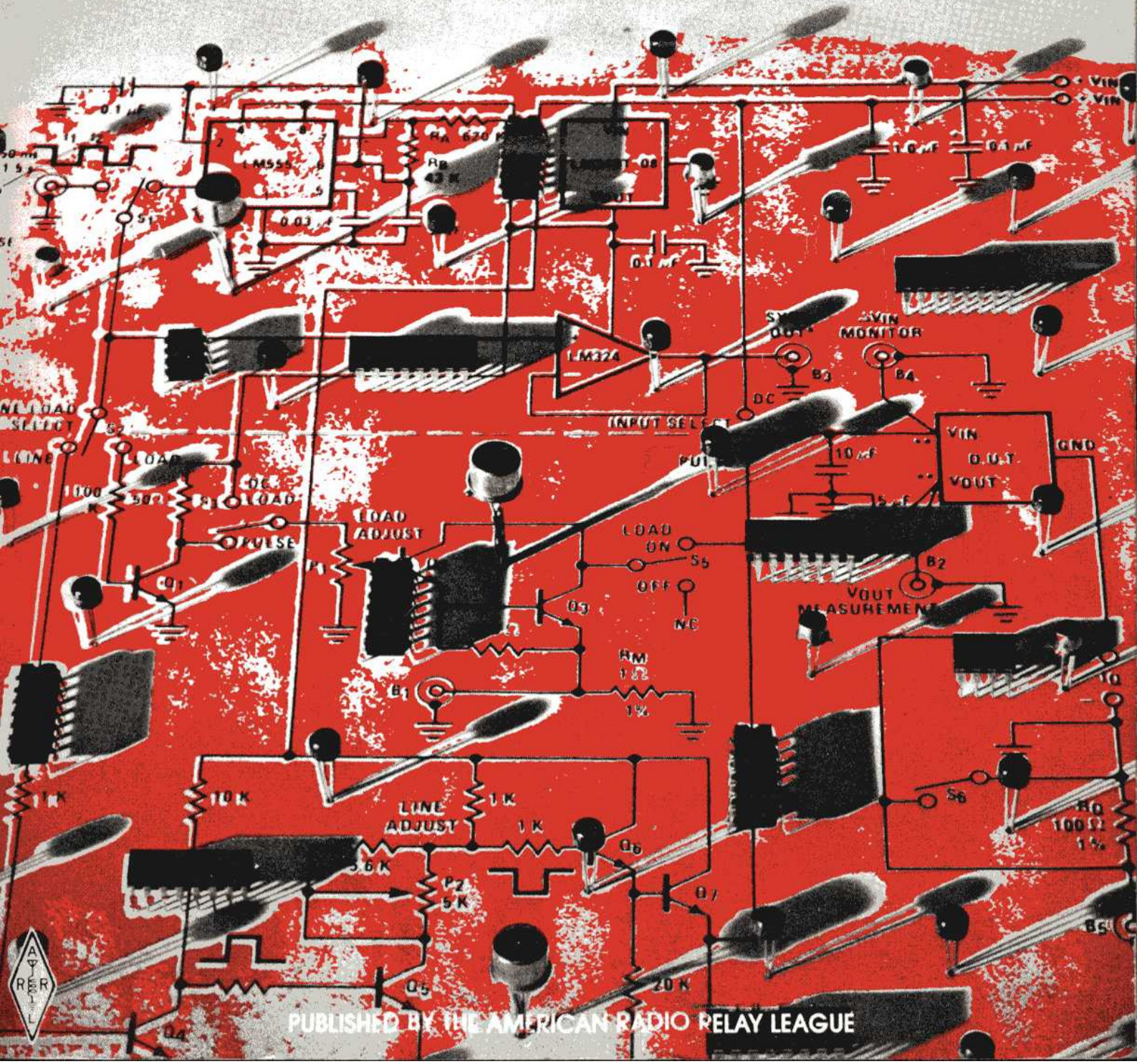
Solid State Design

FOR THE RADIO AMATEUR



PUBLISHED BY THE AMERICAN RADIO RELAY LEAGUE

Solid State Design for the Radio Amateur

By Wes Hayward, W7ZOI
and
Doug DeMaw, W1FB



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Second Printing

Foreword

This book was first released in 1977 as a theoretical and practical guide for the radio amateur interested in using solid-state devices in RF design work. It gained a large, immediate following not only among amateurs, but among professional RF designers as well.

In this second printing, the occasional errors and omissions which inevitably creep into a work of this magnitude have been corrected, making the publication even more valuable to its intended audience.

It is our hope that this book will provide today's readers with a thorough understanding of a technology which has left its indelible mark on radio-communication.

David Sumner, K1ZZ
Executive Vice President

Acknowledgment

This book not only reflects the recent work of the writers, but also the assistance of others. Without their help the book would not have been easy to prepare. It is impossible to list all of those who contributed, but I would like to mention a few and express my gratitude to them. Assistance in the construction of many of the projects was provided by Terry White (KL7IAK), Jeff Damm (WA7MLH), and Deane Kidd (W7TYR). I am grateful for discussions with members of TERAC (Tektronix Employee's Radio Amateur Club, K7AUO) and for the photography done by Denton Bramwell. Special thanks goes to Mike Metcalf, W7UDM. He not only provided assistance and advice, but offered a number of his designs for our use. Discussions with my professional colleagues in the Communications Division at Tektronix have been helpful and enlightening. Additional thanks go to Linley Gumm (K7HFD), Fred Telewski (WA7TZY), and Larry Lockwood (W7JBY). Mention should be made of the liberal policy at Tektronix which allowed me to use its test equipment and computer facilities to generate data which would not have been available otherwise.

Special recognition is given to my friend and co-author, Doug DeMaw, W1FB. His candid views of my circuits and his tolerance of my forthright reviews of his work have, hopefully, led to designs which reflect sound engineering practice and ease of duplication.

Finally, I would like to express my deep appreciation for the patience and assistance given by my wife, Shon, and our sons, Ron and Roger. Not only did Shon devote several hundred hours of typing time to the project, but she maintained an attitude of understanding and encouragement toward the book. The boys willingly gave up my time that could have been spent with them. They even breadboarded a few of the circuits described!

Wes Hayward, W7ZOI
Beaverton, Oregon

No book of this kind is possible without the good will and assistance of the many people who work in the electronics industry as professional engineers and technicians. In our effort to make this publication useful and informative to the reader it was necessary to consult with numerous key people in the semiconductor manufacturing field. I would like to express my gratitude to the personnel at RCA, Motorola and National Semiconductor Corp. who provided direct consultation for some of my circuits, data sheets, booklets and engineering samples of their various solid-state components.

Without the generosity of Bill Amidon of Amidon Assoc. the circuits which contain ferrite and powdered-iron toroid cores would not have been so numerous. International Crystal Mfg. Co. and John Beanland (G3BVU) of Spectrum International, were responsible for many of the components used in these circuits.

I wish to recognize the contributions of personal time and materials received from several members of the ARRL hq. staff, and finally I want to acknowledge the many hours without compensation that were invested by co-author W7ZOI during tape-letter and telephone exchanges of technical data. His intense motivation to make this an outstanding contribution to the amateur's technical library led to many debates between the authors, and subsequently, a volume which will expand the technical knowledge of the reader.

Doug DeMaw, W1FB (ex-W1CER)
Co-Author

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Chapter 1

Semiconductors and the Amateur

From the start, amateur radio has been a pastime wherein those involved have communicated with one another by means of short waves, and at the offset via long-wave paths. During recent years much of the equipment built by amateurs has been for use at hf, vhf and above. Homemade gear has been assembled for two primary reasons – economics and the need for equipment with specific features or qualities not found in commercially manufactured amateur equipment. A third and important stimulus has been the amateur's quest for knowledge of how circuits operate. Individual creative needs lure still others into the field of design, where the pride of achievement comes from the act of *doing*. Generally speaking, communication is for these fellows a means to an end – not an end in itself. This volume is aimed at those amateurs who are not disposed to sitting in front of store-bought equipment and simply communicating with others who are similarly inspired.

Emphasis is placed here on methods which are currently popular in the amateur community among experimenters and designers. It is beyond the scope and size of this book to offer a complete treatment of solid-state design principles for communications, but in the broader sense the reader is referred to many general texts which treat most of the subjects covered here in somewhat greater depth. For the most part, the topics treated in this publication are those which the authors have been involved with for the past several years while working with semiconductors as amateurs. All of the construction projects illustrated herein have been built, tested and subjected to normal and sometimes stringent on-the-air use. Circuits which are shown schematically, but which do not relate directly to a given construction project, are proven ones, and will provide good performance.

Our present world of solid-state device technology has been a springboard for experimenting amateurs in their development of simple and complex circuits for communications. The vacuum tube moves gradually into the shadows as the semiconductor advances in character and capability. Industrial designers are using transistors and ICs in nearly all applications where they perform as good as or better than tubes, and in small-signal work transistors fill that role handily. Furthermore, the overall efficiency of a solid-state piece of equipment versus that of a comparable unit employing vacuum tubes is markedly greater. Reliability is still another part of the design rationale when using semiconductors. Last, but definitely not least, practical miniaturization when semiconductors are used far surpasses that which can be achieved with tubes. Amateurs have long been aware of the foregoing contrasts in active devices, and have forged ahead with enthusiasm as they designed and built transmitting and receiving equipment for their own use. This volume is intended as a guidepost for those amateurs who have embraced the technology of solid-state circuit design. It is hoped that this primer in circuit design and application will serve as the basis for greater achievement by the reader, and that it will inspire further study and experimentation for many.

Simplicity Versus Complexity

In general, the writers have attempted to emphasize methods which are, at least conceptually, straightforward. Frills have been incorporated only where they might serve specific needs in operating the equipment. In most cases the nonessential circuits can be deleted without causing a degradation in overall utility. Such features as side-tone monitors, break-in delay TR switching, and VOX are among those frills being discussed.

There is a tendency among some amateur experimenters to oversimplify their designs. That approach can lead to a piece of gear which does not function as desired. The equipment might even be plagued with spurious output and distortion. Designs are provided in this book which are clean in operation, and are generally more efficient than some of the most simple circuit configurations; e.g., the one-transistor crystal-controlled transmitter.

Historically, amateurs have viewed the complexity of a piece of gear as being commensurate with the number of active devices in the circuit. For example, the five-tube receiver of the middle 1950s was considered by some a "simple design." Conversely, those 15- and 20-tube multiconversion "superhets" were regarded as complex pieces of station apparatus. Such a point of view is no longer appropriate, for nowadays, the number of active devices has little bearing on the cost or complexity of a particular design. Most modern transistors are relatively inexpensive, as is true of ICs and diodes. One can view the addition of one or a few more solid-state devices to a circuit with the same casual outlook that is taken when adding a resistor or capacitor. Indeed, in many instances the addition of active circuitry may allow the builder to leave out a collection of passive components,

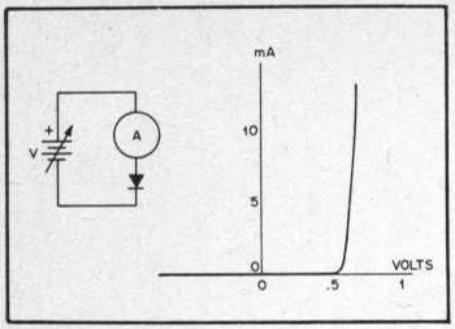


Fig. 1 — Current flow in a diode versus the applied voltage.

thereby enhancing miniaturization, lowering cost and contributing to improved performance. Thus, counting the number of transistors or ICs in a circuit is not a recommended way of judging the simplicity of a circuit.

Another matter of concern to the builder is being able to make the circuit perform correctly after it is built. Quite often a circuit which contains only a small number of components will work just as well as, or better than, a similar circuit which uses many more parts, or even some sophisticated integrated circuits. There is irony in the fact that some simpler circuits will require adjustment by means of sophisticated laboratory equipment in order to effect proper operation, while the seemingly more complex version may function perfectly when power is first applied. Casual observation should not be relied upon in the determination of circuit complexity.

The Design Approach

There are a number of techniques which can be used by the amateur or professional designer when building a piece of equipment. For many amateurs the approach has been purely an empirical one. That is, the circuit must perform a specific function, so the amateur tackles the assignment on an experimental basis. He may peruse the available literature (application notes, data sheets, magazine articles) until he spots a circuit similar to what he has in mind. The circuit will be duplicated, except for subtle changes in component values. Then, measurements may be performed to discover whether or not the circuit functions "as advertised." On the other hand, the professional engineer, if he is worldly wise in his field, will follow a totally different path. From the data sheets he will choose a device which appears to be appropriate for a given application. He will then design a circuit around the component, say, a transistor. He will utilize advanced analytical methods, often based on the availability of a computer. In this manner he will fully understand and establish the circuit performance prior to building it. After the circuit is built in physical form, there is seldom a significant difference between the predicted and actual performance.

The two procedures just discussed are clearly extreme examples. Moreover, in the real world of electronics the two will merge. The more skilled amateur will engage in considerable analysis of his design before starting construction. As a result, he will spend less time to obtain proper circuit operation once the last wire has been soldered in place. In reality, a professional designer is likely to spend a great deal more time experimenting with his circuits than we may suspect, and in particular where rf circuits are concerned. Because of the experimental aspects of such work, amateur radio often serves as an excellent background for professional design efforts.

In this book the authors attempt to approach solid-state design work from the middle ground. There are a number of circuits which can be "lifted" directly for use in amateur applications. Regardless, an attempt is made to provide straightforward mathematical procedures and circuit models, both of which should enable the amateur designer/experimenter to gain a better understanding of the work he is undertaking. It is hoped that the fallout from his design work will assure improved equipment performance.

Basic Transistor Modeling

It is not appropriate now to include a detailed discussion of the solid-state physics which are the basis of transistor operation. The reader is referred to the series by Stoffels which appeared in *QST*, and which is available as a reprint.[1] It will serve as an excellent introductory treatise on the topics that will be highlighted in this book. In this section we will discuss some simplified "models" that can be used in the analysis of many communications circuits.

The term "model" may sound unfamiliar when used in a commentary about electronics, even though we are familiar with the expression in other ways. Certainly, as youngsters most of us have built scaled-down *models* of aircraft, ships or cars. We not only ended up with an attractive replica of the item we were modeling, we learned something about the original after which the model was patterned, and in particular about its structure.

Models are often used in the analysis of electronic circuits for the purpose of describing various components in terms of simpler and more basic circuit components. The junction diode serves as an excellent illustration of this method. A

[1] Reprint available from ARRL for $1.

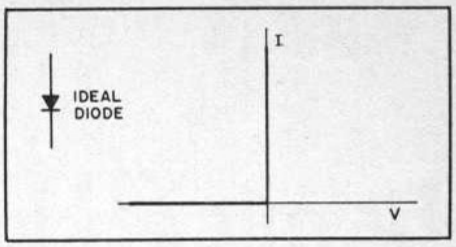


Fig. 2 — Current flow in the "ideal" diode.

physicist would examine a diode with bias provided from a battery and would proceed with a fairly complicated analysis in order to describe the diode operation. First, he would describe the electric fields resulting from the applied voltage. Then he would proceed to calculate the density of electrons and holes within the semiconductor material, the rate at which they are created (from knowledge of the material temperature), how the charges move through the material, and the rate at which they combine with one another. Such calculations would give him a rudimentary knowledge of what is happening inside the diode.

For the physicist or device engineer the preceding calculations (and many more) are significant. Were the circuit designer to go through such an exercise in analysis each time he wished to use a diode, he would be seriously encumbered. His only concern is with the behavior of the device when viewed from its *two external terminals*.

The current flowing in a diode is given by the well-known diode equation

$$I = I_s\,(e^{qV/kT} - 1) \qquad \text{(Eq. 1A)}$$

where I_s is the diode saturation current in amperes, V is the bias voltage across the diode, q is the fundamental electronic charge, k is Boltzman's constant and T is the temperature in degrees Kelvin. For room temperature (about 300 degrees K), the fraction $kT \div q$ has the value of 26 millivolts. A germanium diode might have saturation currents in the neighborhood of 10^{-8} A while a silicon diode would be typified by values closer to 10^{-13} A. This equation is plotted for a typical silicon diode in Fig. 1.

This information can be used directly by the designer, and often it is. However, in many situations much less refined information is sufficient for design purposes.

Fig. 2 illustrates a simplified version of the curve shown in Fig. 1. This shows how the diode has been replaced by an "ideal" diode, the behavior of which

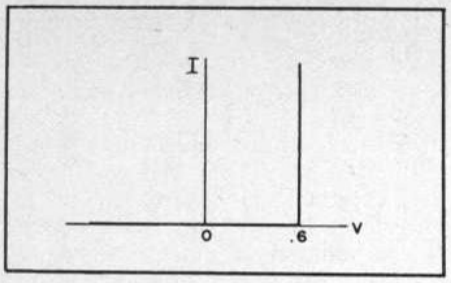


Fig. 3 – Current flow in a perfect diode with offset.

can be described easily. When the diode is *reverse biased*, there is absolutely no flow of current. However, when the diode is *forward biased* (a more positive potential applied to the p- than to the n-material of the diode), the current which flows is determined totally by the circuit external to the diode. The so-called perfect diode is a *model* we can use to describe the conduct of *real* diodes in many circuits. The use of a model leads to simplified analysis. Another diode model is shown in Fig. 3, where a battery has been connected in series with a perfect diode. With a forward bias of approximately 0.6 volt, current will begin to flow, still being limited by the external circuitry. Germanium diodes start to conduct at a somewhat lower applied voltage, in the region of 0.2 to 0.4 volt.

If two silicon diodes are connected back-to-back as shown in Fig. 4, a system behavior would prevail which could be analyzed using the model given. This arrangement provides a three-terminal device which looks strangely familiar. It resembles an npn bipolar transistor! Indeed, if an npn transistor were examined by means of an ohmmeter – connecting only two transistor terminals to the meter at one time – it would appear to be nothing but a pair of back-to-back diodes.

A transistor, conversely, has a property which makes it quite different from a pair of isolated diodes. The characterization can be seen when one of the diodes within it (base-emitter junction) is forward biased while the other (base-collector junction) is reverse biased. Under these conditions current will flow in the collector terminal! This would not occur when using a pair of reverse-connected diodes.

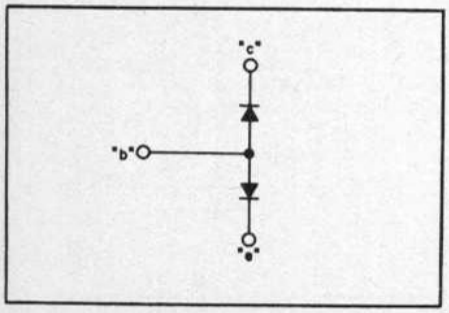


Fig. 4—The basic transistor is formed by back-to-back diodes.

Current flow in the collector is not highly dependent upon the voltage supplied to the collector. It is, however, quite dependent upon the *current* flowing in the base-emitter diode. This parameter is a relatively linear one – the collector current is directly proportional to the base current. The ratio of I_c/I_b is the *beta* of the transistor.

Using the Information

By using the foregoing information, we can construct a simple transistor model (Fig. 5). A new element has been introduced – the *current generator*. It is shown in a circle with an arrow which indicates the direction of current flow. The battery we used with our simplified silicon-diode model has been included in the base leg of the transistor model, for it is significant when describing transistor operation. A battery has been omitted in the collector circuit because the collector-base diode is reverse biased in the typical application. Amplification is implicit in this model, as the current generator in the collector represents not a constant current, but a *dependent current* where the pertinent independent variable is the base current.

The model illustrated in Fig. 5 is not complete for many situations. If we backtrack momentarily to Fig. 1, where a real diode is depicted, it can be seen that the current does not increase infinitely as forward bias is applied. The current increase is sharp and pronounced with increasing voltage, but is finite in nature. This characteristic can be depicted in a transistor model by inserting a resistance in series with the base. The magnitude of this resistance can be given approximately by

$$R_b = \frac{26\beta}{I_{e(dc)}} \qquad \text{(Eq. 1B)}$$

where the dc emitter current is in mA, R_b is the base resistance in ohms, and β (beta) is the current gain introduced above.

A matter of significance which is not covered in Fig. 5 is the frequency effect on transistor gain. It should be noted that at low frequencies beta is constant, with typical values ranging from 10 or 20 to several hundred. However, as the operating frequency is increased in MHz the beta of the transistor tends to decrease. At an ac operating frequency called the f_T of a transistor – sometimes called the gain-bandwidth product – the beta (current gain) is unity, or 1.

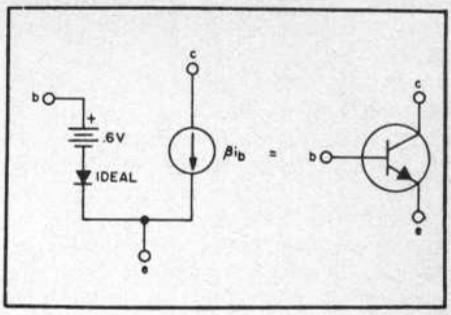


Fig. 5 – Initial transistor model.

At operating frequencies below the effective f_T the current gain is often well approximated by $\beta = f_T \div f_{op}$, where f_T is the gain-bandwidth product and f_{op} is the chosen frequency of operation. For example, a 2N3904 would have an effective beta of 10 at 30 MHz since its f_T is 300 MHz.

Fig. 6 shows a composite transistor model which is suitable for approximate analysis of circuits which employ bipolar transistors at both low and high frequencies. This illustration is highly simplified. Models used by modern circuit designers may contain a dozen or more elements instead of the few depicted in this example. It is not surprising that sophisticated methods lead to amazing accuracy in predicting actual circuit behavior. What *is* spectacular is the fact that for many routine kinds of circuits the simplified model of Fig. 6 will provide surprisingly accurate results – often at very high frequencies.

At low frequencies the beta of a 2N3904 is 100 typically. Hence, if this transistor were biased for an emitter current of 10 mA, the base resistance, R_b, would be 260 ohms.

Biasing of Bipolar Transistors

The simplified model of a transistor presented in Fig. 6 can be used as a tool in the analysis of circuits such as amplifiers and switches. When a transistor is

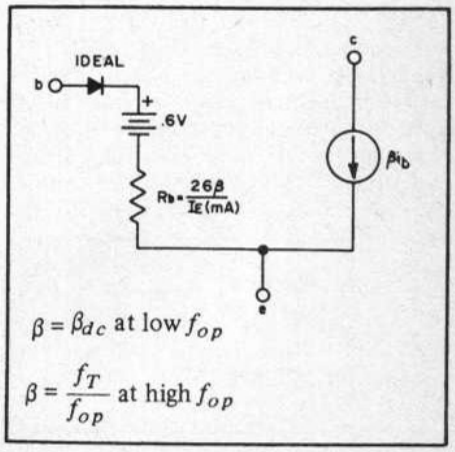


Fig. 6 – Transistor model used for circuit analysis at high and low frequencies.

used as an amplifier, it is usually biased with dc voltages in such a way that the applied ac signals cause the existing (quiescent) dc currents and voltages associated with the transistor to be varied slightly. It is these variations that are usually of interest when an amplifier is built.

In this section various methods for biasing bipolar transistors will be considered. This will serve not only the purpose of reviewing these concepts, but will illustrate how the simple model can be used as a means of circuit analysis.

As an example, a simple audio amplifier will be studied. A likely transistor for this application is the 2N3565 which has an f_T of about 60 MHz and a dc beta of 100. In the example, the amplifier will be biased for a dc collector current of 1 mA with the emitter grounded and the collector at +6 volts. Shown in Fig. 7 is a possible amplifier circuit, a simplified version of the schematic diagram showing only the dc part of the circuit, and finally, the dc portion of the circuit with the simple model substituted for the more conventional transistor symbol.

First of all, since the collector current is to be 1 mA, and the voltage at the collector +6 volts, the value of R_c is determined. In this case, it is given by

$$R_c = \frac{12\text{V} - 6\text{V}}{.001\text{A}} = 6000 \text{ ohms} \qquad \text{(Eq. 2)}$$

Further, knowing that the collector current is 1 mA, the base current to yield this value must be 1 mA/beta = 10 μA. Knowing this value, the net resistance in series with the base can now be determined. The value of R_b was given earlier as

$$R_b = \frac{26\beta}{I_{e\,(\text{mA})}} = 2600 \text{ ohms} \qquad \text{(Eq. 3)}$$

The net resistance in series with the base will be

$$R_{net} = \frac{12\text{V} - 0.6\text{V}}{10^{-5}\text{A}} = \frac{11.4\text{V}}{10^{-5}\text{A}}$$
$$= 1.14 \text{ megohms} \qquad \text{(Eq. 4)}$$

R1 is merely this value less R_b, or 1.137 megohms. In practice the builder would probably take a one-megohm resistor from the parts box for use at R1 with minimal problems being encountered, assuming that the transistor parameters used in the calculation are accurate.

In the real world, the biasing scheme outlined in Fig. 7 will sometimes work, but presents a number of problems. The main deficiency of such a design is that the dc beta of a given transistor type can vary considerably. For the 2N3565 used in the example, a beta of 100 might be typical, but values as high as 300 are frequently encountered. Assuming that the value of beta is 300 and that a one-megohm resistor was used at R1, the base current would be 11.4 μA and the collector current would tend to be $11.4 \times 10^{-6} \times 300 = 3.42$ mA. This much current flowing in the 6000-ohm collector resistor would lead to a voltage drop across the resistor of 20.5 volts, which might suggest that the collector voltage would be negative. This is not possible (because of the ideal diode built into the collector of the transistor model). In reality, the voltage of the collector will drop to zero, or ground, and then go no farther. The collector current will now be determined purely by R_c, and in this case will be 2 mA instead of the 1 mA desired originally. Clearly, with the collector at ground potential, with excess base current keeping it there, the transistor is not going to function well as an amplifier. This condition, where the collector voltage is less than the base voltage, is called *saturation*. The originally analyzed case with the collector voltage larger than that of the base is called the *active* region.

The problems outlined above, which resulted from a beta that was higher than expected, can be circumvented by the use of other circuit configurations or the addition of other components. Shown in Fig. 8 is a variation which is still less than optimum but will at least ensure that the transistor is biased in the active region. Here, the voltage source used to drive the base-bias resistor is the collector of the transistor rather than the 12-volt supply, as originally used. This arrangement has the advantage that negative feedback is applied to the base. That is, if the beta were higher than the desired 100, this would cause the current in the transistor to increase beyond the 1-mA design goal. However, as the collector current increases, a larger *IR* drop occurs across R_c, resulting in decreased collector voltage. This, in turn, decreases the base current, causing the collector voltage to stabilize at some value larger than zero, but still less than the desired 6 volts. The transistor will always be biased in the active region with this scheme.

The reader might find it instructive to assume that the transistor beta is 200 and analyze the circuit of Fig. 8 by using the simple model. The result for this problem is that $V_c = 3.94$ V, $I_c = 1.34$ mA and $I_b = 6.68$ μA. (Hint: The solution of two simultaneous equations is required.)

Shown in Fig. 9 is a circuit which is more typical of the techniques used for biasing transistors in well-designed amplifiers. In this scheme, the base is connected to a voltage divider formed by the 10,000- and 5,000-ohm resistors. A capacitor has been added from the emitter to ground. A capacitor has a characteristic that prevents the voltage impressed across it from changing instantaneously. Hence, for ac signals applied to the amplifier, the emitter

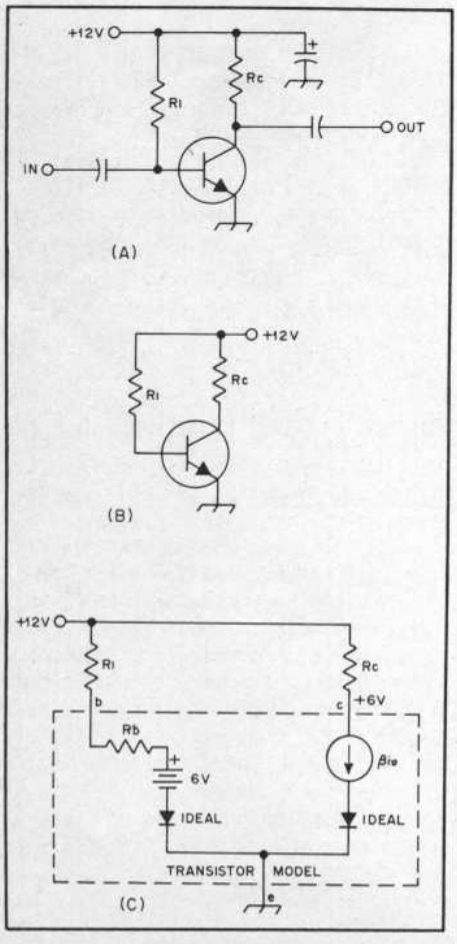


Fig. 7 – Representations for the analysis of a transistorized amplifier.

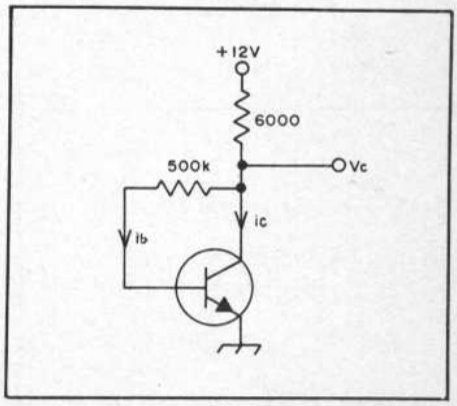


Fig. 8 – Bias arrangement to ensure that the transistor is in the active region.

may be regarded as being at ground potential. However, the dc voltage certainly will not be at ground.

In Fig. 9B the dc part of the circuit has been drawn, omitting the details associated with the ac part of the amplifier. Using classic circuit theory, it may be shown that the voltage divider consisting of R1 and R2 may be replaced with a lower voltage V' in series with a resistance R' where

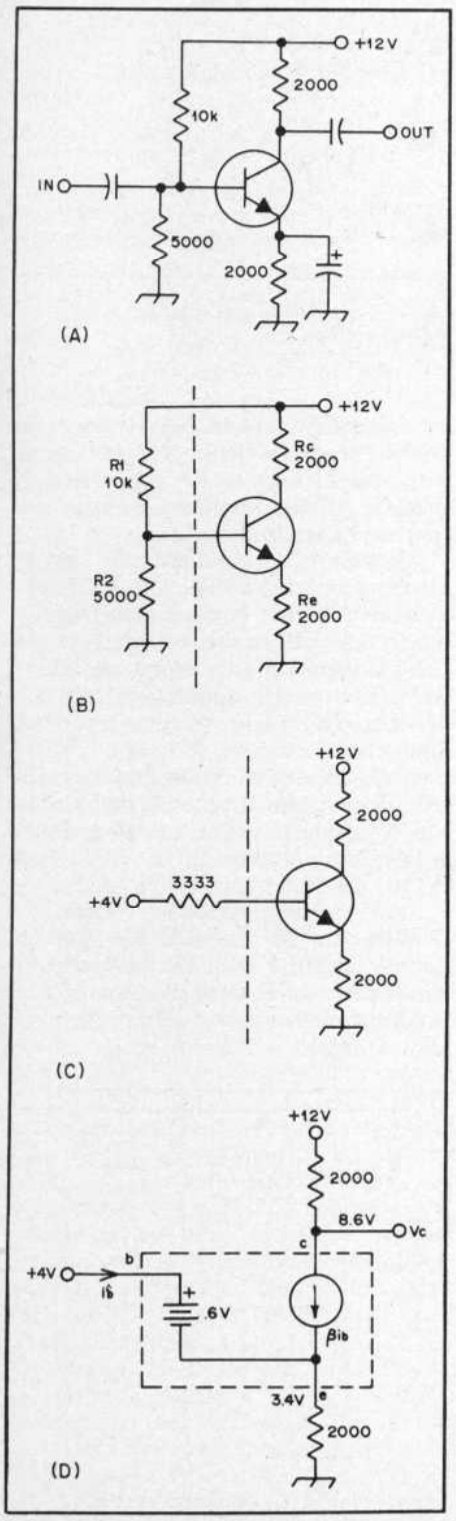


Fig. 9 — Typical bias arrangement for a well-designed amplifier.

$$V' = V_{cc} \times \frac{R2}{R1 + R2} \qquad \text{(Eq. 5)}$$

and R' is the parallel equivalent of R1 and R2. This equivalent circuit is shown in Fig. 9C.

Presented in D of Fig. 9 is a schematic diagram which results when a simplified model of the transistor is substituted in the amplifier circuit. Note here that the model used is even simpler than the one employed earlier, and that the resistance of the base-bias divider, R', has been omitted. These changes will be justified in the following text.

Noting the equivalent circuit of Fig. 9D, it can be seen that the emitter voltage is 0.6 lower than that of the base, or in this case, 3.4 volts. The dc current flowing in the emitter is hence, by Ohm's Law, 3.4 V ÷ 2000 ohms = 1.7 mA. We see from the model that the emitter current is the sum of the base and collector currents. However, the collector current = beta times the base current, and beta is typically a fairly high value. Thus, the emitter current is approximately equal to the collector current. Using this approximation, the collector current is also 1.7 mA. It is significant to note that the value of beta was not even used in the calculation of the emitter and collector currents.

If the beta of the transistor used in the circuit of Fig. 9 was 100, the base current would be 1.7 mA/100 = 17 μA. This current flow through R', the equivalent resistance of the R1-R2 voltage divider, would case a voltage drop of only .02 volt, causing the base voltage not to be 4 volts, but 3.98 volts. This is close enough to 4 volts that the more detailed calculation is not necessary. Generally speaking, the current flowing through the R1-R2 voltage divider (0.8 mA in the example) should be large in comparison with the expected base current. As long as this constraint is maintained, the simplified analysis is justified.

Throughout the text many circuits are presented, using this bias method, many of them containing dc voltage measurements at various points. The reader who is unfamiliar with biasing calculations is encouraged to use these examples as problems to test his understanding of the foregoing concepts.

Typically, the amateur designer biases his amplifiers with the thought that only a single power supply will be available — usually +12 volts. This constraint is the result of the ultimate desire for using the gear in mobile or portable applications where only one power source is available. However, in modern industrial circuits it is common to find a number of power supplies available in a given piece of equipment. For example, in the typical Tektronix

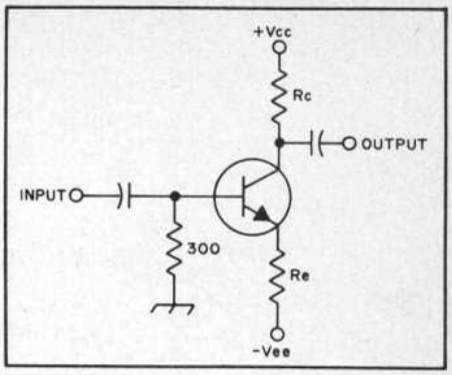


Fig. 10 — Dual supply biasing.

7000-series oscilloscope, voltages of +50, +15, +5, −15 and −50 volts are available to the designer. The access to a large number of supplies greatly simplifies design problems, especially where critical dc biasing situations are concerned. Shown in Fig. 10 is the method for biasing the simple amplifier just considered, when two supplies are available. Since the base is virtually at dc ground potential, the emitter voltage is −0.6 volt. The emitter and, hence, the collector current are given approximately by

$$I_e = I_c = \frac{V_{ee} - 0.6}{R_e} \qquad \text{(Eq. 6)}$$

The collector voltage is merely $V_c = V_{cc} - R_c I_c$.

A special type of diode, which is used frequently as a reference element in a voltage-regulator circuit, is the Zener diode. This component is merely a diode which is operated with a reverse bias that is allowed to increase until the reverse-diode breakdown potential is reached. This voltage is usually quite stable with temperature, and is relatively independent of the current flowing through the diode. Shown in Fig. 11 is a simple model for a Zener diode.

Presented in Fig. 12 is a method for biasing a transistor amplifier when using a Zener diode. In the example, an 8-volt Zener diode is used, yielding I_c = 1 mA, and V_c = 6.6 volts. The approximate design equations are given in the figure.

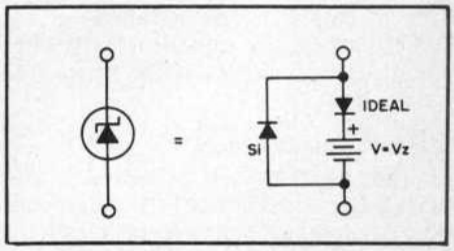
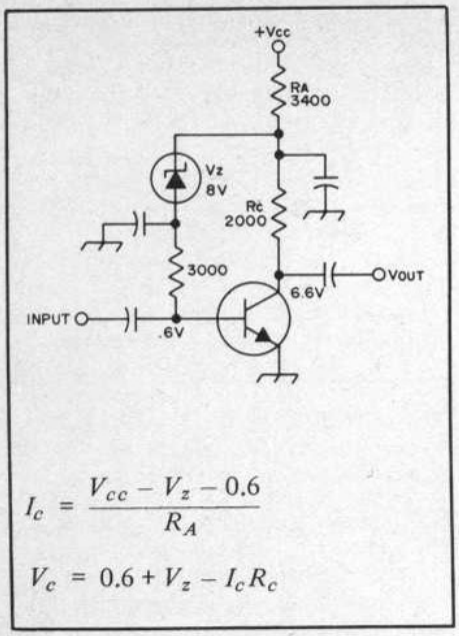


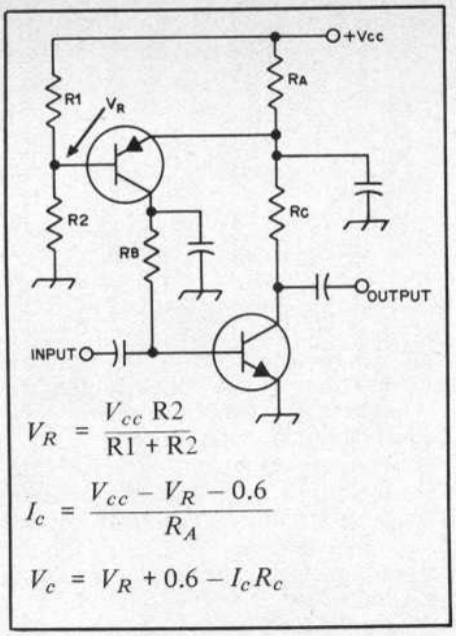
Fig. 11 — Zener diode model.



$$I_c = \frac{V_{cc} - V_z - 0.6}{R_A}$$

$$V_c = 0.6 + V_z - I_c R_c$$

Fig. 12 – Amplifier bias using the Zener diode.

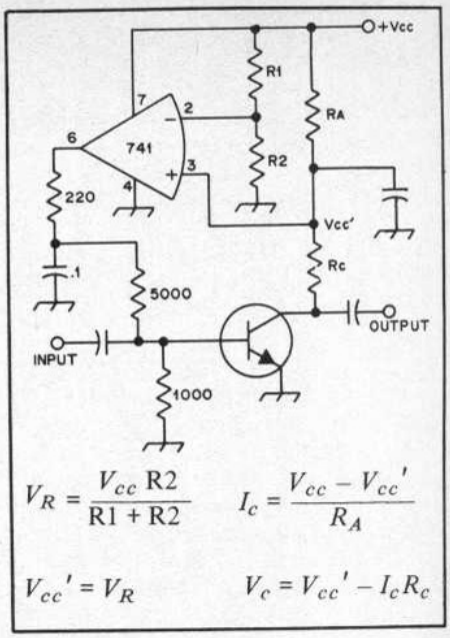


$$V_R = \frac{V_{cc}\,R2}{R1 + R2}$$

$$I_c = \frac{V_{cc} - V_R - 0.6}{R_A}$$

$$V_c = V_R + 0.6 - I_c R_c$$

Fig. 13 – Separate transistor acting as a bias source.



$$V_R = \frac{V_{cc}\,R2}{R1 + R2} \qquad I_c = \frac{V_{cc} - V_{cc}'}{R_A}$$

$$V_{cc}' = V_R \qquad V_c = V_{cc}' - I_c R_c$$

Fig. 14 – An operational amplifier supplying the bias voltage.

Shown in Figs. 13 and 14 are two additional methods for biasing small-signal amplifiers. One scheme uses another transistor, in this case a pnp silicon device such as the 2N3906, while the other technique uses an inexpensive 741 type of operational amplifier. The appropriate design equations are presented with the figures.

The last three biasing schemes may at first sight appear to be absurd, overly complicated and expensive. However, they all have a significant advantage which may not be apparent to the beginner. The asset is that the bias is quite stable and well regulated even though the emitter of the amplifier is at ground potential. This can be of extreme significance when the transistor must be operated at ultra-high frequencies (e.g., 1296 MHz), or if the amplifier is to be used as a relatively high-power output Class A amplifier at rf. In both of these situations it can be difficult to obtain suitable-quality bypass capacitors for the emitter which would allow the simpler methods outlined in Fig. 9 to be used. Furthermore, the transistors used in these applications may cost ten to twenty dollars. In such a situation, it is worth the investment of an extra dime for a Zener diode, a pnp transistor or a quarter for a 741 operational amplifier. As outlined in an earlier section, the true complexity of a circuit is difficult to judge by casual observation.

The Small-Signal Model

The simple models presented in the preceding sections have been general purpose in that they can be used not only for the analysis of the dc biasing conditions, but for the behavior of the amplifier with applied signals. The ability to do analysis at high frequencies was implicit in the model because transistor beta was allowed to decrease linearly with frequency, reaching unity at the f_T of the transistor. The models used by the design engineer are much more complicated, often containing upward of two dozen components, including many capacitive elements. The general procedures are, nonetheless, the same, although the mathematics are sufficiently complicated to require computer-based analysis at times.

Even though the models presented above are quite simple when compared with those used by industry, further simplification can be realized if only small ac signals are considered in the analysis. As an example, consider the simple audio amplifier presented first in Fig. 9 and repeated in Fig. 15, with the circuit redrawn to include the general model. If this circuit is investigated, with respect now to the application of small ac signals, considerable simplification can be realized.

Capacitors C1 and C2 serve as dc blocking units. That is, the dc voltage may be different between the two terminals of the capacitor. However, a small ac signal presented to one end of the capacitor will appear unattenuated at the other side of the capacitor. Similarly, capacitors C3 and C4 are included merely to insure that the emitter of the transistor and the power-supply terminal are at ground as far as ac signals are concerned.

If the interior of the transistor model is investigated, a further reduction can be realized. The 0.6-volt battery in series with the base may be eliminated, since small changes in base potential will be transmitted through

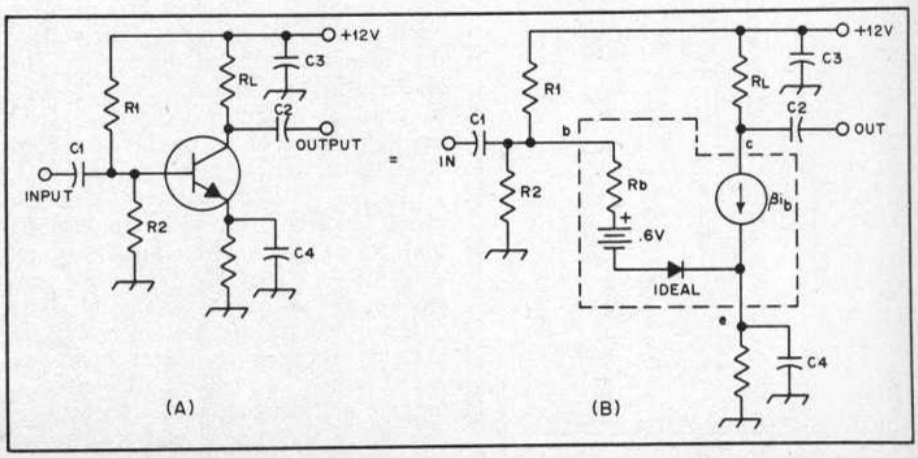


Fig. 15 – The transistorized amplifier redrawn to include the transistor model.

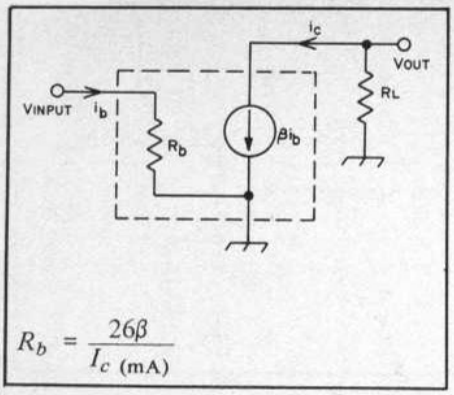


Fig. 16 – Small-signal model of the audio amplifier.

the battery. Similarly, the ideal diode in the base is no longer of practical value, for the dc bias in the transistor will always keep this diode turned on as long as the input signals are kept small with respect to the dc levels present. Shown in Fig. 16 is the small-signal equivalent of the amplifier circuit of Fig. 15. Clearly, this circuit will be much easier to analyze than would be the case if the more complete model were used and all external components were retained.

Consider that an ac input voltage of 1-mV rms is applied to the circuit of Fig. 16. The input current will be $E_{in} \div R_b$. If the transistor has a beta at the operating frequency of 100 and is biased for 2 mA of emitter current, the input resistance of the transistor, R_b, will be 1300 ohms. Hence, the current flowing into the base will be .001 V ÷ 1300 ohms = 0.77 μA. The current flowing *into* the collector will be beta times this value, or 77 microamps. If a 2000-ohm load resistor, R_L, is used, the voltage across the resistor will be $-I_c \times R_L = -(77 \times 10^{-6} \times 2 \times 10^3) = -0.154$ V. The voltage gain is 154.

The minus sign in the output is of significance. This can be seen from a close examination of the model. A current flowing *into* the base of the transistor leads to a larger current flowing *into* the collector. This current will flow through the load resistor in the direction indicated by the arrow. With one end of R_L grounded, the current flow in the indicated direction will mean that the collector end of R_L is going to be negative. Since we are dealing with ac signals, this minus sign indicates merely that the output voltage will be 180 degrees out of phase with the input voltage.

