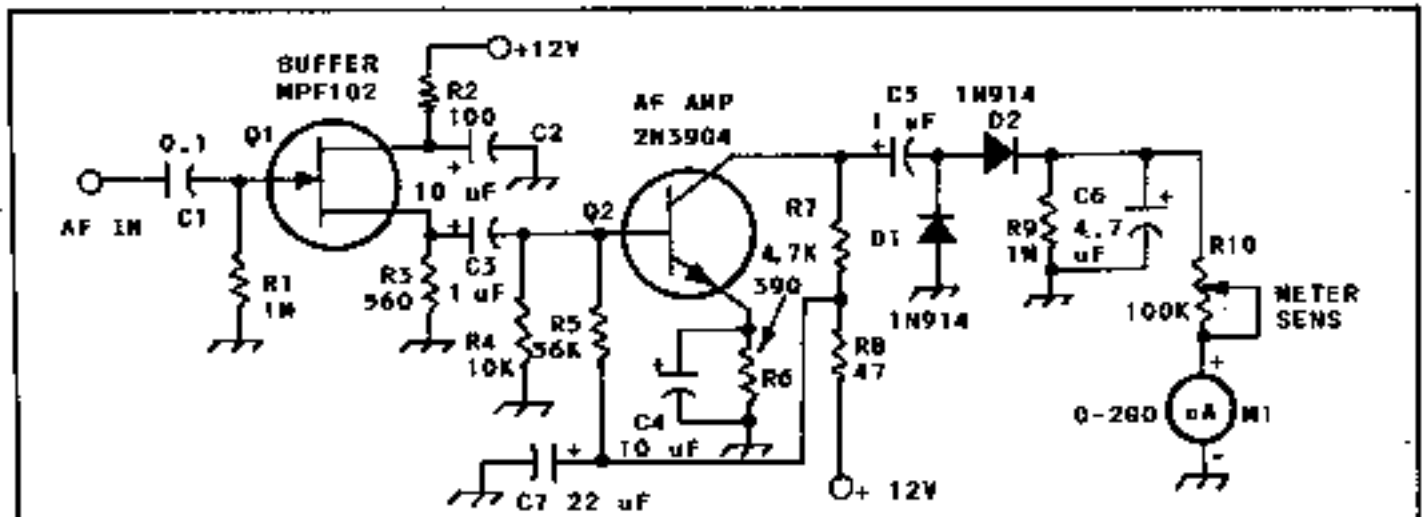
Fig 3-16 -- Practical audio S meter. R10 is a PC-mount control. Polarized
capacitors are tantalum or electrolytic, 16 V or greater. See text for M1
data.

Q1 of Fig 3-16 functions as a source follower isolation stage between the audio pick-off point and amplifier Q2. The 1-megohm Q1 gate does not load the output of the audio stage being sampled. Q2 amplifies the audio energy, which is rectified by voltage doubler D1 and D2. R9 and C6 establish a time constant for M1. The decay time for this circuit may be adjusted to your liking by changing the value of C6. Greater capacitance increases the decay time. Without this network the needle in the meter would gyrate wildly as the audio level changed. R10 is adjusted for the desired amount of meter deflection. I set mine to read S9 for 50 uV of receiver input signal. A calibrated signal generator is required for calibration. Alternatively, you may observe the S meter on your transceiver while monitoring a signal of medium strength. Note the reading, then attach the antenna to your home-made receiver and adjust R10 for the same signal-strength reading. It is important to bear in mind that all S meters provide relative signal-strength readings. Few of them track accurately over a wide range of the scale. Ideally, an S meter would read S9 on all bands for an antenna input signal of 50 uV. Each S unit below S9 would equate to 6 dB and the dB scale above S9 would be accurate. Factually, most S meters give different readings on different bands for a given signal strength. This is because the gain distribution in a receiver is often different from band to band. S meters are useful, however, for observing signal levels when a friend wants to compare antennas, etc. A calibrated step attenuator at the receiver input enables you to tell the other amateur exactly how many dB his signal has increased or decreased during antenna tests. This assumes that QSB is not present!

The circuit in Fig 3-16 may be used with a regenerative, DC or superhet receiver. The audio sampling point for the circuit must be located ahead of (before) the audio gain control. If not, the S meter reading will change with the setting of the gain control.

M1 in Fig 3-16 is a microammeter. Many of the low-cost edgewise signal-level meters found on the surplus market will do the job. Some are available with calibrated scales. Most of these meters have movements that are from 100 to 500 uA. Any meter with a sensitivity of up to 500 uA may be used. You may need to draft your own S-meter scale if you use an instrument with a numbered, linear scale. Fig 3-17 shows the PC-board pattern and parts-placement guide.

There is no reason why you can't use the Fig 3-16 circuit in an RF-derived S-meter application. This would require that you sample the output of the receiver last IF amplifier. In this situation you must change C1, C3 and C5 to units of lower capacitance. C1 becomes a 100-pF component. C3 and C5 would be changed to 0.1 uF. You may want to increase the Q2 amplifier gain by replacing R7 with a 1-mH RF choke, or you may elect to replace this resistor with a tuned circuit for the appropriate IF. If a tuned circuit is used, change C5 to a value between 10 and 56 pF, with the larger value being used for the lower IFs, such as 455 kHz. C4 and C7 should be changed to 0.1 uF. Likewise with C2. No other modification is necessary. You may substitute a 2N2222, 2N4400 or similar device for the 2N3904 (Q2).

Fig 3-17 -- Scale etching template (A) for the S meter
board as seen from the etched side. The artwork at B is
an X-ray view of the etched side of the board showing
the placement of the components. This circuit board is
available from FAR Circuits in Dundee, IL for $3 plus
$1.50 shipping.
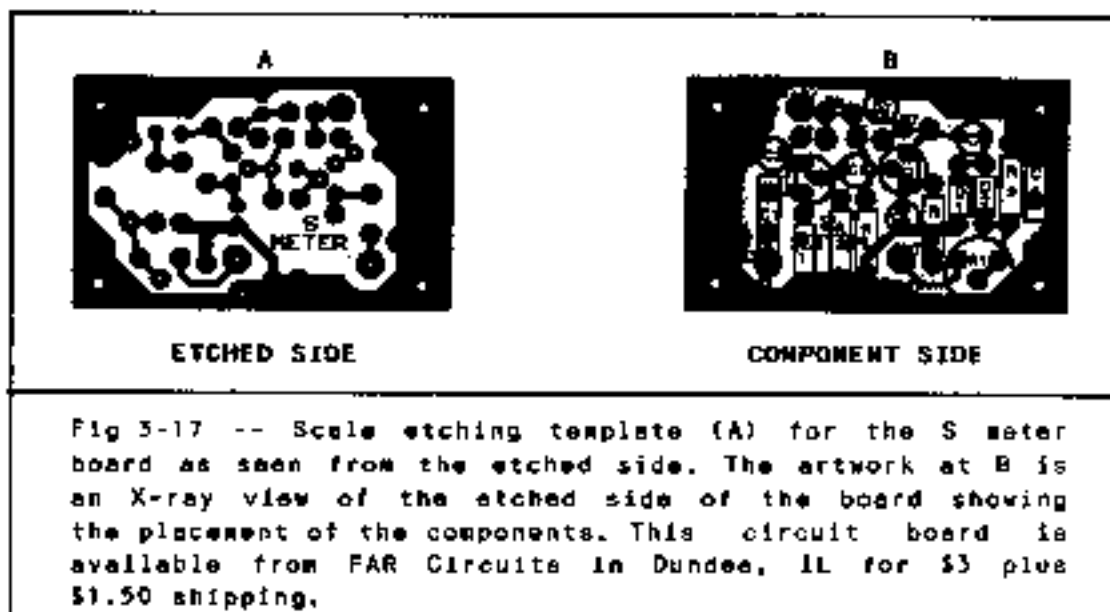
AGC circuits that use op amps and include S-meter circuits are present-
ed in **Solid State Design for the Radio Amateur** and in **W1FB's Design
Notebook**. Both books are available from The ARRL, Inc.

Digital Readout for QRP Receivers

It is a dedicated undertaking to build one's own frequency counter
for use for reading the frequency of a receiver. Complete construct-
ion data for such an instrument is presented on page 25-13 of **The
ARRL Handbook**, 1990 edition. A PC board for this project is available
from FAR Circuits for $8.95 plus $1.50 shipping.

I have used a nice alternative to the more expensive counters. It is
a small circuit that is well suited to QRP gear. This circuit was
designed by N0HPK and N1FB. It appeared in **QST** for February of 1989.
It features a 4-digit display and is suitable for use from 1 to 40
MHz with 100 Hz of resolution. A complete kit of parts for this low-
cost counter is available from A & A Engineering, 2521 W. La Palma,
Unit K, Anaheim, CA 92801. The price is $39.95 plus $2.50 for shipp-
ing. Order no. 165 kit. A & A also offers a low cost LED bar-graph
kit that is suitable for S-meter or relative RF power indication.
This visual voltmeter has a 20 LED readout. It sells for $19.95 in
kit form.

In order to use a frequency counter as a digital frequency display
in receivers it is necessary to include a mixer circuit and a post-
mixer low-power RF amplifier. Direct readout is made possible by mix-
ing the receiver local oscillator output with the BFO frequency in
a single-band, single-conversion receiver. This combination yields
the received signal frequency. When working with multiband receivers
you may elect to use the local-oscillator frequency and beat it with
the frequencies of crystal-controlled oscillators to indicate on the
display the receive frequency for several bands. The mixer module
needs to be well shielded and isolated in order to prevent the circuit
from producing an interfering signal at the receive frequency.

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The Subject of Preamps

Does your receiver need a preamp? Some amateurs believe that adding an outboard RF preamplifier will enable them to hear weak signals that might not otherwise be copied. Is this true? The answer can be "yes" or "no," depending upon the receiver with which the preamp is used. Certainly, a preamplifier contributes to the overall gain of a receiver. Typical gains are 10 to 25 dB. This is dependent upon the transistor or MMIC (monolithic microwave integrated circuit) used. Performance is based also on the type of preamp circuit you adopt. For example, a common (grounded) gate JFET can yield 10-12 dB of gain, whereas the same FET could provide 25 dB of gain in a common-source circuit.

The fact of the matter is that a preamp can actually ruin receiver performance. If the receiver already has a gain stage ahead of the mixer, a preamp can degrade the receiver dynamic range and cause spurious products to appear when strong signals are present. Also, a noisy preamp can degrade the SNR (signal-to-noise ratio) of a receiver. It is evident, therefore, that great care must be taken when we consider adding a preamplifier. Elevated S-meter readings occur when preamps are used, but that weak signal may be no more readable than it was before the preamp was turned on.

When to Use a Preamp

The primary purpose of a preamp is to override the mixer noise. Most mixers have a poor NF. For example, if you are using a CA3028A IC as a mixer it may provide up to 15 dB of conversion gain. But, the noise figure may be 8 dB. This can be complicated by losses in the tuned circuit or filter that is used between the mixer and the antenna. In this situation we gain an advantage by adding a low-noise preamp. Not only have we enhanced the SNR of the receiver, but we have the required gain to overcome the input-circuit losses. By way of example, I built a superhet that used a balanced active mixer with no RF amplifier ahead of it. Two-pole bandpass filters were used between the antenna and mixer. On 40 and 80 meters the weakest discernible signal from my generator was 2 uV -- dreadful. I added a broadband MMIC type of preamp (Motorola MWA110) ahead of the filters. It provided 15 dB of gain with a 4-dB NF. Upon adding the preamp I could easily discern a 0.1-uV signal.

It is important to recognize that very low noise figures are not required below, say, 14 MHz. Normally, atmospheric and man-made noise that is picked up by the antenna tends to negate the benefits of a low-noise receiver. An effective NF of 6 dB is well below the noise from your antenna. As the operating frequency is increased above 10 MHz we need to be more dedicated to ensuring a low NF. This is a particularly critical matter at VHF and UHF.

It is important to provide your preamp with the proper load impedance. Typically, the input circuit is designed for 50 ohms. Likewise for the output circuit unless the load is a mixer that presents a different load impedance. Mismatching at the preamp reduces the effective

Fig 3-18 contains examples of RF preamplifiers that are suitable for use in receiver front ends. Example A shows a common-gate JFET with tuned input and output circuits. This amplifier is known for its stability which is dependent in part upon keeping the gate lead very short. The relatively low gain of this amplifier also aids the stability. R1 at A and B functions as a VHF parasitic suppressor. A miniature 850-mu ferrite bead may be used in place of R1.

The circuit at B in Fig 3-18 can produce 20 dB or greater gain. It is, therefore, more prone to instability than is the circuit at A. Care must be taken to keep the input and output tuned circuits shielded from one another to minimize mutual coupling, which encourages self-oscillation. The gate may be tapped down on L2 to discourage self-oscillation. In a like manner you can tap the drain down on L3 to aid stability. Normally, we would not tap the JFET elements below the coil mid points. The coil taps may be determined by experiment if instability becomes a problem.

The broadband amplifier at C and D in Fig 3-18 is easy to build and is stable when it is terminated at each end with a 50-ohm load. A ferrite bead can be added at the amplifier input (D) to discourage VHF self-oscillation if the load is not 50 ohms. The innards of U1 consist of a bipolar transistor and bias resistors. The MWA110 draws 10 mA when used as shown. Monolithic ceramic capacitors are used to minimize stray inductance. These chips are soldered directly to the PC-board conductors as shown. Double-sided PC board material is required to ensure that the input and output PC foils form a 50-ohm strip line. Line width is approximately 1/8 inch. Glass epoxy board material is used. U1 is installed on the ground-plane side of the board. It is pushed snugly against the ground plane to cause the case to be grounded. A small dab of solder may be added to ensure positive contact between the case and the ground plane. Three 1/8-inch holes are bored through the board. A pass-through wire is soldered to the board on each side of each hole to make the ground conductors on each side of the PC board common to one another.

Be aware that the Fig 3-18C amplifier has no selectivity. It can produce a flat 15 dB of gain from 100 kHz to 400 MHz if built as shown in drawing D. This means that antenna energy across that frequency range is amplified by U1. A 50-ohm bandpass filter may be used at the amplifier input to provide selectivity. Good performance may be had with no input filtering if a tuned circuit or bandpass filter is used between this preamp and the mixer input. The 4-dB NF of the MWA110 makes it suitable for preamp use at HF. Other units in the MWA series have higher noise figures. I do not recommend them as receiver preamps. It will be necessary to use band switched input filters at U1 if you desire filtering ahead of this preamp for multi-band use. Bandpass filter tables are provided in the **ARRL Electronics Data Book**. The tables were developed by W7ZOI.

I made the PC board illustrated in Fig 3-18D by removing the necessary copper with a hobby motor and a small cone-shaped carborundum bit. A neater job will result if you etch your board.

Introducing IF Selectivity

Standard procedure calls for including a selective filter in the
IF system of a superheterodyne receiver. Commercially made filters
are entirely predictable in terms of performance, but they are costly.
It seems a bit absurd to install a $100 IF filter in a home-made
superhet that cost $15 to build. Low-cost alternatives are worth
considering. We can make our own filters by using surplus computer
crystals, which often sell for as little as $1 each. The appendix
section in **W1FB's Design Notebook** contains two excellent articles
that describe how to construct crystal ladder filters. They were
written by W7ZOI. I have made 250-Hz bandwidth ladder filters from
Radio Shack TV color-burst crystals. My four-pole CW filter cost
$12! The design was based on the above-referenced articles. Fig 3-19
contains examples of IF filters you can build without spending an
inordinate amount of money.
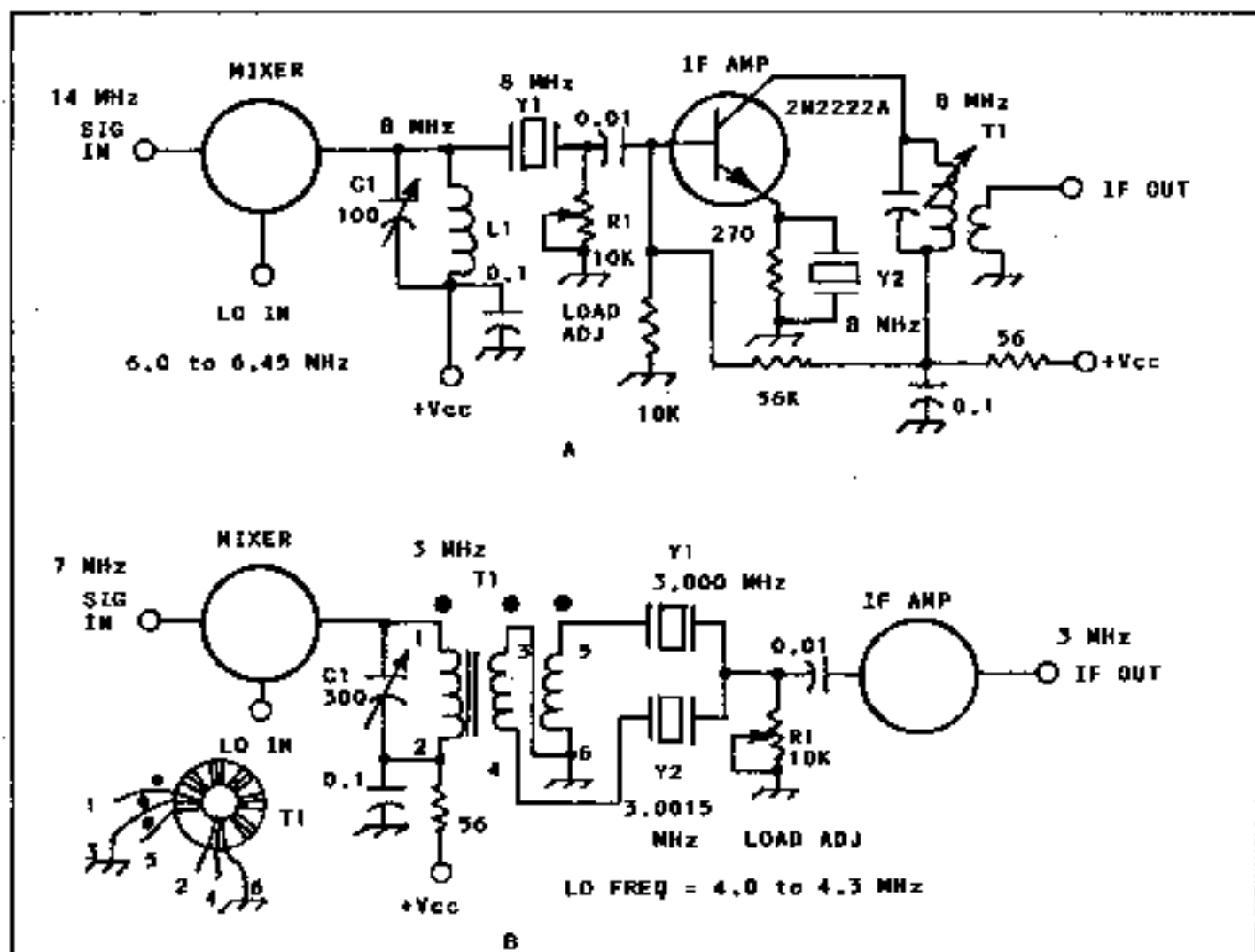
Fig 3-19 -- Examples of two simple crystal filters. Circuit A has two crystals
of the same frequency, whereas circuit B has crystals that are separated
by 1.5 kHz in a half-lattice arrangement. Two frequency examples are given.

The circuit at A in Fig 3-19 is perhaps the easiest to build because it uses two crystals that are on the same frequency. This makes it practical to employ low-cost computer crystals. Y1 allows IF energy to pass at only the crystal frequency. The available selectivity is dependent upon the Q of the particular crystal you use. The Q factor is determined by the internal series resistance of the crystal. The ladder-filter articles in W1FB's Design Notebook show how to measure the series resistance and determine the Q of a crystal.

R1 in Fig 3-19A may be adjusted while listening to a strong signal. The control is adjusted for minimum ringing, which is an annoying "boinging" sound on signal peaks. Noise bursts will also cause the circuit to ring. Once you find the best setting for R1 you may replace the control with a fixed-value resistor of the appropriate value.

Y2 in Fig 3-19A prevents the IF amplifier from producing gain at any frequency other than that of Y2. At 8.0 MHz this crystal acts as a bypass capacitor for the emitter resistor, and this allows the amplifier to provide gain at 8 MHz. At other frequencies the IF amplifier has degenerative feedback (unbypassed emitter resistor) and the gain falls off markedly. L1 and T1 for this circuit can be 10.7-MHz IF transformers. Add sufficient external capacitance to resonate them at 8 MHz.

Fig 3-19B shows a half-lattice crystal filter. This circuit requires crystals that are offset by 1.3 to 1.5 kHz. This provides a fairly symmetrical bandpass response if the electrical and physical symmetry of the circuit around the filter is good. T1 is a tuned trifilar transformer that has an inductance of 15 uH. The inset drawing shows how the T1 windings (three) relate to the circuit. The black dots indicate the polarity of phasing of the windings -- an important factor. T1 is wound on an Amidon Assoc. or equivalent T68-2 powdered-iron toroid.

R1 in Fig 3-19B is adjusted for minimum ripple across the nose of the filter response. Ripple can be explained easily by imagining a dip in signal strength as you tune across a signal. There will be a peak, then a dip which is followed by another peak. Multipole filters, unless terminated properly, may have many dips over the nose of the response curve. They may be a great as 6 dB with a poorly terminated filter. Tune across a steady, unmodulated signal while observing your S meter. There should be but one dip for a half-lattice filter. Adjust R1 and repeat the test. Set R1 for the least amount of dip between the two peak responses, then replace R1 with a fixed-value resistor of equivalent value.

The frequencies listed in Fig 3-19 are simply random choices on my part to illustrate the frequency relationships for two popular surplus computer crystals. These are listed in the Digi-Key no. 906 catalog which you may request by phoning 1-800-344-4539.

Crystals from 450 kHz to 10 MHz may be used in the Fig 3-19 circuits. The lower the crystal frequency the greater the circuit Q, and hence the sharper the filter response. Bandwidths for these circuits are

sufficiently wide to allow SSB and CW reception. Moving the Y1-Y2 crystal frequencies closer together in circuit B of Fig 3-19 will narrow the response and make the filter more desirable for CW reception. I suggest a Y1-Y2 separation of 700 Hz to 1 kHz for CW use.

LC filters are practical at low IFs, such as 455 kHz. It is possible to cascade transistor IF transformers to create an LC filter that is suitable for SSB reception. Fig 3-20 illustrates how this is done.
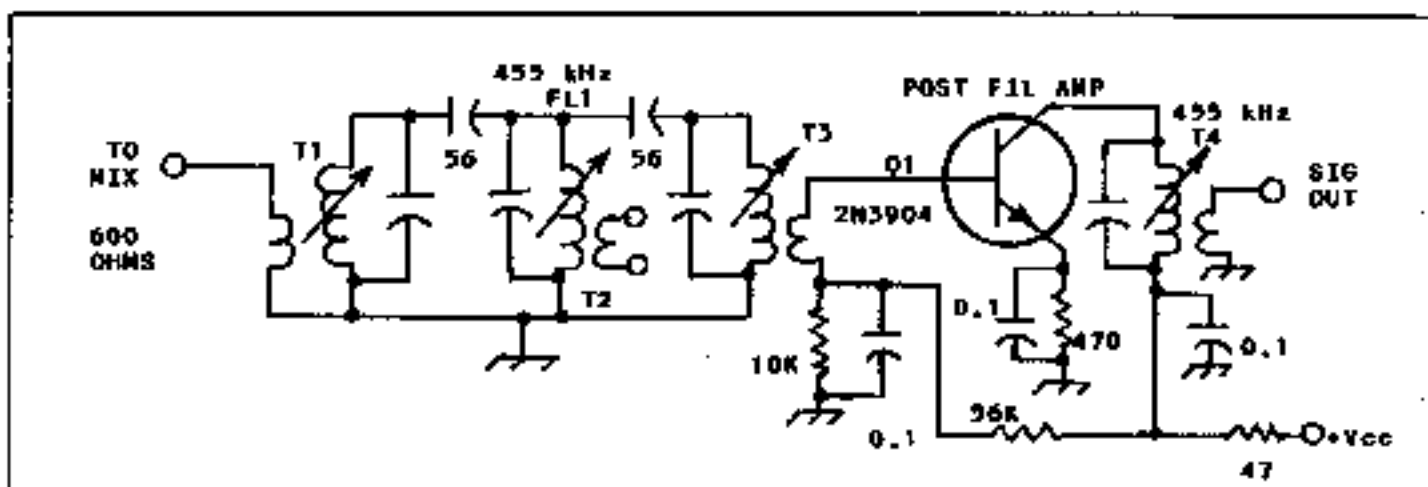

Fig 3-20 -- Example of a filter that uses three transistor-radio IF transformers to provide SSB bandwidth. As many as four transformers may be used for FL1 before insertion loss becomes prohibitive. The filter input and output impedance is 600 ohms. Q1 compensates for the loss introduced by FL1.

The shield cans for the IF transformers are grounded to provide isolation between the three filter poles. If you observe objectional ripple with FL1, reduce the value of the two top-coupling capacitors. Select values that eliminate the dips in filter nose response. T1, T2 and T3 will need readjustment if you change the top-coupling capacitors. Peak all four transformers at 455 kHz.

The input of FL1 in Fig 3-20 should be terminated in 600 ohms. For example, if your mixer is a dual-gate MOSFET, simply use a 620-ohm resistor for the mixer drain load. No tuned mixer output circuit is needed owing to the selectivity provided by the filter. You may use the IF transformers from a discarded transistor radio. They are designed for a 10,000-ohm primary and a 600-ohm secondary impedance. The exception is the last IF transformer, which looks into the usual diode AM detector. It has a different transformation ratio, usually. Use that transformer for T4.

Most IF transformers in transistor AM radios have the same pinout and characteristic impedance. Therefore, if you buy new units they should work as well as those in a transistor radio. These IF transformers have built-in capacitors. You need not add external capacitance to resonate the transformers. A 455-kHz BFO can be made by using a 455-kHz IF transformer with a JFET or bipolar transistor. A practical circuit for this type of circuit is presented in the diagram for the simple superhet that appears later in the chapter.

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Audio-Frequency Selectivity

Audio filters provide a means to obtain CW and SSB selectivity when
the use of IF filters is not possible or practical. An audio filter
may consist of inductors and capacitors (passive filter) or it may
be structured from capacitors, resistors and transistors or ICs. The
latter type is called an active filter. The term "active" means that
an operating voltage is necessary. By comparison, most passive filters
are physically larger than active ones, and they introduce circuit
loss (insertion loss). Active filters, on the other hand, are capable
of providing gain if they are designed for greater than unity gain.
An example of a simple tunable RC active bandpass filter is given
in Fig 3-21.
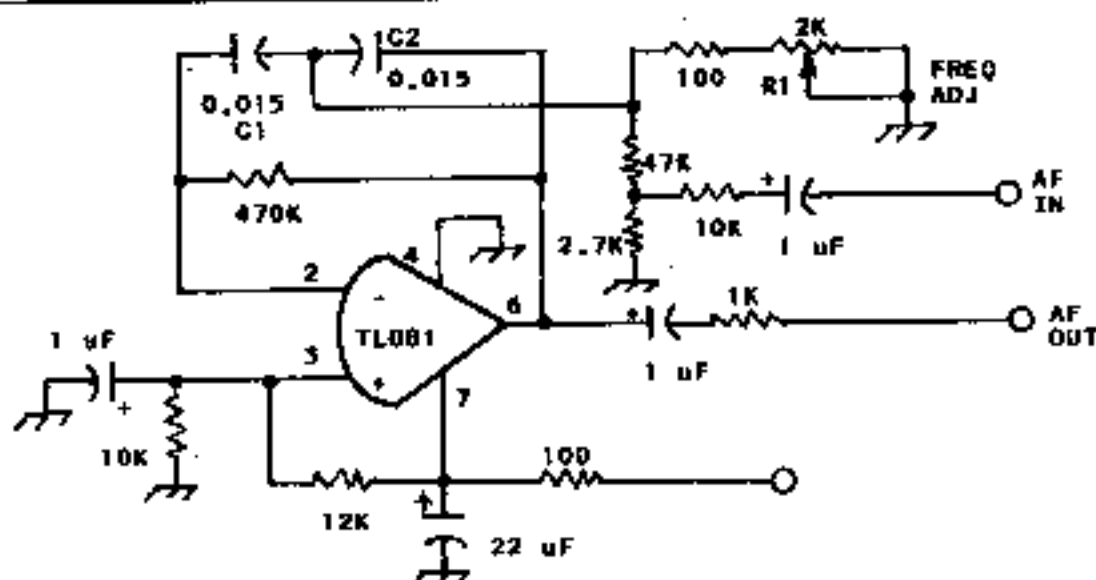
Fig 3-21 -- Practical example of a frequency-variable RC active CW filter.
It has a bandpass response which rejects frequencies above and below the freq-
uency of interest. R1 is adjusted for the desired CW beat-note frequency.
C1 and C2 are polystyrene capacitors that need to be closely matched (1%)
in value. Although a 741 op amp may be used in this circuit without changes,
the TLO series biFET op amps are less noisy and are a better choice for low-
level applications in the receiver audio system.

Bandpass audio filters are the most popular for CW work because they
reject QRM above and below the frequency of choice. Low-pass active
filters are preferable for SSB reception because they eliminate high-
pitched "chatter" while allowing the desired low-frequency audio to
pass to the phones or speaker. Although high-pass RC active filters
are practical, they find little application for signal reception.
A bandpass filter is a combination of high- and low-pass filter RC
active filters. Explicit design data for RC active filters is provided
in The ARRL Handbook and in Solid State Design for the Radio Amateur.

Advantages of Audio Filtering

Audio filters have the ability to lift weak signals above the atmos-
pheric and man-made noise that arrives via the antenna. An unreadable
CW signal can become a Q5 signal when a selective audio filter is

used. This is especially important to those of us who operate QRP, since many of the signals we deal with are rather weak. Also, a DC receiver requires overall selectivity that does not exist without some type of audio filter. Bandwidths as narrow as 200 Hz may be used for dedicated weak-signal reception, but the narrower responses tend to cause operator fatigue after an extended period of operating. This may be relieved somewhat by changing the CW note pitch from time to time. The Fig 3-21 circuit makes this possible. Fixed-tuned RC active filters do not make this psychological "fix" possible.

When an audio filter is used late in the receiver circuit (near the audio output circuitry) it greatly reduces wide-band receiver noise. This form of noise often originates in the IF-amplifier section of a superhet. It may become manifest also in the audio preamplifier stages. The effect of the filter is an improvement in the signal-to-noise ratio of the receiver.

Operator fatigue can result also from excessive filter ringing (explained earlier in this chapter). The narrower the filter bandwidth and the greater the number of filter poles, the greater the chance for ringing. This irritating pinging sound is especially difficult to endure over long operating periods when we wear headphones.

Greater selectivity (sharper filter response skirts) may be obtained from the Fig 3-21 circuit if you cascade two of the filters. This requires the use of a dual 2000-ohm control at R1. Care must be taken to ensure that all of the polystyrene capacitors are closely matched in value. If not, the filter response will be broadened because the filter response peaks will be offset from one another. It is important also to use high-Q capacitors in audio filters. Polystyrene and high-quality mylar capacitors are good choices. I have found it easier to select closely matched capacitors from a group of polystyrene units than when sorting through other types of capacitors.

A Switched Variable-Selectivity AF Filter

Fig 3-22 contains the circuit for a practical RC active CW filter that has four bandwidths from which to choose. This is an adaptation of a filter that was produced by Ten-Tec Corp. some years ago. A PC board for this filter is available from FAR Circuits.

The filter, as shown, has four selectivity positions. The 3-dB bandwidth for each position is listed in Fig 3-22. A second switch may be added to permit bypassing the filter during SSB reception. The audio energy is then routed around the filter. A second switch section would be used to remove the operating voltage from the circuit. If this is done I recommend that you use shielded audio cable or RG-174 coaxial line for the audio cables that go to and from the switch.

Although 747 op amps are specified in Fig 3-22, the filter noise (called "popcorn") will be substantially less if you use TL082s for U1 and U2. It is especially important to use matched capacitors and resistors in this filter, because the four filter-pole frequencies must be the same.

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C1 2200 R3 470K C5 2200 R8 470K C8 2200 R12 470K

220K C2 U1A 4 R5 220K C6 U1B 7 220K C9 U2A 1 4

AF IN R1 1 — 12 — 10 R10 2 — 12

22K R2 2 + 9 0.1 C4 R7 22K 5 + 13 0.1 C7 22K R11 220K R13

2200 747 0.1 2200 R6 22K C10 0.1

22K R4 22 uF 220K R9 220K 220K

C3 R18 22K

+12 V

R14 220K

2 S1

1 3

SELECTIVITY

AF OUT 4 U2B 13 R15 22K

0.1 C14 0.1 C13 10 + 6 2200

470K R17 — 7 R16 470K C11

-5 dB BANDWIDTH

1 -- 350 Hz
2 -- 160 Hz
3 -- 130 Hz
4 -- 110 Hz

Center Freq = 750 Hz
Filter Gain (AV) = 2.1
$Q_L = 3$

2200 C12

Fig 3-22 -- Practical circuit for a switched four-pole RC active CW filter that may be used with DC or superhet receivers. The 2200-pF capacitors in the frequency-determining circuits are matched polystyrene units. The 22K- and 470K-ohm resistors should also be closely matched in value. A single-pole, four position rotary switch is used for S1. The filter center frequency may be changed by selecting component values that can be determined from the equations provided in **The ARRL Handbook.**

Perhaps you are wondering how filters such as those shown in Fig 3-22 compare in performance to crystal or mechanical IF filters that have narrow bandwidths. It is difficult for a person with an untrained ear to discern the difference between the two types of filtering. An IF filter has the advantage of filtering out unwanted adjacent-frequency energy before it reaches the IF amplifiers. This provides a cleaner signal at the receiver output, but it does not "launder" the overall energy to reduce wide-band noise that is generated after an IF filter. Some modern receivers employ a technique that is known as "tail-end" filtering of the IF. A second IF filter is used in the IF system. It follows the last IF amplifier and the center frequency is the same as the first IF filter. However, the effective bandwidth of the second filter is slightly greater than that of the first filter. Tail-end filters also reduce wide-band noise. Owing to the high cost

of factory made crystal filters, it is more practical to use a
1st IF filter with SSB bandwidth and employ a CW-bandwidth RC active
filter after the 1st audio amplifier. This arrangement is, of course,
not possible for a DC receiver. Audio filtering is the only option
we have.

Best RC active filter performance results when we install the filter
near the start of the audio-amplifier section. Strong-signal over-
loading of the filter can result when we insert it at the receiver
output or just ahead of the audio power amplifier. Fig 3-23 contains
a scale etching template for the Fig 3-22 filter, along with a parts
placement guide.
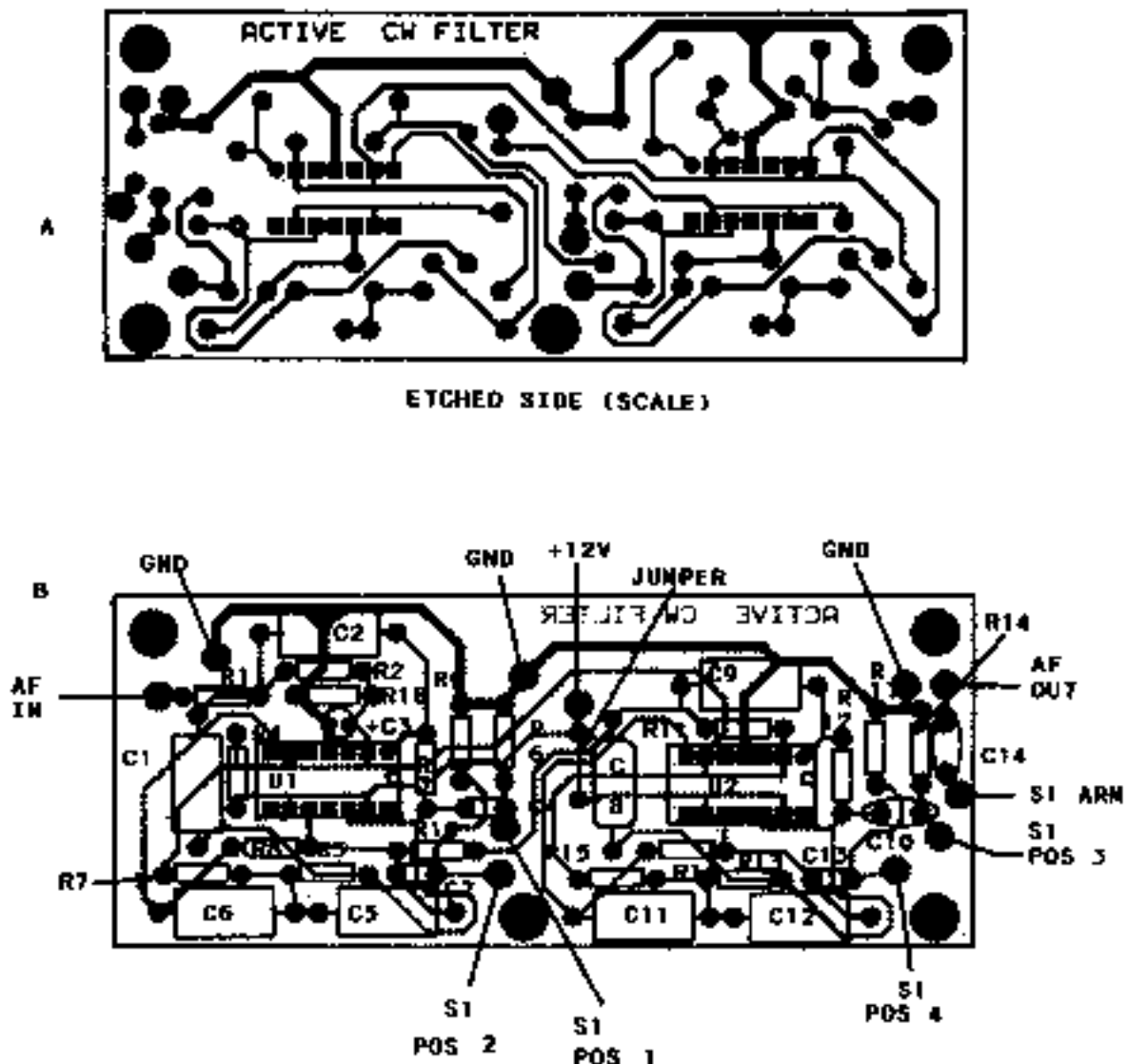
Fig 3-22 -- A scale etching pattern is shown at A. An Xray view of the board
as seen from the component side is illustrated at B. PC boards are available
from FAR Circuits for $5 plus $1.50 shipping.

Regenerative Receivers

I feel it is worth digging into the past to briefly discuss the matter of regenerative receivers. They still have a place in some areas of QRP operation. If they serve no other purpose, they represent a fine starting point for an experimenter. Regenerative receivers are similar to DC receivers in many ways. The major difference is that the regen detector serves also as the oscillator. The overall circuit is somewhat simpler than that of a DC receiver, which means the parts count is smaller and the unit can be constructed in a rather compact manner.

A regeneration control is necessary to permit us to bring the detector to the point of self-oscillation. This provides the required beat note for CW reception and the necessary carrier for SSB reception. You may also receive AM signals by adjusting the regeneration control so that the detector is on the edge or fringe of oscillation. Regeneration occurs when sufficient feedback energy is introduced to cause oscillation.

As is the situation with DC receivers, single-signal reception is not possible with a regen receiver. In other words, energy that is present either side of the desired frequency can be heard. This has the effect of introducing QRM that would not be present in an SSB type of receiver. Fig 3-23 shows the circuit for a simple regenerative receiver. Nothing has been included for providing audio selectivity. The Fig 3-22 circuit, or two poles of that filter, may be added to enhance CW reception.
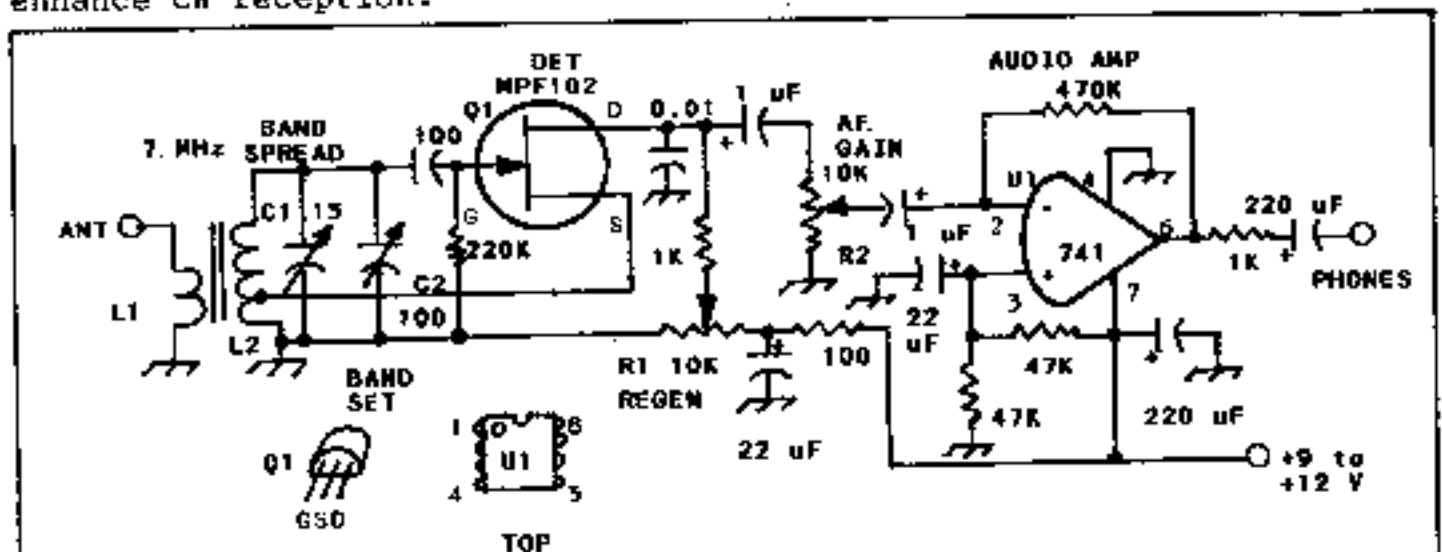
Fig 3-22 -- Schematic diagram of a simple regenerative 40-meter receiver. Polarized capacitors are 16-V or greater tantalum or electrolytic. C1 and C2 are miniature air variable capacitors for panel mounting. Resistors are 1/4-W carbon film or carbon composition, except for R1 and R2 which are carbon composition controls (panel mounted). L2 is a 6-uH toroidal inductor. Use 35 turns of no. 28 enam. wire on an Amidon T50-2 toroid. Tap L2 at 8 turns above the grounded end. L1 has 3 turns of no. 28 enam. wire. Other VHF JFETs may be used at Q1, such as the 2N4416.

The little receiver in Fig 3-22 covers a wide frequency range. The band-set capacitor (C2), which you may think of as a coarse tuning control, provides coverage from 6.5 to 17 MHz. This enables you to receive the 40, 30 and 20 meter bands by setting C2 to the proper part of the spectrum, then spreading the amateur band by means of C1. You may think of C1 as the fine tuning control.

Adjust R1 (clockwise) until the CW signals have a beat note. They will sound mushy until the detector begins to oscillate. Use no more regeneration than is needed to produce a clear signal. You will find it necessary to readjust R1 when you tune the receiver to a new frequency -- especially when the new frequency is well removed from the original one.

Excessive loading by the antenna can prevent Q1 from oscillating. Use the least number of L1 turns that provide good reception. The greater the number of L1 turns the tighter the antenna coupling and hence the greater the loading effect.

Regenerative receivers and DC receivers radiate a signal at the frequency to which they are tuned. Addition of an RF amplifier stage will greatly reduce the radiation by virtue of isolation. Your nearby DX operator may have problems with "that carrier on the frequency" if you monitor his operations! It is good to be aware that you can cause QRM with these receivers if you live near another amateur.

You can use the Fig 3-22 receiver on other bands by changing the value of the L2 inductance. For example, double the inductance for 80 meters or halve it for reception on 18, 21, 24 and 28 MHz. Maintain the same turns ratio for L1 and L2 if you change the inductance.

Some Practical Aspects of DC Receivers

The beginning QRPer who is not endowed with a strong technical background is wise to build a few DC receivers before launching into a superhet project. Generally speaking, DC receivers work well and are easy to make operate. A DC receiver that percolates properly exhibits very clean audio output, especially when it is operated from a battery. I have heard many DC receiver owners comment about the crystal clarity of the signals from their receivers. This is due in part to the gain distribution throughout the system. Numerous stages are not cascaded to cause overdriving of any one stage. This minimizes distortion that might otherwise be present.

A major problem we encounter when using simple DC receivers is unwanted AM detection. This shows up as a blanketing of the receiver by strong in-band or out-of-band commercial short-wave stations. This malady is most likely to occur in a DC receiver that uses a single-ended detector, such as an MPF102 or 40673. A singly balanced detector of the CA3028A IC type, or push-pull MPF102s or similar, helps to resolve the problem, but a doubly balanced detector will usually work the best with regard to AM rejection. The NE602, MC1496 and similar mixer ICs are good choices. Doubly balanced diode detectors are good

performers as well, but they require substantially more LO injection than do the IC types of detector. Approximately +7 dBm of LO power is needed for diode mixers and product detectors. This equates to roughly 6 mW of LO power, since 0 dBm is 1 mW at 50 ohms.

If we use a high order of tuned-circuit selectivity ahead of the detector we can mimimize the AM detection problem for out-of-band broadcast signals. A double tuned input circuit is helpful in this case. The AM blanketing problem is the most severe on 30 and 40 meters. I have seldom experienced these problems when listening to 80 or 20 meters with a DC receiver.

A practical circuit for a simple DC receiver is shown in Fig 3-23. It used a dual-gate MOSFET as the product detector. Devices such as the RCA 40673 or Texas Instruments 3N211 or 3N212 are suitable at Q1. Best performance occurs when the peak-to-peak LO injection at gate no. 2 is 5 to 6 volts. Values greater than that can damage the
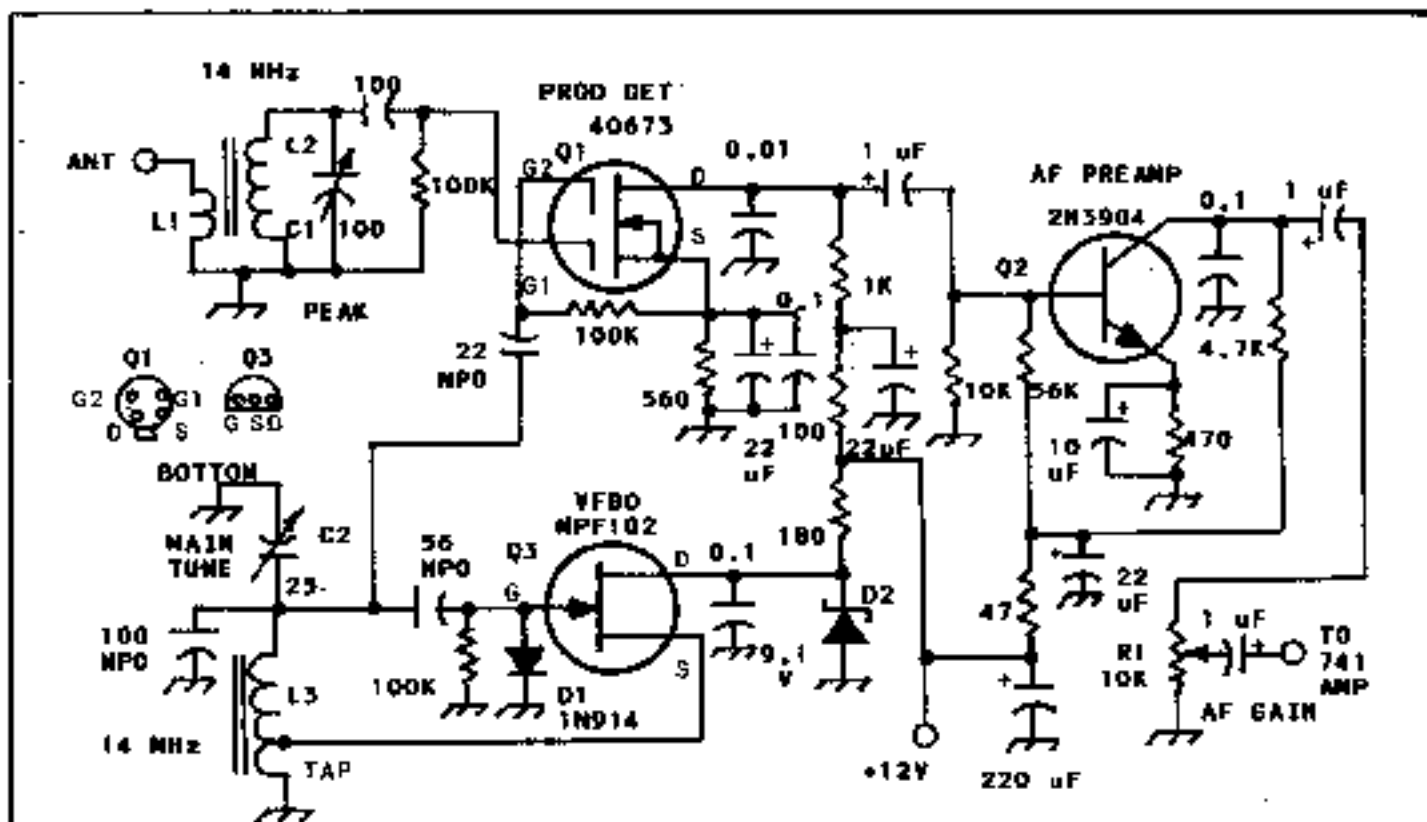
Fig 3-23 -- Practical circuit for a 20-meter DC receiver that uses a single-ended product detector. Fixed-value capacitors are disc ceramic except those with polarity marked, which are 16-V electrolytic or tantalum. Fixed-value resistors are 1/4-W carbon film or composition. C1 is a 100 pF ceramic or mica trimmer. C2 is a miniature air variable (use a vernier drive). L2 is a 2.5-uH toroidal inductor. Use 25 turns of no. 26 enam. wire on an Amidon T50-6 toroid. L1 has 2 turns of no. 26. L3 is a 1.2-uH inductor. Wind 17 turns of no. 24 enam. wire on a T50-6 toroid. R1 is an audio taper carbon composition control. D2 is a 9.1-V, 400-mW Zener diode. NPO means zero temperature coefficient.

gate insulation and cause the MOSFET to develop a short circuit.

The Fig 3-23 can can be used on any frequency from 1.8 to 14 MHz by merely changing the inductance values of L2 and L3. VFBO stability may become a challenge at 18 MHz and higher, although the receiver is entirely capable of good performance above 14 MHz. When you change the L2 and L3 inductance be sure to maintain the same L1-L2 turns ratio. The tape on L3 should always be located at a point that is 1/4 the total number of turns. This will ensure adequate feedback for the VFBO (variable frequency beat oscillator).