Power delivered to a resistive load, R, is given as $P = V^2 \div R$, where the voltage is the rms value. Using this equation, the input power delivered to the base is $(.001)^2/1300 = 7.69 \times 10^{-10}$ watt. The output power is similarly $(0.154)^2/2000 = 1.19 \times 10^{-5}$ watt. The ratio of these powers is the power gain, in this case 15,400. This can be expressed in dB with the expression G_p (dB) = 10 log P_{out}/P_{in}, or in this case 41.9 dB.

The use of small-signal models is quite universal in almost all areas of circuit design, and the science has been well developed by using advanced matrix methods. This discipline is often described under the name "two-port network theory." Although the mathematics are complicated enough that such methods are not appropriate for a book aimed at the radio amateur, they are still exceedingly powerful, and do not require the use of a computer except in some of the more specialized cases. Some of the basic two-port network concepts are presented in the appendix, and have been used for many of the more refined designs in this book.

Even though the full utilization of modeling methods is probably beyond some amateurs, the limited models can still be of extreme utility. When a circuit is first encountered, the builder should study the circuit and evaluate the biasing conditions. After this is done, the equivalent small-signal circuit may be redrawn, either on a sheet of paper or mentally. Through this process surprisingly complex circuits may often be analyzed with ease.

Biasing and Modeling Field-Effect Transistors

Although the workhorse of modern communications technology is the bipolar transistor discussed in the preceding sections, a device of increasing popularity is the field-effect transistor (FET). There are several methods which are used to construct FETs, leading to various schematic symbols and design approaches. The popularity of the FET with radio amateurs is, in large part, due to their similarity of behavior to the more familiar vacuum tube.

The basic dc characteristics of an n-channel junction FET are outlined in Fig. 17. Probably the two most significant dc parameters are I_{dss} and V_p. The current, I_{dss}, is that which will flow in the FET if the gate and source are tied together and the drain is biased at a voltage higher than the magnitude of V_p. The parameter V_p is called the pinch-off voltage and is the voltage applied to the gate with respect to the source, which will cause the drain current to go virtually to zero.

Probably the easiest method for designing the biasing of a JFET (junction FET) into the active region is to use a graphical technique to determine the value of a suitable source resistor. The circuit is shown in Fig. 18, and a suitable graph is shown in Fig. 19. In the graph we have assumed that the values for I_{dss} and V_p are, respectively, 10 mA and −6 volts. The curve of Fig. 17 is approximated in the graph with a straight line. If it is desired to bias the FET to a drain current of 5 mA, a *load line* is drawn from the origin to the 5-mA point on the FET characteristic curve. The voltage at this point is −3. The slope of this line is 3 V ÷ 5 mA, corresponding to a resistance of 600 ohms. This is thus the value of resistor which would be chosen for the source bias. While this method is approximate, it should suffice for most amateur applications.

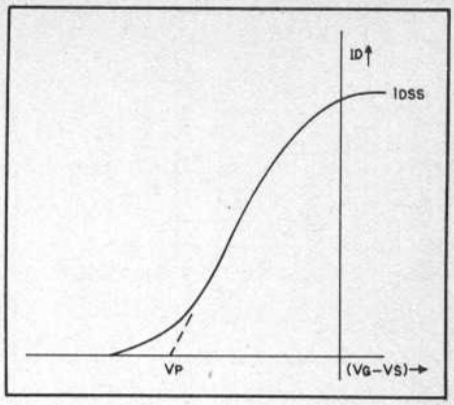


Fig. 17 – Basic dc characteristics of the junction FET.

Shown in Fig. 20 is a simple small-signal model for a JFET. Like the models used for the bipolar transistor, the basis which leads to a description of amplification is a *dependent-current generator*. However, where the bipolar transistor had a current generator in the collector circuit which was dependent

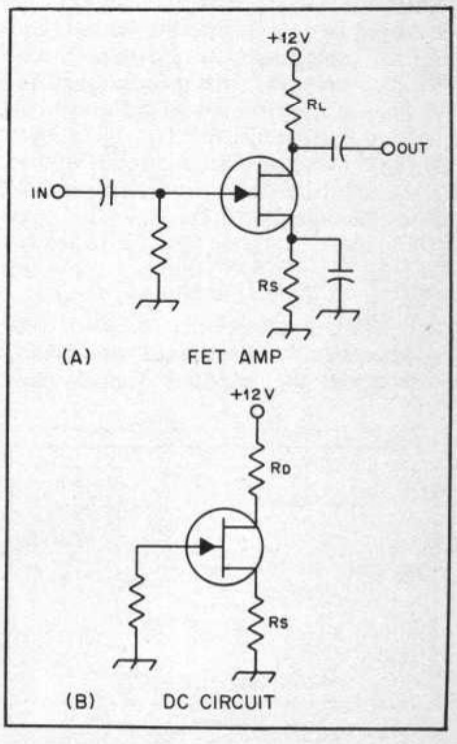


Fig. 18 – FET biasing schematic.

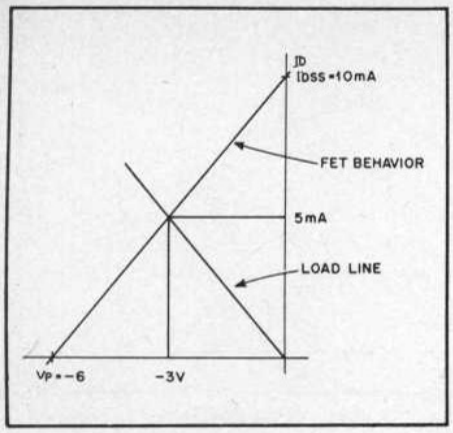


Fig. 19 – FET behavior with biasing.

upon the base *current*, the generator in the FET is dependent upon the *voltage* on the gate of the FET. Since the input resistance of a typical FET is extremely high, the input can be fairly well represented with an open circuit. The constant relating drain current to gate-source voltage is the transconductance and has the units of mhos (= 1 ÷ ohms). Typical values might be 4000 micromhos, or .004 mho for a popular FET like the MPF102 or the 2N4416.

Shown in Fig. 21 is a typical audio amplifier which uses an FET with the constants of the foregoing examples. In this circuit a large resistor is used to connect the gate of the FET to ground, to ensure that the proper bias conditions are maintained. Using the analysis methods just outlined, the dc drain voltage would be found to be +7, the dc source voltage would be +3, and the voltage gain would be 4. (Note that the transconductance of a typical bipolar transistor is *much* higher than that of an FET.) Although the voltage gain of the FET is only 4, the power gain is virtually infinite. This is because a finite power output is delivered to the 1000-ohm drain resistor, but the input to the FET is essentially an open circuit, which will not accept power.

Negative Feedback and the Integrated-Circuit Operational Amplifier

Although the transistors and FETs outlined in the previous sections are used for the predominant applications in communications equipment, in many areas integrated circuits have gained wide acceptance. Of the many ICs available, undoubtedly the most generally useful type is the operational amplifier, or "op-amp," with the most common example being the μA741. In recent years these devices have become so common in industry and in amateur work that their prices have dropped to very low levels. With such a low cost (usually 50 cents or less in small quantities), they can be used with the same casualness that one would exercise in adding a transistor or a capacitor to a circuit.

While 741 op amps have been used widely in amateur circles, they have also been used improperly in many situations. The misuses have resulted from a lack of understanding of the principles and consequences of feedback and an incomplete understanding of a proper equivalent circuit to use in circuit design and analysis.

Shown in Fig. 22 is the circuit symbol for an integrated op amp of the 741 type along with a suitable equivalent circuit or model. There are several differences here from the models used with transistors and FETs. First, the output is not a current source, but a *voltage* source. Second, the op amp is a *differential amplifier*. That is, the output voltage is directly proportional to the *difference between the two input voltages*. The constant of proportionality is the open-loop voltage gain, A_o. Finally, the equivalent circuit of Fig. 22 is reasonably accurate for both dc conditions and for small-signal analysis.

The two inputs are labeled with *a*+ or *a*–. The + input means that an increase in the voltage at this terminal causes an increase in the output. This + terminal is called the noninverting input. The – input, or the inverting input terminal, exhibits the opposite behavior. That is, an increase in its potential leads to a decrease in the output potential. The impedances seen at the two input terminals are high, typically. They are not as high as experienced with FETs, but are high enough to make the model of Fig. 22 valid in most applications.

The value of A_o is typically high – 10,000 to 100,000, or even more. However, this is the gain at dc and very low ac frequencies. As the frequency increases, the value of A_o starts dropping, decreasing by a factor of two for every doubling of the frequency. The 741 op amp has a gain of approximately 1000 at 1 kHz, and the voltage gain drops to unity at frequencies of about 500 kHz.

There are some limitations to the performance of an op amp, and they are fairly obvious. Mainly, the output voltage cannot go higher than the positive supply voltage, V_{cc}, nor can it go lower than V_{ee}. Actually, with 741-type op amps, one is safe to assume that the output can approach each supply within about 2 volts. If two supplies of + and –15 volts were used, as is the usual case with industrial equipment, the output might be expected to swing from –13 to +13 volts. If a single 12-volt supply was used, as is the typical situation in most amateur applications, the output could be expected to range from +2 to +10 volts or a little higher.

In discussing op amps, it is generally easier to describe the behavior if two supplies are used. Hence, for the typical amateur application where a single supply is to be used, a "synthetic ground" will be created with a resistive divider. All voltages in the rest of this discussion will be with respect to this level. The circuit is shown in Fig. 23. Note that this would be exactly the same as working with + and –6-volt supplies, derived from a floating 12-volt battery.

The behavior of an op amp will be described in terms of a number of circuit situations. The experimentally inclined amateur might wish to breadboard some of these in order to obtain a better feel for the phenomenon.

In the first experiment (Fig. 24) the noninverting input of the amplifier is "grounded" and a signal, E_{in}, is applied through a 10-kΩ resistor to the inverting input. The output is described by the equation, noting now that $V_+ = 0$, leaving $V_{out} = -A_o V_{minus}$. Assume for this experiment that A_o is 1000. If E were set at a positive 1 mV, the output

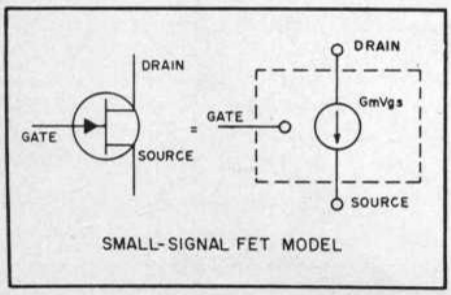


Fig. 20 – Small-signal model of JFET.

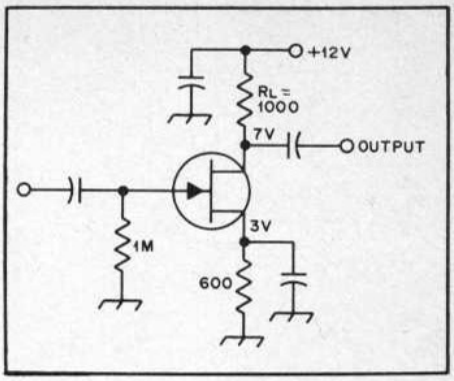


Fig. 21 – Audio amplifier using a JFET.

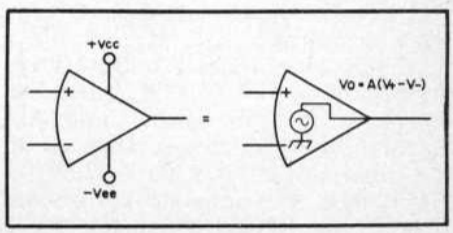


Fig. 22 – Operational amplifier model.

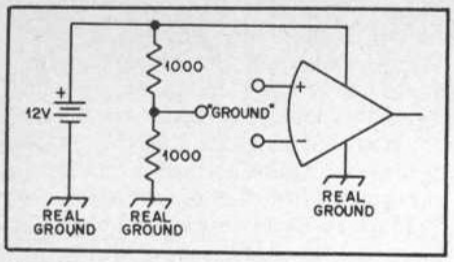


Fig. 23 – Synthetic ground for an operational amplifier.

will be −1 volt. Similarly, if E were set at a negative 1 mV, the output would be 1-volt positive with respect to the synthetic ground.

It is also instructive to examine the input resistance of this composite amplifier. The op amp itself has virtually an open circuit at its input. Hence, no current will flow in the 10-kΩ resistor, and the resistance seen at the driving source, E, is essentially infinite. This may seem like a redundant statement at this point, but later experiments will lead to different results.

Consider now the modification of the first experiment where a feedback resistor is added. This is presented in Fig. 25, where E is now +1 volt. As the input voltage is increased toward this 1-volt level, the voltage at the inverting input will also tend to increase. This input change will be reflected through the amplifier and amplified by a factor of A_o, making the output try to go negative. However, as the output voltage decreases, a negative voltage from the output is applied through the feedback resistor to the input. Since this fed-back input signal opposes the original driving signal, it is not immediately clear just where either the V_{minus} input or the output voltage will end up.

This is one of those situations where the use of a little elementary mathematics cannot be avoided. The procedure in setting up the equations is really quite straightforward and should not frighten any amateur who has taken high-school algebra.

Although the value is not yet known numerically, the voltage at the inverting input is specified as V_{minus}. The current flowing into the overall circuit is $(E - V_{minus})/R_i$. Since the op amp itself appears as an open circuit, no current flows into its terminal. However, there will be current flowing in the feedback resistor with a magnitude of $(V_{minus} - V_{out})/R_f$. These two currents must be equal since the total current entering a point in a circuit must be zero. This gives us the equation

$$\frac{(E - V_{minus})}{R_i} = \frac{V_{minus} - V_{out}}{R_f} \qquad \text{(Eq. 7)}$$

but, V_{out} is known: $V_{out} = A_o(V_+ - V_{minus}) = -A_o V_{minus}$.

This value for V_{out} is now substituted in the first equation and the equation is solved for V_{minus}. The net result is

$$V_{minus} = \frac{R_f E}{R_i (A_o + 1) + R_f} \qquad \text{(Eq. 8)}$$

Noting again that $V_{out} = -A_o V_{minus}$, we can solve for the *closed-loop voltage gain*.

$$G_V = \frac{V_{out}}{E} = \frac{-1}{\frac{1}{A_o} + \frac{R_i}{R_f}\left(\frac{A_o + 1}{A_o}\right)}$$

$$= -1.994 \qquad \text{(Eq. 9)}$$

For large values of A_o, we see that the last equation reduces to

$$G_V \approx R_f \div R_i = \frac{20\text{k}}{10\text{k}} = 2 \qquad \text{(Eq. 10)}$$

It is also instructive to calculate the input resistance of the circuit of Fig. 25. The effective input resistance is just $R_{in} = E \div I_{in}$. But, the input current, I_{in}, is just given by the expression $I_{in} = E - V_{minus} \div R_i$ where V_{minus} was arrived at in an earlier equation. Using this expression and noting the values used in the diagram of $A_o = 1000$, $R_i = 10$ kΩ and $R_f = 20$ kΩ, we calculate that the effective input resistance is 10,019.98 ohms. Of this, 10,000 ohms is attributed to the input resistor, R_i. The other 20 ohms is the effective resistance seen at the inverting input of the operational amplifier. Generally, the input resistance of such a circuit at the inverting input is R_{in} at V_{minus} port $\approx R_f \div A_o$. It can also be shown that the output resistance of an amplifier is reduced when negative feedback is introduced. To do this, we would have to modify our model to include some finite output resistance in series with the voltage source now used.

While the foregoing analysis may appear to the amateur, who is uncomfortable with simple mathematics, to be nothing but a bunch of esoteric gibberish, the results are really profound and should be treated as such! In the beginning of the problem, we took an amplifier which had a high, but perhaps ill-defined, gain with input and output resistances which might be quite unknown. However, by applying feedback we ended up with a total circuit whose gain was determined by the ratio of two resistors and an input resistance which was well defined. Since the open-loop gain of the amplifier was variable with frequency, but the final expression for gain (Eq. 10) does not contain the open-loop gain, the ultimate amplifier response is virtually independent of frequency.

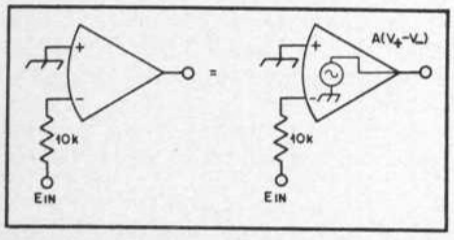


Fig. 24 – Operational amplifier connected in the inverting configuration.

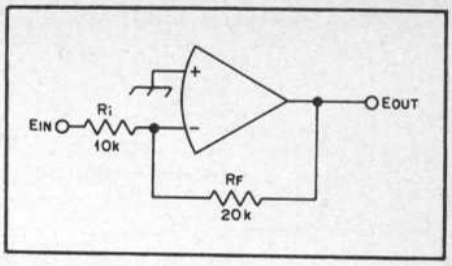


Fig. 25 – Operational amplifier with feedback.

There is another way to view the previous amplifier, which is extremely useful in the casual design of circuits with feedback. Viewing Fig. 25, while disregarding the mathematics for awhile, we see that the input signal causes a current to flow in R_i and some small voltage to appear at the inverting input. However, with negative feedback the output voltage moves around in such a way that the voltage difference between the two inputs is maintained essentially at zero.

This general view may be used to easily analyze a noninverting feedback amplifier. Consider the circuit shown in Fig. 26, where feedback is used but the input signal is applied to the noninverting input. With the input signal initially equal to zero, the output voltage will adjust itself until the voltage at V_{minus} is also zero. This will occur for $V_{out} = 0$. Now, assume that E_{in} is increased to 1 volt. The output voltage will move in such a manner that the voltage at V_{minus} is also +1 volt. But, this will occur when the output voltage is 3 volts. The only place current can come from to put the inverting input at 1 volt is from the divider formed by R_f and R_i being fed by V_{out}. In general the

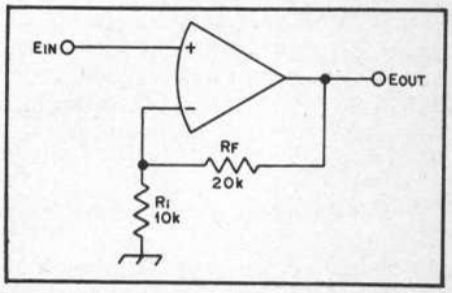


Fig. 26 – Non-inverting amplifier with feedback.

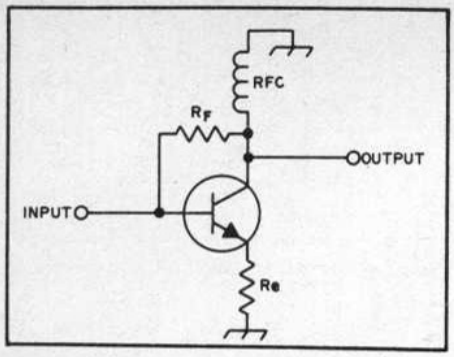


Fig. 27 – Transistorized amplifier with feedback.

gain of a non-inverting amplifier is

$$G_V \approx 1 + \frac{R_f}{R_i} \text{ for large } A_o \qquad \text{(Eq. 11)}$$

Although it will not be shown at this time, feedback of this kind has the effect of increasing the input resistance seen at the non-inverting input, while still decreasing the output resistance. Again, these effects cannot be demonstrated mathematically with the model used due to the initial simplifying assumptions which were used.

Although the details will not be presented until later chapters, feedback may be applied to simple one-transistor amplifiers in order to realize the same advantages achieved with an operational amplifier. Shown in Fig. 27 is the small-signal equivalent of a circuit of this kind. With the proper choice of feedback resistors, this amplifier may be designed such that the input and output impedances are both very close to 50 ohms and the gain is flat from under 1 MHz to the low vhf region if a good transistor is used. Feedback is one of the most powerful tools available to the amateur or professional designer.

Chapter 2

Basics of Transmitter Design

The basic element of any amateur radio station is the transmitter. In years past, the transmitter found in the usual "ham shack" was a large unit, often mounted in a floor-to-ceiling rack cabinet. This "machine" was decorated with a large collection of knobs and meters, all serving a necessary function. Some of the more elegant units even had windows which were covered with glass or a wire mesh, which allowed the final amplifier tubes to be monitored visually. Too much color on the plates indicated that perhaps the tubes were being pushed a little too hard.

Times have changed and the modern homemade transmitter is often a small unit, designed with a minimum number of panel-mounted controls. If the builder acquires a flair for miniaturization, the QRP transmitter can be very small indeed.

In spite of the variations in size, and the fact that most of the modern equipment built by the radio amateur is solid state, there are many similarities. Shown in Fig. 1 are block diagrams for cw transmitters of varying degrees of complexity. These range from the simple crystal-controlled transmitters to a frequency-synthesizer-based unit. All of these examples could be realized with modern solid-state technology or the vacuum-tube methods of the past. In this, as well as the following chapter, all of the systems outlined in the figures will be discussed. An attempt is made to expand those areas where minimum information has been published previously. Many of the basics are reviewed also.

Crystal Oscillators

The workhorse of modern communications equipment is the crystal oscillator. In the simplest kind of transmitter, a crystal oscillator may serve as a complete circuit. More often, such oscillators are used to drive additional amplifiers to provide increased power output. In the more advanced amateur transmitters, crystal oscillators are used in conjunction with mixers and VFOs in a superheterodyne circuit design. Ultimately, the most advanced designs will use a crystal-controlled oscillator as the reference for a frequency synthesizer.

The crystals used in communications technology are usually made from quartz, where the basis of operation is the piezoelectric effect. Materials which exhibit this effect have the characteristic that when subjected to an electric field, a mechanical stress occurs within the crystalline material. The mechanical displacement resulting from this stress is often in a direction different from that of the electric field. Depending upon the nature of the crystalline material and the physical size and mounting, a quartz crystal will exhibit mechanical resonances in much the same way that the strings of a musical instrument have mechanical resonances. The unusual characteristic of piezoelectric devices is that not only can an electric field cause a stress which will excite an internal mechanical resonance, but the presence of mechanical stress will generate an electric field. The net result with a quartz crystal is that we end up with a small device consisting of nothing more than a piece of quartz with two electrical connections which, electrically, behaves just like a tuned circuit. The equivalent circuit for a quartz crystal is shown in Fig. 2.

The values associated with the equivalent L and C values are often much different than those we would experience in circuits built with discrete components. For example, the series inductance, L_s, may approach one henry, with a series capacitance of a few femtofarads (10^{-15} farad). The parallel capacitance, C_p, is typically around 6 pF. While not shown in the figure, there are also loss elements in a more complete equivalent circuit, which will give rise to a finite Q. The typical Q of a crystal which might be used in amateur transmitters would be around 50,000. In some special crystals, Qs of over 1,000,000 are achieved.

There are dozens of circuits which can be used to make oscillators with quartz crystals. We will present a few of them here.

Shown in Fig. 3 is a circuit using a bipolar transistor. Here, a transistor is biased in the usual way, and is operated much like an LC tuned oscillator in the common-base mode. However, the usual base-bypass capacitor is replaced with a crystal which operates as a series-tuned circuit. With a 12-volt supply, this circuit will deliver a typical power output of 20 mW or so. The signal on the collector is approximately 10- to 15-volts pk-pk.

In this oscillator stray and transistor internal capacitances provide feedback for oscillation. Proper feedback is maintained by adjusting the external capacitor at the emitter of the transistor. This capacitor should be one which will exhibit some 200 ohms of reactance at the operating frequency (e.g., 100 pF at 7 MHz). The tuned-collector circuit is resonant at the operating frequency. This circuit may be hesitant about oscillating at the lower frequencies, especially at 160 and 80 meters. In these cases, it is often possible to make an excellent oscillator by adding a ca-

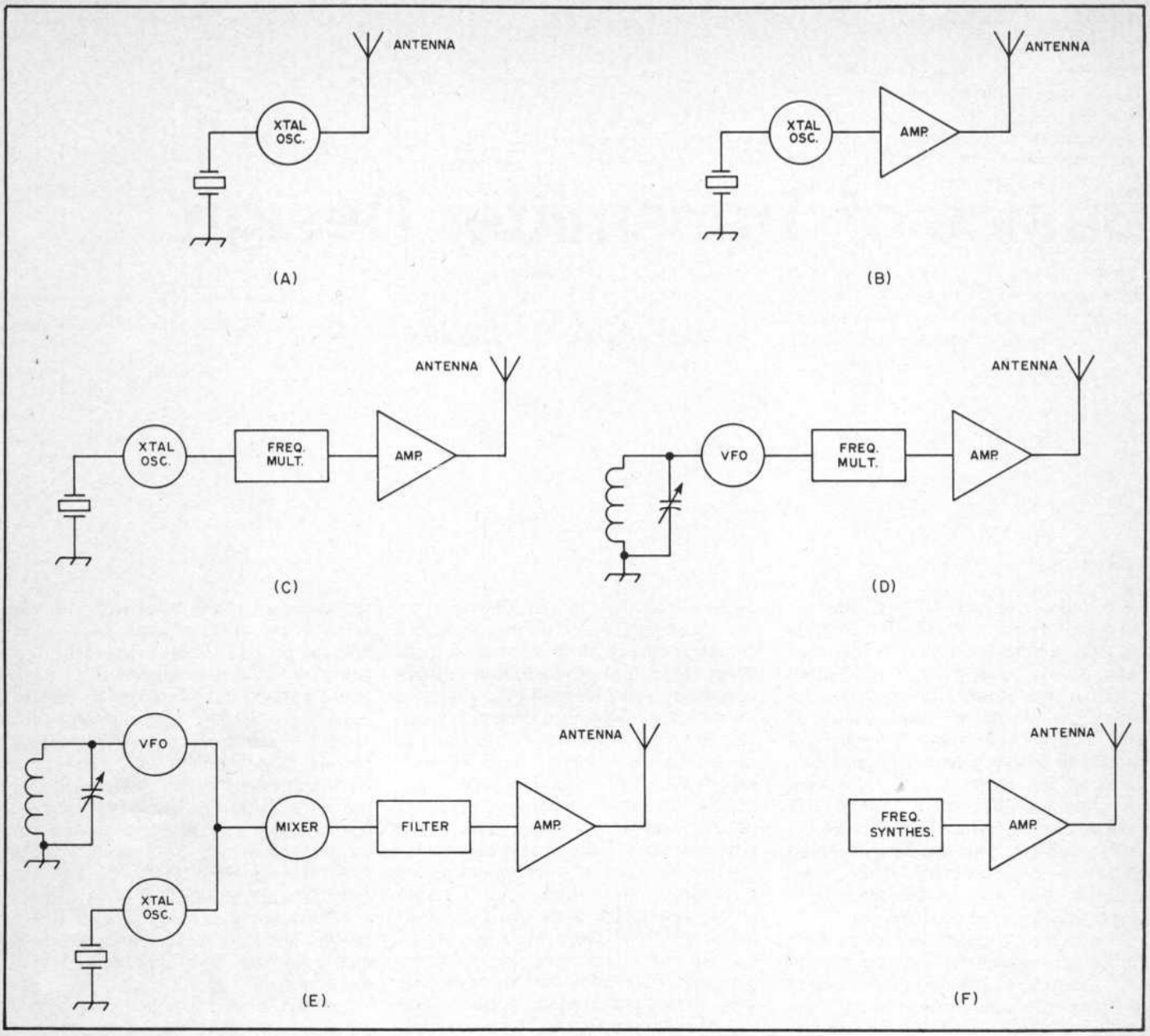


Fig. 1 — Block diagrams of various cw-transmitter formats.

pacitor between the base and the emitter. Typically, a capacitive reactance (X_C) of 500 ohms is sufficient.

One useful characteristic of this circuit is that it will operate on the overtone modes of a crystal. An overtone is merely an oscillation which uses a harmonic resonance of the crystal. That is, a violin string can be made to oscillate at frequencies higher than the one typically associated with the length and tension in the string. It is the existence of these harmonics, along with the fundamental, which adds character to the sound, differentiating the violin from a simple audio oscillator. In a similar manner, a crystal can be made to oscillate on higher overtones. Because of the mechanical boundary conditions imposed upon the crystal, overtone oscillations will occur only at *odd multiples* of the fundamental frequency. Furthermore, the high Q of a crystal (in comparison with that of a violin string) allows the overtone oscillation to occur alone, without the presence of the fundamental.

An example of a third-overtone crystal oscillator is the circuit of Fig. 3 with all constants set for 21 MHz. However, the crystal is a 7-MHz fundamental unit. The output of the overtone oscillator will be at 21 MHz. Absolutely no output will be detected at 7 MHz!

When crystals are purchased, they will usually be fundamental-mode devices up to a frequency of around 20 MHz. From 20 to 60 MHz, third-overtone units are typical. Some 5th-, 7th- and even 9th-overtone crystals are used in communications equipment. In many cases a crystal will exhibit a higher Q at its overtone frequencies than at the fundamental.

Shown in Fig. 4 is a simple crystal oscillator using a junction field-effect transistor (JFET). This circuit will operate on crystal overtones as well as at the fundamental of the crystal, depending upon the tuning of the output circuit. The simplicity of this circuit makes it appealing, although the cost of a JFET is usually higher than that of a good bipolar transistor.

The JFET oscillator is converted easily to a simple variable-crystal oscillator (VXO) by paralleling the crystal with a 100-pF variable capacitor. The ability to "pull" the frequency of a crystal is, generally, limited to fundamental-mode oscillations in this circuit. Using a 14-MHz fundamental-mode crystal (International Crystal, type EX),

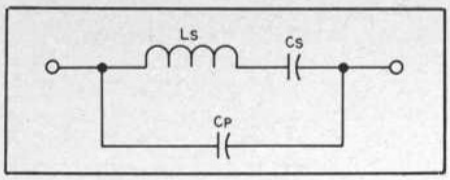


Fig. 2 — Equivalent circuit for a quartz crystal.

a frequency shift of 8.4 kHz was measured. On the other hand, using a 7-MHz crystal, only 1.4 kHz of shift was measured. Although the ability to "VXO" a crystal is highly dependent upon individual crystal characteristics, the technique is still useful. For example, an oscillator like that shown in Fig. 4, operating at 18 MHz and followed by a suitable frequency-multiplier chain, could yield an excellent exciter for 2-meter cw. That approach could be used for the hf bands also, even though the tuning range would be limited.

The bipolar-transistor oscillator of Fig. 3 can also be pulled by means of external components. This is most easily done by adding an inductor in series with the crystal. The inductance value will depend upon the individual crystal and the "pull" amount desired, but is typically a few microhenries (μH) per kHz of shift when using a 7-MHz crystal. A simple means of utilizing this VXO capability is to mount a slide switch across the inductor. This will, in effect, give the builder the ability to shift his oscillator frequency down enough to dodge QRM, certainly a desired objective with a crystal-controlled QRP transmitter as the example. Up to 15 kHz of shift in a 7-MHz crystal-controlled oscillator has been measured with this circuit.

Shown in Fig. 5 is a JFET VXO. In this circuit the system is optimized for maximum frequency shift with standard crystal types, while maintaining a fairly constant output voltage. This required the use of a dual-section variable capacitor for tuning, and careful component mounting was necessary to minimize stray capacitance. The inductor is a high-Q slug-tuned unit. Probably, the Q of the coil is not as critical as is the self-capacitance. A toroidal inductor on a relatively high permeability powdered-iron core (such as the Amidon Assoc. E series) might work well. Experimentation is clearly required on the part of the builder. A frequency shift of 12 kHz with a 6-MHz crystal, and a shift of 23 kHz with an 11-MHz unit was obtained, confirming that the maximum shift available is around 0.2 percent of the crystal frequency. A VXO of this kind would provide the basis for a number of interesting transmitters or transceivers.

The VXO of Fig. 5 was breadboarded and tested with a number of different crystals. An experimental change from the circuit shown was the use of a hot-carrier diode in place of the 1N914 and smaller inductance values at L. The output is surprisingly constant over most of the tuning range of a given crystal, with variations less than 1 dB being typical. Using a 10-MHz crystal, a 17-kHz shift was measured with a 16-μH slug-tuned inductor. Several overtone crystals were operated on their fundamental modes, and spectacular results were noted in some cases. For example, a 54-MHz third-overtone crystal was operated at 18 MHz with the 16-μH inductor. An excess of 150 kHz of shift was noted! The tuning was nonlinear, with most of the range being compressed near the low-C end of the variable capacitor spread.

Two more oscillators using bipolar transistors are shown in Fig. 6. Neglecting slight differences in biasing, the circuits are essentially identical. They offer the advantage of requiring no tuned circuit for operation. Both are fundamental-mode oscillators.

All of the circuits shown are aimed at reasonable stability, but have relatively low output power. It is possible to bias many of these circuits higher to obtain outputs of up to perhaps 1/4 watt. However, thermal stability is often severely degraded, chirp is introduced if the oscillator is keyed, and the user

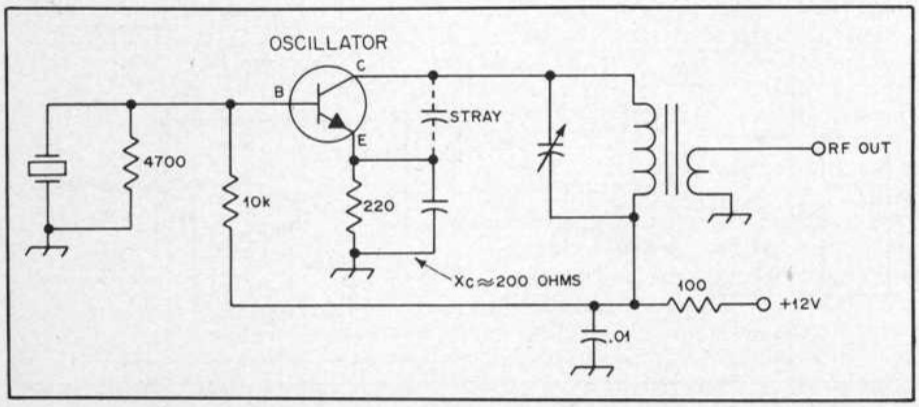


Fig. 3 — Circuit for a bipolar-transistor crystal oscillator.

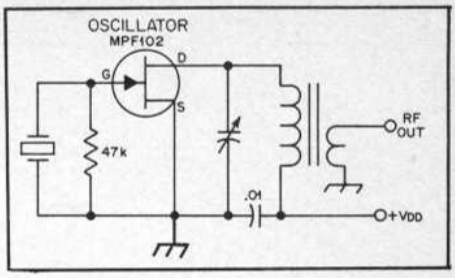


Fig. 4 — Crystal oscillator which employs a JFET.

stands a chance of damaging the crystal from excessive rf current. It is *not* recommended that a single oscillator stage be used as a simple transmitter. The addition of an amplifier is so straightforward, and the system efficiency is so much better, that the minimal simplicity is not of value.

Most crystal oscillators which use bipolar transistors will operate fairly well with hundreds of different transistor types. Generally, the only requirement other than the usual voltage-breakdown and maximum-current criterion is that the transistor have as high an f_T as possible. This is met easily for oscillators in the hf region with transistors likethe 2N3904, 2N4124, 2N706, 2N2222A, 2N3563 and others. For overtone oscillators operating well into the vhf region, one should select transistors with an f_T of 1 GHz or higher. The 2N5179 is excellent in such applications.

Designing Untuned Buffer Amplifiers

While the output of a low-frequency crystal oscillator may be as high as 50 milliwatts (mW) or more, the output from a VFO or mixer in a heterodyne exciter may be much less. An amplifier is needed to build up the power. Also, amplifiers help isolate an oscillator from the effects of changing load, such as might result from keying or modulation. These chores are usually handled by means of a Class A buffer/amplifier. In this section, the basics of untuned amplifiers will be presented. The following section will review the design of tuned Class A amplifiers.

This presentation is, by necessity, oversimplified. A more exhaustive treatment would carry us well beyond the scope of this volume. An attempt is made at justifying some rules of thumb which will be used later in the text. The reader who is not familiar with basic transistor concepts is urged to review a good basic treatment of the subject. The series of articles in *QST* by Stoffels is excellent.[1]

Consider first the simple amplifier shown in Fig. 7. This amplifier operates in Class A, which means that collector

[1] Available in reprint form from ARRL for $1.

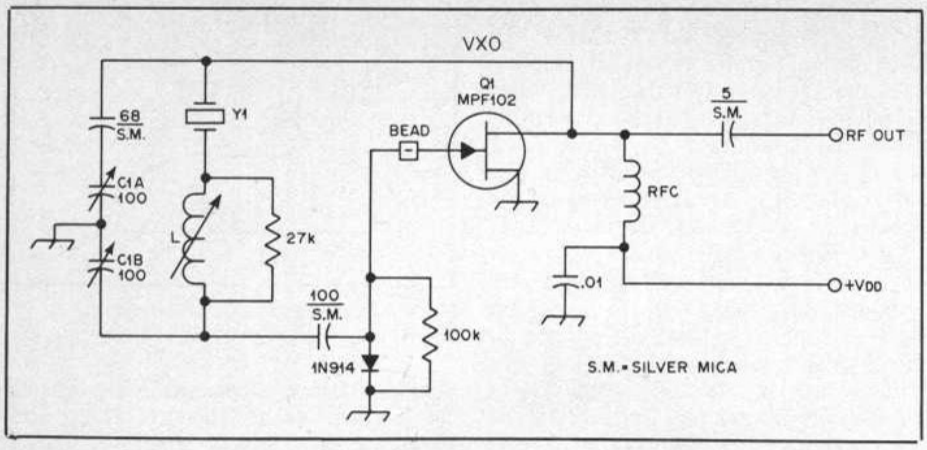


Fig. 5 — VXO circuit for pulling the crystal frequency.

current flows during the *entire* drive cycle.

First, we will review the biasing. The base is driven from a voltage source of 4. Since we are using a silicon transistor, the emitter voltage will be less than the base by about 0.7 volt, or 3.3 volts. The emitter current is 3.3 ÷ 500, or 6.6 mA. Since the collector current is virtually the same as the emitter current, the collector voltage is $V_{cc} - I_cR_c$ = 8.7 volts.

This arithmetic is based on the assumption that the base is biased from a *true* voltage source. It's wise to confirm this. The current in the base-voltage divider is 12 V ÷ 15 kΩ or 0.8 mA. If the β of our transistor is 100, the base current is $I_b = I_c \div \beta$ = 6.6 mA ÷ 100 = 66 μA. Since the current in the divider is 10 times this value, our bias divider is indeed "stiff" enough.

These bias calculations describe the operation of the transistor at dc. Our interest, however, is in the behavior of the circuit for an ac signal. For rf signals the emitter is essentially ac-grounded through the emitter bypass capacitor. Recall that a capacitor is a device which has the characteristic that the impressed voltage cannot change instantaneously. Any rf signal that appears at the emitter of the transistor will be connected to the capacitor directly. Since the voltage at this point cannot change instantaneously (i.e., at an rf rate), all ac parts of the emitter current flow through the capacitor rather than through the 500-ohm emitter resistor. Thus, we treat the amplifier as a grounded-emitter stage.

The input resistance of a grounded emitter amplifier is approximated by $R_{in} = 25\beta \div I_e$, where the emitter current is in mA. The beta used in this equation is not the dc-current gain we used in the preceding bias discussion, but is the *ac-current gain*, which is well approximated by $\beta_{AC} = f_T \div f_{op}$, where f_{op} is the operating frequency of the amplifier and f_T is the usual gain-bandwidth product. If we use a transistor with a 150-MHz f_T at a frequency of 7 MHz, the ac beta is about 20. Hence, the input resistance of the amplifier is 75 ohms. The .01-μF capacitor in the input merely serves to block dc. That is, it allows a difference in dc voltage to exist between the amplifier input and the output of the previous stage, but offers essentially no impedance to the flow of rf currents.

Let's assume that the amplifier is driven with .01 volt (10 millivolts). The input current (rf only) will be $I_b = E_{in} \div R_{in}$ = .01 ÷ 75 = 0.133 mA. The collector signal current is then $I_c = \beta I_b$ = 20 (0.133 mA) = 2.66 mA. This current flows through a load resistor of 500 ohms. Again, using nothing but Ohm's Law, we see that the output voltage is 1.33. The small-signal voltage gain is 1.33 ÷ .01, or 133.

What would happen if we increased the input drive from 10 mV to 0.1 volt? If we were to follow the foregoing analysis again, we would calculate an ac current of 26.6 mA in the collector. However, the dc current is only 6.6 mA. There is no way that this can happen in a *linear* amplifier. On positive peaks of the input voltage, the collector voltage would be driven down until it was nearly at the voltage of the emitter. This condition is called ***saturation,*** and is typified by reduced current gain. On negative peaks of the input signal the collector current decreases from the dc level of 6.6 mA until it is zero. The current *can't* go *negative* in a transistor. At this point, there is no collector current flowing; hence, the output voltage equals the supply voltage of 12. We see that our amplifier is clipping the output waveform on *both* positive and negative peaks. What can be done to avoid this distortion? There are three possible solutions. First, we can reduce the drive level. Second, we can increase the dc current flowing in the stage while simultaneously reducing the output load resistance. Finally, we can introduce some negative feedback in the amplifier, thus bringing about a reduction in stage gain.

Let's consider the feedback solution by analyzing the modified circuit of Fig. 8, where emitter degeneration is introduced. First, we note that the dc resistances are the same as before. Therefore, the dc bias current has not changed from the previous 6.6 mA. Using this value we find the dc voltage across the capacitor, labeled V_x in the schematic, to be 2.64 volts. This point is bypassed, so it cannot change in potential when rf is applied. Assume now that a signal of 0.1-volt peak is applied to the input (0.2-volt pk-pk). As the input voltage goes from 0 to 0.1 volt, the base voltage will increase by 0.1 volt. The emitter voltage will also increase and follow the base, going from the dc level of 3.3 to 3.4 volts. Noting that the V_x point in the emitter circuit is bypassed, the emitter current will increase to an instantaneous value of (3.4 − 2.64) ÷ 100 = 7.6 mA. The collector current is essentially the same. Hence, the collector voltage will drop to $V_{cc} - I_cR_c$, or 12 − 7.6 (0.500) = 8.2 volts. But, the dc voltage was 8.7 volts. Hence, the voltage change is 0.5 volt. The small-signal gain is now 0.5 volt peak ÷ 0.1 volt peak = 5. Note that the voltage gain is now the ratio of the collector load to the unbypassed part of the emitter resistor.

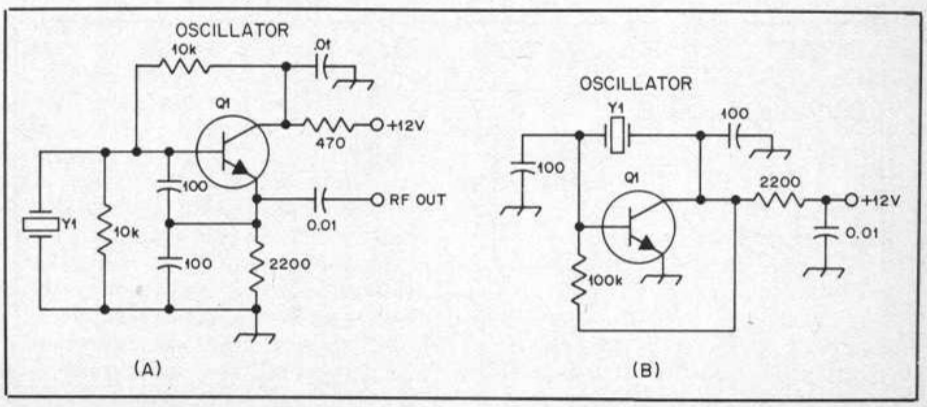


Fig. 6 — Crystal oscillators which use no tuned circuits.

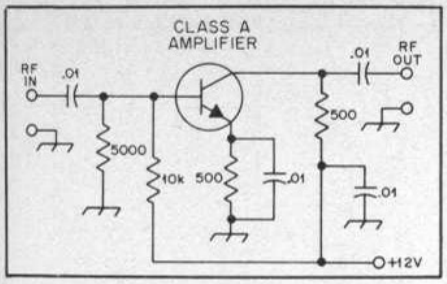


Fig. 7 – Circuit of a simple Class A rf amplifier.

We can extend this simple argument to show that the input resistance of the amplifier has increased. With an input signal of 0.1 volt peak, the collector current increased from 6.6 to 7.6 mA (1 mA). Since the high-frequency beta of the transistor is 20, the base-current increase is 1 ÷ 20 mA. The small-signal input resistance is given by

$$R_{in} = \frac{\Delta V}{\Delta I} = \frac{0.1}{\frac{1}{20} \times 10^{-3}} = 2000 \text{ ohms} \qquad \text{(Eq. 1)}$$

where the deltas signify a small change.

In general, the input resistance of a transistor with emitter degeneration is βR_e, where R_e is the unbypassed portion of the emitter resistance.

By using emitter degeneration we have realized a number of goals. First, the distortion is removed, for the signals are significantly less than the dc bias conditions in the amplifier. We have substantially increased the input resistance, making the amplifier much more effective as a buffer. Finally, we have realized a gain which is dependent upon resistor values, rather than upon transistor characteristics. As a bonus the bandwidth of the amplifier will be significantly higher.

In the form shown in Fig. 8 our amplifier is not especially useful, for the output is not connected to anything. All of the output power is being delivered to the 500-ohm collector value. Suppose we coupled the amplifier capacitively to a following stage with an input resistance of 500 ohms. The net load on the amplifier is now the *parallel combination* of the *two* loads, or 250 ohms. With a reduced collector-load resistance the voltage gain has dropped to 2.5.

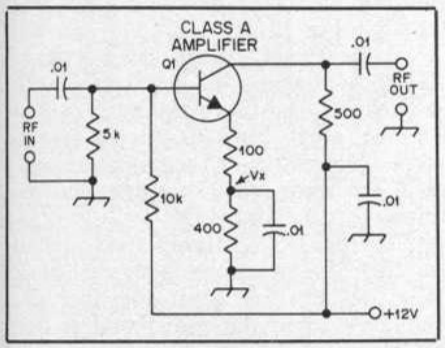


Fig. 8 – Class A amplifier using emitter degeneration.

The original voltage gain of 5 could be regained by replacing the collector resistor with a large inductor (i.e., an rf choke). An inductor is merely a component which resists any change in current flowing through it. (Note the analogy of an inductor to a capacitor. The L is to current what a capacitor is to voltage, with regard to circuit behavior.) With an inductance supplying the dc current to the collector, but resisting any changes in current, all signal current must flow into the external load. In this case, the load would be the 500-ohm input to the next stage.

The simple amplifier could be modified further by the addition of an emitter follower, as shown in Fig. 9. Since the emitter follower has a bypassed collector with the output signal taken from the emitter, we have a stage with *unity* voltage gain, but a very high input resistance.

From earlier calculations, we found the dc collector voltage of Q1 to be 8.7 volts. Hence, the emitter potential of Q2 is 0.7 volts less than this, or 8.0 volts. The current in the emitter of Q2 is 20 mA. When a drive signal is applied to this two-stage amplifier, the emitter of Q2 will follow the base, being 0.7 volt lower. For positive-going excursions of the output, signal current will be supplied to the external load and to the 400-ohm emitter resistor from Q2. On negative-going output excursions, however, current is pulled out of the external load resistance and is allowed to flow into the 400-ohm emitter resistor. In this case, the maximum current we can handle on the negative-going excursions is 20 mA, peak. In general, the standing dc current in the follower must exceed the peak signal current that the emitter follower is required to deliver.

Shunt Feedback

The amplifiers just discussed use *emitter degeneration,* or *series feedback*. Another type of feedback that is quite useful is shunt feedback, and is used typically with operational amplifiers. An example of an rf buffer amplifier using shunt feedback is shown in Fig. 10.

Recall that a silicon transistor has an input offset of about 0.7 volt. That is, the base of a conducting transistor is 0.7 volt above the emitter. Also, note that a common (grounded) emitter amplifier is an *inverting* amplifier. This means that an increase in base voltage leads to a decrease in collector voltage. With these ideas in mind, let's analyze the circuit of Fig. 10.

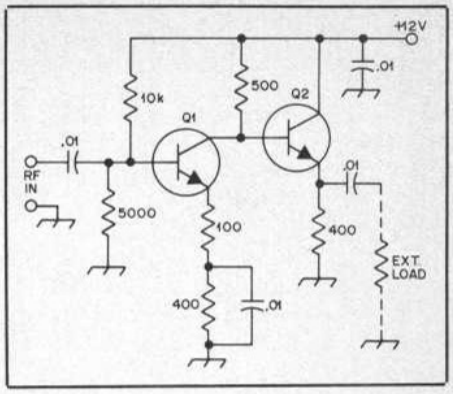


Fig. 9 – Class A amplifier followed by an emitter follower.

With the emitter of Q1 grounded, the base potential must be 0.7 volt. This means that there will be 0.7 mA of current flowing through R1. Where can this current come from? It certainly can't be coming from the transistor – dc current flows *into* the base of an npn transistor rather than *out* of it. The current must be supplied by R2, a 5-kΩ resistor. This resistor must also supply the base current to Q1. This current, as we will show, is small enough in comparison with the 0.7 mA that we can ignore it. With 0.7 mA flowing in R2, we must see a voltage drop of 5 kΩ × 0.7 mA, or 3.5 volts across R2. The output dc voltage at the emitter of Q2 must therefore be 4.2. Again, noting that Q2 will have a 0.7-volt offset, the collector of Q1 must be at 4.9 volts.

Consider now an input signal applied to the amplifier of 0.1 volt peak (0.2-volt pk-pk or 70-mV rms). As the input signal increases from zero to +0.1 volt, the current through R3 will go from zero to some positive value. This current would tend to flow into the base of Q1. However, this would cause the collector voltage of Q1, and hence the amplifier output voltage, to drop dramatically. This drop leads to a decrease in the current flowing in the feedback resistor, R2. The output voltage will drop until the net current flowing into the *node* at the base of Q1 is just enough to keep the potential of the base at 0.7 volt.

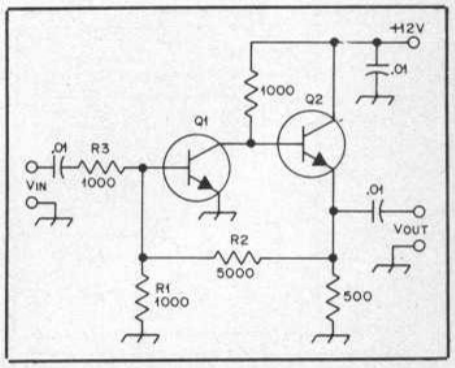


Fig. 10 – Rf buffer using shunt feedback.

That is, the effect of the input signal is to replace current flowing in the feedback resistor with current flowing from the input resistor. The input voltage is maintained at 0.7 volt in this amplifier when the output drops from the dc value of 4.2 volts to 3.7 volts. The voltage gain is

$$G_v = \frac{\Delta V_{out}}{\Delta V_{in}} = \frac{(3.7 - 4.2)}{0.1 - 0} = -5 \qquad \text{(Eq. 2)}$$

The minus sign indicates that the amplifier is inverting.

Note that the gain depends upon the resistors and not upon transistor characteristics. That is G_v = R2 ÷ R3 = $-R_{fb} \div R_{in}$. Also, the potential at the base of Q1 has been maintained essentially at 0.7 volt because of the feedback. This means that the input resistance looking into the base is virtually zero. The input resistance of a shunt fed-back amplifier approximates the value of the input resistor (R3). This well-defined input-resistance characteristic is independent of the load effects at the output, making such an amplifier ideal for buffering and isolation purposes.

Consider, finally, what would happen at the output if we were to increase the load, or ask the follower for more output current. This might correspond to keying a following stage. Owing to the feedback, the output voltage will adjust until the input offset of 0.7 volt at Q1 is maintained, with Q2 delivering whatever current is needed to do this. Hence, output impedance is reduced by shunt feedback.

The examples discussed here have demonstrated the use of series and shunt feedback. Admittedly, the analysis was highly simplified. What is, perhaps, surprising is that in many cases the simplistic analysis presented is more than adequate for design purposes. Many of the buffer amplifiers used in the projects described later were designed by using these methods rather than a more elegant approach.

Irrespective of the accuracy of the analysis, we can certainly use the results qualitatively to improve our intuition about feedback circuits. This will guide us in our experimental efforts. Negative feedback, (series or shunt) will always *decrease* the amplifier gain. It will also increase the bandwidth. Series feedback will have the effect of raising the input impedance while shunt feedback will decrease both the input and output impedances. Feedback amplifiers will be discussed in more detail in the chapter on ssb methods.

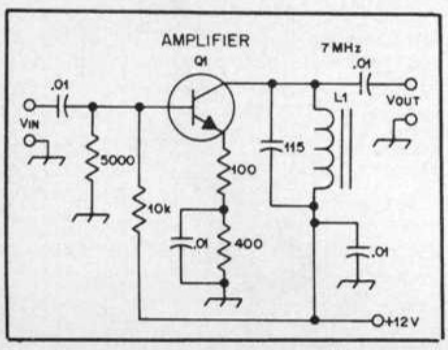


Fig. 11 — Class A amplifier with emitter degeneration and a tuned collector circuit.

Tuned Buffer Amplifiers

The previous circuits used resistive loads. However, most buffer amplifiers will be tuned. The use of resonant circuits improves the performance in a number of ways. Higher gain is possible, selectivity is introduced into the response of the circuit, and finally, higher power outputs are possible, since a high standing current can be used while maintaining a high collector voltage.

In this section, we will extend the designs described earlier to the case of tuned output loads. The rudimentary details of how a tuned circuit is treated analytically and how it is used for impedance matching will be presented. The first example is shown in Fig. 11. This circuit is nearly identical with that discussed in Fig. 8 where emitter degeneration was introduced.

Before considering the behavior of the amplifier of Fig. 11, we should review the nature of a simple tuned circuit. A toroidal inductor has been used. Toroids have distinct advantages for the experimenter. First, the magnetic field of a toroid is contained almost completely within the core. As a result, minimal magnetic energy from the tuned circuit will couple into other parts of the circuit to cause instability. This is not the case for a solenoidal inductor. The second advantage is that the inductance of a toroid is described by a simple and quite accurate equation which can simplify things for the designer. Knowing the number of turns (N) on a toroid the inductance is $L = KN^2$, where K is a proportionality constant. For the Amidon T-50-2 core used in our design example, the constant is 5 nanohenrys (nH) per turn squared. Thus, the inductance of a 30-turn winding on this core is L = 5 nH/t^{-2} $(30t)^2$ = 4500 nH = 4.5 μH. Data are presented in the appendix for a number of popular toroid cores available to the amateur.

This amplifier will be operated at 7 MHz. The capacitance required to resonate the inductor on 40 meters is given as $C = 1 \div (2\pi f)^2 L$. In this case, C = 115 × 10^{-12} farad, or 115 pF. In practice, one might use a 180-pF mica-compression trimmer capacitor. Alternatively, a low-capacitance variable could be paralleled with a fixed-value mica capacitor.

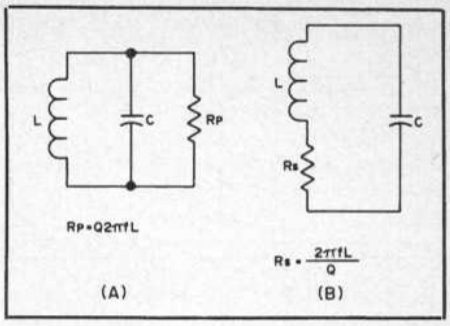


Fig. 12 — Schematic representation of circuit losses.

We have now described the superficial details of our circuit by specifying the inductance and capacitance. However, additional information is needed for circuit analysis.

If a quantity of energy is injected into a tuned circuit, that energy will remain *stored* for a reasonable time. A voltage across the capacitor will cause current to flow in the inductor. However, current flowing in the inductor will lead to a voltage being developed across the capacitor. If there were no loss elements in the tuned circuit the energy would remain stored forever. Any real tuned circuit, however, does have losses. In the hf region and at lower frequencies the predominant losses are associated with the inductor.

The presence of losses leads us to define a pertinent term, Q, which is a figure of merit for a resonator. Formally, Q is defined as the total energy stored in a tuned circuit, divided by the energy lost in one cycle of oscillation. It may be shown mathematically that this Q is also related to the bandwidth by $Q = f \div \Delta f$, where Δf is the 3-dB bandwidth.

For circuit applications, still another means is needed to model the losses of a tuned circuit. This can be done by assuming that our real and lossy tuned circuit is replaced by a perfect one with a resistor, either in series or parallel with the inductor. Using this representation it may be shown that these resistors are related to the Q and the inductance by $Q = R_p \div 2\pi fL = 2\pi fL \div R_s$. Schematics showing these loss resistances are presented in Fig. 12.

If the tuned circuit has no other elements attached to load it, the Q realized is called the unloaded Q, or Q_u. On the other hand, if energy is extracted from the LC combination and used for some other purpose, the resulting Q is the loaded Q.

For the T-50-2 toroid used in our amplifier example, the typical Q at 7 MHz is 150. (Q is a dimensionless number.) Since the inductance is 4.5 μH, the parallel-equivalent loss resistance, R_p, is given as $R_p = Q2\pi fL$ = 29.7 kΩ. If we were to shunt the tuned

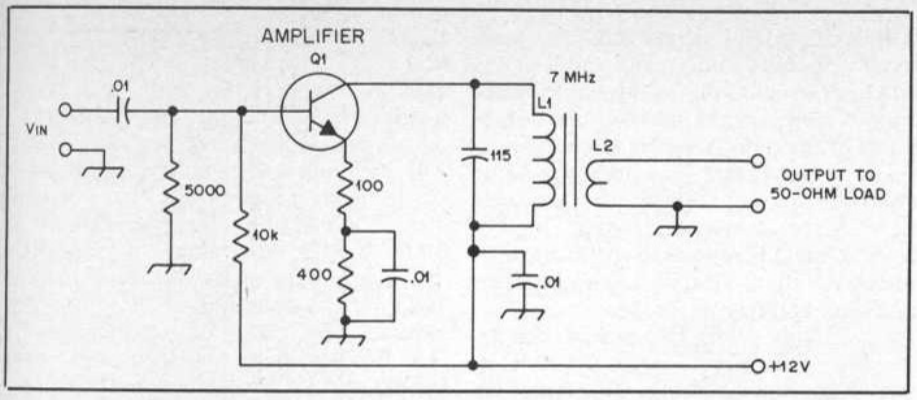


Fig. 13 — Output coupling from a Class A amplifier using a toroidal transformer in the collector circuit.

circuit with an external 5-kΩ resistor the net parallel-equivalent resistance across the coil would be 4.28 kΩ. Hence the *loaded Q* would be 4.28 k ÷ $2\pi fL$ = 21.6. The loaded Q is *always less* than the unloaded Q.

How do we treat this parallel combination of an inductor, a capacitor and a resistor when they appear in a circuit? In general, it would be necessary to consider the parallel combination of all of the impedances in order to arrive at a suitable equivalent impedance for use in an analysis. However, at resonance the case is simplified considerably, for the parallel capacitor and the inductor have the effect of canceling each other, in terms of reactance leaving the parallel resistor as our equivalent impedance. Indeed, this is the definition of resonance.

We are now in a position to return to the original amplifier of Fig. 11 and to calculate its gain. At resonance, the tuned circuit appears to be a 29.7-kΩ resistor. The voltage gain of the circuit is 29,700 ÷ 100, or 297. This gain is extremely high. In fact, it is so high that the chances of instability are very good. Ignoring this potential problem, we note that this high gain is obtained while keeping 12 volts of bias on the collector, and several mA of current flowing. This could not be realized without a tuned circuit.

In order to extract some energy from the output of the amplifier, assume that a 5-turn link is wound over the toroidal inductor (see Fig. 13). An asset of toroids is that almost unity coupling is provided between various windings on the core. It was this unity coupling that led to the simple N^2 inductance formula described earlier. Another feature is that impedances terminating one winding are transformed to the other winding according to the *square* of the ratio of the turns. Hence, if a 50-ohm resistor is placed across the 5-turn link, this has the same effect as a parallel load resistor across the tuned circuit where: $R_L = (30 \div 5)^2 \times 50 = 1800$ ohms. This external load appears in parallel with the 29.7-kΩ resistor which represents the core losses, resulting in a net load of about 1.7 kΩ. With this load, the voltage gain of the circuit (collector voltage divided by base voltage) is 17, a high but probably stable value. Further, the loaded Q of the resonator is $Q_L = R_L \div 2\pi fL$ = 1,700 ÷ 198 = 8.5, where R_L is the net load resistance. The loaded bandwidth will then be about 800 kHz.

Assume now that the amplifier is excited by a 0.1-volt peak signal at the input. The ac signal at the collector will be 1.7 volt, peak. The rf collector current is just 1.7 V ÷ 1.7 kΩ, or 1 mA. Since this is well below the dc current standing in the stage, the linearity should be excellent.

Since the turns ratio on the tuned transformer is 6:1, the voltage across the 50-ohm load resistor is just 1/6 the collector voltage, or 283-mV peak.

If this amplifier were driven from a low-impedance source, the net voltage gain would be only 2.8, and we would not consider this to be much of an amplifier. However, the buffering is quite good since the input resistance was 2 kΩ (see previous section). If the input to this amplifier were impedance matched, the gain would be a little over 25 dB — a very respectable value.

Power Output

It is interesting to calculate the maximum power output which can be obtained from this stage while maintaining Class A operating conditions. In the example outlined above, we saw an ac signal on the collector of 1.7-volts peak. The dc collector potential was 12 volts. Hence, the instantaneous voltage on the collector would vary from 10.3 to 13.7 volts at a 7-MHz rate. Note that the collector potential exceeds the +12-volt dc bias.

The emitter dc voltage was 3.3 volts. As an approximation we will neglect the fact that the emitter is not totally bypassed. The maximum signal voltage we could expect to see on the collector would be (12 − 3.3) = 8.7 volts peak. That would be the signal which would cause the transistor to just go into saturation on negative peaks. The positive voltage peak would be (12 + 8.7) or 20.7-volts peak. The pk-pk signal is just twice 8.7, or 17.4 volts.

If our amplifier is to stay linear (barely) during this voltage excursion, the current must be fluctuating from zero to twice the dc value of 6.6 mA. Now we ask what the proper load resistance would be to obtain these swings in voltage and current simultaneously. This is given again by Ohm's Law, as R_L = (8.7-V peak) ÷ (6.6-mA peak) = 1.32 kΩ. If we increased our link from 5 to 6 turns, the load presented to the collector would be 1.25 kΩ, a close approximation. With this load the maximum power output will be given as $I^2 R = (6.6 \times 10^{-3})^2 \times 1.250$ kΩ = 54-mW peak, or 27-mW rms.

In this example we will not, in practice, be able to obtain quite this much output. This is because on negative-going output peaks, when the transistor approaches saturation, the emitter voltage will rise above the 3.3-volt dc level. On the other hand, if this amplifier were slightly overdriven, the dc collector current would rise above the 6.6-mA bias level and some additional power output could be obtained. This nonlinear mode of operation is often used in cw applications.

In most linear applications it is desirable to maintain Class A operating conditions where the stage current does not fluctuate with drive level. While ssb

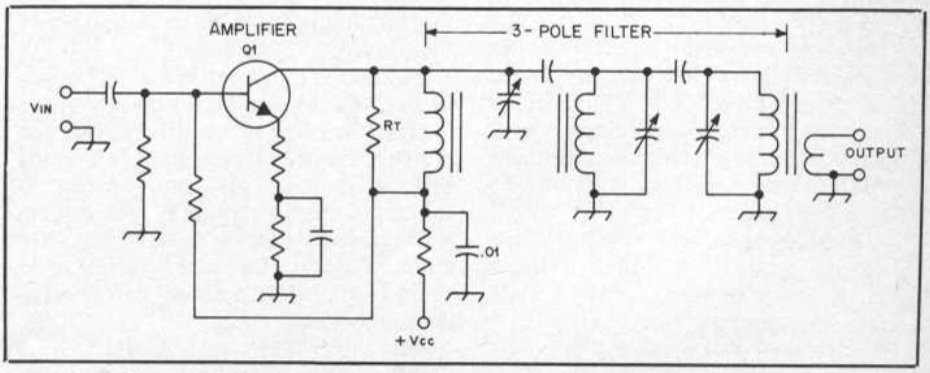


Fig. 14 — Buffer amplifier with a three-pole output filter.

is the obvious application, in some cases this is also advisable during cw operation. The reason is that a linear amplifier tends to maintain the selectivity inherent in the resonators. On the other hand, if the amplifier is allowed to saturate additional loading occurs across the tuned circuits, and that decreases the selectivity. That can have the effect of increasing spurious output, especially when the stage is driven by a mixer or frequency multiplier.

Load Resistance

Now that we have analyzed an example we are in a better position to ask a more general question. That is, what load resistance should be used for a specific power output? If we consider only amplifiers which have bypassed emitters, the load required is that resistance which will allow the collector to fluctuate with a peak voltage excursion equal to the difference between the supply, V_{cc}, and the emitter voltage, V_e. For a given resistance the power delivered to that load is ${V_{rms}}^2 \div R$, or ${V_{peak}}^2 \div 2R$. Solving this for the load resistance, we have $R_L = (V_{cc} - V_e)^2 \div 2P_o$. The Class A amplifier should be biased to a current equal to the peak signal current, which is $I_{dc} = 2P_o \div R_L$.

Although we have been discussing Class A amplifiers predominantly, the expression for load resistance is quite general and applies as well to Class C amplifiers. In the typical Class C power amplifier, the emitter is at dc ground, leading to the well-known expression $R_L = {V_{cc}}^2 \div 2P_o$.

The last two expressions may be combined to show that the maximum efficiency of a tuned Class A amplifier is 50 percent. In practice, efficiencies near 30 percent are more common, especially if good linearity is desired, as would be the case with ssb.

It is desirable when seeking selectivity to use circuits other than a single *LC* combination in the output of a Class A stage. An example might be the first buffer amplifier following a mixer in a heterodyne exciter of the kind that might be used in a ssb transmitter. Although rather complex, the same basic principles apply. The gain of the amplifier is determined by the impedance "seen" when looking into the input of the more complex filter.

In some cases a filter may require a given termination at its input in order to provide the desired selectivity. In this case it may be required to resistively terminate the collector of an amplifier in order to present the proper load to the following filter. The gain of such an amplifier will depend upon the resistor and filter characteristics. This situation is illustrated in Fig. 14. The appendix contains a catalog of two- and three-section filters for the amateur bands. They are suitable for such applications.

Earlier, it was shown that shunt feedback in an amplifier has the effect of decreasing the output resistance of that circuit. Therefore, by careful use of feedback the output impedance of an amplifier can be adjusted to provide the proper input termination for a multi-pole filter. Designs of this kind are practical and will be covered in the ssb chapter.

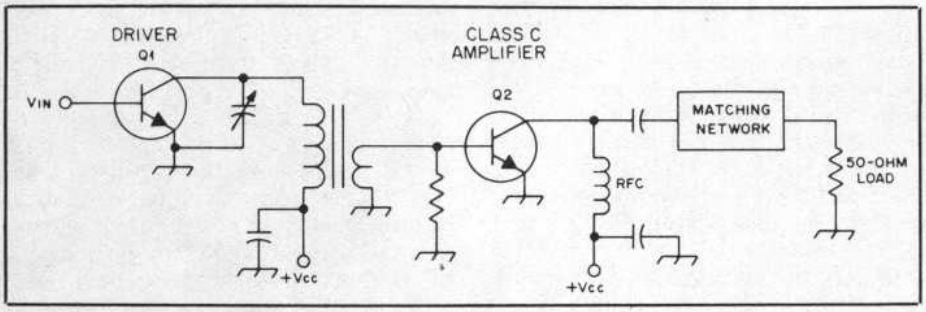


Fig. 15 — Class C amplifier which is link coupled to a driver stage.

The Medium-Power Class C Amplifier

When high output power is desired for the final stage of a QRP transmitter, or from the driver in a medium-power cw transmitter, a Class C amplifier is usually chosen. While these stages lack the envelope linearity needed for ssb, they offer high power gain, high power output and good efficiency. In this section, amplifiers with an output up to 2 watts will be considered.

A Class C amplifier is defined as one where collector (or plate) current flows for less than half of the drive cycle. The normal transistor amplifier operated with no reverse bias on the base is actually a Class C amplifier, since there is in effect a built-in bias in the transistor. That is, the base voltage must exceed 0.7 volt positive before conduction occurs.

At this point, we will shift gears slightly, away from a simple analytic treatment toward a more empirical approach to the design problems. In general, the small-signal approximations used in the previous text are not too accurate in the description of Class C amplifiers. Nonetheless, we can extend our previous understanding to describe qualitatively a high-power Class C amplifier. For example, the gain of such a stage is still determined by the high-frequency beta of the transistor, which is in turn, a function of the f_T of the device. The maximum output power will be limited by the load impedance we present to the collector. As the dc current level is increased, and hence, the power level of the stage, the input resistance decreases.

Shown in Fig. 15 is a Class C amplifier coupled by a link from an earlier stage.

Starting at the input, the first consideration is to determine the turns ratio of the input transformer. If the base of the power amplifier were a simple resistive input, as is essentially the case with a Class A stage, the turns ratio would be determined by the simple impedance-matching criterion outlined earlier. However, the input to the Class C amplifier is not, in the general case, a *pure* resistance. At low frequencies a better model for the input would be a silicon diode with some series resistance. Unlike the usual silicon diode, however, the one used in our model (representing the power transistor) will have a low reverse-breakdown voltage. Typical values will be 3 to 5 volts. The input link must be chosen to deliver current to the base on positive peaks of the driving voltage. However, the open-circuit voltage from the link must be low enough that the reverse breakdown of the diode is not exceeded. The driver should have a power output consistent with the expected power gain of the Class C stage. That is, if an output of 1 watt is desired, and we expect a gain of 16 dB in the Class C amplifier, we should have 25 mW available from the previous stage.

If the reverse breakdown of the base-emitter junction of the power amplifier is exceeded, the result is not an instantaneous catastrophe: The transistor does not go up in smoke. However, the long-term result is just as devastating. Prolonged operation with the input diode being switched into breakdown will lead to a deterioration in the current gain of the transistor. Hence, the power output will continually drop off.

This effect can be observed easily with small-signal transistors operating at very low frequencies. A simple experiment can be done to demonstrate it. Start with an inexpensive plastic transistor, for this is a destructive test. Measure the dc beta of the transistor at a collector current of, say, 10 mA. Then

apply a reverse bias to the emitter-base junction with current limiting to keep the "Zener" current at around 10 mA. Operate the transistor for about an hour in this manner. Then, again measure the dc beta. A degradation will usually be noted. Low-level transistors are often used as Zener-diode substitutes by operating the e-b junction as outlined. This practice is generally fine. However, once used as a Zener diode, the device should be retired from service as a *transistor.*

It is generally more difficult to observe this phenomenon at high frequencies. It is straightforward, however, if the experimenter is fortunate enough to have a high-frequency oscilloscope in his shop. This problem is generally limited to transmitters on the lower-frequency amateur bands, usually at and below 7 MHz. The reverse base breakdown is prevented by choosing carefully the turns ratio in the driving circuitry and by keeping the value of the shunt resistor at the base fairly low. A resistor in series with the base should be avoided.

Returning to Fig. 15, the resistor shunting the base serves two functions. First, it provides a load for the driver during the negative voltage excursions of the driving signal, and hence, prevents the reverse breakdown from occurring, as outlined. Second, it absorbs some drive energy that might otherwise find its way into the base. Since part of this energy could result from feedback in the amplifier as well as from the driver, the resistor decreases stage gain and tends to stabilize the amplifier. If instability is ever noted in a Class C stage, the first thing to do in order to "tame the beast" is to decrease the ohmic value of this shunting resistor.

As a rule of thumb with amplifiers operating from 12- to 15-volt supplies, the driving link is approximately 1/10 the number of turns used in the primary of the driver transformer. Typical values for the base resistor in 1- to 2-watt amplifiers is 18 to 100 ohms.

When the operating frequency of the amplifier is increased to roughly a tenth of the f_T of the transistor (or higher), the input of the transistor ceases to look like the simple diode model outlined previously. Charge-storage effects within the transistor make the input appear much more like a resistive input shunted with a capacitance. Modern transistors designed specifically for rf power applications have the input resistance and capacitance specified by the manufacturers. As odd as it may seem the reduced power gain and more stable input characteristics which occur at high frequencies often make it much easier to build amplifiers which operate toward the high end of the spectrum. (We've encountered many more stability problems with 1-watt amplifiers on the 160-meter band than we have seen at 144 MHz!)

Output Circuit

Designing the output circuit is similar to the procedure described for the Class A amplifier in the previous section. With no drive power present, the base of the Class C amplifier is at ground and the transistor draws virtually no current. Only when drive is applied does any collector current flow. This current in the collector will cause the voltage at the collector to depart from the quiescent value of V_{cc}. If we assume that the collector voltage varies from 0 to twice the V_{cc} level while delivering the desired output power, the load needed at the collector is given by the familiar relation $R_L = V_{cc}^2 \div 2P_o$.

There are a number of networks which can be designed to transform a 50-ohm termination to any desired practical resistance. These are outlined in chapter 4. For stages operating at power levels such that R_L is 50 ohms or higher, link coupled output networks can often be used satisfactorily. For higher powers other networks are recommended.

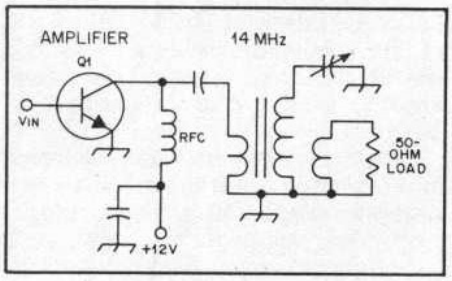


Fig. 16 — Example of a link-coupled output network.

As an example of a link-coupled output network consider the stage shown in Fig. 16. We will design for an output power of 1/2 watt at 14 MHz, and we will use a transformer with 15 turns as the major resonant winding on an Amidon T-50-2 toroid core. The inductance is $(15)^2 \times 5$ nH-t^{-2}, or 1.13 μH. This will resonate at 20 meters with a capacitance of 115 pF. We will design for a loaded Q of 6 and a supply voltage of 12.

The inductive reactance of the coil is 99.4 ohms. With a Q_L of 6, we thus want the 50-ohm link to present a parallel resistance across the coil of $Q_L 2\pi fL$, or 596 ohms. Noting that impedances transform in proportion to the square of the turns ratio, we see that the output link should have 0.29 the number of turns of the main winding (4.35 turns). We will use 4 turns.

With a 12-volt supply, the load we want to present to the transistor is $V_{cc}^2 \div 2P_o$ = 144 ohms. The turns ratio between this winding and the 50-ohm winding is $\sqrt{144 \div 50} = 1.7$. Since the 50-ohm winding has 4 turns, we calculate that the transistor winding should be 6.8 turns. A 6- or 7-turn link will do the job. The parallel resistance representing the unloaded Q of the coil has been neglected since the loaded Q of 6 is much less than the inductor unloaded Q.

Once a suitable network is designed and implemented, the maximum power output is defined. To realize this output the stage must be driven adequately. If the drive is less than that required for full power output, the collector voltage will not swing from ground to twice V_{cc}, but something less, centered around V_{cc}. Such operation is typical for linear amplifiers used for ssb applications. However, for cw use, the amplifier is usually driven to full output since this results in maximum efficiency.

Components

The collector rf choke is a component which is often treated too casually. The choke should have a low dc resistance, for any IR drop in the choke will subtract from the available supply voltage. The inductance of the choke should not be excessive. Too much inductance will cause resonances to exist with the capacitors in the output network which are much lower than the output design frequency. Since the typical transistor has a gain which is increasing dramatically at lower frequencies, these resonances can lead to instabilities. A reasonable rule of thumb is that the output rf choke should have a reactance at the operating frequency which is between 5 and 10 times R_L. An additional (and wise) precaution is to parallel the usual 0.1-μF bypass capacitor with an electrolytic capacitor of around 10 μF.

The general criteria for selecting transistors for amplifiers of this kind are f_T, breakdown voltage, power dissipation and maximum current. The f_T should be well above the operating frequency; however, not by too much. It is sometimes quite difficult to use vhf power transistors on the lower hf bands due to the tremendous gain available, which causes instability problems. The collector breakdown voltage should be twice the supply to be used, although this rule can sometimes be violated because the transistor is not conducting during the period when the highest collector voltages are present. In general, the power dissipation of the transistor should be *at least* as high as the output power desired. This also implies that a heat sink may be necessary if it is needed to realize the dissipation rating. The maximum collector-current capability of the transistor should be at least

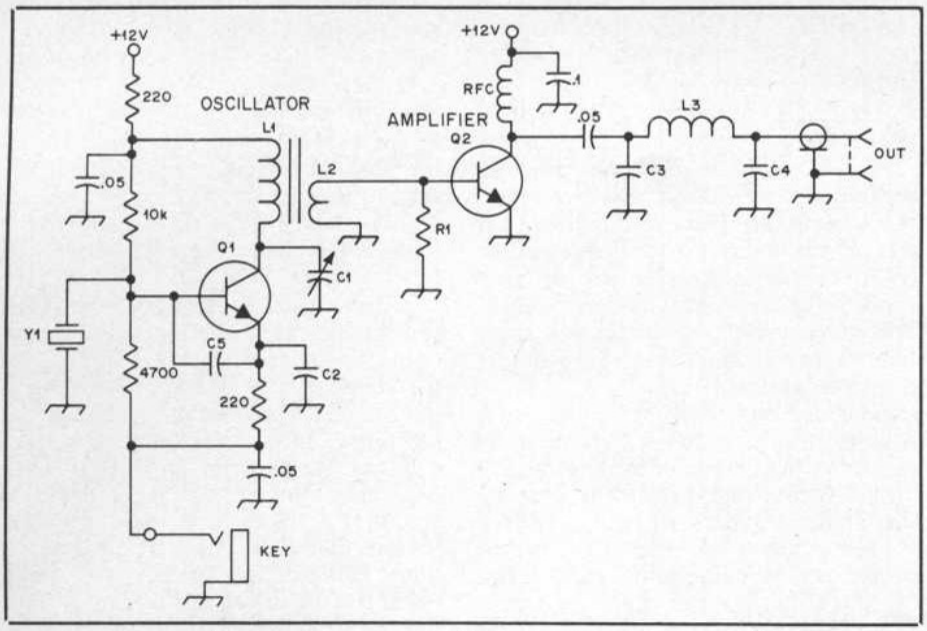


Fig. 17 — Schematic diagram of the universal QRP transmitter. Resistors are 1/2-watt composition. C1 is a trimmer capacitor. C3 and C4 are silver-mica capacitors. Remaining capacitors are disk ceramic, 50 volts or greater. See text for Q1, Q2 types. Component values not on the diagram are listed in Table 1.

twice the dc current expected.

The efficiencies of Class C amplifiers in the 1- to 2-watt category vary considerably, but are usually around 60 percent. Efficiencies of over 75 percent are not uncommon. If the efficiency is under 50 percent, a better output transistor might be in order.

A Universal QRP Transmitter

The ideas outlined previously can be applied to the design of a simple two-stage transmitter for the hf or 160-meter bands. Although the seasoned QRP operator may scoff at a non-VFO transmitter, the use of crystal control can lead to simplicity as well as an uncompromisingly clean signal. The design lends itself well to the later addition of a VFO.

The essential details of the transmitter are shown in Fig. 17. Only a few of the component values are specified on the schematic. The rest vary from band to band and are summarized in Table 1.

The transmitter is near the ultimate in simplicity, consisting of a crystal-controlled oscillator driving a single-stage power amplifier. The crystal oscillator is keyed in all versions but the 10-meter one. In the output stage a pi network is used to match the 50-ohm antenna to the collector of the amplifier. In this case the word "match" is a bit of a misnomer, for the network shown presents no impedance transformation. When the output is terminated in 50 ohms, a load resistance of 50 ohms is presented to the collector of the final. However, the network acts as a low-pass filter to attenuate harmonics. The maximum power output which can be expected is about 1.44 watts when using a 12-volt supply. Indeed, the measured output is just about 1-1/2 watts on all bands except 10 meters, where the power is still over 1 watt.

In the schematic, a capacitor, C5, is shown from the base of the oscillator to the emitter. This capacitor is used only on the 160- and 80-meter bands.

On the bands up to 14 MHz, fundamental-mode crystals were used. In the test units, HC-6 type plated crystals were chosen. Several surplus FT-243 style 7-MHz crystals were used in the 40-meter unit. They all oscillated readily and keyed well.

On the 10- and 15-meter bands, third-overtone crystals were required. Since most 40-meter crystals will oscillate readily on their third overtone, the 7-MHz crystals also operate well in the 15-meter transmitter. When FT-243 crystals were used, the 21-MHz output was excellent, as was the keying.

The reader will note that only one design is presented for both the 10- and the 15-meter bands. The circuit functions well on both of the bands by merely retuning C1, the capacitor which resonates the crystal oscillator.

A minor problem was observed with the 10-meter design. It was found that there was a slight chirp when the oscillator was keyed. This was eliminated by rebiasing the stage for reduced output, but the drive to the final was then inadequate. Best 10-meter operation of this rig resulted from keying only the final, as shown in Fig. 18. Here, a pnp transistor is used as a switch, allowing the key to remain at ground potential. An even better solution would be to modify the design with a keyed Class A buffer between the oscillator and the output amplifier. This approach was taken in a 6-meter transmitter described at the end of this chapter.

The number of transistors which can

Table 1

	C1	C2	C3	C4	C5	L1	L2	L3	R1	RFC
160 M	400 pF MAX	1800 pF	1800 pF	1800 pF	360 pF	73t No. 28 T-50-2	8t	30t No. 26 T-50-2	18Ω	50 μH
80 M	400 pF MAX	100 pF	750 pF	750 pF	200 pF	43t No. 26 T-50-2	5t	21t No. 22 T-50-2	39Ω	25 μH
40 M	180 pF MAX	100 pF	470 pF	470 pF	—	35t No. 26 T-50-2	4t	14t No. 22 T-50-2	39Ω	15 μH
20 M	60 pF MAX	33 pF	210 pF	210 pF	—	27t No. 24 T-50-6	3t	12t No. 22 T-50-6	47Ω	15 μH
15/10 M	60 pF MAX	33 pF	105 pF	130 pF	—	17t No. 24 T-50-6	3t	9t No. 22 T-50-6	47Ω	15 μH

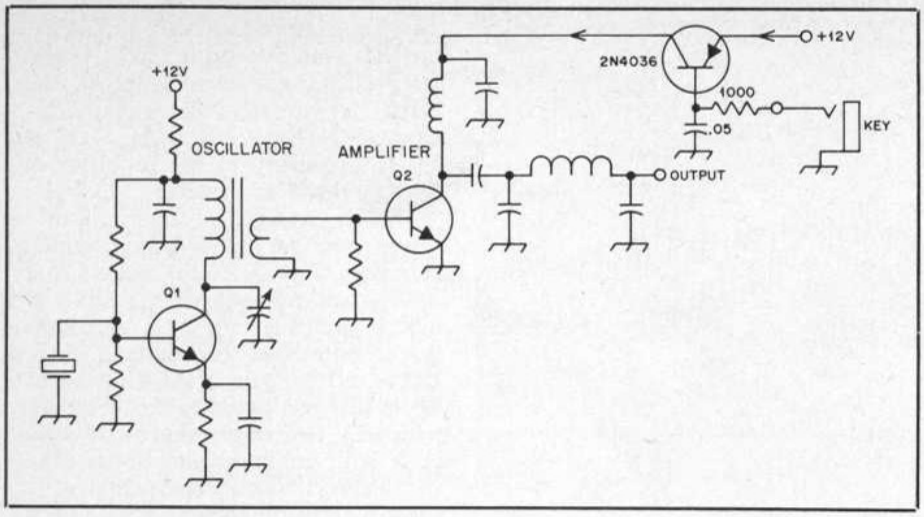


Fig. 18 — Modification of the keying circuit for the 28-MHz version of the QRP transmitter.

be used in this design is nearly endless and is growing daily. In test units built, the oscillator was either a 2N2222A or a 2N3904. These devices are inexpensive and readily available. Other good candidates would be the 2N4124, 2N3641, 2N3563, 2N3866, 2N3692 or 2N706, to mention only a few.

In all of the units built, the final amplifier was a Motorola 2N5859. This is a TO-5 device similar to the RCA 2N5189. The differences between the two are minimal. The 2N5859 is perhaps a bit "hotter," with the 2N5189 being slightly more rugged. A small smokestack type of heat sink was used on the output transistor in all units.

When 2N5859s were used, they appeared to operate reliably when the transmitter was terminated properly in a 50-ohm antenna with a VSWR of under 2:1. However, the potentially destructive testing procedure to be described in the following section showed that the transistors would not survive a severe mismatch. A Motorola 2N3553 was substituted in several of the units and the power output was the same. The output transistor could not be destroyed under the worst mismatch that could be found. Additionally, the higher power dissipation and breakdown voltage ratings of the '3553 allow the transmitter to be operated at up to 28 volts, a level at which several watts of output power can be obtained. In this case, careful heat sinking is required. While this transistor is specified as a vhf power device, the cost is only $2.30 in single lots.

Shown in Fig. 19 is a printed-circuit layout for the universal transmitter. This board is single sided and is only 2 X 3 inches. The builder may want to make the board slightly larger if it is to be used on 160 or 80 meters, where the components are bigger. Likewise, the 10-meter version could be reduced in size, if desired.

Tuning of this family of transmitters is straightforward. After the unit is built and carefully inspected to ensure that the parts are in the proper slots, a dummy load, power supply and crystal

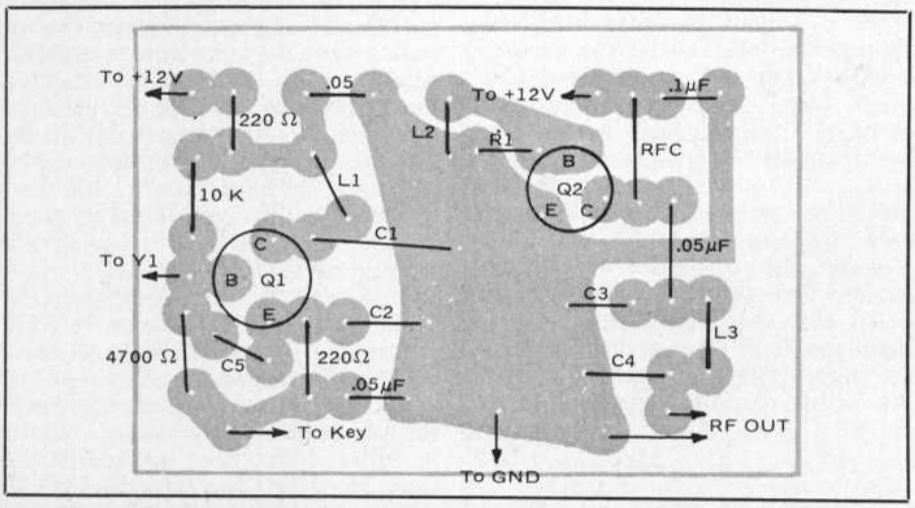


Fig. 19 — Scale layout of the universal QRP transmitter pc board.

are connected. Some means of monitoring the transmitter output is needed. Such a QRP power meter is described in a later chapter, although a suitable substitute would be a 51-ohm, 1-watt resistor as the output termination with a VTVM/rf-probe combination for measuring output. Ideally the power supply should be current limited to around 0.25 A. With the power on and the key closed, the oscillator tank is tuned for maximum power output. The keying is monitored in the station receiver, just to be sure it's clean. That's it! Debugging, should problems occur, is covered in the next section.

Fig. 20 shows a photograph of the 160-meter board. Shown also is a box which contains the 20-meter version. The packaged unit contains a slide switch which transfers the antenna and the 12-volt supply to the final stage during transmit intervals. The rear of the box contains a pair of bnc coax connectors for the antenna and receiver as well as banana jacks for the dc power input. Dc voltage is always applied to the crystal oscillator. This allows the operating frequency to be spotted by merely hitting the key.

The 20-meter version was used for a couple of months of casual operation in the spring of 1974 by W7IYW. Although only one crystal was available, contacts were made with KH6, UAØ, JA, ZL, VK, KX6 and G as well as with a few stateside amateurs. The 3-element Yagi antenna (at 80 feet) and an excellent location helped. Similar results can be expected with a dipole or ground plane vertical in a typical location, although the contacts will not come as easily, and the reports are sure to be down by a couple of S units.

Construction Methods, Testing Techniques and "Bandaids"

In the earlier sections of this chapter, the discussion has been rather basic with emphasis on the fundamentals. One design example was presented in the preceding section, but not very much has been said about construction and debugging of solid-state circuits. There are a few rules which make a profound difference in the performance obtained.

Once a design has been transferred to a hardware form, it still may not function exactly as originally envisioned by the designer. Indeed, it is only in rare cases that debugging of some sort is not required. Some problems will be covered in this section. The reader is referenced to a *QST* paper on this subject which is especially good.[2]

As one reads the various amateur publications, he soon realizes that

[2] DeMaw, "How to Tame a Solid-State Transmitter," *QST* for Nov. 1971.

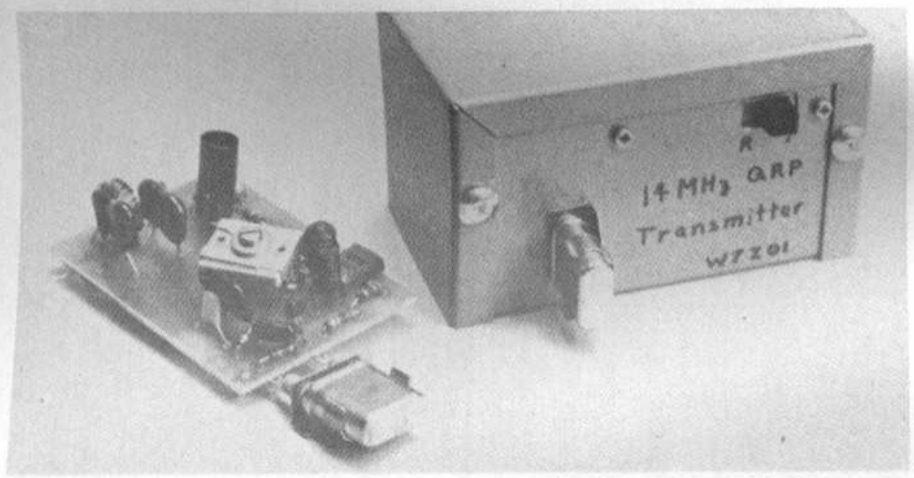


Fig. 20 — Photograph of the assembled QRP transmitter for 20 meters. At the left is a 160-meter version.

almost all of the equipment built by today's amateur experimenter is fabricated on etched circuit boards. One might assume that this is done merely to allow easy duplication and repeatability of performance and to impart a pleasing appearance. After all, that's what the professionals do. In reality there is a bit more to it than this, especially when rf circuitry is concerned. A proper pc layout has the major advantage of presenting a low impedance return to ground wherever it is desired. This characteristic provides ample justification for using pc-board methods when building rf circuits!

The amateur magazines and reference books contain data for layout and etching of pc boards. These will not be repeated in detail here, for the methods are straightforward and easy to apply in the home. The builder is, however, cautioned to keep the basic goal of proper grounding in mind when designing a layout, even if it means that some of the aesthetic qualities of the board might be sacrificed.

The best way to ensure a clean ground plane for an rf circuit is to use double-sided board (copper on both sides). This may present a minor problem to those who frequent only the local outlets where single-sided board is sold. However, when surplus outlets are investigated one finds that double-sided board is the rule rather than the exception. If modern electronic equipment is studied by the reader, he will notice that single-sided boards are seldom used. The norm these days for densely packed circuits is multi-layer boards, often containing up to 6 or even more individual layers.