Some of the circuit features are pertinent to all DC receivers. For example, regulated dc is supplied to Q3. This prevents frequency changes when there is a variation in supply voltage, such as occurs during mobile operation. A low noise audio preamplifier should follow the product detector. The output of the product detector (Q1 drain) is bypassed to remove VFBO energy that would otherwise reach the first audio amplifier. The bypass capacitor aids also in attenuating high-frequency hiss noise and high-pitched heterodynes. The 0.1-uF bypass capacitor at the collector of Q2 (Fig 3-23) also bypasses high-frequency -audio. The value is chosen to suit the listening needs of the user. Larger or smaller values may be used.

Note that the source of Q1 is bypassed for RF and audio. Since this stage involves both types of energy it is important to ensure maximum conversion gain at audio. The 22-uF capacitor is included for audio purposes.

C1 in Fig 3-23 may be replaced with a small panel-mounted air variable for operation on 160 or 80 meters. At frequencies higher than 4 MHz the trimmer can be set for mid band without a need for further adjustment. The tuned-circuit Q at 4 MHz and lower is such that peaking is necessary when making frequency changes greater than 15-25 kHz. As C1 is adjusted there will be some frequency shift (pulling) of Q3. This is because both tuned circuits are on approximately the same frequency and the VFBO tank is coupled directly to Q1 and Q3. This could be resolved by adding a buffer after Q3 and taking the LO injection from it (see section on VFOs).

You can recognize quickly that the Fig 3-23 circuit is anything but complex. It represents a good starting point for your first effort at building a receiver. Improved performance can be had from circuits that are somewhat "busier." Notably, a doubly balanced product detector will minimize or prevent AM detection. A VBFO with two stages of buffering and amplification will provide greater oscillator stability. An RC active audio filter between Q2 and the succeeding audio amplifier will provide CW selectivity and improve the overall receiver SNR. Also, an audio-derived S meter (Fig 3-16) can be connected to the collector of Q2, if desired.

The audio level for this receiver is dependent upon the circuit you add after R1. A 741 or TL0 series op amp (see Fig 3-22) is adequate for headphone use. The op amp can be followed by an LM386 IC to provide sufficient volume for a loud speaker.
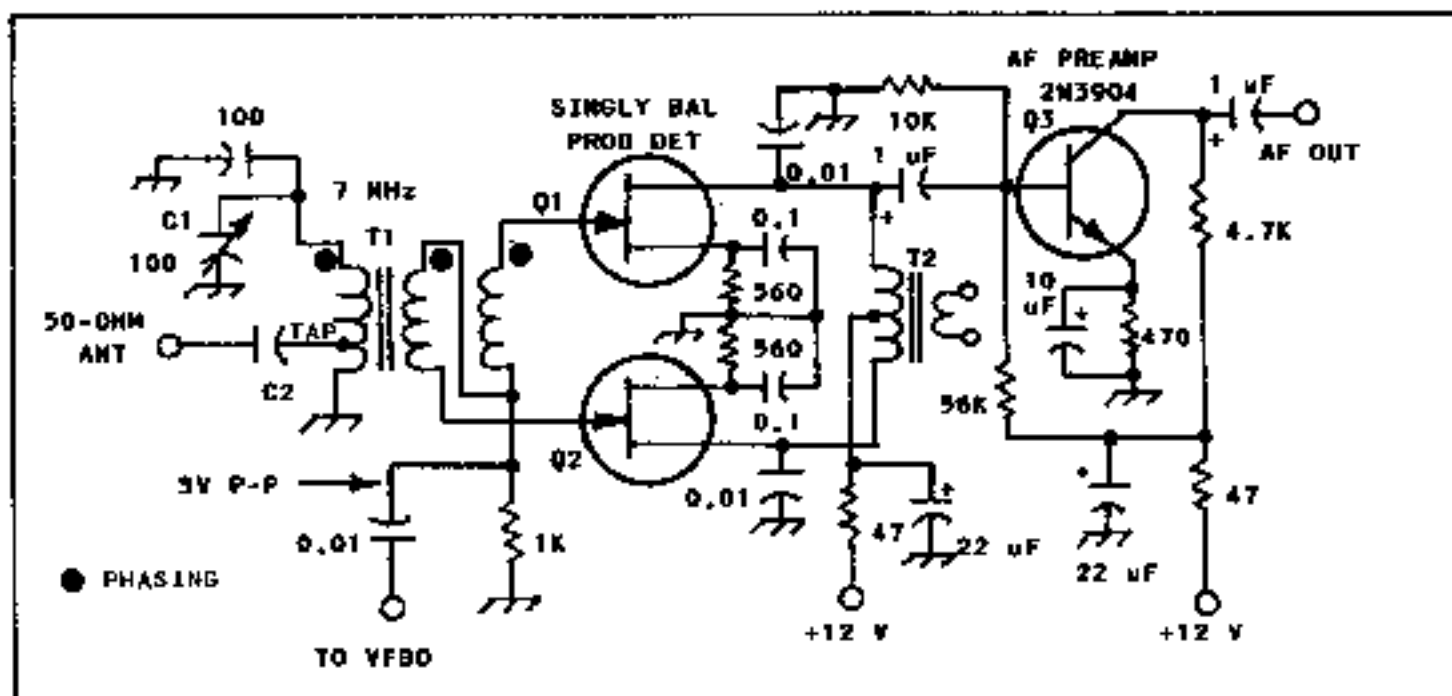
Fig 3-24 -- Practical circuit for a doubly balanced detector that uses two JFETs.
T1 is a trifilar-wound toroidal inductor. For 40 meters use 3 uH (29 turns of
no. 28 enam. wire on an Amidon T50-2 toroid). Tap the primary at 2 turns above
ground. C1 may be a trimmer or panel-mounted air variable. T2 is an audio-output
transformer (1000 ohms, ct, to 8 ohms). A 10K-ohm to 2K-ohm interstage trans-
former may also be used at T2. T2 may be removed from a junk transistor radio.

Fig 3-24 shows an improved product detector. It is effective for reduc-
ing AM detection. T1 may be wound for other amateur frequencies. It
contains a trifilar winding (three wires in parallel, twisted 6-8 times
per inch). Q1 and Q2 may be MPF102s, but better performance will result
if 2N4416s are used. The more closely Q1 and Q2 are matched the better
the performance. An alternative VFBO injection method is to ground
the bottom of the T1 secondary windings, lift the 0.1-uF source capac-
tors above ground (then join them) and apply the VFBO injection to
the paralleled Q1, Q2 sources. This method is preferred by some design-
ers. The conversion gain of this detector is about 10 dB.

Doubly Balanced Detectors

A doubly balanced mixer, modulator or detector does not offer greater
conversion gain than we can obtain from simpler mixers. In fact, a
diode type of balanced product detector exhibits a gain loss (known
as conversion loss). For a doubly balanced diode-ring mixer or detector
the loss is approximately 8 dB. The VFBO injection needs to be +7
dBm for a doubly balanced diode detector. The principal advantage of
a doubly balanced detector is that it offers the best port-to-port
isolation of the common types. Specifically, very little energy that
reaches one port appears at the other two ports. The dynamic range
of a balanced diode mixer or detector is excellent compared to other
kinds of mixers or detectors. A low-noise RF preamplifier is usually
necessary ahead of a diode detector to enhance the receiver NF. This
is seldom necessary when we use doubly balanced active detectors below
14 MHz.

Practical examples of passive and active doubly balanced detectors
are presented in Fig 3-25. Matched diodes are necessary in the circuit
at A if you construct a doubly balanced mixer or detector. Also,
electrical symmetry of the overall detector circuit is essential.
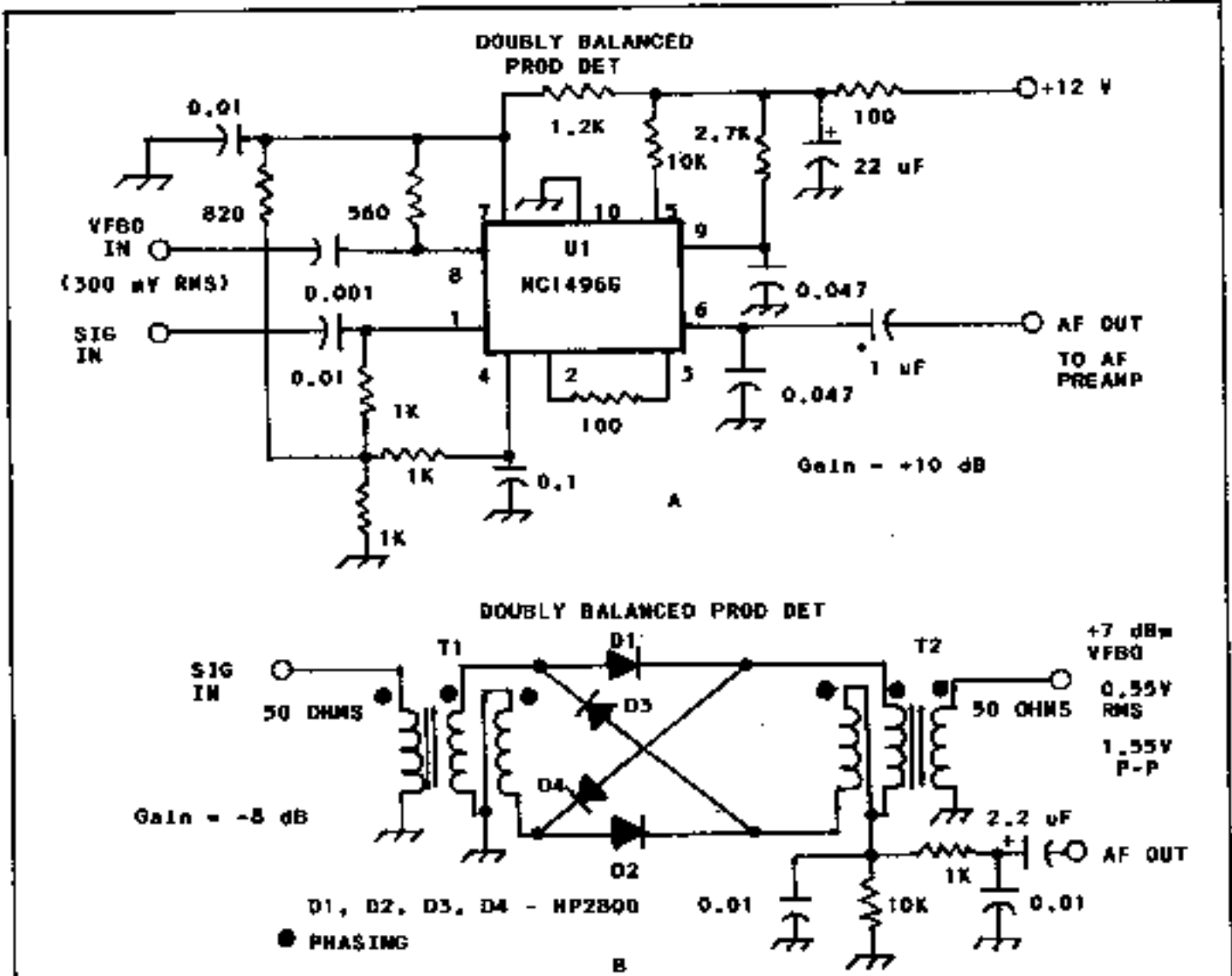
Fig 3-25 -- Practical examples of a doubly balanced active product detector
(A) and a passive doubly balanced ring-diode detector (B). VFBO injection
levels are indicated. Both detectors are quite immune to AM detection. D1
through D4, incl., should be matched diodes. 1N914 diodes may be used in
place of the HP2800 hot-carrier diodes if the forward resistance (8-10 ohms
typically) of the diodes is matched with an ohmmeter. T1 and T2 at B are
trifilar-wound broadband transformers. Use 15 trifilar turns of no. 28 enam.
wire on an Amidon FT-37-43 ferrite toroid (850 mu). Symmetrical layout is
vital to balanced performance for both of these product detectors. Long leads
should be avoided.

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Singly Balanced Product Detector

Transistor-array ICs, such as the low-cost RCA CA3046 or Motorola
MC3346P, are excellent devices for use as doubly balanced detectors.
They may be used as DC receiver front ends or as product detectors
in superhet receivers. A practical circuit is shown in Fig 3-26. The
circuit uses three transistors as the detector (arranged like the
inner workings of the CA3028A). A fourth transistor serves as an audio
preamplifier and the fifth transistor is used as a dc switch for muting
the receiver.
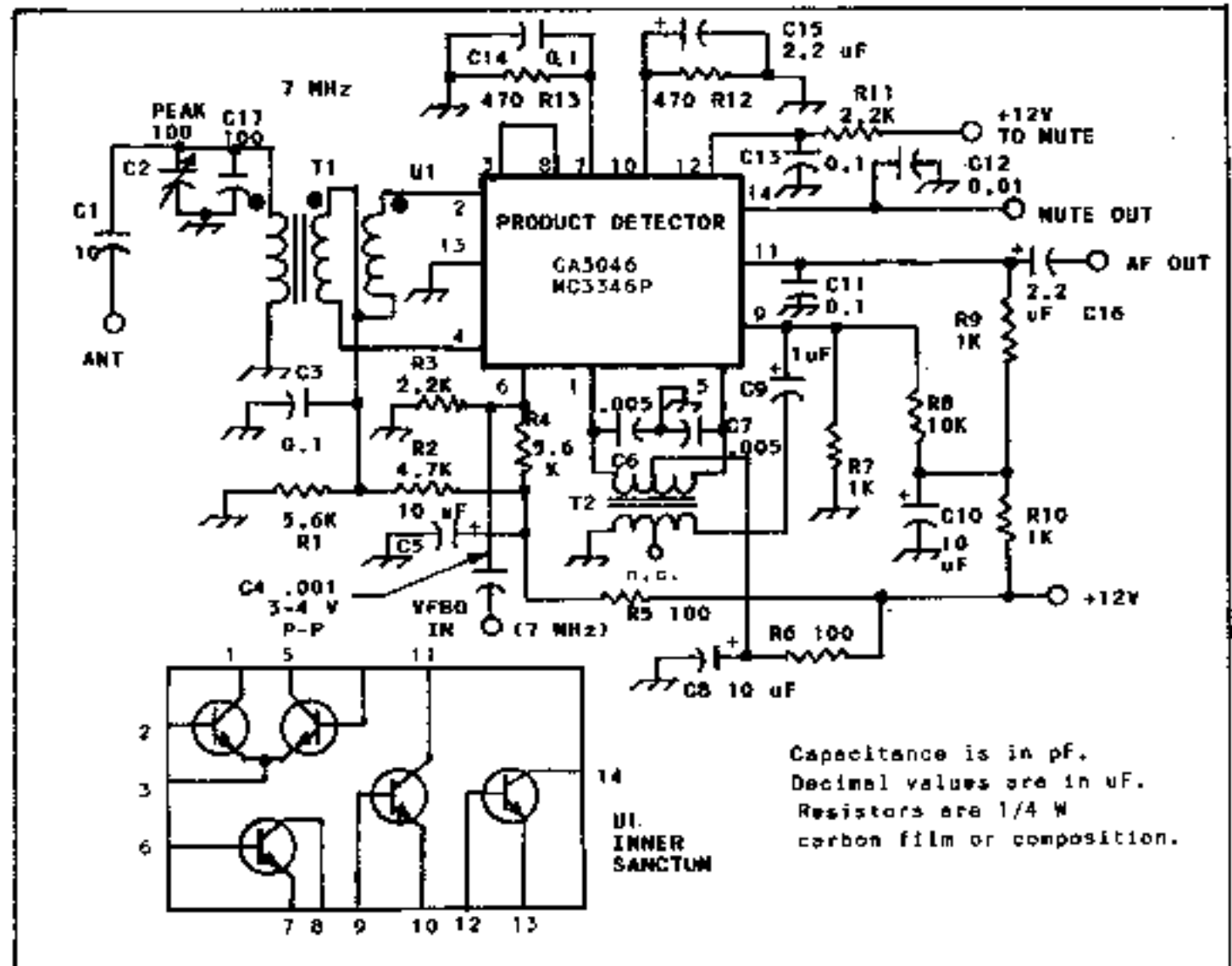
Fig 3-26 -- Schematic diagram of an IC product detector for use
as a DC receiver or superhet product detector. If used in a super-
het, T1 is changed for resonance at the IF. The BFO is fed to
C4. A ceramic or plastic trimmer is used for C2. Fixed-value
capacitors are disc ceramic. Polarized units are tantalum or
electrolytic. T1 has 36 primary turns of no. 30 enam. on a T50-
2 toroid. Bifilar-wound secondary has 8 turns of no. 28 enam.
wire. T2 is a 10K-ohm to 2K-ohm PC mount audio transformer (avail.
from Mouser Electronics).

The Fig 3-26 circuit may be used on other frequencies by simply changing the T1 inductance and the VFBO injection frequency. The T1 turns ratio must be maintained as specified for all frequencies. The inset drawing of Fig 3-26 shows the inner workings and pinout for the transistor-array IC. Various other applications become apparent as we study the inner circuit. This chip is suitable for the basis of an entire regenerative receiver. It would be a fine basis for a QRP transmitter as well.

When working with a CA3046 or MC3346P we must be sure that pin 13 is grounded directly. This is because pin 13 is connected to the IC substrate. Elevating this part of the circuit above ground can destroy the IC.

A PC board or complete parts kit for this project may be obtained from A & A Engineering (W6UCM), 2521 W. La Palma Blvd., Unit K, Anaheim, CA 92801. I obtained excellent performance from my original circuit, and found the A & A version to be very good in both performance and quality.

A complete DC receiver can be built around this circuit. I suggest that a 2N3904 audio preamp be used after U1. It can be followed by an RC active CW audio filter and an LM386 audio power amplifier. One of the VFOs described earlier in this chapter will serve as the VFBO. Unwanted AM detection is minimal with this circuit. With respect to overall sensitivity, I was able to clearly discern a 0.1-uV signal from my URM-25 generator.

Universal DC Receiver

Fig 3-27 contains the circuit for a high-performance DC receiver that has a low parts count. It uses the popular NE602 doubly balanced mixer IC. Q1 is a low-noise audio preamp. It is followed by a single-pole RC active low-pass audio filter. This helps eliminate audio energy above 700 Hz for improved CW reception. A 741 op amp (U2) provides robust output for a pair of headphones. You may add an LM386 or other audio power IC after U2 if you desire loudspeaker operation.

The VFBO is contained on the PC board. Although an MPF102 is indicated for Q2 in Fig 3-27, a 2N4416 is suggested for better overall performance. The device transconductance is more predictable and it has a better pinchoff characteristic.

Jumper wire W1 may be removed to provide receiver muting during transmit periods. An external relay or PNP transistor switch can be used to complete the circuit during receive. Only the audio amplifier, U3, is turned off during transmit. This ensures that the Q2 oscillator remains operational at all times. This eliminates short-term drift when changing from transmit to receive.

A PC board for this project is available from FAR Circuits for $6 plus $1.50 for shipping. Fig 3-28 has a scale etching template for those who wish to make their own PC board.
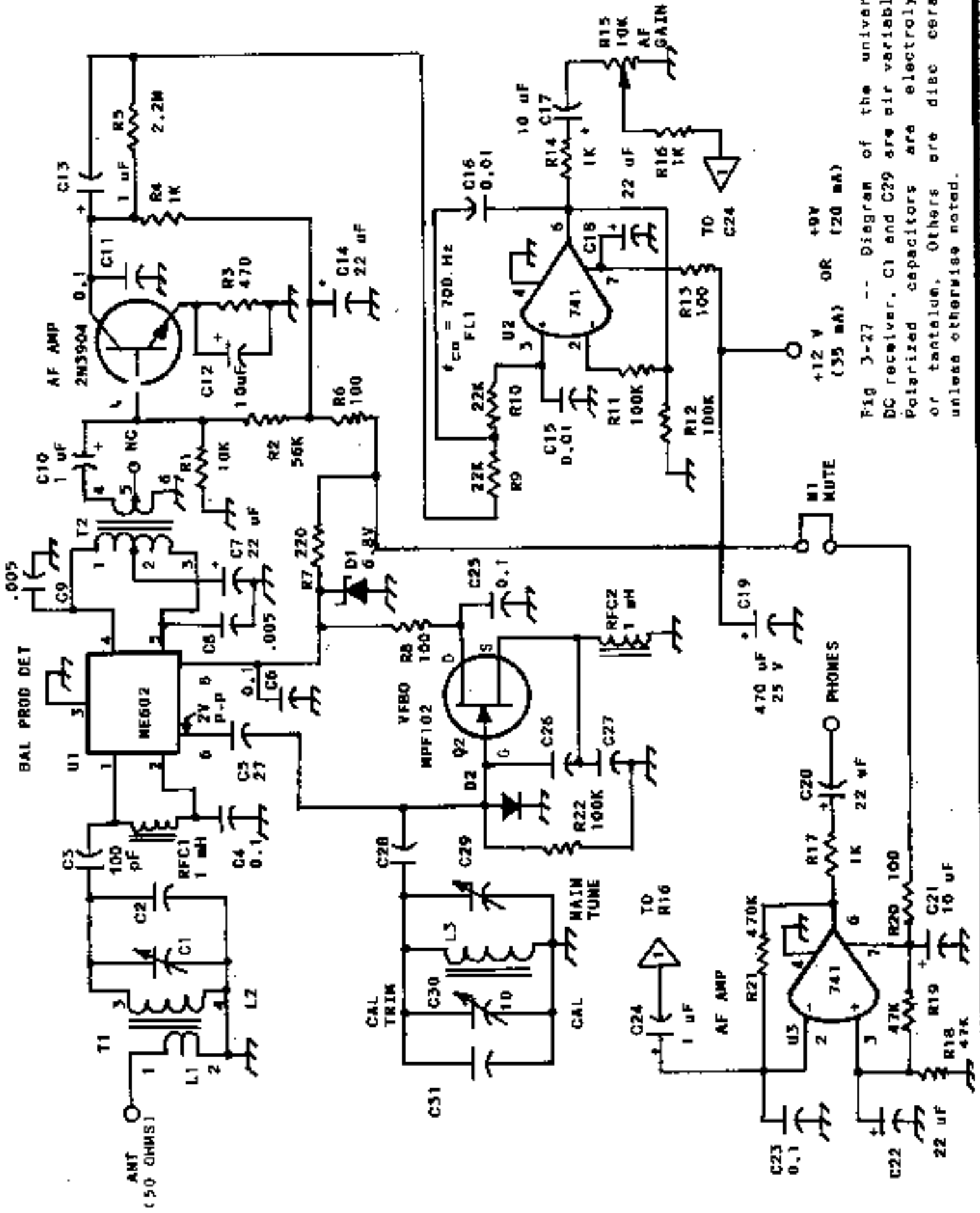
Fig 3-27 -- Diagram of the universal DC receiver. C1 and C29 are air variables. Polarized capacitors are electrolytic or tantalum. Others are disc ceramic unless otherwise noted.

R15 in Fig 3-27 is an audio taper carbon composition control. RFC1 and RFC2 are miniature RF chokes, such as those sold by Mouser Electronics (no. 43LR103). Audio transformer T2 has a 4000-ohm primary and a 600-ohm secondary (Mouser no. 42TL021). The secondary center tap is not used. Other interstage transformers will work in this circuit, such as one with a 10K-ohm primary and a 1K-ohm or 2K-ohm secondary.

The operating voltage for U1 is lowered to 6.8 by virtue of Zener diode D1. A 1N914 diode (D2) is used to stabilize the VFBO bias at Q2. NPO capacitors are used at C5, C28 and C31. Polystyrene capacitors are suitably stable for you to use at C26 and C27. Silver micas may used at C26 and C27, but they may not be as temperature stable as the polystyrene units. Table 3-2 contains the component values for the frequency-determining circuits.

TABLE 3-2

BAND (m)	C26,C27	C28	C1,C29	C2, C31	L1	L2,L3
80	0.001	100	50	100	4 t no. 26	8.uH, 41 t no. 26 enam. on a T68-6 toroid.
40	560	56	25	56	3 t no. 26	3.8 uH, 31 t no. 26 enam. on a T50-6 toroid.
30	560	56	25	56	2 t no. 24	1.8 uH, 21 t no. 24 enam. on a T50-6 toroid.
20	220	27	15	56	2 t no. 24	1.4 uH, 16 t no. 24 enam. on a T50-6 toroid.

These values provide full coverage of each band specified. The decimal value capacitors are in uF. Others are in pF. Toroid cores are Amidon Assoc. or equiv. Micrometels Corp. units. No. 6 has a yellow color code. L3 should be coated with two applications of polystyrene Q Dope or an equivalent high-Q coil cement. C1 can be a trimmer capacitor for 30, 40 and 20 meters. If a trimmer is used, adjust it for peak response at the center of the band.

If you wish to add a brute-force type of RF gain control to the Fig 3-27 receiver you can install a 500-ohm carbon composition control between the antenna and L1. The antenna connects to the arm of the control. L1 is attached to the high end of the control and the bottom end of the potentiometer is grounded. An RF gain control is helpful when very strong signals tend to overload the receiver. This style of control circuit does, of course, create a mismatched input impedance. A 50-ohm step attenuator is a better device to use for this purpose.
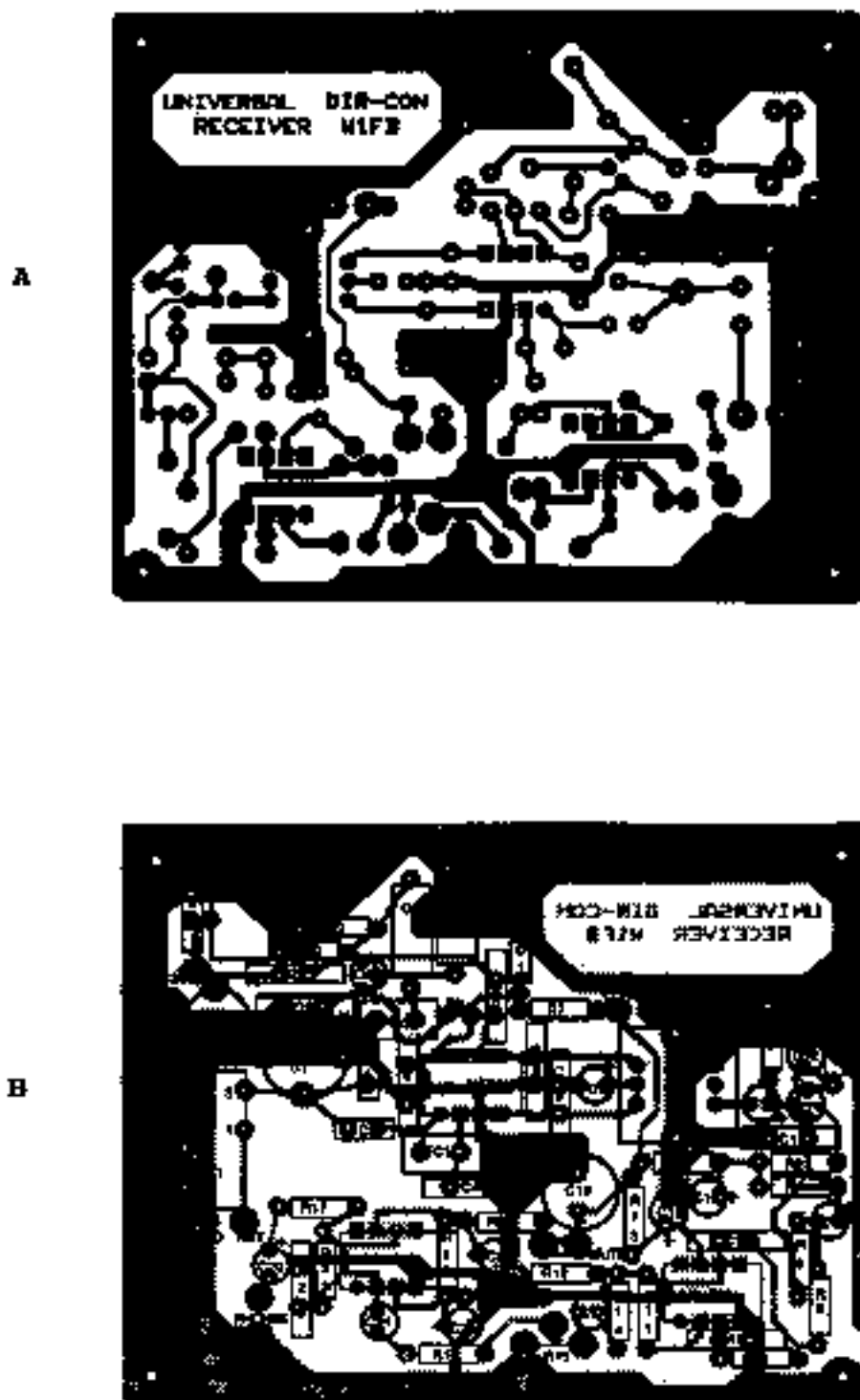
Fig 3-28 -- A scale etching template for the universal DC receiver is shown at A, as viewed from the etched side of the PC board. An Xray view of the PC board, as seen from the component side, is seen at B.

Performance Hints

Op amp U2 tends to latch up without some bias applied to pin 3. Latch up occurs in some op amp circuits if certain precautions are not taken. The phenomenon causes the op amp to saturate. This can happen instantly, or it can build up over seconds or minutes. The stage gain wanes until there is no output. Turning off the operating voltage and returning it restores performance until saturation occurs again. R5 in Fig 3-27 prevents U2 latch up. BiFET op amps are more prone to latch up than are 741 types of amplifiers, at least in certain kinds of circuits.

Overall receiver gain for the Fig 3-27 circuit is determined mainly by the values used at C12, R2 and R21. Greater U3 gain is possible by increasing R21 from 470k-ohms to, say, 1 megohm. The higher gains may tend to make the op amp unstable, which can cause it to self-oscillate at audio frequency. The gain of Q1 can be reduced by using a smaller capacitance value at C12 or a larger resistance value at R2.

The cutoff frequency for low-pass filter U2 can be changed by using different component values for C15, C16, R9 and R10. Equations for component selection are provided in **The ARRL Handbook** (see active filters section).

Make certain that the VFBO injection voltage to pin 6 of U1 does not exceed 3 volts P-P. Excessive injection power can cause strong harmonic currents to develop within the NE602 detector. This gives rise to all manner of spurious signal responses in the receiver tuning range. Too little injection power, on the other hand, reduces the conversion gain of the chip.

I measured the sensitivity of the Fig 3-27 receiver at 7 MHz. I was able to inject a 0.2-uV signal at L1 and locate it by tuning across the 40-meter band. A 10-uV signal from my URM-25 generator quieted the receiver background noise completely. The sensitivity may be somewhat lower at 14 MHz.

I detected a slight trace of unwanted AM detection when nearby out-of-band commercial signals were equivalent to 20 dB over S9 on my FT-102 S meter. The AM signal was very subliminal and did not cause a problem. The audio output from the DC receiver is very clean, except when R15 (gain control) is fully advanced for an S9+ signal. Very strong signals cause U3 to overload when R15 is set for maximum gain. All tests were made with 8-ohm earphones.

Although U2 rolls off the audio response at 700 Hz, SSB signals sound good, owing to the rolloff being gradual with only one section of filtering. High-pitch heterodynes are, however, greatly attenuated.

VFBO frequency stability is excellent. I measured only 5 Hz of short-term drift (1 minute) after each turn-on of the receiver at 40 meters. Long-term drift was less than 100 Hz over a one-hour period at a room temperature of 70 degrees F. I mounted L3 vertically on the PC board

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after applying two coats of General Cement polystyrene Q Dope. A dab
of epoxy cement was used to firmly affix the bottom on the L3 toroid
to the PC board. I also cemented C26 and C27 to the PC board to keep
the capacitors from moving about and causing frequency changes.

You can obtain better bandspread by using smaller values of capacit-
ance for the main tuning capacitor, C29. A 15-pF air variable permits
coverage from 6.9 MHz to 7.38 MHz (including CHU) on 40 meters. In any
event, you will want to use a vernier drive with C29 to make tuning
smoother.

Loudspeaker operation is possible by merely adding an outboard LM386
or similar audio IC after U3. Alternatively, you can buy a Radio Shack
no. 277-1008 mini audio amplifier with its built-in speaker. The unit
operates from a 9-V battery and costs under $15.

A NO FRILLS SUPERHET

Circuit simplicity is an objective of most of us who build QRP gear.
Not only are uncomplicated circuits easy to build and get working,
but the fewer the active stages the lower the current drain. Keeping
the parts count low reduces the cost of such a project. The receiver
described here was designed with those objectives in mind. Admittedly,
fewer parts means a tradeoff in performance and not as many operating
frills or conveniences; but adequate service can usually be obtained
from a minimum-parts receiver.

Fig 3-29 contains the schematic diagram of a simple superhet I design-
ed especially for this book. It serves as a tunable IF with which
converters may be used to obtain coverage of the HF bands. It tunes
from 1.7 to 2.1 MHz, thereby allowing a 400-kHz tuning range in the
chosen HF band when a down-converter is used.

The receiver can be used as it stands for reception of the 160-meter
MF band. The LO tuning range can be reduced for 160-meter-only use
(1800 to 2000 kHz). This may be done by using an LO tuning capacitor
that has the appropriate maximum capacitance.

No RF amplifier is required for the specified tuning range in Fig
3-29. The noise figure of the CA3028A active balanced mixer is more
than adequate for MF reception. The effective noise figures of the
outboard down-converters establishes the overall receiver NF during
HF reception.

You will observe that there is but one IF amplifier, and there is
no AGC. A 455-kHz Collins-Rockwell mechanical filter is specified
in Fig 3-29 for obtaining IF selectivity. I chose a used filter that
has an SSB bandwidth of 2.7 kHz. I bought my filter at a flea market
for $10. My filter has a center frequency of 450 kHz (an upper side-
band filter) and the part number is F450Z-7F. A good filter for this
receiver is the F455FD-25, no. 526-9692-010, which is a low-cost unit
offered by Collins. It has a 3-dB bandwidth of 2.5 kHz. You may wish
to consider one of the latest Collins low-mass filters. These are
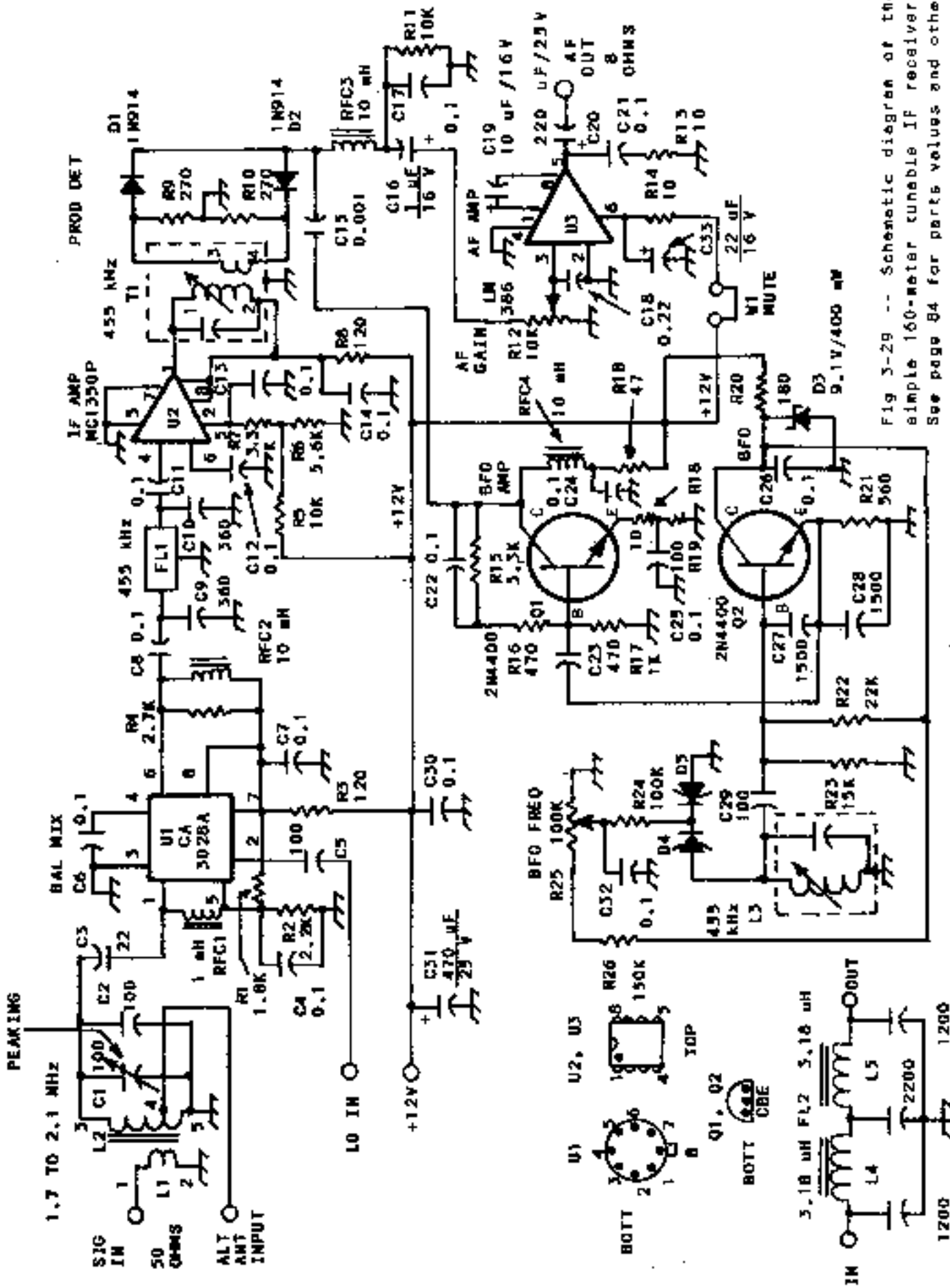
Fig 3-29 -- Schematic diagram of the simple 160-meter tuneable IF receiver. See page 84 for parts values and other information about this circuit.

Fig 3-29, continued -- Fixed-value capacitors are disc ceramic. Fixed-value resistors are 1/4-W carbon film or composition. Decimal value capacitors are in uF. Others are pF. Polarized capacitors are electrolytic or tantalum, 16 V or greater.

C1 -- Miniature air variable, panel mounted.
C29 -- NPO ceramic.
C27, C28 -- Polystyrene capacitor.
D1, D2 -- Small-signal silicon diode, type 1N914 or equivalent.
D3 -- 9.1-V, 400-mW Zener diode.
D4, D5 -- VVC tuning diode, Motorola type MV2109 (available from Hosfelt Electronics in Steubenville, OH). Range approx. 15-50 pF with voltage change from 1 to 9 volts dc.
FL1 -- IF filter (see text).
L1 -- Input link, 5 turns no. 28 enam. wire over grounded end of L2.
L2 -- 24 uH inductor. Use 65 turns of no. 28 enam. wire on an Amidon T68-2 toroid (red code). If tap is used, locate at 6 turns above grounded end of L2.
L3 -- Primary winding of a 455-kHz transistor IF transformer (see text). J.W Miller Co. no. 2067 or equivalent. Secondary not used.
RFC1 -- Miniature 1-mH RF choke (Mouser Electronics no. 43LR103).
RFC2 - RFC4, incl. -- Miniature 10-mH RF choke (Mouser Electronics no. ME434-1120-104K).
R12 -- Audio taper, 10K-ohm carbon composition potentiometer, panel mounted.
T1 -- Intermediate stage (white core) 455-kHz transistor IF transformer (30K to 500 ohms). Mouser Electronics no. 42IF102 or J.W. Miller 2067, or equivalent.
U1 -- RCA or Motorola differential amplifier mixer IC, TO-5 case.
U2 -- Motorola 8-pin DIP IF amplifier IC. Mount in socket.
U3 -- National Semiconductor LM386 audio power IC, 8-pin DIP. Mount in an IC socket.
W1 -- Jumper wire on PC board.
R25 -- Linear taper 100K-ohm potentiometer, panel mounted.

Transistors Q1 and Q2 need not be 2N4400s. You may use 2N4401 devices or 2N2222 or 2N2222A transistors without making circuit changes. Other transistors with similar characteristics may also be used.

Many of the small parts used in this circuit may be taken from a junked broadcast band transistor radio. For example, the oscillator section of an AM transistor radio tuning capacitor can be used for C1. The ceramic and electrolytic capacitors may be taken from the same radio for use in this receiver. The specified values for the electrolytic capacitors in Fig 3-29 are not critical. Variations as great as 30% will be okay. The tuned circuit and FL2 capacitor values are critical. These values should be maintained. Collins Rockwell 455-kHz IF filters are suitable for obtaining selectivity at FL1. These components are

available with bandwidths of 2.5 kHz (no. 526-8635-010) for SSB and 500 Hz (no. 526-8634-010) for CW. These rectangular mechanical filters have a volume of 2.2 cubic inches and are listed as low-cost filters.

Alternative 455-kHz IF Filters

If you are unable to locate a bargain-price mechanical filter and are unwilling to pay the price for a new unit, you may wish to employ a home-made half-lattice crystal filter. Two 455-kHz-range crystals with a frequency spacing of 1.5 or 1.8 kHz would be suitable for SSB and CW reception. Use a spacing of 700-800 Hz for a CW lattice filter. You may want to consider building a 455-kHz ladder filter for the Fig 3-29 receiver. Complete design details for this are given in the appendix of **W1FB's Design Notebook** (ARRL).

Another method for constructing a 455-kHz IF filter is to use three 455-kHz IF transformers in series or cascade with very light capacitive top coupling between them. You can create an SSB-bandwidth bandpass filter by adopting this technique.

Still another consideration is to change the receiver IF to some other frequency, such as 9 MHz, and alter the LO frequency accordingly. By changing the IF it becomes necessary to change the U2 IF transformer frequency, plus make similar modifications to the BFO, Q2.

More About the Receiver Design

D1 and D2 in Fig 3-29 function as a passive product detector. Output from the detector is routed to audio amplifier U3. It provides ample output for headphone operation, but is not adequate for driving a speaker, except for moderately strong signals. A 2N3904 audio preamp may be inserted between R12 and the input of U3 to boost the audio level sufficiently for weak-signal speaker reception. I added a little preamp of this type to my version of the receiver. It is built on a scrap of PC board (ugly construction) and is mounted near U3.

The BFO circuitry may appear grossly overcomplicated at first glance. True, a pair of 455-kHz crystals could have been used with a single 2N3904 beat oscillator, but I wanted to have my BFO variable in order to simulate IF-shift tuning. The varactor or VVC diodes, D4 and D5, allow the beat frequency to be shifted plus and minus 4 kHz. Also, new 455-kHz crystals are rather expensive! Broadband amplifier Q1 boosts the BFO level to 5 volts P-P before it is injected at the product detector. L3 is a 455-kHz transistor radio IF transformer. Likewise for T1 in Fig 3-29.

You will see a half-wave, low-pass filter at the lower left in Fig 3-29. It may be used at the input (terminal 1) of L1/L2 to provide additional front-end selectivity. This filter will help to reject stray LO energy from the down-converters you elect to use.

LO injection to pin 2 of U1 should be restricted to 2-V P-P. Excessive LO injection to the CA3028A will cause all manner of spurious HF-band

responses in the receiver tuning range, owing to the generation of harmonic currents within the mixer chip.

An L2 coil tap is shown in Fig 3-29 for use as an alternative antenna input, should you prefer this method to winding the L1 link. If the tap is used it should be located near the grounded end of L2 at 10% of the total coil turns.

A common problem that occurs when a receiver has no AGC is that very loud signals sound distorted. This is because the ratio of the IF signal to that of the BFO is too great. This may be solved by using a 500-ohm control at the input to L1, connected as an RF gain control. It is adjusted so that a strong signal sounds normal. I modified the Fig 3-29 IF circuit as shown in Fig 3-30. R1 functions as an IF gain control. The receiver is operated at maximum or near-maximum audio gain and the IF gain control is used to set the receiver gain for a comfortable listening level. This technique is the one we used many years ago when copying CW or SSB with a commercial AM/CW type of ham receiver that did not have a product detector.
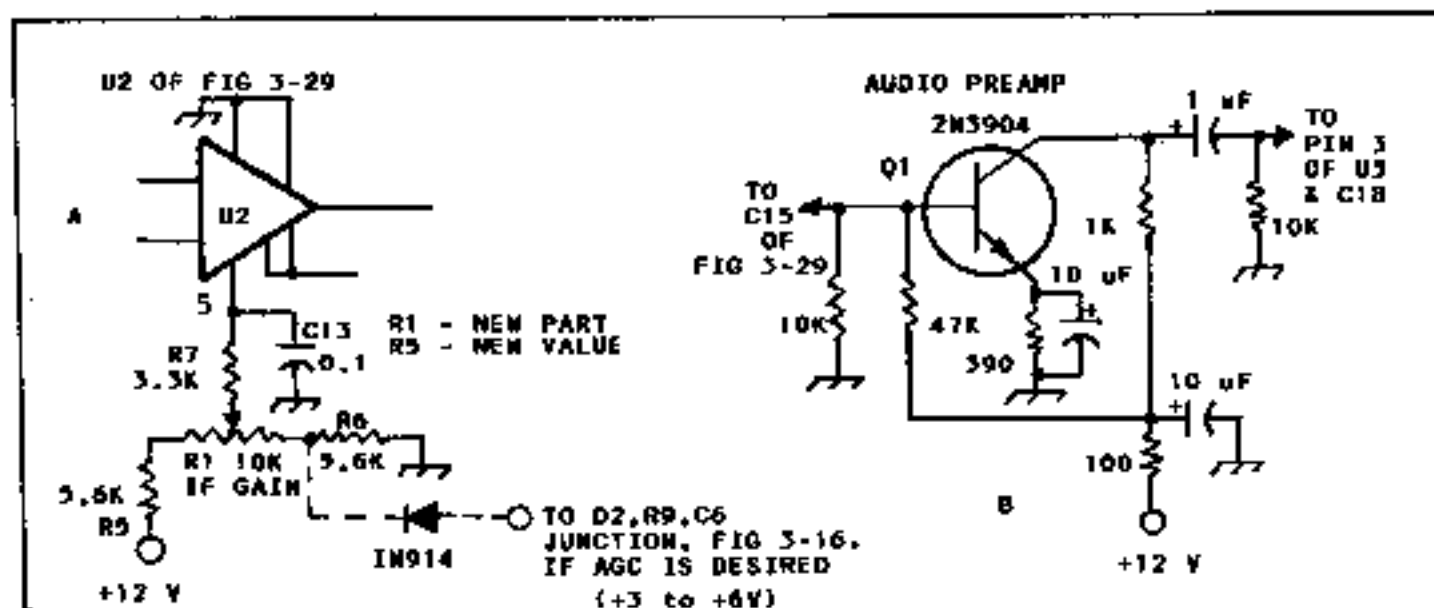
Fig 3-30 -- Circuit A shows how to add an IF gain control to the Fig 3-29 receiver. The audio preamplifier discussed earlier in the text is shown at B. It is added between the product detector and audio-output IC.

An enterprising experimenter can add an AGC circuit to this receiver by sampling and rectifying some of the audio energy ahead of the AF gain control, then applying it to pin 5 of U2 in accordance with the Motorola specifications for the MC1350P. This will eliminate the need for R1 in Fig 3-30. Maximum IC gain occurs when there is +5 volts on pin 5 of U2. Minimum stage gain takes place when +12 volts is fed to pin 5.

W1 in Fig 3-29 may be opened to allow an external control circuit to mute the receiver during transmit periods. If a CW sidetone signal

is supplied to pin 3 of U3, W1 is kept in place. Muting may then be done by opening the +12-volt feed to U1 and U2.

Receiver Local Oscillator

The stable VFO in Fig 3-13 is set up to be used with the Fig 3-29 receiver. No circuit changes are required other than shunting a 560-ohm, 1/4-watt resistor from the VFO output port to ground. The VFO output network is designed to look into a 500-ohm load. The resistor provides a constant load for the VFO (which C5 and pin 2 of U2, in the series arrangement, do not). Q3 in Fig 3-13 may otherwise break into self-oscillation for lack of a correct load.

Check to make certain there is no more than 3 volts P-P at pin 2 of the mixer (U2) when the VFO is connected to the receiver. If there is too high an injection level, reduce the capacitance of C5 in Fig 3-39 to obtain 2 volts P-P. If there is too little injection voltage, increase the C5 capacitance.   The P-P voltage may be measured by means of a scope. If you use an RF probe and VTVM to read the RMS voltage at pin 2 of U1, adjust the injection for 0.7 to 1 V RMS.

Some Final Thoughts

The sensitivity of the Fig 3-29 receiver is good. A 0.1-uV signal from a generator is audible in the headphones when the front end of the receiver is peaked by means of C1. Any signal that is 1 uV or greater can be copied Q5 if band noise on 160 meters is low.

The local oscillator should be in a shielded box for best stability. I made a box for my VFO by soldering together some sections of double-sided PC board. I mounted the VFO on the main chassis with two no. 6 spade bolts.

PC board artwork for this project is provided in Fig 3-31. Circuit boards are available from FAR Circuits (see earlier references) for $7.25 plus $1.50 shipping.

The audio-derived S-meter amplifier in Fig 3-16 may be used with this receiver. It connects to the collector of Q1 in Fig 3-30 B. The audio preamp provides more than ample drive for the S meter. However, it should be possible to connect the S-meter amplifier to RFC3 and C15 in Fig 3-29 and still obtain sufficient output voltage for the S-meter circuit, if the preamp is not included.

The pinout for most transistor radio 455-kHz IF transformers is the same. Use transformers that have white-coded cores. You can obtain T1 and L3 from a junk radio, or you may buy the transformers new from one of many surplus dealers.

If you use a mechanical filter at FL1, find out what the correct value C9 and C10 capacitors is. R4 should be chosen for the characteristic impedance of the filter. The values listed in Fig 3-29 are correct for my filter, but may not be for other filter numbers.