The easiest way for the amateur to use double-sided board, especially if one-of-a-kind boards are being built, is to use one foil for nothing but the ground plane. All soldering pads and runs are on the other side of the board. Once the side containing the "meat" of the circuit is etched and washed, and the resist is removed, the holes are drilled in the board. Then, a large drill is used as a counter-sink to remove the copper from around all of the holes on the solid-foil side of the board. Then the components are inserted, being mounted on the ground-plane side of the board, and soldering can commence. Whenever a connection to ground is desired, the component is soldered directly to the ground foil with the shortest possible lead length on the part. Numerous examples are shown in the photos throughout the book.

All of the etched boards used in the illustrative examples of this book were built in the home lab. The resist materials used were small pads or strips of Scotch brand electrical tape, or masking tape. In some cases a resist-ink pen was used. Ferric chloride was used as the etchant. The resist material used to protect the ground plane during etching was a layer of enamel spray paint, or full-width strips of masking tape or Scotch electrical tape.

A series of *QST* articles featured circuit boards which are not etched.[3] Instead, a hacksaw was used to cut a series of shallow grooves in the board, through the foil. This leaves a checkerboard pattern of copper islands to which components may be soldered. Some of the equipment described in later chapters was built using a modification of this method. Double-sided board was chosen, and a hacksaw was used to create the matrix of islands. However, the components were mounted on the groundplane side of the board. Holes were drilled in exactly the same way as with an etched board. If the copper islands are kept fairly small, the method seems to work quite nicely up through the vhf spectrum. The hacksaw can even be used for some "casual" micro-strip uhf circuits for the 432-MHz band. No matter which method is chosen, keep the grounds short and clean, and many of the problems outlined next will never occur!

As an example of an rf circuit to debug, consider the rf power amplifier shown in Fig. 21. We'll assume that a driving power from a VFO or mixer of 1 mW is available, and that an ultimate power output of 2.5 watts is desired. Hence, a total gain of 34 dB is needed. While this gain could easily be obtained with only two stages, the use of a third stage will give us a much better chance of realizing unconditional stability. Two Class A stages are used to drive a Class C power amplifier. The base of the final amplifier is matched by means of an *L* network, and a single pi network is used for the output impedance transformation.

The first step in testing such a design is to get a source of rf drive. Although the VFO which will eventually be used could serve to excite the amplifier, an equal approach would be to use an existing QRP transmitter. For example, one of the units from the preceding section would do the job, except that the power output is too high. This is easily remedied with a step attenuator of the kind outlined later on. The attenuator is adjusted for 1 mW of output, and we are ready to proceed.

Only the first stage is attached to the signal source. The output link from L1 is attached to a short length of coaxial cable which is run to a simple power meter. Power is applied to the first stage and C1 is tuned for maximum power output. Here is where some of the more subtle effects may rear their ugly heads. As C1 is tuned there should be a single well-defined peak, assuming the tuned circuit cannot be tuned to a harmonic of the input frequency. If the tuning is not smooth and well defined, the stage may be self-oscillating. The power output should disappear completely, of course, when the input drive is removed. At this time the stage should be checked for spurious output. The best amateur instrument for this is probably an absorption wavemeter. Another useful tool is a bc-band receiver. If low-frequency oscillations are taking place, spurious responses may be heard while tuning from 550 to 1650 kHz.

The Bandaids which may be applied to cure unwanted oscillations are many and varied. If spurious outputs (spurs) are noted in the low-frequency region or near the operating frequency, they may often be eliminated by placing a resistor in series with the base and/or the collector of the stage, typically 10 to 22 ohms. Also, reducing the stage gain may help a great deal. In this case the gain

[3] DeMaw and McCoy "Learning to Work with Semiconductors," *QST* for April through November, 1974. DeMaw and Rusgrove, "Learning to Work with Semiconductors," *QST* for April through November, 1975.

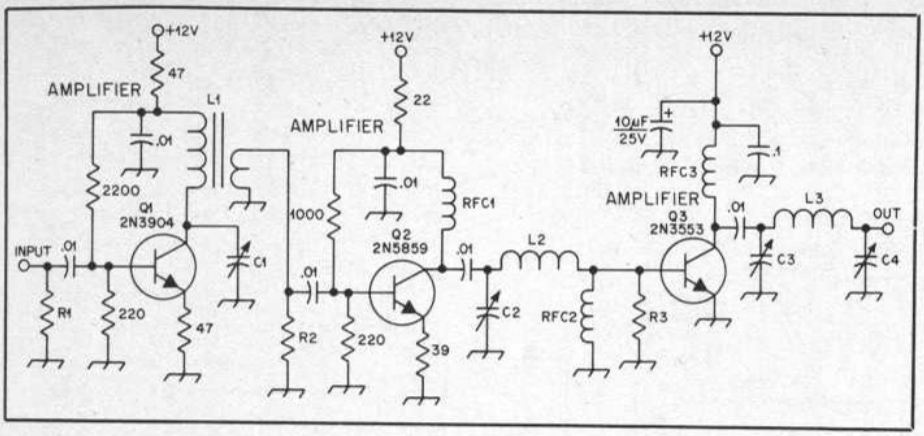


Fig. 21 — Circuit of a three-stage amplifier for use with text discussion of debugging.

can be lowered easily by increasing the value of the 47-ohm emitter resistor. Varying the value of R1 should have little effect when the stage is being driven from our 50-ohm attenuator. However, it may add greatly to the stability when the VFO is tied into the system later.

If vhf parasitics are observed with the wavemeter, they can be cured by means of the base or collector resistors mentioned above. Another solution is the use of a ferrite bead in either of these positions. If a clean layout is used, and proper bypassing is insured, vhf spurs are rarely a problem in hf transmitters.

Since we are using three stages in this amplifier, and ultimately need only a gain of 34 dB, probably a good amount for the first stage would be 13 dB. Hence, an emitter resistor which would yield an output of about 20 mW should be chosen.

Once the first stage is operating properly, the second stage is built and connected. Since its output is meant to drive the base of the final stage, probably the most effective way to test the system would be to build the final amplifier, but leave the output transistor temporarily out of the circuit. With power applied to the first two stages of the amplifier, the voltage is monitored across R3 with an rf probe and a VTVM. Typically, R3 will be approximately 39 to 56 ohms, or perhaps even less. C2 is tuned for maximum power delivery to R3. The tuning of C1 is also checked. As before, tuning should be smooth. If spurs are observed the same Bandaids are applied to the second stage. The power delivered to R3 should be around 200 mW. If this level is exceeded, the emitter resistor at Q2 can be increased in value. Also, R2 is chosen to obtain the desired output from Q2.

When the first two stages are operating properly, it will be time to add the final amplifier. Transistor Q3 is placed in the circuit, a 50-ohm power meter is used to terminate the rig, power and drive are applied, and the system is tuned. As before, all tuning is for maximum output. C2 will require retuning because the termination of the second stage has changed with the addition of the final-amplifier transistor. It may be desirable to increase the value of R3 in order to get more drive into the final amplifier. On the other hand, if there is the slightest sign of instability, the value of R3 should be reduced. Great care should be taken to ensure that the lead length of the emitter of the final stage is as short as possible. If the mounting method in a heat sink is such that a long lead is needed for Q3, make the connection with a relatively wide strap. A scrap of pc board or flashing copper can be used effectively for this. The 2N3553 used at Q3 has an f_T of 400 MHz. If the emitter lead were as much as half an inch in length, vhf oscillations could almost be guaranteed. They would be observable with a wavemeter coupled near the final amplifier. However, they might not be observed at the output port due to the low-pass nature of the output network.

If low-frequency oscillations are noted, they cannot be cured by adding the series resistance recommended for the first two stages, for such resistors would absorb too much power. The low-frequency spurs which might be occurring in the PA can be related to problems with the rf choke in the circuit. As suggested earlier, this choke should have a reactance (at most) of ten times the load resistance of the output stage. The electrolytic capacitor bypassing the supply to the last stage is then effective in killing the low-frequency spurs. If all else fails, a little resistance in parallel with the collector rf choke can be used to stop a low-frequency spur.

Most likely the amplifier is operating nicely now. If the foregoing verbiage seems extensive, it is because of our attempt to cover all bases. However, if careful construction practices are used (good grounding) and the gain-per-stage is kept down to a reasonable level, stability and smooth spur-free operation should be obtained without much trouble. When the board is mounted in the metal enclosure, and the transmitter is driven by the VFO (or whatever), it may be necessary to check the alignment again, and ensure that stability has been retained. The pc board sitting on the bench may behave in a cleaner manner than the same board inside a metal enclosure. This is because energy may be radiated from the free board. However, when inside the metal box that radiated energy is *reflected back* into the box where it may interact with various parts of the circuit to cause unstable operation.

One final test remains before the rig can be considered finished and ready for use. This is related to the output termination used for testing. Typically, the load is a 50-ohm resistor of appropriate power dissipation, along with some means for rf-voltage detection. This load, if purely resistive, looks like 50 ohms at *all* frequencies. Hence, the transmitter is terminated properly, not only at the operating frequency but at other frequencies. On the other hand, the typical antenna appears to be 50 ohms (or thereabouts) at only one, or perhaps a few discrete frequencies. Elsewhere within the spectrum, it will be highly reactive. In some cases this can lead to instabilities, especially if emitter degeneration is used in the final stage.

Testing for this condition is realized easily with a common ham-shack accessory — a Transmatch or antenna tuner. Connect the transmitter to an absorptive type of bridge (see later chapter for details). The output of the bridge is fed to a Transmatch for the band in use, with the output of the Transmatch connected to the previously used 50-ohm wattmeter. The Transmatch is tuned for a balanced condition of the bridge. Then the bridge is removed from the system. An rf probe and VTVM are connected to the output of the transmitter and power is applied to the system. The rf voltage observed should be nearly identical to that observed with the broadband termination. When the various adjustments in the transmitter are tweaked, they should produce a smooth, stable variation in output, identical to that observed with the broadband termination. Any departures from these results are indicative of stability problems. Incidentally, if the power observed in the wattmeter is not close to that measured earlier, the Transmatch may need a bit of work.

If the experimenter has both courage and a replacement for the output transistor, there is another worthwhile experiment which can be done with the

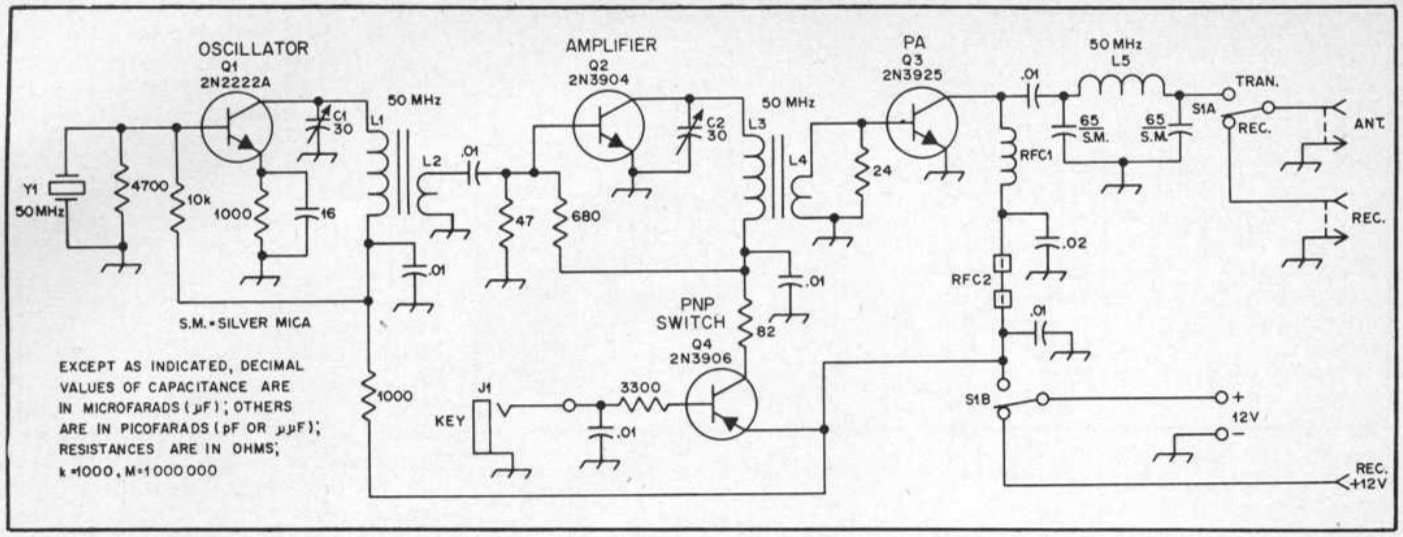


Fig. 22 — Schematic diagram of the 6-meter QRP transmitter. Resistors are 1/2-watt composition. Capacitors are disk ceramic unless otherwise noted.
C1, C2 — 30-pF trimmer capacitor.
J1 — Two-circuit phone jack.
J2, J3 — Phono jack or SO-239 fitting.
J4 — Insulated jack for 12-volt input.
L1 — 10 turns No. 28 enam. wire on Amidon T-37-6 toroid core.
L2 — 1 turn same wire over L1 winding.
L3 — 9 turns No. 28 enam. wire on T-37-6 toroid core.
L4 — 2 turns same wire over L3 winding.
L5 — 6 turns No. 22 enam. wire on T-50-6 toroid core.
RFC1 — 15-μH choke.
RFC2 — Two Amidon miniature ferrite beads on wire lead.
Y1 — 50-MHz, third-overtone crystal (International Crystal Mfg. Co. type EX or equiv.).

test setup outlined. The game is quite simple: Grab the controls on the Transmatch and twist them to grossly improper settings. That is, settings which would yield very high VSWR at the input to the Transmatch. If the output transistor survives this rather violent and potentially destructive test, the project is pretty well finished. It is then safe to use the transmitter in a fairly casual way, even with in-line type VSWR bridges for antenna adjustments. If the output stage does not survive, the blown transistor is replaced. The transmitter is still quite usable, but should be used only with something close to a proper termination. Furthermore, the rig should be used only with Transmatches which are tuned with an absorptive bridge.

A 6-Meter QRP CW Transmitter

When the universal QRP rigs described earlier were built, it was intended to include a 6-meter version along with the other designs. However, when construction was started, several problems occurred. The most severe one was that the 50-MHz crystal oscillator could not supply sufficient output to drive the final stage when it was biased to yield good stability. The next attempt was to try to combine two of the single-sided boards used for the rest of the "universal" rigs. This also caused problems — the grounding was not good enough. Finally, it was decided to build a separate rig for 6 meters, apart from the designs for the lower bands, using double-sided board. The result is shown in Figs. 22 and 23.

A three-stage circuit is used for the 6-meter design. The crystal oscillator is a third-overtone circuit of the kind outlined earlier. The emitter resistor was increased from the usual 220 to 1000 ohms in order to reduce the crystal current and improve the stability. The crystal oscillator is not keyed.

Oscillator output is taken from a one-turn link and is applied to a keyed Class A buffer. This stage operates with fairly high gain due to the grounded emitter. Bias stability is achieved through the negative feedback at dc realized with the biasing scheme shown. The current is 15 to 20 mA, and the rf output from the buffer is about 50 mW.

The final amplifier is a Class C 2N3925. This device is specified for 12-volt operation as an rf power amplifier in the 175-MHz region, and is capable of several watts of output. In this design, the power output was held down to a bit over 1 watt in order to permit battery operation. The design of this stage was performed using the guidelines offered earlier, with the exception that some additional decoupling was included in the form of a pair of

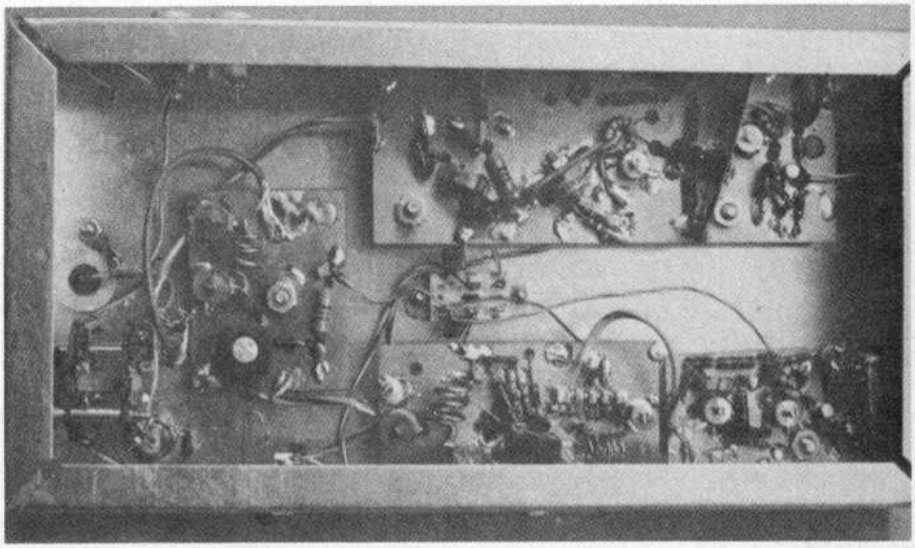


Fig. 23 — Photograph of the vhf cw transmitter. The circuit board at the upper right contains the 1-watt 50-MHz transmitter of Fig. 22. The crystal oscillator is at the right end of the board and the output circuit is at the left. The stud-mount transistor is bolted to a small piece of circuit board, the latter of which is soldered to the main board. The remaining three pc boards form a similar design for the 2-meter band. The wafer switch accommodates T-R switching and band changing.

ferrite beads on the collector supply line. A 2N3553 would probably serve nicely as a substitute for the output transistor used.

The transmitter was enclosed in a small aluminum chassis box along with a switch for transmit-receive switching. Also included in the box is a crystal-controlled transmitter for 144 MHz. The design is similar to that described for 50 MHz. They can be seen in the photograph of Fig. 23. An alternative approach to packaging would be to include a simple crystal-controlled receiving converter in the box with the transmitter.

Using only a 2-element Yagi antenna, this transmitter has yielded several contacts over 1000 miles away. The reports were always complimentary. A frequent comment was that the rig provided "The cleanest cw signal ever heard on 6." Perhaps this is not as much a testimonial for this transmitter as it is a commentary on the poor-quality cw signals often found on 6 meters!

Chapter 3

More Transmitter Topics

Emphasis in this chapter will be on the more elaborate and practical considerations of transmitter design. We will treat VFOs, frequency multiplication and mixing – all means of adding frequency coverage to a transmitter, beyond that which is reasonable for the crystal-controlled rigs in the previous chapter.

Several design examples are given. They are intended to illustrate the methods outlined in the text and are also suitable for duplication. Additional examples are given in later chapters.

Building and Using VFOs

In chapter 2 emphasis was placed on the use of crystal-controlled oscillators. The approach is ideal from a cost and circuit-simplicity outlook. However, there are occasions in operating where a VFO provides a necessary flexibility which is not possible with VXOs and simple crystal oscillators. A VFO permits greater effectiveness during low-power work, especially if crowded band conditions prevail. However, inclusion of a VFO compromises miniaturization and battery drain. Also, frequency stability is more difficult to realize when a VFO is used in preference to a crystal oscillator – notably when the equipment is designed for field use where the temperature environment may change markedly. It is of paramount importance, therefore, to design for the best stability possible with ordinary circuits and components.

VFO Design Philosophy

As the radio amateur reviews the ham magazines, he finds a large number of VFO designs. The more extensive the search, the less rigid may be the conclusions reached. Some of the popular circuits have names like Colpitts, Clapp, Seiler, Vackar and Hartley. Many of these designs are given in standard reference books.

VFO performance requirements are varied and many, and depend upon the intended application. For use in a typical transmitter the major need is that the oscillator have good long-term stability. By *long term* we mean that the oscillator should have a constant average frequency for periods of a second and longer. For critical receiver applications, and for most transmitters, the oscillator should have good short-term stability and low noise. In this chapter we have concerned ourselves mainly with the long-term stability matter – the "wanderies." The problems of short-term stability, phase noise, and the "wobblies," as well as a-m types of noise, are covered in the receiver chapters.

Fig. 1 shows the block diagram of an oscillator. The basic components are a resonator (tuned circuit), an impedance-matching network, an amplifier and a second impedance-matching network. The two matching networks may include phase-reversing properties, depending on the nature of the amplifier. Typically, these networks are merely

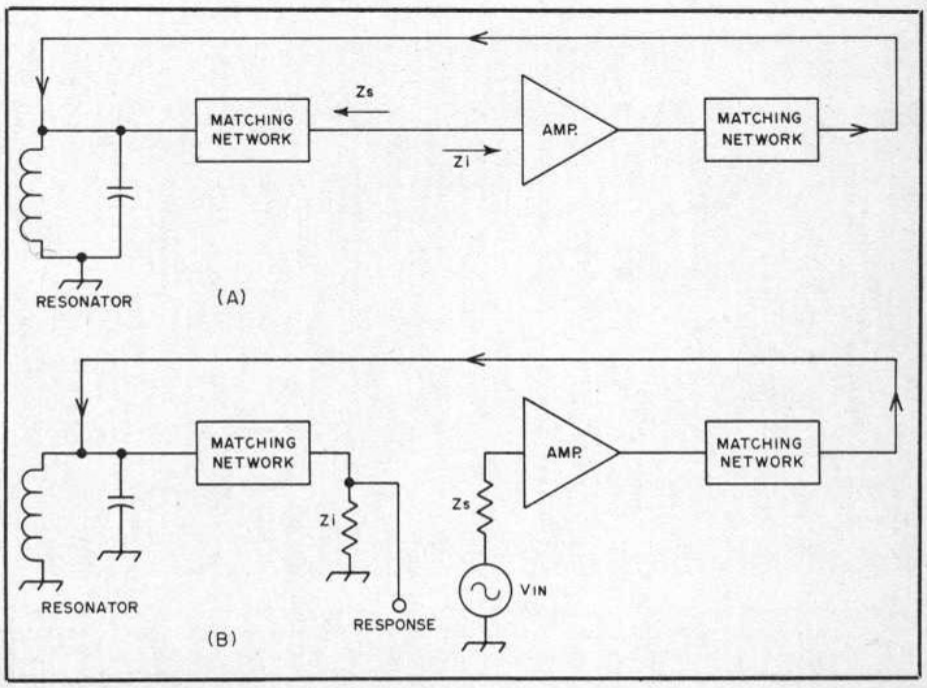


Fig. 1 – Block diagram of an *LC* oscillator.

capacitors between the tuned circuit and the amplifying bipolar transistor or FET. The usual tuned circuit contains an inductor and capacitors, with the impedance-matching capacitors often being part of the resonator. Furthermore, the parasitic capacitors of the transistors are, to some extent, part of the resonator. The better oscillators are those which use high-quality components throughout, such that changes in temperature do not change the frequency of the resonator. The sources of heat which can cause this drift include not only the external environment, but the heat created by the rf energy circulating in the loss elements of the tuned circuit.

There are a number of methods for matching into and out of the tuned circuit. The gentlemen who have studied the various methods now have their names attached to the configuration that they found most interesting. In general, the configuration chosen by the builder is secondary to considerations of component quality and fundamental design.

The conditions for oscillation in a circuit of the type shown in Fig. 1 are described by the Barkhausen criterion. These conditions are related to Fig. 1B where the feedback loop is opened at one point. Assume that the loop is opened at the input to the amplifier and that a signal is applied to the input of the amplifier. The conditions for oscillation (when the loop is closed later) are (1) The output signal after amplification and filtering should have an amplitude which is greater than the original signal and (2) the phase of this output signal should be exactly the same as that of the input signal.

The first criterion specifies the gain needed in the amplifier. It's just that amount required to overcome the losses in the resonator. The second criterion defines the frequency of oscillation. The oscillator operating frequency will be that at which the phase shift in the resonator is proper to fulfill the requirement.

These are general conditions. They have applied here to the design of VFOs. However, they may also be applied to crystal oscillators, or to audio oscillators which use *RC* networks. While we will not attempt such an analysis in this text, many of the guidelines which follow result from a careful application of this theory, along with empirical observations.

Design Guidelines

Some of the more common VFO circuits, such as the Colpitts and Clapp varieties, can be made stable enough for most amateur work, and the output levels will be ample for ordinary applications. This is true even though unity-voltage-gain buffering may be used after the oscillator. In cases where additional driving energy is required, a simple Class A low-level amplifier can be included.

The solid-state VFO offers a distinct advantage over a tube type of VFO – reducing heating. The efficiency is better, and 60-Hz fm is not as likely to occur in a transistorized VFO, because there are no filaments to heat. Finally, miniaturization is greatly enhanced by employment of transistors as opposed to tubes in VFO circuits.

It is beyond practicality to describe all of the VFO circuits which can provide good stability. Additional data not offered here can be obtained from *The Radio Amateur's Handbook*. We shall emphasize several circuits, all of which are easy to build and adjust.

Long-term stability is attainable by adhering to some simple guidelines. Rule No. 1 is to use only that amount of feedback necessary to assure quick oscillator starting and minimum pulling by external load changes. Rule No. 2 is to bias the oscillator at a power level no greater than that needed for a specific output amount – generally, 10 mW or less of output power. The higher the dc input power to the oscillator, the greater the internal heating. Therefore, the rf currents flowing in the frequency-determining components (*L* and *C* units) will be more pronounced. The higher the rf current flow, the greater the internal heating of capacitors and magnetic core materials. This leads to unwanted changes in operating frequency. So, in the present vernacular, *keep it cool*!

Components

Temperature-stable capacitors should always be used in a VFO except where drift compensation is desired. Among the best low-cost capacitors available to amateurs are the dipped silver-mica and polystyrene varieties. The latter, generally speaking, have a much tighter tolerance to changes in temperature, and are highly recommended. Silver-mica capacitors are rather unpredictable with regard to temperature effects. Some may exhibit positive drift, while others from the same manufactured batch may change value in the opposite direction. Still others may be very stable in the presence of changing temperature. This phenomenon has not been noted when using polystyrene capacitors in ARRL lab experiments. NP0 ceramic capacitors are used in some VFO circuits, single or in combination with micas or poly units, with good results.

The VFO inductor should be rigid and of relatively high *Q*. Whenever possible, the coil should be without a magnetic core (iron or ferrite), as temperature changes will affect to some degree the permeability of the core material. Such changes will shift the inductance and, hence, the frequency. No matter what materials are used, the wire on the coil form should be cemented securely to the form by means of *Q* dope or some other high-dielectric compound. The inductor should not be mounted near any component that radiates heat.

Toroidal inductors (magnetic core) are perhaps the most prone to changes in characteristics as the ambient temperature shifts. They should be used only in VFOs that will be operated in a fairly constant temperature environment. The most stable toroid core material is the SF kind (Amidon type 6). Slug-tuned inductors are a better choice than toroids. They should be chosen and operated so that the slug barely enters the coil winding at resonance. The farther into the winding the slug is placed, the more pronounced the unwanted temperature effects.

The variable capacitor in a VFO should be mechanically stable, and should rotate smoothly with minimum torque applied. A double-bearing type of capacitor is recommended. Brass or iron capacitor plates are less subject to temperature effects than are aluminum plates. Air-dielectric trimmers are preferred over those with ceramic or mica materials.

If a bipolar transistor is used as the active element in a VFO, it should have an f_T considerably higher than the VFO operating frequency, say, a 250-MHz f_T for a 7-MHz VFO. This minimizes phase shift in the transistor. Furthermore, the small-signal beta should be 10 or greater to minimize the amount of feedback needed for reliable oscillation. When an FET or MOSFET is used in a VFO, it should also be a high-frequency device, and the transconductance should be 2000 or higher. A 2N4416 or MPF102 JFET is suitable for VFOs operating below 30 MHz. An RCA 40673 or 3N200 is fine for VFOs which employ MOSFETs.

Other Considerations

Lead lengths in a VFO should be as short as possible. Excessive lead lengths become unwanted "parasitic" inductances. In circuits where very low values of *L* are used, long connecting leads become a significant part of the tuned circuit and can degrade the *Q*. As a result, the VFO may not oscillate, or when the chassis is stressed the leads may move and cause shifts in the operating frequency. In some designs the circuit-board foils become part of the tuned-circuit inductance, so the layout should be planned for short, direct connections.

Double-sided pc boards are not recommended in VFOs . . . at least not in the frequency-determining part of the circuit. The pc board, if double-sided,

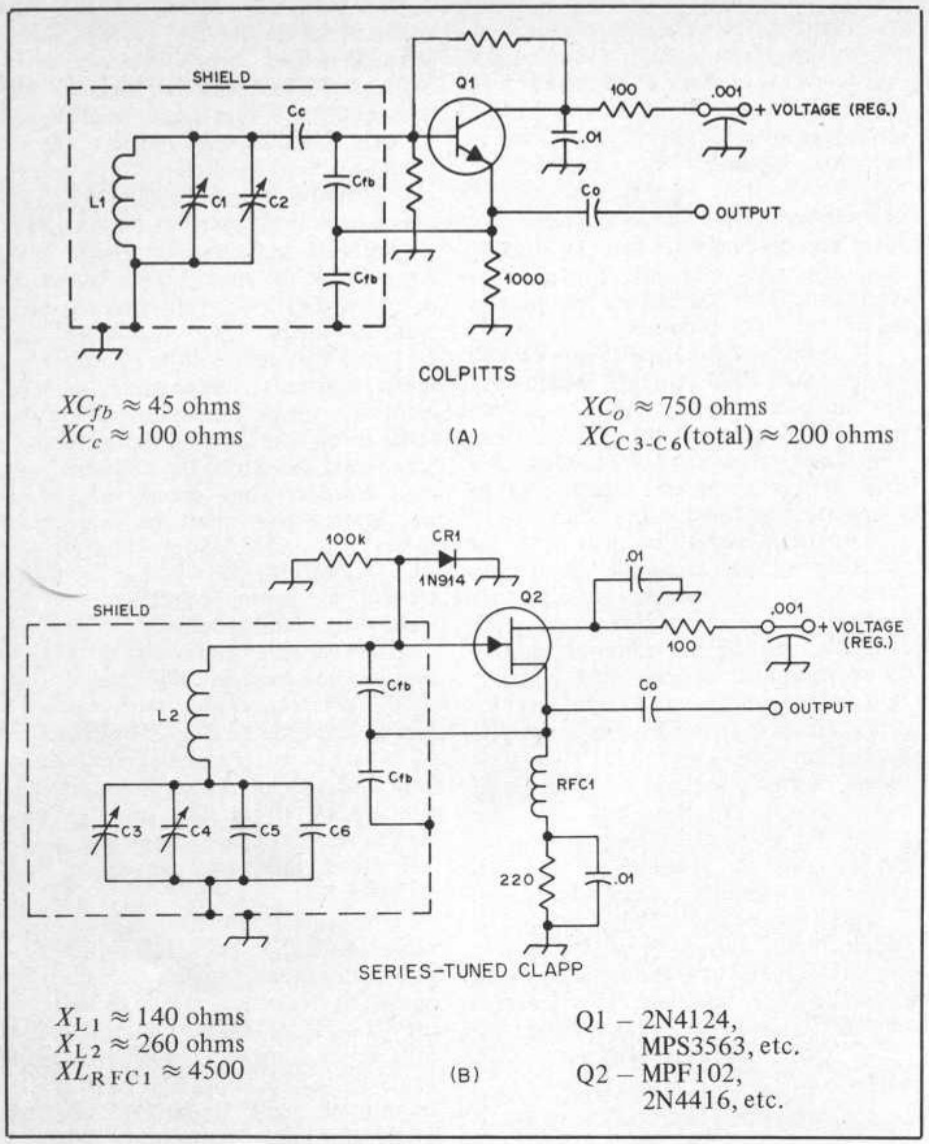


Fig. 2 – Schematic diagram of two common VFO circuits. Reactance values are given for the critical components.

provides numerous unwanted capacitances wherever the circuit foils are formed. The dielectric material of the pc board (phenolic or glass epoxy) is not especially stable with regard to changes in temperature and humidity, and drift can result from the double-sided board approach. Also, capacitors formed in that manner will be relatively low in Q, and this can lead to poor oscillator performance.

Finally, the VFO should be contained in an enclosure to isolate it from stray rf which originates in other parts of a receiver or transmitter. This also provides thermal isolation. Unwanted rf coupling can seriously affect VFO performance. It should be noted that VFOs can oscillate at some lf, hf or vhf point other than the desired one, while still performing at the chosen frequency. The amplifier following a VFO should be operated into a constant load impedance and the output examined by means of a high-frequency scope (if available). The waveform should be nearly a pure sine wave. Random oscillations above the VFO operating frequency will be superimposed on the fundamental waveform. The measures prescribed earlier (ferrite beads, bypassing, addition of low-value resistors) for correcting instability are applicable in VFOs as well.

The operating voltage for a VFO should be regulated and well filtered. In most amateur circuits a Zener-diode regulator will suffice. It is not uncommon to see regulation applied to the VFO *and* its buffer stages. The practice is a good one to avoid load changes caused by voltage fluctuations, as they may pull the oscillator. Three-terminal IC voltage regulators are also well suited to this application. Some of the newer units are no larger than a plastic transistor.

Examples which show two of the oscillators under discussion are given in Fig. 2. Approximations are given for the reactances of L and C in significant areas of the circuit. These are ball-park values, and will enable the builder to scale either circuit to a selected tuning range in the hf or mf spectrum. At Fig. 2A, C1 can be the main tuning capacitor, with C2 serving as a padder for calibrating the VFO to the dial readout. The absolute values of C1 and C2 will be dependent upon the size of coupling capacitor C_c and both C_{fb} capacitors. It will be necessary to determine the combined series capacitance value of C_c and both C_{fb} units, then add that value to C1 and C2 to find the tuning range of the oscillator. L1 is a fixed-value component in this case.

Generally speaking, the output capacitor, C_o, should be as small in value as possible, consistent with adequate output voltage to excite the following stage (buffer or amplifier). The fixed-value capacitors just discussed should be polystyrene types for best frequency stability, but selected silver micas can be used if the builder is willing to solder-and-try until some stable ones are found.

The circuit of Fig. 2B shows a Clapp VFO which is a series-tuned form of the Colpitts. It has been proved quite stable when used from 1.8 to as high as 10 MHz. The advantage in using a series-tuned gate tank is that greater inductance is required than with the parallel-tuned type of tank. This means that stray inductances have less effect upon circuit performance – an advantage. At 7 MHz the circuit at A requires approximately 3 μH for L1. Conversely, the circuit at B will have an L2 value of roughly 6 μH at 7 MHz.

Capacitors C3 through C6, inclusive, are in parallel at the bottom of L2 in Fig. 2B. The advantage in using several capacitors instead of one or two is that the rf current is divided among them, which lessens the internal heating of any one capacitor. This greatly enhances stability. Similarly, the builder could use paralleled capacitors for the C_{fb} units for the same reason.

If the Barkhausen criteria for oscillation outlined earlier are examined, we see that they predict the signal in an oscillator will always be increasing. This is, of course, impossible. Something is required in any oscillator to limit the amplitude of oscillation.

In the FET oscillator of Fig. 2B, the output of the circuit is stabilized by means of diode CR1. The diode rectifies

the rf signal from the tuned circuit and charges the capacitors to some dc value. This bias reduces the gain of the amplifier until the output voltage is stabilized. The oscillator would operate without this diode. However, the limiting bias would then be developed in the gate-source diode of the FET. This not only tends to create harmonics in the output, but loads the tuned circuit. Further, since the source of the FET is not tied to ground, the oscillator will operate at higher amplitudes. The larger circulating currents in the tuned circuit will degrade stability.

With both circuits of Fig. 2 it is wise to apply the least amount of operating voltage practical. That is, use no more regulated voltage than is necessary to assure reliable operation and adequate rf output. The lower the voltage the better the stability, generally speaking. When FETs are used, the supply should exceed the pinch-off voltage of the device. A good voltage range is from 6 to 9, regulated. The tuned-circuit components should be housed in a shield enclosure, as shown by the dashed lines. It is good practice to enclose the entire oscillator circuit in a metal compartment when space permits.

Practical examples of VFO circuits are presented later in this chapter. Information concerning the design of buffer stages was provided in chapter 2.

Any of the circuits shown may be tuned with varactor diodes instead of the more common mechanically variable capacitor. There are, however, some problems which may occur. First, the diode should always be biased in such a way that the rf voltage does not cause the diode to conduct. The simplest way to realize this is to utilize two varactor diodes in a back-to-back arrangement, as shown in Fig. 3. While this arrangement decreases the net capacitance of the diodes by one-half, it prevents significant current from flowing in them. The second precaution that should be taken is to ensure that the variable biasing voltage is as clean and stable as possible. Any drift or noise on this controlling voltage will show up as instability or fm noise on the oscillator frequency.

Some Other VFO Circuits

Shown in Fig. 4 and the photograph is an adaptation of a Sieler-type oscillator developed by W2YM (*QST* for Dec., 1966). While silver-mica capacitors are shown in the circuit, we later replaced them with polystyrene units, resulting in an improvement in stability. The constants given are for 3.5-MHz operation.

While a MOSFET was used in the original W2YM circuit, this oscillator also functions well with a JFET. It may be scaled to a number of other frequencies. The constants for several

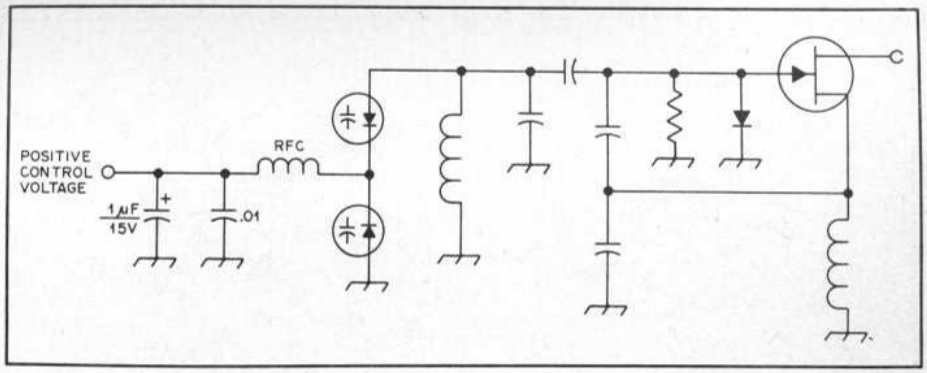


Fig. 3 — VFO circuit showing varactor-diode tuning.

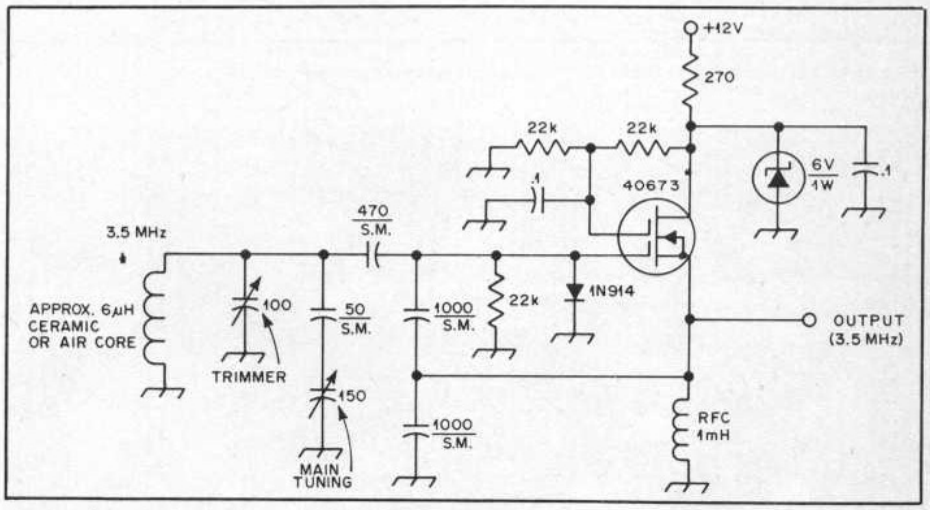
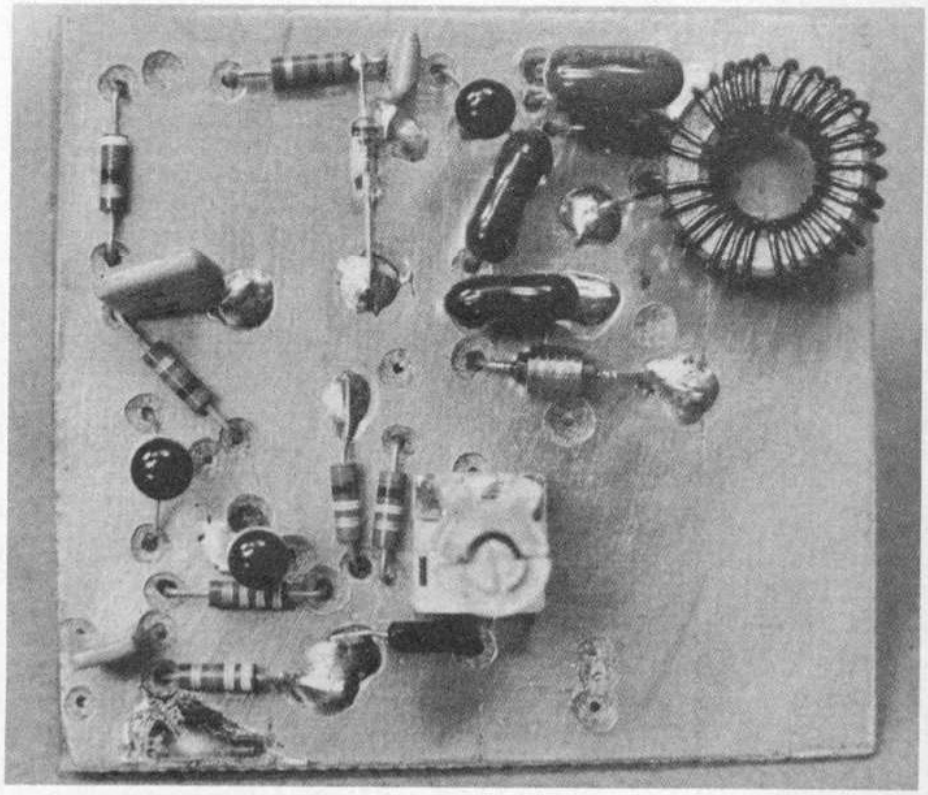


Fig. 4 — Circuit of a dual-gate MOSFET VFO.



Here is the simple 80-meter VFO. The T-68-2 toroid inductor is seen at the upper right, and the JFET oscillator is at the top center. At the lower left is a two-stage buffer amplifier with feedback. The air trimmer is switched into the circuit by means of a diode, providing a frequency offset function when desired.

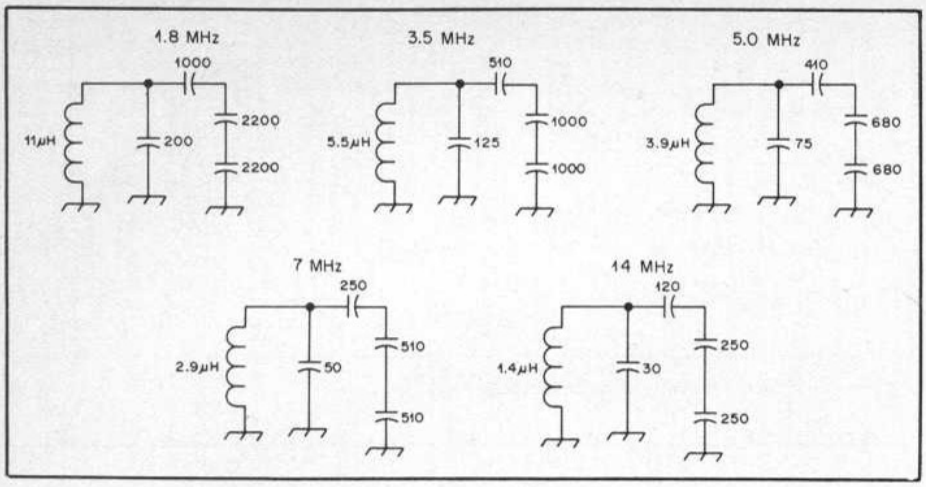
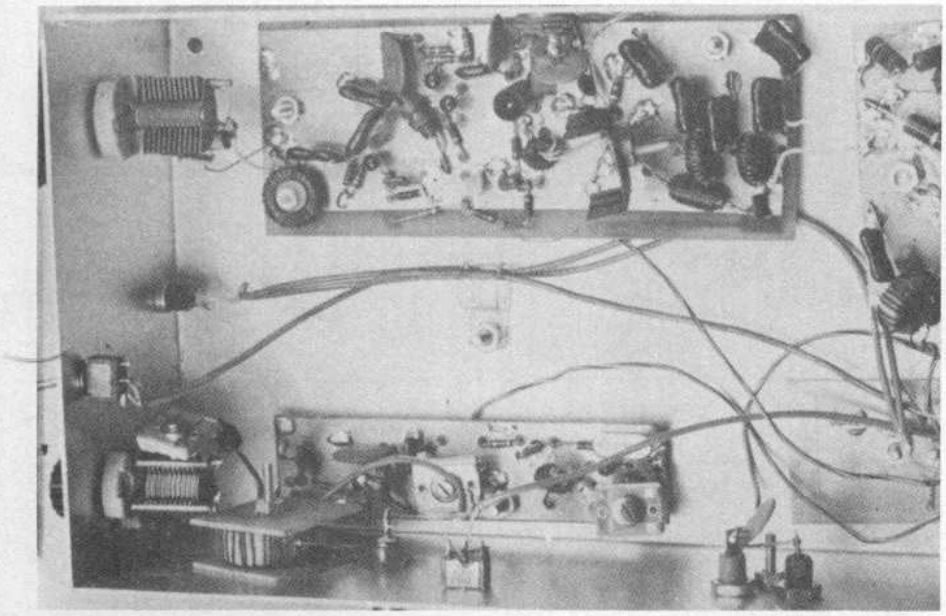


Fig. 5 — VFO *L* and *C* constants for various operating frequencies.



Layout of the 160-meter transmitter with VFO. The top circuit board contains the entire transmitter. The VFO section is at the left. Seen at the bottom of the photograph is a crystal-controlled 160-meter converter with a 7-MHz i-f. Front panel controls are for VFO tuning, VFO spotting, and T-R control. A receiver antenna trimmer is also on the front panel. The remaining circuitry is for a solid-state power amplifier and T-R relay.

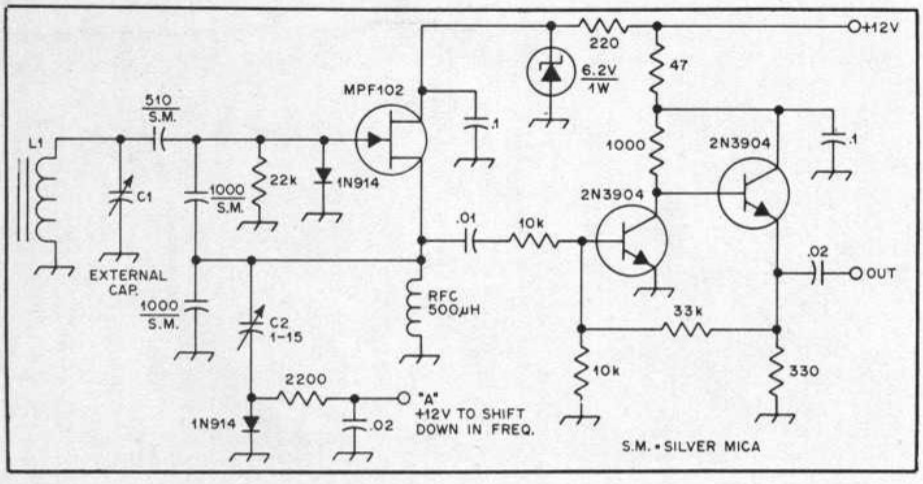


Fig. 6 — Schematic diagram of the 80-meter JFET VFO. C1 is the main-tuning capacitor, the value of which is selected for the desired tuning range. C2 is adjusted for the desired offset amount, and is an air-dielectric trimmer. L1 is a T-68-2 toroid core wound with 30 turns of No. 22 enamel wire.

other frequencies are shown in Fig. 5.

When miniaturization is more significant than extreme long-term stability, toroid inductors can be used. Shown in Fig. 6 is an 80-meter VFO which was developed for use in a compact portable transceiver (described later in the book). A JFET has been used in the W2YM circuit. An additional feature of this design is the inclusion of a diode switch to shift the frequency slightly. When the diode has no external bias applied at point A, the small variable capacitor, C2, will charge to a dc voltage such that virtually no current flows in the capacitor. However, when +12 volts are applied to point A, rf current will flow in C2, making it part of the resonant circuit. A decrease of up to 2 or 3 kHz can be realized, depending upon the setting of C2.

Shown in Fig. 7 is a simple Hartley oscillator. This circuit is of significance for two reasons. First, it is easily scaled to just about any frequency in the hf spectrum or lower. Second, it demonstrates that component quality and proper application of design fundamentals are more significant than a detailed oscillator configuration.

This oscillator was first breadboarded using a large piece of Miniductor coil stock and a 200-pF double-bearing air capacitor, tuned to resonance at 3.5 MHz. The small 1-10 pF capacitor was adjusted for easy starting, but was replaced later with a 5-pF ceramic NP0 unit. Even though the oscillator was tested on the open workbench with no shielding, in a room where the temperature was changing rapidly, the maximum drift observed over a two-hour period was 50 Hz. The air capacitor was then replaced partially with a fixed-value silver-mica unit, resulting in degraded stability. A similar degradation was observed when the air-core inductor was replaced with one wound on a T-68-2 toroid core. Good stability was maintained, however, when most of the capacitance was replaced with paralleled 47-pF NP0 ceramic

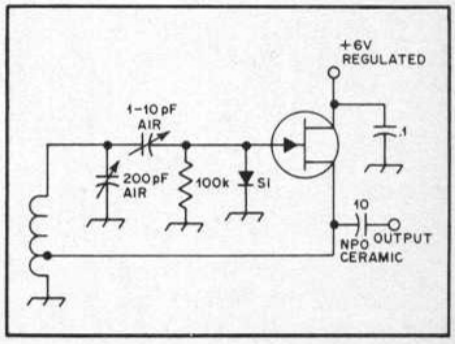


Fig. 7 — W7ZOI high-stability Hartley VFO circuit.

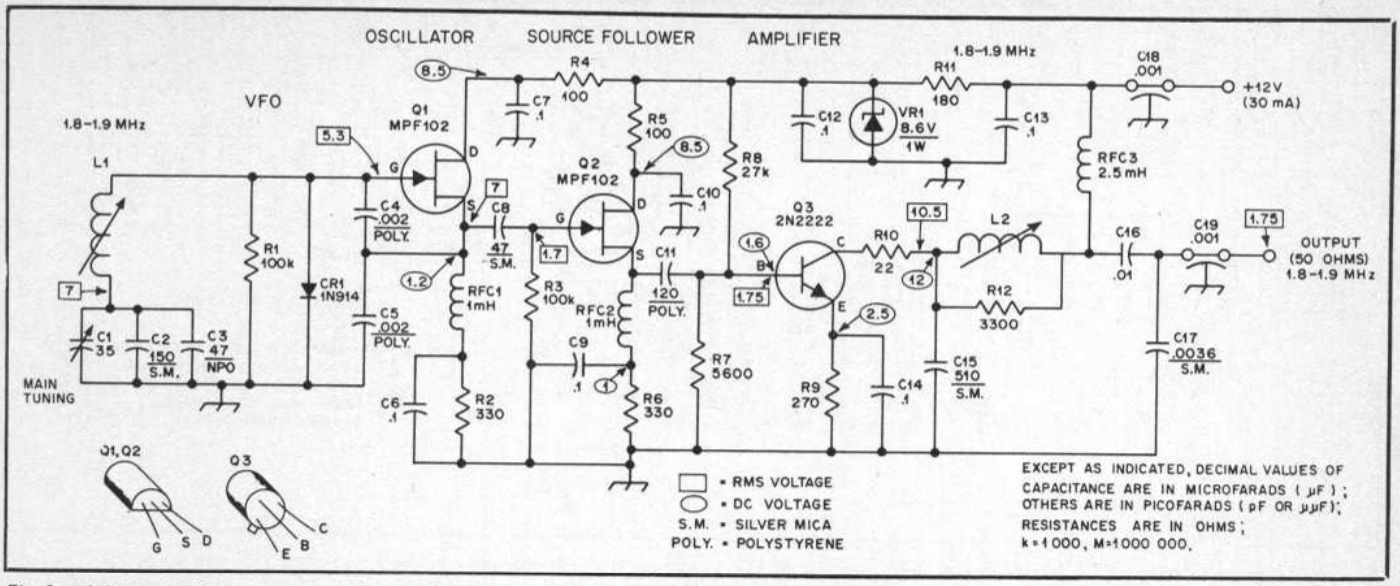


Fig. 8 — Schematic diagram of the 160-meter VFO. Capacitors of fixed value are disk ceramic unless otherwise indicated. Resistors are 1/2-watt composition. Numbered components not appearing in parts list are numbered for pc-board layout purposes only. Rms voltages were measured a VTVM and diode probe.

C1 — 35-pF air variable (Millen 28035MKBB or equivalent).
C18, C19 — .001-μF feedthrough capacitor.
CR1 — Small-signal high-speed silicon diode, 1N914 or equivalent.
L1 — Slug-tuned high-*Q* inductor, 25 to 58 μH (Miller 43A475CBI, Q_u = 180 at 2.5 MHz).
L2 — Slug-tuned, pc-board-mount inductor, 10 to 18.7 μH (Miller 23A155RPC or equivalent).
Q1, Q2 — Motorola JFET.
RFC1, RFC2 — Miniature 1-mH rf choke (Millen J301-1000 or equiv.).
RFC3 — Miniature 2.5-mH rf choke (Millen J302-2500 or equiv.).
VR1 — 8.6-V, 1-W Zener diode.

units. The tap on the coil was 1/4 of the way up from the grounded end.

A Practical High-Stability VFO

The circuit of Fig. 8 is patterned after the VFO used in a W1CER 10-watt cw transmitter for 160 meters which was described in *QST* for November of 1974. Stability is such that in this model the drift could not be measured with ordinary laboratory-style frequency counters during tests in a relatively constant temperature environment (68 to 78 degrees F). From a cold start (no dc applied) to an "on" condition exceeding two hours, the frequency remained constant at plus or minus one Hz. The operating voltage was keyed while monitoring the cw signal from the VFO, and a chirpless note characteristic was observed. While the builder may not be able to duplicate this stability, the circuit should still yield much better than typical performance.

With the *LC* constants shown the VFO tunes linearly from 1.8 to 1.9 MHz. An imported vernier mechanism with a 0-to-100 dial scale provided 1-kHz readout increments. Increased frequency coverage can be had by employing a main-tuning capacitor which has a greater maximum capacitance amount.

A Clapp circuit is used to permit a greater amount of inductance at L1 than would be possible with a parallel-tuned gate tank. The advantages of this were covered in the VFO philosophy section of this chapter. To enhance stability there are a number of polystyrene capacitors employed in key parts of the circuit, and a 1N914 diode is used as a gate clamp.

Q2 presents a high-impedance load to the oscillator, which minimized loading. It has a broadly resonant source circuit from which energy is taken to drive Q3, a Class A bipolar-transistor amplifier. Regulated voltage is supplied to the oscillator, buffer and biasing network of output stage Q3. The collector tank of Q3 is designed for a 50-ohm output impedance, and is a pi network. Although the load seen by a typical transmitter VFO is on the order of 500 to 1000 ohms, assuming a low-level Class A amplifier follows the VFO assembly, the mismatch is intentional. The low-impedance output at Q3 is less likely to "recognize" load changes than would be the case if a 500- or 1000-ohm characteristic were chosen. In fact, when placing a dead short across the operating VFO (C19 to ground), maximum frequency shift was only 10 Hz.

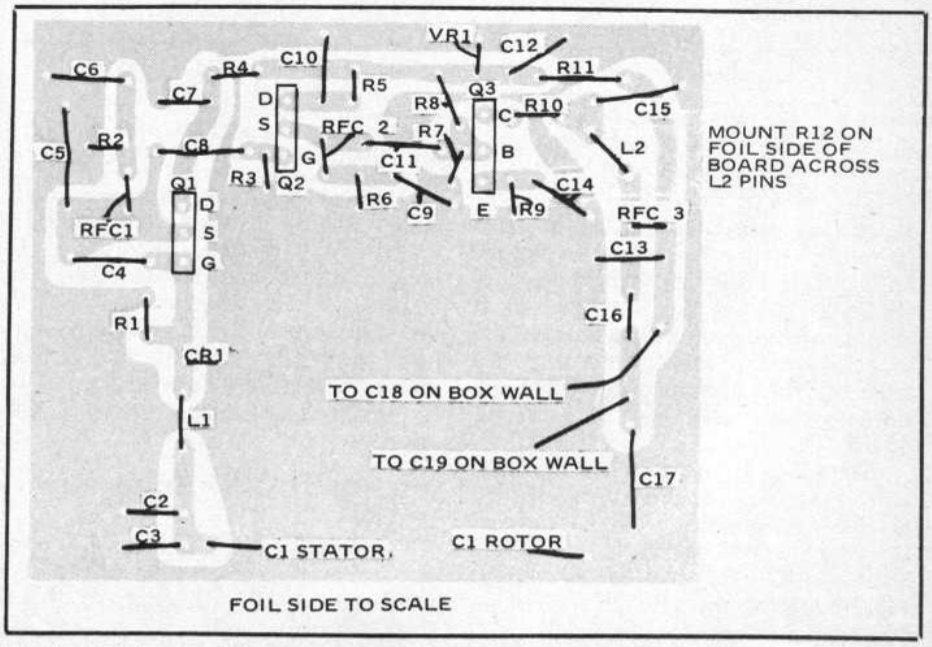


Fig. 9 — Scale layout of the VFO circuit board.

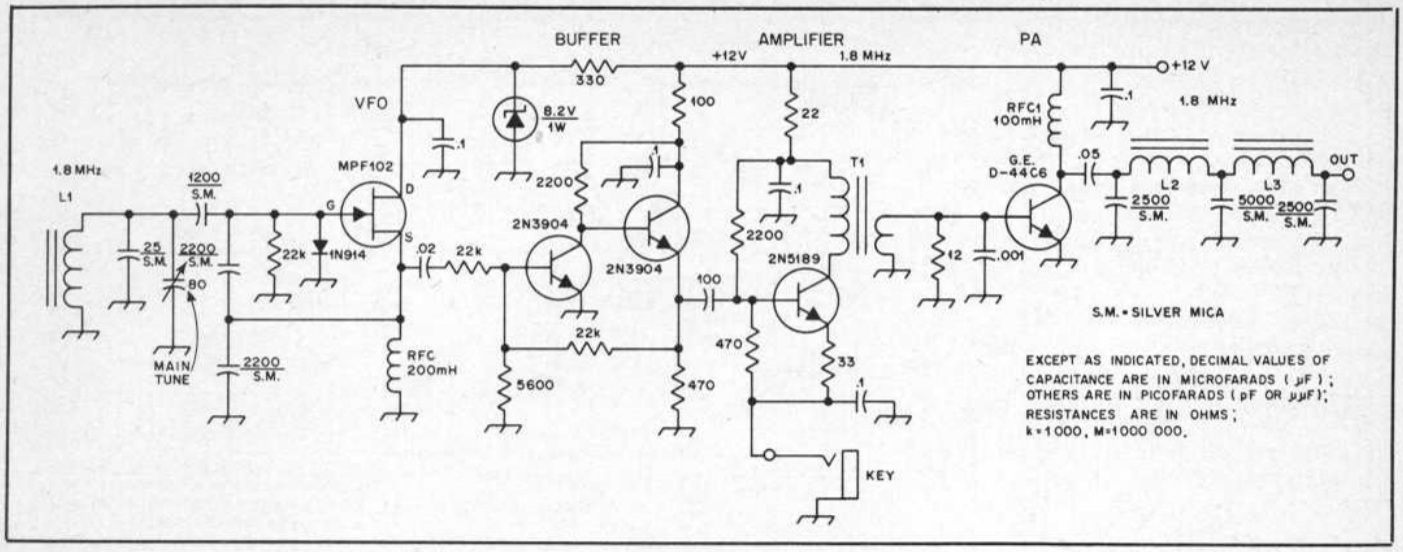


Fig. 10 — Schematic diagram of the 160-meter QRP transmitter. Capacitors are disk ceramic unless otherwise noted. C1 is an 80-pF air variable (main tuning). L1 is a T-68-2 toroid core with 45 turns of No. 26 enamel wire. L2 and L3 are Amidon T-50-2 toroid cores wound with 23 turns of No. 26 emanel wire. RFC1 must be able to pass 0.5A of dc current. T1 is an Amidon FT-37-61 ferrite toroid (μ_i = 125) with 25 primary turns of No. 26 enamel wire. The secondary contains 4 turns of No. 26 wire. Resistors are 1/2-watt composition.

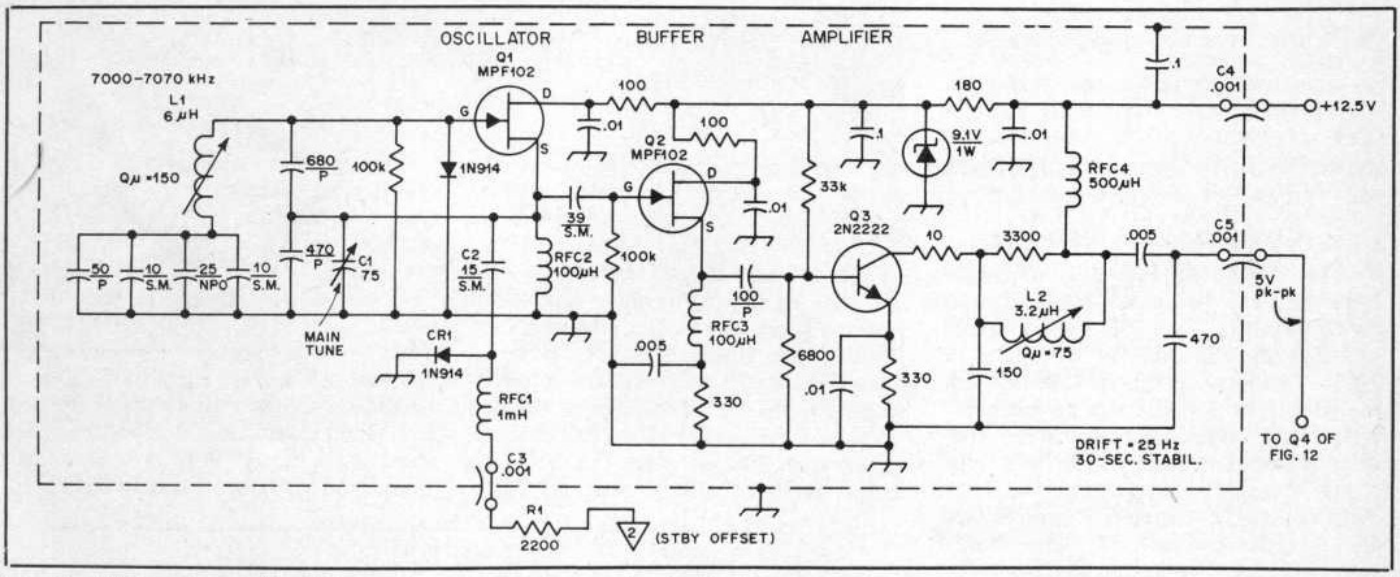


Fig. 11 — VFO portion of the QRP transmitter. Parts values are given in Fig. 12.

The pi-network output tank is a simple low-pass filter which attenuates harmonic energy. The broadbanding resistor, R12, does not significantly degrade the filtering action of the tuned circuit. Measurements showed that the second harmonic was down some 38 dB from the fundamental output, and the third harmonic was down in excess of 45 dB.

The VFO is enclosed in an rf-tight box made of double-clad pc-board material. C18 and C19 are feedthrough capacitors which are installed on the box wall. C19 is part of the output capacitance of the pi network. A pc-board layout is provided in Fig. 9.

Although the VFO is designed for 160-meter use, it can be used in combination with a frequency-multiplier stage for 3.5-MHz operation. Alternatively, it can be modified for higher operating frequencies by taking the reactances of the various components and calculating new L and C values (see Fig. 2). The pc-board pattern is suitable for other operating frequencies.

A 1-Watt 160-Meter Transmitter with VFO

There has been a rebirth of interest in the 160-meter band. While the number of QRP enthusiasts on 160 is small, the band offers excitement and challenge to the low-power enthusiast. Many of the regular operators on "top band" are accustomed to receiving weak signals. Hence, they are able to dig into the noise for a contact.

Shown in Fig. 10 is the circuit for a simple VFO-controlled rig for 160 meters. The design is straightforward and illustrates many of the circuits discussed so far. The VFO is adapted from the one shown in Fig. 4. The VFO is followed by a feedback amplifier with a closed-loop gain of unity. This drives a Class A keyed buffer amplifier. This stage differs slightly from those discussed earlier because a broadband, untuned output transformer is used. This output transformer is much like a tuned toroid, except that the unit is wound on

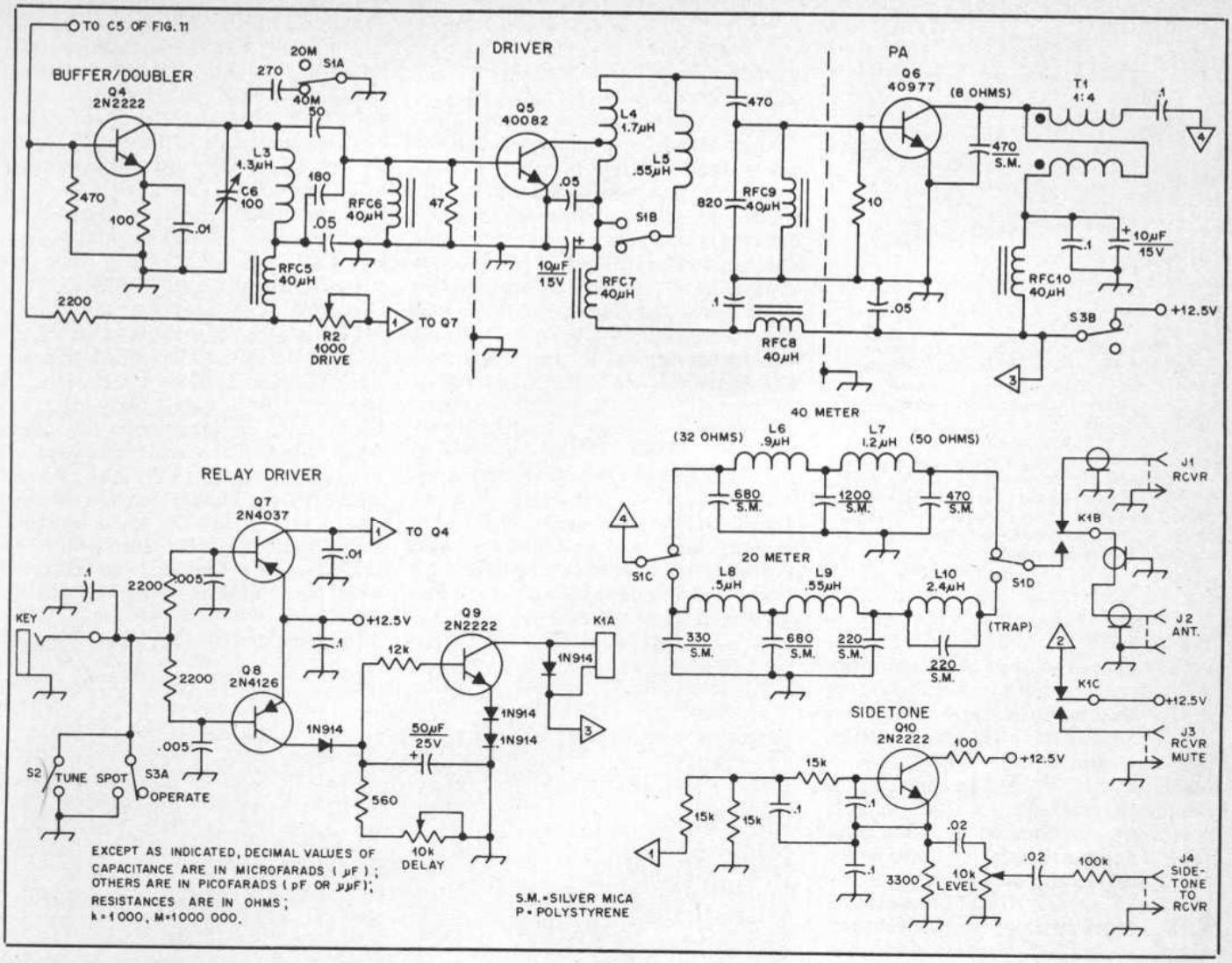


Fig. 12 — Schematic diagram of the QRP transmitter. Fixed-value capacitors are disk ceramic unless otherwise indicated. Capacitors with polarity marked are electrolytic. Fixed-value resistors are 1/2-W composition unless noted differently. S.M. means silver mica, P means polystyrene. Triangles containing numbers indicate circuit connections which are joined directly. Numbered components not listed in caption are so identified for text reference only.

C1 — Small 78-pF air variable. (Miller No. 2109 dual-gang miniature with only 78-pF section connected was used here.)
C3-C5, incl. — .001-μF feedthrough type.
C6 — 100-pF mica compression trimmer.
CR1 — Silicon switching diode, 1N914 or equivalent.
J1-J4, incl. — Panel-mount coaxial jacks of builder's choice.
K1 — Two-pole, double-throw, 12-volt, low-current relay. (24-V P&B KHP17D12 used here, with spring tension reduced for fast pull-in at 12 V.)
L1 — Slug-tuned coil with Q_u of 80 or more, 6 μH nominal. (Miller 42A686CBI used here.)
L2 — Pc-board-mount slug-tuned coil, 3.2 μH nominal. (Miller 23A476RPC used here.) J. W. Miller Co., P.O. Box 5825, Compton, CA 90224.
L3 — 17 turns No. 26 enam. wire to occupy total area of Amidon T-50-6 toroid core (1.3 μH).
L4 — 21 turns No. 26 enam. wire to occupy total area of T-50-6 toroid core, tap at 6 turns from collector end.
L5 — 12 turns No. 26 enam. wire to occupy total area of T50-6 toroid core.
L6 — 11 turns No. 20 enam. wire to occupy total area of T-68-2 toroid core (0.9 μH).
L7 — 13 turns No. 20 enam. wire to occupy total area of T-68-2 toroid core (1.2 μH).
L8 — 8 turns No. 20 enam. wire to occupy total area of T-68-6 toroid core (0.5 μH).
L9 — 10 turns No. 20 enam. wire to occupy total area of T-68-6 toroid core (0.55 μH).
L10 — 25 turns No. 26 enam. wire to occupy total area of T-50-6 toroid core (2.4 μH).
Q1, Q2, Q8 — Motorola transistor.
Q3, Q4, Q9, Q10 — Surplus 2N2222 or equivalent.
Q5, Q6, Q7 — RCA transistor.
R2 — 1000-ohm linear-taper control.
RFC1-RFC4, incl. — Miniature rf choke (Millen J301 series or equivalent).
RFC5-RFC10, incl. — 40-μH low-Q rf choke. Five turns No. 26 enam. wire on Amidon jumbo ferrite bead.
S1 — Subminiature slide switch, S1A and S1B each spdt. S1C and S1D single dpdt unit. (Radio Shack switches. See text.)
S2 — Spst miniature toggle switch (Radio Shack).
S3 — Dpdt miniature toggle switch (Radio Shack).
T1 — Broadband 1:4 toroidal transformer. Ten bifilar-wound turns No. 24 enam. wire, 8 twists per inch, to occupy entire area of two Amidon FT-61-301 ferrite toroid cores (stacked one atop the other). Amidon Associates, 12033 Otsego St., N. Hollywood, CA 91607.

a ferrite core. Most of the toroids used by builders of solid-state gear are of powdered iron and have a relative permeability of around 10 or less. The ferrite core used here (available from Amidon Associates) has a permeability of 125. The reason that high permeability is desirable for a broadband design is that high inductance may be realized with a relatively small number of turns. With a small number of turns the capacitance between turns is low enough that self-resonances are avoided. Broadband performance is enhanced further by the fact that ferrites exhibit a permeability which is a decreasing function of frequency. The transformer is a conventional type in contrast to the



Fig. 13 – Close-up view of the 20/40-meter solid-state transmitter. The cabinet is home-made from 1/16-inch aluminum stock. A cover was made from perforated aluminum which was obtained at a flea market. Kurz-Kasch aluminum knobs are used on the controls. The large knob on the vernier drive was cut down on a lathe to make it thinner, and to permit the set screw to mate with the drive shaft. An SWR-indicator meter is seen at the upper right. Green tape labels identify the controls on a green panel.

transmission-line types described later in this book.

The Class C output amplifier differs from those described earlier. First, the GE D-44C6 used for the final stage has an F_T of over 50 MHz: The available gain is high. This could lead to instabilities. Stability was obtained by the addition of a small value of capacitance across the base-emitter junction. The second departure from the norm was in the design of the output network. We are ahead of ourselves a little here, for such designs have yet to be described. However, in this case we used what appears to be a typical half-wave filter. This is merely a double pi network, each section having a Q of 1. Usually it is designed for a termination of 50 ohms. In this case an impedance of 50 ohms is then presented to the collector. The unusual aspect of the network shown is that it was designed for a termination of 35 ohms. This was done so that a number of available 5000-pF silver-mica capacitors could be utilized. We then take advantage of the characteristic of the half-wave filter wherein it behaves like a half wavelength of transmission line. The result is that a 50-ohm termination on the output yields a 50-ohm load which is presented to the collector. More data will be presented later about the design of these networks.

The output of this transmitter is approximately 1.2 watts into a 50-ohm load, and it is flat across the entire 160-meter band. However, it should be operated through a Transmatch so the rig will always see something close to a 50-ohm termination.

A 3 X 6 X 13-inch chassis was used to house the transmitter, a crystal-controlled converter, an rf power amplifier with an output of 6 watts and suitable T-R switching. The receiving converter will be described in chapter 4.

This package is similar to the unit described earlier for the 6-meter band: All of the required circuits are contained in one box (see photograph). The items needed to complete the station are a receiver in the hf range, a power supply, a Transmatch and keyer. This station design has worked well for bands which are operated on a sporadic basis. One-hundred sixty meters is used only during the winter months, but 6 meters finds heavy use during the late spring and summer months. A similar unit for 2-meter cw is used in the ARRL June and September vhf QSO parties.

A 20- and 40-Meter CW Transmitter with VFO

Fig. 11 shows the VFO used in our 10-watt two-band transmitter. It is patterned after the 160-meter VFO of Fig. 8. Only the L and C values have been changed to increase the operating frequency. A different pc-board pattern is used, but only to enhance miniaturization. C2, CR1, RFC1, C3 and R1 have been included to offset the VFO during receive periods. In that manner the VFO can be kept operational during standby to assure stability (avoiding warm-up drift). Measured drift with this model (at 7 MHz) was 25 Hz over an ambient temperature range of 68 to 75 degrees F. Stabilization occurred in 30 seconds after turn on.

The offset circuit is actuated by application of 13-volts dc during standby. CR1 acts as a switch when saturated, placing C2 in parallel with C1. The amount of frequency shift can be set by selecting a suitable value for C2.

This design was described originally by W1CER in *QST* for March, 1975. A low-power Bruene-style SWR bridge has been added in the cabinet for utility when afield. The circuit was described in *QST* for April, 1959. Also, R1 was changed from 10,000 ohms to the value shown in Fig. 11. The lower resistance value cured a slight chirp which occurred during the first cw character when the break-in delay circuit was actuated.

Fig. 12 contains the circuit diagram

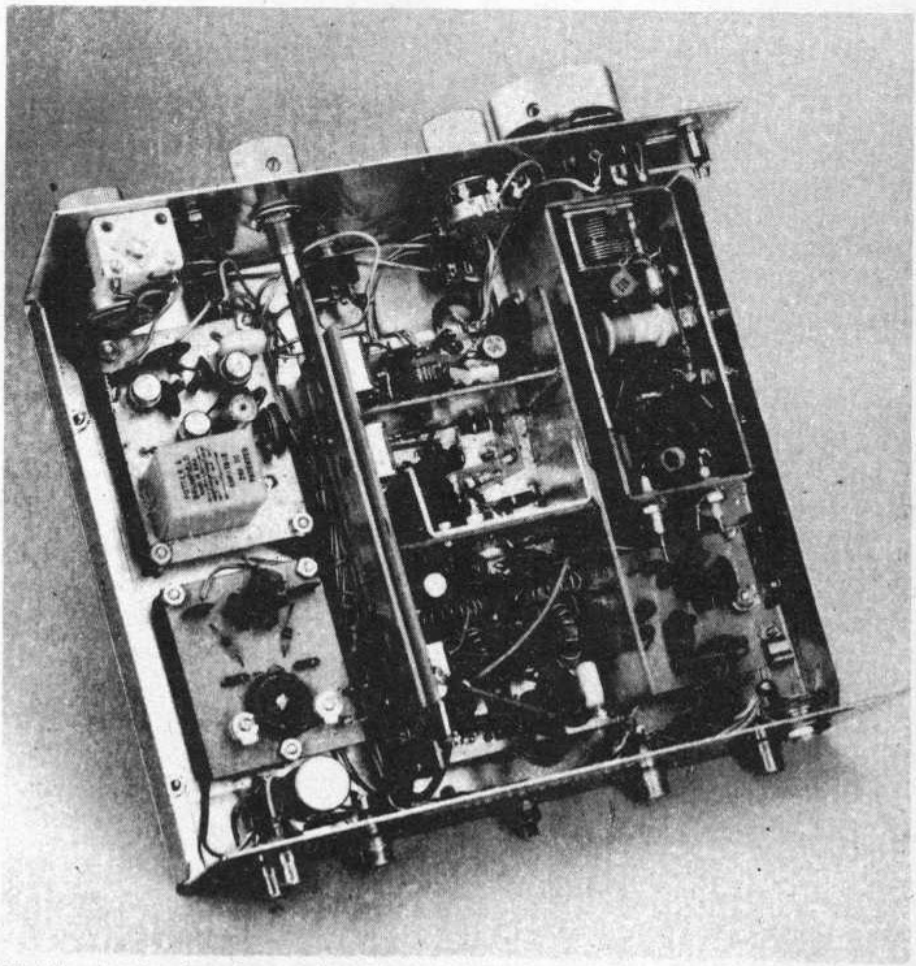


Fig. 14 – Interior view of transmitter. The VFO box is at the upper right with its aluminum cover removed. Directly below the VFO is the sidetone module. The large assembly occupying the center of the chassis is the rf power strip. Three miniature slide switches are ganged by means of a pc-board strip (left on power module). At the upper left is seen the break-in delay assembly. Below it is the SWR-indicator module.

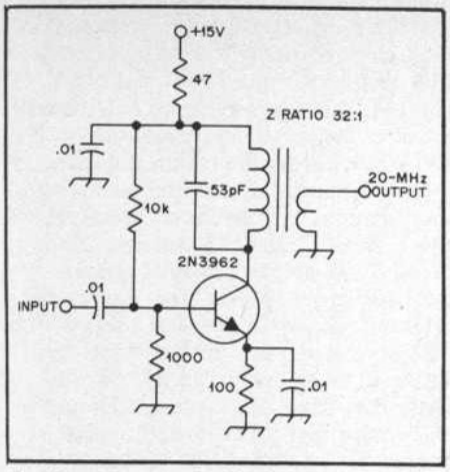


Fig. 15 — Diagram of a bipolar-transistor frequency multiplier.

of the main section of the transmitter, plus peripheral items. The break-in delay and side-tone circuits can be eliminated if manual switching is desired, and if side tone is not needed. The functions of K1 can be effected by means of a two-pole double-throw switch.

A power output of 7 watts is available from this circuit, indicating a PA efficiency (Class C) of 70 percent. This power plateau is ample for most field work. During a two-week DXpedition (ZF1ST) this transmitter was used to work the world on 20 and 40 meters. Simple dipoles were erected near the seashore on Grand Cayman Island, neither of which was more than 25 feet above ground. Power consumption at 13 volts is just under 2 amperes.

The PA tank circuit consists of two double-section pi networks, fixed-tuned, and serving as half-wave filter-matching networks. Because these are low-pass filters, a slight amount of 7-MHz energy appears at the transmitter output during 20-meter operation. Therefore, a 40-meter trap is used (L10) to provide clean output at 14 MHz. Drive control R2 was included to permit very low-power experiments (QRPp), and to reduce transmitter output when driving external high-power amplifiers. Band changing is made possible by ganging three miniature slide switches which are mounted on the amplifier-compartment wall and operated by means of a strip of pc board which is coupled to a knob on the front panel (push-pull action). Photographs of the interior and exterior of the equipment are shown in Figs. 13 and 14. With the VFO *LC* values given, the tuning range is 7 to 7.070 and 14 to 14.140 MHz. Increased range can be obtained by making C1 larger in capacitance.

Frequency Multipliers

The designs offered in the preceding pages have utilized oscillators which operate at the same frequency as the output of the transmitter (Fig. 12 excepted). Certainly for the usual crystal-controlled rig, this presents no problems. However, for work in the amateur bands above 7 MHz it is better practice to operate the VFO at a lower frequency. The output of the oscillator is applied to a stage which multiplies the frequency of the input driving signal. The major advantage of such a scheme is that the frequency multiplier provides excellent buffering. Stray rf from the final amplifier of a small transmitter has minimal effect if it is coupled into an oscillator operating at a different frequency. Of equal significance is that the builder can take full advantage of the harmonic relationship between the lower amateur bands and can build multi-band transmitters with relative ease.

Most of the active devices used in electronics are linear in nature, at least for small signals. Mathematical analysis will show that the output of a linear amplifier contains only those frequencies present at the input, and nothing more. Other frequencies, such as the harmonics we consider here, arise only from departures in linearity.

Most writers state that optimum performance is obtained from a multiplier when it is biased and driven in a way that the distortion products are maximized. However, the discussion usually ends there. The reason for this lack of data is really fairly obvious when one considers the measurements needed. The equipment required to evaluate a frequency multiplier is elaborate and expensive. Only in recent years has this gear become commonly available in even the better equipped electronics labs.

In an attempt to fill this gap, a number of experiments were performed using state-of-the-art instrumentation. The basic unit was a Tektronix 7L13 spectrum analyzer in a model 7704 oscilloscope mainframe. Even though sophisticated measurement gear was used to obtain the data which follows, the results are applicable to the amateur experimenter with his limited measurement capability.

The first experiment was to evaluate a frequency multiplier of the type found in many published designs, Fig. 15. A garden-variety silicon transistor was biased for 7 mA of dc collector current with no rf drive. With high-value rf-drive signals, the current may increase to 15 or 20 mA. The multiplier output contained a powdered-iron toroid, resonated at 20 MHz. The performance as an amplifier, frequency doubler or a tripler, could be evaluated by applying drive from a signal generator at 20, 10 or 6.7 MHz, respectively. The generators

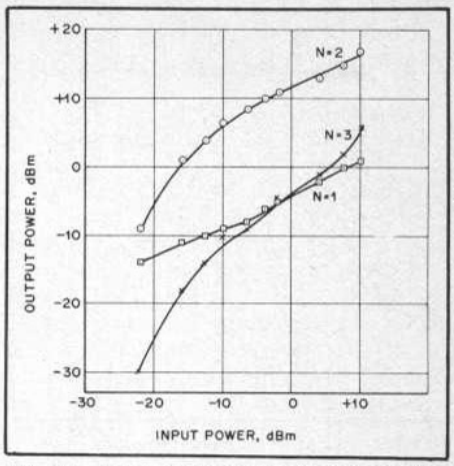


Fig. 16 — Input power versus output power in dBm for a bipolar-transistor frequency doubler. See text for explanation of the curves.

used in the experiments had 50-ohm output impedances; hence, the stage showed no instability. The circuit provided a gain of 24 dB when operated as an amplifier.

Shown in Fig. 16 are the results obtained when the stage was operated as a frequency doubler. The curves show output power as a function of input power. The data form may not be familiar to the amateur. The powers are plotted in dBm, the unit which is used for most rf measurements within the electronics industry. Power in dBm is power referenced to 1 mW. Hence, 0 dBm is 1 mW, −30 dBm is a microwatt and +20 dBm is a tenth of a watt. The other atypical part of the data is that the component powers at the various frequencies of interest are plotted individually. This allows us to compare

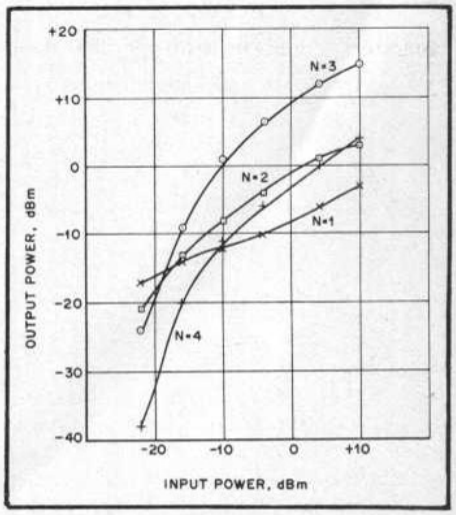


Fig. 17 — Input power versus output power for a bipolar-transistor frequency tripler. See text for data concerning the curves.

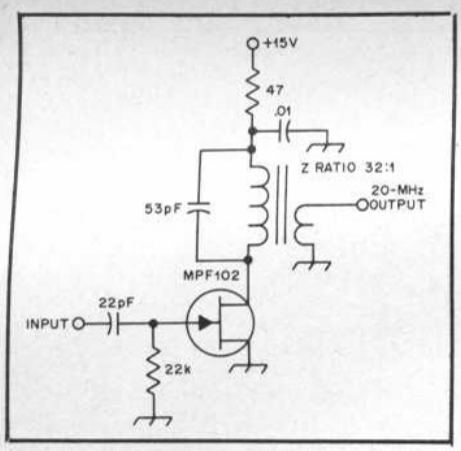


Fig. 18 — Circuit of an FET frequency multiplier.

the desired doubler output ($N = 2$) with the fundamental feedthrough ($N = 1$) and with the third harmonic of the drive frequency ($N = 3$). The input power is not that actually delivered to the stage, but the power available from the generator. There is a difference between the two.

The results are quite revealing. We see that the doubler (Fig. 16) can provide output powers of up to 50 mW (+17 dBm) with a gain of 7 dB. However, the multiplier is not very clean. The best suppression of undesired components in the output is only 16 dB. This occurs at outputs below the maximum obtainable, a less than desirable situation when sophisticated test equipment is not available for evaluation. The performance could be improved substantially by increasing the selectivity of the output tuned circuit. This is most easily realized by tapping the collector a few turns from the V_{cc} end of the output tuned circuit. A double-tuned circuit at the output, if designed properly, would lead to an acceptable doubler.

Shown in Fig. 17 are the results obtained when the stage was operated as a tripler. Performance is even worse than that of the doubler. The best suppression of undesired outputs was 12 dB. This circuit would provide marginally acceptable performance only if a double-tuned output tank were used.

The next experiment is outlined in Fig. 18, where a JFET was evaluated. The first FET tried is typical of those used by the amateur, a 2N4416 with a pinch-off of about 5 volts. The results were discouraging. At high drive levels, the maximum output obtained was only +4 dBm, with spurious output down only 12 dB when operated as a doubler. Surprisingly, the results as a tripler were slightly better. With a drive of 10 volts pk-pk the output was still +4 dBm and the worst spur, the feedthrough of the 6.7-MHz drive, was down 16 dB.

The FET was changed to a 2N4302. This device has a relatively low transconductance and more significantly a pinch-off of only 1.5 volts. When operated as a doubler the output power was quite low, only +1 dBm. However, all spurs were over 18 dB below the desired output. This occurred, again, for a 10 volt pk-pk drive. The performance as a tripler was extremely poor, although the behavior as a X-4 and as a X-6 multiplier was reasonable. This high-order multiplication is not recommended unless high-quality test equipment is available for evaluation and alignment.

In view of the foregoing, it is no surprise that some amateurs encounter problems in building and adjusting gear for the higher hf bands. Furthermore, the problems are not limited to homemade equipment! A prime area where problems arise is in a 2-meter fm rig for which a signal of 6, 8 or 12 MHz must be multiplied many times to arrive in the proper part of the vhf spectrum. Those vhf rigs which use double-tuned circuits throughout the multiplier chain usually have spurious outputs which are at least 45 or 50 dB down. Others rarely fare as well!

All is not lost. The preceding pessimism was intended to encourage the experimenter to strive for good designs. The key to building clean multipliers is balanced circuitry: At least some of the undesired output frequencies should be cancelled. Shown in Fig. 19 is a simple two-diode frequency doubler which was evaluated. Also presented are the classic waveforms for this circuit. We are much more familiar with this configuration as a full-wave power-supply rectifier than in rf circuits, but the same basics apply. The output rf choke will short the dc part of the output signal, effectively moving the zero reference up in the lower curve from the position shown.

The balanced diode doubler shown is not included merely as an example of the effect of balanced circuitry. Shown in Fig. 20 are the output powers vs. available drive power for this circuit. While the diode doubler has a loss of 7.5 dB or more, the fundamental feedthrough is as much as 41 dB down! Note that there are no tuned circuits in this multiplier. The performance appeared to be essentially the same over an output range of 1 to 50 MHz. The input transformer consists of seven trifilar turns of No. 28 wire on a ferrite toroid, 0.375-inch OD, and a permeability of 125. The diodes are silicon switching types of the 1N914 or similar variety. If a smaller core and hot-carrier diodes are used, the circuit will perform well into the vhf range.

This simple diode doubler is used in a direct-conversion transceiver described later. Although a couple of tuned circuits are used in later stages for impedance matching, no attempt was made to achieve good selectivity in the transmitter. Still, the 80-meter component in the output was measured at 52 dB below the desired 7-MHz signal.

The use of balance to remove undesired frequencies from the output of a multiplier can be extended to stages with reasonable power output capability. Two examples are shown in Fig. 21. A push-push doubler is shown at A. It uses a pair of 2N3904 tran-

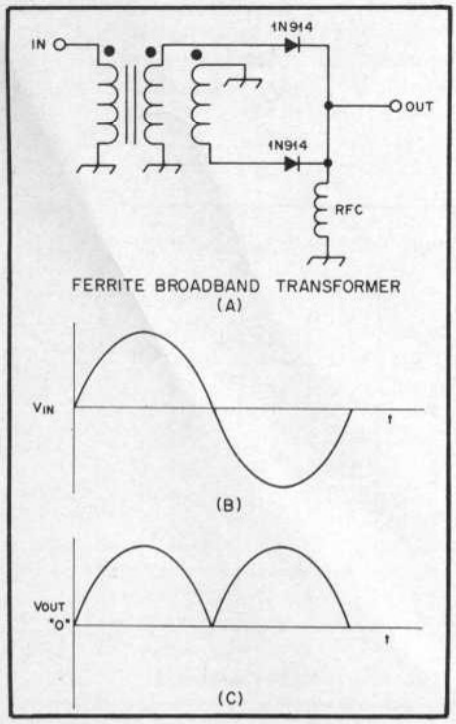


Fig. 19 — Illustration of a diode frequency doubler.