A

B
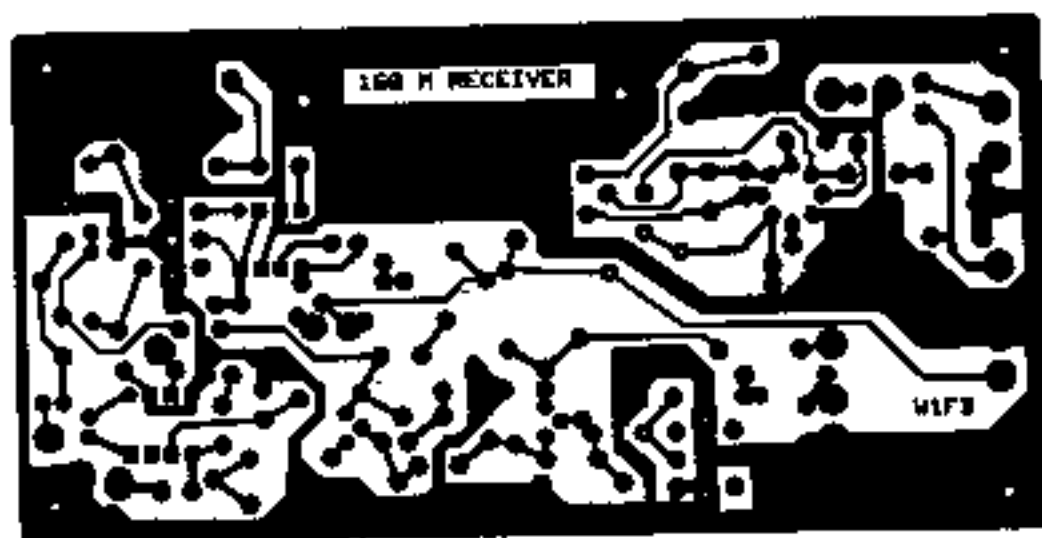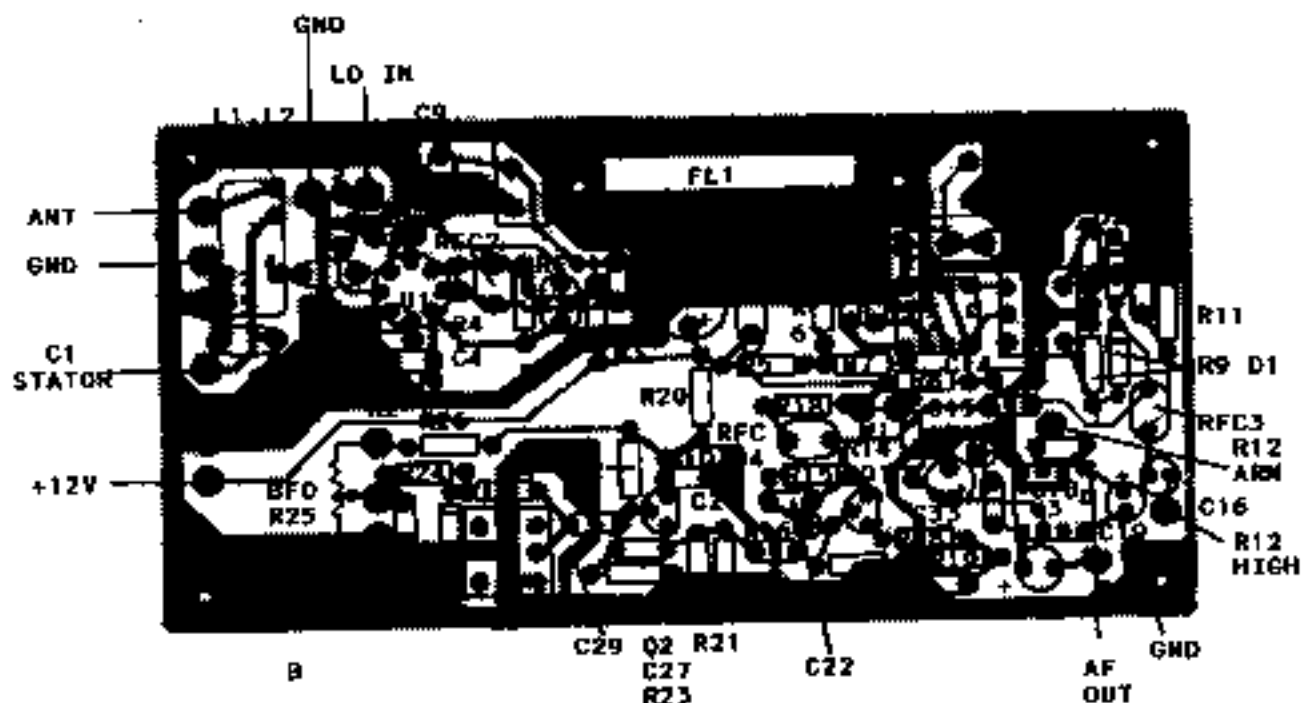
Fig 3-31 -- A scale etching pattern (etched side) is shown at A. Illustration B is an X-ray view of the PC board as viewed from the component side. **NOTE**: To conform to the input tuned circuit (L1, L2) in Fig 3-29, move C3 from the L2 coil tap (no. 4) to coil lead no. 3.

THE MATTER OF CONVERTERS

HF-band converters may be used to cover amateur bands that are not included in the tuning range of a receiver. A converter can be used ahead of almost any amateur receiver, and it may be set up to work over various receiver tuning ranges or bands. The converter LO frequency is selected for the receiver tuning range over which the desired converter coverage range is spread. For example, the 20-meter band may be heard by using a 20-meter converter and tuning the main receiver from, say, 3.5 to 3.9 MHz, the converter LO should operate at 17.9 MHz (14 + 3.9 = 17.9 MHz and 14.4 + 3.5 = 17.9 MHz). It can be seen from the foregoing example that the 20-meter band tunes backward from the 75/80 meter band. If the converter LO was made to operate at 10.5 MHz (also acceptable), the 20-meter band would tune from the low to the high end in accordance with the tuning on 75/80 meters. Thus, you have two choices when selecting the LO frequency for a converter.

Simple converters are often used for receiving 160, 80, 40, 30 or 20 meters. It is not necessary to use an RF preamplifier ahead of an active mixer (one with conversion gain) for these bands, although a preamplifier usually results in improved 20-meter performance, owing to improved NF. A low-noise preamp is necessary above 14 MHz for good weak-signal reception.

A Simple but Practical HF Converter

Fig 3-32 contains a circuit for a simple converter that is capable of good performance when used ahead of a receiver that has reasonable overall sensitivity. It may be used with the Fig 3-29 superhet receiver.

Fig 3-32 -- Diagram of a simple HF-band converter. See the table on page 90 for key parts values. This circuit operates on 80, 40 or 20 meters. See fig 3-33 for data about operation above 14 MHz.
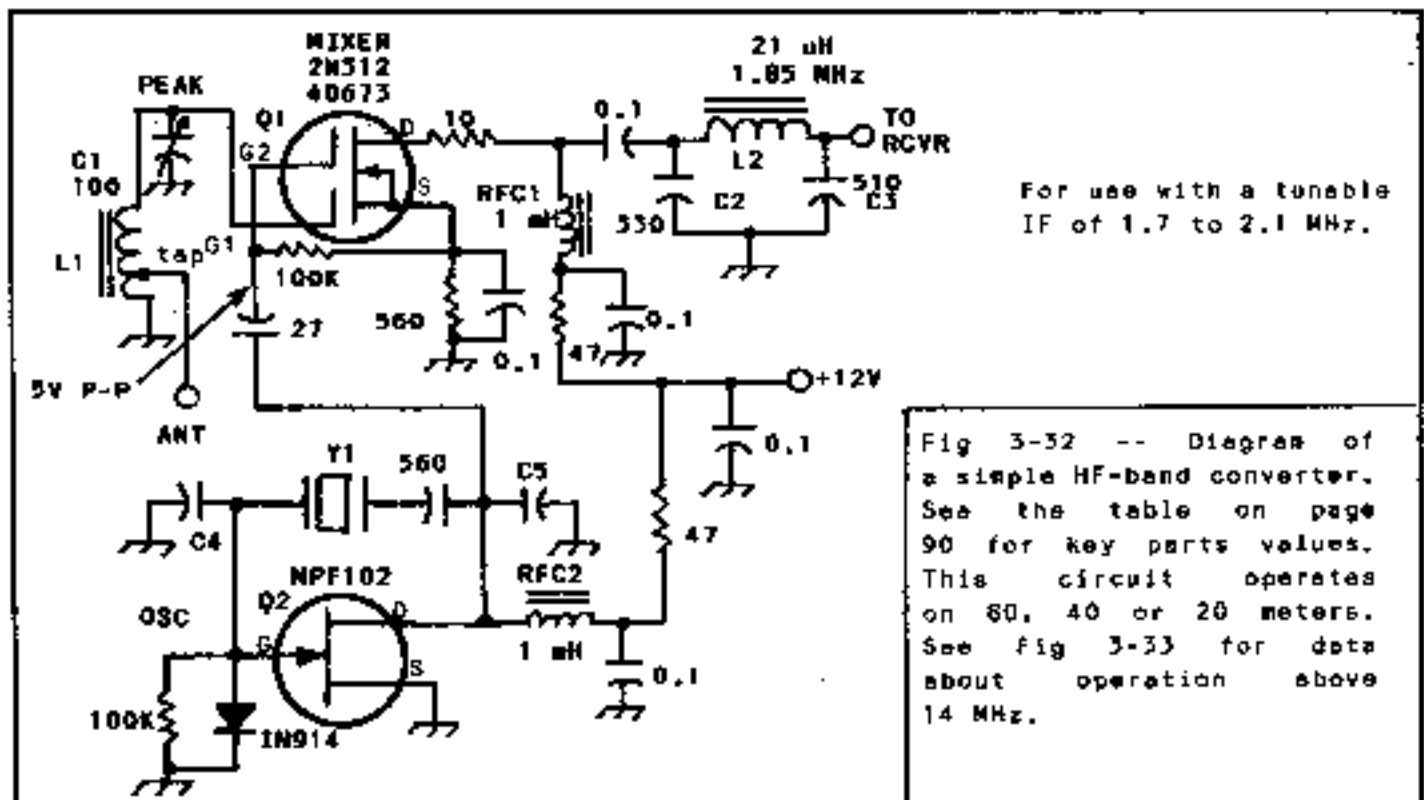
There is no PC artwork for the circuit in Fig 3-32. It can be built easily on a FAR Circuits universal breadboard or you may use the W7ZOI "ugly construction" technique.

Q1 is a single-ended mixer, which does not provide "world beater" performance with respect to high dynamic range, but it works fine otherwise. Any VHF dual-gate MOSFET can be used at Q1. If you prefer an antenna input circuit that uses a link rather than the coil tap shown, use 10% of the total L1 coil turns when winding the input link over the grounded end of L1.

A low-pass filter is used at the mixer output to minimize the passage of the LO frequency to the main receiver. A Pierce oscillator is used at Q2. C4 and C5 are the feedback capacitors. The 10-ohm resistor at the drain of Q1 is a VHF parasitic suppressor. Y1 is a fundamental crystal, 30 pF load capacitance. RFC1 and RFC2 are miniature Mouser Electronics RF chokes. C1 is a 100-pF miniature air variable. A 100-pF trimmer may be used if you are interested in a small portion of the 80 or 40 meter band. Repeaking is necessary for large changes in frequency, except for 20-meter operation.

TABLE 3-3

BAND (m)	C4, C5 (pF)	Y1 (MHz)	L1	L2
75/80	100	5.6	22 uH, 20 turns no. 28 enam. on an Amidon FT-37-61 ferrite toroid.	21 uH, 19 turns no. 28 enam. on an FT-37-61 toroid.
40	47	9.1	6 uH, 35 turns no. 28 enam. on an Amidon T50-2 toroid core.	SAME
20	22	16.1	3.4 uH, 25 turns no. 26 enam. on a T50-6 toroid.	SAME

Coil and capacitor data for the HF converter in Fig 3-32. The tap on L1 is 10 percent of the total L1 turns. Locate tap near the grounded end of L1. C2 and C3 are silver mica or polystyrene. C4 and C5 are NPO disc ceramic.

Converter Operation Above 14 MHz

Third-overtone crystals are necessary for converter operation above 14.0 MHz. This requires changing the Q2 circuit in Fig 3-32. The output of Q2 must be tuned to the desired crystal overtone. Also, an RF preamplifier is used ahead of the Q1 mixer in order to provide a low noise figure. The RF amplifier is a grounded-gate JFET that has a low noise figure and a gain of approximately 10 dB. Make certain that gate no. 2 of Q1 in Fig 3-32 has approximately 5 volts P-P of injection. If there is too little LO voltage, change the 27-pF gate 2 coupling capacitor to an appropriately larger value.
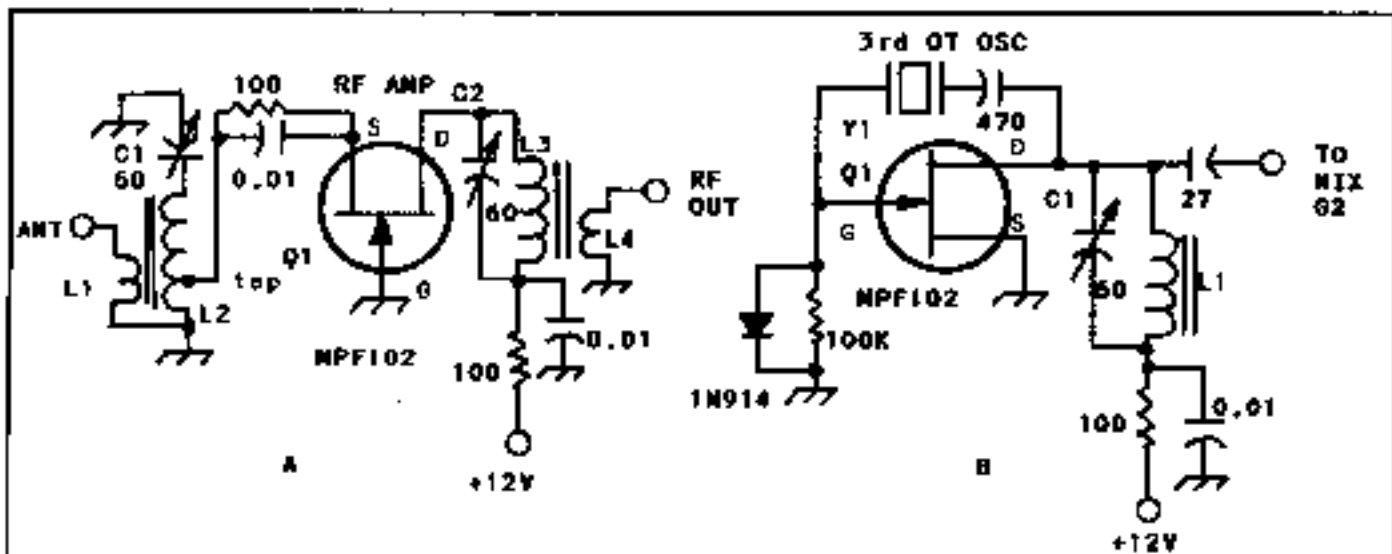
Fig 3-33 -- The circuit at A is a grounded gate RF amplifier for use ahead of the mixer in Fig 3-32 for operation above 14 MHz. The L2 coil tap is 1/4 the total coil turns, above the grounded end. L1 and L4 contain 10% the number of L2 or L3 turns and are wound over the lower ends of L2 and L3. C1 and C2 are ceramic or plastic trimmers. The gate lead of Q1 must be kept short to ensure circuit stability. The oscillator at B is for use in place of the Q2 circuit in Fig 3-32 for operation above 14 MHz. Y1 operates at the its overtone. C1 and L1 are resonant at the desired overtone. A ceramic or plastic trimmer is used at C1. Component values for these circuits are provided in Table 3-4.

TABLE 3-4

BAND (m)	L1,L4 (A)	L2,L3 (A)	L1 (B)	Y1 (MHz)
17	3 turns no. 28 enam.	2 uH, 25 ts. no. 28 enam. Tap L2 at 6 ts.	1.5 uH, 22 ts. no. 28 enam.	20.10
15	2 turns no. 28 enam.	1.5 uH. 22 ts. no. 28 enam. Tap L2 at 5 ts.	1.2 uH, 20 ts. no 28 enam.	23.10
12	2 turns no. 28 enam.	1.1 uH. 19 ts. no. 28 enam. Tap L2 at 5 ts.	0.9 uH, 17 ts. no. 28 enam.	26.90
10	2 turns no. 26 enam.	0.75 uH. 16 ts. no. 26 enam. Tap L2 at 4 ts.	0.65 uH. 15 ts. no. 26 enam.	30.10

Coil winding and crystal frequency data for the circuits in Fig 3-33. All coils are wound on Amidon Assoc. T37-6 (yellow) toroid cores. L1, L4 and the tap on L2 are located near the bottom ends of the related coils. L1 and L4 are wound over the main windings of L2 and L3.

Diode-Switched Two-Band Converter

Although the two-band converter in Fig 3-34 may seem rather "busy" with regard to circuit details, it is a fairly simple example of how band-pass filters may be used. Diodes are used to select the two filters for the band of operation. This eliminates the need to use a mechanical switch and long RF leads to and from the switch. Switching in this example is done only at dc, so the switch leads can be any convenient length.

The Fig 3-34 converter is designed for use on 40 and 75 meters. It is configured to operate with the tunable IF receiver in Fig 3-29. Additional bands can be added if the band switch contains additional contacts to accommodate the extra filters and oscillators. I chose to use separate oscillators in order to avoid the complication of diode switching the crystals and tuned circuits for a single transistor oscillator.

The bandpass filter design was established by W7ZOI some years ago. Data for these filters are presented in **The ARRL Electronics Data Book** (all editions). Forward bias (dc) is applied to the various switching diodes by means of S1 to turn on the diodes that are associated with FL1 or FL2. When bias is applied the desired diodes conduct and act as switches. When they are not turned on they present a high resistance to the RF energy and are effectively open-circuited.

A singly balanced JFET mixer, Q1 and Q2, is coupled to the filters by way of broadband, trifilar-wound transformer T5. Another trifilar, broadband transformer (T6) couples the mixer to the 50-ohm receiver input circuit. Although MPF102 transistors are specified for Q1 and Q2, slightly better performance can be expected from a pair of 2N4416 FETs. Dual-gate MOSFETs such as the 3N212 or 40673 may also be used in the mixer by tying gates no. 1 and 2 together and treating the FET as a single-gate device. No other circuit changes are necessary if this is done.

The band-pass filters require high-Q, stable capacitors in the tuned circuits. Silver mica or polystyrene capacitors are good choices in this part of the circuit. Miniature Murata trimmer capacitors are used at C3, C7, C11, C15, C20 and C26.

Q3 and Q4 in Fig 3-34 are fundamental oscillators. The feedback for these oscillators is regulated by means of C21 and C27. You may find it necessary to experiment with the capacitor values in order to ensure rapid starting of the oscillators. The values listed should be fine for crystals of normal quality and activity. Some crystals may require more feedback to make them oscillate. The collector tuned circuits must be tuned to resonance in order to make the crystals oscillate. You can add 60-pF trimmers in parallel or in series with Y1 and Y2 if you wish to move the frequency slightly for dial the purpose of receiver dial calibration. The series connection raises the operating frequency. Parallel use lowers the frequency.
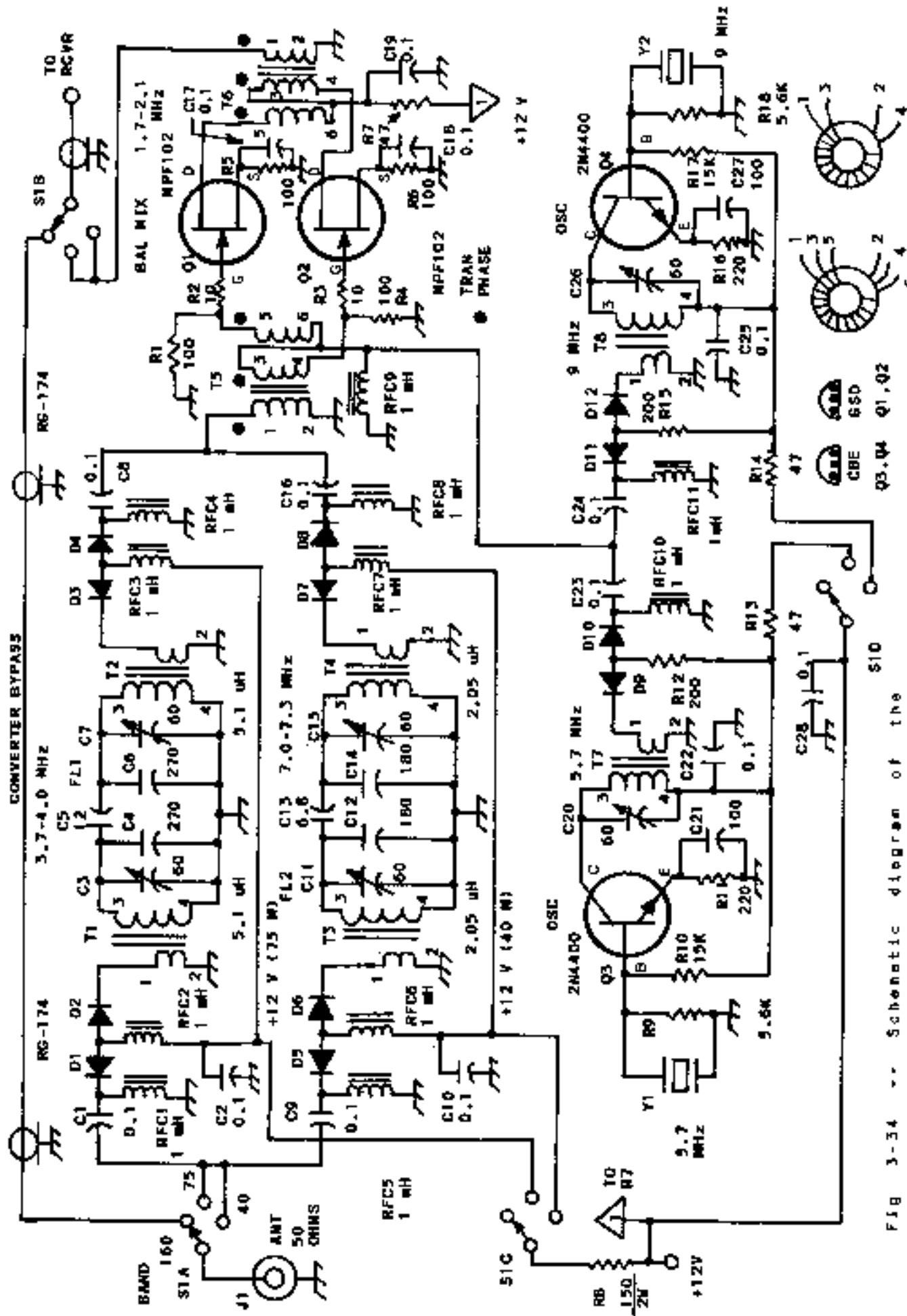
Fig 3-34 -- Schematic diagram of the two-band HF converter that may be used with the Fig 3-29 160-meter receiver. Parts list appears on the next page.

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Fig 3-34, continued.

Fixed-value capacitors are disc ceramic unless otherwise noted. The resistors are
carbon film or composition, 1/4 watt, except for R8, which is a 2-W unit. Trimmer
capacitors are miniature 60-pF Murata types or equivalent.

D1-D12, incl. -- Silicon small-signal diodes, type 1N4148 or 1N914.
RFC1-RFC11, incl. -- Miniature 1-mH RF choke. Mouser Electronics no. 43LR103 or
equiv.
S1 -- Four-pole, three-position phenolic or ceramic wafer switch. Mouser   no.
10YX043 or equiv.

T1, T2 -- Link windings have 2 turns of no. 26 enam. over grounded ends of main
windings. Main windings contain 16 turns of no. 26 enam. on Amidon FT-50-63 ferrite
toroid cores.

T3, T4 -- Links have 3 turns of no. 26 enam. wire over grounded ends of the main
windings. Main windings contain 23 turns of no. 26 enam. wire on Amidon Assoc.
T50-6 powdered-iron toroids.

T5, T6 -- Broadband transformer. Use 10 trifilar-wound turns of no. 26 enam. wire
on an Amidon FT-37-43 ferrite toroid.

T7 -- 15 uH primary, 28 turns of no. 26 enam. wire on an FT-50-63 ferrite toroid.
Link has 5 turns of no. 26 enam. wire.

T8 -- 6.3-uH primary, 36 turns of no. 28 enam. wire on a T50-2 toroid. Link has
6 turns of no. 26 enam. wire.

Y1, Y2 -- fundamental crystal, 20 pF load capacitance.

Losses occur in the diode switches and the band-pass filters. This
degrades the receiver NF. As a result, the Q1, Q2 mixer noise is
excessive. A low-noise preamplifier is necessary ahead of the filters
to compensate for the circuit losses and to improve the NF. Fig 3-
18C and D shows the MWA110 preamplifier I use at the input of the
the Fig 3-34 converter. It has a gain of 15 dB and a NF of 4 dB.
You may prefer to use a different type of preamp, but the MWA110
unit is broadband and requires no tuned circuits, and hence my choice.
With the preamp installed ahead of the Fig 3-34 converter, and with
the converter used with the Fig 3-29 receiver, a 0.1-uV signal from
my generator is audible in the speaker. Without the preamp a 1-uV
input signal is necessary to hear it in the speaker.

Adjustment of FL1 and FL2 is accomplished by applying a signal from
a generator at midband. The trimmers are adjusted for a peak signal
response. Repeat the process three times to compensate for interact-
ion of the filter sections. I bypass the MWA110 preamp when the system
is used for 160-meter reception.

In an ideal situation the preamp would be located between the output
of the filters and the input to T5 in Fig 3-34. This can be done
by opening the PC-board foil at terminal 1 of T5.
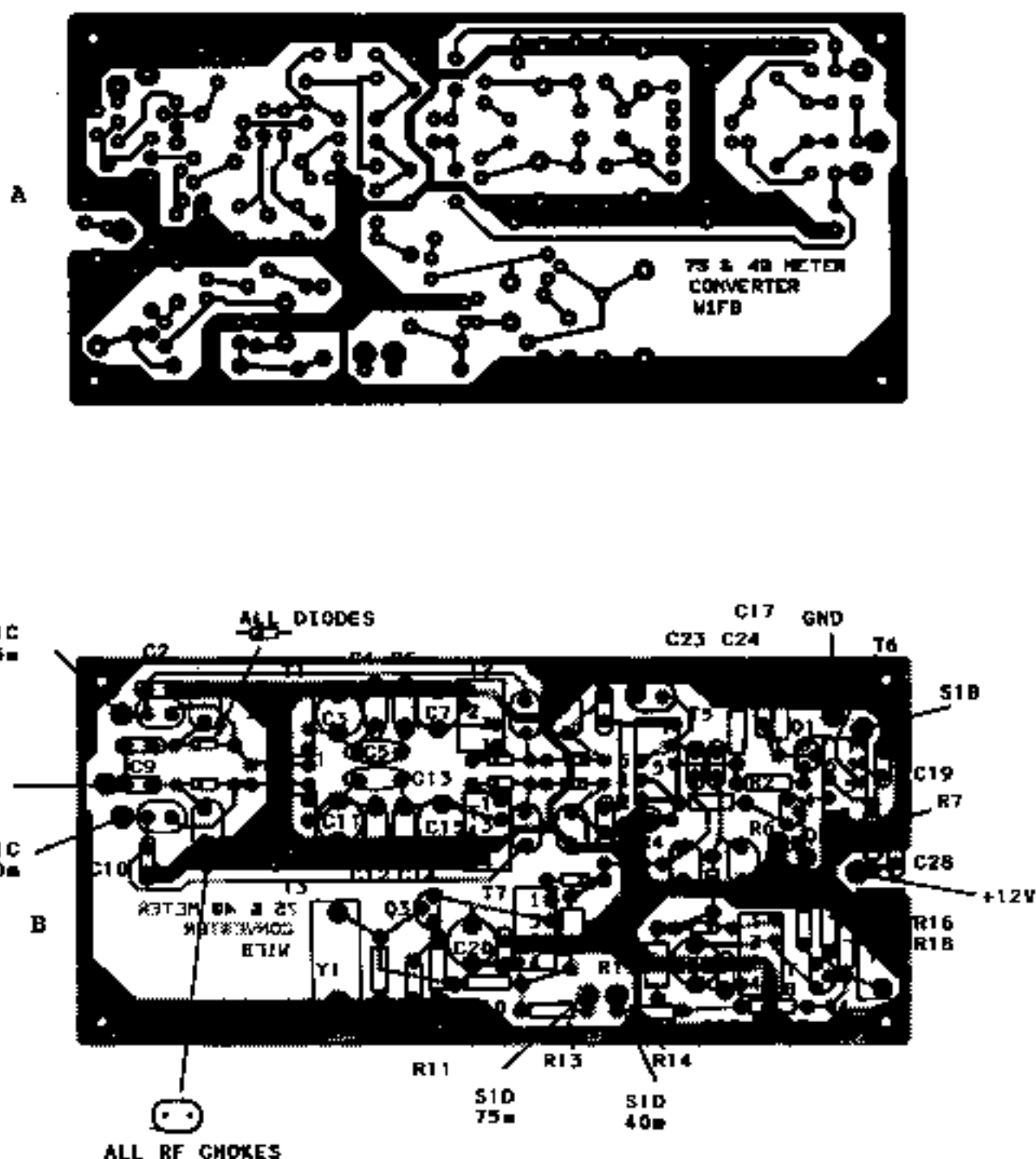
Fig 3-35 -- A scale etching template (etched side) is shown at A. The draw-
ing at B is an Xray view of the PC board as viewed from the component side.
This circuit board is available from FAR Circuits for $7.25 plus $1.50 for
shipping. RF chokes and diodes at B are not identified because all units
of each are the same type. Observe diode polarity. Y1 and Y2 hole spacing
if for International Crystal Co. PC crystal sockets.

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Alternative Broadband Preamp

ICOM Company uses an interesting broadband preamp in some of its HF transceivers. The circuit uses two JFETs in push pull. Broadband input and output transformers couple the amplifier to the input and output loads. This preamp is switched in and out of the receiver circuit at the will of the operator. Effective gain is approximately 10 dB and it has a low NF. Fig 3-36 shows the circuit. I have supplied the data for the devices and the transformers. That information was not available to me as this was written.
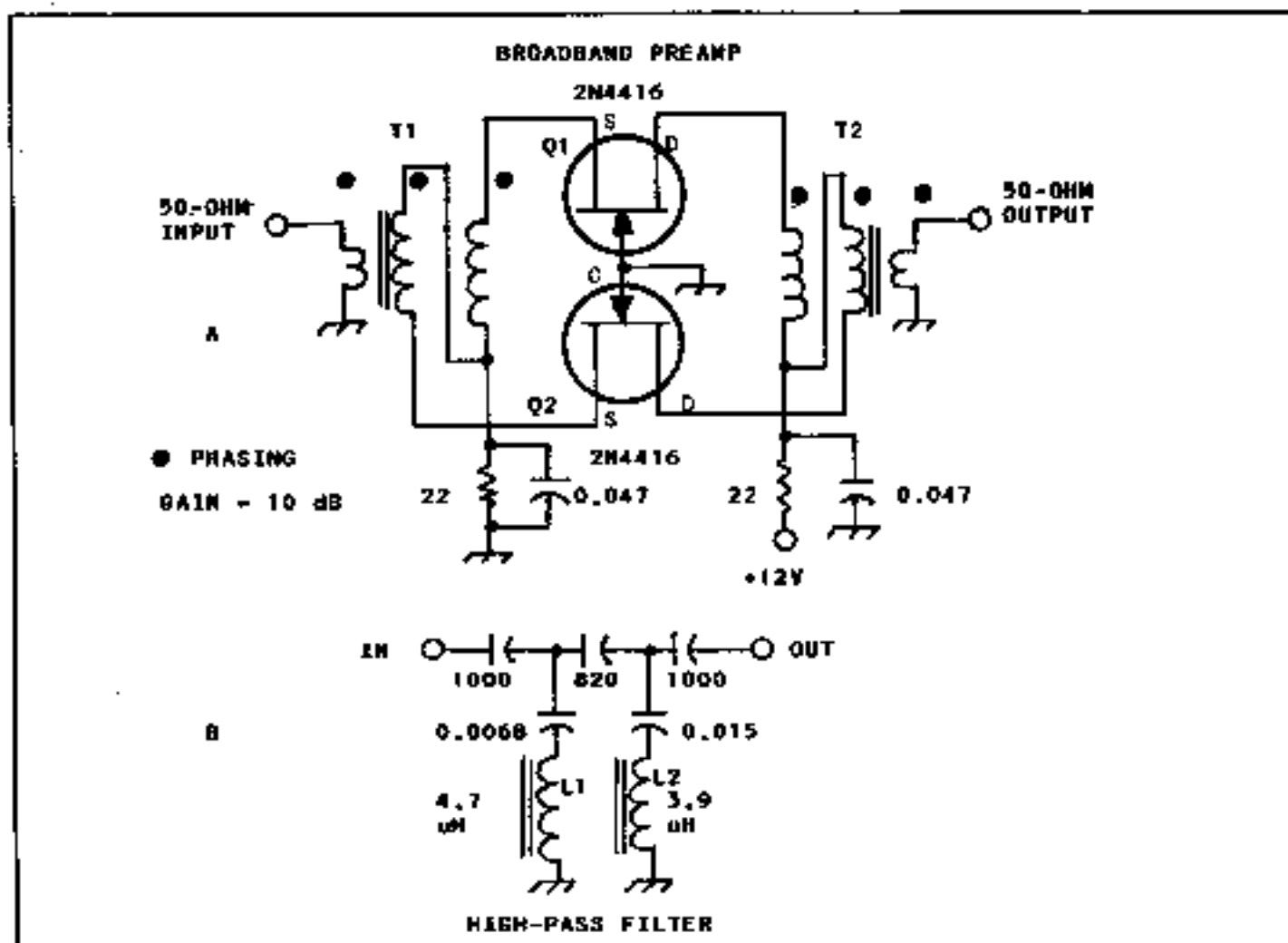
Fig 3-36 -- Schematic diagram of the ICOM broadband receiver preamp (A). The drawing at B is the ICOM high-pass filter that is used between the antenna and the preamp or receiver input circuit (switchable). It prevents commercial broadcast band energy from affecting receiver performance. It may be used with any receiver that has a 50-ohm input impedance. T1 and T2 at A contain 12 turns bifilar turns of no. 26 enam. wire on an Amidon FT-37-43 ferrite toroid core (850 mu, 3/8 inch OD). The input and output links for T1 and T2 have 4 turns of no. 26 wire for T1 and 2 turns of no. 26 wire for T2. L1 at B has 31 turns of no. 26 enam. wire on an Amidon T50-2 toroid and L2 has 28 turns of no. 26 enam. wire on a T50-2 toroid.

CHAPTER SUMMARY

Certainly, this chapter is not all-inclusive. I have tried to present a sufficient number of practical circuit ideas and design suggestions to enable you to "learn by doing," in the event you are a newcomer to circuit construction. If you are experienced at designing and building QRP equipment, chances are that you will find some new ideas here. But most importantly, try your hand at equipment construction. Don't be afraid to experiment!

A vast number of other receiver designs for QRP enthusiasts are presented in The ARRL's **Solid State Design for the Radio Amateur** and in **W1FB's Design Notebook.** I recommend that you add those volumes to your QRP library.

CHAPTER 4

QRP TRANSMITTERS & TECHNIQUES

The trend in QRP transmitter design and construction is generally toward equipment for CW operation. This may be due in part to the relative simplicity of a CW transmitter, plus the effectiveness of CW communications (at low power) compared to the success of SSB operation. In the latter situation we must contend with heavy phone QRM and high-power signals, whereas band occupancy and signal bandwidths are less troublesome in the CW bands. Another advantage associated with CW QRP transmitters is reduced current drain over a similar transmitter for SSB use. This goes hand in hand with battery operation, where current drain needs to be minimized.

QRP phone operation begins to make sense when we operate a 5- or 10-W SSB rig on 10 or 15 meters. The QRM, however intense at times, is more tolerable than on 20, 40 and 75 meters. Also, propagation on 10 and 15 meters is such that a low-power signal will get through, even when simple antennas are used. These considerations make these two bands ideal for QRP voice operation. The same is generally true of SSB operation on 12 and 17 meters, where QRM is usually at the lowest level for the HF bands, collectively.

I would like to again take this occasion to encourage you to build your own QRP gear. Not only will you save a lot of money by not buying commercial equipment, you will experience the incomparable joy of communicating over vast distances with equipment that you built and perfected. The learning experience is, by itself, a reward that is worthy of the effort. You won't have to send this kind of gear back to the factory for repairs. You can service it yourself!

Transmitter Output Power

We QRPers often think in terms of "miles per watt." This is part of the overall QRP challenge. The question becomes one of deciding how much output power is necessary to achieve our objectives. Many things influence watts versus miles. Certainly, band conditions at a given time are a primary factor. The frequency of operation also determines how far our signal will travel and be copied. The antenna quality must also be plugged into the equation. It goes without saying that a gain type of antenna that is high above ground (and matched to the feed line) is capable of making your QRP signal sound like a QRO one. This does not mean that a dipole that is not high above ground can't open the door to DX contacts when conditions are good: It does mean that you may have to work harder to attract attention when you call CQ, and your signal report may not cause you to strut to and fro in your ham shack. The real name of the game is "being copied Q5." There is no reason why you should not be as proud of an RST 559 report as one that comes back as RST 599.

You can have endless fun with power levels as low as 50 mW. Many of us enjoy building one-transistor crystal-controlled CW transmitters that emit flea power. Rigs of this variety provide a delightful kind of operation that is filled with challenges. Imagine earning your WAS award on 15, 20 or 10 meters with a battery-powered 50-mW transmitter! Certainly, this is a possibility if you are a dedicated ham. WAS, WAC and even DXCC achievements are entirely possible when using a transmitter in the 1- to 5-watt power class. In this context I recall working 42 countries in 30 days with a 2-watt QRP CW rig on 40 meters. My antenna was a square 40-meter loop that was erected vertically. The bottom side of the loop was only 6 feet above ground. Operation from Newington, CT netted many reports of RST 569 to RST 599. Reports as low as RST 349 were received also.

Your choice of transmitter power level will need to be founded on acceptable power consumption, transmitter complexity and the expense of the project. I recommend that newcomers start with, say, a one-transistor mW-class transmitter. Next, add an amplifier stage and raise the output power to 0.5 watt. Later on you can use this system to drive a 5-watt power amplifier. In other words, build the transmitter in stages and enjoy operating as you proceed.

Your QRP transmitter need not be a work of art. Some amateurs shy away from equipment construction because they feel that the end product should resemble a commercially manufactured unit. Some of us tend to place too much emphasis on aesthetics. An ugly transmitter or receiver is capable of working as well as or better than a unit to which you impart the commercial look. Many of my transmitters are simply tacked together on a scrap piece of PC board while using the W7ZOI "ugly construction" method.

TRANSMITTER PERFORMANCE

The on-the-air character of your QRP transmitter is your finger print, so to speak. You should strive to equal the performance of a properly operating commercial rig when you commit your signal to the airways. There is no justification for having a chirpy or clicky CW signal. Although these are common maladies, they can be resolved. We should be mindful also of the spectral purity of our transmitter emissions. Some designers of QRP transmitters ignore the harmonic currents in the output of the transmitter. In an effort to minimize the parts count and save money, they skimp on the harmonic suppression circuit. A single pi-network output circuit seldom offers ample harmonic suppression to enable the equipment to comply with FCC regulations. Not only that, excessive harmonic output leads to TVI and RFI problems. Your QRP transmitter should be designed to ensure that all spurious emissions are 40 dB or greater below peak output power.

Some builders attempt to use vacuum-tube design principles when they construct solid-state transmitters. Not only does this approach result in dreadful efficiency, it can cause self-oscillations and destruction of the transistors. Transistors are low-impedance devices. Tubes have high-impedance input and output characteristics. It is worth saying again that maximum power transfer occurs only when unlike

impedances are matched. We need to remember also that transistors amplify current while vacuum tubes amplify voltage. Low impedances associated with transistors are the result of relatively low operating voltage and moderate to high device currents. Tubes, on the other hand, operate at high voltages (comparitive) and low current.

An ideal CW output waveform has slightly rounded leading- and trailing edge corners. A rise and fall time of 5 ms will yield a clickless note. Examination of the keyed output waveform can be accomplished with a scope. You can vary the shaping-component values until you obtain a click-free waveform.

Chirp generally results from oscillator keying or keying a buffer stage that presents a widely changing load characteristic to the oscillator. Incorrect oscillator feedback also causes chirp. You may experience chirp also if your crystal is sluggish. If unwanted RF gets into the oscillator circuit (such as from the PA stage) it can cause the keyed signal to be chirpy. Stage isolation is, therefore, an important consideration. This does not mean that each stage needs to be in its own shielded compartment. Rather, it suggests that care must be exercised in the layout (straight-line layout is best) of the PC board and each stage should have an RC or LC decoupling network in its Vcc lead. This prevents stray RF from migrating from stage to stage along the +V supply line.

Preventing Self-Oscillation and Device Damage

High-Q circuits contribute to transistor self-oscillation. We can prevent this condition in some circuits by applying de-Qing techniques. For example, something as simple as a 10- or 15-ohm non-inductive resistor, when placed in a collector lead near the transistor, can prevent VHF parasitic oscillations. A miniature ferrite bead will often resolve the same problem. These components ruin the circuit Q at VHF and UHF while allowing HF energy to pass unimpeded. This technique is illustrated in Fig 4-1A.

The Vcc decoupling method discussed above may also be employed. Fig 4-1B shows the details for this preventive or corrective measure. The decoupling components need to be located as close to the transmitter as practicable. Likewise for parasitic-suppression components.

It is vital that the input and output parts of a transistor amplifier circuit be physically isolated from one another, within practical construction constraints. When the input and output circuits are close to one another there can be unwanted mutual coupling, which allows positive feedback to occur, and this causes oscillation. In a worst case example we may need to install a shield baffle or divider across the transistor to help isolate the input circuit from the output one. This is often required in order to stabilize VHF and UHF small-signal amplifiers. It may be necessary even at 21 or 28 MHz. The greater the amplifier gain the more pronounced the likelihood of self-oscillation.

Fig 4-1C illustrates a common method for taming an RF power amplifier

that uses broadband input and output transformers. A capacitor that is connected from collector to ground will help prevent VHF parasitic oscillations while suppressing VHF harmonic currents. This capacitor should have a capacitive reactance that is four to five times greater than the collector impedance in order to prevent signal loss at the operating frequency. For example, if the collector impedance is 15 ohms ($Z = VCE^2$ divided by 2Po) the capacitor XC should be no less than 50 ohms. Po is the amplifier output power and VCE is the collector to emitter dc voltage. Thus, for operation at 7 MHz the collector shunt capacitor must be 380 pF or less.
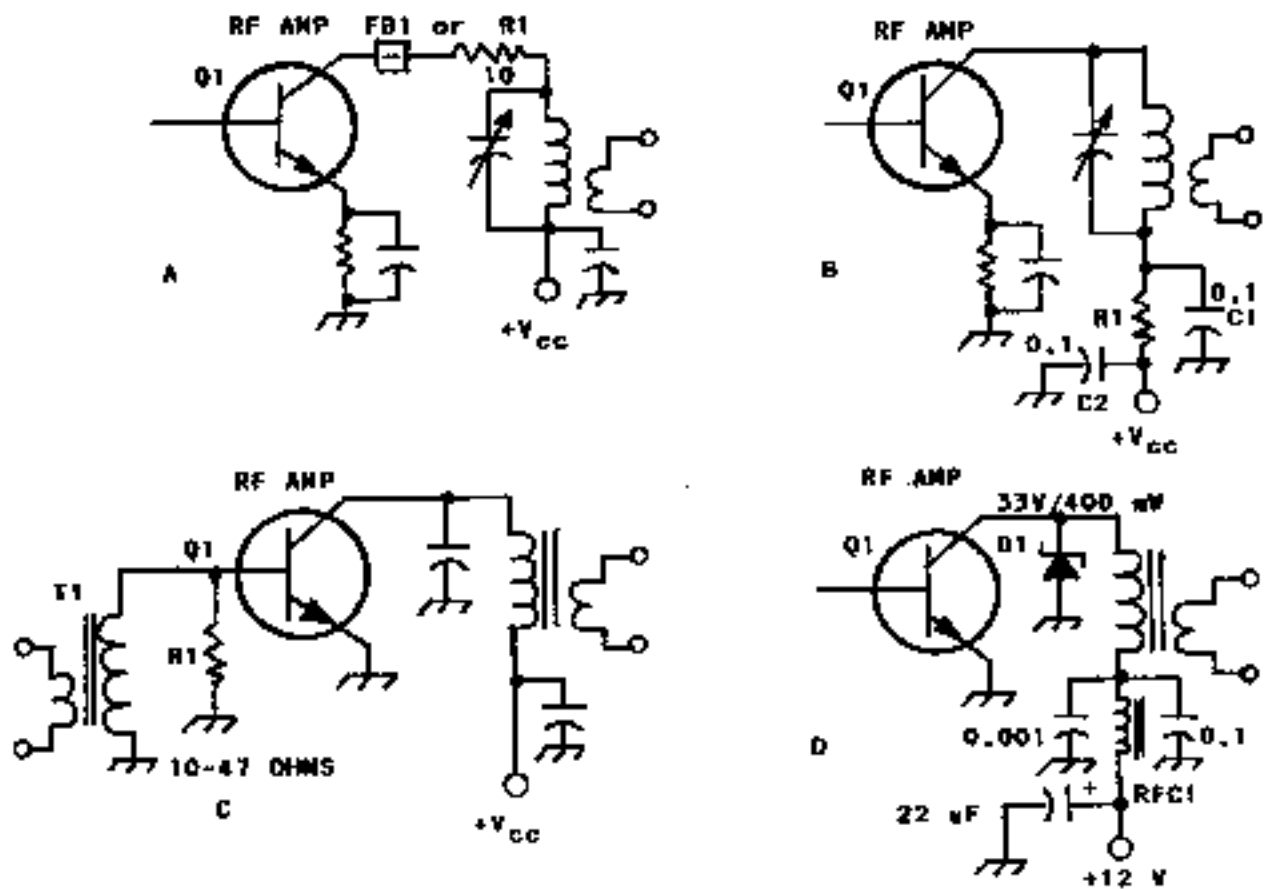
Fig 4-1 -- Circuit A shows suppression of VHF and UHF parasitics. FB1 is a mini 850-mu ferrite bead. Alternative component R1, which is a 1/4-W carbon composition resistor, may be used in place of FB1. Circuit B illustrates how a Vcc decoupling network (C1, C2 and R1) is used. R1 is typically 33 to 100 ohms for small-signal amplifiers that draw up to approximately 25 mA. Circuit C shows how a 1/4-W resistor may be used to lower the Q of the broadband input transformer. Alternatively, a mini ferrite bead may be used over the grounded-end lead of the T1 secondary. Use the largest value of resistance that prevents amplifier instability. R1 does consume some of the RF input power. Circuit D shows a decoupling circuit for power types of RF amplifiers. RFC1 (5 to 10 uH) is used in place of a resistor to prevent voltage drop to Q1. D1 protects Q1 from dangerous peak RF and dc voltages.

Fig 4-1D provides information about decoupling the Vcc line when the transistor draws a substantial amount of current. A resistor in place of RFC1 would cause a prohibitive voltage drop in the Vcc line to Q1. An RF choke for this circuit can be made from an 850-mu 1/2-inch OD ferrite toroid by winding five or six turns of heavy gauge magnet wire on the core. The wire gauge chosen must be capable of passing the Q1 current without causing a significant voltage drop. Three values of decoupling capacitor are listed. The 0.001-uF unit provides effective VHF bypassing and the 0.1-uF capacitor is effective at HF and MF. The 22-uF capacitor works effectively as an audio, VLF and LF bypass. These capacitors help to prevent amplifier self-oscillation from audio to VHF.

Zener diode D1 in Fig 4-1D may be added to protect Q1 from damage in the event of self-oscillation or high SWR, at which time peak-to-peak collector voltage can exceed the maximum safe device value. D1 also protects the transistor from dangerous spikes that may appear on the Vcc line. The D1 internal capacitance acts as a collector bypass capacitor. Its value is minimal and does not affect the amplifier performance at HF. D1 does not conduct until its barrier voltage is reached by RF or dc voltage. A suitable rule of thumb when selecting this Zener diode is to select a conduction voltage that is roughly three times the VCE. Remember that the sine-wave excursion at the Q1 collector, when it is amplifying RF energy, can swing to twice the supply voltage. Hence, a 24-V Zener diode would not be suitable in this circuit.

Transmitter Gain Distribution

The design of a multistage transmitter is not a casual matter. All too often an amateur will "marry" circuits from a collection of published articles, only to find that the transmitter performs poorly. Excessive drive is as bad an event with transistors as it is with tubes. Too little RF excitation, on the other hand, results in low output power. Ideally, we would study the transistor data sheets and learn how much input power (Pin) is needed to obtain the rated output power (Po). If the driving power to a particular RF amplifier is too great we can drive the device into saturation (the point at which no further output power is available no matter how much we increase the driving power). Excessive collector current can result and the stage becomes prone to the generation of excessive harmonic currents. The transistor can become destroyed immediately in a worst-case situation.

A viable method for optimizing your transmitter is to test it stage by stage. Observe the output power of the stage under test by using a scope or RF probe and VTVM. Terminate the stage with a resistor that matches its normal in-circuit load impedance. Adjust the output power of the preceeding stage until no further increase of output power is noted for the stage under test. Now, reduce the drive slightly below the saturation point. Proceed with this method until all stages of the transmitter have been set just below the saturation point. This will ensure a cleaner output signal from each stage and it will protect the transistor from damage.

Broadband versus Narrow-Band Amplifiers

You may question the usefulness of broadband amplifiers with respect
to narrow-band ones. Each has its positive qualities. It is necessary
to realize that when we use a broadband amplifier we must trade gain
for bandwidth. Therefore, the gain for a narrow-band amplifier might
be 12 dB. The same transistor, if used in a broadband amplifier for
the same frequency, might yield only 8 dB of gain. A narrow-band
amplifier is one that has a tuned circuit at the input, output or
both.

Narrow-band RF amplifiers are more prone to instability than broad-
band ones. This is because they have greater gain and because they
may not be matched properly to their input and output loads. This
mismatch condition is the result of haphazard circuit design. This
is one reason why we can't apply vacuum-tube design techniques to
transistors.