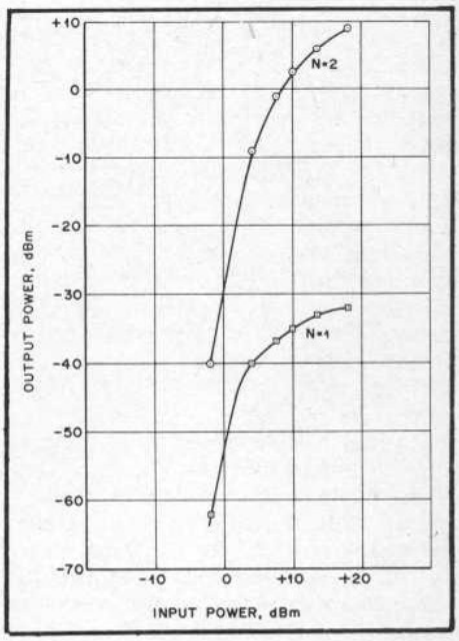


Fig. 20 — Input versus output power for a broadband balanced diode frequency doubler. See text for data on the curves.

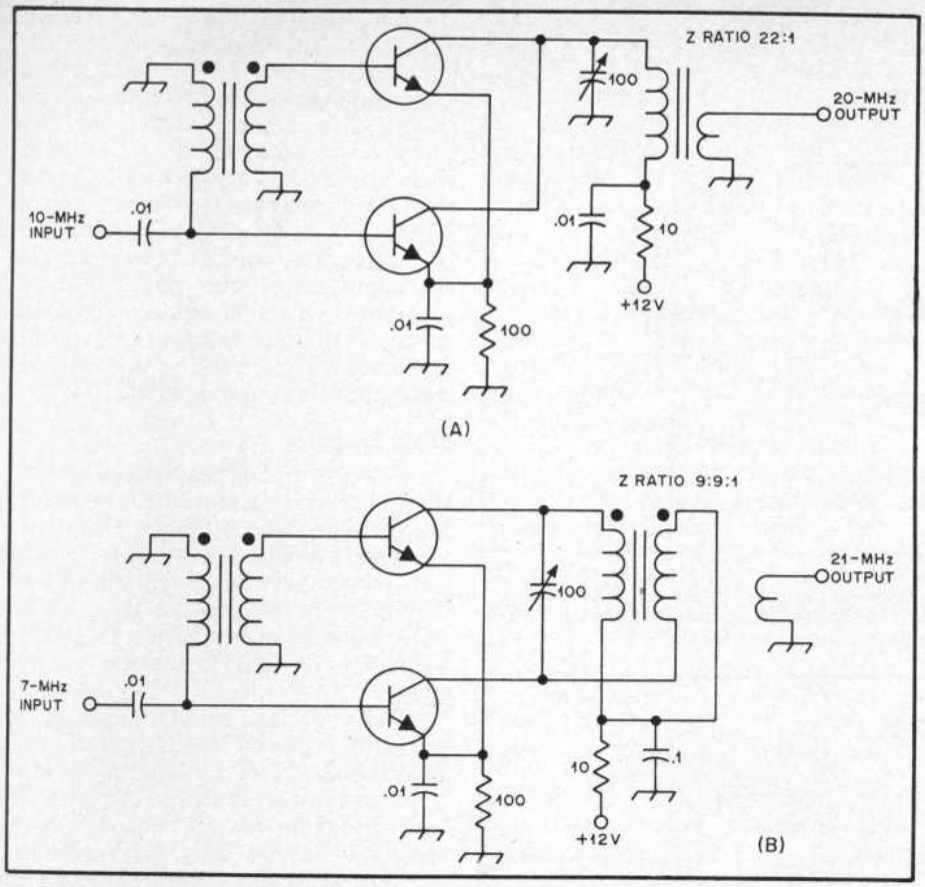


Fig. 21 — A push-push doubler is shown at A. The circuit at B is a push-pull tripler.

sistors. For simplicity, only a bifilar winding is used as the input transformer. This is otherwise identical to the transformer used with the diodes. With 10 mW of drive at 10 MHz, the output is tuned to 20 MHz with a resonant circuit using a powdered-iron toroid. The measured output power was 50 mW. The spur components at 10, 30 and 40 MHz were, respectively, down 50, 40 and 31 dB. The collector efficiency was 42 percent.

Also shown is a push-pull tripler (Fig. 21B). This is identical to the doubler except that a balanced output circuit is used, tuned to 21 MHz. The output power was 32 mW with 10 mW of drive. The spurs at 7, 14 and 28 MHz were suppressed by 30, 55 and 46 dB, respectively, and the efficiency was 26 percent.

If proper methods are used, these balanced multipliers may be used into the lower uhf region. A small cw transmitter was built with a 54-MHz crystal oscillator and three cascaded push-push doublers. All interstage networks are single-tuned, and a low-Q double-tuned filter is used on the output to yield 20 mW at 432 MHz with only one detectable spur, 55 dB down.

Several ICs lend themselves well to clean frequency multiplication. This is because of the excellent inherent matching between monolithic transistors, and this enhances the balance. ICs investigated include the Motorola MC1496G, the RCA CA3046 and RCA CA3028A.

The MC1496 is a double-balanced modulator which is quite useful for mixing applications. It is used as a doubler by injecting the fundamental drive signal to both input ports simultaneously. Although the drive level is a little critical, 60 dB of fundamental attenuation was observed with a single-tuned output circuit. The MC1496 is covered in more detail as a mixer in a later section.

The CA3046 is an array of five transistors. Hence, four of the transistors may be used to form a pair of multipliers of the type described in Fig. 21. Other array-type ICs are worthy of experimentation.

The CA3028A is a general-purpose IC consisting of a differential pair of transistors with a single transistor serving as a current source for the differential pair. If the current source is biased into saturation, the differential pair will serve well as a low-power push-push doubler. This is depicted in Fig. 22.

In general, any of the balanced multipliers outlined may be used. They all offer performance which is significantly better than usually realized with single-ended configurations. However, there are problems encountered with balanced multipliers which are sometimes difficult to diagnose without the aid of sophisticated instrumentation. These are related to imperfections in balance.

Improper balance will result from two major causes. First is the problem of device similarity. For example, the push-push doubler of Fig. 21 will not perform as desired if one of the transistors has twice the current gain of the other. For this reason, it is best to use matched devices whenever these circuits are chosen. This is best realized through the use of integrated circuits such as the CA3046 transistor array or the CA3028A differential amplifier. Even if a perfect match is obtained between the two devices in a balanced multiplier, less than optimum suppression of the fundamental drive frequency will result if there is an asymmetry in the driving waveform. For this reason, the preceding stage driving the multiplier should be a tuned amplifier, or should be a fairly clean Class A amplifier. An alternate might be the use of a low-pass filter such as the unit described at the end of the next section.

It is not imperative that an IC be used in a push-push doubler, respective to matched transistor characteristics. Fig. 23 illustrates how a pair of 2N2222A transistors is connected in push-push style and driven by a JFET source follower. T1 is tuned to 7 MHz, providing push-pull drive to the doubler transistors. Some forward bias is used on the doubler bases to increase the stage gain, but when driven the 2N2222As operate in the Class C mode — essential to doubler action. R1 and R2 are chosen in accordance with the driving voltage available. In this example

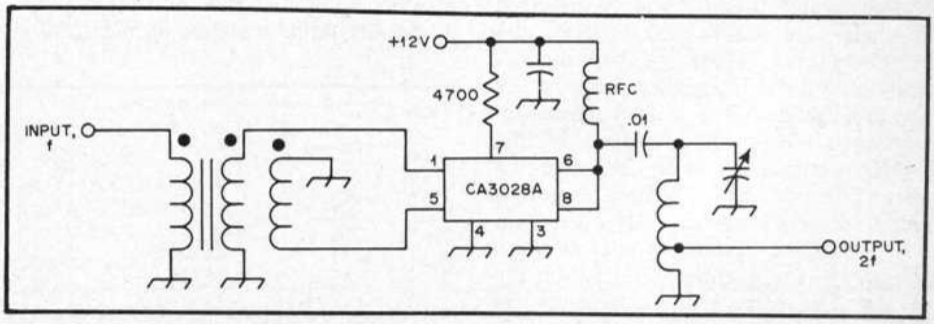


Fig. 22 — Schematic illustration of a CA3028A push-push frequency doubler.

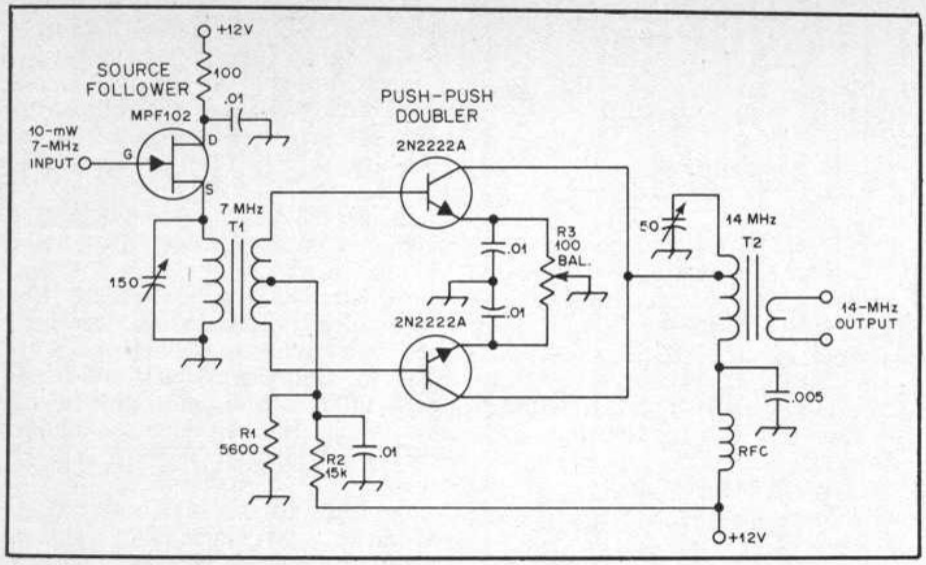


Fig. 23 – A push-push frequency doubler using discrete bipolar transistors, and driven by a tuned-source JFET follower. Power output is approximately 20 mW. T1 has an impedance ratio of 1:1, primary to total secondary. T2 is tuned to the doubler output frequency. R3 is adjusted for dynamic balance of the two bipolar transistors (see text).

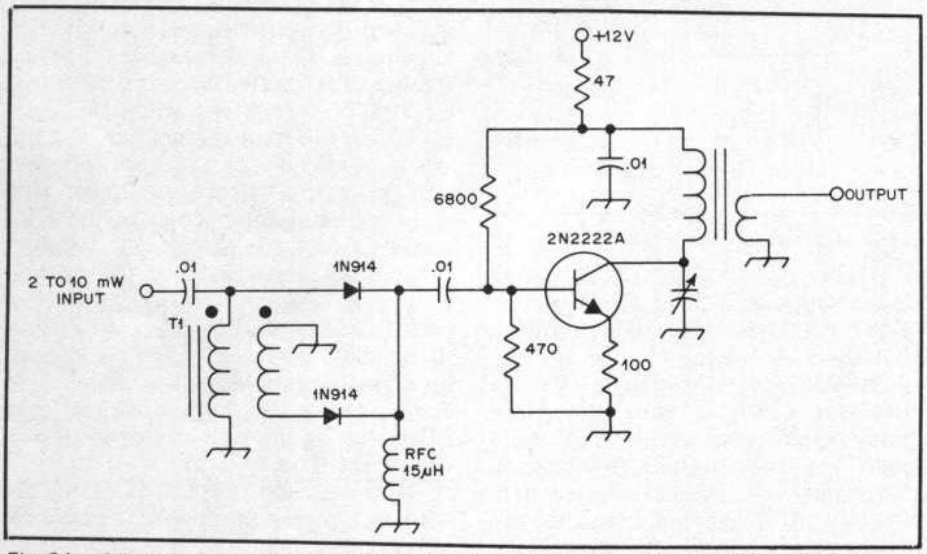


Fig. 24 – Schematic diagram of a diode doubler followed by a low-level amplifier. T1 is an FT-37-61 ferrite toroid containing 10 bifilar turns of small enamel wire.

the FET was driven by a VFO from which the output was approximately 10 mW.

Dynamic balance of the 2N2222As is effected by means of control R3. The output waveform (14 MHz) is observed on a scope and T2 is adjusted to resonance. Then, R3 is set for best waveform purity at 14 MHz. Unless the doubler transistors are widely different in their electrical characteristics, the balancing control will provide the desired effect. In laboratory tests of the circuit (Fig. 23), the output waveform contained no visible evidence of the 7-MHz component after R3 and T2 were adjusted as described here. A Tektronix 453 scope (50 MHz) was used

If sufficient driving power is available – 50 mW or more – the center tap of the T1 secondary winding can be connected directly to ground. Forward bias will not be required to ensure adequate output from the doubler.

For the above reasons, the diode doubler described earlier has appeal. Shown in Fig. 24 is a general-purpose frequency doubler. The previously described diode circuit is followed here by a tuned amplifier. With 5 to 10 mW of driving power, this "gain block" will provide up to 20 mW of output. The diodes should be matched by means of a VOM for forward resistance. However, with unlike silicon diodes in the circuit, the suppression of the fundamental drive was measured at better than 40 dB down. With matched diodes, the suppression was nearly 60 dB. A number of these stages could be cascaded to form a multiband transmitter, starting with a VFO at 160 or 80 meters.

Although a ferrite toroid was used in the input transformer, this could be replaced by a bifilar link on a previous tuned circuit. The output tuned circuit is chosen for the band of interest, and the output turns ratio is about 10.

Mixer Design

Although transistors have been used, the transmitters described thus far have been rather classic in design. That is, we have started with an oscillator which was (crystal-controlled or variable in frequency) followed by an amplifier. In some cases there has been a frequency multiplier or two somewhere in the chain.

Today we find another approach to transmitter design which is becoming predominant. This is depicted in Fig. 25. Instead of working directly with an oscillator at the output frequency, or at some sub-multiple of it, two oscillators are heterodyned in a mixer. The output of the mixer is tuned to a frequency which is the sum or difference of the two input frequencies.

There are a number of advantages to using a mixer. First, stability is often improved. The reason for this is that one of the oscillators may be a highly stable crystal-controlled unit, while the other is variable in frequency. The VFO in the system may often be operated at a relatively low frequency. This will enhance its stability. Furthermore, this oscillator can run continuously. Hence, one has to worry about warm-up drift only once per operating session rather than every time a transmission is started. Another asset of a heterodyne approach to transmitter design is that functions of keying and modulation are well isolated from the critical variable frequency oscillator. Finally, the mixer allows the frequency of a transmitter to be controlled from the same oscillator that is used to control a companion superheterodyne receiver, making full transceive operation practical.

In spite of the advantages listed for

Table 1

	M = 0	*1*	*2*	*3*	*4*	*5*
N = 0	0	9	18	27	36	45
1	5	4/14	13/23	22/32	31/41	40/50
2	10	1/19	8/23	17/37	26/46	35/55
3	15	6/24	3/33	12/42	21/51	30/60
4	20	11/29	2/38	7/47	16/56	25/65
5	25	16/34	7/43	2/52	11/61	20/70

Most cases show two numbers, representing sum and difference frequencies.

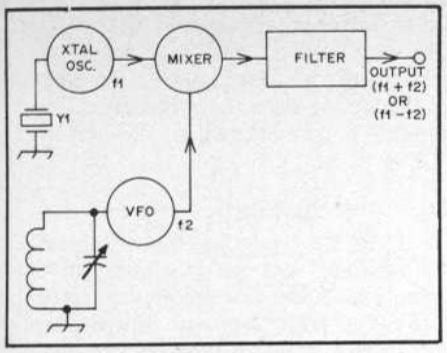


Fig. 25 — Block diagram of a heterodyne frequency generator.

mixers in a transmitter lineup, there are problems which make the design less than trivial. In many ways, the problems are akin to those encountered in our study of frequency multipliers. The mixer and the circuits following it should be designed in such a way that only the desired frequency is dominant in the output. Generally, if we have two input frequencies, f_1 and f_2, a mixer output will contain components at $Nf_1 \pm Mf_2$ where N and M are integer numbers starting at zero!

Let's consider an example, one which is typical because it is based on the frequencies used in many 20-meter receivers. Assume that we have a VFO in the region of 5 MHz, and the crystal-controlled oscillator is at 9 MHz. Some of the possible output frequencies are shown in Table 1.

The list was stopped arbitrarily at N and $M = 5$. However, it goes on (and on and on). Clearly, the spurious response is potentially worse than was the case with frequency multipliers where the only possible output frequencies were of the form $N \times f$.

If we study the list, remembering that our desired output is the 1:1 response at 14 MHz, we see that mere filtering is not ample. For example, we see a 3:0 response at 15 MHz, and a couple of different spurs at 16 MHz, as well as a 1:2 response at 13 MHz. In spite of this, clean spur-free mixers can be built. The key to the design is the same as we encountered in building frequency multipliers — balance. That is, circuits are chosen which cause some components to be canceled in the output. With most mixers we will consider, the fundamental driving frequencies and their odd harmonics are well suppressed in the output, sometimes by as much as 60 dB. The additional spurs may be suppressed by filtering and a judicious choice of input frequencies.

Shown in Fig. 26A is the circuit for a double-balanced mixer using an MC1496G. By double balance, we mean that information at *both* of the input ports is suppressed from appearing in the output. (Often, in the generation of single sideband by the phasing method, a pair of balanced modulators is used with a common output. This is *not* what is usually meant by "double balance.") The internal workings of the MC1496 are shown at Fig. 26B. One signal is injected differentially on the bases of a pair of common-current sources. Since emitter degeneration is used at this input, it is usually the best point for applying a low-level signal where it is desired to preserve linearity. For example, this would be the place to apply a low-level ssb signal if such a transmitter were being built.

The collectors of the two signal-carrying input stages are then routed through four switching transistors. The stronger local-oscillator signal is applied to the bases of these switching transistors. Using the component values suggested in the Motorola applications literature, the maximum current that ever need be switched is around 1 mA. Hence, fairly small local oscillator injection voltages are required to achieve proper switching action. Usually, signals of the order of 100 to 300 mV (rms) will be sufficient. In cw transmitters, the lower level signal can be as much as 100 mV. In linear applications, however, the signal at pin 1 should be less than this by 10 or 20 dB. Often, in linear applications, better distortion characteristics will be obtained by biasing the IC to larger currents. This is realized by decreasing the 10-kΩ resistor that connects to pin 5. The standing current in the IC is essentially twice the current flowing into pin 5. The Motorola data state that the chip should not run with more than 10 mA.

Shown in Fig. 27 is the internal circuitry (A) and a mixer application of the RCA CA3028A (B). Although

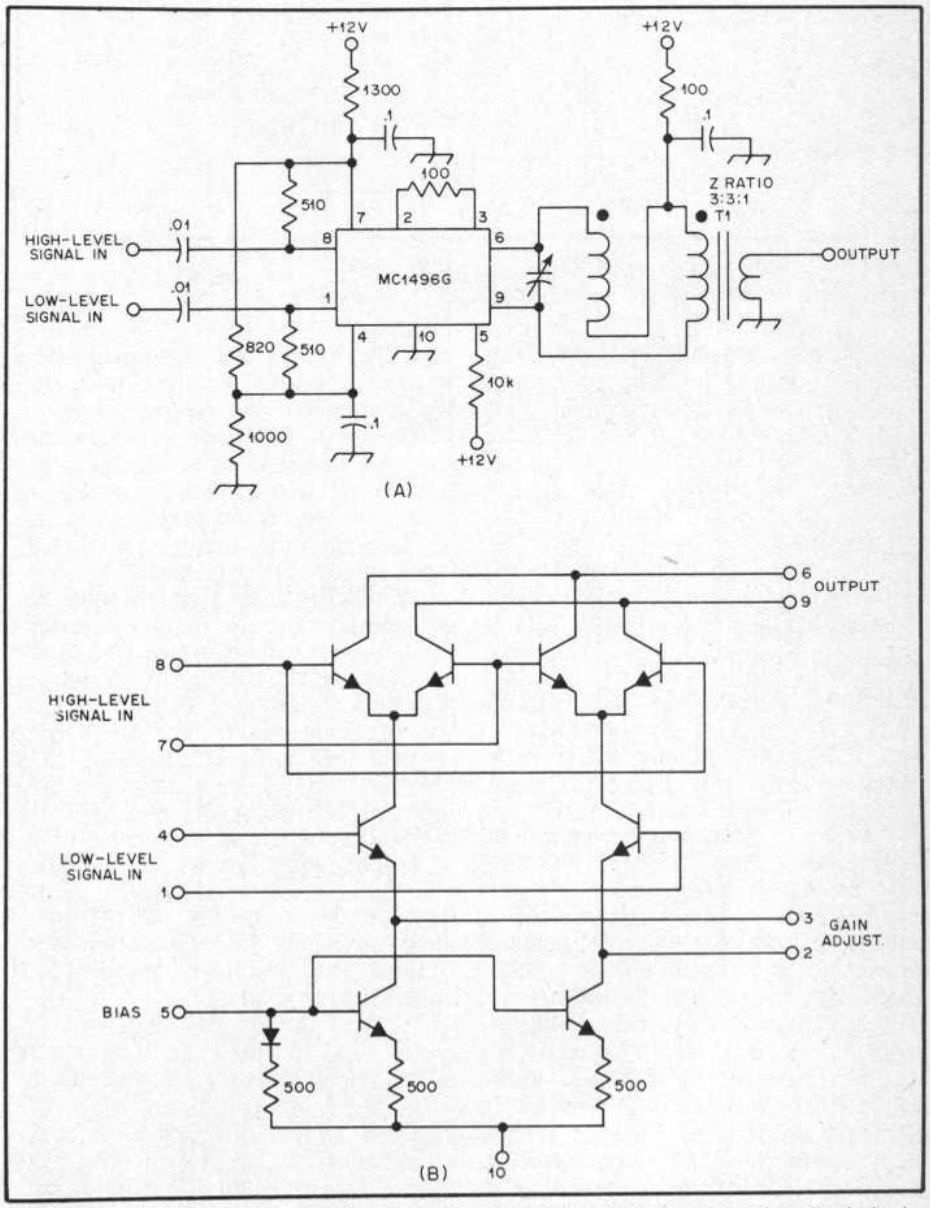


Fig. 26 — An IC mixer is shown at A. T1 is a toroidal bifilar transformer, tuned to the desired output frequency. The internal circuit of an MC1496G is shown at B, courtesy of Motorola.

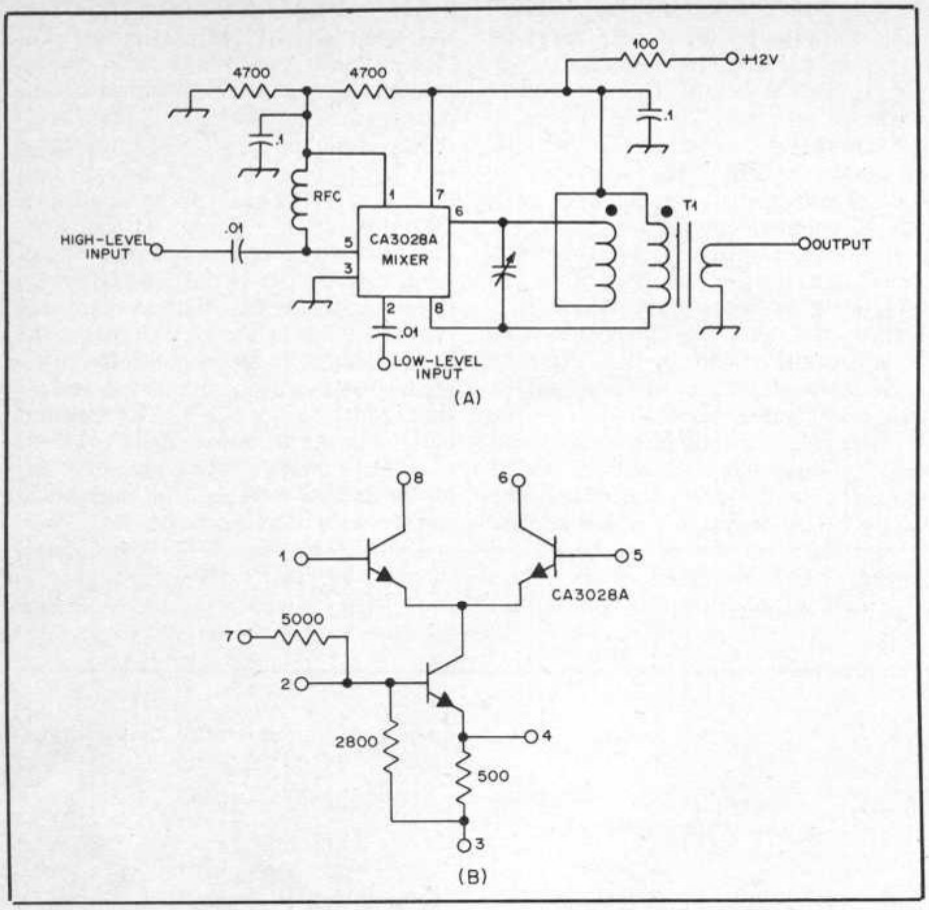


Fig. 27 — Circuit of a CA3028A single-balanced mixer at A. T1 is the same as for the circuit of Fig. 26. At B is the internal circuit of the CA3028A, courtesy of RCA.

simpler than the previous circuit, this configuration has the disadvantage that it is only a single balanced mixer. That is, signals applied on pin 2 of the IC are suppressed in the push-pull output. However, the push-pull drive applied between pins 1 and 5 is not suppressed in the output.

Fig. 28 presents the internal circuitry (A) and a suggested mixer circuit (B) for the TI SN-76514. This chip is similar to the MC1496 in its operation, although the role of the rf and LO ports is reversed. The SN-76514 should be an easier "pill" to apply than the MC1496 since all of the biasing resistors are contained on the chip: One pays for this convenience by reduced versatility.

In the sample circuits presented for the MC1496G and SN-76514, the outputs are taken differentially between two collector terminals. However, if a builder is willing to accept reduced conversion gain, and this is usually acceptable, output may be taken from only one collector. The balanced properties of the chip will be retained so long as proper collector bias is maintained. Using this design philosophy, it would be convenient to build a two-band transmitter. The band switching would be simplified by attaching a band-pass filter for each band to the two output collector points. This is shown in Fig. 29. Appropriate band-pass filters may be selected from the "filter catalog" presented in the appendix.

One of the really classic approaches to mixer design is to use diodes as the mixing elements. Two examples of diode-type mixers are presented in Fig. 30. Like the other examples presented, these mixers are balanced. The two-diode mixer is single-balanced while the diode-ring mixer is double-balanced. Diode mixers exhibit loss instead of the gain associated with the other mixers presented. Impedance matching is critical in diode mixers, and some spur responses are not well suppressed. On the other hand, diode mixers come into their own in broadband applications and in situations where wide dynamic range is desired. Most mixers of this kind utilize hot-carrier diodes, such as the HP-2800. However, for the hf region silicon switching diodes are often satisfactory. They should be matched for similar foward resistance.

FETs of the junction and the MOS types may be used in transmitting mixers. However, they are used ideally in balanced configurations. While the dual-gate MOSFET is popular as a receiver mixer, it has the problem that harmonics of the local oscillator, particularly even-order ones, are easily created within the device. This can cause serious problems with spurious responses unless good balancing techniques are used and careful filtering is applied. Additonal information on mixers is presented in the receiver chapter.

Frequency Synthesis

When we hear the term "frequency synthesizer," we may think of the techniques used for frequency control of 2-meter fm equipment. Narrow-band vhf-fm is a mode of amateur communications which requires great frequency accuracy and stability. Hence, it is ideal for synthesis techniques. However, frequency synthesis is by no means limited to 2-meter fm. It appears that such methods will become predominant as the major means of frequency control in all high-performance amateur equipment.

In the general sense, frequency synthesis is any process which electronically operates on one or more frequencies to produce other frequencies. The mixers and frequency multipliers we have discussed earlier are examples of simple forms of synthesis. There are, however, other methods which can be applied.

It would be folly to attempt a complete treatment of synthesizers. Such a discussion would take us well beyond the relatively empirical scope of this volume. Nonetheless, synthesis methods are becoming so popular that some explanation is required. We will confine our discussion to two types of synthesizers which are of interest to the experimentally inclined amateur.

The major advantage of frequency synthesis is stability. If one begins with a highly stable crystal oscillator as the reference frequency, the output of the synthesizer using this reference will have a stability which is dependent upon the characteristics of the quartz crystal rather than a less stable VFO. If the system is well designed, the stability will be quite good. One of the simplest synthesizers the amateur can build consists of nothing more than a pair of crystal-controlled oscillators and a mixer. Each oscillator contains a bank of switchable "rocks." The advantage of a scheme of this kind is that the stability of crystal control is retained while great frequency accuracy is obtained. An additional characteristic, which may or may not be an asset, is the digital nature of the "tuning." Such digital techniques are useful for portable equipment designed for cold-weather conditions.

As an example of this type of synthesizer, consider the block diagram of Fig. 31. Here, the two crystal oscillators are operated at 20 and 27 MHz. Each oscillator has five crystals available. The two reference frequencies are

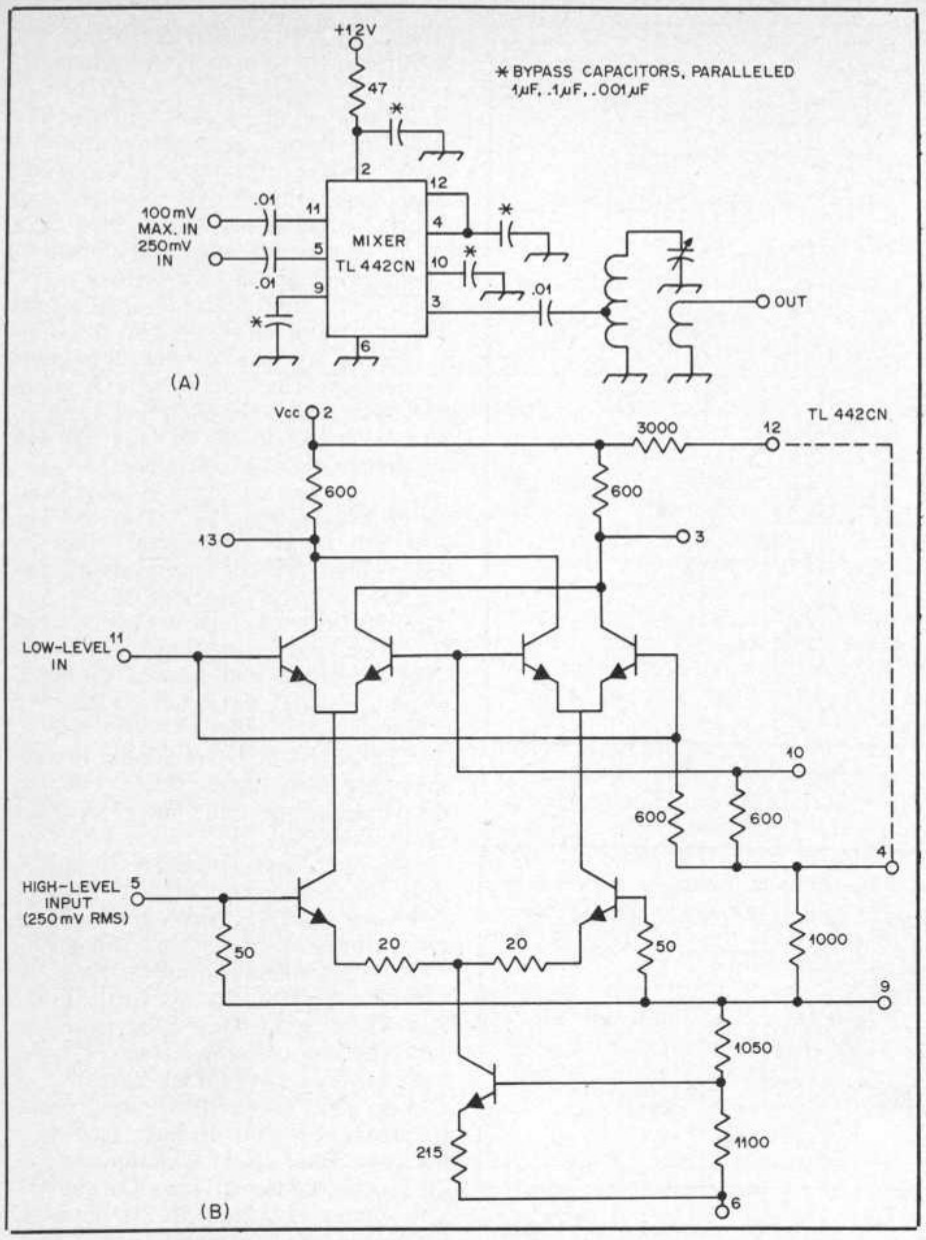


Fig. 28—Circuit for a doubly balanced SN-76514 mixer, at A. The circuit at B shows the internal workings of the IC, courtesy of Texas Instruments. The SN-76514 mixer IC has been reidentified as TL-442-CN by Texas Instruments. It may be procured under either part number.

applied to a mixer with an output at 7 MHz. A low-pass filter at the output ensures that none of the higher-order spurs are present. With an investment in only 10 crystals, 25 discrete frequencies in the 40-meter band will be available. A module of this sort would not be expensive to build, for CB crystals could be used in the 27-MHz oscillator.

Although frequency synthesizers using banks of crystal-controlled oscillators are fairly common, they are not as practical as might be desired. This is because a large number of crystals are required if versatility is desired. The techniques used to avoid this deficiency are usually based upon the phase-locked loop (PLL).

There are a number of circuits which will serve as the critical element in a PLL (the phase detector). A phase detector is a three-port circuit, much like a mixer. At two of the ports (the inputs) two signals *at the same frequency* are applied. At the third port, a dc voltage appears. This voltage is proportional to the phase difference between the two input signals.

A simple PLL is shown (Fig. 32) in block-diagram form. The system includes the phase detector, a reference oscillator, a voltage-controlled oscillator (VCO) and a loop filter. The phase-detector operation was defined above, and the reference oscillator could be, as an example, a stable crystal-controlled oscillator. The VCO is merely a VFO with the usual mechanically tuned capacitor replaced with a varactor diode. As the voltage on the varactor is changed, the effective width of the depletion region of the diode changes, causing the diode capacitance to change. The loop filter is essentially a low-pass filter which tends to remove any ac components from the output of the phase detector.

How is this system used to control frequency? The key to understanding PLL operation, at least on a rudimentary basis, is to recall that frequency is merely the rate of change of phase. That is, the phase of a signal from a highly stable oscillator is a constantly changing parameter. Once during each cycle of oscillation the phase returns to some "zero-degree" reference point. Recall that the phase detector is a circuit which compares the phase *difference* between two signals. If the outputs of our two oscillators (the reference and the VCO) are exactly at the same frequency, there will be some dc voltage at the detector output. This dc level is proportional to the constant phase difference, whatever it may be, between the two oscillators.

Assume now that the VCO starts to drift a little with respect to the frequency of the reference. Say, for example, the VCO tends to move in frequency by 1 Hz from that of the reference. If the two frequencies were indeed different by 1 Hz, the phase difference would be continually changing. That is, it would be a 1-Hz ac signal. However, in our PLL, this never happens. As soon as the phase starts to shift the resulting dc signal from the phase detector is amplified (and filtered) in the loop filter and then applied to the VCO. The change in the dc control voltage on the varactor diode of the VCO is just that required to bring the frequency of the VCO back to that of the reference oscillator. The control voltage may be different from that present before the VCO started to drift, but the frequencies will be the same.

The simple PLL shown in Fig. 32 has one flaw which may not be apparent immediately. It will, however, become painfully clear when one attempts to build such a unit. Assume, for example, that the crystal reference is at 1 MHz

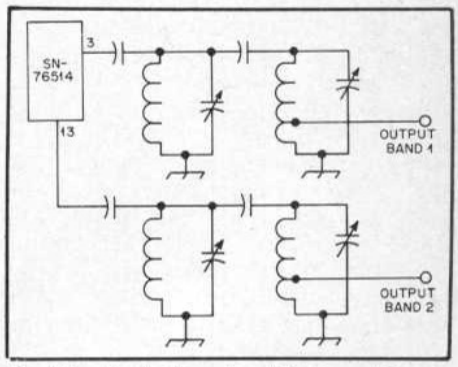


Fig. 29 — Illustration of an IC mixer with output on two frequencies. Pins not indicated on the SN-76514 are connected as shown in Fig. 28.

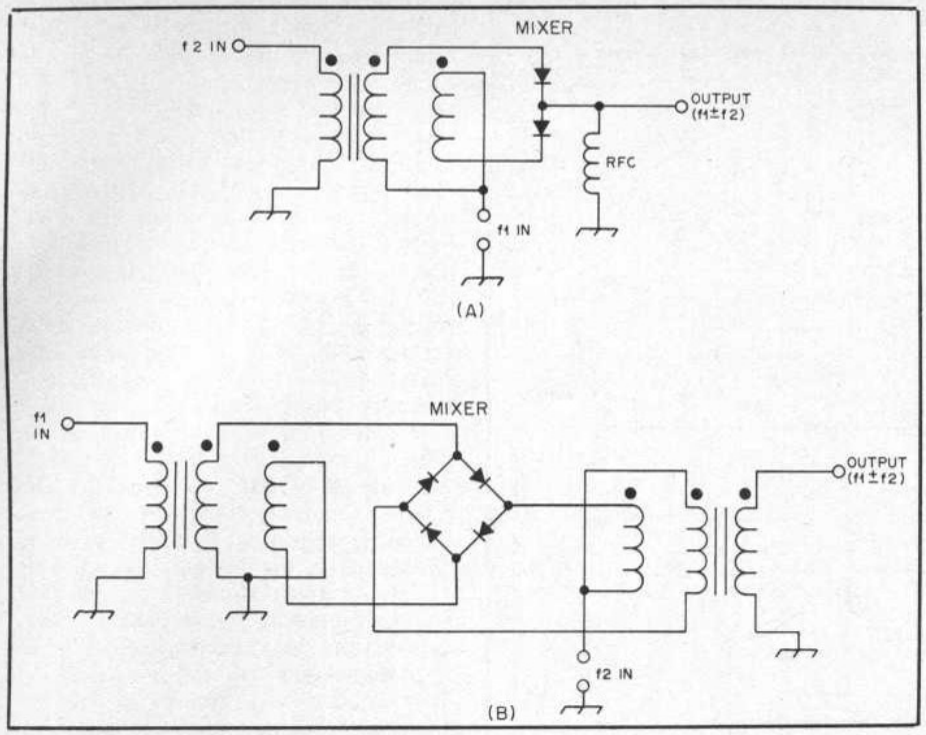


Fig. 30 – At A, a two-diode mixer. A four-diode mixer is shown at B. The transformers are trifilar-wound on ferrite toriod cores.

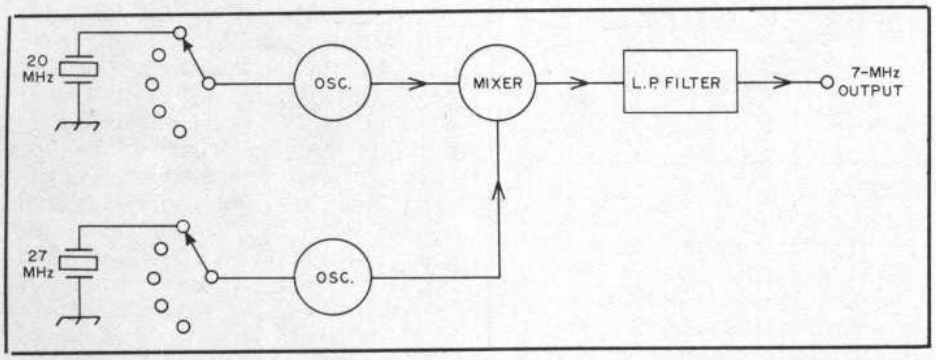


Fig. 31 – Representation of a simple 7-MHz synthesizer.

and that the VCO is capable of tuning from 0.9 to 1.1 MHz with the available voltages. Most likely, what will happen when power is first applied to the circuit is that the VCO will start oscillating at one end of the control range or the other, and it will stay there. With a 100-kHz difference in frequencies, there will be no dc control voltage emanating from the phase detector – just the ac signal at 100 kHz. What we must do is to initially "perturb" the VCO until it is momentarily at the same frequency as the reference. Then a suitable dc phase-controlled signal will exist which will cause the PLL to "lock up" and control the VCO. This perturbation is usually realized by additional circuitry which will cause the VCO to sweep over its range prior to lockup.

A simpler and more convenient approach to this problem is to replace the phase detector with a phase-frequency detector. This circuit provides a dc output which is a function of frequency difference prior to lockup. This signal, in combination with the smoothing effects of the loop filter, will in effect generate the required sweep voltage. Once the VCO is near the frequency of the reference, normal phase-detector operation commences. An example of such a detector is the MC4044. The detailed operation of this digital circuit is rather complicated, but is well outlined in the Motorola literature.

Shown in Fig. 33 is a simplified phase-frequency detector which is built from a pair of D flip-flops and a NAND gate. A D type of flip-flop is a fairly simple device in comparison to many of the digital circuits used extensively in modern electronics. Whenever the positive edge of a pulse appears at the clock input, C, the logical state present at the D terminal is transferred to the output, Q. $\overline{Q}$ is merely the opposite logical state from that at the Q terminal at any instant. A logical zero applied at the reset input, R, will always return the Q output to a logical zero.

In the circuit shown, an SN-7474 dual-D FF is used, in conjunction with a single NAND gate from an SN-7400 quad two-input NAND-gate package. The D terminals are always tied to a logical one. Hence, whenever a positive-going pulse appears at either clock input, that flip-flop is set into a high output state. Each of the two flip-flops is clocked by one of the two input frequencies. The NAND gate is wired such that both flip-flops are reset to zero whenever both Q1 and Q2 are simultaneously at a logical one.

Several sets of possible waveforms are shown in Fig. 33. At B are two different input frequencies with f_1 higher than f_2. The appropriate levels for Q1 and Q2 are also shown. Of significance is the high *average* level of Q1. When this is smoothed out in the loop filter, we will have a dc signal coming from Q1 which tells us that f_1 is higher than f_2.

The curves at C are similar, except that here f_2 is higher than f_1. We see that the average value of Q2 is much higher than Q1.

Fig. 33D and E depicts f_1 and f_2 equal in frequency, but out of phase with each other. As shown in the curves, the outputs at Q1 and Q2 will tell us what nature and magnitude of the phase difference is actually present. In the case where exact phase coincidence occurs, the outputs from Q1 and Q2 will both be very short positive pulses.

Fig. 34 shows how this phase-frequency detector is interfaced with the loop filter. Note that the outputs used from the detector are Q1 and $\overline{Q2}$. The circuit is meant to be generally descriptive of the operation and lacks many of the interfacing details necessary to provide stable operation. These details will depend upon the final system configuration.

As we study the simple PLL of Fig. 32, the first impression we get is that the system is redundant. That is, why would one use an oscillator at 1 MHz to control another? Why not take the output directly from the reference oscillator, dispensing with all of the other circuitry? While the present system is an illustration, simple loops of this kind are

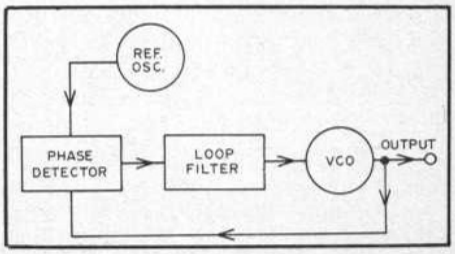


Fig. 32 – Illustration of a basic PLL circuit.

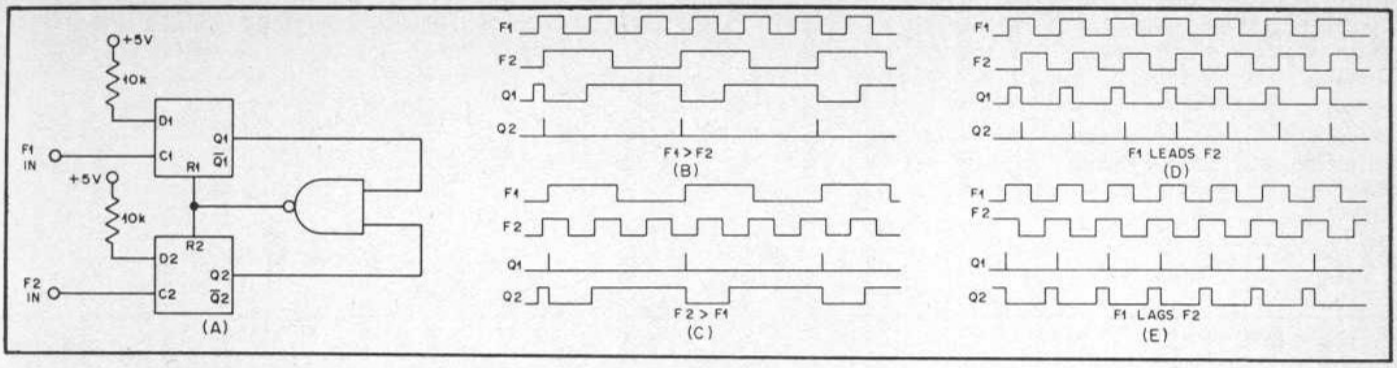


Fig. 33 — Representative phase-frequency detector using an SN-7474 IC and 1/4 of an SN-7400 IC. See text for details of illustrations B through E.

of value in some advanced systems. For example one could arrange the circuitry and choose a proper phase detector such that the outputs from the two oscillators were 90 degrees out of phase. The two outputs could then be used for generation of ssb by the phasing method.

A much more significant application of a simple loop of this kind relates to the noise characteristics of an oscillator. When we think of an oscillator, we envision a device which has an output at one discrete frequency. Perhaps we acknowledge the existence of a few harmonics, but take a simplistic view of the typical oscillator. Usually, this is justified. However, if one attempts to build equipment which approaches the state of the art (whatever that means), the noise characteristics of the oscillator must also be considered.