Broadband amplifiers enable us to have transmitter stages that do
not need to be band switched. The exception is when the broadband
amplifier has harmonic filters that need to be switched. Fig 4-2
shows a broadband and a narrow-band RF amplifier.
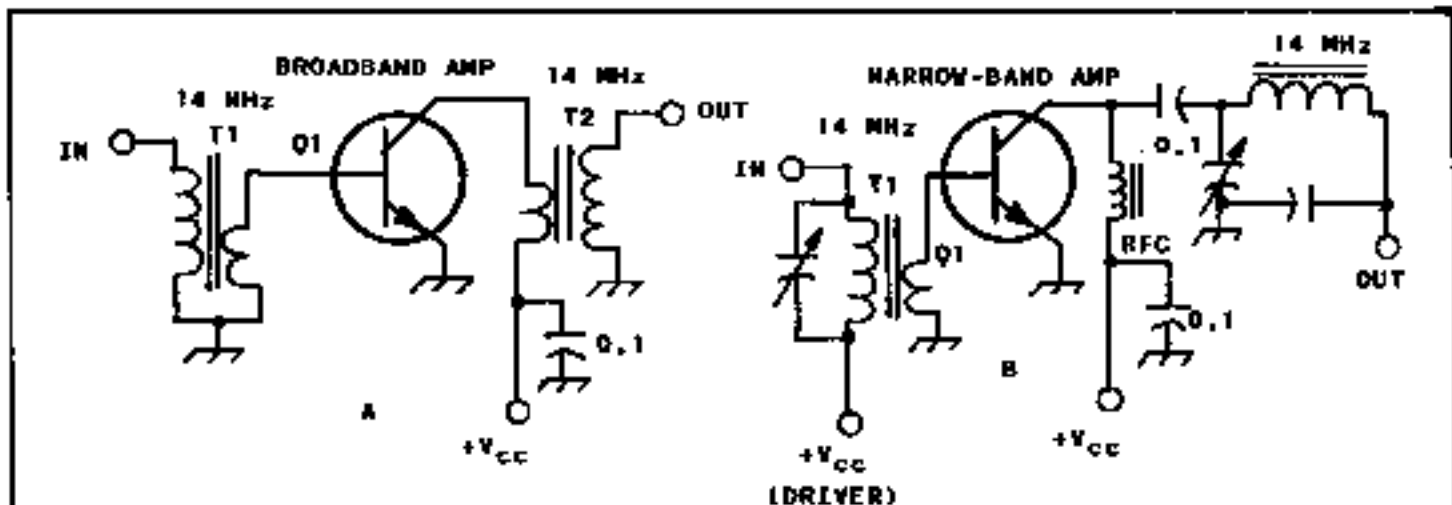
Fig 4-2 -- Examples of broadband (A) and narrow-band (B) RF amplifiers.
T1 and T2 at A are untuned. T1 at B and the output pi network are resonant
at the operating frequency. The broadband and tuned transformers and networks
must be designed to match the transistor input and output ports to the input
and output load impedances.

The circuit in Fig 4-2A uses broadband transformers at the input
and output ports of Q1. Generally, these are ferrite-core devices
(toroid or balun core). For MF and HF operation the core should
have an initial (u_i) permeability of 800 to 900. A u_i of 125 is suit-
for VHF amplifiers. Powdered-iron cores are not suitable for use
in broadband transformers because of their low permeability. The low
permeability dictates that an excessive number of wire turns be used
in order to obtain suitable performance at the low frequency end
of the transformer operating range.

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The amplifiers in Fig 4-2 are configured for class C operation. In practice, either of them may be biased for class AB operation by applying approximately +0.7 V of forward bias the the transistor base via the input transformer (T1) secondary winding.

Linear broadband amplifiers (class A or AB) produce less harmonic output than is generated in a class C amplifier. If the linear amplifier is not driven into distortion, we can often use it without a harmonic filter at the output port. This assumes that the input waveform is essentially free of distortion. Class C RF amplifiers, on the other hand, generate a considerable amount of harmonic current. It is not unusual to find the second- and third-harmonic energy only 10 to 15 dB below the fundamental frequency at the collector of a class C amplifier. It is for this reason that TVI and RFI can be a problem, even though the transmitter has an output power of one or two watts. This can occur when the output network is not capable of suppressing the harmonic currents. Under no circumstances should you exclude an effective harmonic filter at the output of the last transmitter stage. This applies to linear and class C types of amplifiers.

Transistor Operating Temperature

Heat is the enemy of many electronics components. This is particularly true of semiconductors. The question becomes one of deciding how hot a particular transistor or diode can become before damage occurs. It is the device junction temperature that determines the safe operating power dissipation. Since we have no convenient way to monitor the internal temperature of a semiconductor, we need to pay heed to the case temperature. Even through some devices are designed to operate at very high temperatures, I prefer to follow my own rule of thumb when in doubt. If the case of a transistor is so hot that it is uncomfortable to the touch, I take measures to cool it down. This requires the use of a heat sink or a larger sink than I am using. I don't worry if the transistor feels quite warm to the touch if it is being operated within the manufacturer's safe limits. I do, however, feel better about the matter if the device is only slightly warm to the touch. It provides a comfortable margin of safety when sufficient heat-sink area provides cool or warm operation.

The transistor that has a heat sink should have the best thermal bond you can manage. The mating surfaces of the transistor and its sink should be reasonably flat and smooth in order to assure good bonding. Also, a thin layer of heat-sink compound (a mixture of zinc oxide and silicone grease) should be applied to the mating surfaces before the device is bolted to the sink.

Small plastic transistors, such as 2N2222s or PN2222s, can be cooled by using a thin layer of epoxy cement to affix a tab of copper, brass or aluminum material to the transistor case. The tab should be clamped securely to the device case until the cement hardens. I have used short pieces of copper or brass tubing as heat sinks for the smaller transistors. I glued them to the device body with epoxy cement.

Selecting an RF Power Transistor

Although we tend to buy transistors that are designed for RF power amplification, and often pay a premium price for them, we can use almost any transistor that has a sufficiently high fT rating. Many audio power transistors work well as RF amplifiers. Those with fT ratings of 100 MHz or higher can provide excellent performance at the MF and HF frequencies. As a rough rule you may use any transistor that has an fT which is five times higher than the proposed operating frequency. In practice, lower fTs can be acceptable at the risk of reduced transistor gain. The fT is the frequency at which the transistor gain is unity or 1. At frequencies above the fT, the transistor exhibits a signal loss rather than a gain.

Some transistors are specified as high-speed switches. This generally implies that the transistor is suitable for RF service if the fT is sufficiently high. In order for a transistor to be a good high-speed switch or RF amplifier it must have low internal capacitance and resistance, since these components establish a time constant. The greater the R and C the longer the time constant and hence the lower the effective operating frequency. This applies equally to diodes.

We need to be cognizant also of the maximum Vce (collector to emitter voltage) and PD (power dissipation) set by the manufacturer. For RF amplifier operation from a +12-V supply, select a transistor that has a 30-V or greater maximum safe Vce. This allows for the RF voltage swing at the collector, which will be up to twice the Vce. PD concerns the operating current versus the operating voltage, and it is expressed in watts at 25 degrees C. There is also a maximum steady-state collector-current rating. This should not be exceeded under any circumstances.

The maximum PD poses a common question: If a transistor is rated at 10 watts maximum PD, how much dc input power may we have in an RF amplifier? I like to stay on the safe side of things by restricting the dc input power to 0.75 or less the PD rating. Therefore, in the case of a 10-W transistor I would restrict the dc input power to no more than 7.5 W. Transistor efficiency being what it is in an RF amplifier, we would expect to obtain approximately 3.5 W of RF output power under this condition. Most transistor RF amplifiers are 50-60 percent efficient, although I have seen efficiencies as high as 70% for class C transistor amplifiers that were optimized. I have obtained even greater efficiency when using a class C power MOSFET RF amplifier.

SWR-protected transistors are available for RF service. These are referred to by some manufacturers as BETs (ballasted-emitter transistor). In effect, numerous bipolar transistors are formed on a common silicon substrate. The transistor bases and collectors are connected in parallel, internally. The emitters each have an internal 1-ohm resistor and the ends of the resistors opposite the emitters are joined to form a single emitter terminal on the case. These resistors equalize the currents of the transistors and prevent any

one transistor from hogging the current. BETs are excellent transistors for us to use for RF power amplifiers if we are willing to pay the price for them. Technically speaking, they are capable of withstanding open or shorted load conditions without being damaged. But as we learned earlier in the chapter (Fig 4-1D), a Zener diode may be bridged from collector to ground to protect the device when the SWR is high. This permits us to safely use nonballasted transistors in our RF amplifiers.

As an experimenter it is worth your while to obtain transistor data books from the semiconductor manufacturers. This will enable you to look up the base diagrams and learn what the device ratings are. Two books that I find invaluable are the **Motorola RF Device Data Book** and the **Motorola Small-Signal Transistor Data Book**. These are available from Motorola Semiconductor Corp., P.O. Box 20912, Phoenix, AZ 85036. Books of this type are available also from other semiconductor manufacturers, such as National Semiconductor Corp. and Siliconix, Inc.

Many small-signal transistors, such as the 2N2222 and 2N4400 devices, are excellent for use as QRP RF power amplifiers. A 2N4400 or 2N4401, for example, is capable of producing up to 0.5 W of output power when operated at +12 V dc. Additional power may be obtained by using two or more of these transistors in parallel. If you plan to use them in this manner, install a 1-ohm, 1/4-W resistor between each emitter and ground. This will help to equalize the current in each transistor. These emitter resistors need not be bypassed because the floating resistors create very little degenerative feedback. Excessive degeneration would cause reduced transistor gain and could encourage unstable operation. A push-pull pair of 2N4400s result in a very nice 1-W RF amplifier. This provides a bargain-price amplifier because these transistors can be purchased for as little as 10 cents apiece. I have actually extracted 1-1/4 watts of output power from a single 2N4401 transistor at 7 MHz. This represents an unsafe output power for the device. I have, however, operated a single 2N4400 or 2N4401 at 0.75 W of output after affixing a small heat sink to the plastic case. I used a 3/8 inch piece of 1/4-inch ID brass tubing as the sink. I flattened the tubing slightly to provide a snug fit between it and the transistor. The space within the heat sink was then filled with epoxy cement and allowed to set. This brings up the subject that follows.

Thermal Runaway

The phenomenon of thermal runaway can best be equated to the well known "domino effect." The affected transistor destroys itself quickly from excessive internal heat. This effect is caused by allowing a transistor to draw excessive current or permitting it to operate without an adequate heat sink. In either situation the culprit is heat. As the transistor heat increases, so does its gain. The greater the gain the greater the current taken by the transistor, and the greater the internal heat. This process continues until the device is destroyed. Excessive heat melts a semiconductor junction and causes an open condition. Excessive voltage, on the other hand, can puncture a semiconductor junction and cause a shorted condition.

Analysis of a defective diode or transistor can be carried out by measuring the resistance (or absence of it) between the elements of the device. This provides a clue concerning the cause of a failure. Adequate heat-sink area and safe operating current prevents thermal runaway.

A Problem with Transistor Gain

We need to be aware that the power gain of an RF transistor is specified at a particular frequency -- generally the upper design frequency for the device. For example, a specified 10-W RF power transistor may be characterized in part by a gain of 13 dB at 30 MHz. At some frequency above 30 MHz the gain will begin to fall off. Further increase in operating frequency will eventually bring the device to its unity-gain point (fT). This is not the case when we decrease the operating frequency. This results in escalating gain above the 13-dB example point. The greater the amplifier gain the more likely the occasion for (1) thermal runaway and (2) instability. In theory, the gain of a bipolar RF power transistor increases 6 dB for each octave the operating frequency is lowered. Thus, if a device is specified to provide 13 dB of gain at 30 MHz, the gain at 15 MHz could be 19 db, and at 7.5 MHz it might be as great as 25 dB! Although the gain does increase as the operating frequency is lowered, it does not equal the theoretical amount. However, there is a significant gain increase and this can cause all manner of problems. Fortunately, this phenomenon does not occur when we use power FETs. The simple cure for the gain-increase problem is to progressively reduce the P_{in} (driving power) to the amplifier as the operating frequency is lowered, thereby keeping the collector current within the safe rating for the device. A better technique is to apply negative feedback, as described in **W1FB's Design Notebook**. This serves as a gain-leveling network, and the driving power can remain essentially the same for any operating frequency below the transistor fT. Some designers use lossy networks at the input of the power amplifier in order to cause a reduction in driving power at the lower end of the operating spectrum. These networks consist of components of inductive and capacitive reactance. At the higher end of the operating spectrum these parts tend to become "invisible" to the driver stage and amplifier.

Transistor Input and Output Capacitance

The published curves for bipolar RF power amplifiers show clearly that the input and output capacitances of the devices change with the operating frequency. What does this mean to us? It poses a knotty design problem if we attempt to include a negative-feedback network that is effective from, say, 1.8 to 29 MHz. These capacitances must not be ignored also when we use LC matching networks at the input and output of the amplifier. Specifically, we must absorb these capacitances into the matching network in order to avoid having them present unwanted XC, which can spoil the matching network. Some devices present very high input capacitances (1000 pF and greater). Generally, the output capacitances are much lower at a given frequency.

Input and output C also affects the performance of broadband RF
transformers, especially at the high-frequency end of the operating
range. The unwanted XC can cause prohibitive input and output SWR
at the base and collector of the transistor. Again, this phenomenon
is not encountered in large measure when we use power FETs as RF
power amplifiers.

Using Power FETs as RF Amplifiers

There are some distinct advantages associated with the use of power
FETs. As I mentioned in the foregoing text, FETs exhibit a relatively
constant input and output capacitance, irrespective of the operat-
ing frequency. Furthermore, power FETs are seldom prone to destruct-
ive thermal runaway.

Another plus feature associated with power FETs is the low driving
power they require in order to produce respectable output power.
Also, the high-order distortion products are lower in amplitude
than is the situation with bipolar devices, especially when the
FET is biased for linear operation. From a purely academic point
of view we may think of a power FET as an electrical matching trans-
former that produces gain. This philosophy has been expressed
to me a number of times by Siliconix applications engineer, Ed Oxner,
KB6QJ.

RF voltage is used to excite a power FET, whereas RF power is requir-
ed to excite a bipolar-transistor amplifier. The gate current of
a power FET can be measured in microamperes, and this makes it some-
what fragile with respect to device damage. Owing to this low gate
current, the input impedance of a power FET is very high (greater
than 1 megohm if nothing is attached to the gate). The output imped-
ance, on the other hand, is the same (low) as for a bipolar trans-
istor. Output impedance is a function of drain current and Vds (Z
= $Vds^2/2Po$, where Po is the device output power in watts).

All is not peaches and cream when we use power FETs. First, they
are especially prone to VHF self-oscillation. Some type of de-Qing
circuit or parasitic suppressor is generally required at the FET
gate terminal in order to discourage VHF oscillation. A 10- or 15-
ohm, 1/4-W noninductive resistor may be installed at the FET gate
to discourage these oscillations.

Power-FET gates are sensitive to excessive P-P voltage. Maximum
swing should be restricted to less that 30 V P-P. Too great a voltage
can perforate the gate insulation and cause an internal short circuit.
Protective built-in Zener diodes are found in some power FETs. They
are bridged between the gate and the source. Other power FETs have
Zener diodes (internal) between the drain and the source to provide
protection from excessive drain-source voltages. These Zener diodes
tend to limit the useful upper frequency of the FET. Some designers
use external gate-source and drain-source Zener diodes to preclude
device damage. These external gate Zener diodes are used in pairs,
and they are connected in parallel for opposite polarity to provide
clamping on both the positive and negative sine-wave peaks. It is

common practice to use 15-V, 400-mW Zener diodes for this purpose (see Fig 4-3).

Typically, the external Zener diode that is bridged from the drain to the source is rated at twice the Vdd. Thus, if the power FET is operated from a +24-V supply, the Zener diode is rated for 48 or 50 volts. Although it adds its junction capacitance to the drain circuit, it does not conduct until its barrier voltage is reached. This protects the FET from excessive drain-source voltage peaks, such as those encountered when the SWR is high or the transistor breaks into self-oscillation.
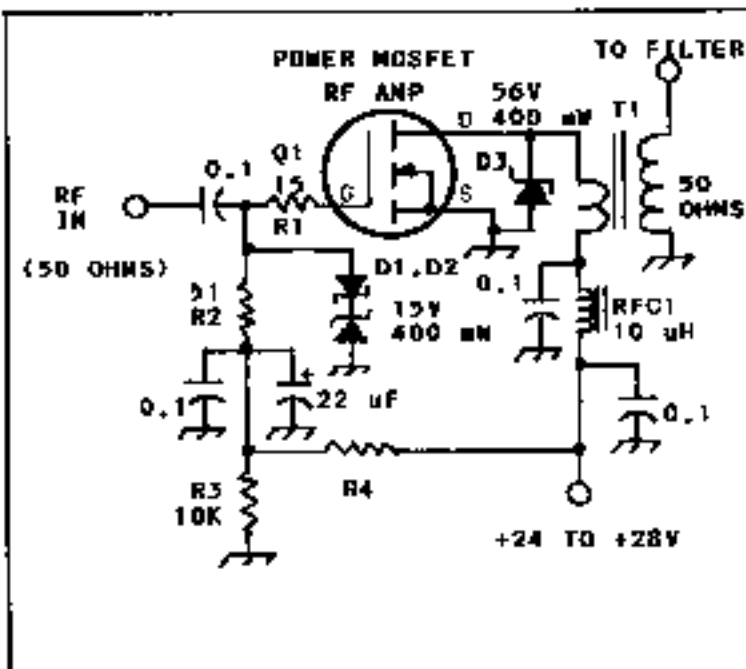
Fig 4-3 -- Example of a power FET RF amplifier. The input impedance is determined by the value chosen for R2. The lower the R2 value the greater the driving power required to provide the necessary RF gate voltage. D1 and D2 may be added to provide protection against excessive gate voltage. D3 can be used to protect Q1 from excessive drain-source voltage (see text). R4 is selected to obtain the desired forward gate voltage (bias). Bias is typically +1 to +6 V. T1 is a broadband matching transformer. R1 is a VHF parasitic suppressor.

Fig 4-3 shows a practical circuit for a power FET amplifier. The bias you apply to the Q1 gate determines the class of operation. No forward bias is used for class C amplifiers. Forward bias is needed for linear operation (class A, AB or B). Gate-bias regulation is not required for FETs, owing to the small gate current. A resistive divider (R3 and R4), along with the 22-uF capacitor, ensures a sufficiently stiff bias voltage. D1, D2 and D3 should not be used in well designed, stable circuits that exist in a controlled electrical environment. However, they are useful for inexperienced designers to provide go-no-go device protection. They may be removed once the circuit is optimized.

The input impedance for Q1 in Fig 4-3 can be elevated by using a resistor of higher value at R1. The higher the resistance the lower the driving power required to provide a given gate-voltage swing. However, the potential for amplifier instability increases as the effective input impedance is elevated. I do not recommend a resistance greater than 300 ohms for R1.

Many non-RF power FETs work well at MF and HF. The IRF511, which is a low-cost switching transistor, performs efficiently up to 10 MHz. A number of audio types of power FETs are suitable up to 30

MHz. Experimentation is necessary in order to learn how well a specified FET can perform at RF. Power FETs are available and meant expressly for RF amplification, but they are costly.

Power FETs are designed mainly for operation at +24 volts and higher. Although you may use them with a +12-V power supply, the efficiency will be poor. It becomes necessary, when using a low Vdd to stand considerable current in the transistor in order to obtain acceptable output power. This is not in keeping with the theme of QRP operation. Despite this limitation I would not rule out the use of power FETs at +12 V. If you are willing to accept reduced output power, you can operate the transistors at normal gate-bias levels.

When you select a power FET for RF service, study the device characteristics. The lower the rated RDS_{on} (drain-source resistance) the better the high-frequency performance or high-speed switching trait. Values less than 1 ohm are best, and RDS numbers below 0.25 ohm are better. RDS is the effective junction resistance when the FET is turned on. An MTM25N10, for example, has an RDS of only 0.07 ohm.

Transmitter Layout

Many of the common ills that we experimenters contend with can be associated with poor layout and construction practices. Long RF leads and ground conductors are inductive. These unwanted inductance invite VHF oscillations and also become part of the inductance in matching networks and transformer leads. Inductive reactances of this type change the impedance of circuits, especially at the upper end of the HF spectrum.

Circuit-board conductors need to be as short as practicable. The greater the conductor width the lower the stray inductance. Try to have your main PC board ground-foil area as large as you can make it. This will discourage the formation of RF "ground loops."

Double-sided PC board aids overall circuit stability in amplifiers. The copper on the component side of the board acts as a ground plane and forms numerous low-value capacitors in combination with the etched foils on the opposite side of the board. These capacitors bypass VHF currents and aid stability. The ground conductor on the etched side of the PC board should be connected to the ground plane on the component side of the board at several points. Small U-shaped strips of copper or brass may be soldered to the ground conductors around the edges of the board.

Double-sided PC board is not suitable for VFO circuits. The small capacitors that are formed by the PC conductors change value with changes in temperature. This causes VFO frequency drift.

Make an effort to lay out your transmitter PC board so that the stages are in a straight line. If, for example, the third stage is wrapped around so that it is close to the first stage, RF mutual coupling can cause one or both stages to self-oscillate.

PRACTICAL TRANSMITTERS

In this section we will discuss various practical transmitters from
the milliwatt to the 5-watt power-output level.   Parts-placement
guides and PC-board patterns are included for most of the projects.
Etched, drilled and plated boards for these projects are available
from FAR Circuits, 18N640 Field Court, Dundee, IL 60118.

If you are not experienced at building circuits I highly recommend
that you commence with the simple circuits in this chapter. The exper-
ience you gain will enable you to handle the more complex circuits
I shall describe. All of the circuits have been proof-built and made
to work as specified. Variations in performance can result   from
the inherent differences in transistor gain and departures   from
the parts values listed. Generally, the change in performance will
result in more or less transmitter output power.

Build a "Mighty Mite"

A proper starting point for beginners is a one-transistor, crystal-
controlled mW transmitter. Your signal from a 25- or 50-mW trans-
mitter is capable of being heard hundreds of miles away if you use
a good antenna. The 40-meter band is perhaps the best one to try
with this type of transmitter. Greater distances can be covered with
low power on 10- and 15-meters when band conditions are good, but
most one-transistor transmitters for those bands would put out very
little power, and they would produce chirpy signals. This is because
the crystal would have to operate on its 3rd overtone. Therefore,
we will consider the 40-meter "Mighty Mite" in Fig 4-4. The output
power is on the order of 30 mW. Some QRPers refer to this general
power level as QRP_p. I can assure you that there's no greater thrill
than to receive an RST 569 signal report from a station 300 miles
away when your transmitter output power is 30 mW! The "miles per
watt" game becomes a wonderful challenge that you should enjoy for
weeks on end. You may use your regular station receiver with this
transmitter.

Fig 4-4 -- Diagram of the Mighty Mite mW CW transmitter. C1 and C3 are mica compression trimmers. L1 (4.78 uH) has 30 turns of no. 26 enam. wire on an Amidon T50-2 toroid core. RFC1 is a miniature RF choke (Mouser Elect. part no. 43LR225). Y1 is a fundamental crystal, 30 pF load capacitance.
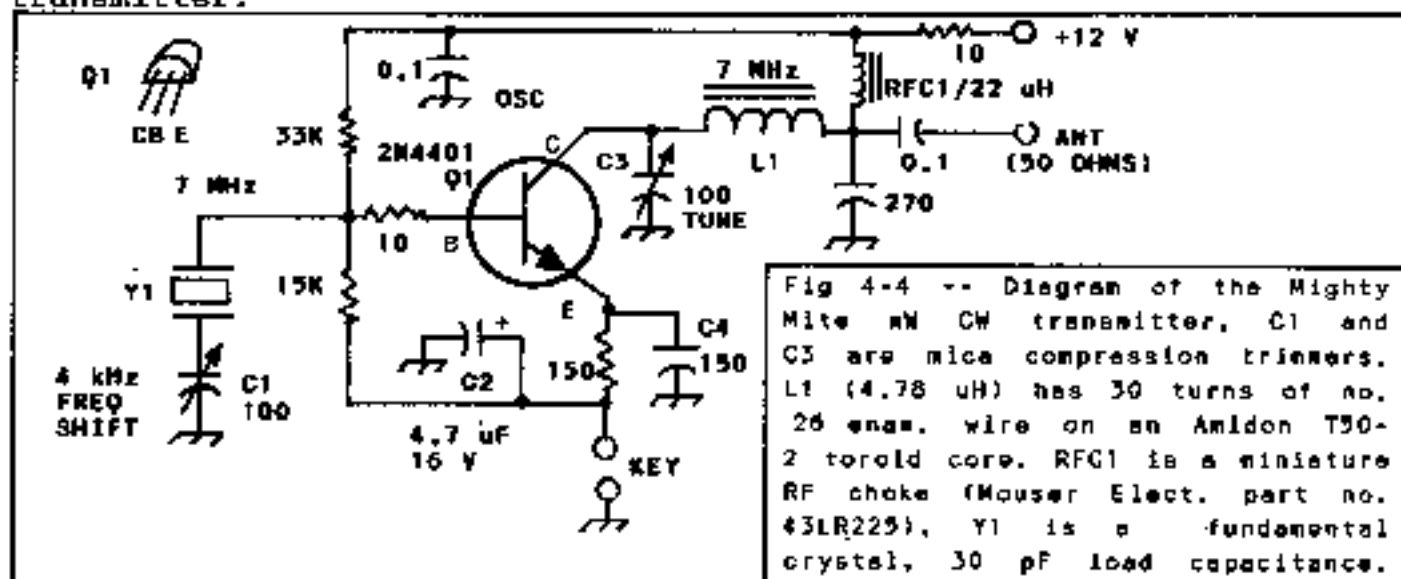
Simplicity is the keynote for the Fig 4-4 circuit. C1 enables you to shift the crystal frequency some 4 kHz. This component can be a trimmer or an air variable. The amount of shift at 40 meters will depend upon the nature of the crystal you use at Y1. Plated AT-cut crystals yield the greatest frequency change. The least suitable crystals are surplus units in FT-243 holders. You can eliminate C1 and ground the lower end of Y1 if you do not wish to shift the crystal frequency.

C4 controls the oscillator feedback. You may need to experiment with the value of this capacitor in order to make your particular crystal oscillate reliably. Values between 100 and 560 pF may be tried if the 150-pF value listed is not suitable for your crystal.

C2 in Fig 4-40 is used to shape the keyed waveform in order to minimize key clicks. It rounds off the corners of the waveform trailing edge. Increased capacitance at C2 softens the waveform and less capacitance hardens it.

C3 tunes the output pi network to resonance. It should be adjusted for maximum Q1 power output, consistent with a nonchirpy CW note. The output tuned circuit has a loaded Q of 4. It is important to ensure that the antenna you use has a low SWR. The pi network is designed to look into a 50-ohm load. A 10-ohm resistor is used at the base terminal of Q1 for the prevention of VHF parasitics.

Owing to the simplicity of the circuit in Fig 4-4, no circuit-board pattern or parts-placement guide is provided. You can tack this circuit together on a small breadboard. You may substitute other transistors for the 2N4401. Such devices as the 2N2222A may be used at Q1.

A more powerful one-transistor transmitter (1.5 W) is described on page 157 of **W1FB's Design Notebook**. It is capable of providing up to 5 watts of output power, depending upon the transistor used.

The "Challenger Two" Transmitter

Another practical example of a low-power CW transmitter is shown in Fig 4-5. The Challenger Two may be used as a stepping stone to greater construction experience. This little transmitter has an output power of 60 mW. A Pierce oscillator, Q1, drives a class C amplifier, Q2. This circuit can be tailored for use from 80 through 20 meters by changing the values of C1, C2, C8 and T1.

You may use 2N4400s, 2N4401s or 2N2222As for Q1 and Q2 without changing the circuit values. C1 and C2 are feedback capacitors. The values of these capacitors may be changed to suit the crystal you plug into the oscillator. The procedure we discussed for the Mighty Mite transmitter may be followed.

Tuned transformer T1 matches the collector of Q2 to a 50-ohm load. The number of turns for the secondary winding, L2, may be altered for matching loads other than 50 ohms.
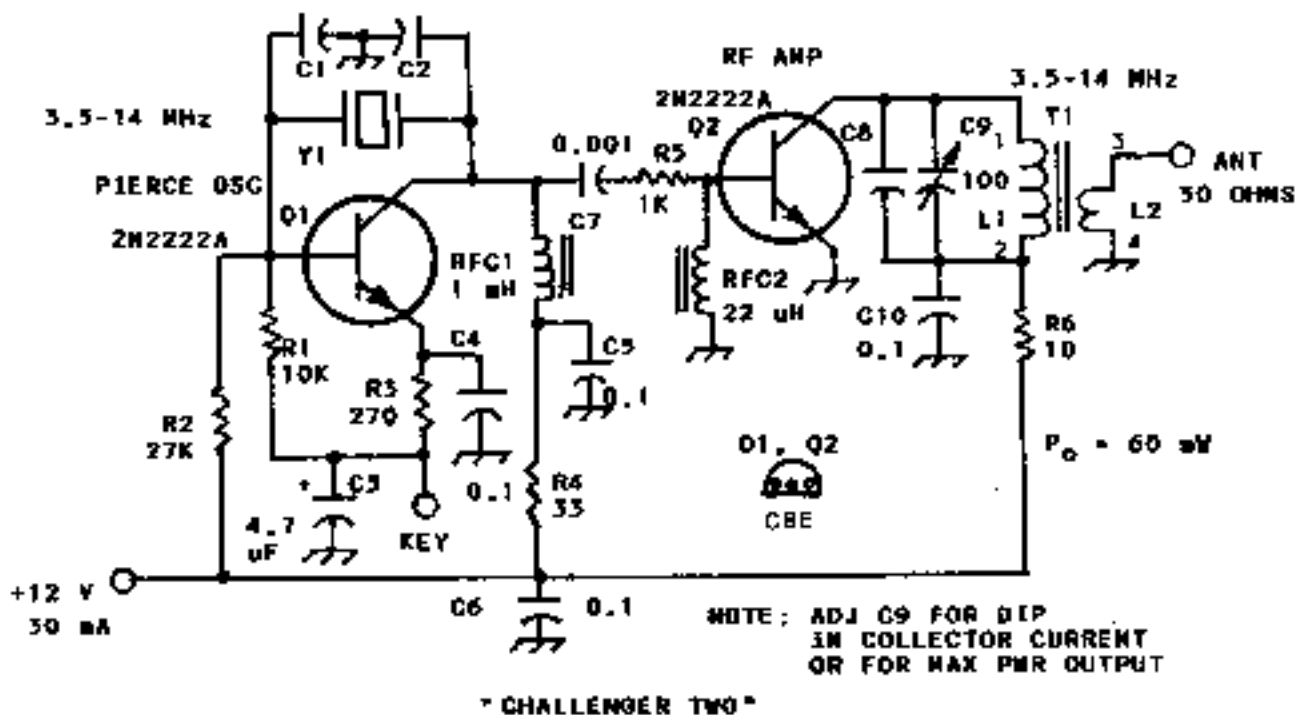
Fig 4-5 -- Schematic diagram of the Challenger Two QRP transmitter. Fixed-value capacitors are disc ceramic. Resistors are 1/4-W carbon film or composition. RFC1 and RFC2 are miniature Mouser Electronics or Oak Hills Research RF chokes. C1 and C2 are used only if they are needed to ensure Y1 oscillation. Values between 10 and 100 pF are typical if additional feedback control is necessary. The higher the operating frequency the lower the capacitance value. CB is 220 pF for 80 meters, 133 pF for 40 meters, 100 pF for 30 meters and 56 pF for 20 meters. L2 has 20% the turns used for L1. For 80 meters use 39 turns of no. 28 enam. wire (7.5 uH) on an Amidon T50-2 toroid core. For 40 meters use 25 turns of no. 26 enam. wire (3 uH) on a T50-2 toroid. L1 has 20 turns of no. 24 enam. wire (1.7 uH) on a T50-6 toroid. At 20 meters use 18 turns of no. 24 enam. wire on a T50-6 toroid (1.3 uH) for L1. C9 is a 100 pF ceramic trimmer (Mouser Electronics no. 24AA054). A mica trimmer may be substituted at C9.

The Challenger Two is keyed by grounding the junction of C3, R1 and R3. If an electronics keyer is used it should be set for positive keying.

The Fig 4-5 circuit may be used to excite solid-state amplifiers that provide greater output power. This circuit may be used as a basic building block for other transmitter projects. I was able to obtain 75 mW (0.075 W) of output from my model of the Challenger Two when operating it into a 50-ohm resistive load. The beta of the particular transistors installed at Q1 and Q2 will determine the net output power. The greater the transistor gain the higher the power-output level.

Maximum power output coincides with the dip in Q2 collector current when C9 is adjusted for circuit resonance. A 50- or 100-mA dc meter can be installed in the +12-V lead to Q2 for observing the collector current. A no. 49 pilot lamp (Mouser no. 35ND014), which is a 60-mA lamp, may be used in place of a meter for visual tuning of C9. Tune

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CHALLENGER TWO

W1FB

TO SCALE

CHALLENGER TWO

W1FB

X1.5 SCALE

GND   C3   GND

KEY
+12V
GND
ANT
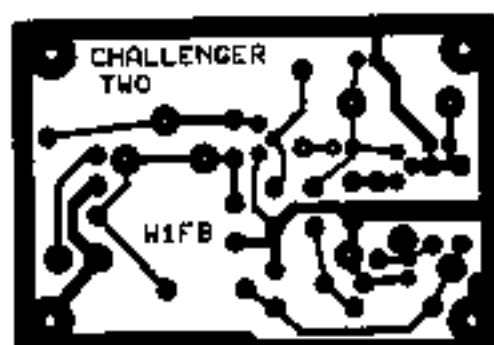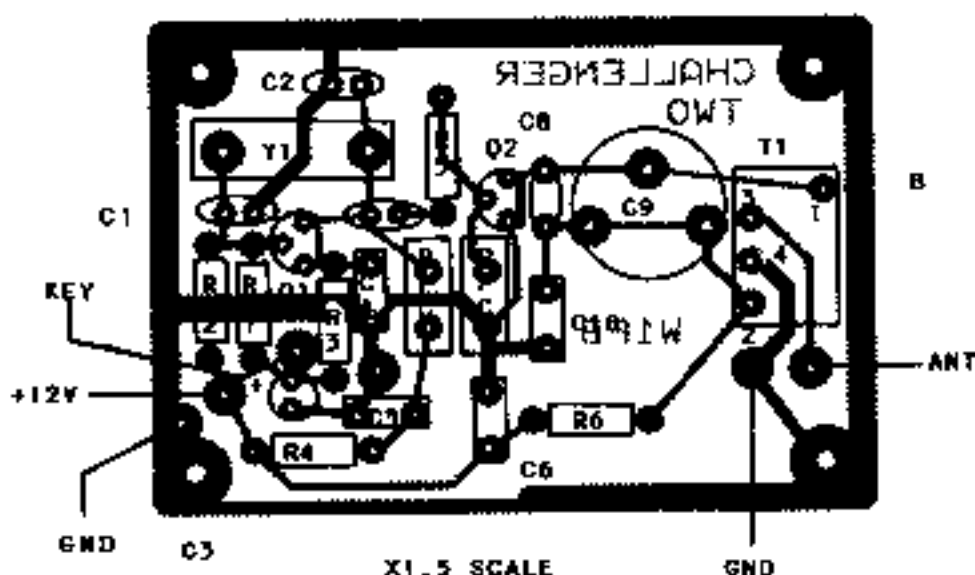
Fig 4-6 -- The PC template at A is shown from the etched side of the board.
Illustration B is an Xray view of the PC board as viewed from the component
side. Circuit boards for this project are available from FAR Circuits for
$3.75, plus $1.50 shipping.

for minimum lamp brilliance. The pilot lamp should be removed from
the circuit after C9 adjustment by shorting across it with a switch.
Fig 4-6 contains artwork for the PC board on which this project is
assembled.

The "Pebble Crusher Two"

Scarcely a well known "gravel grinder," the Pebble Crusher Two offers
greater output power than the previous two units, and it provides
an additional exercise in the "how to's" of transmitter building.
This little widget is designed to provide 0.5 W of RF power. All of
the parts are inexpensive and you should be able to construct this
unit in two hours or less.

The circuit in Fig 4-7 is actually an extension of the milliwatter
in Fig 4-4. Design data are provided for operation on 40 meters. You
can scale the circuit for operation on 160, 80, 30 and 20 meters by
using the XC and XL values of the tuned-circuit components shown in
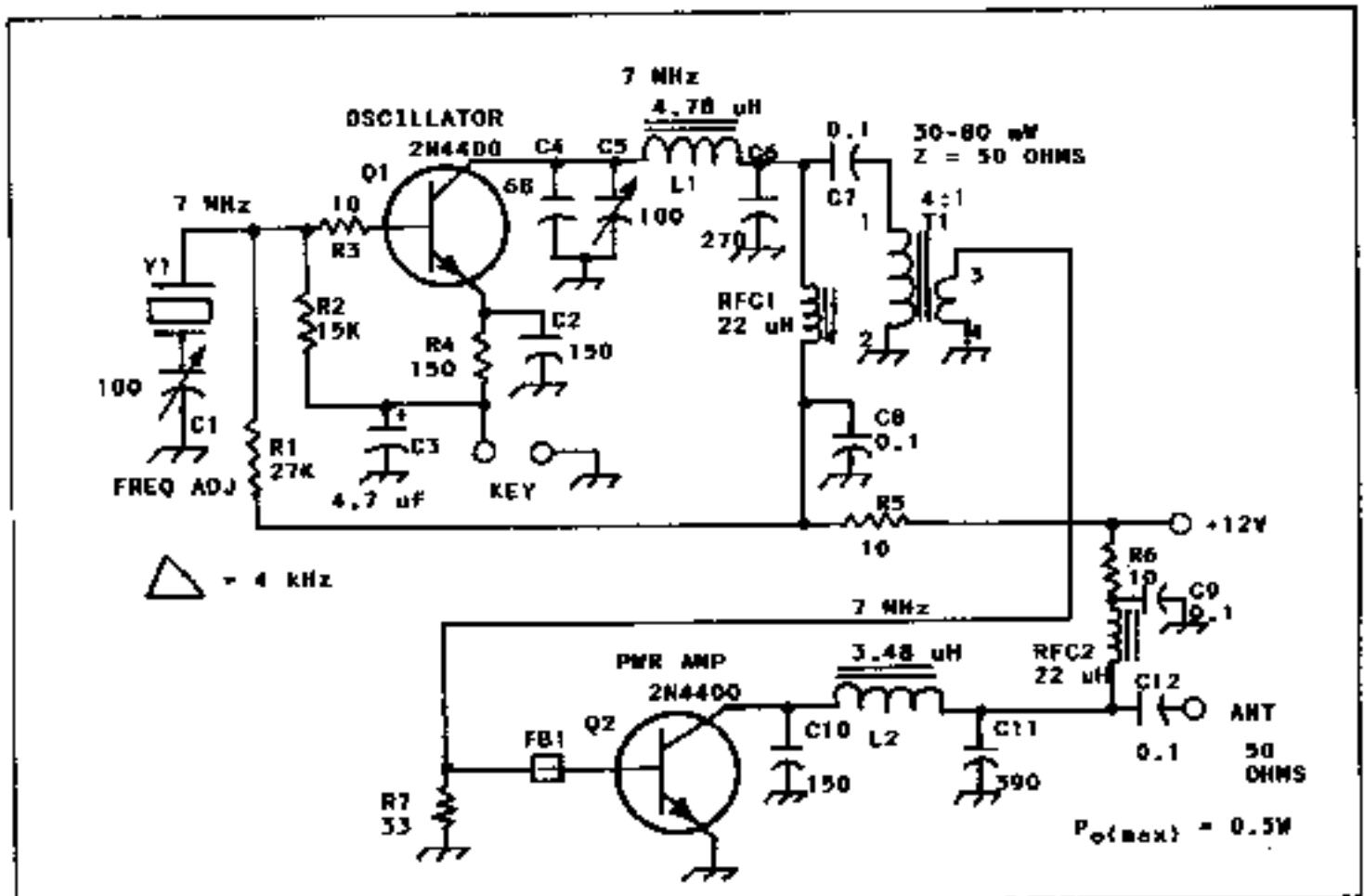
Fig 4-7 -- Schematic diagram of the 40-meter Pebble Crusher transmitter. Fixed value capacitors are disc ceramic except for C3, which is tantalum or electrolytic. C1 is a trimmer or air variable. C5 is a Mouser Electronics ceramic trimmer, no. 24AA034. Resistors are 1/4-W carbon film or composition. RFC1 and RFC2 are miniature encapsulated Mouser or Oak Hills Research RF chokes. L2 consists of 31 turns of no. 26 enam. wire on an Amidon T50-6 toroid. T1 is a broadband 4:1 impedance ratio toroidal transformer. Use 12 turns of no. 26 enam. wire (primary) on an Amidon FT-37-43 ferrite toroid. The secondary has 6 turns of no. 26 wire. L1 has 30 turns of no. 26 enam. wire on a T50-6 toroid. FB1 is a miniature Amidon ferrite bead, 850 mu.

Fig 4-7 to learn the C and L values for other amateur bands. The values of C4, C6, C10, C11, L1, L2 and Y1 must be changed for other bands of operation. No other changes are required.

An advantage in the Fig 4-7 circuit is that the driving power to Q2 is relatively clean with regard to harmonic currents. This means that Q2 is not supplied with unwanted energy that would otherwise be amplified with the desired signal energy. This enables Q2 to operate more efficiently and its harmonic output is reduced significantly. The Q1 collector tuned circuit is a low-pass filter. T1 matches the output of Q1 to the input of Q2 (50 ohms to 12 ohms). FB1 prevents VHF oscillation at Q2 and suppresses VHF harmonics from Q1.

C5 should be adjusted for a chirp-free CW note and for a transmitter

maximum current of 85 mA. This represents 0.5 W of output power. At one time during the testing period for this circuit I observed 1.25 watts of output power -- a bit much for a 2N44001 Q2 runs just slightly warm to the touch when it is delivering 0.5 W of output power. As a safety measure I added a small heat sink to Q2. It is a 3/8-inch piece of 1/4-inch ID brass tubing that is cemented to the Q2 case, as is discussed on page 106.

It should not be difficult to work some DX on 40 meters with the Pebble Crusher. A good antenna is needed, of course, along with normal band conditions for the time of night when you operate. My first QSO with this circuit was on 7015 kHz. I called a station in Hawaii and received an answer to my call. My signal report was RST 559. I was using a 40-meter inverted V antenna with the center at 50 feet. I was in MI when I worked the KH6.
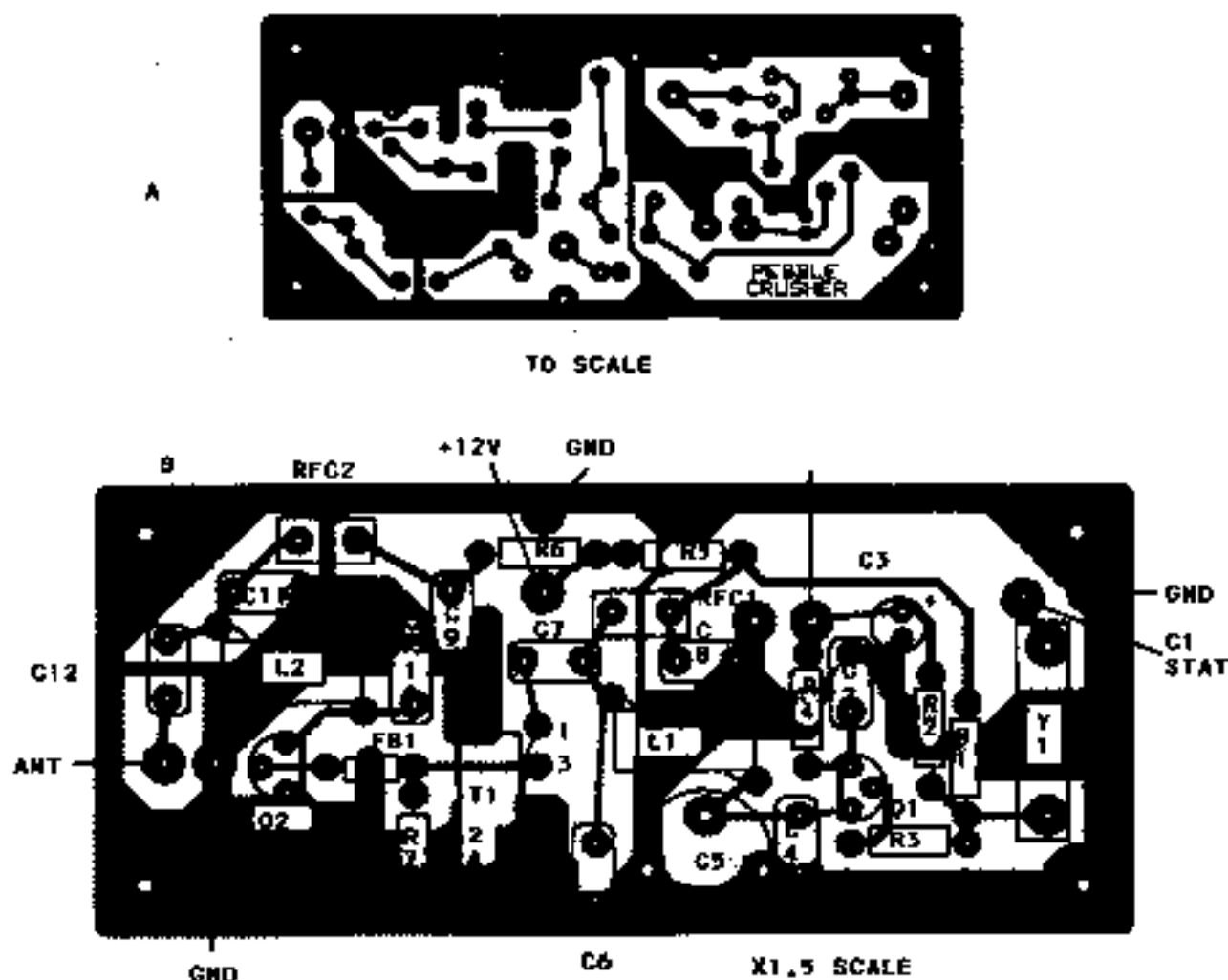
Fig 4-B -- A scale etching template, as seen from the etched side, is shown at A. Illustration B is a 1.5 times scale Xray view of the PC board as seen from the component side. PC boards for this project are avail. from FAR Circuits for $3.75, plus $1.50 shipping.

A VXO Transmitter/Exciter for 30 M

The CW transmitter in Fig 4-9 is called the "EX-1" and it may be altered for use on other HF bands from 160 to 20 meters by making a few changes in C and L values at key points in the circuit. This practical example is specified for use on 30 meters.

The Fig 4-9 transmitter may be used by itself, or it can serve as an exciter for the 5-W amplifier described later in this chapter. The circuit has some features that are not found in the previous three transmitters. A VXO with greater frequency range has been added. A power-output control, R16, has been added. Also, a PNP keying switch, Q3, is used to shape the output waveform and eliminate clicks. Both the leading and trailing edges of the keyed waveform are rounded off because of the capacitor and resistor values chosen for use at Q3. This keying switch may be used to trigger other circuits, such as a sidetone oscillator or break-in delay or QSK TR circuit.

Q1 is a Pierce oscillator that is keyed with amplifier Q2. The CW note has no discernible chirp. C2 and RFC1 form a reactance network that allows Y1 to be shifted up to 10 kHz at 10.1 MHz. C2 is a panel-mounted and isolated miniature air variable capacitor. C1 and C4 in Fig 4-9 are feedback capacitors. Their values are selected to ensure reliable oscillator starting and minimum chirp. Some experimentation with these values may be necessary, depending upon the traits of the crystal used at Y1.

Q2 is an untuned amplifier. Output for Q4 is taken from the C8-C9 capacitive divider. The ratio of these capacitors may be changed to increase or decrease the drive to Q4. R16 is a panel-mounted control that regulates the drive to Q4, and hence the transmitter output power. This is useful when the EX-1 is used as an exciter for an outboard RF power amplifier.

Q4 can handily produce up to 0.5 W of power output. Addition of a small heat sink at Q4 will permit output power up to 0.75 W. The PA stage, Q4, operates class C for maximum efficiency. A fixed-tuned pi network with a loaded Q of 4 is used to match the Q4 collector to a 50-ohm load. A scope may be used to measure the output power. A peak-peak voltage of 14 is equal to 0.5 W. If you use an RF probe and VTVM to read the RMS voltage at the transmitter output, the meter should indicate 4.9 V. Measurements are taken with a 51-ohm, 1/2-W resistor as a dummy load.

The Q1 VXO circuit will shift the crystal approximately 1 kHz above the marked frequency and up to 9 kHz below the marked frequency for plated fundamental crystals that are made for a 30 pF load capacitance. AT-cut crystals in HC-6/U style holders yield the greatest frequency swing.

Complete kits for this circuit (80, 40, 30 or 20 meters) are available from Oak Hills Research, 20879 Madison St., Big Rapids, MI 49307. Send $1 for a parts and kit catalog (Ph. 616-796-0920).

EX-1 EXCITER/TRANSMITTER
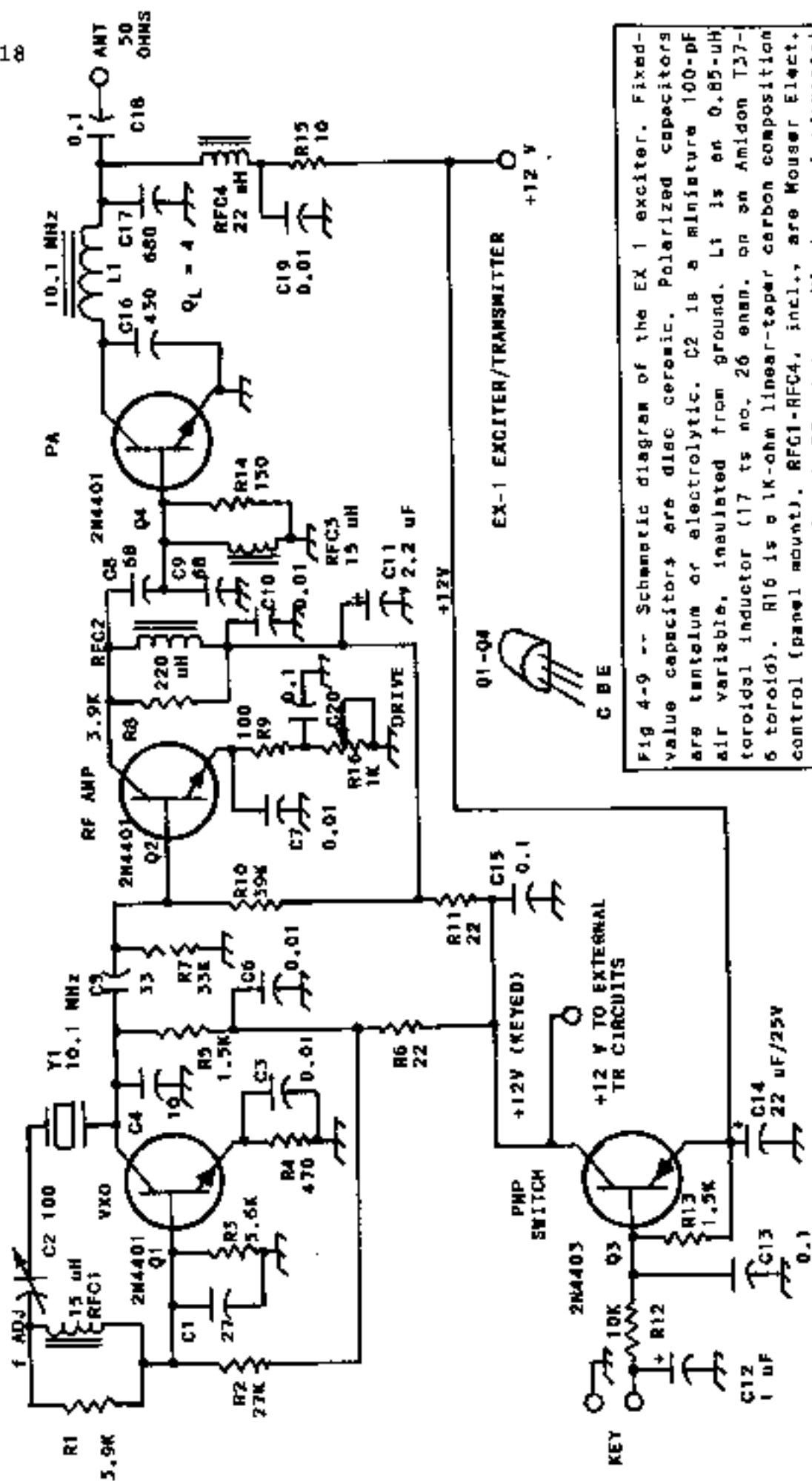
Fig 4-9 -- Schematic diagram of the EX 1 exciter. Fixed-value capacitors are disc ceramic. Polarized capacitors are tantalum or electrolytic. C2 is a miniature 100-pF air variable, insulated from ground. L1 is an 0.85-uH toroidal inductor (17 ts no. 26 enam. on an Amidon T37-6 toroid). R16 is a 1K-ohm linear-taper carbon composition control (panel mount). RFC1-RFC4, incl., are Mouser Elect. miniature, encapsulated RF chokes. Y1 is a fundamental crystal for the frequency of choice. 30 pF load capacitance. See text for kit availibility information.

Q1-Q4

C B E

A

TO SCALE

B
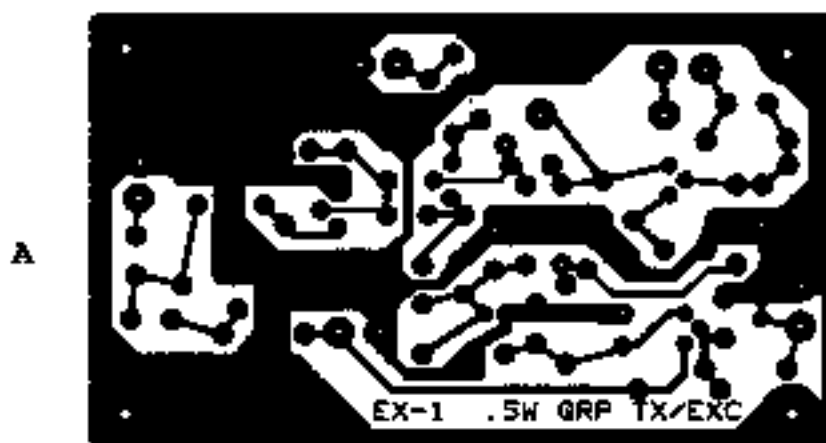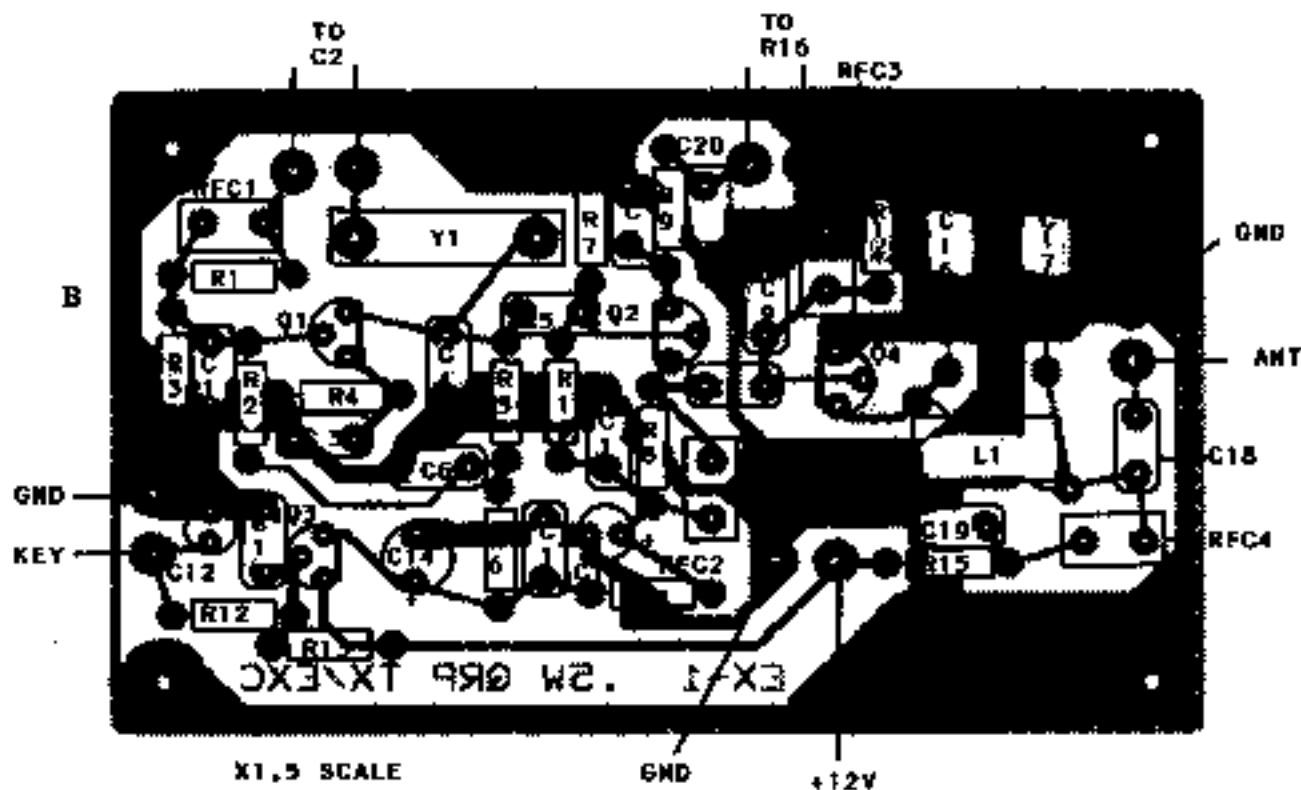
Fig 4-10 -- Illustration A is a scale template for the EX-1 exciter as viewed from the etched side of the PC board. An enlarged Xray view of the board (B) shows parts placement on the component side of the board. Circuit boards for this project are available from FAR Circuits for $4.25 plus $1.50 shipping.

The 10- and 15-Meter "Lil' Slugger"

Excellent DX results can be had on 10 and 15 meters with low power. A beam antenna provides better results than we can expect from a dipole or vertical ground plane, but DX is not difficult to work with any of these antennas.

Fig 4-11 contains a practical circuit for a three-stage, 1-W CW transmitter that you can build for 10 or 15 meters. Circuit boards and PC patterns are not available for this project. I did build and test it on a breadboard, using the familiar W7ZOI "ugly construction" method In fact, it went together easily on one of the FAR Circuits Universal Breadboards. Numerous QSOs were garnered with the Lil' Slugger while using my 160-meter horizontal loop with tuned feeders. I worked 14 countries on 15 meters the first day I fired up the transmitter.

An overtone oscillator is used at Q1 in Fig 4-11. The 33-pF emitter bypass capacitor is critical with regard to proper feedback for the oscillator. You may want to experiment with this value if you have trouble making your oscillator start in a reliable manner. The value listed was fine for all crystals that I used from my collection of plated units in HC-6/U holders. C1 and L1 are tuned to the desired overtone (3rd) of the crystal, Y1. Oscillation does not occur until this circuit is tuned to resonance. The oscillator is not keyed in order to prevent chirp. Backwave from the transmitter (key-up output at the operating frequency) is down 40 dB from peak carrier level. This is entirely adequate. The other station will not hear your backwave, unless he lives very near to you!

Link coupling from L1 is used to route RF energy to the broadband buffer/amplifier, Q2. C2 at Q2 bypasses unwanted VHF harmonic currents. It should have an XC that is 400 ohms or greater in order to prevent loss of the fundamental energy. A 10- or 15-pF capacitor is okay for 15 meters and a 5-pF capacitor is suitable for 10 meters. I suggest you eliminate C2 alltogether if you do not observe TVI from this transmitter.

The final amplifier, Q3, operates class C for maximum efficiency. Three transistors are specified, and each of them is fine for this part of the transmitter. The MPSU02 is a discontinued Motorola device, but they are available as surplus for as little as 50 cents apiece. This audio device has an fT in the VHF spectrum. It has high gain, which makes it easy to drive at HF. A 5-element low-pass harmonic filter is used at the transmitter output to suppress all harmonics by 40 dB or greater from peak carrier value.

A PNP bipolar dc switch, Q4, is used to key Q2 for CW service. The base capacitors and resistors were chosen to provide a click-free CW note. You can use almost any PNP transistor for the keying switch. A sidetone oscillator can be triggered from the collector of Q2. Likewise if you decide to use the QSK TR circuit described later in the chapter. The break-in delay TR module (also later in the chapter) can be actuated by connecting its input to the key jack in Fig 4-11, J1.
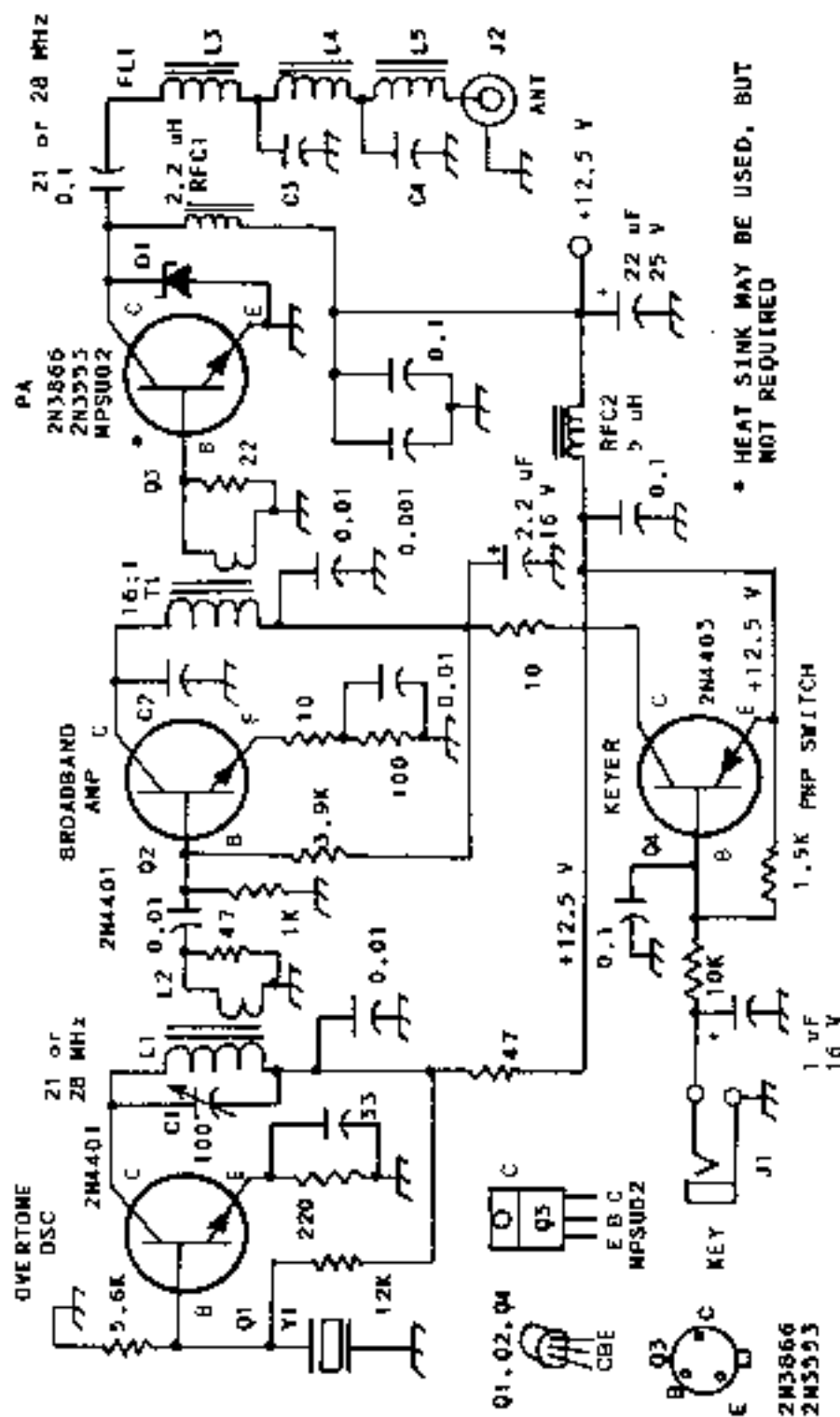
Fig 4-11 -- Schematic diagram of the Lil' Slugger 10- or 15-meter CW transmitter. Fixed-value capacitors are disc ceramic, except those with polarity marked, which are tantalum or electrolytic. Resistors are 1/4-W carbon film or composition. C1 is a ceramic or mica compression trimmer. This circuit appears also in W1FB's Design Notebook and is reproduced by courtesy of The ARRL, Inc. C3 and C4 - 180 pF for 15m and 130 pF for 10m. D1 is a 33-V, 400-mW Zener diode. L1 = 0.8 uH for 15m (14 ts no. 26 enam. on a T50-6 toroid). Use 2 ts of no. 26 wire for the link. L1 - 0.48 uH for 10m (10 ts no. 24 wire on a T50-6 core). Link = 2 ts. L3 and L5 - 0.20 uH for 15m (8 ts no. 24 wire on a T50-6 core). L3, L5 for 10m - 0.19 uH (7 ts no. 24 wire on a T50-6 toroid). L4 - 0.55 uH for 15m (12 ts no. 24 wire on a T50-6 core). L4 - 0.4 uH for 10m (10 ts no. 24 wire on a T50-6 core). T1 - 12 pri ts of no. 26 wire on an FT-50-43 toroid. Sec has 3 ts of no. 26 wire. RFC1 is a miniature Mouser Elect. unit. RFC2 has 10 ts no. 26 wire on a T24-43 toroid. The crystal, Y1, is a 3rd-overtone type, 30 pF load capacitance.

A Deluxe 5-W QRP Transmitter

The transmitters we have discussed earlier in this chapter are relatively simple. We shall next examine a practical CW transmitter that has certain convenience circuits which are not found in the previous designs. The Deluxe QRP Transmitter has a sidetone oscillator, a spotting circuit, frequency range and a 5-W power amplifier. It may be configured for 80, 40, 30 or 20 meters by changing the parts values in the LC networks. Some additional minor changes are necessary in order to alter the operating frequency.

The practical circuit for this transmitter is presented in Fig 4-12. Q1 is a JFET VXO that has provisions for two crystals. More crystals may be added by changing S1 to a multiposition unit. Stray capacitance must be kept at minimum in this part of the circuit so that the shift range of the crystals will not be compromised at the upper end of the crystal operating range. Select a switch that has a low capacitance between contacts and ground. The leads from the crystals to S1 must be as short as practicable. The Q1 VXO can "rubber" the frequency approximately 15 kHz per crystal at 14 MHz, 10 kHz at 10.1 MHz, 7 kHz at 7 MHz and 1.5 kHz at 3.5 MHz. C4, a panel-mounted air variable, is used to shift the crystal frequency. It should have a low minimum capacitance (10 pF or less) in order to avoid restricting the upper frequency range of the VXO. This VXO is as stable as a straight crystal oscillator, which makes the circuit ideal for portable work where wide temperature excursions are common.

VXO output energy is taken from the Q1 gate circuit via C5. Waveform purity is best at the gate, as opposed to the usual source takeoff point for Colpitts FET oscillators. By starting with a clean waveform we aid the overall spectral purity of the transmitter output energy.

Q2 is a tuned class A amplifier. Its linearity, along with the tuned collector tank, helps to insure waveform purity. The driver, Q3, is also a linear amplifier. A pi-network output circuit is used at Q3 to reduce harmonic currents and to provide a match between Q3 and the input of Q4.

The final amplifier, Q4, can be a Motorola MRF475, a NEC 2SC1909 or a 2SC2092. All three devices have the same lead arrangement and case style (TO-220). R14 is chosen to ensure amplifier stability. Use the highest resistance value that permits Q4 to operate stably. The lower the R14 value the greater the driving power needed to obtain 5 watts of output power from Q4. This base resistor permits some of the driving power to be dissipated within it.

Negative feedback is used at Q4 to provide gain leveling and to aid amplifier stability. A one-turn winding in T2 samples the output RF energy and feeds it back to the Q4 base via the inductive reactance formed by FB1 and FB2. R15 (with winding 1,2 of T2) sets the feedback value. Higher resistance at R15 results in less feedback and vise versa. The greater the negative feedback the lower the transmitter output power. T2 matches the Q4 collector to the input of the half-wave output filter (L3, L4), which is 50 ohms.

Effective Vcc decoupling is provided by means of C23, C24, C25, C26, RFC4 and RFC5. The rigorous decoupling is necessary to prevent Q4 RF currents from reaching the earlier stages of the transmitter via the +12-V PC-board bus. Without this precaution the transmitter would break into self-oscillation when it is keyed. W1 may be opened by way of a second switch section at S2. The +12-V feed to Q4 would be opened when S2 is in the SPOT mode. This will help to reduce the level of the spotting signal in the receiver.

Transistors Q5 and Q6 function as a sidetone oscillator to permit monitoring your sending during transmit. Output from this oscillator is routed to the audio-output amplifier in your receiver. R20 sets the audio level for comfortable listening. This circuit was developed by W7ZOI for use in some of his QRP transmitters. It is a simple and inexpensive multivibrator.

A PNP keying switch, Q7, is used to turn on and off the VXO and amplifiers Q2 and Q3. Gating diode D2 allows the current to flow to Q1 and Q2 during transmit, but it blocks dc to Q3 when S2 is in the SPOT mode. This minimizes the strength of the SPOT signal in the receiver, which would otherwise be a trifle overwhelming.

C30, C32, C33, R20 and R21 shape the keyed output waveform from Q4. Increasing the resistance of R21 will further soften the keyed waveform by rounding off and elongating the trailing edge of the waveform.

SWR protection for Q4 in Fig 4-12 is provided by Zener diode D3. It conducts when the dc or RF voltage reaches +33 volts. C19, which is in parallel with D3, reduces VHF harmonic currents that are present at the collector of Q4. C19 has minimal effect on the amplifier HF-band performance. The C19 value is different for each band of operation. It has an XC of 48 ohms or greater.

You can operate external convenience circuits with the Deluxe QRP Transmitter. For example, the break-in delay TR circuit (later in the chapter) can be actuated by connecting its input circuit to the junction of C32 and R21. The QSK TR circuit (later in the chapter) can be triggered by connecting its +12-V terminal to the collector of Q7. It is necessary to keep track of the current taken by exterior modules in order to avoid exceeding the maximum current rating of Q7. Larger transistors, such as the TIP33-C, may be substituted at Q7 to increase the keyed-current capability of the PNP switch.

If you wish to include a DRIVE control in the Fig 4-12 circuit, you can lift the grounded end of R5 and insert a 1K-ohm control between the bottom end of R5 and ground. Bypass the junction of R5 and the control with a 0.1-uF ceramic capacitor. Output power from Q3 may be increased by using a lower resistance value at R2. Do not use less than 22 ohms. A lower value for R12 will increase the collector current for Q3. This may require the addition of a push-on crown type of heat sink for Q3. Do not increase the RF output from Q3 beyond the point where no further output can be obtained from Q4 (saturation). Do not increase the capacitance of C5 in an effort to boost the transmitter output power. This will cause chirp and may prevent Q1 oscillation.

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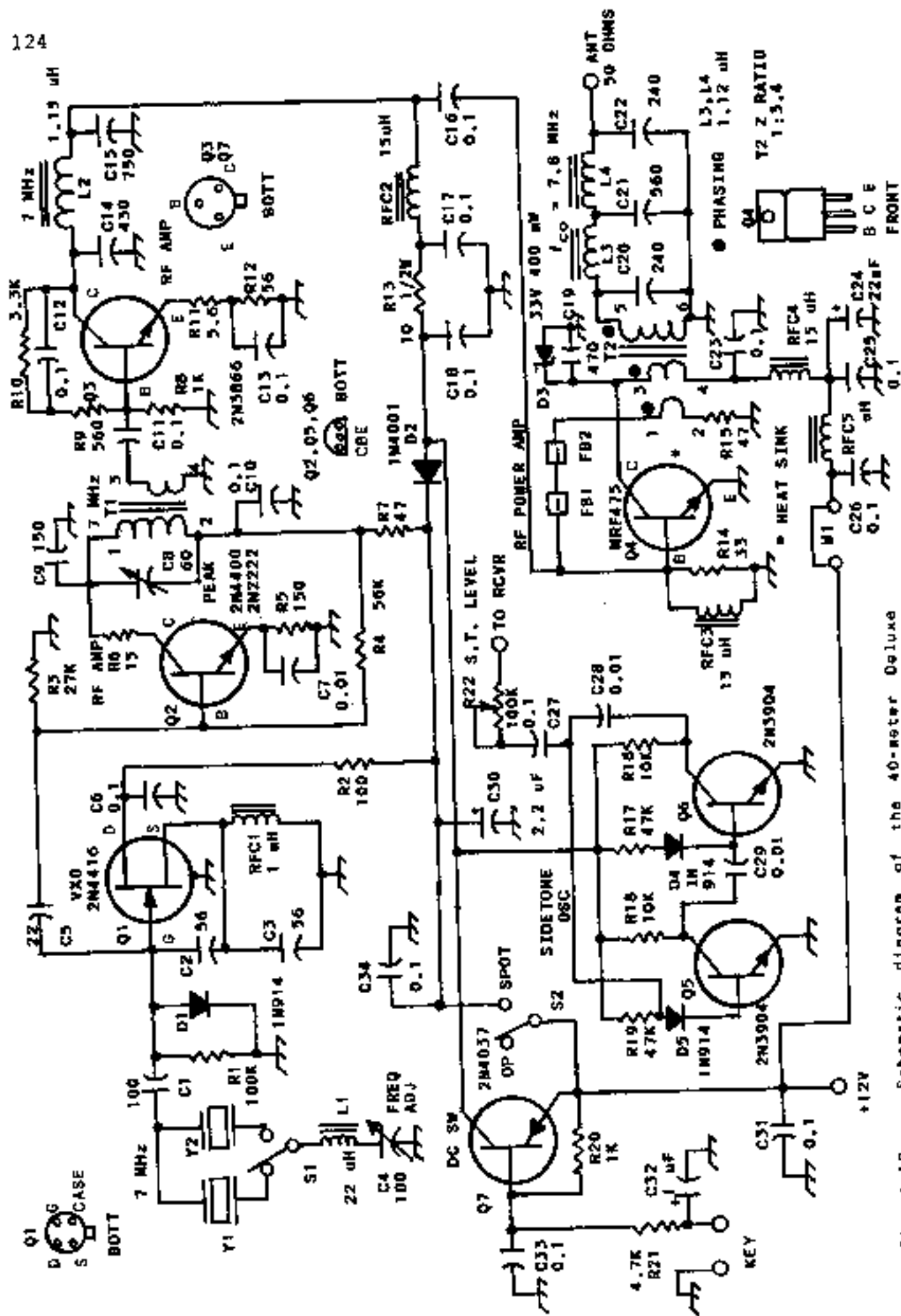
Fig 4-12 -- Schematic diagram of the 40-meter Deluxe QRP Transmitter. Fixed-value capacitors are disc ceramic. Polarized capacitors are tantalum or electrolytic. Resistors are 1/4-W carbon film or composition unless otherwise

Fig 4-12, continued.

C4 -- Miniature 100-pF air variable, panel mounted.
C14, C15, C19, C20, C21, C22 -- Silver mica or polystyrene.
C28, C29 -- Polystyrene or mylar high-Q capacitor.
D1 -- Silicon high-speed switching diode, type 1N914.
D2 -- 1-A, 50-PRV rectifier diode.
D3 -- 33- or 36-V, 400-mW Zener diode.
L1 -- Miniature 22-uH RF choke (Mouser or Oak Hills Research).
L2 -- Toroidal inductor, 1.15 uH. Use 19 turns of no. 26 enam. wire on an Amidon
T37-6 (yellow) toroid.
L3, L4 -- Toroidal inductor, 1.12 uH. Use 17 turns of no. 24 enam. wire on a T50-
6 core.
RFC1 -- Miniature Mouser or Oak Hills encapsulated RF choke.
RFC2-RFC5, incl. -- 15-uH toroidal RF choke. Use 15 turns of no. 28 enam. wire on
an Amidon FT-24-43 ferrite toroid (850 au).
S1, S2 -- Miniature SPDT toggle or wafer switch. S2 may be a momentary type for
added convenience.
T1 -- Tuned toroidal transformer. Primary is 3 uH, with 27 turns of no. 28 enam.
wire on an Amidon T37-2 (red core) toroid. Secondary has 6 turns of no. 26 wire.
T2 -- Broadband transformer, 1:1.8 turns ratio. Core is an Amidon 850-mu ferrite
balun type, no. BN-43-202. Winding 1,2 is 1 turn of no. 26 enam. wire. Winding 3,4
has 3 turns of no. 26 enam. wire. Winding 5,6 uses 6 turns of no. 26 enam. wire.
Install on PC board (vertically) with lead end of core toward board. Affix to board
with a few drops of epoxy cement.
W1 -- Jumper wire.
Y1, Y2 -- Fundamental plated crystal. HC-6/U holder, 30 pF load capacitance.
FB1, FB1 -- Miniature Amidon ferrite bead, FB-43-301.

Construction Thoughts

Fig 4-13 contains a scale etching template and a parts placement guide
for the Deluxe QRP Transmitter. The assembled PC board should be mount-
ed in its case with the L1, Y1 and Y2 end of the board near the front
panel. This will help to keep the S1 leads short and direct. Metal
standoff posts are used to attach the PC board to the chassis or bottom
surface of the box. This aids in providing effective grounding, which
minimizes the formation of RF ground loops.

L2, L3, L4, T1 and T2 are mounted vertically rather than flat on the
PC board. Secure them to the PC board with a couple of drops of epoxy
glue after you have finished checking the circuit for proper perform-
ance. This will eliminate stress on the wire leads if there is vibrat-
ion. Such stress, over a period of time, can cause the coil and trans-
former leads to break.

Operation of this transmitter on 80, 30 or 20 meters may be achieved
by making suitable changes in the tuned circuits. C3 is 56 pF for 3.5
MHz, 22 pF for 10.1 MHz and 15 pF for 14 MHz. L1 is 100 uH for 3.5
MHz, 22 uH for 10.1 MHz and 15 uH for 14 MHz. Calculate the reactances
of C2, C3, C9, C14, C15, C19, C20, C21 and C22 for converting these
parts to other operating frequencies. Likewise for T1, L2, L3 and L4.
No other changes are necessary.

A

DELUXE 5 WATT
QRP TRANSMITTER

SHOWN TO SCALE

X1.5 SCALE

U-SHAPE
HEAT SINK

B

KEY   GND   +12V S2 ARM   ST OUT   GND   ANT
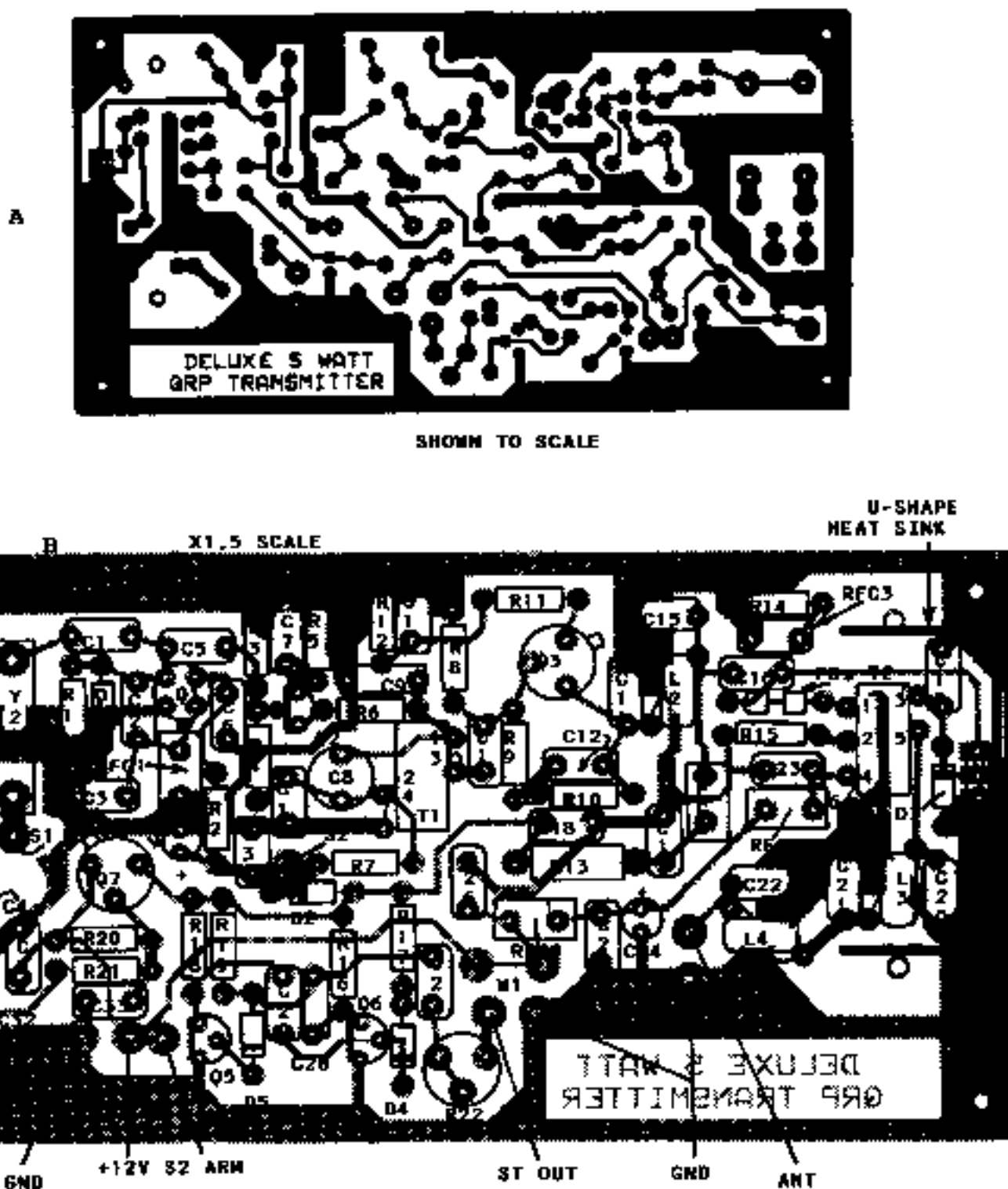
Fig 4-13 -- A scale etching pattern, as viewed from the etched side, is shown at A. An expanded Xray view of the board is seen at B, as viewed from the component side of the board. PC boards for the Deluxe QRP Transmitter are avail. from FAR Circuits for $7 plus $1.50 shipping.

Q4 in Fig 4-12 requires a heat sink. The transistor is attached to a U-shaped home-made aluminum or brass heat sink. This assembly is mounted vertically on the PC board as indicated at B in Fig 4-13 at the right side of the drawing. The heat sink is 1-3/8 inches wide and is 1-3/4 inches high. The "wings" of the U channel are 1/2 inch wide. No. 6 spade bolts are used to affix the heat sink to the PC board. The heat sink is common to the Q4 collector, which makes it mandatory to keep it insulated from ground. No. 16 gauge or thicker stock should be used for the heat sink.

RF POWER AMPLIFIERS

Broadband class C RF power amplifiers seem to be the choice of most QRPers. This is because this type of amplifier is easy to build and make operate. Class-C operation yields the best overall efficiency for bipolar transistors. As we learned earlier, we must trade gain for bandwidth when the broadband mode is chosen. Typical class-C amplifier efficiency for broadband circuits is on the order of 50-60 percent. Most class AB linear amplifiers fall into the same efficiency class.

Class-C efficiencies are as great as 75 percent for narrow-band RF amplifiers in which bipolar transistors are used. Efficiencies approaching 90 percent have been reported to me by Ed Oxner, KB6QJ, who is an applications engineer at Siliconix, Inc., respective to class C power FET amplifiers. I have observed 80 percent efficiency with a 40-watt, 40-meter power FET narrow-band class C amplifier I developed.

Although narrow-band RF amplifiers operate in an efficient manner, they are tricky to use. This is because impedance matching becomes difficult, stability is not easy to achieve, and transistor destruction happens quickly if self-oscillation occurs. Those who attempt to apply vacuum-tube design principles when building solid-state amplifiers learn quickly that transistor destruction can happen instantly!

Push-pull transistor amplifiers are preferred at power levels above five or ten watts. Single-ended amplifiers may be used for moderately high power amounts, but the collector impedance becomes so low that matching to the load becomes an insurmountable task unless high-voltage, low-current amplifiers are employed. For example, a single-ended 50-W amplifier that operates from a 12-V power source has a collector impedance of 1.44 ohms! Not only would matching the amplifier to a 50-ohm load be next to impossible, the inductance of the PC-board conductors in the collector circuit would be significant, and this would tend to seriously compromise the matching network.

Push-pull operation results in higher collector impedances. There is another advantage associated with push-pull amplifiers: They cancel even harmonics, and this makes harmonic filtering a less difficult task.

Finally, it is much easier to band switch a broadband amplifier than it is to change bands in a narrow-band circuit. Typically, we need

only switch the harmonic filters in a broadband RF amplifier. This
becomes an easy assignment because the filters are designed for a
50-ohm input and output impedance. A mechanical switch or relay system
can be used for switching low-impedance circuits, and small coaxial
cable is suitable for the leads between the filters and the switch.

Class AB versus Class C Operation

Class C amplifiers are ideal for CW and FM transmissions. However,
we need to employ linear amplifiers for the amplification of AM and
SSB signal energy. Most solid-state linear amplifiers are biased for
AB operation, at least for power levels above approximately 1 watt.
Class A amplifiers are generally used in low-level linear-amplifier
circuits, since the resting collector current is low and heat sinks
are not required.

Biasing for linear service required the application of positive voltage
to the transistor base or gate, depending upon the type of device
used. This is called "forward bias." The bias voltage causes the trans-
istor to draw standing or resting collector or drain current when
no signal is present. This affects the overall efficiency of the trans-
mitter (reduces it) and causes the transistors to operate warmer than
for class C service. A class C amplifier has the opportunity to cool
down during key-up periods, whereas this is not true of linear amp-
lifiers. Substantially larger heat sinks are required for linear power
amplifiers, owing to the increased heat buildup. Fig 4-14 illustrates
a simple method for applying forward bias to a bipolar transistor
linear amplifier. D1 has a threshold voltage of +0.7, which causes
the Q1 collector current to flow when no signal is present. D1 provides
a stable bias source by functioning also as a reference or regulator
diode. The 22-uF capacitor stores dc voltage and aids bias stabiliz-
ation. Many commercial amplifiers employ bias circuits that are more
sophisticated. They use regulator ICs and diodes. Also, voltages great-
er than +0.7 are often used to improve the amplifier linearity. Much
depends upon the particular transistor used and the design objective.
The method shown in Fig 4-14 is adequate for amateur service at the
QRP power level.
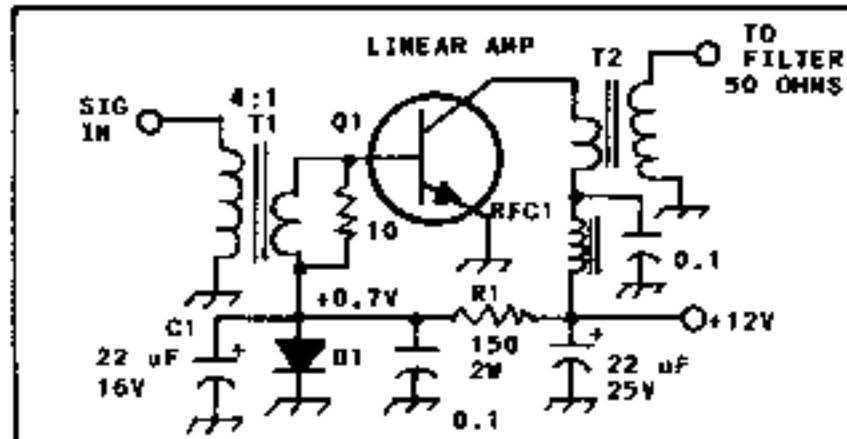
Fig 4-14 -- Example of a single-ended class AB linear RF amplifier. D1 acts as a regulator and establishes a +0.7-V forward bias for Q1. C1 helps stabilize the bias voltage. R1 is chosen to allow D1 to draw 50 mA. D1 is a 1N4001 rectifier diode.

The 10-ohm base resistor in Fig 4-14 is for de-Qing the input circuit. Its value may differ from that specified, as discussed earlier in this chapter. T2 is wound to ensure an impedance match between the Q1 collector and a 50-ohm load. RFC1 can be a 10-uH choke that is capable of accommodating the current taken by Q1. This amplifier may be equipped with an MRF475 or 2SC2092 for use in a DSSC or SSB exciter. A DSSC generator is described later in this chapter.

A 5-Watt Class C QRP Amplifier

Fig 4-15 contains a schematic diagram for a class C amplifier that may be used from 1.8 to 30 MHz as a single-band device. A separate harmonic filter is required for each band of operation. You can use this circuit for coverage from 160 through 10 meters by band-switching the harmonic filters. If you elect to do this it will be necessary to assemble the filters on a separate PC board. This board should be located close to the amplifier board in order to minimize the length of the lead from T2 to the filter board. A two pole, multi-position wafer switch (located near the filter PC board) may be used for switching the filters. This switch should be a two-wafer type to permit isolating the filter input circuit from its output circuit. This will help to minimize leakage that could otherwise compromise the effectiveness of the filters, especially at the high end of the amplifier operating range. A grounded metal shield plate, if installed between the switch wafers, will further aid this cause. RG-174 mini 50-ohm coaxial cable is suitable for connecting the filters to the band switch. The shield braid of each lead must be grounded at each end of the cable.

Power input for the Fig 4-15 amplifier is 0.5 W maximum. Under these conditions Q1 is saturated. The effective driving power that reaches the base of Q1 is on the order of 0.25 W, owing to losses in T1 and the dissipation of RF power in the 10-ohm base resistor. Amplifier gain is 13 dB when the base is driven by 0.25 W. This circuit may be modified for linear service by adopting the technique shown in Fig 4-14.

The component numbering does not commence at 1 because this module was used as part of a complete transmitter I once developed. C26 should be eliminated or set for 10-meter operation if you choose to band switch this amplifier.

T1 is a broadband toroidal transformer. C20 is a reactance-compensating capacitor for the upper end of the operating range. C21, C22 and C22 provide effective bypassing from LF through the HF spectrum.

Negative feedback is used to stabilize the amplifier and ensure that the gain is essentially constant from 1.8 to 30 MHz. Output energy is sampled by a one-turn winding, leads 1 and 2, in T2. FB1 and FB2 act as inductive reactances to lessen the feedback at the upper end of the HF spectrum. The value of R17 determines the feedback power.

A five-element harmonic filter, FL1, ensures that all spurious output energy is 40 dB or greater below peak output power. It has a low-pass response.

T2 matches the 14.4-ohm collector impedance to 50 ohms for matching the amplifier to the harmonic filter. C25 and C25 are silver mica or polystyrene capacitors. Disc ceramic capacitors are generally unsuitable, owing to their wide capacitance variations with respect to the marked values.
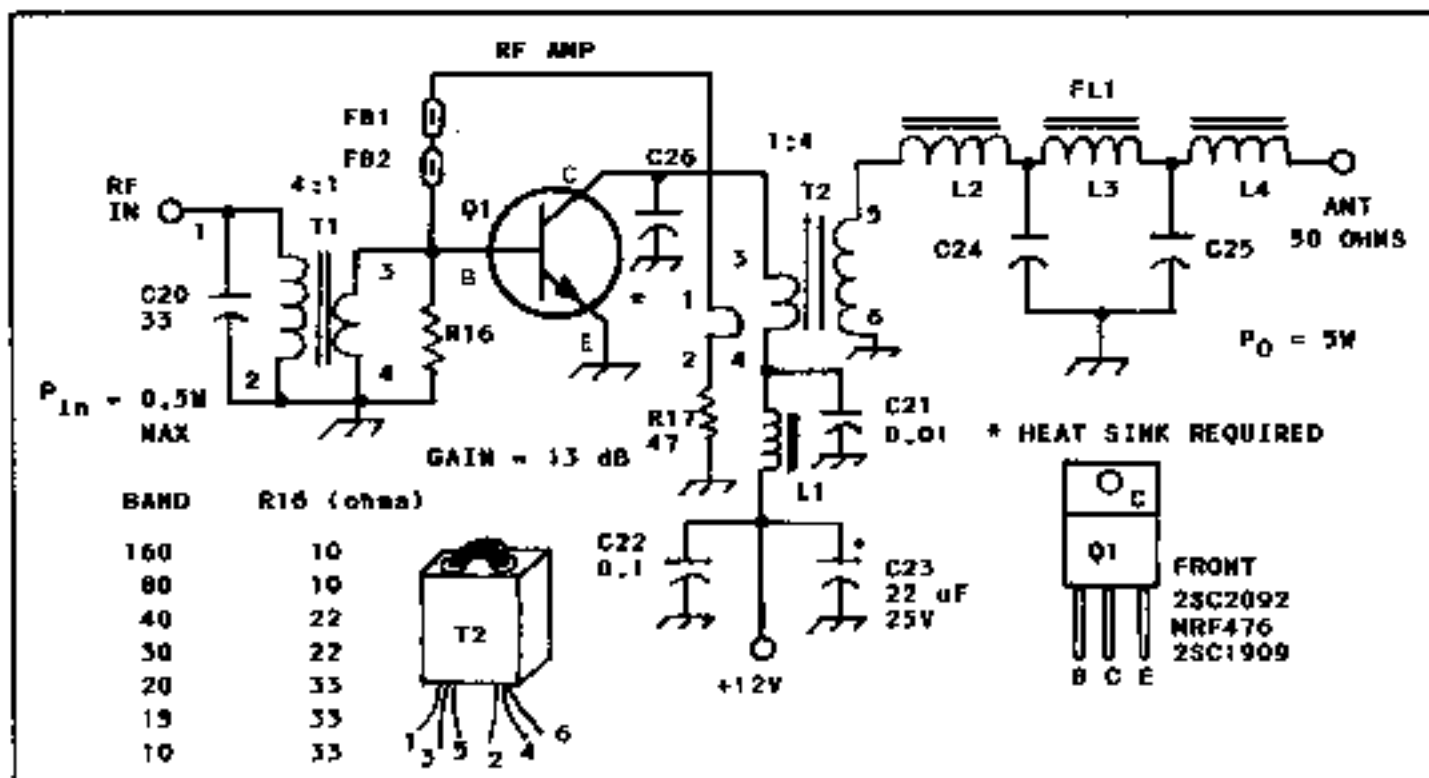
Fig 4-15 -- Schematic diagram of the PA5-HF 5-W class C amplifier. Fixed-value capacitors other than C23, C24 and C25 are disc ceramic. C24 and C25 are silver mica or polystyrene (see text). C23 is tantalum or electrolytic. Resistors are 1/4- or 1/2-W carbon film or composition. FB1 and FB2 are miniature 850-mu Amidon ferrite beads. L1 is 15 uH (10 turns no. 24 enam. wire on an Amidon FT-37-43 ferrite toroid). T1 has 12 primary turns of no. 26 enam. wire on an FT-37-43 ferrite toroid. Use 6 secondary turns of no. 26 enam. wire. T2 has one turn for winding 1,2. Winding 3,4 has 3 turns and winding 5,6 has 6 turns. All windings use no. 26 enam. wire looped through an Amidon BLN-43-202 ferrite balun core.

Table 4-1 contains component data for the 5-W amplifier. Nonstandard capacitor values may be obtained by paralleling standard values.

TABLE 4-1

BAND	C24, C25 (pF)	C26 (pF)	L3	L2,L4
160m	1800	1500	5.5 uH. 33 ts no. 26 enam. on a T50-2 toroid.	2.70 uH. 23 ts. 26 enam. on a T50-2 toroid.
80m	910	910	2.69 uH. 24 ts. no. 26 enam. on T50-2.	1.5 uH. 17 ts. no. 26 on T50-2.
40m	560	470	1.7 uH. 23 ts. no. 26 on a T37-6 core.	1.06 uH. 19 ts. no. 26 on T37-6.

TABLE 4-1, continued.

BAND	C24, C25 (pF)	C26(pF)	L3	L2,L4
30m	423	330	0.88 uH. 17 ts. no. 26 on a T37-6 core.	1.5 uH. 22 ts. no. 28 on a T37-6 core.
20m	286	220	0.41 uH. 11 ts. no. 26 enam. on T37-6.	0.86 uH. 17 ts. no. 26 on T37-6.
17m	218	180	0.3 uH. 10 ts. no. 24 on a T37-6 core.	0.66 uH. 15 ts. no. 26 on a T37-6 core.
15m	198	150	0.41 uH. 11 ts. no. 24 enam. on T37-6.	0.71 uH. 15 ts. no. 26 on a T37-6 core.
12M	167	130	0.35 uH. 11 ts. no. 24 enam. on T37-6.	0.6 uH. 14 ts. no. 26 on a T37-6 core.
10m	124	100	0.26 uH. 9 ts. no. 24 enam. on T37-6.	0.44 uH. 12 ts. no. 24 on a T37-6 core.

Component values for FL1 in Fig 4-15. Nonstandard capacitor values may be obtained by paralleling standard values. A 10 percent variation from the listed values is acceptable.

A heat sink is necessary for Q1. It must be insulated from ground because the tab on Q1 is common to the collector. Alternatively, you can insulate Q1 from the sink by means of a TO-220 mica insulator and a 4-40 nylon screw and nut. With either arrangement be sure to use a thin layer of heat-sink compound between the transistor and the heat sink. Your home-made heat sink can be made from 16-gauge or heavier aluminum or brass sheeting. It is U-shaped and measures 1-3/4 X 2-1/4 inches. The lips of the U are 1 inch high.

Although the component is not shown in Fig 4-15, you can add a 33- or 36-V, 400-mW or 1-W Zener diode for SWR protection of Q1. The diode should be bridged between the collector foil and ground with the shortest span practicable. The diode may cause some loss of output power at the upper end of the HF range (15, 12 and 10 meters), depending upon the internal capacitance and resistance of the diode used. In an ideal situation the Zener diode is not used on any band. It is not needed if you make certain that the SWR is no greater than 1.5:1. If you're in doubt, use a Transmatch between the amplifier and the feed line to ensure a 1:1 SWR.

Use RG-174 coaxial line to connect the exciter to the PA5-HF amplifier. Likewise between the amplifier and the antenna relay. You may use operating voltages from 11 to 13.6 with this circuit. Greater power output occurs at the higher supply voltages. Regulated voltage is necessary if you do not operate from a car battery.

A

ETCHED SIDE TO SCALE

B

INSULATED
HEAT SINK
MTG HOLE

HEAT SINK
AREA

GND

ANT

INPUT   GND

PCB MTG
HOLE

C20

1 4

T1

3 2

R16

C23

GND
+12V

RFC5

C21

3   2
5   T2   4
1   6
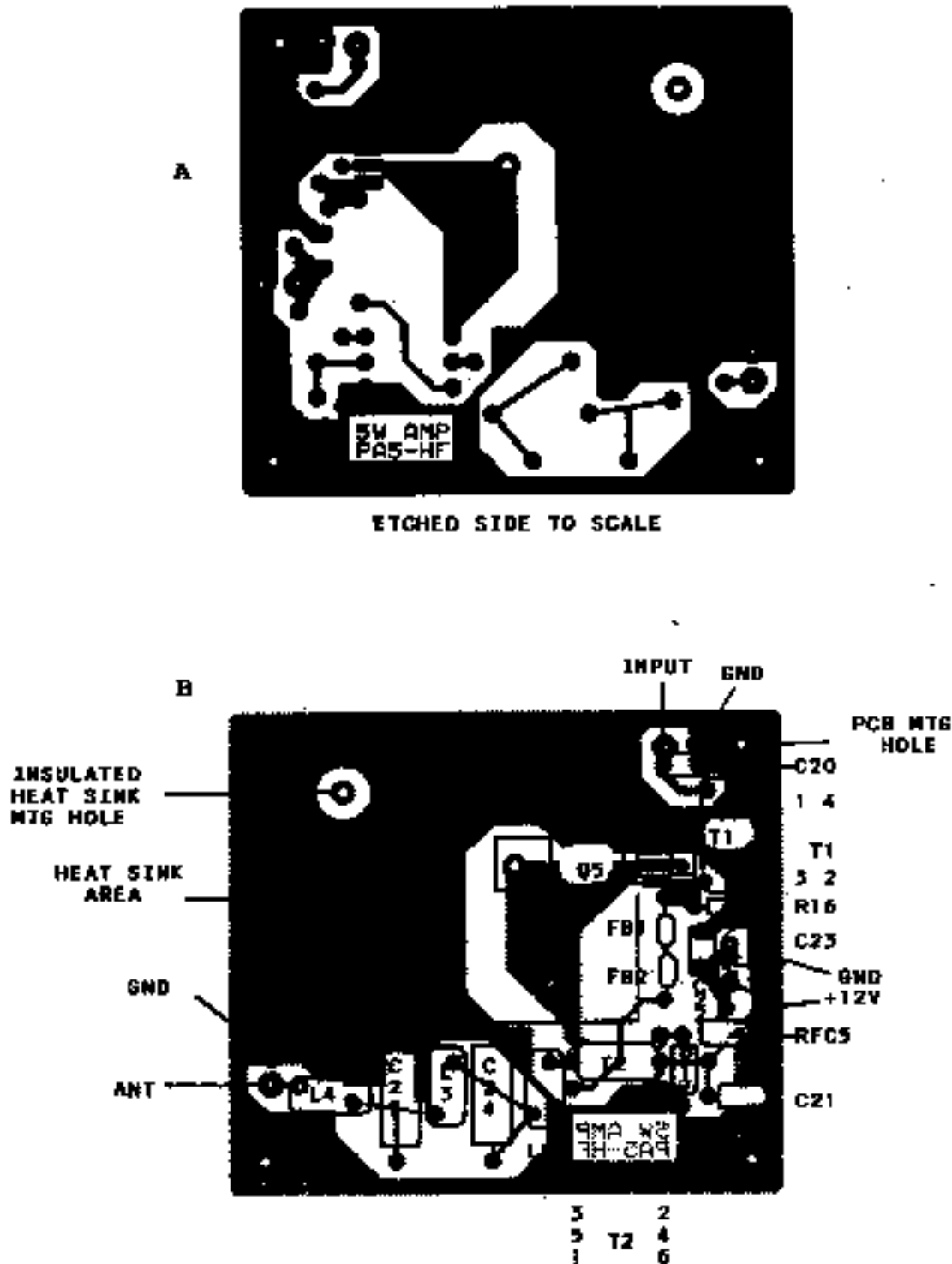
Fig 4-16 -- Artwork for the 5-watt PA5-HF class C amplifier. Illustration A is a scale etching pattern as viewed from the etched side of the board. Drawing B is an Xray view (to scale) of the board, as seen from the component side. PC boards are available from FAR Circuits for $4.50, plus $1.50 shipping. Complete kits may be purchased from Oak Hills Research, Big Rapids, MI 49307 (catalog $1).