In our earlier discussion of VFOs we mentioned long-term stability (the "wanderies") and short-term drift (the "wobblies"). Long-term drift is an instability which usually has its origin in thermal effects. Short-term wobblies, on the other hand, originate from noise in the oscillator. Random variations in the output of the amplifying device used in an oscillator will cause minor variations in the phase (and hence, the frequency) of an oscillator. The net result is that our oscillator seems to provide a discrete frequency which is modulated by noise. In this case, the modulation appears as a variation in phase of the oscillator. This pm or fm — the distinction between the two is essentially nonexistent — causes sidebands next to the "carrier." These noise sidebands may be the ultimate limitation in the design of a wide dynamic-range receiver, as one significant example.

Unfortunately, the short-term and long-term variations in the frequency are not necessarily related. That is, one may fight for long periods of time to remove the wanderies from a VFO, only to find that he has designed a highly stable noise source. Noise considerations are of major significance in the design of any phase-locked loop. In many synthesizers used by amateurs, a PLL has been used to achieve a degree of long-term stability at vhf which surpasses that found on even the lower hf bands, but creates a signal which is excessively noisy. Casual application of PLL techniques can be quite disastrous.

On the other hand, a PLL can be used to clean up residual phase noise in an oscillator. The simple loop of Fig. 32 could be a good example. If the reference oscillator were quite stable (long term) and noise free, essentially all of this cleanliness could be impressed upon the output of the VCO which might otherwise be much less than clean. However, only those noise sidebands on the VCO which are separated from the VCO carrier by a frequency difference less than the bandwidth of the loop filter will be suppressed by the PLL.

Let's now consider a somewhat more complicated synthesizer based upon the PLL shown in Fig. 35. This unit is typical of many units which have been implemented for 2-meter fm use. We have shifted our reference frequency down to 1 kHz. This is easily done by starting with a crystal-controlled oscillator at 1 MHz, then applying the resulting signal to a divide-by-1000 circuit. Typically, this would consist of three SN-7490 decade dividers. Simlarly, the output of the VCO is applied to a frequency divider. Let's assume for the moment that the VCO operates in the 6-MHz region and that the divider is set up to divide by 6000. If the VCO were right at 6 MHz, we would have two 1-kHz signals being applied to our phase-frequency detector. The phase-proportional detector output would now be filtered in the loop filter and applied to the VCO. The VCO would move to the exact frequency required to achieve lock, where both inputs to the

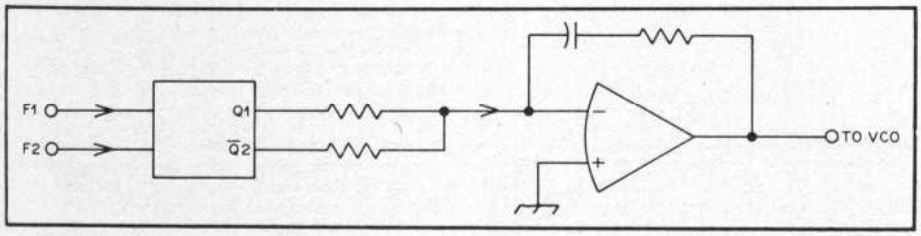


Fig. 34 — Simplified schematic diagram of a loop filter for use with a phase-frequency detector.

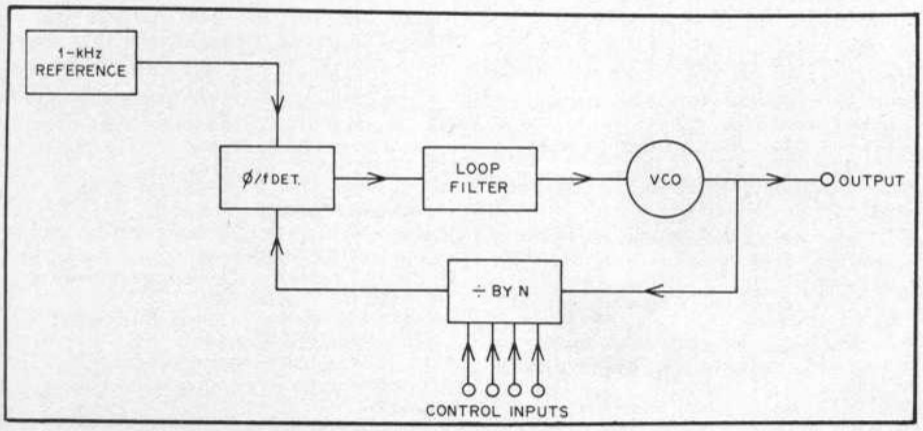


Fig. 35 — Block diagram of a divide-by-N synthesizer.

phase detector have a stable, well-defined phase difference.

A system of this kind is made "tunable" over a band of discrete frequencies by replacing the VCO-driven frequency divider with one which is programmable. That is, from the front panel of our synthesizer we could set switches which would cause the divider to, for example, divide by 6132 instead of 6000, causing the VCO to lock up at 6.132 MHz. By changing the division ratio we pick the desired output frequency. In some kinds of synthesizers the divider in the reference-frequency chain is also programmable.

It is worthwhile to consider the operation of the detector in more detail. The reference frequency in this case is 1 kHz. As a result, once every millisecond the digital phase detector is pulsed by the reference. The phase detector serves the function of telling us whether the similar pulse from the programmable divider ar-

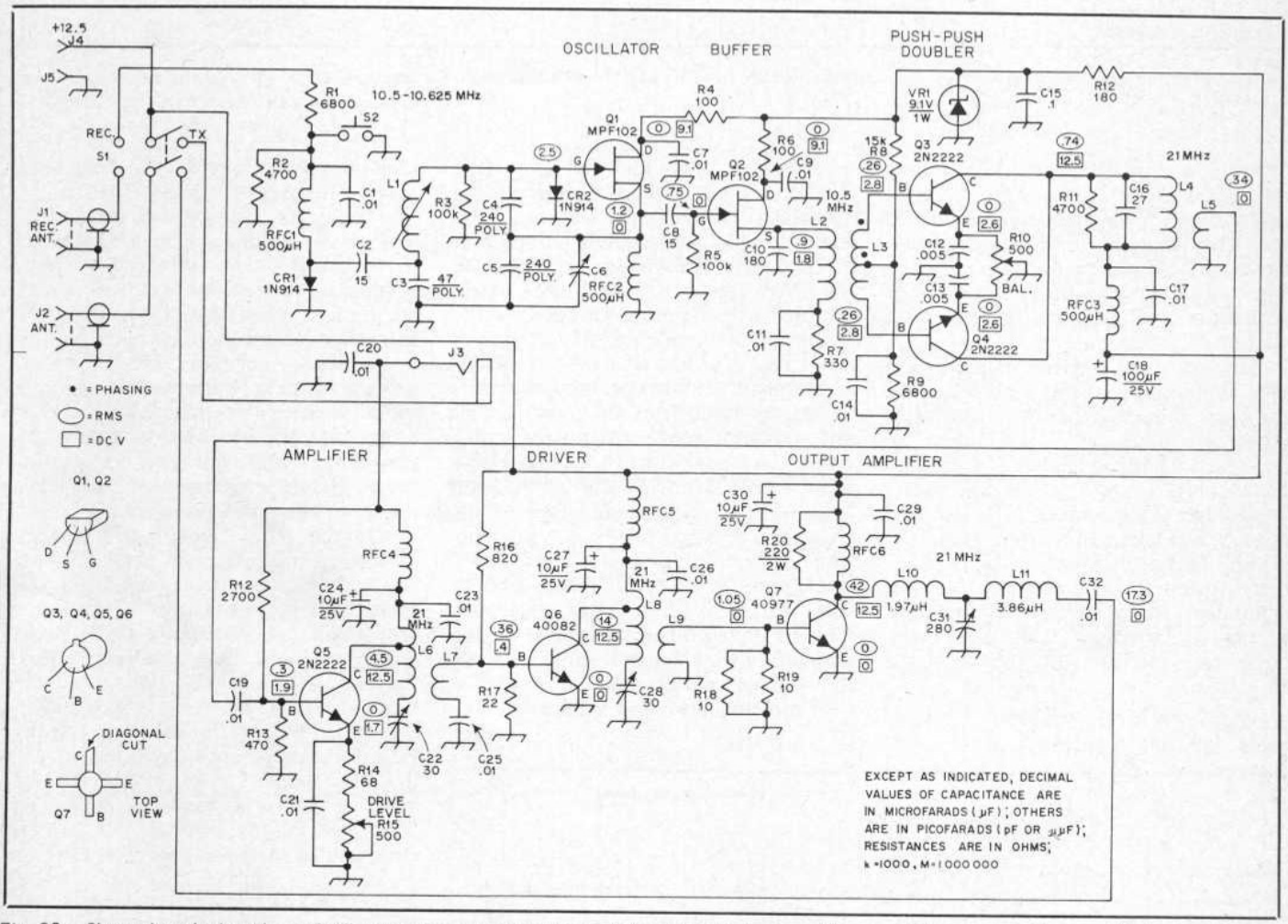


Fig. 36 – Shown here is the schematic diagram of the 15-meter transmitter. Fixed-value capacitors are disk ceramic unless specified otherwise. Fixed value resistors are 1/2-W composition unless noted otherwise. Numbered components not appearing in the parts list are identified for pc-board layout purposes only.

C3 – 47-pF polystyrene.
C4, C5 – 240-pF polystyrene.
C6 – 4- to 53.5-pF variable (Millen 22050 or equiv.).
C18 – 100-µF electrolytic, 25 volts.
C22, C28 – 2.7- to 30-pF variable (Elmenco) 461 or equiv.).
C24, C27, C30 – 10-µF tantalum or electrolytic, 25 volts.
C31 – 25- to 280-pF variable (Elmenco 464 or equiv.).
CR1, CR2 – 1N914 or equiv.
J1, J2 – Coaxial connector, type SO-239.
J3 – Phone jack (Radio Shack 274-280 or equiv.).
J4, J5 – Binding post.
L1 – 6.05- to 12.5-µH adjustable coil (Miller 42A105CBI or equiv.).
L2 – 17 turns No. 28 enam. wire on Amidon T-50-6 core.
L3 – 10 turns No. 28 enam. wire, center tapped, wound over L2.
L4 – 17 turns No. 28 enam. wire on an Amidon T-50-6 core.
L5 – 5 turns No. 28 enam. wire wound over L4.
L6 – 30 turns No. 28 enam. wire on an Amidon T-50-6 core. Tap 10 turns above C23 end.
L7 – 4 turns No. 28 enam. wire wound over L6.
L8 – 30 turns No. 28 enam. wire on an Amidon T-50-6 core. Tap 7 turns above C26 end.
L9 – 3 turns No. 28 enam. wire wound over L8.
L10 – 22 turns No. 28 enam. wire
L11 – 29 turns No. 22 enam. wire on an Amidon T-68-6 core.
Q1, Q2 – Motorola MPF102 JFET or equiv.
Q3, Q4, Q5 – 2N2222 transistor.
Q6 – RCA 40082 transistor.
Q7 – RCA 40977 transistor.
RFC1, RFC2, RFC3 – 500-µH rf choke (Millen J-302-500 or equiv.).
RFC4 – 16 turns No. 28 enam. wire on an Amidon FT-50-61 core.
RFC5 – 11 turns No. 22 enam. wire on an Amidon FT-50-61 core.
RFC6 – 6 turns No. 22 enam. wire on an Amidon FT-50-61 core.
S1 – Dpdt miniature toggle switch.
S2 – Spst momentary-contact push-button switch.
VR1 – Zener diode, 9.1 volt, 1 watt.

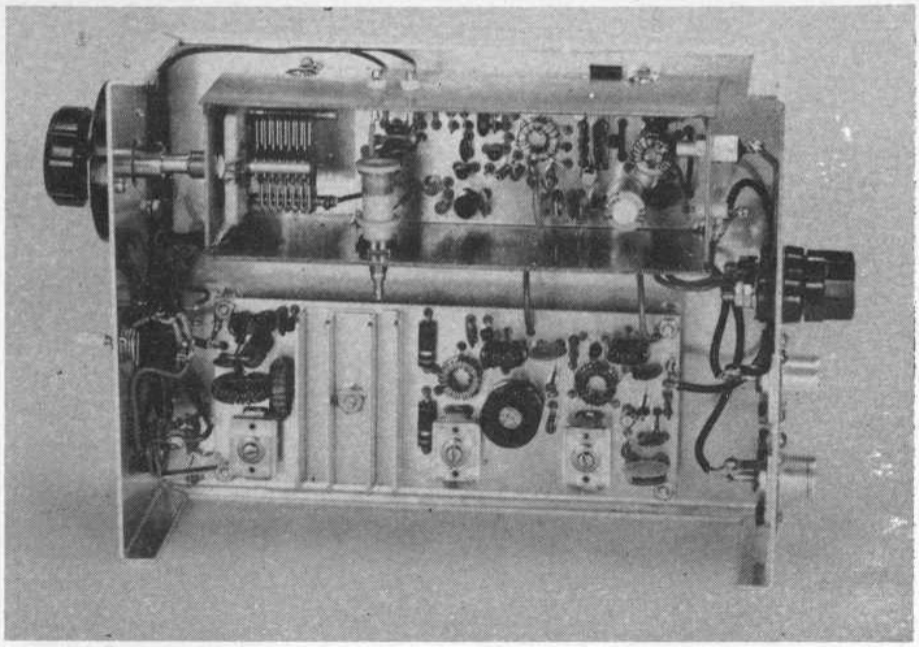


Fig. 37 — Interior view of the transmitter. The VFO is in the compartment at the top. The rf power strip is the lower module.

rived before or after the reference pulse. The output signal is a short pulse of the right polarity to ultimately cause the VCO to shift as needed to assure phase coincidence. The average of these pulses is our dc level. The purpose of the loop filter is to remove, as much as possible, the pulse or ac variations in the signal applied to the VCO. However, in designing the loop filter, we now encounter problems. First, if we are going to effectively filter out a series of pulses occurring at a 1-kHz rate, we must use a low-pass filter with a bandwidth of well under 1 kHz. This, unfortunately means that it is difficult to change frequencies. When we switch the programmable divider to a new ratio, the VCO will "hunt" for a short period, being driven by the proper frequency difference signal from the phase-frequency detector. If the loop filter bandwidth is as narrow as 1 Hz, the loop may take over a second to settle at a new frequency. A compromise bandwidth is usually used.

No matter how narrow the filter there will be some pulse or ac component which will be applied to the VCO. Hence, the VCO is being frequency modulated by our 1-kHz reference. With a suitably narrow loop filter the resulting sidebands are fairly well suppressed. However, when the VCO output is used to drive a frequency multiplier chain, as would be the case with a 2-meter fm transmitter, the suppression of the residual reference sidebands deteriorates. In general, the residual sidebands will come up by 6 dB every time the frequency is doubled.

Another problem arises when we are forced to use an exceptionally low loop order to achieve good suppression of the bandwidth. As mentioned earlier the inherent noise sidebands in an oscillator can be suppressed only at separations from the carrier which are less than the loop bandwidth. In the case described the PLL does essentially nothing to make the VCO output quieter. When the VCO is applied to a multiplier the amplitude of the noise sidebands will again grow, just as the reference frequency sidebands did. The design of the VCO in the example considered must be extremely well done if the ultimate result is to be tolerable.

One final point should be made about the design of the loop filter. In reference sidebands at the VCO, one might be tempted to use a compliciated, multisection low-pass filter of the kind used for audio filtering in a direct-conversion receiver, except, of course, having a lower cutoff frequency. In general, this approach is not viable. The reason is that any filter will exhibit maximum phase shift in any region where the attenuation is changing rapidly with frequency. The ultimate result of this phase shift is that the entire PLL may oscillate. These oscillations are detected experimentally as an ac component on the "dc" signal being applied to the VCO.

While it is hard to generalize, the better PLL designs are those which use the highest possible reference frequency. Furthermore, it is desirable to operate the VCO at the highest reasonable frequency. Finally, heterodyning the VCO output to a desired output frequency is recommended over frequency multiplication. This is because of the degradation of the noise sidebands inherent with multiplication.

Frequency-synthesis techniques offer great promise for future amateur equipment. However, great care is required in the design if high performance is desired.

A Deluxe 15-Meter CW Transmitter with VFO

This circuit was described originally in *QST* for January, 1976, by WA1LNQ. Power output is approximately 6 watts across 50 ohms when using a 12-V dc supply (1.3 A), and 7 watts of output can be had at 13 volts. Frequency coverage is from 21.0 to 21.250 MHz with the constants specified in Fig. 36. An interior view is given in Fig. 37, and the outside of the assembled unit is shown in Fig. 38.

The series-tuned VFO is fashioned after the circuit of Fig. 8, and the push-push doubler follows the lines of the circuit in Fig. 23 of this chapter. Stability is excellent at 21 MHz (less than 70 Hz from cold start to stabilization, requiring approximately two minutes).

A spectral analysis of the 21-MHz rf output (at the 6-W level) shows the second harmonic to be down 45 dB, and the third harmonic is 55 dB down from 21 MHz. The cw note is free of clicks and chirp.

The VFO offset circuit (C2 and CR1) is used to kick the operating frequency 100 kHz off the desired frequency during receiving periods. This prevents interference from the VFO while in the receive mode, and enables the VFO to remain operational at all times, thereby ensuring nearly drift-free VFO operation.

The matching networks and tuned circuits of the overall transmitter are sufficiently broad in response to permit the full 250-kHz operating speed without need to retune the stages. R11 across L4 helps to provide flat response from the VFO chain.

Circuit-board templates and a parts layout are available from the ARRL for $1.25 and a large s.a.s.e.

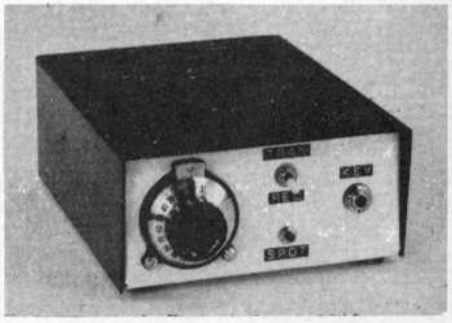


Fig. 38 — Exterior view of the transmitter

Chapter 4

Power Amplifiers and Matching Networks

Practical amplifiers and some "cookbook" equations will be presented in this chapter for those who wish to design their own impedance-matching networks. Concerning the latter, only simple math is needed to solve for the various impedance combinations germane to solid-state amplifier circuits. It is recommended that the builder/designer obtain one of the low-cost engineering-function electronic calculators for the work treated in this book. The resolution is far superior to that which can be realized with a slide rule, and answers to problems can be obtained more rapidly with a calculator.

Despite the large variety of networks available for impedance-matching in transmitters, all of these designs have some common characteristics. First, most of the networks used by the amateur are essentially low-pass types. That is, at frequencies well above the design center the networks offer significant attenuation. As a general rule of thumb, one can assume that the ultimate attenuation will be 6 dB per octave per reactive element in the network. For example, a common network found in the amateur solid-state transmitter is the double-pi network (low Q), containing two inductors and three capacitors. If such a design were "cut" for 7 MHz, the attenuation at 14 MHz would be around 30 dB. It could be higher than this if the network had a high- loaded Q. Another characteristic of the common impedance-matching networks is that they are "singly loaded." This fact requires some elaboration: Assume that a low-power transmitter was being designed for an output of 1 watt with a 12-volt dc supply. Hence, the required load resistance which must be presented to the collector is $V_{cc}{}^2 \div 2P_o = 72$ ohms. A suitable network would be a pi type, designed to transform a 50-ohm antenna termination to the needed 72 ohms. What this means is that if one end of the network is terminated in a 50-ohms resistor, a resistance of 72 ohms is "seen" looking into the other end. The amplifier behaves as if a 72-ohms resistor were coupled capacitively to the collector. However, the network *is not* being driven from a 72-ohm source. Typically, the output impedance of the amplifier will be much higher than this, perhaps several hundred ohms.

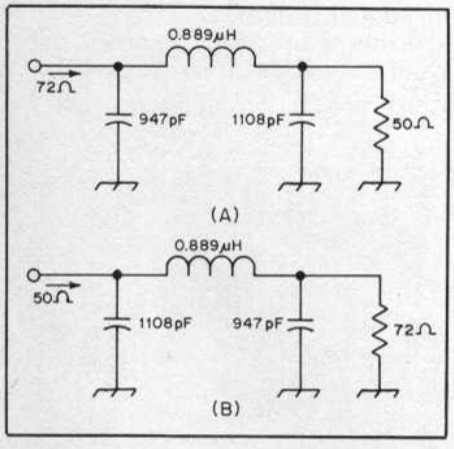


Fig. 1 — Transposition of a pi network to illustrate effect of resistive termination.

Networks which are used for impedance matching are called "singly loaded," since it is necessary that only one end of the network be properly terminated in order to realize the required impedance transformation and filtering characteristics. Not all LC networks are singly loaded, however. The classic double-tuned circuits which one might find in the front end of a receiver are doubly loaded designs. That is, both the input and output of these networks must be terminated properly in order to achieve the filtering desired.

A characteristic of the filters in this section is their reciprocal nature. That is, even though the networks are singly terminated designs, it does not matter which end of the network is terminated resistively. For example, the pi network just mentioned was designed such that a 50-ohm resistor appears as a 72-ohm resistance at the other end. However, with the same network a 72-ohm resistor at the high-Z end would appear as a 50-ohm impedance at the low-Z end, with no difference in filtering properties. This is illustrated in Fig. 1, where the constants are for 7 MHz and the design Q is 3.

Once the desired resistances for each end of a network are determined, the network is then "designed." Inductors and/or capacitors are placed either in series between the two ends of the network, or are connected as shunt elements to ground. In the strictest sense only two reactive components are

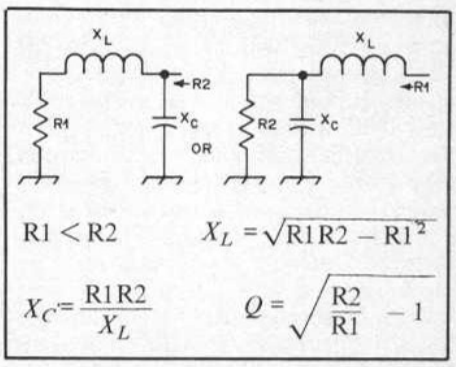
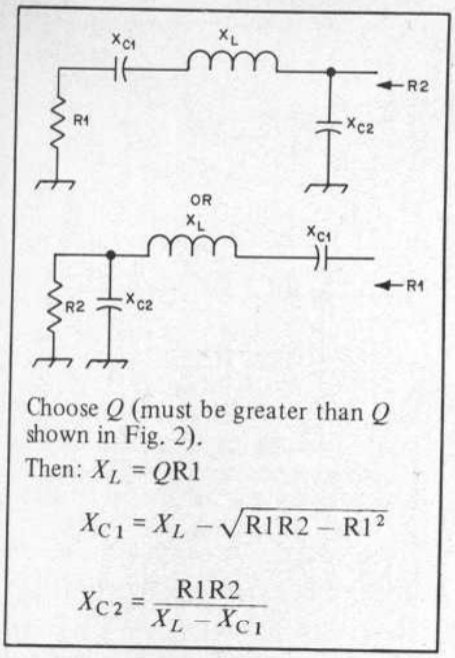


Fig. 2 — The L network and equations for using it.



Choose Q (must be greater than Q shown in Fig. 2).

Then: $X_L = QR1$

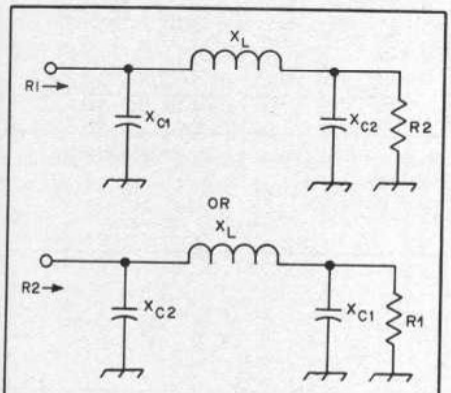
$$X_{C1} = X_L - \sqrt{R1R2 - R1^2}$$

$$X_{C2} = \frac{R1R2}{X_L - X_{C1}}$$

Fig. 3 – Example of a controlled-Q L network with equations.

required to perform any arbitrary impedance transformation. Such a design is realized most directly through the use of a Smith chart.

This simplified approach is sometimes dangerous, for it leaves the designer with no control over the Q of the network. If a three-element network is used, the designer has control over the impedance transformation, frequency and network Q. Occasionally, one will find networks with many additional components. The advantage of such designs is improved harmonic attenuation and greater bandwidth. In all of the three-element networks described next, it is necessary for the designer to specify Q at the beginning of the calculations.



Choose Q and R1 greater than R2.

$$X_{C1} = \frac{R1}{Q}$$

$$X_{C2} = R2\sqrt{\frac{R1/R2}{Q^2 + 1 - R1/R2}}$$

$$X_L = \frac{QR1 + R1R2/X_{C2}}{Q^2 + 1}$$

Fig. 4 – Pi-network configuration with design equations.

The L Network

This network is a classic for antenna matching, but also finds application for base and collector matching in solid-state transmitters with powers up to a few watts. It is not recommended for high-power amplifiers. The network is shown in Fig. 2 with the design equations. Note that R2 must be greater than R1. The Q of the network is given, although the designer has no control over this parameter. Q is an increasing function of the impedance-transformation ratio. This accounts for the undesirability of the network for high-power designs.

The Controlled-Q L Network

Some of the problems encountered with the standard L network can be minimized by adding a capacitor in series with the existing inductor. A Q is first chosen. Then, the equations shown in Fig. 3 are applied.

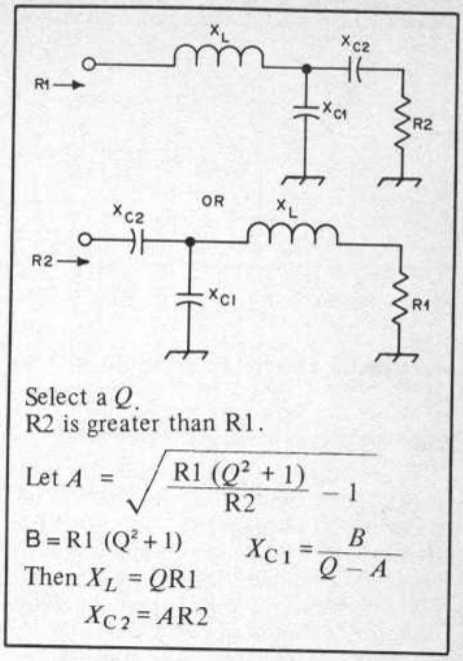
The Pi Network

A very familiar circuit is the pi network. It has served in the output tank of nearly every tube type of transmitter built in the last 20 years. A wide range of terminations can be accommodated, including those with substantial reactance, and the low-pass nature of the network provides excellent harmonic attenuation. The design equations are presented in Fig. 4. Manipulation of the equations will show that the impedance-matching range of the pi network is not unlimited. It may be shown that $Q^2 + 1$ must be greater than R1 ÷ R2. For example, a 10-to-1 transformation is not possible in a network with a Q of only 2.

Although useful in some transistor circuits, the pi network is not as popular as it was in tube-circuit days. The primary problem is that the component values dictated by the equations are sometimes less than practical. For example, it's not unusual when designing an 80-meter transistor transmitter to require inductors of 0.5 μH and capacitors of .01 μF. Networks other than the pi will lead to more practical component values for the same Q and impedance transformation. To generalize, the pi might be best for impedances of 50 ohms and higher on both ports.

The *L-C-C* Type T Network

One of the most practical networks for the low impedances common to transistors is a T network. It uses a pair of capacitors and a single inductor. Generally, the component values are practical if large-value mica-compression trimmers are used. This network is limited to the case of R2 being greater than R1. The equations defining this network are given in Fig. 5. The flexibility of this network is why it is often seen in manufacturers' data sheets for rf power transistors.



Select a Q.

R2 is greater than R1.

$$\text{Let } A = \sqrt{\frac{R1\,(Q^2 + 1)}{R2} - 1}$$

$$B = R1\,(Q^2 + 1) \qquad X_{C1} = \frac{B}{Q - A}$$

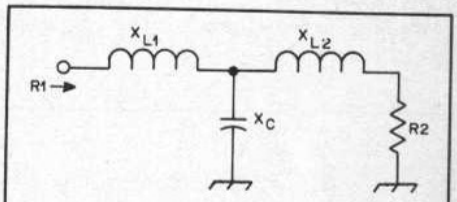
Then $X_L = QR1$

$X_{C2} = AR2$

Fig. 5 – The *L-C-C* matching network with related equations.

The *L-C-L* Type T Network

If two L networks are combined back-to-back, one obtains either a pi network or the T network shown in Fig. 6. This network has the advantage that the component values are often practical for solid-state citcuits. However, the difficulty in obtaining variable inductors with a wide tuning range makes the previous *L-C-C* T network more popular. The two-inductor T network, nonetheless, offers the ad-



Choose Q.

$$\text{Let } A = R1\,(Q^2 + 1)$$

$$B = \sqrt{\frac{A}{R2} - 1}$$

Then $X_{L1} = R1Q$

$X_{L2} = R2B$

$$X_C = \frac{A}{Q + B}$$

Fig. 6 – Circuit and equations of the *L-C-L* T network.

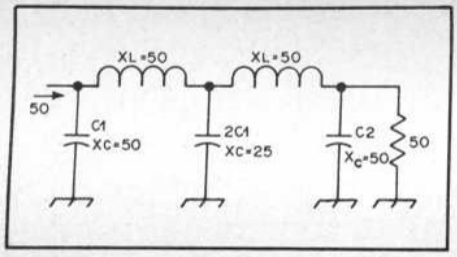


Fig. 7 – Half-wave filter network circuit.

vantage of excellent harmonic attenuation.

Additional Harmonic Attenuation

The primary purpose of the networks just presented is the transformation of impedances. If some of the circuits offer superior suppression of frequencies above their design center, that is certainly a point in their favor. However, it should not be a criterion for choosing one network over another, for harmonic attenuation is easily achieved after a transistor has been matched to 50 ohms.

A popular method for realizing additional harmonic rejection is adding a pi network in the 50-ohms line to the load. A convenient network is the symmetrical pi (50 ohms, in and out) with a Q of 1. In Fig. 4 we saw that this simplified pi section also has easy design equations. In this special case, we have $X_{C1} = X_{C2} \equiv X_L = R$, where R is the termination, usually 50 ohms. If two of these filters are cascaded, we have a network called the *half-wave filter*, shown in Fig. 7. This name results from the properties the network shares with a half wavelength of transmission line. That is, the phase shift through the network is 180 degrees and, more significantly, whatever impedance is used to terminate one end of the network is the impedance "seen" at the other end.

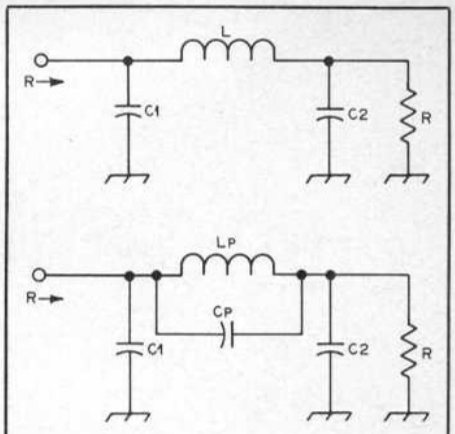
Presented in Table 1 are values for the components needed to build low-power half-wave filters for the amateur bands from 1.8 to 50 MHz.

Networks like the half-wave filter are modified easily to provide infinite attenuation at specific frequencies higher than the design center of the filter. This is realized by considering only half of the filter of Fig. 7. This symmetrical pi network with a Q of unity has the design parameters of $X_{C1} = X_{C2} = X_L = R$. At the design center frequency we can modify the filter by replacing any of the elements with more complicated LC combinations which have the same reactance. For example, the inductor which has a typical reactance of +50 ohms at the design frequency could be replaced with a trap consisting of a parallel LC combination. The behavior at the design frequency would be the same if the reactance of the series element were still +50 ohms. However, by properly choosing the components in the trap the filter will show virtually infinte attenuation at the frequency f_p, where the trap is self-resonant. The design equations for this case are shown in Fig. 8.

Broadband Matching Transformers

In the preceding section several impedance-matching networks were presented. One thing a careful observer might have noted was that the networks would be cumbersome to band switch. This difficulty can be avoided through the use of *broadband* matching transformers. Although these devices have appeared frequently in amateur literature in connection with solid-state linear amplifiers, they may be used equally well with Class C amplifiers at low or high power levels. Like the narrowband networks of the previous section, broadband transformers may be considered as singly terminated reciprocal networks.

Of the broadband rf transformers there are basically two types. One is essentially a conventional transformer which has been adapted for the low impedances common to high-power amplifiers (more on these transformers



Simple π section with $Q = 1$ and no Z transformation.

$X_{C1} = X_{C2} = X_L = R$

Let the L be L_o

Modified π section with trap freq. f_p

$X_{C1} = X_{C2} = X_{LO} = R$

$$L_p = L_o \left(1 - \frac{f_{op}^2}{f_p^2}\right) \text{ and}$$

$$C_p = \frac{1}{L_p (2\pi f_p)^2}$$

Fig. 8 – Modification of the half-wave filter to provide added harmonic attenuation.

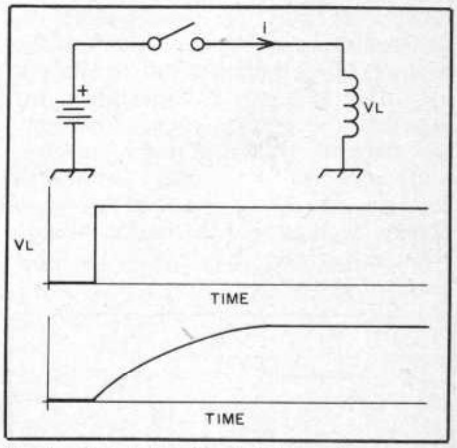


Fig. 9 – Principles of an ideal transformer, with waveforms.

$Q_L = 1$
$R_{in}, R_o = 50$ ohms
$X_L = 50$ ohms
$X_{C1}, C_3 = 50$ ohms
$X_{C2} = 25$ ohms

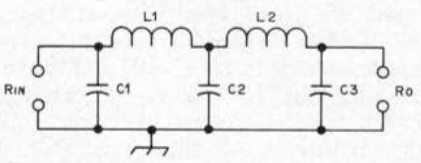


Table 1

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

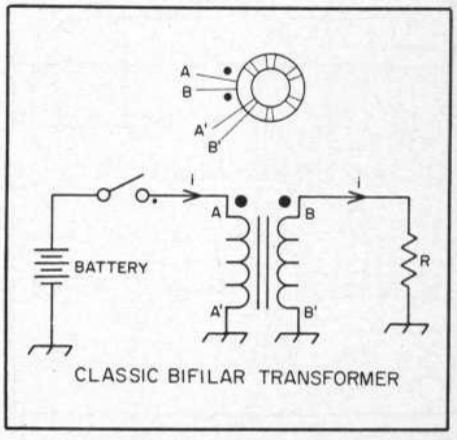


Fig. 10 – Illustration of current flow in a bifilar-wound transformer.

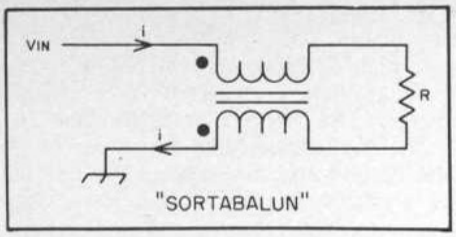


Fig. 11 – Circuit for an isolation transformer.

later in this chapter). The other configuration is the broadband transmission-line transformer. These transformers act as conventional transformers at their lower operating frequency, but act as transmission lines near their upper frequency limit. To attempt a complete explanation would be beyond the scope of this presentation. Hence, we will provide an overview and a few rules of thumb for the construction of the transmission-line transformers.

It is well known that a quarter wavelength of transmission line exhibits impedance-transformation properties. If a $\lambda/4$ length of line with a characteristic impedance Z_o is terminated with a resistance R1, a resistance R2 is seen at the other end of the line, where Z_o^2 = R1R2. For example, if a 35-ohm resistive load, such as the base of a ground-plane antenna, is placed at one end of a $\lambda/4$ length of 52-ohm coax cable, a resistance of 77 ohms is presented at the other end, offering a good match for RG-11 cable. The same principles apply for other kinds of lines (in this case, twisted pairs of insulated wire). Although the pitch of the twist can have some effect on characteristic impedance of the line, as does the wire diameter and insulation thickness, we will ignore these effects for the most part. One can assume generally that a twisted pair of plastic-covered hookup wire will have a Z_o of about 100 ohms. Similarly, a twisted pair of No. 24 enameled wires, twisted to about five turns per centimeter, will end up near 50 ohms. Twisted pairs are formed easily by clamping one end of the pair in a vise. The other end is hooked through a "fishhook" formed from large-diameter wire which is inserted in the chuck of a hand drill. With the wire held taut, the drill is operated until the proper pitch is obtained.

Twisted pairs could be used directly for transmission-line transformers except for a couple of problems. First, a quarter wavelength of line at, say, 80 meters is less than practical. This is where a toroid core comes in. The second problem is that the impedances usually needed for solid-state power amplifiers often dictate the use of low-impedance transmission lines, with Z_o well below 50 ohms. For example, an amplifier designed for an output of 6 watts from a 12.5-volt dc supply would require a load resistance of 12.5 ohms. A 50-ohm output termination could be transformed to 12.5 ohms by a line of Z_o = 25 ohms. This 25-ohm line is realized easily by paralleling two 50-ohm lines. Often, for the really low impedances needed for base matching, the required low-impedance lines are formed by paralleling as many as four or five line pairs.

We will now depart momentarily from our consideration of transmission lines and review the behavior of an ideal transformer. Consider first the relatively simple case of a single inductor, for example, a winding on a ferrite toroid. Recall that an inductor is a component in which the current flow cannot change instantaneously. If our hypothetical inductor is connected directly to a battery, the voltage across the battery immediately appears across the inductor. However, the current flowing in the inductor is initially zero: After all, the current was zero prior to application of the battery. The waveforms are shown in Fig. 9 along with the circuit. The fact that the current builds up slowly is a result of the changing magnetic field in the core. This changing field induces a voltage across the coil which impedes the flow of a net current. The current in the coil will, however, grow in time, leveling off at the level dictated by the internal resistance of the coil and of the battery. If we had ideal components with no internal resistance, the current would grow linearly forever.

Consider now the bifilar-wound transformer shown in Fig. 10. Again, we connect a battery to the primary of this transformer. In this case, however, current *can* flow instantaneously. As soon as the smallest current begins to flow in the primary, AA′, the resulting magnetic field causes a voltage to appear across the secondary, BB′. This voltage causes a current to flow through the resistor loading the secondary. This secondary current, in turn, establishes a magnetic field which *opposes* the field caused by the current in the primary. Hence, with a net magnetic field of zero, there is no inductive voltage to oppose current flow in the primary. The current flow is exactly the same as if the resistor were connected directly to the battery.

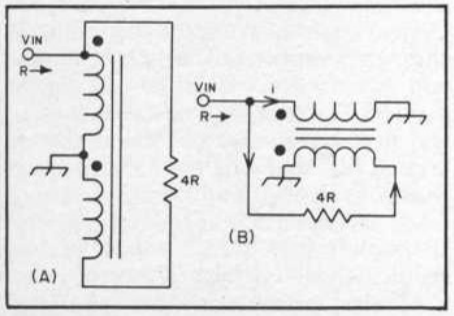


Fig. 12 – Circuit of a 4:1 step-up transformer.

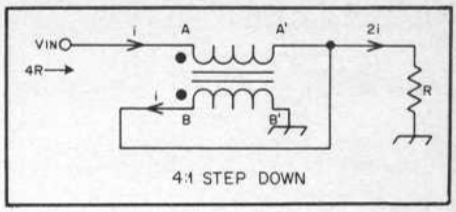


Fig. 13 – A 4:1 transformer which has frequent use in collector matching.

Since transformers work only on *changing* magnetic fields, the transformer will eventually cease to work when the core saturates. However, with ac signals, such as the rf of our present concern, the fields are always changing at rf rates.

It is important to note the direction of current flow and the dots in the figure which indicate voltage polarities. That is, a positive-going voltage applied at one dot will lead to a positive-going voltage at the other dot. The directions indicated for instantaneous current flow are those required for transformer action.

It is instructive to consider some of the transformer configurations which are of practical utility in rf design. Only some of the more straightforward types will be presented. Shown in Fig. 11 is an isolation transformer. This configuration is often called a balun, although it does not really deserve this name, for the transformer does not force the

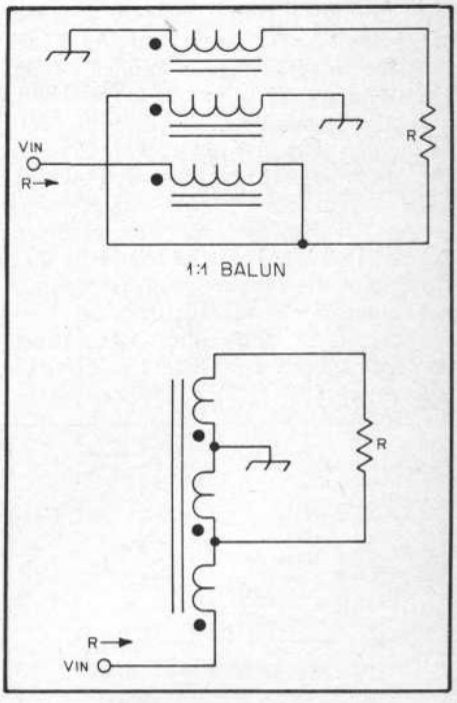


Fig. 14 – A 1:1 balun transformer.

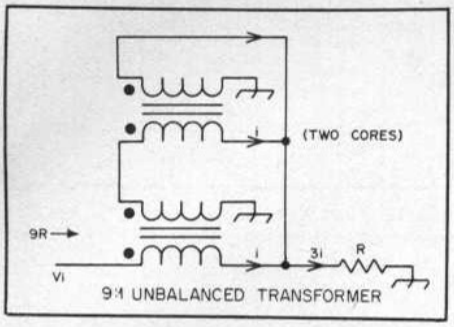


Fig. 15 — Illustration of a 9:1 unbalanced transformer.

voltage applied across the resistor, R, to be balanced with respect to ground. Indeed, if the end of the resistor connected to the primary were grounded, the input voltage would appear across R except that there would be a phase reversal. On the other hand, if R consisted of a pair of resistors in series, with their junction grounded, the input voltage would appear as a balanced, equal voltage across the balanced load. Because of the similarity to a balun transformer, WA6RDZ has suggested that this configuration be called a "sortabalun." Note in Fig. 11 that the current in the transformer is in the proper direction to preserve transformer action. However, any voltage common to both leads at one end of the sortabalun or the other will see a very high inductance, with minimal resulting current flow. Hence, the excellent isolation properties.

Presented in Fig. 12 is a 4:1 step-up balun (for real) transformer. The transformer is drawn in two different ways to emphasize the variety of approaches one can use in the analysis of such components. The sketch at A shows that the drive voltage is applied across one winding of a center-tapped coil, with the termination across both parts of the coil. The diagram at B emphasizes the direction of current flow which *must* exist for proper transformer action. Clearly, there is twice as much current flowing from the source as that flowing in the resistive load, implying a 4:1 impedance transformation.

Several of the other transformer configurations are presented in Figs. 13 through 16. The 4:1 type in Fig. 13 is commonly used for collector matching in medium-power amplifiers. Two transformers of this kind may be cascaded for a 16:1 transformation for matching 50 ohms to the base of a high-power stage. The 1:1 balun transformer (Fig. 14) is often used with balanced antenna systems. Note that this is a real balun rather than a "sortabalun."

The last two figures show transformers which use two toroid cores. The 9:1 single-ended configuration is useful for base matching in medium-power amplifiers. The 4:1 balanced-to-balanced configuration is sometimes used with push-pull high-power amplifiers. Typically, this 4:1 transformer is combined with an isolating sortabalun at the end, which must ultimately be terminated. Point "X" may be grounded if it is necessary to force balance. With care, either the 9:1 or 4:1 transformers can be wound on single cores. The reader is referred to *Motorola Applications Note AN-593* for this subject.

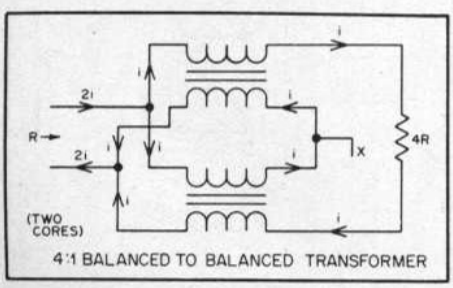


Fig. 16 — A 4:1 balanced-to-balanced transformer.

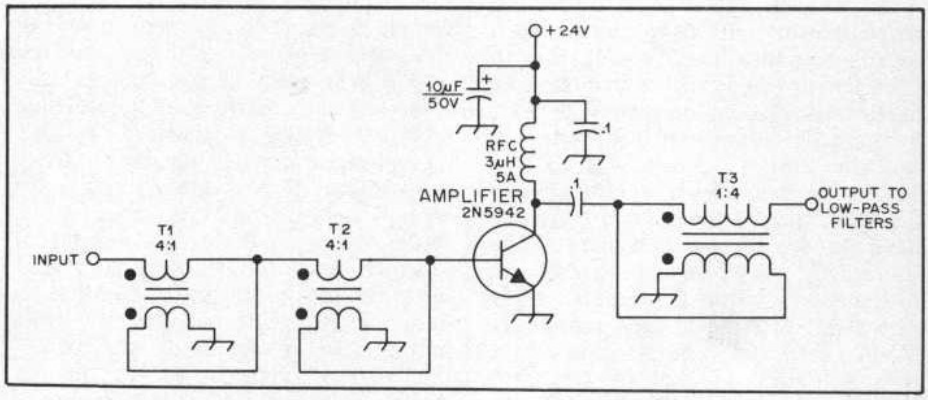


Fig. 17 — Circuit of a 25-W cw amplifier (see text).