You may convert the Fig 4-15 amplifier for linear operation by lifting lead no. 4 of T1 above ground (R16 remains connected to lead no. 4) and applying forward bias in the manner shown in Fig 4-14. A heftier heat sink will be needed if you make this change. The lips of the U-shaped heat sink can be extended to a height of 3 inches if the circuit is used as a linear amplifier.

T1, T2, L1, L2, L3 and L4 are mounted vertically on the PC board. They may be affixed to the board by means of epoxy glue after the amplifier is checked out and operating.

Motorola's 20-Watt HF-Band Linear Amplifier

Although an amplifier that delivers 20 watts of RF power to a 50-ohm load is not in keeping with true QRP operation, it has value when the band conditions are poor, or when greater output power is needed for emergency communications. The circuit in Fig 4-17 was designed by Helge Granberg, K7ES, who is an applications engineer at Motorola in Phoenix, AZ. Complete data are provided in Motorola's **AN-779** application note. This material appears also in Motorola's **RF Device Data** book.

The amplifier utilizes four low-cost plastic transistors in TO-220 style cases. The push-pull class AB driver uses two MRF476s and the push-pull PA has two MRF475s. Operating voltage is +12 to +13.6. The driver and PA transistors are configured for negative feedback of the type found in the Fig 4-15 amplifier. Broadband transformers are used in the amplifier to match impedances and couple the driver to the PA.

Outboard harmonic filtering is <u>mandatory</u>. Do not use this amplifier on the air without including low-pass filtering at the output port. Filter data may be obtained from the normalized tables in **The ARRL Handbook**, all recent editions. I recommend a 5- or 7-element Chebyshev low-pass filter of the pi type (Fig 80, page 2-51 in the 1990 Handbook). All you need do is choose a filter cutoff frequency that is 10 to 15 percent above the highest operating frequency in MHz, then divide this number into the normalized numbers in the tables to obtain the filter capacitance in pF and the filter inductance in uH. The procedure is very simple. Various filter ripple factors are listed in the tables. I generally design for a 0.01 ripple factor, but good performance will result with any of the factors listed.

The PC board for the 20-watt amplifier is double sided. This provides a ground plane that aids stability. The amplifier input and output impedance is 50 ohms. The amplifier operating range is 1.6 through 30 MHz. Input power is on the order of 1/4 watt (250 mW) for 20 watts of output power. This equates to an overall gain of 25 dB.

Details for the broadband transformers are presented clearly in **AN-779**. PC boards for this amplifier are available from FAR Circuits for $10.60 plus $1.50 shipping. Complete kits for this project are available from Communications Concepts, Inc., 508 Millstone Dr., Xenia, OH 45385 (catalog available). CCI sells numerous amplifier kits for higher output power levels as well.

Fig 4-17 -- Schematic diagram of the 1.6- to 30-MHz linear 20-watt amplifier. This is a reproduction of the diagram in AN-779 (courtesy of Motorola, Inc.). Therefore, the symbology differs from that used elsewhere in this book. See text for information about circuit boards and parts kits for this amplifier.
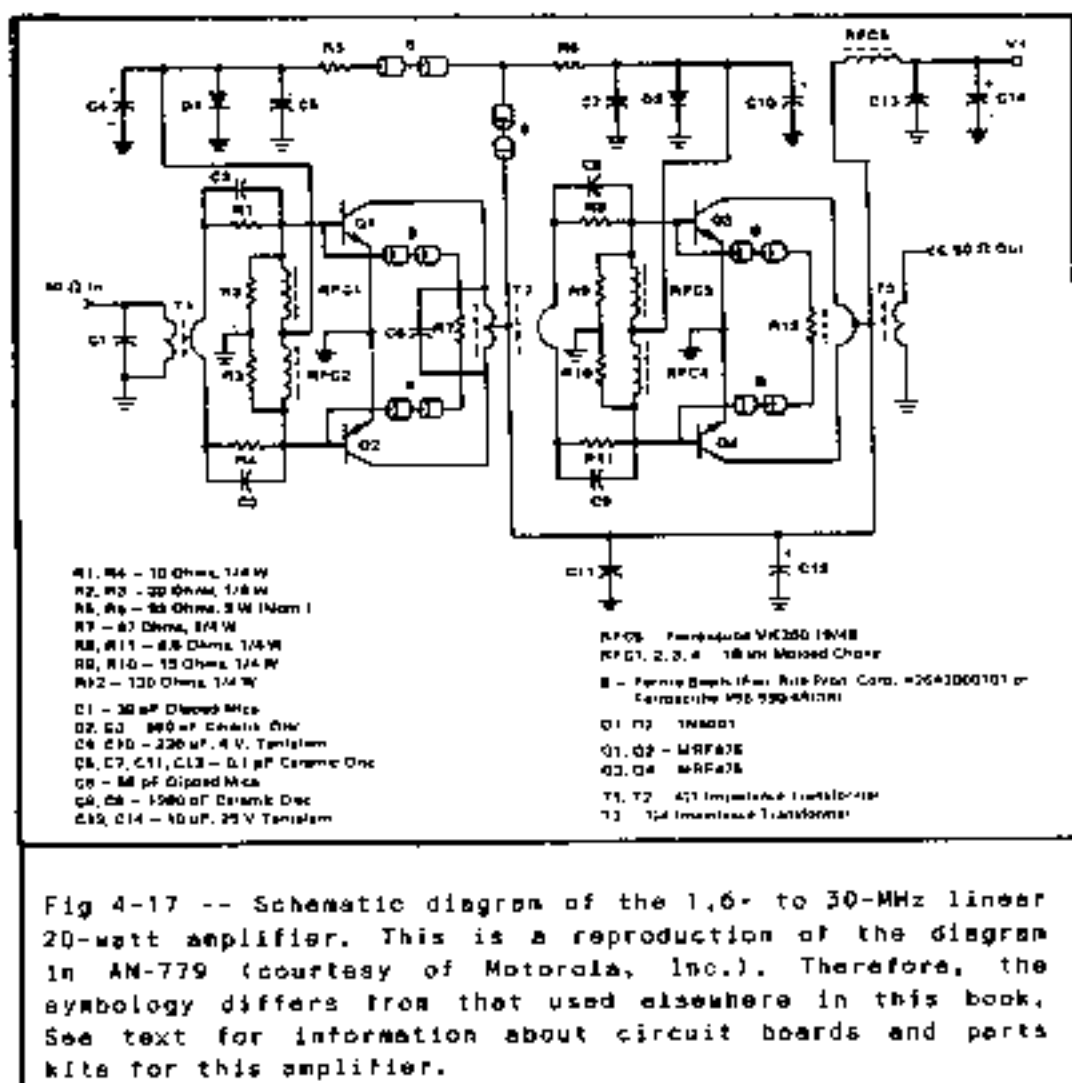
A General-Purpose Broadband Amplifier Strip

Broadband linear amplifiers are useful as low-level power strips in amateur solid-state transmitters. A principal advantage associated with their use is that they can be designed to cover a wide range of frequencies without the need to employ band switching. That is, a given broadband amplifier might provide 40 dB of gain from 1.8 to 30 MHz or higher without requiring tuned circuits. A circuit of this kind can be used easily between an oscillator and an RF power amplifier. Broadband amplifiers are useful also as instrumentation amplifiers when we attempt to measure low-power signals with a frequency meter or scope.

Fig 4-18 contains a practical circuit for a 40-dB broadband amplifier that provides relatively constant output power from 1 MHz to 40 MHz. The gain slumps gradually above and below those frequencies.

Fed-back class A amplifiers are used throughout the circuit. Conventional 4:1 impedance ratio broadband transformers are used between the stages. Negative and degenerative feedback is used to level the gain, enhance stability and ensure that each stage has a 50-ohm input Z.

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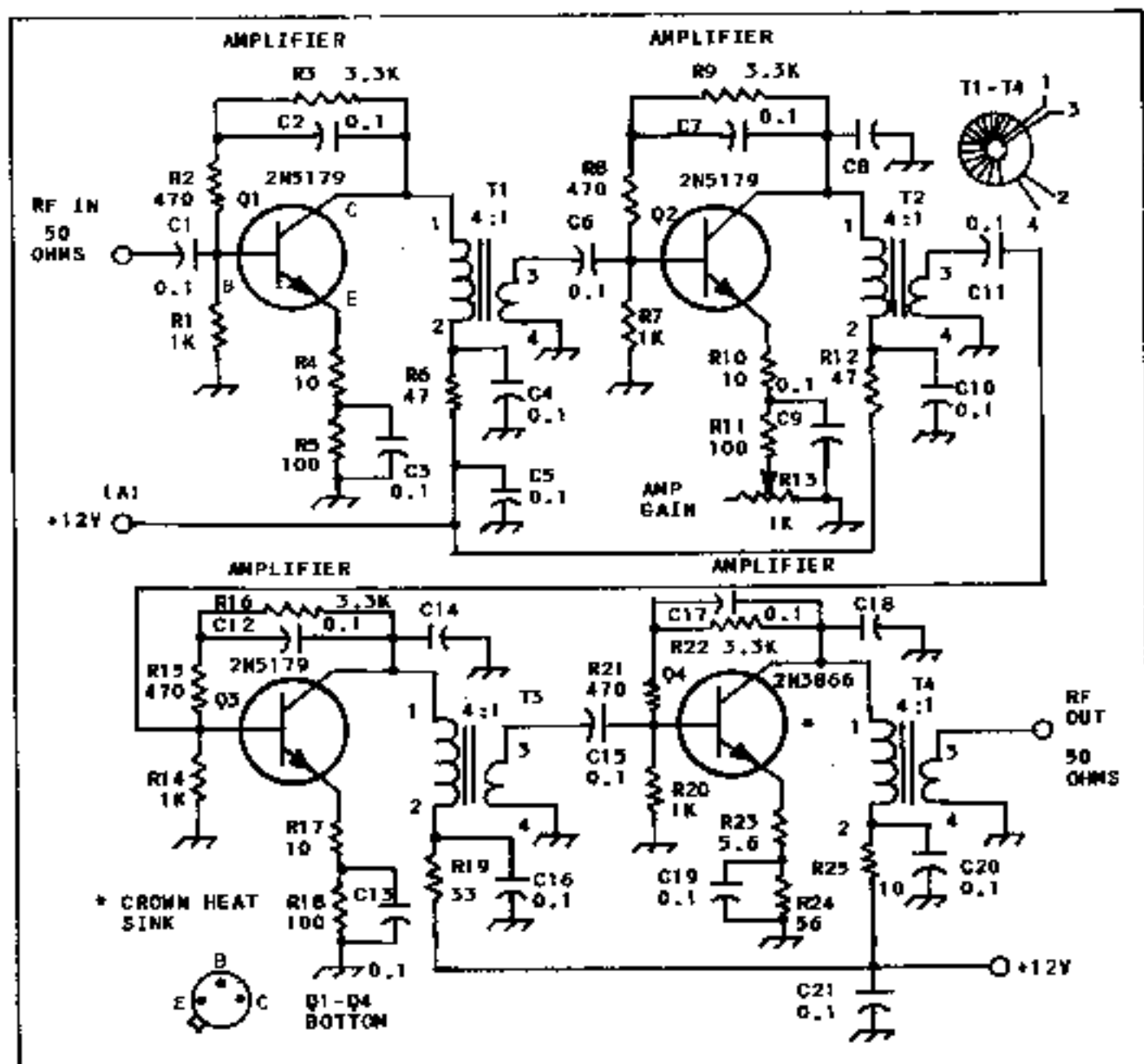
Fig 4-19 — Schematic diagram of a practical broadband preamplifier for use from 1.0 to 40 MHz. Capacitors are disc ceramic. C8, C14 and C18 are shown only as optional units if it is desired to roll off the high-frequency response for a special application. Resistors are 1/4-W carbon film types. Q1, Q2 and Q3 are CATV high gain, high f, transistors to ensure good upper-frequency response. Other transistors, such as 2N3904, 2N2222A or 2N4400, may be substituted, but the upper frequency gain and response will be somewhat degraded. A push-on crown heat sink is needed on Q4. T1, T2 and T3 have 12 primary turns of no. 26 enam. wire on Amidon FT-37-43 ferrite (850 mu) toroids. The secondaries have 6 turns of no. 26 enam. wire. T4 has 12 turns of no. 24 enam. wire on an Amidon FT-50-43 toroid. The secondary has 6 turns of no. 26 enam. wire. R13 is a panel mounted carbon composition control, linear taper. The circuit constants are based on an earlier design by Wes Hayward, W7ZOI.

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ETCHED SIDE TO SCALE
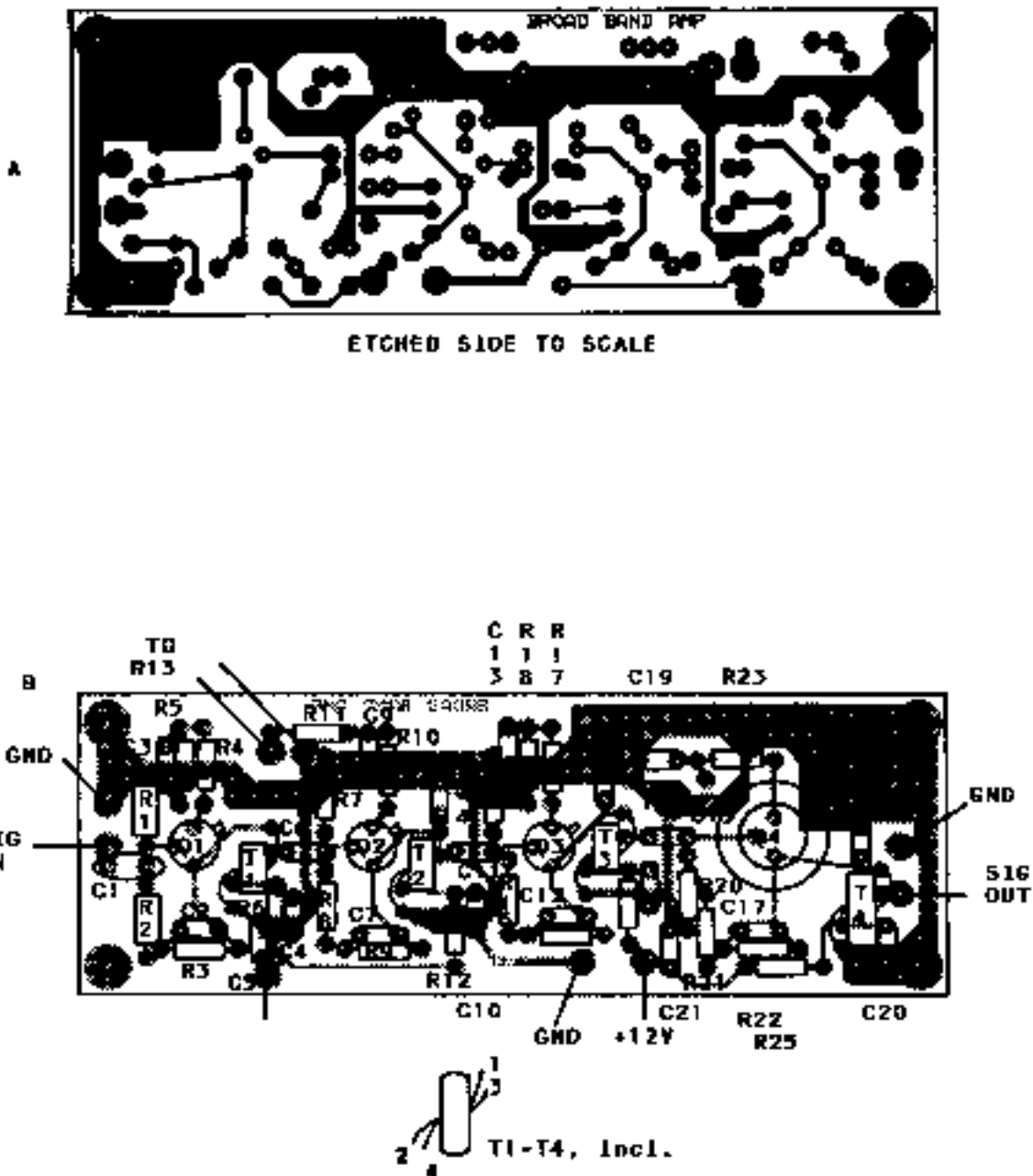
Fig 4-19 -- A scale etching template for the broadband amplifier is shown
at A. This is a double-sided PC board. An Xray view of the board, to scale,
is seen at B. It is shown from the component side of the board. PC boards
for this project are available from FAR Circuits.

A +35 dB MMIC Type of Broadband Amplifier

A pair of NEC uPC1651G MMICs are shown in cascade in Fig 4-20. The
input and output impedance is 50 ohms. Stability is achieved by using
monolithic ceramic capacitors, which are soldered directly to the
PC board conductors, as shown at A in Fig 4-20. The conductors that
carry the signal energy are 50-ohm strip lines. The board pattern
is shown at 2X scale in Fig 4-20. This amplifier is flat from 1.6
to 500 MHz.

Fig 4-20 -- PC board pattern (A) and schematic diagram (B) of the MMIC broadband amplifier. Double-sided board material is required in order to form the 50-ohm strip lines that carry signal energy. Copper or brass U-shaped clips join the ground conductors on both sides of the board, and are labeled "A." Surface-mount ceramic capacitors are used for coupling and bypassing. R1 and R2 are 1/4-W carbon film or composition. This circuit and general layout was developed by H. Johnson, W4ZCB.

DSSC Generator for QRP Voice Communications

You may entertain a notion that QRP phone operation would be fun to try. A double sideband, suppressed-carrier (DSSC) transmitter is easy and inexpensive to construct. Although such a system transmits upper and lower sideband energy at the same time, the carrier can be nulled to 40 or more dB. The absence of a carrier greatly reduces interference to others who may be operating on nearby frequencies. In other words, they will not be annoyed with a heterodyne from your rig. SSB is, of course, the best road to travel in the interest of spectrum conservation, but a 5-watt DSSC signal is seldom offensive in our bands where most operators use 100 to 1.5 kW of peak output power. A 5-watt signal is 23 dB weaker than a 1000-watt signal.

One advantage associated with DSSC transmitters is that the VFO is on the same frequency as the transmitter output signal. No expensive crystal filter is needed, and the parts count is small. A practical DSSC generator is shown in Fig 4-21. The microphone amplifier is a 741 op amp. The audio gain control sets the power output level of the generator. In order to suppress the carrier it is necessary to use a balanced modulator (D1 and D2), just as we must do when we build an SSB generator. Diode balance is assured by (1) using matched diodes for D1 and D2, (2) employing a trifilar-wound transformer (T1) and adjusting the balance control (R8) for suppressing the carrier. It should be possible to obtain 40 dB of carrier suppression with the Fig 4-21 circuit, assuming that symmetrical layout is used.

A class A broadband amplifier, Q2, boosts the output from the balanced modulator. This stage may be coupled to the broadband amplifier in Fig 4-18 to obtain approximately 0.5 watt of DSSC power. It is then a simple matter to add a linear amplifier, such as that in Fig 4-15 (modified as described for class AB service) to obtain 5 watts of output power. The VFOs described earlier in this chapter may be used to generate the desired operating frequency.

The Fig 4-21 circuit is arranged for push-to-talk operation. This helps to simplify the circuit by avoiding the use of a VOX circuit. CW operation is possible by placing the mic amplifier, U1, in standby by means of S1. PNP switch Q1 keys amplifier Q2 during CW operation. If you wish to have the CW feature it will be necessary to unbalance D1 and D2 in order to produce a carrier. This may be done by connecting an RF choke (1 mH) to either end of R8 and feeding +12 volts to the choke through a 5.6k-ohm resistor. An SPST switch can be added to actuate the CW-mode components. This applied dc voltage disturbs the balance of the modulator by allowing dc to flow through one of the diodes.

The circuit in Fig 4-21 can be used also as the core of an SSB exciter. A crystal or mechanical filter is added at the output of Q2 and the LO input is supplied with energy from a crystal oscillator that uses upper- and lower-sideband crystals that are suitable for use with the crystal filter you select. The filter bandwidth should be in the range of 2.0 to 2.4 kHz. Circuits for SSB transmitters are presented in **The ARRL Handbook** and **Solid State Design for the Radio Amateur.**
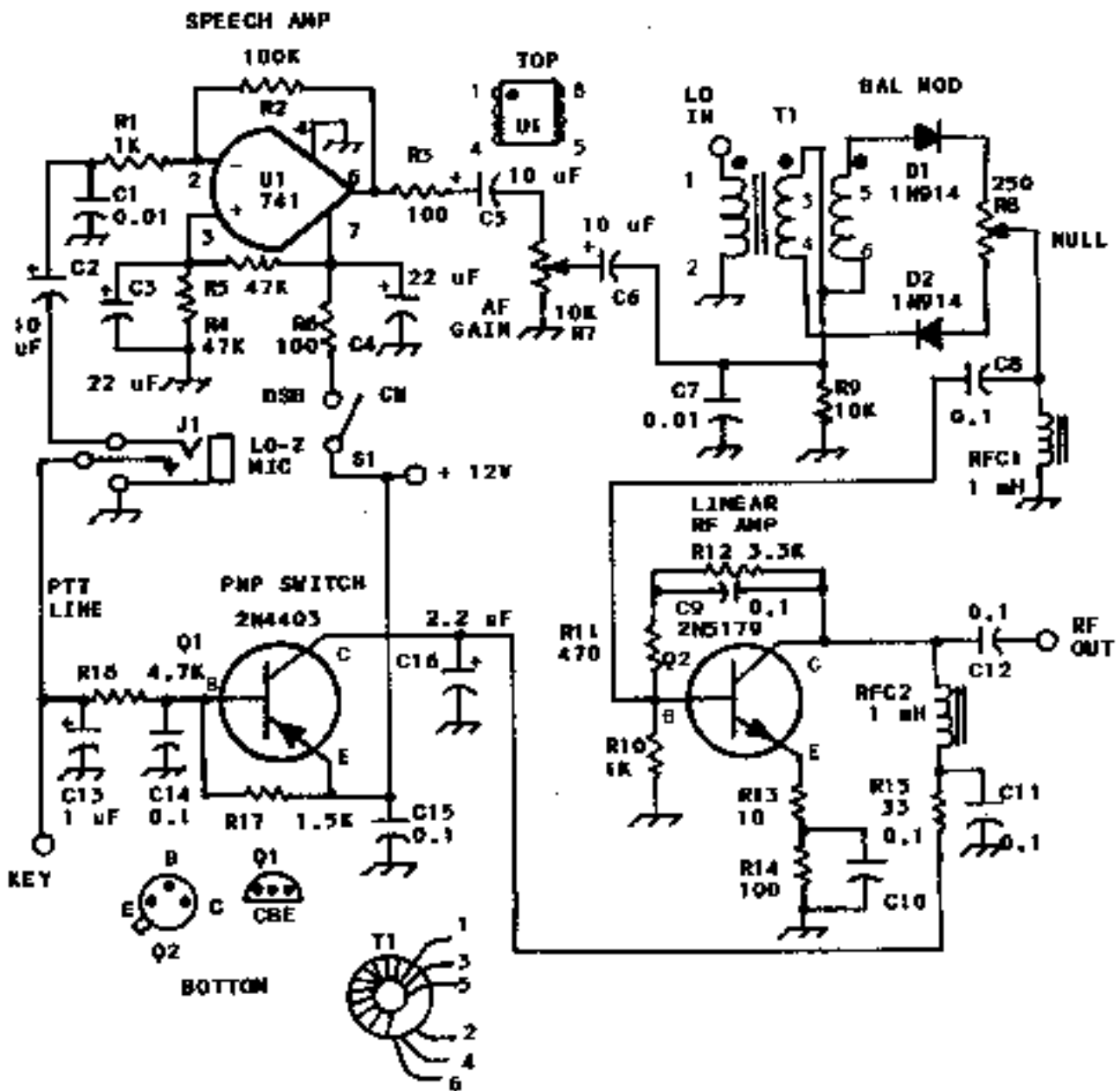
Fig 4-21 -- Schematic diagram of a practical DSSC generator. Capacitors are disc ceramic except those with polarity marked, which are tantalum or electrolytic. Fixed-value resistors are 1/4-W carbon film or composition. R7 is an audio-taper, panel-mounted carbon composition control. R8 is a PC-mount carbon composition control. D1 and D2 should be matched for equal forward resistance by means of an ohmmeter. RFC1 and RFC2 are miniature Mouser Electronics or Oak Hills Research RF chokes. T1 is a broadband, trifilar-wound transformer. Use 12 trifilar turns of no. 28 enam. wire on an Amidon FT-37-43 (850 mu) ferrite toroid (observe winding phase, indicated by the dots). S1 is a SPST toggle and J1 can be a three-circuit mic jack of your choice. Output impedance of Q2 is 200 ohms.

The PTT circuit in Fig 4-21 keys Q2 on and off between the transmit and receive modes. Other circuits in your complete DSSC transmitter may also be controlled from this PTT line.

Fig 4-22 provides a scale etching pattern and parts-placement guide for the DSSC generator. Single-sided board material is used for this project.

Fig 4-22 -- Scale etching template is shown at A. Drawing B is an Xray view of the PC board as seen from the component side. C10, C13 and R17 are not number- ed at B, owing to limited space. R6, R11 and other resistors shown in that manner are mounted vertically. PC boards for this project are available from FAR Circuits for $4.50 plus $1.50 shipping.
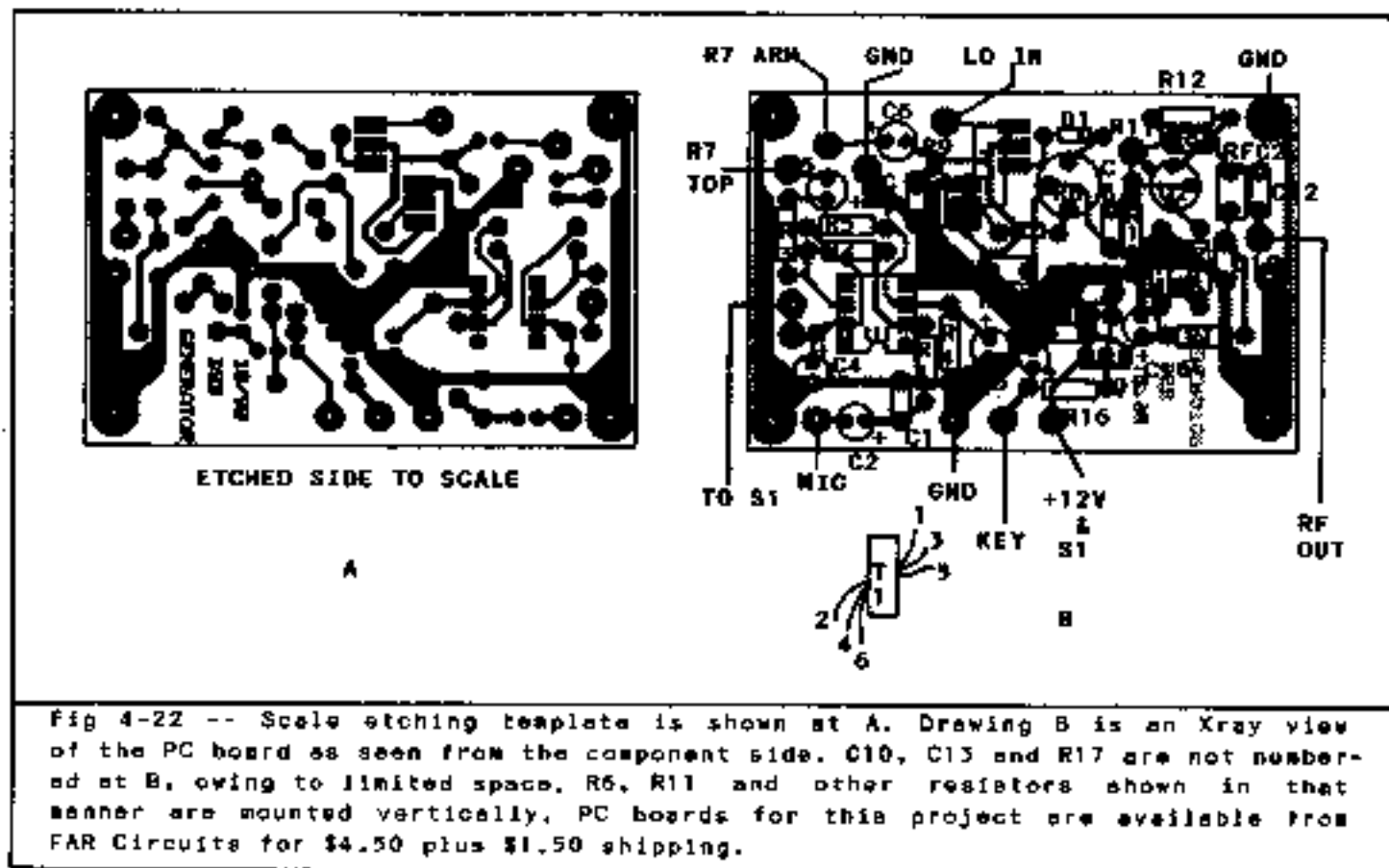
TRANSMIT-RECEIVE CIRCUITS

Various methods are used by QRPers to switch the antenna between the receiver and transmitter. Some of these circuits have provisions also for muting the receiver during the transmit period. The question arises whether to use full break-in (QSK) or to use break-in delay of the type found in VOX circuits. It is purely an operator's choice. Many CW operators prefer full QSK, which is beneficial especially when hand- ling traffic or doing emergency work. Break-in delay circuits trigger a changeover relay, and the time constant is usually adjustable in order to control the drop-out time of the relay. A 1-second drop-out time is the choice of many for a break-in delay TR circuit. The main disadvantage of the latter system is that the operator may miss the first letter or two of the other person's message if he comes back too quickly. This is not true of full break-in.

Both types of TR circuits are presented here. You will need to consider the pros and cons of these circuits before making a choice.
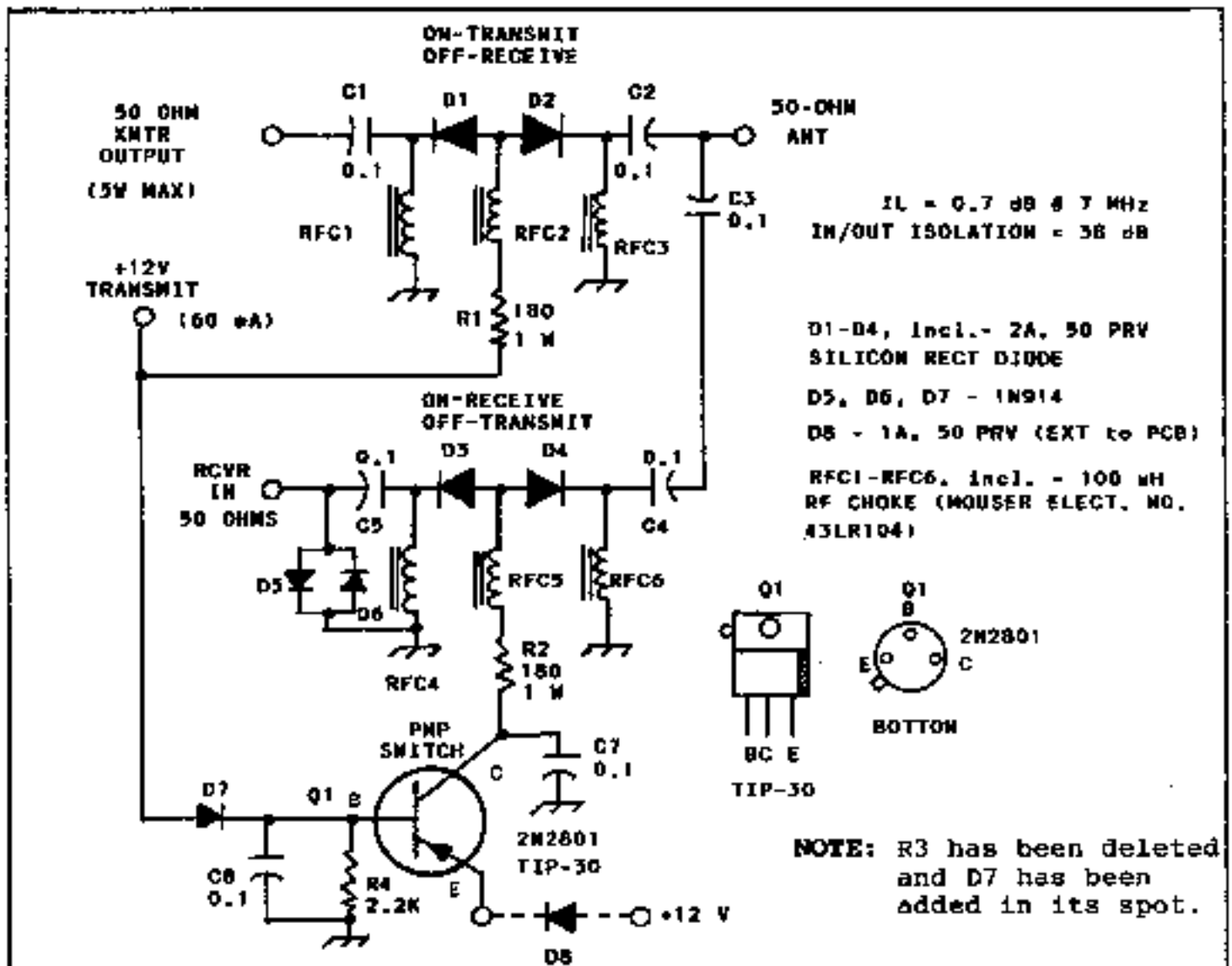
Fig 4-23 -- Schematic diagram of the W1FB full QSK TR circuit. Capacitors are disc ceramic. Resistors are 1/4-W carbon film or composition except those with wattage indicated. D1 through D4 are 2-A, 50- or 100-PRV rectifier diodes, D5 and D6 are 1N914 or equivalent diodes. Q1 may be any PNP transistor that can safely pass 300 mA or greater dc current. Data for the RF chokes is listed above.

Full QSK Diode TR Circuit

Fig 4-23 shows a circuit I developed for full QSK. It can handle up to 5 watts of RF power. Ordinary rectifier diodes are used for the switches. As is true of any passive circuit, some insertion loss will occur. This switch has a 0.7-dB loss. Therefore, if we apply 5 watts of RF energy at the input of D1, we will have about 4.75 watts at the output of D2. This is an acceptable tradeoff for the benefits of the circuit. The physically smaller the diodes the better the input-output isolation of the switch. This is because the smaller rectifier diodes have smaller junctions, and hence less junction capacitance. The application of reverse bias (+12 V to the D1, D2 cathodes) when

D1 and D2 are in the OFF mode would further enhance the input-output isolation. I find that 38 dB of isolation is ample for my needs.

It is necessary to bias the diodes so that they are fully turned on. If not, the switch will be quite lossy. There is approximately 30 mA of dc flowing through each switching diode in the circuit of Fig 4-23. This 60-mA current drain does not make the circuit attractive for battery-operated QRP gear, unless you plan to use your car battery as a power source.

The RF chokes need to have an XL of 600 ohms minimum in order to ensure that the 50-ohm antenna line is not affected by the chokes. Since three RF chokes are effectively in parallel when the diodes are turned on, the resultant   overall XL is 200 ohms (4 X 50 ohms), which is the least acceptable XL value for a 50-ohm line. The dc resistance of the RF chokes must be low in order to assure proper turn-on for the diodes.

Q1 needed to be included in order to provide the correct TR logic for this circuit. In the TRANSMIT mode Q1 is turned off by the positive forward bias supplied to its base. In the RECEIVE mode, R4 causes Q1 to turn on and pass dc current to D3 and D4.

D5 and D6 are protective diodes for the receiver input circuit. This safety measure is included in the event D3 and D4 should become defective.

The series switching diodes do not cause TVI when they are fully turned on. The output waveform, during tests, was as clean as the input waveform when using the circuit values listed in Fig 4-23.
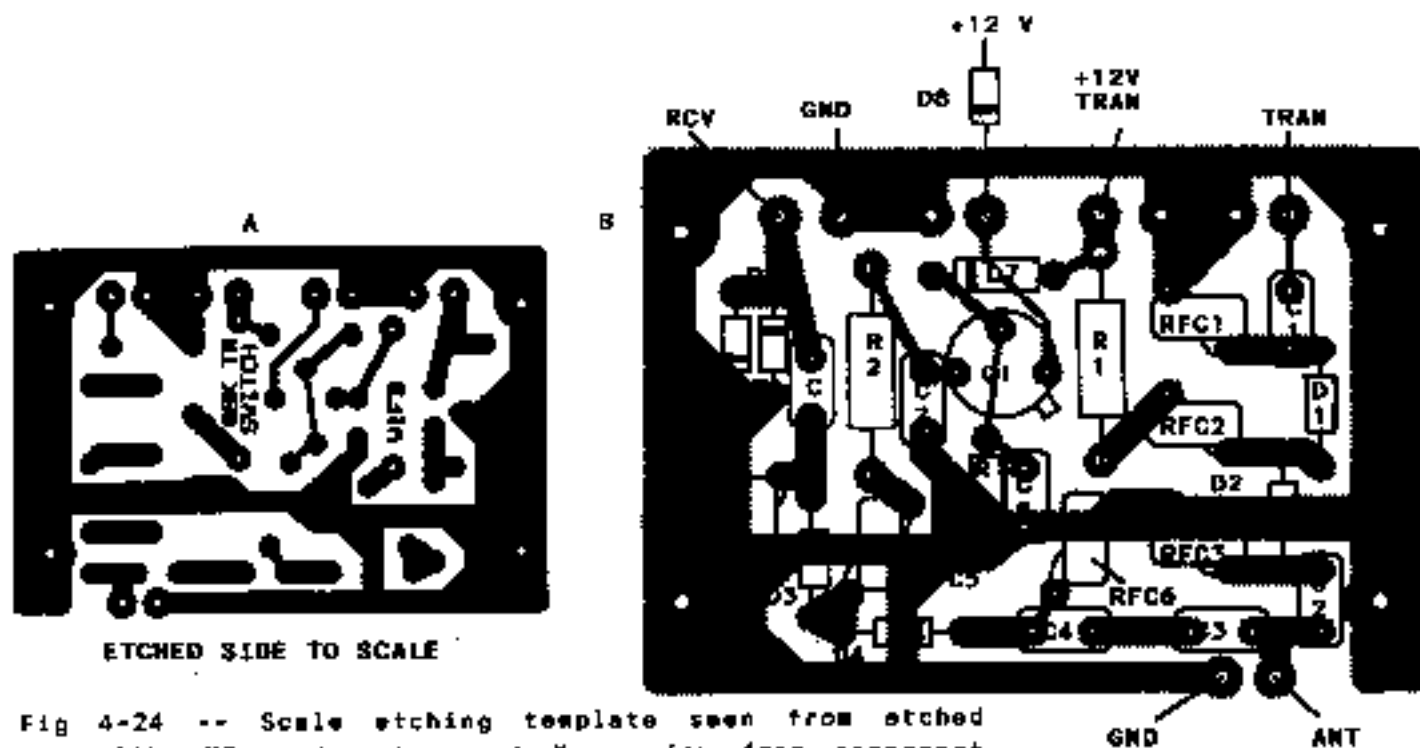
Fig 4-24 -- Scale etching template seen from etched side (A). X2 parts placement Xray view from component side is seen at B. Boards avail from FAR Circuits for $3.75 plus $1.50 shipping.

A Break-In-Delay TR Circuit

Fig 4-25 contains a circuit diagram for a practical delay type of
TR circuit. Q1 is a PNP dc switch and Q2 is an NPN relay-driver switch.
When the junction of R1 and C1 is grounded, Q6 conducts and forward
is applied to the base of Q2, causing it to conduct. When Q2 conducts
it allows current to flow through K1, thereby causing its contacts
to close. This is the TRANSMIT mode. When the R1-C1 junction is opened,
Q1 ceases to conduct, but the charge in C3 keeps Q2 turned on until
the charge decays, at which time K1 opens. R3 and R4 determine the
discharge rate for C3. Adjustment of R4 sets the drop-out time for
K1.

D1 prevents positive voltage transients from harming Q1. D2 establish-
es a 0.7-V bias for Q2. Without D2 the relay does not release, owing
to leakage current in Q2. D3 clips dc voltage peaks that occur when
the field of K1 collapses.

K1 is 12-V DIP relay of the DPDT type. One set of contacts may be
used for receiver muting. The remaining contacts are used for antenna
switching. The relay is not suitable for full break-in keying, even
though R4 can be adjusted to permit full QSK. Not only would relay
contact bounce cause the keyed waveform to be poor, the relay would
wear out quickly if it were cycled in accordance with the segments
of each keyed character. A reed relay or relay with mercury-wetted
contacts can, however, be keyed directly for full QSK up to 25-30
WPM.
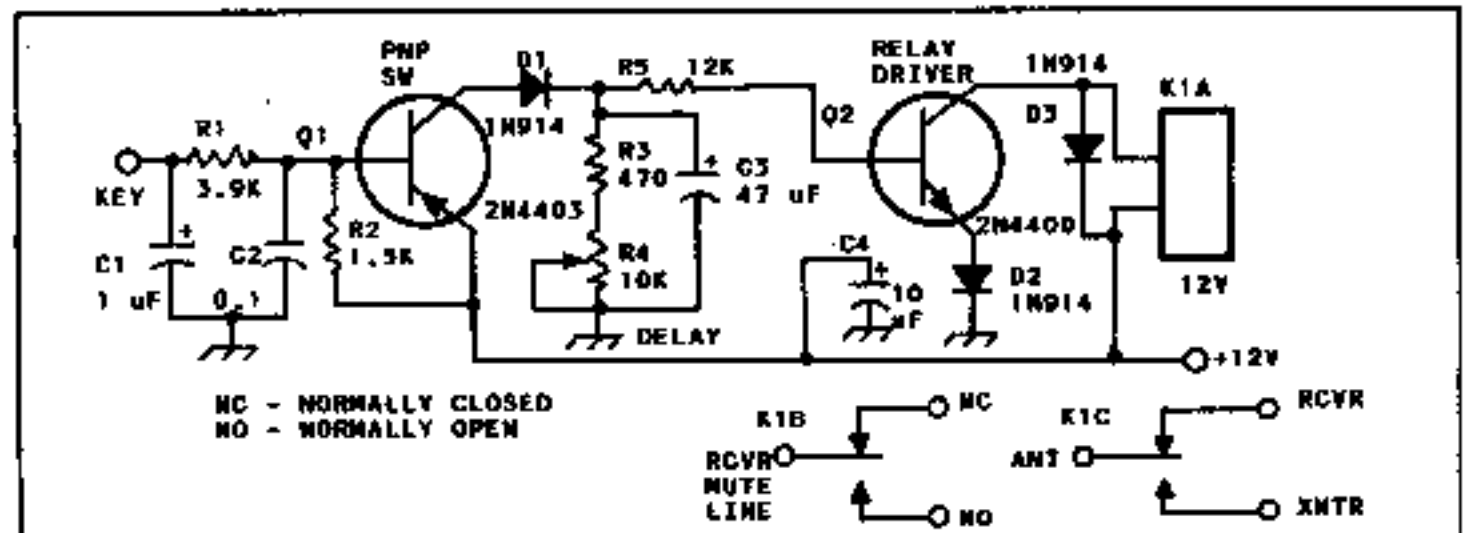
Fig 4-25 -- Schematic diagram of a practical break-in delay TR module. All
capacitors (except C2) are tantalum or electrolytic. C2 is a disc ceramic
unit. Fixed-value resistors are 1/4-W carbon film or composition. R4 is a
10K-ohm trimmer type of control, PC-mounted. K1 is a PC-mount DIP relay,
2PDT, 12 V dc (Omron no. G2V-2B4P-US or equiv., avail. from Hosfelt Elect-
ronics, Steubenville, OH 43952. Catalog avail. via 1-800-524-6464).

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Fig 4-26 -- Scale etching template (A) for the TR-IA break-in delay board as seen from the etched side. Drawing B is an Xray view of the PC board as viewed from the component side. PC boards for this project are available from FAR Circuits for $3.50 plus $1.50 shipping. Complete kits are available from Oak Hills Research ($1 for catalog).
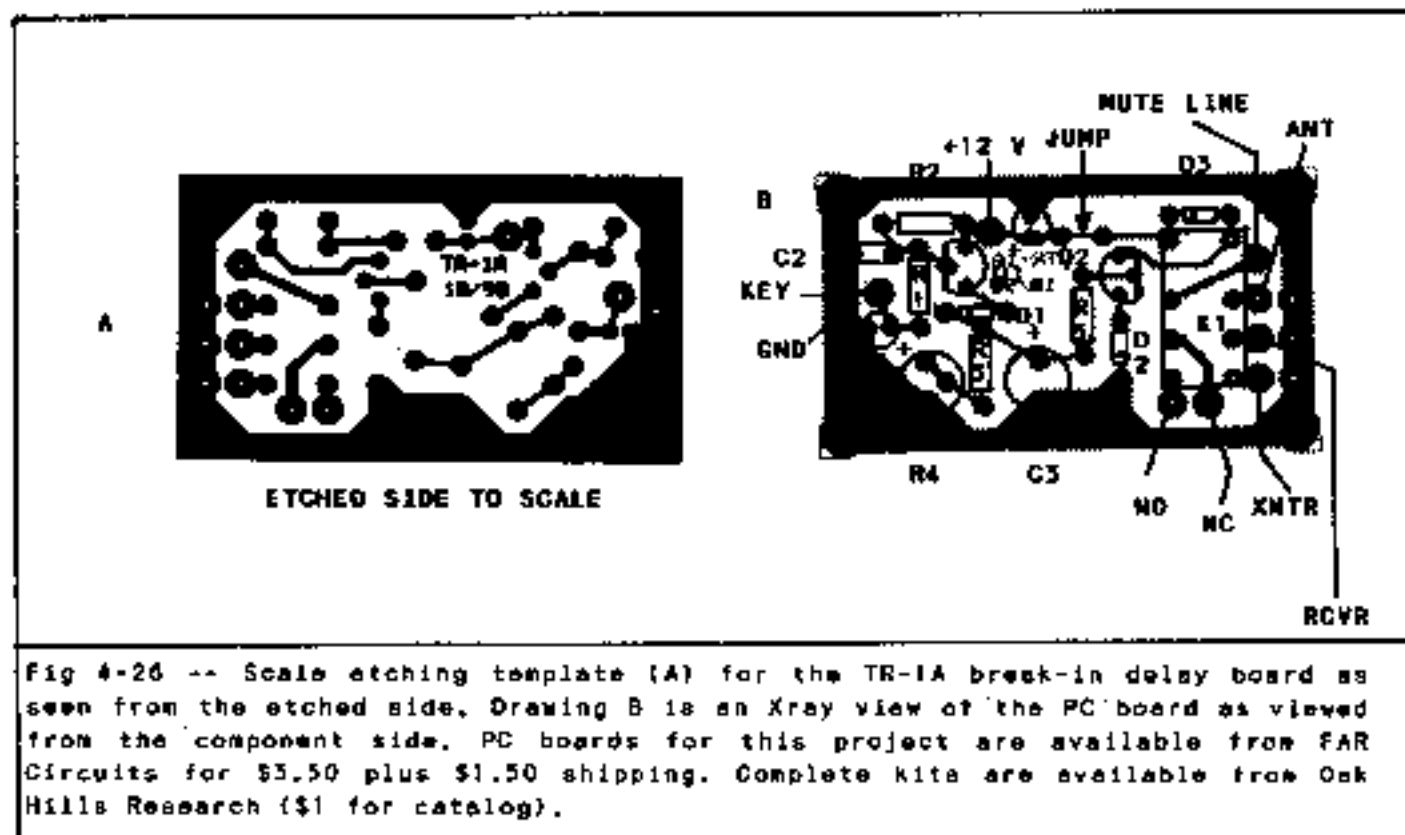
Another QSK Circuit

The circuit in Fig 4-27 may be applied to QRP transmitters for single-band operation. It is a method that was developed by W. Hayward, W7ZOI. Antenna energy is sampled at the collector of the transmitter PA, routed through a series-resonant circuit (tuned to the operating frequency) and then to the receiver input port. Two silicon switching diodes are included to protect the receiver during transmit periods. These diodes (D1 and D2) conduct during TRANSMIT and establish an RF level of 0.7 V RMS maximum, which will not harm the receiver input circuit. Loss through the series-resonant circuit is minimal. The output filter for the PA is in line with the receiver antenna line during RECEIVE, and this is beneficial in adding additional select-ivity at the input to the receiver. C1 in Fig 4-27 must be absorbed into the filter input circuit. Its value is subtracted from the normal value for C2, since it is effectively in parallel with C2 when D1 and D2 conduct. The reactance of C1 and L1 is 450 ohms. Thus, for 40 meters we will use a 50-pF capacitor.
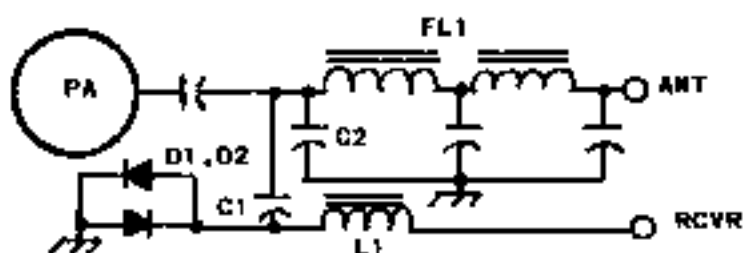
Fig 4-27 -- Example of a simple one-band TR circuit. C1 and L1 have a 450-ohm reactance. See text for more information.

SIDETONE CIRCUITS

It is convenient and often necessary to include a sidetone oscillator
in the transmitter circuit so that we can monitor our CW sending.
Audio output from the sidetone circuit is routed to the audio output
stage of the receiver -- an audio amplifier that is not muted along
with the rest of the receiver circuit. Sidetone energy is applied
after the audio-gain control in order to prevent the sidetone level
from changing when the AF gain control setting is altered.

Various circuits are used for generating a sidetone. The frequency
of the tone is usually between 500 and 1000 Hz. This depends upon
operator preference. A 700-Hz tone is the choice of many amateurs,
since it closely matches the CW offset found in many transceivers.
Most sidetone oscillators are turned on and off by keying the +voltage
line to the oscillator. Various waveforms are generated by sidetone
oscillators. They range from a sine wave to a square wave to a sawtooth
type, depending upon the circuit used. Fig 4-28 shows various simple
circuit you may use for providing a sidetone signal.
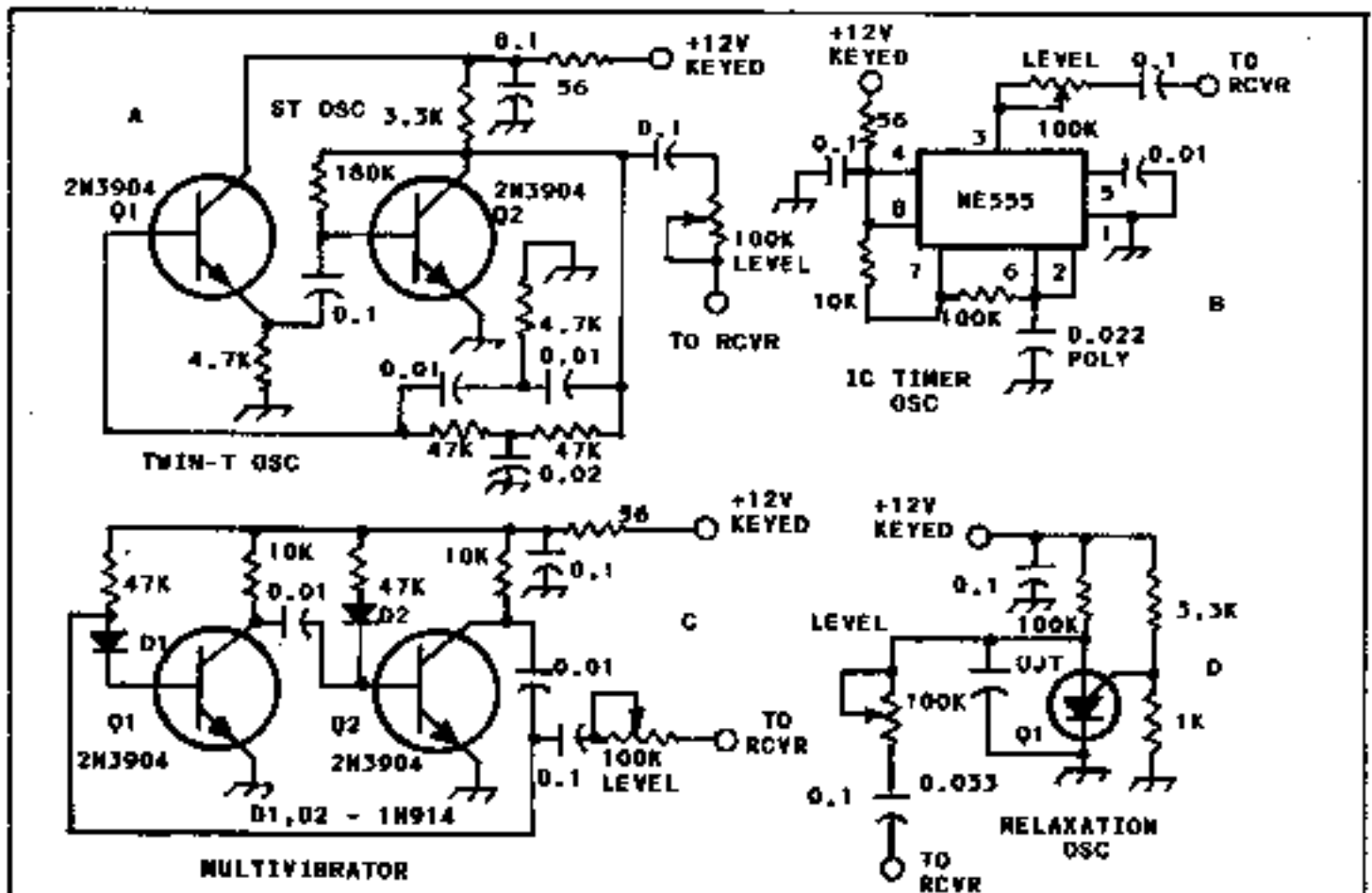
Fig 4-28 -- Four audio sidetone oscillators. Freq. determining capacitors
are mylar or polystyrene. Fixed-value resistors are carbon film or composition,
1/4-W. Variable resistors are PC-mount carbon. Q1 at D is a Motorola programm-
able unijunction transistor, Motorola MMBPU131 or equiv.

CHAPTER SUMMARY

I have attempted to present practical information in progressive form within this chapter. The stepping stones are in place toward fun and success as a QRPer. Transceivers have not been presented because many such projects have appeared in the pages of **QST** and other amateur publications. Receivers and transmitters from this book can be used in combination to provide transreceivers which can, in effect, perform the same function. A compact 40-meter transceiver of simple design is presented in **W1FB's Design Notebook**. PC boards are available for that project by ordering them from FAR Circuits.

I hope this section of the book has encouraged you to build your own QRP transmitters. These proven circuits do not contain expensive parts, and they are within the ability of any amateur who has the courage to heat a soldering iron and stuff components in the holes on a PC board. Most of the parts can be garnered at ham-radio flea markets and from the many surplus electronics mail order houses.

CHAPTER 5

QRP ACCESSORIES

QRP operators are concerned about many things that require small items of accessory equipment for monitoring equipment performance. This is particularly true during field operation where adjustments and repairs are often necessary, owing to changes in ambient temperature, transport shocks and such. We are always concerned about obtaining maximum transmitter output power, which hinges in part upon maintaining a low antenna SWR. Voltage monitoring is important when we operate from a battery power supply.

It is convenient to combine as many pieces of accessory gear into one box as is practicable. For example, a dummy antenna, SWR indicator and 100-kHz frequency standard may be housed in a single small cabinet. I once built a gadget of this type which contained (1) a dummy antenna, (2) a 100-kHz standard, (3) an SWR indicator and (4) a relative field-strength meter. It was called an "Omnibox." A single 200-uA meter was switched to provide field-strength, SWR and RF power (across the 50-ohm dummy load) readings. A transistor radio 9-V battery was used to power the 100-kHz oscillator and its amplifier. This piece of gear proved invaluable during QRP Field Day operation and when I took my station with me on camping trips. The various circuits I use in my Omnibox are described in this chapter. You may wish to package all of them in one case, or you may prefer to build only one or two of the circuits.

Field Strength Meter

Perhaps you are wondering what use we can find for a field-strength meter. There are a number of applications that come to mind. An example is the use of this instrument for simply monitoring the transmitter output power when no other indicator is available. I often place my FS meter near the feed line in my camper during portable operation. I set it for midscale deflection when the transmitter is keyed. An occasional check of the meter reading assures me that all systems are functioning okay. I have used the FS meter, when afield, as a visual indicator when peaking the tuned circuits in my QRP transmitter. There have been times when I forgot to take my SWR indicator to a campsite. I was able to adjust my antenna tuner for maximum transmitter output power by tuning its circuits for maximum FS. The FS meter was placed near the antenna and a remote meter was plugged into the FS meter to permit observing the signal increase at the operating position. An FS meter is useful also when adjusting an antenna loading coil or feed-point matching section. It is better, of course, to use an SWR meter for these applications, if you have one available.
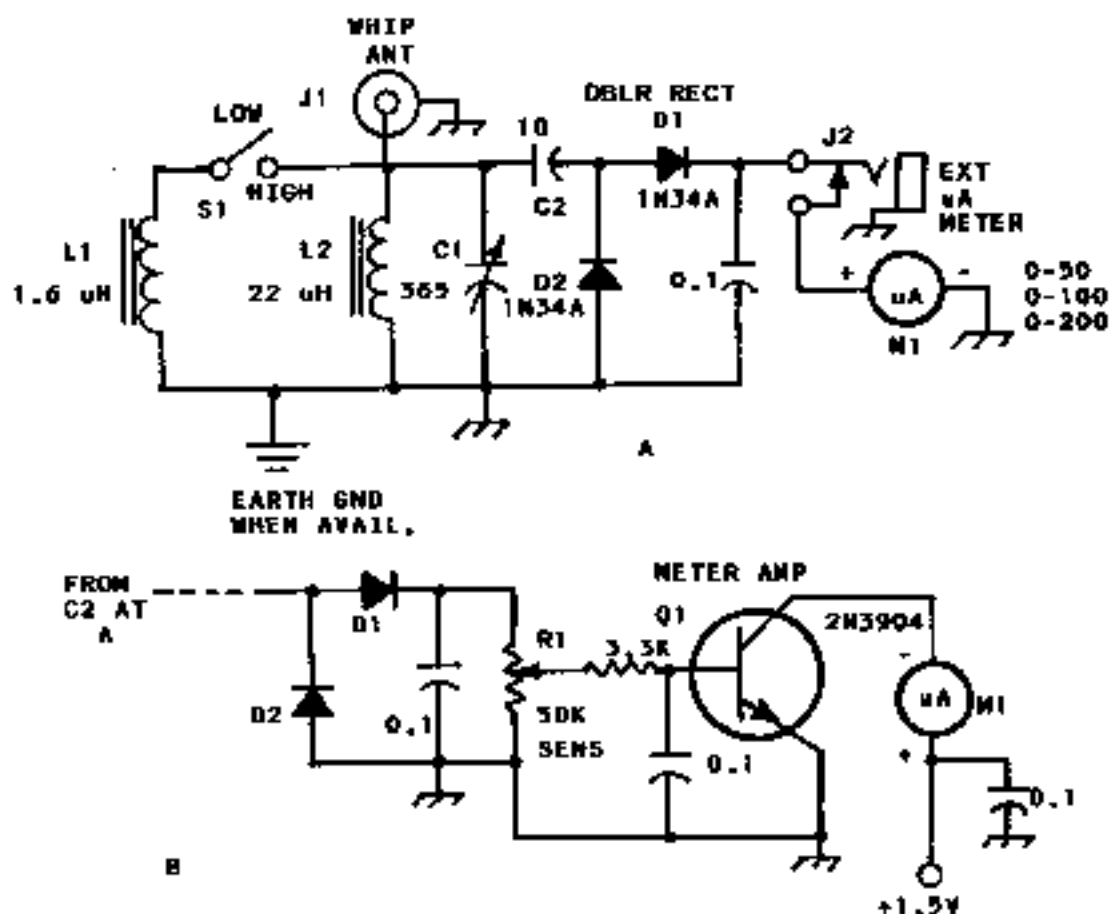
Fig 5-1 -- Simple tunable diode FS meter (A). C1 is a 365-pF air variable. Other capacitors are disc ceramic. D1 and D2 are germanium diodes, type 1N34A or 1N60. J1 is a phono jack or BNC connector. J2 is a standard closed circuit phone jack. L1 has 20 turns of no. 26 enam. wire on an Amidon T50-6 toroid. L2 contains 17 turns of no. 24 enam. wire on an Amidon FT-50-61 toroid (125 mu). M1 may be any dc meter rated for 200 uA or less. Circuit B shows how to add a transistor meter amplifier for greater sensitivity. R1 is a 50K-ohm linear taper carbon composition control, panel mounted.

Fig 5-1A shows a diode type of FS meter that covers from 1.8 through 30 MHz by means of two inductors. The low end of the range uses L2. For the upper end of the HF spectrum it is necessary to place L1 in parallel with L2 by means of S1, an miniature toggle switch. L1 and L2 in parallel equals 1.49 uH. A short whip antenna (3 feet long) is plugged into J1. D1 and D2 operate as a voltage doubler to enhance the sensitivity of the instrument. An external meter can be plugged in at J2 (see text). An earth ground is recommended, when you can provide one. A reference dipole can be used with the circuit at A if a small link is wound at the grounded end of L2 (approx. 3 turns). The coaxial feed line is then connected to the input link.

Fig 5-1B shows how to add a transistor meter amplifier for improved sensitivity. A 1-mA dc meter can be used at M1 to obtain sensitivity that is comparable to the circuit in Fig 5-1A. R1 serves as a sensitivity control for the circuit at B. C1 at A may be detuned to reduce the needle deflection at M1, if no sensitivity control is used.
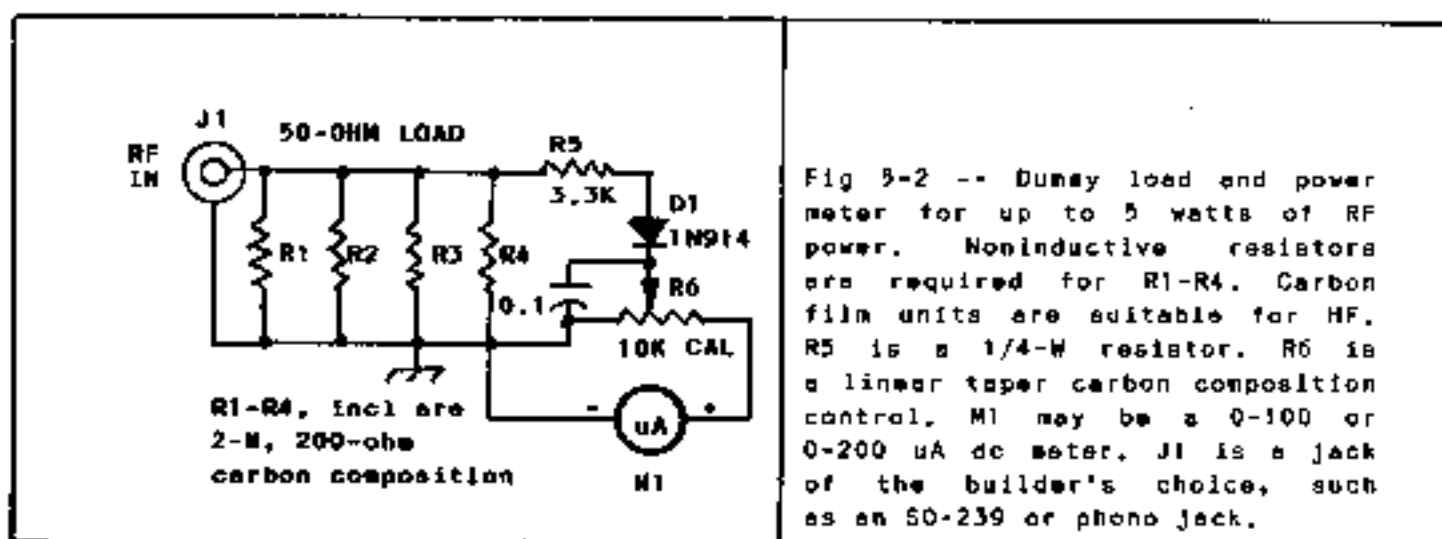
Fig 5-2 -- Dummy load and power meter for up to 5 watts of RF power. Noninductive resistors are required for R1-R4. Carbon film units are suitable for HF. R5 is a 1/4-W resistor. R6 is a linear taper carbon composition control. M1 may be a 0-100 or 0-200 uA dc meter. J1 is a jack of the builder's choice, such as an SO-239 or phono jack.

Dummy Load/RF Power Meter

Fig 5-2 contains a circuit that may be used as a 5-W dummy load and RF power meter. Short bursts of power up to 10 watts may be applied to the load resistors without damaging them. Do not exceed a key-down period greater than 15 seconds for power amounts above 5 watts. Allow 30 seconds for the load to cool if powers above 5 watts are used.

An RF probe and VTVM may be used to calibrate M1 in watts. Attach the RF probe to the junction of R4 and R5 then calculate the power by means of $P(watts) = E(rms)^2$ divided by R(ohms). R6 must be returned to the same position used during the calibration process, whenever future power measurements are undertaken. Note the meter reading for M1 under various power levels and draft a chart for future use. By using various settings of R6 and making meter-reading charts, you can develop ranges of 0-1 W, 0-5 W and 0-10 W. This requires a different calibration mark for the R6 knob, respective to each power range.

It has become difficult to locate 2-W carbon composition resistors. Newark Electronics still lists them in its catalog. Combinations of 1-W carbon resistors (the net resistance being 50 ohms) may be used instead of the 2-W units specified. Keep the resistor leads short to minimize unwanted stray inductance. The load can be made by sandwiching the resistors between two pieces of double clad PC board. Holes are drilled in the board material and the resistor leads are pushed through the holes and soldered to the board. The two end plates should be snug against the resistor bodies.

QRP SWR Meter

Fig 5-3 contains a schematic diagram of a QRP SWR meter. The circuit follows the classic design by Warren Bruene of Collins Radio. A toroid is used for the T1 core. This transformer samples the 50-ohm RF line and causes RF current to flow through the secondary winding. D1 and D2 convert the RF current to dc. C1 and C2 are adjusted for bridge balance (forward and reflected power modes). This is a 50-ohm bridge and it is capable of full-scale meter deflection at 1 watt. It may be used for power levels up to 100 watts.

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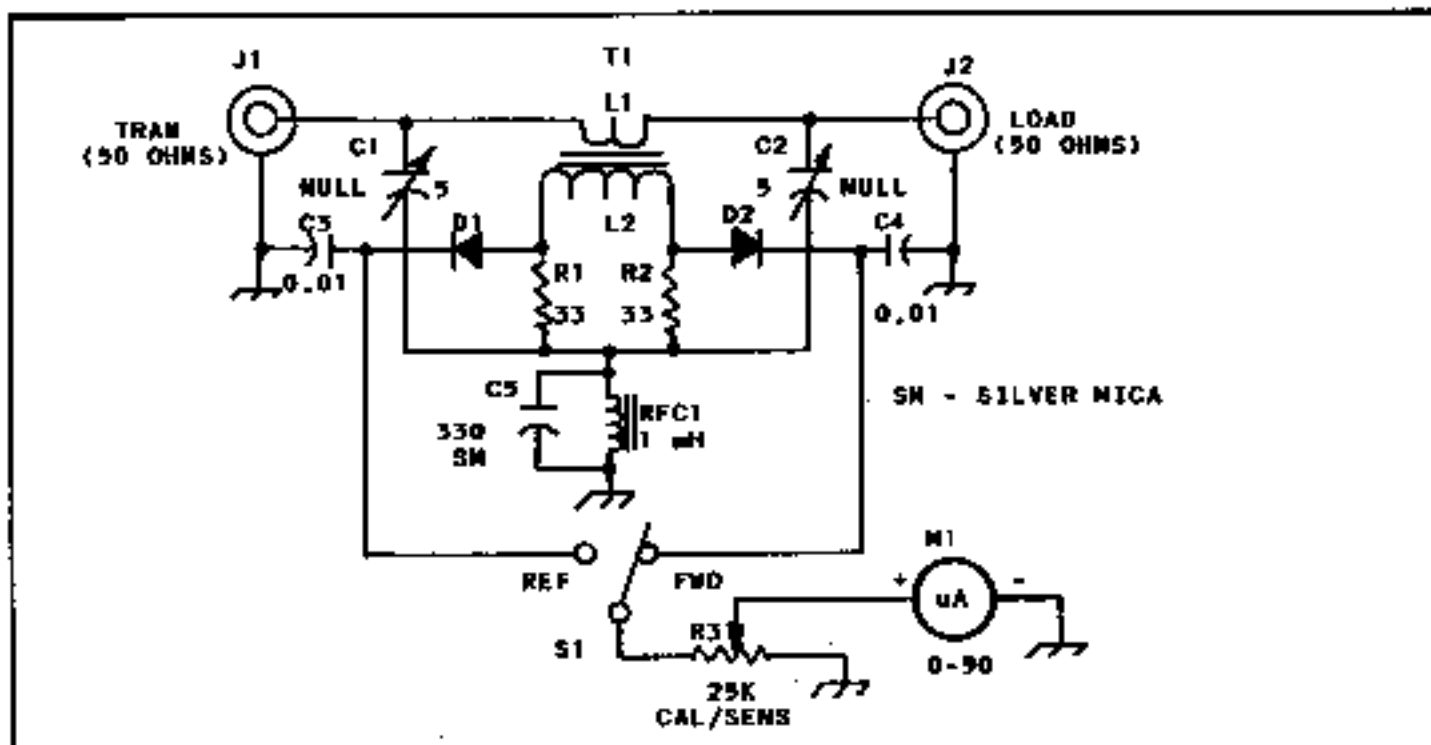
Fig 5-3 -- QRP SWR bridge circuit. C1 and C2 are 5-pF miniature ceramic, air
or glass piston trimmers. Minimum C should be 1 pF or less. C3 and C4 are
disc ceramic. D1 and D2 are HP2800 or equivalent hot-carrier or Schottky diodes.
Diodes should be matched for equal forward resistance. Matched 1N914s may
be used. J1 and J2 are coaxial connectors of the builder's choice. R1 and
R2 are 1/2-W carbon composition, 5% tolerance (should be closely matched).
R3 is a panel-mount 25K-ohm carbon composition linear-taper control. RFC1
is a miniature RF choke. Meter M1 may be 100 uA if reduced sensitivity is
acceptable. T1 has 60 turns of no. 30 enam. wire on an Amidon T68-2 toroid
(red) for L2. L1 is two turns of no. 26 enam. wire over exact center of L2
winding.

Physical and electrical symmetry are important when you construct
the SWR bridge in Fig 5-3. This helps to ensure balance of the bridge
circuit. This instrument needs to be constructed on a single-sided
PC board, using PC-board conductors of uniform width in the RF part
of the circuit. Minimum foil length is the criterion to follow when
designing the PC board. The conductors need to be 1/8 inch or greater
in width in order to minimize stray inductance. This bridge is shown
photographically in **Solid State Design for the Radio Amateur**, page
150.

Calibration requires attaching a 50-ohm resistive load at J2. Power
is applied at J1 and S1 is placed in the REF position. Adjust C2 (use
an insulated screwdriver) for a null in M1 reading. The needle should
drop to zero. Now, reverse the connections at J1 and J2, place S1
in the FWD position and adjust C1 for a null. Repeat these steps until
the bridge is balanced. This procedure should be followed with the
signal source at 14 MHz. When using the instrument, adjust R3 for
a full-scale reading with S1 in the FWD position. Set S1 in the REF
position and read the relative reflected power. Adjust your antenna
tuner for a zero REF reading at M1.

An excellent QRP SWR bridge circuit that first appeared in **SPRAT**, the UK QRP journal, appears in **W1FB's Design Notebook**. The presentation is on page 173, and a PC pattern is provided on page 174. PC boards for that project are available from FAR Circuits.

A 100-kHz Frequency Standard

It is helpful to have a secondary frequency standard available for checking the dial calibration of our home-made receivers. In fact, some countries require that we have secondary standards before they will issue a visitor's license to operate. I learned about this requirement during two QRP DXpeditions in the West Indies. Receivers tend to drift, particularly in the heat of the tropics, and I was fortunate for having taken my 100-kHz standard on those trips. The VFOs in receivers and transmitters tend to drift considerably also when the gear is stored or used in a tent on hot summer days. The Fig 5-4 circuit is simple and easy to build. It should be helpful when you go afield with your QRP station.
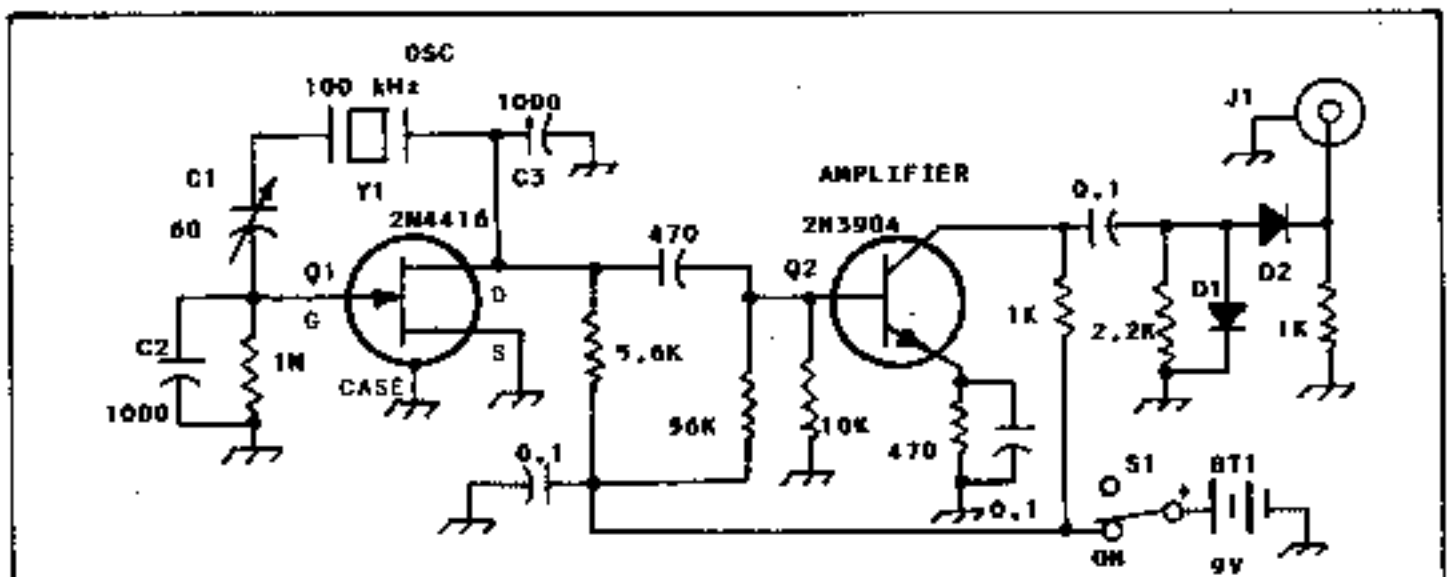
Fig 5-4 -- Schematic diagram of the 100-kHz frequency standard. C1 is a 60-pF ceramic or plastic trimmer. Other capacitors are disc ceramic. D1 and D2 are 1N34A or 1N914 diodes. BT1 is a 9-V transistor-radio battery. Resistors are 1/4-W carbon film or composition. A phono jack is used at J1 and S1 is a miniature SPDT toggle or slide switch.

A Pierce oscillator is used at Q1 in Fig 5-4. C1 is adjusted for zero beat with WWV. C2 and C3 are feedback capacitors. The indicated value may need to be changed if your crystal fails to oscillate. Values in a range from 560 to 1500 pF are typical.

Q2 amplifies the 100-kHz energy. D1 and D2 generate harmonic energy which makes the 100-kHz markers audible in the upper part of the HF spectrum. A short antenna or clip lead is attached at J1 to permit the marker energy to radiate. You may connect the standard to your receiver directly by using a wire from J1 to the receiver input jack.

A Novel Resonant Antenna Tuner

Certainly, we need to have a method for matching our QRP transmitters
to loads that are not 50 ohms. Various names have been assigned to
devices that tune out unwanted reactance that appears at the feed
line to the antenna. Among the names we hear are Transmatch, antenna
matcher, antenna tuner and ATU (antenna tuning unit). In effect, these
devices do not tune an antenna, nor do they correct a mismatch that
may be present at the antenna feed point. But, they do enable the
transmitter to look into a 50-ohm termination when we use them between
the transmitter and the feed line. This is a healthy situation for
solid state amplifiers, and it does help to ensure maximum power trans-
fer to the feeder. Fig 5-5 shows a practical circuit that I like to
use.
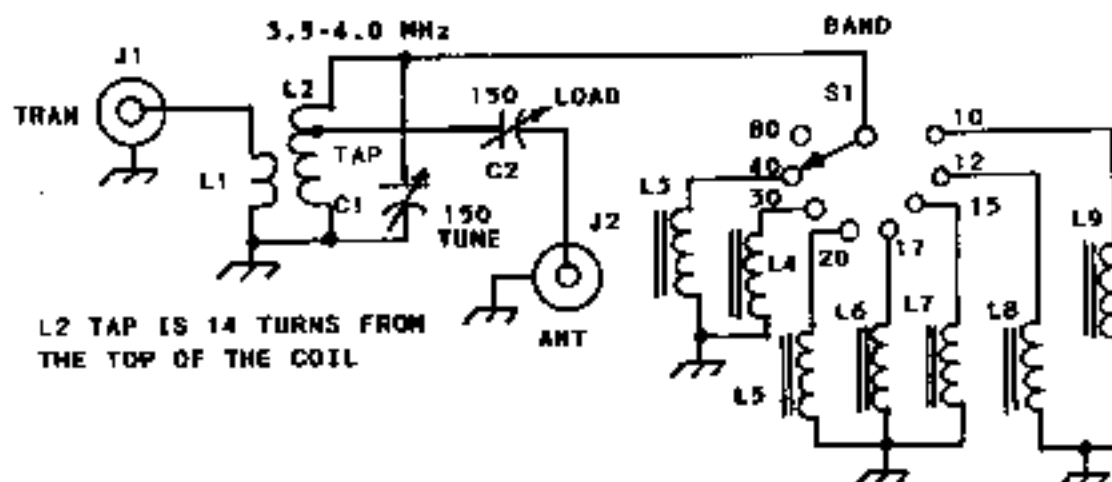
Fig 5-5 -- Circuit for the resonant antenna matcher. C1 and C2 are miniature
140- or 150-pF air variables. C2 requires an insulated shaft coupler. J1 and
J2 are phono jacks or SO-239 connectors (builder's choice). L1 has 5 turns
of no. 24 enam. wire over the grounded end of L2. Use 50 turns of no. 22 enam.
wire, close wound, on a 1-inch OD coil form (25 uH) for L2. Winding length
is 2 inches, requiring a 2-1/2 inch coil form. L3 is an 8.8-uH coil (39 ts.
no. 26 enam. on an Amidon T68-2 toroid). L4 has 3.7 uH of inductance (30 ts.
no. 26 enam. on a T50-6 toroid. For L5 (2.9 uH) use 27 ts. no. 26 enam. on
a T50-6 toroid. L6 is 1.39 uH (18 ts. no. 26 enam. on a T50-6 toroid. L7 has
17 ts. no. 26 enam. (1.2 uH) on a T50-6 core. L8 is 0.9 uH (15 ts. no. 24
enam. on a T50-6 core. L9 (0.7 uH) has 13 ts. no. 24 enam. on a T50-6 toroid.
S1 is a single pole, 8 position phenolic or ceramic wafer switch, nonshort-
ing type.

Only L1 and L2 are used for 80 meters. The tap point and turns ratio
remains effectively the same for each band. The inductance of L2 is
reduced when L3 through L9 are placed in parallel with L2. This does
not degrade the circuit Q. You may use only those shunt coils that
suit the bands you operate. Matching is accomplished by applying RF
power at J1 and connecting the feed line at J2. Use an SWR indicator
between the transmitter and J1. Set the SWR meter for REF power. Now

adjust C1 and C2, alternately, for a low SWR. It should be possible to obtain an SWR of 1:1 or nearly so. There is interaction between C1 and C2, thereby necessitating working back and forth between the two controls.

An insulated shaft coupler for C2 can be fashioned from a piece of 3/8 inch OD dowel rod. Bore a 1/4 inch hole in each end of a 1-1/2 inch length of dowel rod. Drill a no. 58 hole through each end of the coupler and through the tuning capacitor shaft. Pass a stiff piece of wire through the hole and bend the ends of the wire to keep it in place. This will keep the coupler attached to the shaft of the variable capacitor. A piece of 1/4-inch metal rod can be used between the coupler and the panel knob. Affix the coupler to this shaft with a piece of stiff wire. Epoxy glue may be used to hold the metal shafts securely in the wooden coupler.

The leads between the toroid inductors and S1 need to be as short as practicable. Likewise with the lead from the S1 arm to the top of L2. Coils L3 through L9 need not be toroidal. You may use pieces of 5/8 inch PVC tubing as coil forms, provided you use the proper number of coil turns to obtain the inductances listed in Fig 5-5.

This resonant tuner helps attenuate harmonic energy, since it is a band-pass type of tuned circuit. It provides additional front-end selectivity when used with your receiver. The circuit may be tuned to the operating frequency (coarse adjustment) by adjusting C1 and C2 for maximum background noise or signal strength while listening to the receiver audio output. A version of the Fig 5-5 matcher that uses coil forms rather than toroids, and with a built-in SWR indicator, is presented in ARRL's **QRP Classics**, page 228.

A 12-V Regulated Power Supply for QRPers

Special attention needs to be paid to the design of ac-operated dc supplies for QRP equipment -- especially for DC receivers that are prone to suffering from common-mode hum. Energy from the DC receiver local oscillator enters the power supply and becomes modulated with 120-Hz hum, and is reradiated into the receiver front end at the operating frequency. The problem becomes more pronounced as the operating frequency is increased. This problem is generally the worst above 7 MHz.

Certain measures become necessary in order to prevent this from taking place when an ac-operated power supply is used. First, a quality earth ground should be attached to the ground bus or cabinet of the power supply. The problem is aided also if you avoid end-fed wire antennas that are 1/2 wavelength (or multiples thereof) long. The end of the antenna or feed line should be at low impedance when it is near the operating position.

Various things can be done to the power supply to prevent common-mode hum problems. Fig 5-6 contains the circuit for a practical power supply that is designed specifically to deal with the malady. Note that the primary of T1 is bypassed to ground. Each rectifier diode is shunted by a 0.01-uF ceramic capacitor. RFC1 keeps RF energy from entering

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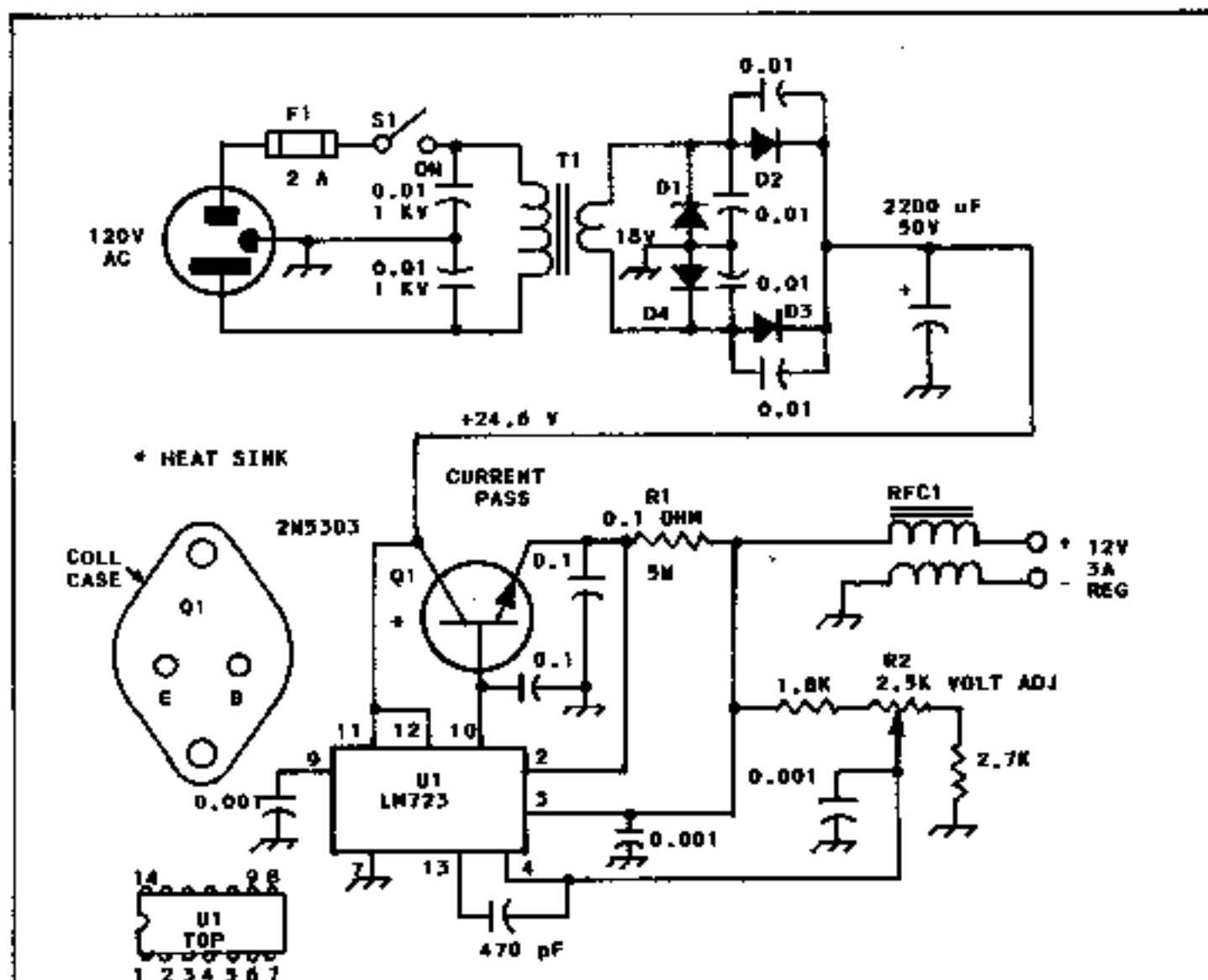
Fig 5-6 -- Practical circuit for a +12-V, 3-A regulated and hum-suppressed QRP power supply. Capacitors are disc ceramic except the polarized one, which is electrolytic. D1, D2, D3 and D4 are 6-A, 50- or 100-PRV rectifier diodes. R1 is a 0.1-ohm, 5-W (or greater) current-sense resistor. R2 is a PC-mount trimmer control or panel-mounted larger control, carbon composition. The other resistors are 1/4- or 1/2-watt carbon film or composition. RFC1 is a bifilar-wound toroidal choke. Use 12 turns of no. 18 enam. wire on an Amidon FT-82-43 ferrite toroid (850 mu). S1 is a SPST toggle. U1 is a 14-pin DIP regulator IC. T1 is an 18-V, 4-A transformer or two Mouser Elect. no. 41FJ020 transformers in parallel. A 24-V, 4-A transformer, such as an All-Electronics no. TX-244 may be used.

the power supply. Additional bypassing is used to prevent self-oscillation of Q1 and U1.

Q1 is rated for 10 watts of dc current. This transistor runs quite warm at full load. Therefore, a husky heat sink is necessary to protect the pass transistor. Use an extruded aluminum sink with fins. It should

measure at least 3 X 3 inches with a height of 1 inch or greater. Remember that the case of Q1 is common to the transistor collector. Therefore, the transistor needs to be insulated from the heat sink if the latter component is bolted to a chassis. A mica TO-3 insulator is required if the sink is grounded. A thin layer of heat-sink grease is used between the transistor and the insulator. Likewise between the insulator and the heat sink. Nylon shoulder washers are used to insulate the transistor case from the sink. Alternatively, you may let the transistor be common to the heat sink if the sink is insulated from the chassis. The problem with this approach is that care must be taken to keep from shorting between the heat sink and the chassis with tools, wires, etc. In any event, keep the heat sink outside the chassis or case in order to allow air to reach it.

The Fig 5-6 power supply can deliver in excess of 3 amperes if the duty cycle is not excessive. I detected no sag in output voltage when I attached a 4-ampere resistive load during the test period. The output waveform remained reasonably clean, although the noise observed on the output line did increase slightly. Certainly, the power supply is adequate for most QRP transmitter and receiver needs.

PORTABLE POWER SUPPLIES

Camping, hiking, mountain climbing, biking and boating dictate the use of batteries for powering QRP equipment. Various options are available to us. Batteries are necessary, and we need to recharge them periodically when afield. This requirement is usually met by taking the batteries to a location where ac power is available, recharging them from a car battery or using Mother Nature's product (sunlight) to energize a solar-electric panel.

I built a 100-mA solar panel for use when I'm away from other sources of energy for charging batteries. The cells were purchased singly from a surplus house. I wired them in series (36 cells) with fine wire and mounted them in an 8 X 10 inch wooden picture frame. A thin foam pad was used under the cells to protect them from breakage. I purchased some 1/8-inch thick clear plastic sheeting (UV resistant) and used this material to fashion a window for the picture frame. Glass is likely to break during portable operation, and hence the plastic replacement sheet. I sealed the picture frame with bathtub caulk to prevent dirt and moisture from reaching the solar cells. An aluminum bracket is mounted on the back side of my solar panel. It is equipped with a nut that fits the stud on my small camera tripod. By using the tripod I can place the solar panel at a location where full sunlight impinges on the cells. The panel is tilted skyward and faces the southern hemisphere when I operate in the northern part of the USA.

A photovoltiac panel that contains 36 cells (0.5-V per cell) will deliver +18 volts in full sunlight on a clear day. The panel acts as a trickle charger on dull days. Typically (allowing for haze), the solar panel delivers +14 to +16 volts to my 12-volt NiCd batteries.

I often keep the solar panel attached to my batteries while I operate. A 1-A, 50-PRV rectifier diode is connected between the output of the panel (anode toward the panel) and my battery. This measure is worthwhile in order to prevent the battery from discharging into the panel during darkness. The diode causes a voltage drop of 0.7, owing to its barrier voltage characteristic.

There are a number of small, commercially built solar panels available today, should you not wish to construct your own. Edmund Scientific Co. has a 100-mA, 12-V panel that is compact and reasonably priced. Also, check the automotive supply outlets, such as the J. C. Whitney Co. in Chicago (catalog available) for small solar panels that are used for trickle charging car batteries.

CHAPTER SUMMARY

It should go without saying that there are a number of QRP accessories that are not described in this chapter. I have presented those items that I use for my QRP work. I did not include data about electronic keyers because there are so many of them that one can buy at a reasonable price. I use a Curtis CMOS keyer and a Bencher paddle for all of my portable QRP work. I have, however, built keyers around the Curtis 8044 IC and installed them in my QRP transceiver boxes. A nice circuit for the 8044 chip is presented in the 1991 edition of The ARRL Handbook, page 29-4. Also, a simple CMOS iambic keyer project is described on page 29-1 of the same book.

You will find numerous QRP transmitters, receivers and accessories in the ARRL QRP anthology (QRP Classics). The book contains reprints of QRP articles that have appeared in QST.

CHAPTER 6

TECHNICAL BITS & PIECES

This final book chapter contains last-minute circuits and information that was not available in final form when the earlier chapters were prepared. Included are basic data concerning antennas that offer good performance for QRP field operation. These wire antennas are easy to construct and deploy. We will also examine a superhet receiver mainframe that represents a few steps in performance above the superhet model I covered in the receiving chapter. It is structured somewhat in "universal" form in order to make it adaptable to different kinds of filters, VFOs and BFOs. It provides a foundation for your experiments with receiver circuits.

QRP ANTENNAS for PORTABLE OPERATION

There are few antennas as simple and easy to erect as the popular Inverted-V dipole. This antenna requires only one support pole. The drooped ends may be affixed with nylon cord to tent stakes, logs, tree trunks or large stones. The lower ends of the dipole can be only a few inches above ground, and good performance will result. Ideally, of course, the higher the overall antenna above ground the better its performance.

An inverted V can be a single or multiband antenna. This depends upon your needs and the feed line used. Single-band use may be had by simply feeding the dipole with 50-ohm coaxial cable, any length. The classic inverted V has an enclosed angle, at the apex, of 90 degrees. It adheres rather well to its radiation pattern if the angle is increased up to 110 degrees. With the specified enclosed angle the antenna is essentially omnidirectional and the polarization is vertical. Depending upon the antenna height and the nature of the earth below the antenna, it exhibits a useful radiation angle. Good signal reports may be expected from DX and local stations.

Wire of large cross section is not required for antennas. I have used countless dipoles and end-fed wires that were made from no. 24 and no. 26 enamel wire when operating afield. Since these antennas are temporary, wire strength and longevity are not a major concern. I have used rubber bands or nylon cord for the end insulators. These temporary antennas roll up into a small bundle for easy transport and storage -- another advantage for the QRPer. The principal performance difference between small and large diameter antenna wire is the antenna Q. The larger the conductor size the lower the Q and hence the greater the antenna bandwidth. Normally, this is not a serious concern.

OVERALL DIPOLE LENGTH

$$FT = \frac{468}{f(MHz)}$$

NYLON GUY LINE

90°

RG-58 COAX

TRAN
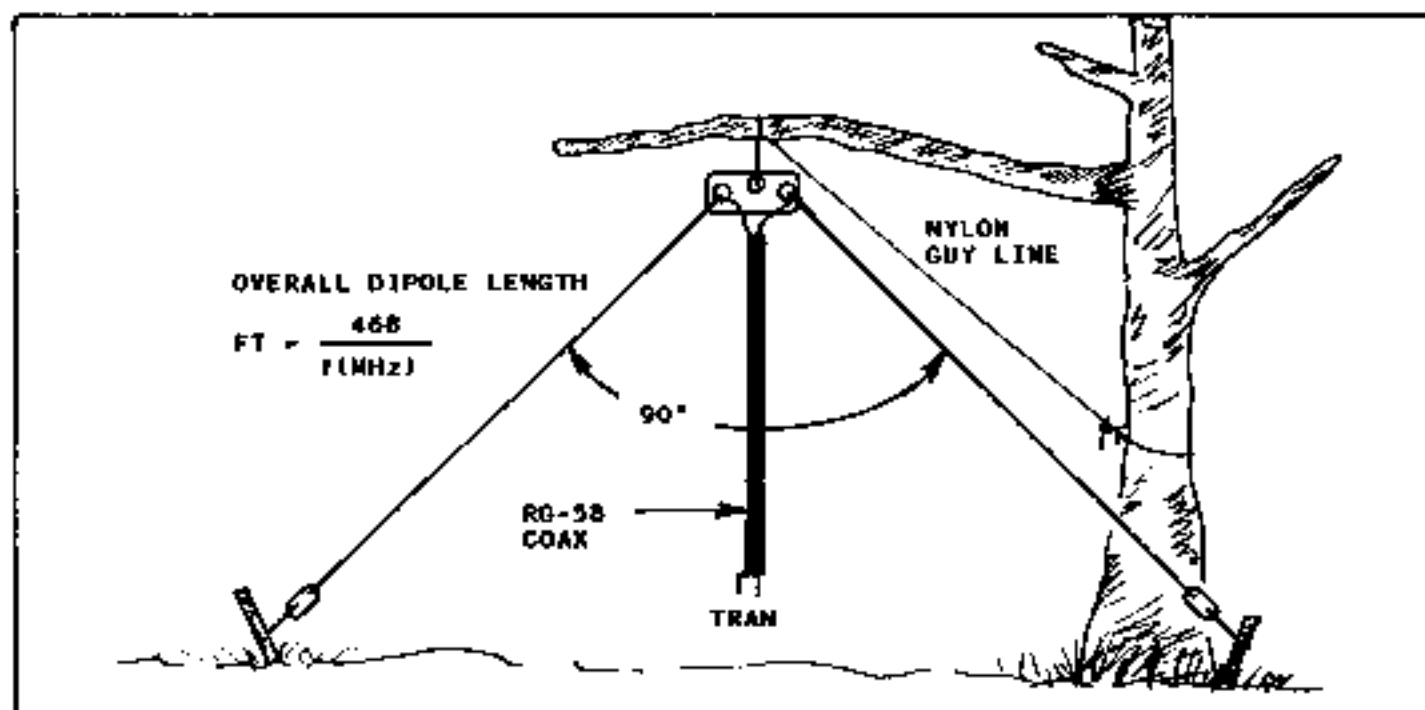
Fig 6-1 -- Inverted V dipole configured for single-band use. A tree limb provides a convenient support for the antenna center. RG-174 miniature coax cable may be used in place of RG-58 if the length is kept under 50 feet (RG-174 is very lossy and not recommended for 14 MHz and higher).

The Fig 6-1 antenna may be converted for multiband use by replacing the coaxial feeder with 300-ohm TV ribbon, adding a 4:1 balun transformer and using a Transmatch between the balun transformer and the transmitter. Open-wire feed line is less lossy than TV ribbon, but it is bulkier and more costly. Clean, dry TV ribbon has approximately the same dB loss per 100 feet as does RG-58.

Multiband Delta Loop

Another high-performance QRP field antenna (home station as well) is the full-wave closed loop. This low-Q antenna is not affected by nearby conductive objects as is the dipole antenna. It works well close to ground, but works best at greater height.

Various loop shapes are possible. The greatest loop gain occurs when it is circular. Next comes the square format and this is followed by the delta or triangular shape. The least gain results when the loop is deployed in a rectangular shape.

Loops work very well on their harmonics and they may be erected vertically or horizontally. Single-band operation is had when the antenna is fed through a 1/4-wave coaxial matching transformer (75-ohm coax) by means of 50-ohm coaxial line. The approximate feed impedance of a square or round loop is 115 ohms, thereby dictating the need for a step-down matching section. The velocity factor of the coaxial matching transformer must be taken into account when cutting the transformer cable. For example, a 1/4-wave coaxial transformer for 7 MHz is 23.19 feet long (L = 246/7 times 0.66). The velocity factor of the coax in this example is 0.66, as indicated.
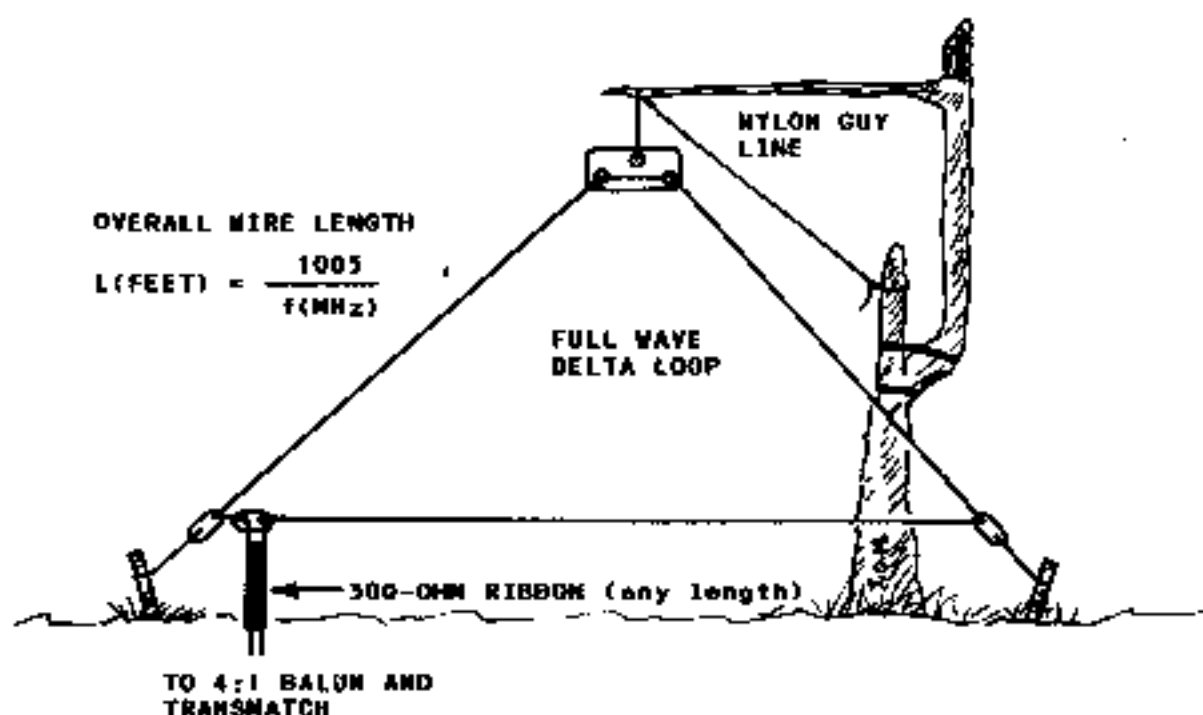
Fig 6-2 -- Example of a delta loop that requires but one support structure. Multiband operation from the antenna fundamental frequency 30 MHz is possible by using tuned 300-ohm feed line, as shown. This antenna works well on frequencies that are not harmonically related to the fundamental frequency, such as 30, 17 and 12 meters. The greater the antenna height the better the DX performance. When fed at the apex rather than at a lower corner the effective height of the antenna is increased. Maximum radiation is broadside to the loop.

A closed loop may be fed at any point along its perimeter, but most amateurs prefer to feed it at the top, lower corner or at the center of the lower wire. Corner feed, as shown in Fig 6-2, provides vertical polarization. Likewise if the loop is fed at the center of either of its sloping sides. Horizontal polarization occurs when the loop is fed at the center of the lower wire.

The Fig 6-2 antenna is ideal for multiband Field Day use. N8HLE and I operated Field Day with one of these antennas in 1985 from a hill on my farm here in MI. Our delta loop was cut for 40 meters. We used it from 40 through 10 meters and worked all of the USA and many DX stations with 5 watts of power on CW. The feeders may be tied together to permit using the antenna as a broadband vertical for the band below the fundamental frequency of the loop. Four 1/4-wave ground radials under the loop, when used in this manner, enhance the performance when the system.

Tree-Supported Vertical

An above-ground vertical ground-plane antenna has appeal for portable operations. Again, only one main support structure is needed. The vertical can be used as a single-band radiator by feeding it with coaxial line, or you may use 300-ohm ribbon line, a balun and a Transmatch to enjoy multiband use.

Although we can employ a shortened trap type of vertical antenna for fixed-station and portable operation, I don't encourage you to use that style of antenna for low-power work. Shortened antennas are less efficient than full-size ones; The QRPer has no decibels to waste! Furthermore, trap antennas and other shortened radiators have a much narrower bandwidth than is true of full-size radiators. Thus, our logical choice is to erect a full-size vertical antenna when this is possible. A ground-plane vertical can be suspended from a tree limb or by means of a nylon guy line that is hung between two trees or man-made structures. An example of a 40-meter vertical, so arranged, is presented in Fig 6-3.

$$L(FT) = \frac{234}{f(MHz)}$$
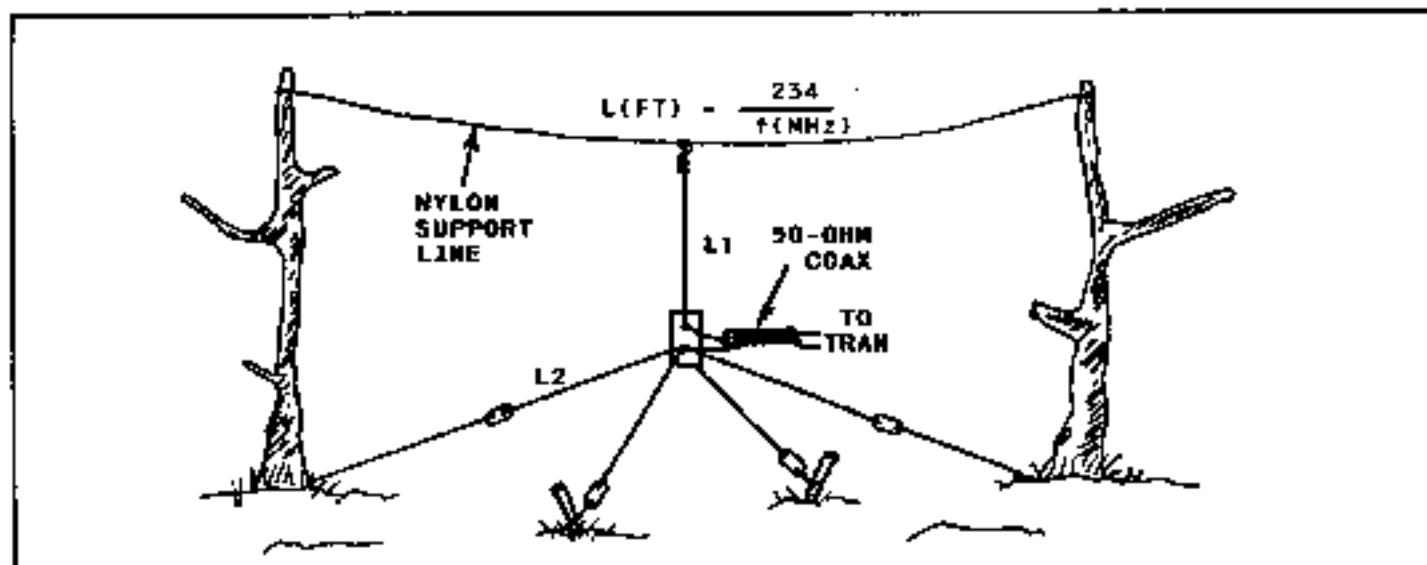
Fig 6-3 -- Example of a method for erecting a wire type of vertical ground-plane antenna. No insulator is needed at the top of L1. A spin-casting fishing rod and reel may be used to place the support line in the tree tops. Alternatively, a bow and arrows can be used for this purpose. The 50-ohm coaxial line may be replaced by 300-ohm TV ribbon or 450-ohm ladder line to permit multi-band use. A balun transformer and a Transmatch are required if balanced feeders are employed. Radials L2 are 5% longer than L1.

A vertical antenna of the type shown in Fig 6-3 is excellent for DX work. It has a low radiation angle and has an omnidirectional radiation pattern. The higher it is above ground the better the performance. A coaxial-cable-fed antenna of this kind may be used on its third harmonic with no changes. Hence, a 40-meter ground plane will work well on 15 meters also. The wire used for L1 and L2 need only be of sufficient diameter to support the weight of the system.

Using a Full-Wave Loop on the Next Lower Band

You may be wondering about the performance of, for example, a 40-meter loop on 80 meters. If the loop remains closed it will be a miserable performer, and typically 3 or more S units inferior to its efficiency on 40 meters. It is, of course, better than no antenna at all, but by no means a world-beater. There is a solution to this problem, and the cure is simple and inexpensive. You can install an 80-meter trap electrically opposite the antenna feed point. This causes the loop

to operate as an open circuit on 80 meters, but at 7 MHz and higher it remains a closed loop. With the 80-meter parallel-tuned trap in place the antenna functions as a half-wave dipole with its ends bent back toward one another. The feed impedance remains low. There is some cancellation with the dipole ends so close together, but performance is otherwise good.

The reactance of the trap should be on the order of 400-450 ohms. Therefore, an 80-meter trap requires 100 pF is a XC of 430 ohms is used for 3.7 MHz. The trap coil therefore needs to have 18.5 uH of inductance to cause resonance at 3.7 MHz. The trap may be checked for resonance by means of a dip meter before it is installed in the antenna. You can make a trap of this type for 80-meter QRP work by using a 100-pF silver mica capacitor in parallel with a T68-2 toroid that has 57 turns of no. 28 enam. wire. A suitable 40-meter (7.1 MHz) trap can be made with a 56-pF silver mica capacitor and a 9-uH inductor. Use 39 turns of no. 26 enam. wire on a T68-2 toroid. The toroid should be coated with low-loss sealant (such as Q Dope or polyurethane lacquer) after it is wound. This will keep the coil turns from shifting and causing changes of the inductance. These traps are adequate for power levels up to 20 watts. Large transmitting capacitors and heavy-gauge wire should be used for QRO operation.

How to Build QRP Balun Transformers

If we were to use a high-power commercial balun transformer with a QRP system it would be the same as "the tail that wagged the dog." Not only are store-bought balun transformers bulky and heavy, they are expensive. We QRPers can build our own low-power baluns quickly and cheaply. Fig 6-4 shows how to wind a 1:1 and a 4:1 style of balun transformer. In-depth data on the design and construction of baluns and other broadband transformers is available in the W2FMI book, **Transmission Line Transformers**, which is available from The ARRL. Also see the W1FB book by Prentice-Hall, Inc. entitled **Ferromagnetic Core Design & Appplications Handbook.**
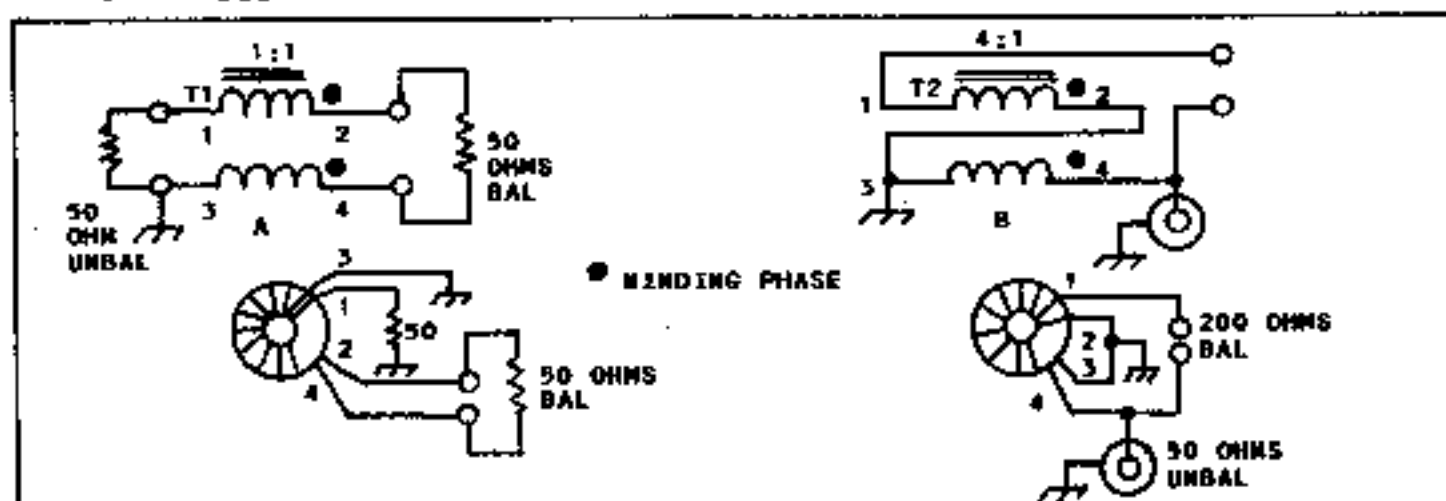
Fig 6-4 -- Circuits for 1:1 (A) and 4:1 (B) baluns. T1 and T2 are Amidon FT-50-43 ferrite toroids (850 mu) wound with 12 bifilar turns of no. 26 enam. wire. Core diameter is 0.5 inch. Transformers suitable for 0-25 watts, 160-10 meters.

The baluns in Fig 6-4 should be protected by coating them with coil dope or polyurethane varnish. They may be weatherproofed by dipping the completed transformer in plastic tool-handle compound. I have used this material as a coating for antenna traps and found it to remain intact for more than five years out of doors.

High-power baluns must be treated differently than those described in Fig 6-4. Larger cores are necessary and the core should be covered with insulating tape, such as 3M glass tape, before the windings are added. Teflon-insulated wire is best for use when winding QRO baluns. These precautions are necessary because of the high RF voltages that can develop from high-power transmitters.

Calculating the Effective Capacitance in VFOs

In order for us to determine the inductance required in the tuned circuit of a VFO we need to know the effective maximum and minimum capacitance of the circuits. The calculations are different for parallel or series tuned VFOs. Fig 6-5 contains this information, respective to a parallel-tuned Colpitts oscillator that uses a JFET device. A value of 6 pF is included in the equations for the input capacitance of the transistor. This figure is valid for most JFETs and MOSFETs.
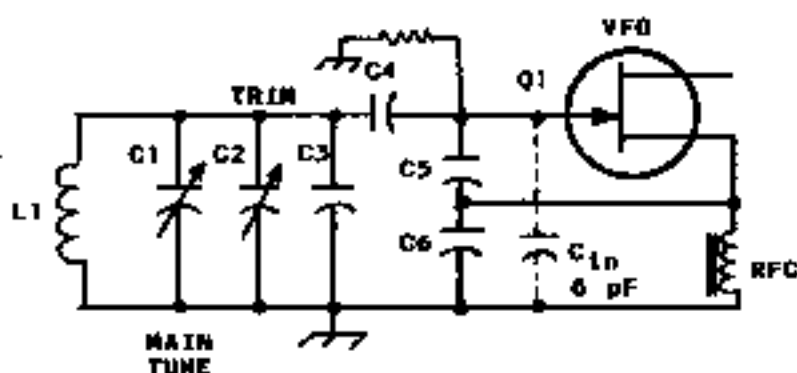
Fig 6-5 -- Method for calculating minimum and maximum effective capacitance for a Colpitts VFO that is set up for parallel tuning. The results provide needed data for determining the required L1 inductance for a given tuning range.

$$\text{Eq 6-1} \qquad C_T = (C1 + C2 + C3) + \cfrac{1}{\left[\cfrac{1}{C4} + \cfrac{1}{\cfrac{1}{\left(\cfrac{1}{C5} + \cfrac{1}{C6} + C_{in}\right)}}\right]} \quad pF$$

A) Assume that trimmer C2 is at midrange.

B) Assume C1 is at max C to find max effective C.

C) Assume C1 is at min C and repeat calculation with Eq 6-1 to find min effective C.

Fig 6-6 -- Method for calculating minimum and maximum effective capacitance for a series-tuned Colpitts VFO. The results provide needed data for determining the required inductance for L1 versus the desired VFO tuning range.

Eq 6-2

$$C_T = \cfrac{1}{\left(\cfrac{1}{C1 + C2 + C3}\right) \cdot A + \cfrac{1}{C4}} \quad pF$$

$$A = \cfrac{1}{\cfrac{1}{\left(\cfrac{1}{C5} + \cfrac{1}{C6}\right)} + C_{in}}$$

A) Assume that trimmer C2 is at midrange.

B) Assume C1 is at max C to find max effective C.

C) Assume C1 is at min C and repeat Eq. 6-2 to find effective min C.

Fig 6-6 shows a series tuned Colpitts VFO and an equation for determining the effective minimum and maximum circuit capacitance. If we use one half the capacitance of the C2 trimmer we then have leeway for calibrating the VFO frequency readout dial. Some stray circuit capacitance always exists. Generally, it amounts to roughly 10 pF.

An Experimental Phasing Type of SSB Generator

Many QRPers like to make their circuits as simple as they can in order to conserve battery power and ensure that the equipment is reasonably compact. The circuit in Fig 6-7 provides the basis for developing a direct-frequency SSB exciter that offers improved performance over a DSSC type of transmitter. The circuit is an adaptation of a two-tube phasing rig that was described by Leo Boisvert, W1HIE, in the February 1961 issue of **The Sidebander**. The original circuit was sent to me by Ken Cornell, W2IMB, with his request that I develop a solid-state version of the unit.

The VFO operates at the transmitter output frequency. It is important to restrict the audio bandpass to the speech range (500-2000 Hz) in order to achieve the best carrier suppression. Measurements at 1000 Hz show the suppression to be 35 dB when the circuit is balanced carefully by means of R1 and R2. The carrier suppression degrades somewhat at the upper and lower limits of the audio range. Careful matching

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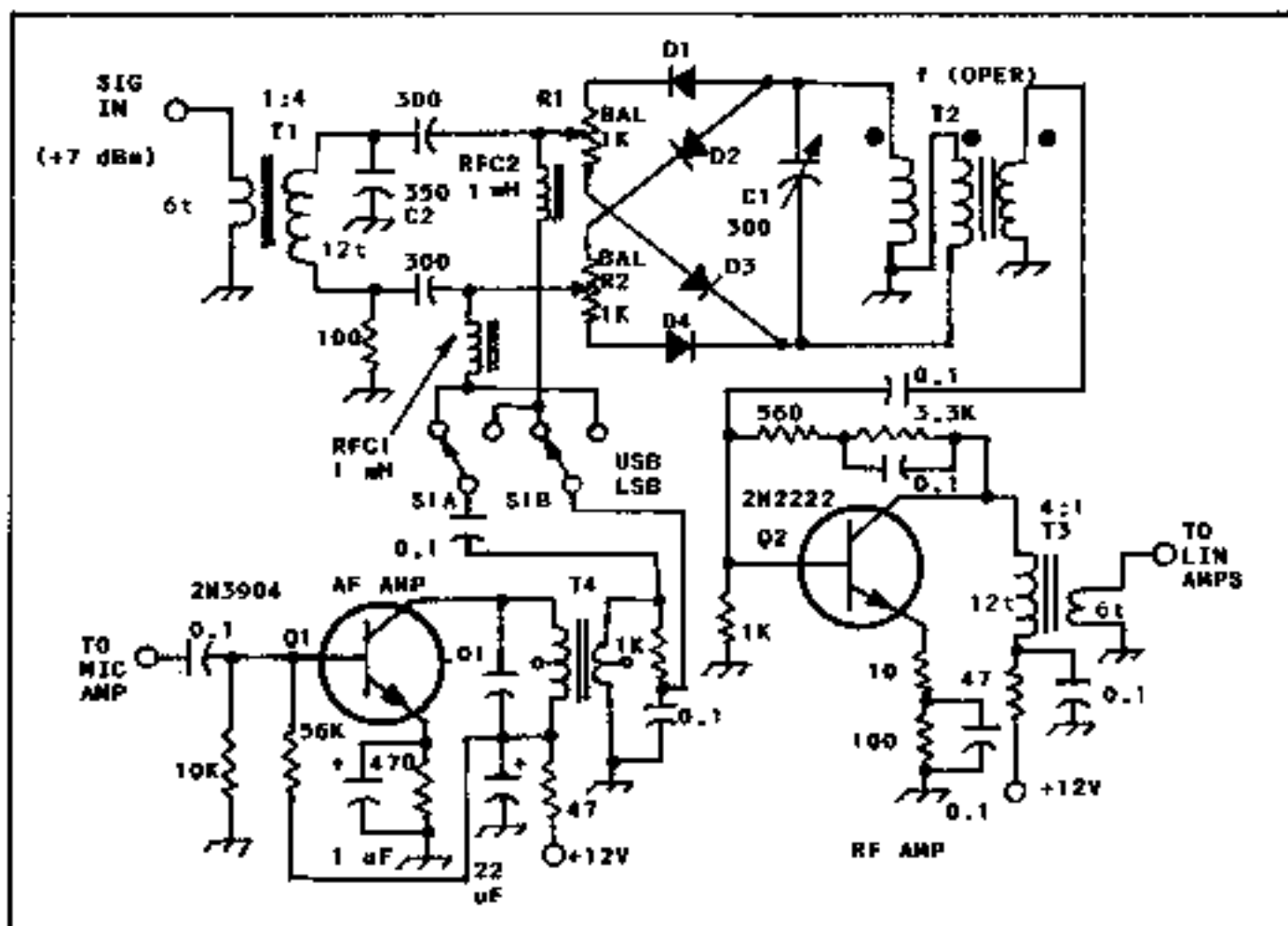
Fig 6-7 -- Schematic diagram of the experimental SSB phasing exciter. Decimal-value capacitors are in uF, Others are in pF. Fixed-value resistors are 1/4-watt carbon film or composition. Polarized capacitors are electrolytic or tantalum. C1 is a 300-pF mica trimmer. D1, D2, D3 and D4 are 1N914 diodes that are matched for forward resistance. S1 is a DPDT toggle. T1 and T3 have 12 primary turns of no. 26 enam. wire on an Amidon FT-37-43 toroid (850 mu). T2 is wound on an Amidon T68-2 powdered-iron toroid and is wound for the operating frequency (trifilar wound) to resonate with C1 at 1/2 or 3/4 its maximum capacitance. T4 is a 10K-ohm to 600-ohm audio transformer (Mouser Elect. part no. 42TL019 or equiv. Center taps not used.) R1 and R2 are PC mount carbon composition controls.

of D1 through D4 is important in order to achieve the best carrier suppression. Physical and electrical symmetry is vital in the area of the balanced modulator. C2 must be changed for the band of operation. The 350-pF value shown is for 75 meters. The XC is 120 ohms.

A class A fed-back linear amplifier is used (Q2) after the balanced modulator. It may be followed by the broadband amplifier strip that is described in Fig 4-18. A linear RF power amplifier can then build the power output to 5 or 10 watts PEP. A mic preamplifier and gain control is required ahead of Q1. C1 and T2 form a tuned circuit that is resonant at the chosen operating frequency.

R1 and R2 are adjusted for minimum carrier bleed-through while observing the S meter. Transmitter tuneup and CW operation is accomplished by unbalancing the circuit with R1 or R2. A low dc voltage can be applied to the junction of D1 and D2 through a 100K-ohm resistor and switch to allow for tuneup and CW operation, thereby enabling the balance controls to remain adjusted for best carrier suppression. There is no PC pattern or circuit board available for this project.

An Advanced Superhet Mainframe

The Fig 6-8 circuit is the big brother to the receiver in Fig 3-34. You may select your own operating and intermediate frequencies. Two IF amplifiers are used for additional gain. AGC is included.

Singly balanced mixer U1 is preceded by a half-wave filter for the band of operation (FL1). It prevents unwanted image responses. A preamp may be added after FL1 for increased sensitivity above 10 MHz.

The IF filter area on the PC board (Fig 6-9) can accommodate almost any IF filter you use. Modify the large conductors with a hobby motor and bit to remove unwanted copper. R3, R5 and C12 are chosen to match the impedance of your filter. Keep in mind that U2 presents an input Z of roughly 1K ohms.

An AGC amplifier/rectifier (Q2, D4 and D5) and S meter circuit are included. AGC is audio-derived, but does cause a pop or click when it engages.

You may eliminate FL1 if you use a double- or triple-tuned input circuit in place of T1. An 8-element fixed-tuned bandpass filter may also be substituted for T1, in which case FL1 will not be needed.

The LO injection power is critical. Try to set the level between 2 and 4 volts P-P (at pin 2 of U1) for best performance. Excessive LO voltage causes U1 to generate harmonic currents that create spurious which will be heard across the receiver tuning range. Too little LO injection, on the other hand, reduces the mixer conversion gain.

The LO and BFO circuits are not included on the Fig 6-9B PC board. This is done to minimize the occasion for stray energy getting into parts of the circuit where it could cause spurious responses. House your LO and BFO circuits in separate shield boxes and feed the output energy to the main PC board via RG-174 mini coax. Suitable VFO and BFO circuits are presented earlier in this book. You may wish to save money by using a tunable BFO (crystals are expensive!) of the type shown in Fig 3-34. Tunable BFOs provide the advantage of being able to reduce QRM (somewhat like pass-band tuning) by moving the BFO frequency about on the IF filter response curve.

T2 is tuned to the IF you select. The Z ratio is 3:1. C4 is chosen to provide resonance with C6 at T1. AGC time constant is set for your liking by means of C31 and R24. Increasing the value of these parts lengthens the AGC decay time.

Q4 is a muting switch. R27 must be grounded to activate the receiver.
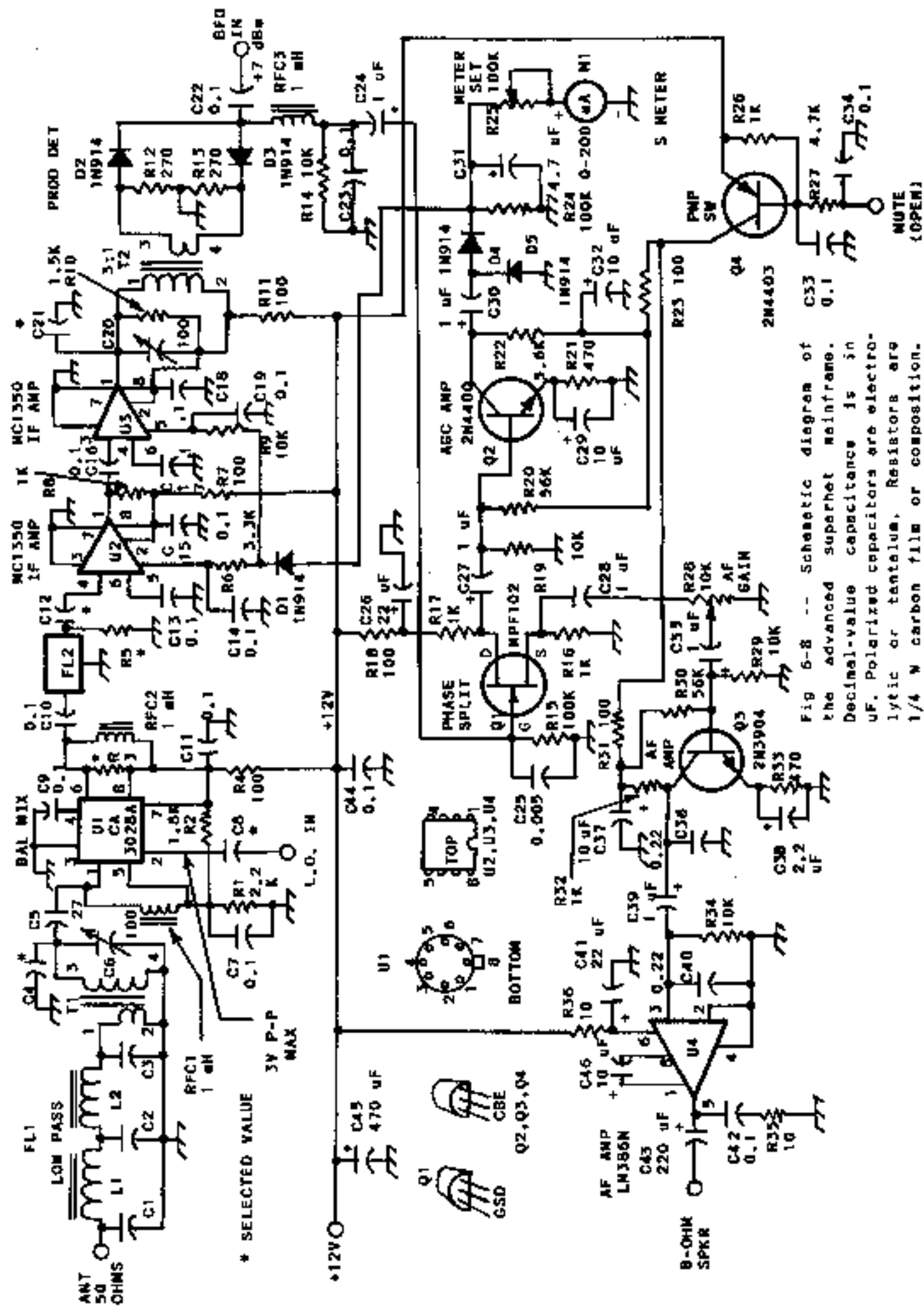
Fig 6-8 -- Schematic diagram of the advanced superhet mainframe. Decimal-value capacitance is in uF. Polarized capacitors are electrolytic or tantalum. Resistors are 1/4 W carbon film or composition.

Refer to the parts description in Fig 3-34 for most of the components in the Fig 6-8 receiver. PC artwork for this project appears below. A 0.1-uV signal is plainly audible at the receiver output for the Fig 6-8 circuit when it is set up for 75-meter operation.
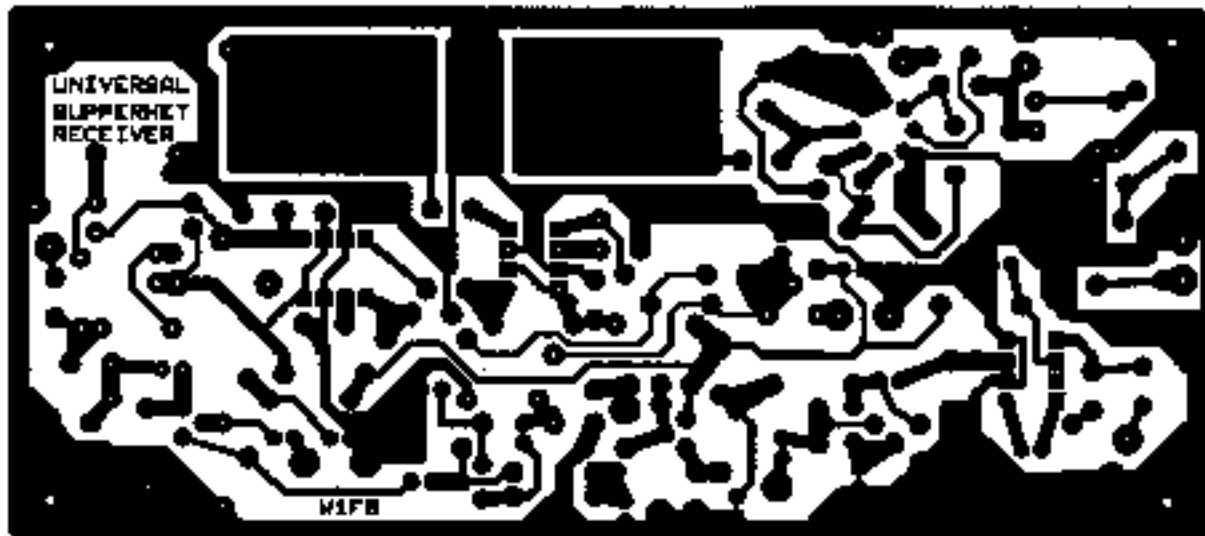
A
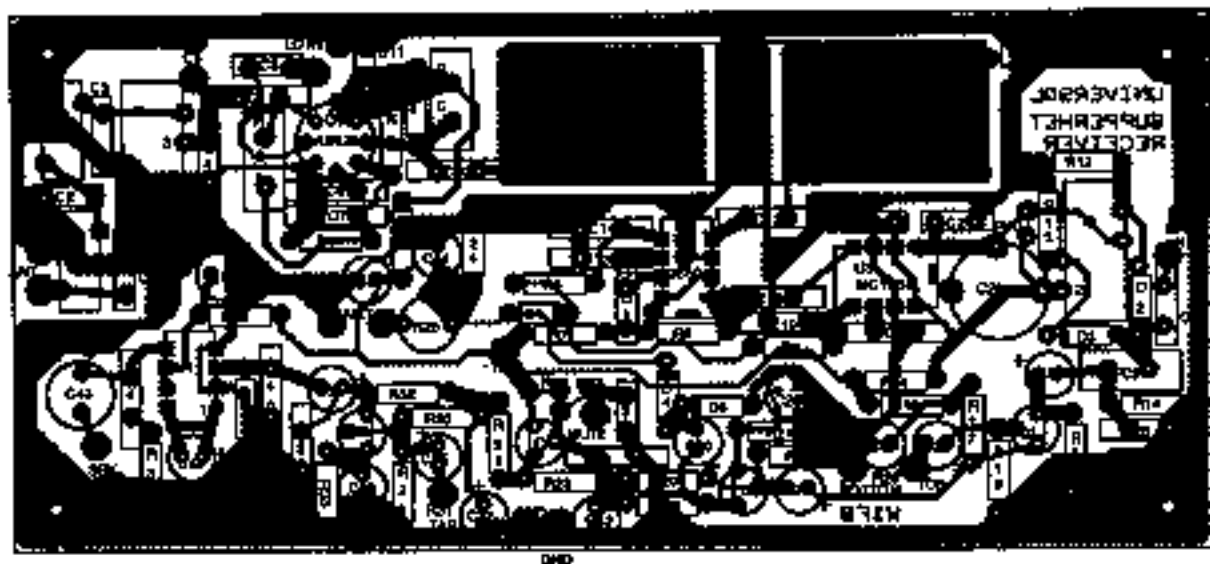
B

NOTE: C46 IS EXT. TO THE PC BOARD

Fig 6-9 -- Scale etching template as seen from the etched side of the PC board (A). Drawing B is an Xray view of the component side of the PC board. PC boards for this receiver are available from FAR Circuits for $12.50 plus $1.50 shipping. Boards are plated and drilled.

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Simple Ladder filters for CW and SSB

Surplus computer crystals come in a wide variety of frequencies that make them suitable for home-made ladder filters. Complete design data for ladder filters, plus a crystal evaluator circuit are found in the concluding section of **W1FB's Design Notebook**. The **QST** article reprints are by W7ZOI and W1FB, and step-by-step instructions are given for determining the Q and series resistance of crystals. This information is used in the filter-design equations that are included.

Fig 6-10 shows two ladder filters I designed for use in a home-made receiver. The crystals were obtained from JAN Crystals, 2341 Crystal Drive, P.O. Box 06017, Fort Myers, FL 33906 (catalog avail.). These crystals are in HC-6/U metal holders. I measured the unloaded Q at 35,683 and the Rs (series resistance) as 68 ohms. The design of the two filters is based on this information. The R(end), which is the characteristic impedance of the filter, is 508 ohms for the CW filter and 4.7K ohms for the SSB filter. The filter input and output broadband transformers, T1 and T2, are wound to interface with the circuit impedances that terminate the filters (mixer and post-mixer or IF amplifier).

I chose 2750 kHz as the filter center frequency in order to avoid having the BFO harmonics fall in the HF bands. BFO harmonics appear as loud unmodulated carriers in the receiver tuning range. Keep this fact in mind when you choose an intermediate frequency. Multiply the BFO frequency up to 11 times to learn where the harmonics fall.
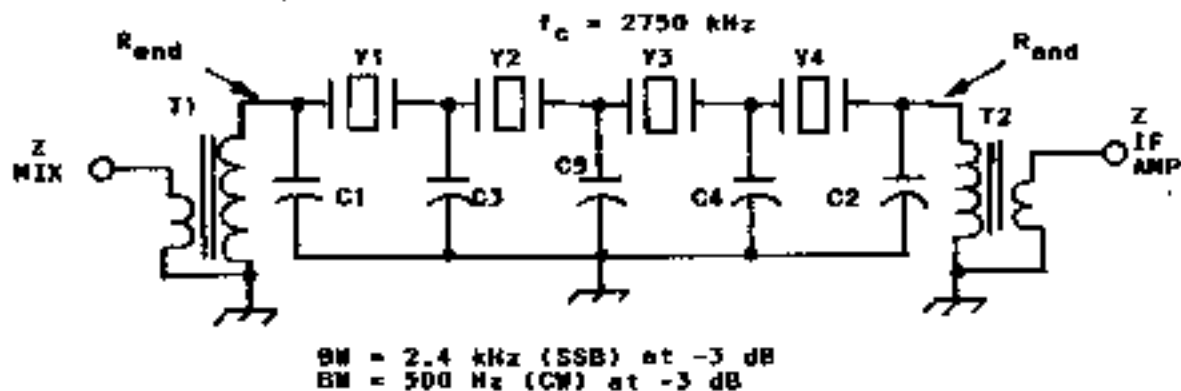
Fig 6-10 -- Circuit for the 4-pole crystal ladder filter. Capacitors are close-tolerance silver mica. T1 and T2 are broadband matching transformers. Main winding has 12 turns of no. 28 enam. wire on an Amidon FT-37-43 ferrite toroid (850 mu). The smaller winding contains the appropriate number of turns to match the input and output loads to the filter R(end). For CW, C1 and C2 are 47 pF, C3 and C4 are 150 pF and C5 is 300 pF. R(end) = 508 ohms. For SSB, C1 and C2 are 6 pF, C3 and C4 are 22 pF and C5 is 51 pF. R(end) is 4.7K ohms. Y1-Y4, incl., are 30 pF load capacitance, matched units (2750 kHz) from Jan Crystals. BFO frequency for CW is 2750.7 or 2749.3 kHz. For upper and lower SSB use BFO frequencies of 2751.3 and 2748.7 kHz. SSB BFO frequencies should fall approximately 20 dB down from peak filter response on the filter curve. Other crystal frequencies are suitable upon determining the crystal Q and series R, then obtaining the parameters from the equations in the referenced book.

Eliminating BC-Band Interference

If you live near a commercial AM broadcast station it is likely that
the strong signal from that station will overload your receiver and
cause spurious responses and desensing to occur. This is especially
true when the receiver front end has limited selectivity, such as
with the superhet receivers in Figs 3-34 and 6-8. A simple solution
is to add a 50-ohm high-pass filter at the input to the receiver.
The filter cutoff frequency should be on the order of 1.7 MHz. Fig
6-11 shows two such circuits. The circuit at A was designed by W7ZOI
and is a 5-pole Chebychev with an fco of 3 MHz. The ripple factor
is 2 dB. This filter was designed for use at 3.5 MHz and higher.
The filter in Fig 6-11B has an fco below the 160-meter band and was
designed by W4ZCB for use with his version of the Fig 3-34 receiver.
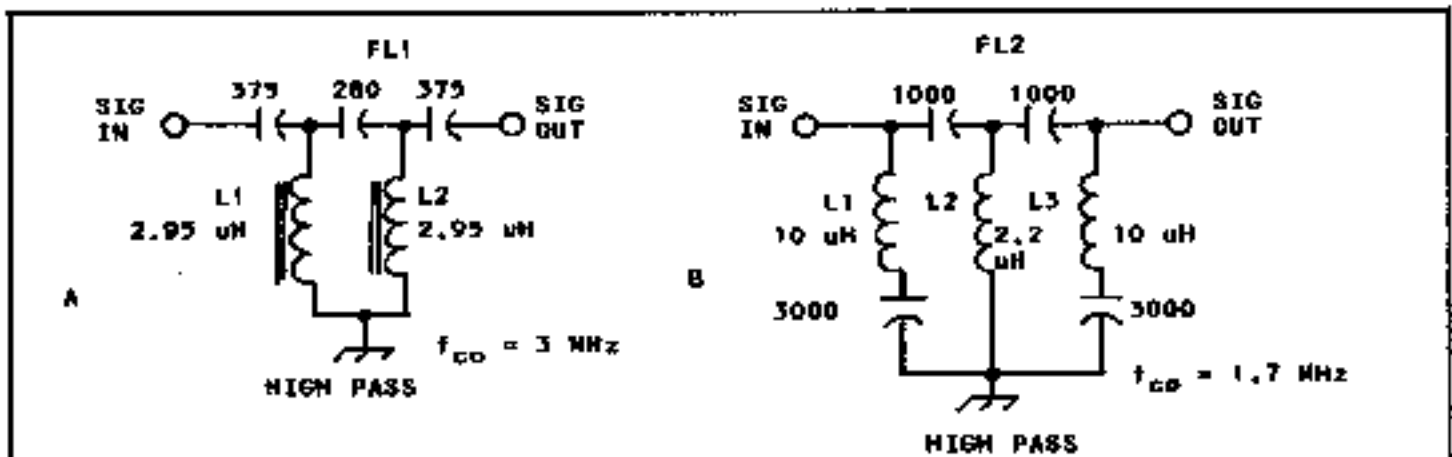
Fig 6-11 -- Practical circuits for two 50-ohm high-pass filters for suppress-
ing commercial AM broadcast band energy at the receiver front end. Circuit
A was developed by W7ZOI. Capacitance is in pF. Silver-mica capacitors are
used in both circuits. L1 and L2 at A are toroidal inductors. Use 31 turns
of no. 30 enam. wire on an Amidon T37-6 toroid core. L1, L2 and L3 at B are
miniature RF chokes. Circuit B was designed by W4ZCB.

Insertion loss for the filters in Fig 6-11 is minimal. It is less than
1 dB in both examples. Correct filter performance, as is true of any
filter, depends upon proper termination at each end of the filter.
The above filters are designed for an impedance of 50 ohms. FL1 and
FL2 in Fig 6-11 should be laid out in a straight line with good separat-
ion between the input and output ports of the filters. The inductors
need to be at least an inch apart to minimize mutual coupling. This
is especially pertinent to FL2. A metal shield between each of the
FL2 coils would further aid the isolation.

Half-Wave Harmonic Filters with Standard Values of C

Tables 6-1 and 6-2 list the inductance values for 5- and 7-element
low-pass filters that are suitable for use at the output of QRP trans-
mitters. These filters were developed by Ed Wetherhold, W3NQN, around
capacitors of standard values. Silver-mica capacitors should be used,
although close-tolerance polystyrene capacitors are also suitable.

TABLE 6-1
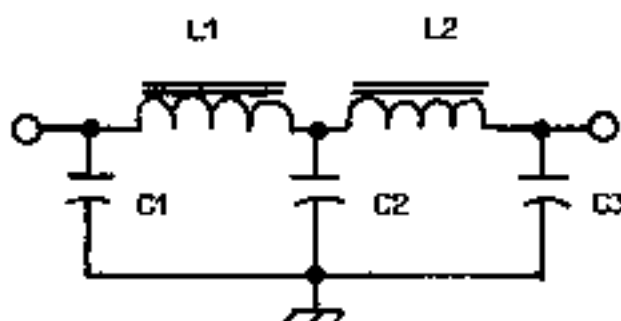
L1   L2

BAND (m)	C1, C3 (pF)	C2 (pF)	L1, L2 (uH)	f_{co} (MHz)
80	560	1200	2.5	4.056
40	240	560	1.12	7.818
30	220	470	0.98	10.39
20	150	330	0.678	15.00
15	100	220	0.462	22.00

Component values for a 5-element half-wave low-pass filter. Inductors are wound on no. 6 powdered-iron toroids (T37-6 for up to 5 watts and T50-6 for for up to 25 watts).

TABLE 6-2

L1   L2   L3
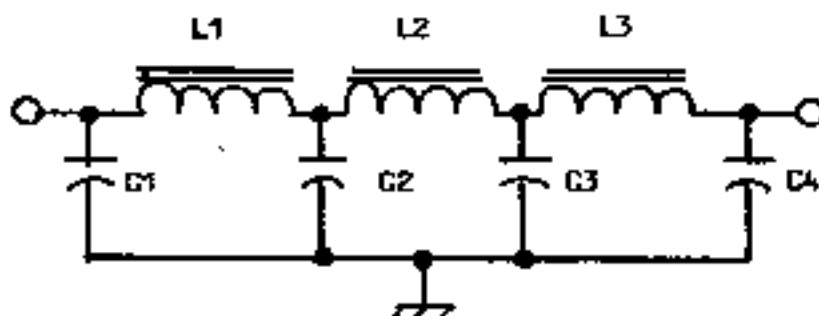
Chebyshev

BAND (m)	C1, C4 (pF)	C2, C3 (pF)	L1, L3 (uH)	L2 (uH)	f_{co} (MHz)
80	510	1300	2.637	3.261	3.81
40	330	750	1.508	1.789	7.23
30	180	470	0.952	1.188	10.33
20	180	390	0.773	0.904	14.40
15	130	270	0.526	0.606	21.48
10	82	180	0.359	0.421	30.90

Component values for a 7-element low-pass harmonic filter. Inductors are wound on no. 6 powdered-iron toroids. Use standard equation for finding the required number of coil turns.

Practical Power FET RF Amplifier

Fig 6-12 illustrates a practical circuit for an RF power amplifier
that can deliver up to 20 watts of output power. It uses a switching
type of TMOS power FET. The device is a low-cost Motorola MTP3055E.
This transistor has a TO-220AB type of plastic case. Maximum VDS is
+60 and maximum continuous drain current is 12 A dc. The feature that
makes it a good device for RF service is the low RDS (drain-source
resistance), which is rated at 0.15 ohm. Input capacitance (Ciss) is
500 pF and output capacitance (Coss) is 300 pF. Transconductance (gfs)
is 4 mhos.
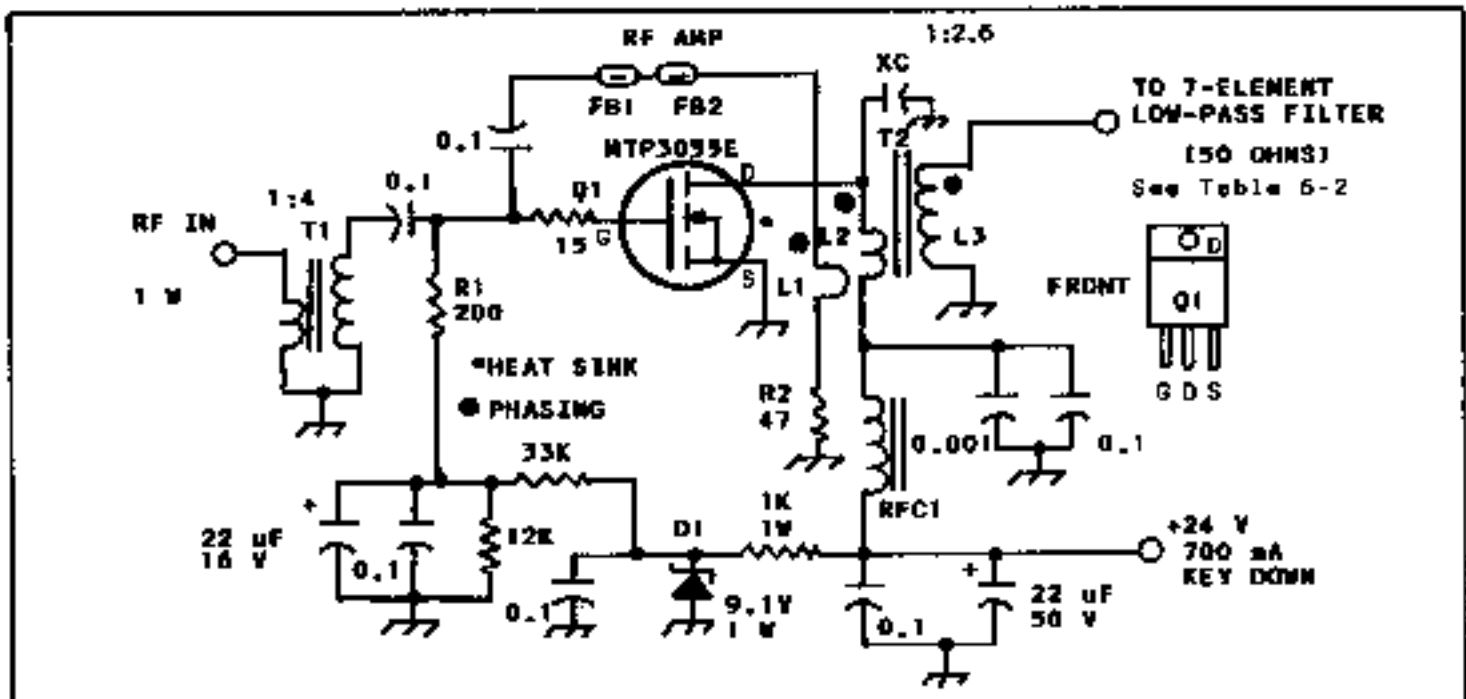
Fig 6-12 -- Schematic diagram of a practical linear amplifier that uses a power
FET. Capacitors are disc ceramic, 100 V or greater, except those with polarity
marked, which are tantalum or electrolytic. Resistors, unless otherwise noted,
are 1/4- or 1/2-W carbon film or composition. Capacitor XC has a reactance
of 150 ohms and is selected for the operating band. For 40 meters use 150 pF
for XC, etc. FB1 and FB2 are miniature Amidon ferrite beads, 850 mu. RFC1 has
10 turns of no. 24 enam. wire on an Amidon FT-50-43 ferrite toroid. T1 consists
of 12 secondary turns of no. 26 enam. wire on an FT-37-43 ferrite toroid. Use
6 primary turns of no. 26 wire. T2 has 1 turn of no. 26 enam. wire for L1.
L2 has 4 turns of no. 24 enam. wire and L3 has 7 turns of no. 26 enam wire.
These windings are looped through an Amidon BN-43-3312 balun or "binocular"
core.

The Fig 6-12 amplifier operates class B when biased as shown. R1 is
selected to establish a 200-ohm input impedance for the amplifier.
Negative feedback is provided by way of L1, FB1, FB2 and R2. Increase
the resistance of R2 to decrease the feedback or vise-versa. The 15-
ohm gate resistor is for VHF parasitic suppression. It should be used
as close to the transistor gate lead as practicable. D1 and its resist-
ive divider provide regulated gate bias for Q1. Idling current is 60
mA. The value of XC is determined by the operating frequency. XC =
150 ohms for this component. It bypasses VHF harmonic currents.

The impedance ratio (1:2.6) for T2 is based on a power output of 15 watts. No SWR protection Zener diode is used from the Q1 drain to ground because the device has a built-in drain-source Zener diode. If you wish to protect the Q1 gate from excess voltage you can bridge a 15-V, 400-mW Zener diode from gate to ground. The cathode of the diode would be connected to the gate of Q1 at the junction of R1 and the 15-ohm gate resistor.

The heat sink for Q1 needs to be at least three inches square and an inch high. An extruded aluminum heat sink with fins is best. It must be insulated from ground unless Q1 is insulated from the heat sink.

You may change the operating class of the amplifier to AB by increasing the resting drain current. Increase the forward gate bias until Q1 draws approximately 100 mA with no signal applied to T1.

There is no PC board pattern or related artwork for this project. I recommend that you use double-sided PC board material. The copper on the component side of the board can then serve as a ground plane to aid stability. It should be an easy matter to lay out a board and remove the unwanted copper with a hobby motor and small cone-shaped abrasive bit.

Although this amplifier can be operated from a 12-volt power supply, the available output power will drop to about 4 watts. Also, it will be necessary to replace the gate-bias resistive divider values to ensure +3 volts or greater at the Q1 gate. Power FETs can be operated class C by eliminating the forward gate bias. The bottom of R1 in Fig 6-12 would be grounded for class C operation, and the bias network would be eliminated. Greater RF driving power is needed for class C service. Also, the harmonic output is greater during class C use.

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Notes

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M. F. "Doug" DeMaw, W1FB was licensed in 1950 as W8HHS while he worked as an R&D engineer for the University of Michigan Willow Run Research Center. Later, he worked as an R&D engineer for Ryan Aeronautical Research Laboratories in San Diego. He served also as a TV broadcast engineer at WWTV, channel 9, and as chief engineer/DJ for WATT radio in Cadillac, Michigan.

Doug founded Avtronics, Inc in 1960 at Traverse City, Michigan, where his firm manufactured low-frequency radio-beacon transmitters and alarm receivers for civilian airports. Upon selling Avtronics in 1963 he established Comaire Electronics in Ellsworth, Michigan, where the firm manufactured VHF and UHF amateur equipment, plus low-frequency radio-beacon transmitters and VHF Unicom transceivers. While operating Avtronics, he established *VHFer* magazine, which he turned over to Loren Parks, K7AAD, in 1965 when he joined the ARRL HQ staff as an assistant technical editor.

In 1968 he was promoted to *Handbook* Editor/Lab Supervisor, and in 1970 he succeeded the late George Grammer, W1DF, as Technical Department Manager/Senior Technical Editor.

While employed by the ARRL, he coauthored *Solid State Design for the Radio Amateur* with well-known QRPer Wes Hayward, W7ZOI. W1FB also wrote two books for Prentice-Hall, Inc and one book for Howard Sams.

W1FB retired early from the ARRL staff in 1983 to return to his native state of Michigan. Since retiring, he has written articles for *QST* and has done article and book editing for the ARRL.

Doug is a Life Member of the ARRL, a past Senior Member of the IEEE and past president of the Michigan QCWA, Chapter 10. He holds a Radiotelephone First Class license and an Amateur Extra Class license. He is presently Chairman of the Lake County Board of Commissioners in Michigan and is a member of the Michigan Association of Counties.