Little has been said about the construction of practical versions of the transformers we have discussed. Fortunately, building them is straightforward. The first step is to obtain suitable toroids. Most of the toroidal cores used in amateur radio are of powdered iron and are used in tuned-circuit applications. However, for broad-band transformers, ferrite cores are preferred (μ of 125 to as great as 950). The main reason for this is that ferrite exhibits a much higher permeability than most of the powdered-iron cores used in the hf region. Because of the high initial permeability, the inductances required for good transformer action are realized with a minimum number of turns. This minimizes problems with self-resonances in the cores. Both ferrite and powdered-iron cores are available from Amidon Associates (see *QST* ads).

The next step is to consider the transmission-line requirements of the transformer. As mentioned earlier, at the higher frequency end of the operating range of most of these transformers the core has minimal effect. It is the transmission line which performs the desired transformation. The core is of significance only at the lower frequencies. The required end impedances of the transformer are first determined. For example, a 4:1 transformer for collector matching in the 6-watt amplifier mentioned earlier must match between 50 and 12.5 ohms. The required Z_o is given by $\sqrt{R1R2}$, or in this case, 25 ohms. This is obtained best with two paralleled 50-ohm lines.

It is helpful to make each twisted pair from enameled wire of two different colors. If this is not possible, it might be worthwhile to paint one of the wires with a suitable coloring agent, or tie a knot at each end of one wire. The twisted pair is made with a hand drill as outlined earlier. Then, two of these pairs are paralleled and twisted loosely with the drill. In this case, a couple of twists per centimeter is probably more than sufficient. This bundle of four wires is then wound through the core several times. The accepted rule of thumb is that the length of the winding should be 1/8 wavelength at the highest operating frequency, although much less wire will often work satisfactorily. Then, after winding, the ends of the wires are stripped of insulation. Assuming the two colors are red and green, the beginnings of the two red wires are twisted together as are the beginnings of the two green wires. The ends of the red and green wires are treated in a similar fashion. Having four wires now, we can assign the green wire as "A" and the red wire as "B" and wire the transformer as shown in Fig. 13. For transformers with lower characteristic impedances, similar procedures are followed with, of course, more than two paralleled twisted pairs.

Several transformers have been built and studied with a network analyzer. In both cases to be described, the toroids

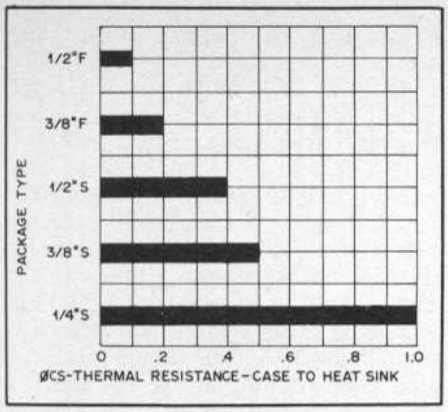


Fig. 18 — Representation of the thermal resistance of a transistor case to the heat sink (see text).

had an initial permeability of 125, and had an OD of 0.375 inch (Amidon Associates FT-37-61). The first case studied was a 4:1 transformer suitable for the output of a 25-watt amplifier with a 24-volt supply. Three turns of two bifilar pairs of No. 24 enamel wire were wound on a stack of four of the toroids. The high-impedance end of the transformer was terminated in 50 ohms, and the input impedance of the low-impedance port was measured. In scanning the range from 3.5 to 21 MHz, the measured impedance varied from 12.5 + *j*3.6 to 13.3 + *j*4.3. The slightly inductive impedance seen should present no problem in an amplifier, for the transistor is slightly capacitive.

The second case studied was a composite 16:1 transformer formed from two 4:1 transformers. The first transformer (50 ohms to 12.5 ohms) used one core wound with six bifilar turns of one twisted pair of No. 26 wire. The second used two twisted pairs on a single core, again only six turns. By the rules outlined above, the first core should have used two twisted pairs, and the second should have had eight! The cores, however, were too small to accept this much wire. In spite of the departure from the design ideals, the 16:1 transformer looked reasonable, although still inductive. With the high-impedance end of the composite transformer terminated in 50 ohms, the impedance seen at the other port ranged from 3.4 + *j*1.4 at 3.5 MHz to 3.2 + *j*4.5 at 21 MHz. The relatively high reactance would probably require some capacitive compensation at the higher frequencies.

A medium-power cw amplifier was breadboarded using the two transformers described above, and is shown in Fig. 17. The transistor used was a Motorola 2N5942. This device is specified for 80-watts PEP linear output, so it was loafing in the 25-watt test circuit. Nonetheless, the performance was just about that expected when tested at 7 and 14 MHz. An output of 25 W was obtained easily on both bands with a 24-volt power supply. The drive required on 20 meters was about 0.5 watt, while 250 mW were sufficient on 40 meters. No instability problems were noted.

High-Power Solid-State Amplifiers

There was a time when transistor transmitters were for low-power enthusiasts. It was not a matter of choice — the only transistors available were low-power devices. Today final stages with an output of 100 watts or more are practical and economical. In a few years the amateur may no longer be able to purchase a transceiver in this power class with even a single tube in the circuit. Through the use of hybrid-power splitters and combiners, a number of amplifiers in the 100- to 300-watt output class have been combined to yield over 1 kW of output. Most of the problems encountered in building a high-power amplifier are similar to those outlined earlier for low-power stages.

Almost all modern rf power devices are specified for operation in the frequency range for which they were designed. Most manufacturers' data sheets include curves of input resistance, input reactance and output capacitance as a function of frequency. Output load resistance is not often specified, since the equation ($R_L = V_{cc}^2 \div 2P_o$) is sufficiently accurate. With transistors specified for the hf region, most of the data are for linear operation. However, the information is close enough for use in designing Class C stages for cw and fm.

Heat Sinking and Mounting

The main difference between a high-power amplifier and one for QRP work is the level of heat sinking required. The efficiencies quoted by manufacturers vary, but a ball-park number might be 65 percent for Class C service, and 30 to 50 percent for Class AB or B linear amplification. The builder should expect that as much power will be dissipated in heat as will be obtained in rf power output. Certain prescribed methods should be followed to ensure long transistor life, as heat in excessive amounts (junction temperature) is one of the major enemies of power transistors.

The thermal resistance (resistance of a material to heat transfer) from a transistor case to the heat sink is anything but incidental. Fig. 18 shows typical values of thermal resistance for different package types when the devices are bolted to their heat sinks in accordance with the manufacturers' specified torque. The latter is usually 6 ±1-inch pounds for 3/8-inch studs, 5

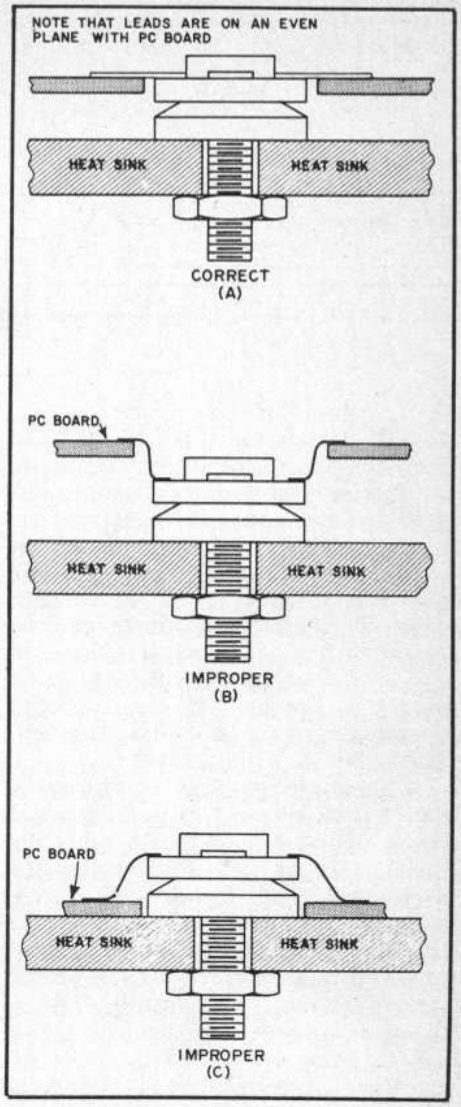


Fig. 19 — Correct and incorrect mounting methods for stud transistors with strip-line connector leads.

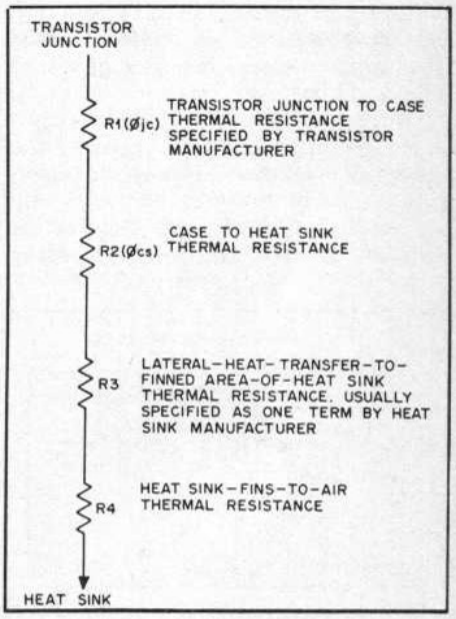


Fig. 20 — Resistances to heat flow when a transistor is joined to a heat sink.

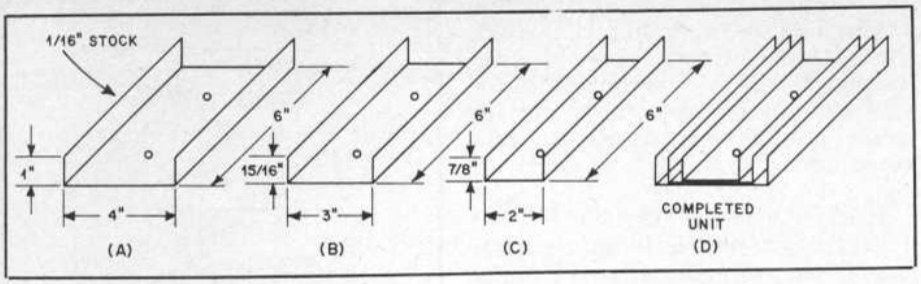


Fig. 21 – Details for forming a homemade high-power heat sink.

±1-inch pounds for 1/4-inch studs, and 8 ±1-inch pounds for 1/2-inch studs.

Thermally induced mechanical stress should not appear anywhere in the transistor. It is for this reason that correct torque is important. Furthermore, the surface of the heat sink to which the transistor case mates must be as smooth and flat as practicable. A thin layer of heat-transfer silicone grease should be coated on the stud and interface portions of the transistor and heat sink prior to mounting.

Strip-line types of transistors (wide, flat emitter, base and collector external leads) should be mounted so that the leads are not stressed. Furthermore, the circuit-board foils to which they connect should be brought as close to the transistor body as possible to prevent unwanted inductances from being formed by the strips (Fig. 19A). When the leads are bent as shown in Fig. 19B and C, stress exists, and may increase when heating occurs. A bad effect from bent emitter strips is that of degeneration caused by the excessive inductance which results. This will lower stage gain, and is a particularly significant matter as the operating frequency is increased to the upper hf region, and at vhf and uhf. Fig. 20 shows the resistances to heat flow which occur when a transistor is joined to a heat sink.

If the foregoing ideal guidelines can't be followed, the amateur can use the following procedure to assure safe operation. Start by bringing the power supply voltage up slowly, and monitor the collector current continuously. Make frequent checks of the transistor and heat-sink temperatures by touching a finger to each element. If the transistor body becomes too hot to endure with comfort, excessive heat will be present. This will indicate that the heat sink is not of adequate area, that thermal bonding is improper, or that excessive collector current is flowing. If a torque wrench is not available, tighten the stud nut just beyond the point where it is finger tight. Transistor mounting and heat considerations are treated in *Motorola Application Note AN-555*, and in *Solid Circuits* by Communications Transistor Corp. of San Carlos, CA.

It is not necessary to purchase heat sinks if aluminum sheeting is available. Large heat sinks can be fashioned from U-shaped pieces of heavy-gauge aluminum, as shown in Fig. 21. Homemade sinks are inexpensive and can be put to use quickly.

The use of wide pc-board foils is recommended in rf portions of the circuit. Wide foils will lessen the unwanted inductance effects, and will make soldering of the transistor strip lines easier. An illustration of the principle is given in Fig. 22. Double-sided pc-board material (copper on both sides) is almost mandatory in the interest of electrical stability. The side opposite the foils and transistor body serves as a ground-plane surface to discourage current loops which can cause feedback. Additionally, the ground plane acts as one plate of a capacitor for each of the etched foils, affording vhf and uhf bypassing throughout the board. This also helps prevent unstable operation. The ground-plane side of the board should be made electrically common to the ground foils on the etched side of the board.

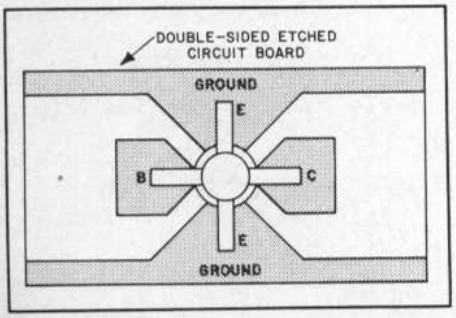


Fig. 22 – Example of a recommended pc-board foil pattern for use with stud-mount strip-line transistors.

Some Electrical Considerations

It is practically impossible to lay down a definite rule for selecting a power transistor which must deliver a specific output power. Commercial designers have, on occasion, pushed power transistors quite hard – extracting power amounts which were as great as 3/4 P_D maximum. That is, a transistor with a miximum safe power-dissipation of 10 W at 25°C might be called upon to deliver 7 watts of rf output when installed on an adequate heat sink with correct mounting techniques. In amateur work that kind of courage is not recommended. A transistor operated within sensible ratings should last for 100,000 hours of "on" time, at the least. That kind of longevity would not be typical of an amateur amplifier if it were "milked" for all it was worth. A good rule of thumb is to select a transistor which has a $P_{D(max)}$ of roughly twice the power it will deliver. It is not especially wasteful of money and device capability to make the safety margin even greater. When more power output is needed than the P_D rule of thumb can assure, use a larger single transistor, or two in push-pull, instead of paralleling two smaller ones. This will reduce cost somewhat, and will make the circuit less difficult to optimize. When two or more devices are used in parallel, layout and load-sharing problems become difficult to predict and control.

It is not recommended that vhf or uhf transistors be used in mf and low-hf band power amplifiers. The gain (Fig. 23) increases markedly as the operating frequency is lowered (6 dB per octave), and this can make stabilization extremely difficult. It is best to utilize transistors which were designed for the frequency range of interest. Furthermore, a power transistor should be operated at a power-output figure which is 75 to 80 percent of the saturated power output. That approach will assure best efficiency and will reduce power drop-off with heating. (Saturated power output is that point where further output can't be obtained with increased drive.)

Gain Compensation

Broadband amplifiers require gain equalization if a wide range of frequencies must be accommodated, say, 1.8 to 30 MHz. It was said earlier that transistors have increasing gain at ap-

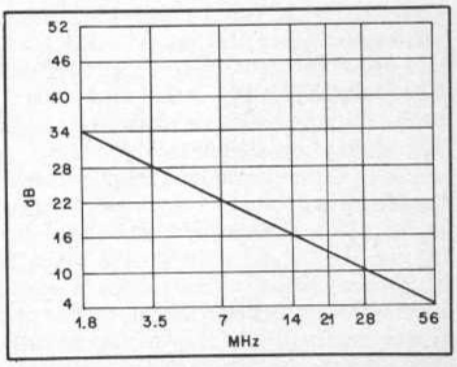


Fig. 23 – Curve showing the 6-dB-per-octave gain characteristic of a transistor.

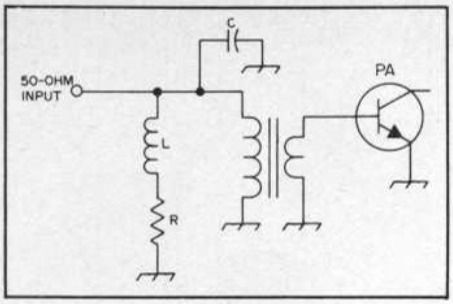


Fig. 24 – Gain-compensating network, labeled L and R.

proximately 6 dB per octave lower, which means that very high gain is probable at the low end of the amplifier range. It is desirable to equalize amplifier gain as much as possible to prevent the necessity for a variable drive-power exciter, and to prevent damage to the transistors from parasitic oscillation or excessive collector current at the low end of the operating range. Two forms of compensation are popular, and each requires some empirical adjustment.

One technique is to add a "losser" network at the input to the amplifier (Fig. 24). Inductance *L* is selected to have low reactance in the range where the gain increase is significant, and as the operating frequency is made lower, the loss through the compensating network increases. Addition of resistance *R* serves a twofold purpose – it lowers the network *Q* and provides a load for the driving power that must be dissipated external to the transistor. It is sometimes necessary to add component *C* to correct for a mismatch caused by the compensating network. Depending on the capability of the exciter with respect to SWR, a moderate amount of mismatch may be tolerable at the low end of the amplifier frequency range. Generally, input SWR should be made lowest at the high end of the amplifier operating range.

Another technique used by some designers to equalize amplifier response is to employ negative feedback (collector to base). The method is illustrated in Fig. 25 for a single-ended amplifier, and in Fig. 26 for a push-pull module. The principle is one of adding an *R-C* network which has the property of increasing the negative feedback as the operating frequency is lowered. The component values depend on the device characteristics, power levels, and impedance characteristics of the amplifier. Therefore, no set rules for component values are offered here. (See chapter 8 for details.) Typical values for a 50-W amplifier might be 10 ohms and 100 pF for the circuits of Figs. 25 and 26, assuming an amplifier bandwidth of 1.8 to 30 MHz. It should be said that any compensating network a builder may add to an amplifier will have some effect on the circuit, and caution should be used when such *L-C-R* sections are included in a design.

Ballasted Transistors

Modern power transistors for linear-amplifier service are emitter ballasted. That is, each emitter or group of emitters in a device (several bipolar transistors are used in parallel on a single substrate) contains a separate series resistance. This feature helps prevent hot-spotting on the chip (second breakdown) which can occur anywhere on the complex internal surface. Hot spotting takes place when one or more of the individual transistors on the substrate "hog" power. The result is failure of the composite transistor. The series resistances tend to equalize the current sharing as changes occur externally, thereby protecting the transistor from damage. The possibility of second breakdown is related mainly to linear transistors (but also affects Class C amplifiers) because forward bias is applied. Therefore, when SWR is high, or when strong self-oscillation takes place, hot-spotting is likely to become manifest. Ballasted transistors are excellent for all classes of operation – A, AB, B and C.

A protective measure for unballasted transistors is seen in Fig. 27. A Zener diode is connected as a peak-voltage clamp from collector to ground. Assuming the maximum collector voltage swing will be twice the supply amount (24 V), VR1 is not part of the collector circuit. However, should a load mismatch occur, or the stage break into self-oscillation, the collector rf voltage will soar to high value. At that point VR1 will conduct at 36 V and clamp the voltage above that value, thereby protecting the transistor. Furthermore,

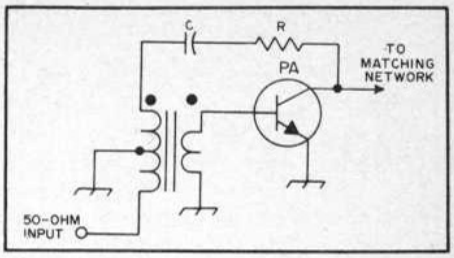


Fig. 25 – Negative-feedback gain compensation using *C* and *R* components.

should voltage spikes occur on the supply line the Zener diode will clamp at 36 V or higher again protecting the transistor. If protection against excessive positive and negative voltage swings is desired, two Zener diodes can be bridged from collector to ground, back-to-back fashion. ARRL lab tests indicate that no degradation in amplifier performance results from use of Zener-diode clamps at hf and mf, provided the diode conduction point is well above the normal rf-voltage peak value. No evidence has been found that VR1 enhances the generation of harmonic currents while in its "off" state.

Protective measures should be assured for any piece of solid-state equipment which operates from a dc supply that is not treated for transient suppression. Notably, mobile gear which uses the automotive ignition supply for operating voltage can be subjected to large voltage spikes that can ruin the transistors or ICs. A good safety precaution is to add an 18-V, 10-W Zener diode from the 12-volt input line to ground. The same principle applies to equipment which is powered by ac-operated dc supplies that have no spike-protector circuits.

Conventional Broadband Transformers

Considerable treatment was given

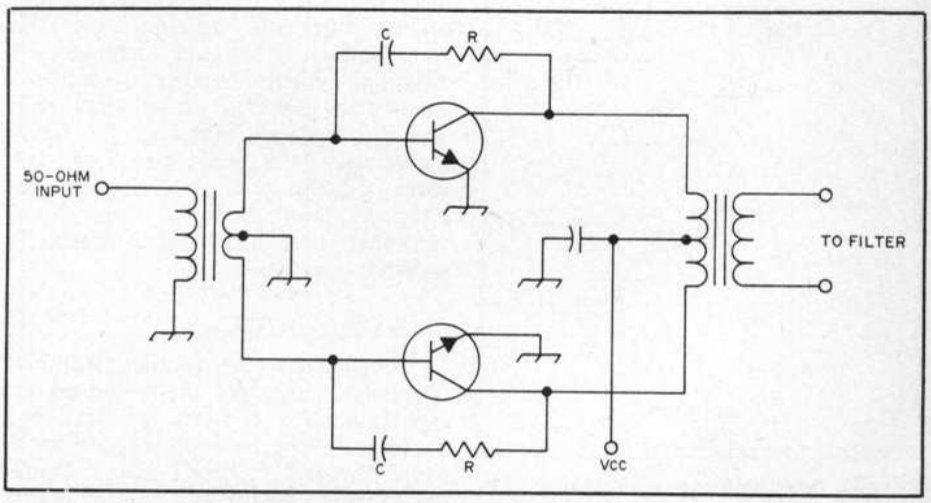


Fig. 26 – Gain compensation networks for negative feedback.

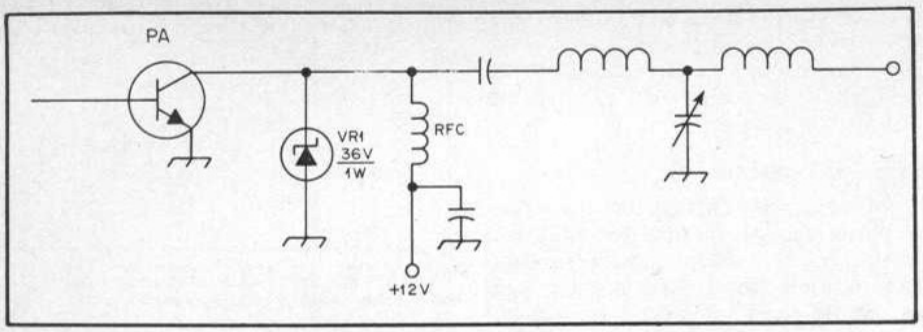


Fig. 27 – Example of a Zener-diode protective clamp at the collector of a power amplifier.

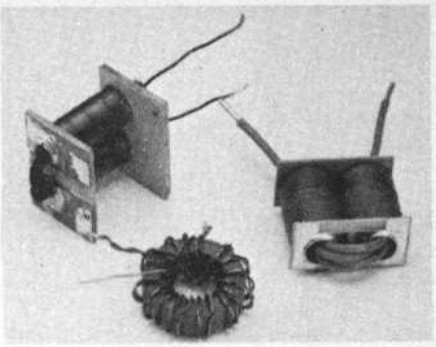


Fig. 29 – Photograph of some conventional and transmission-line transformers. The unit with the twisted wires (center) is a transmission-line transformer.

earlier to the design and use of transmission-line transformers for broadband applications. It is worth mentioning that conventional broadband transformers are also suitable for many amateur circuits. A number of commercial manufacturers are using conventional transformers in their power blocks, and with good results.

Most broadband rf transformers of the "conventional" type are toroidal and use iron or ferrite cores. However, ferrite rods can also be used as the core material in conventional broadband transformers. The self-shielding properties of toroid cores are preferable in most amateur work, however.

Fig. 28A shows the electrical representation of a conventional broadband transformer. L1 is a small-diameter brass or copper tube which is U-shaped, and over which several high-μ toroid cores have been placed (permeability = 950 in most designs for mf and hf). The ends of the tubes are soldered to the pc-board plates as shown at B. U-shaped L1 functions as a 1-turn secondary winding, and is hooked to the bases of push-push amplifier transistors when T1 is used as an input transformer. Alternatively, T1 can be used as an output transformer, in which case, the ends of L1 connect to the collectors of the amplifier, or to the balanced winding of a collector rf choke. L1 establishes the turns ratio of the transformer by virtue of its being a 1-turn winding. Insulated hookup wire is passed through the tubes of L1 and serves as the primary winding (L2) of an input transformer, or as the secondary of an output transformer. The number of turns used will depend upon the impedance-transformation ratio needed.

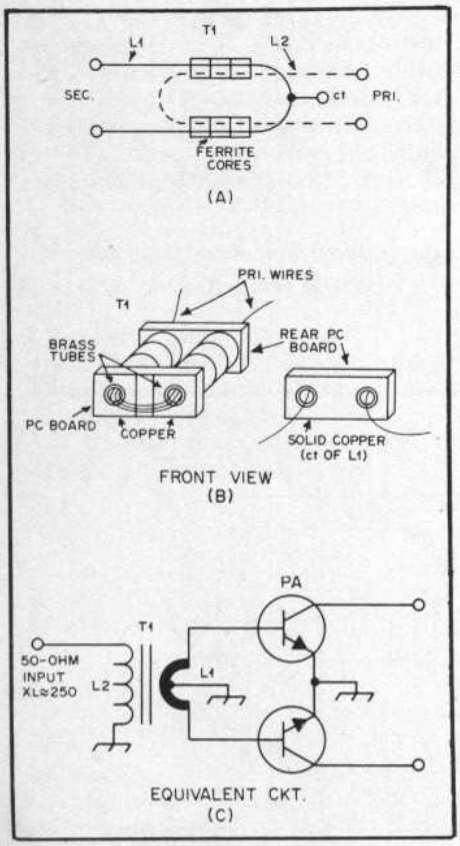


Fig. 28 – Circuits of a conventional broadband transformer with sketch of how they are constructed (see text).

The number and size of the ferrite cores used will be related to the power level of the amplifier and the desired reactance of the windings. A good rule of thumb is to make the transformer windings exhibit four to five times the impedance of the circuit to which the transformer is connected. Thus, a winding that connects to a 50-ohms load should look like, say, 250 ohms at the lowest operating frequency.

One advantage of the conventional transformer of Fig. 28 (and many transmission-line transformers) is that excellent symmetry results from the construction style, and symmetry is essential when obtaining electrical balance in push-pull power amplifiers. The pc-board end plates of the transformer can be soldered directly to the main pc-board pads to which they relate. A photograph of some conventional broadband transformers is shown in Fig. 29.

Other Considerations

Occasionally, the amateur will use transistors which have the f_T and power capabilities for rf-power applications, but lack the specifications needed for a really complete "paper" design. These devices can often be used for amplifiers by making reasonable estimates of the parameters. Some of the guessing procedures will be outlined without detailed justification. First, the f_T of the device being well above the operating frequency (a factor of three, four, or more) will ensure that a reasonable power gain is available. The V_{CEO} of the device should exceed the operating voltage, V_{cc}, by a factor of two or more in cw and linear applications. Ideally, the beta of the transistor should hold up well at the desired collector current. If these criteria are met, the operating parameters are easy to guess. The output resistance needed is $V_{cc}^2 \div 2P_o$. Usually, the output capacitance (C_o) can be ignored: It can be absorbed in the output tuning network of a narrowband design. Broadband designs may present more problems, however. The input resistance is related to the current gain at the frequency of operation and is inversely related to the output power. For amplifiers in the 20- to 70-watt output region, one can arrive at a satisfactory design by assuming an input resistance of around 2 ohms. If an *L-C-C* type of T network is used for matching, with a design Q of 5, input resistances of less than 1 ohm may still be accommodated without excessive network Q (see Fig. 5). It is possible to neglect the input reactance of the base, allowing the reactance to be absorbed in the impedance-transforming network. As a

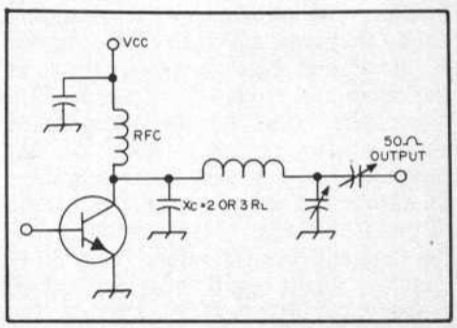


Fig. 30 – Circuit of the modified *L-C-C* network.

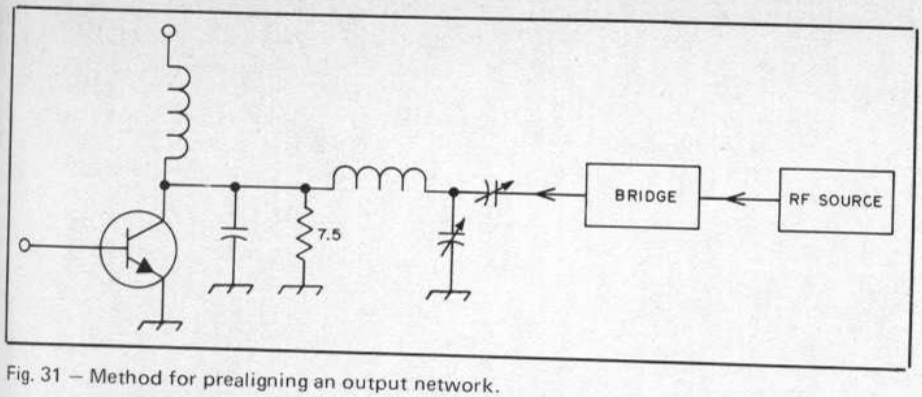


Fig. 31 – Method for prealigning an output network.

conservative rule of thumb, one should never design for an output power exceeding the heat dissipation of the transistor being used. *Less* is a better and safer assumption.

As was outlined earlier, there is a wide variety of networks from which to choose for impedance matching. However, the *L-C-C* type of T network is an excellent first choice for base and collector matching, owing primarily to the range of impedances which may be accommodated with a given network design, and to the practicality of the component values. It is worthwhile to modify the output network slightly by adding some additional capacitance in parallel with the collector. A reasonable value is a reactance of two or three times the output resistance, Fig. 30. This added capacitance will have little effect at the design frequency, but will significantly aid in the suppression of vhf parasitics. This is of major significance if a vhf power device is used in the hf region.

One can pretune the networks to the design frequency and impedance before power and drive are applied to an amplifier. This prealignment is done easily with a 50-ohm impedance bridge and a low-level rf source. (A suitable bridge is described in a later chapter.) As an example, assume that an amplifier will deliver an output of 50 watts with a 28-volt power supply. The collector load resistance will be $V_{cc}^2 \div 2P_o = 7.8$ ohms. The network is designed and a reasonably close-value resistance is connected temporarily to the circuit as shown in Fig. 31. In this example, an appropriate resistance would be a pair of paralleled 15-ohm carbon resistors. The network is adjusted for a bridge null, indicating that 50 ohms exists at the output port. The 7-1/2 ohm resistor is then removed from the circuit!

Shown in Fig. 32 is the input part of a power amplifier. The rf choke serves as a dc path for the flow of base current. Since the input resistance of the transistor is very low, the reactance of this choke is not critical and is usually four or five times the input resistance. However, the Q of this choke should be quite low, often less than 1. This is realized by shunting the choke with a low-value resistor, less than the reactance of the choke. Even lower values (down to an ohm or two), comparable to the value of the transistor input resistance, will add to the stability of the amplifier. If this practice is followed, the input network may be prealigned with a bridge without substitution of extra base resistance.

Once an amplifier is built and prealigned, the moment of truth comes when dc power and rf drive are applied. The output is terminated in a 50-ohm resistive load with means for measuring power output. The light bulb load of the tube era has *no place* in the modern amateur lab, and should not be used as an rf termination! A current-limited power supply should be used. Initially, the voltage is reduced to half of the normal operating level in the case of high-voltage amplifiers (e.g., 28 volts). For stages operating from 12 volts it is suitable to begin experimentation at that level. A low amount of rf drive is applied and the output is noted. The networks are adjusted for maximum output, always keeping an eye toward signs of instability. This procedure is repeated at increased power-supply voltages and rf drive levels, keeping the networks tuned for maximum power output. The collector current should be monitored for any tendency toward thermal runaway, and the device and heat-sink temperature should be monitored.

If the amplifier has forward bias, as is typical of linear amplifiers, *careful* attention should be devoted to monitoring the current during application of rf drive, and afterward. Many amplifiers which perform well in ssb service may not be capable of withstanding the tremendous power dissipation levels incurred during cw testing or two-tone evaluation.

A final problem which can occur with high-power amplifiers should be mentioned. Often the collector current in a high-power amplifier is several amperes. With such a high current it can be extremely difficult to decouple the amplifier from the remaining circuit. Additional decoupling networks may be

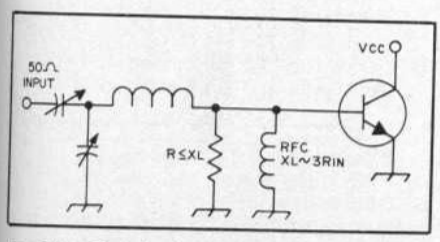


Fig. 32 – Circuit of the input part of an amplifier.

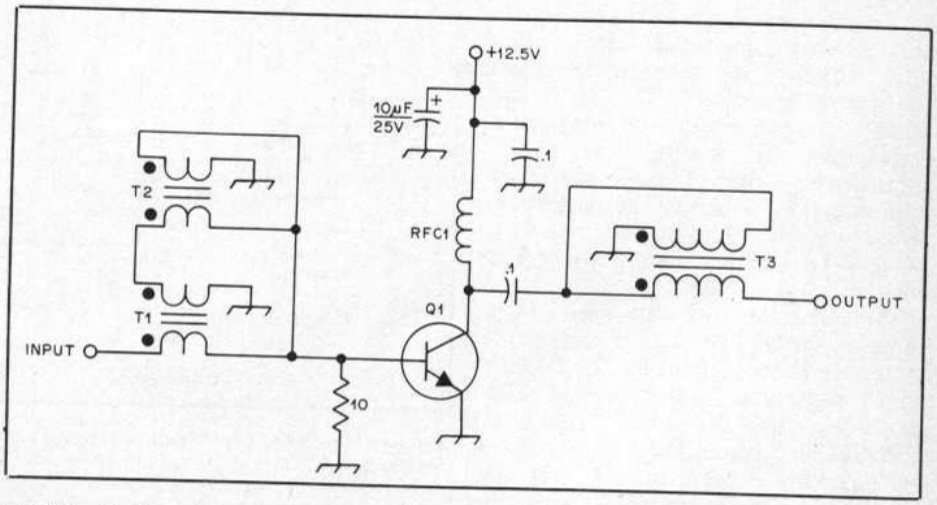
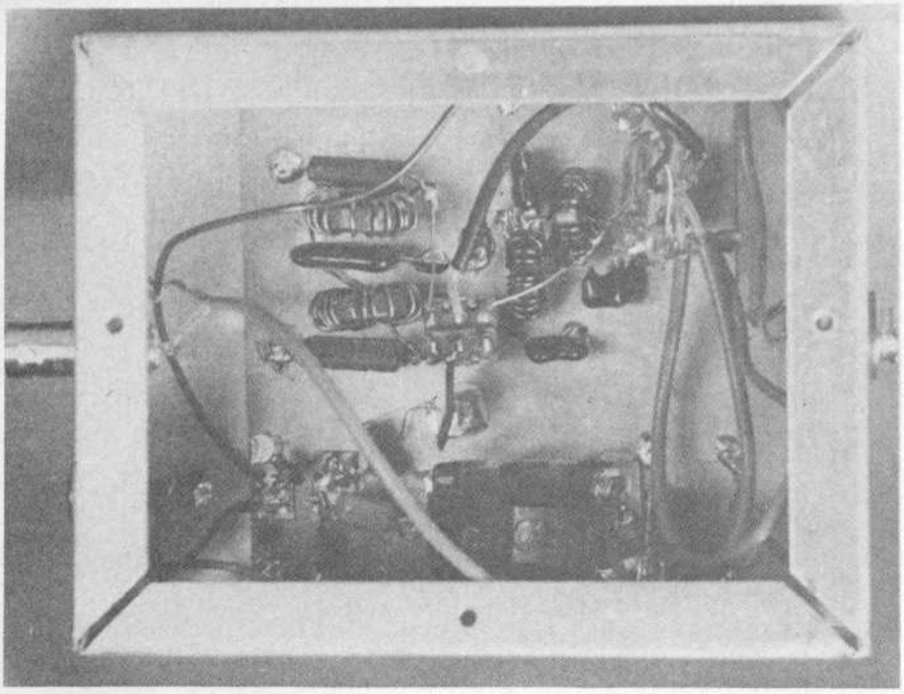


Fig. 33 – Single-ended 4- to 6-W amplifier. RFC1 is a 25-μH choke capable of passing 1 ampere. See text for discussion of Q1. T1, T2 and T3 contain 7 bifilar turns of two twisted pairs of No. 26 enamel wire on Amidon FT-37-61 toroid cores.



General-purpose 6-watt rf amplifier which uses a single transistor. The amplifier is seen at the bottom of the photograph. This WA7MLH unit contains half-wave output filters for 80 and 40 meters, plus a small relay which, when actuated, bypasses the amplifier for QRP operation.

required. In the home station it is worthwhile to operate a high-power final stage from a power supply separate from that used to power the rest of the station.

Broadband Utility Power Amplifiers

Many QRP transmitters built by the experimenter have an output of a watt or less. The amplifiers shown in Figs. 33 and 34 are designed to complement such rigs, providing outputs of four to six watts, while not presenting a strain on the pocketbook. Both designs use broadband matching transformers of the type outlined in a section earlier. They are suitable for the amateur bands up through 20 meters.

The simpler of the pair of amplifiers (Fig. 33) has a single-ended design using one transistor. All three transformers are wound identically. T1 and T2 are wired as a composite 9:1 step-down transformer such that the base of the transistor is driven from a source of approximately 6 ohms. The output resistance of 12 ohms is matched to a 50-ohm termination with T3, which is wired as a 4:1 step-up.

Several transistors were tried in the single-ended configuration. Excellent results were obtained with the GE D446C, which is available for just over $1. This device has an f_T of 50 MHz, a 30-watt collector dissipation, and a V_{CEO} of 45 volts making it ideal for rf-power applications on the lower bands. With this transistor, output powers of 6 watts have been obtained on 80 meters, with 4-1/2 to 5 watts being more typical for 40 meters. The power gain is roughly 10 dB on 40 meters. It approaches 16 dB at 3.5 MHz. Versions of this amplifier have been used by West Coast amateurs for the output of QRP transceivers which were designed specifically for Field Day use. Since the efficiency is about 50 percent, the amplifiers are ideal for the 10-watt input limit in the QRP category. The RCA 2N5321 is worth investigating as a substitute in this circuit.

Shown in Fig. 34 and in the photograph is a push-pull version. Although slightly more complicated than the single-ended amplifier, this scheme is worth consideration. First, the push-pull version has the advantage of twice as much power dissipation. Furthermore, even-order harmonics are suppressed by the balanced circuit. Finally, and this is of significance when using inexpensive transistors not intended for rf power application, a higher output-load resistance may be used. This allows reasonable efficiency to be maintained without requiring that the transistors have good saturation specifications.

In the push-pull amplifier T1 steps the input 50-ohm drive down to 12.5 ohms in a single-ended manner. T2 then provides drive to the balanced bases. The third core in the input section of the amplifier ensures that the load presented to T2 is balanced. Each transistor sees a driving impedance of 6-1/4 ohms. In chapter 2 it was noted that at low frequencies a problem sometimes encountered with power stages is breakdown of the emitter-base diode of the transistor. The use of push-pull circuitry prevents this from happening, for each transistor acts like a negative-clamping diode for the other.

In the output T4 plays two roles. First, it provides a path for the dc bias to reach the collectors. Saturation of the toroid is no problem at high currents, since the current through the opposing windings sets up opposing fields. The second purpose of T4 is to ensure that the collectors are balanced. T5 transforms the balanced drive from the collectors to a single-ended 50-ohm termination. Note that an impedance of 25 ohms is presented to each collector.

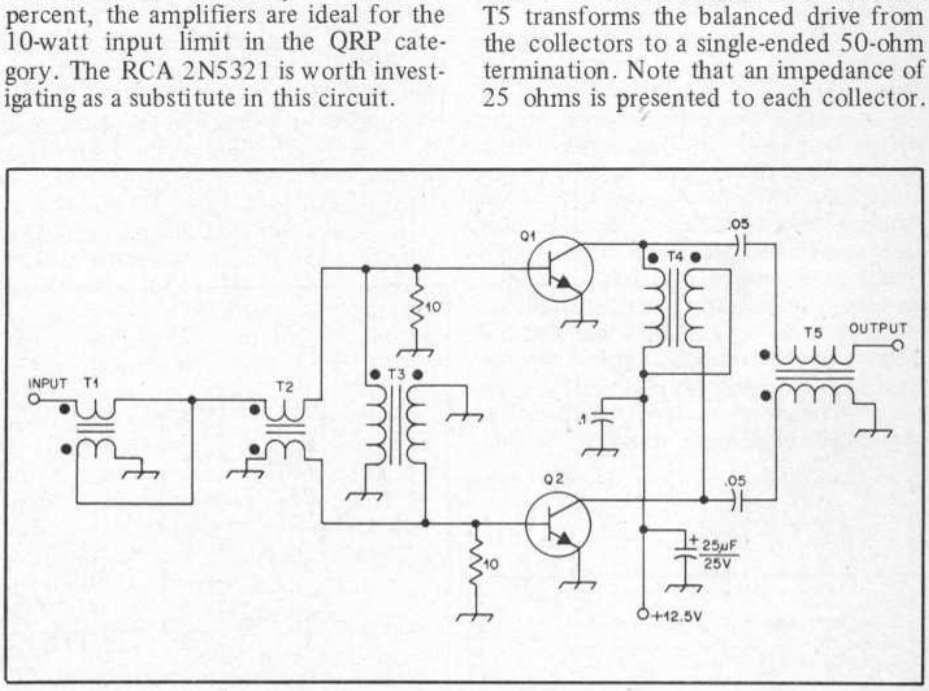
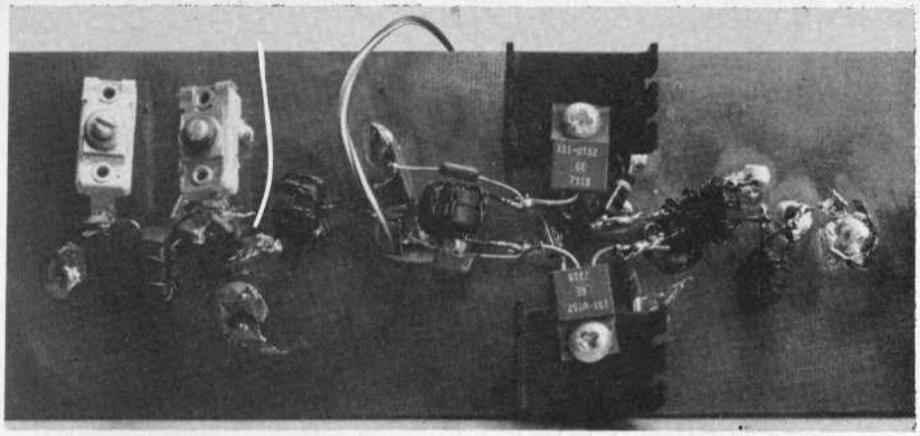


Fig. 34 — Circuit of a push-pull broadband amplifier of the 4- to 6-W class. Filtering is necessary at the output of this amplifier and the one in Fig. 33 to prevent harmonics from being radiated by the antenna system. Q1 and Q2 are GE D44C6 units. T1, T2 and T3 contain 6 bifilar turns of No. 26 enamel wire (twisted pairs) on Amidon FT-37-61 toroid cores. T4 is the same as T1, but two cores are stacked. T5 has 6 bifilar turns of a single twisted pair of No. 26 enamel wire on two stacked FT-37-61 toroid cores.



General-purpose broadband push-pull amplifier. This view shows the breadboard version of the circuit. The transistor mounting bolts affix the transistors to the heat sink and extend through the pc board. Insulating washers are used. The network at the left was used for filtering during initial tests on 20 meters. Power output is 6 watts from 1.8 to 14 MHz. In excess of 20 watts can be provided by this amplifier at 7 MHz if a 24-volt power supply is used.

The push-pull amplifier was tested on 40 and 20 meters. At 7 MHz, the measured output was 5-1/2 watts with a drive of 0.5-watt. The efficiency was 59 percent. These measurements were with V_{cc} = 12.5 volts. With a 24-volt supply, over 12 watts of output were obtained with 0.5 watt of drive power. On 20 meters, 5 watts of output were seen a 12.5-volt supply. However, 1 watt of drive was required. While only 7 dB of power gain is marginal, it is still useful. The amplifier should perform well on 80 meters, and delivered 18 W on 160 while using a 24-V supply.

Both amplifiers should be followed by a filter to remove harmonics. The half-wave filters described earlier should be adequate. WA7MLH built one of the single-ended amplifiers with half-wave filters for 80 and 40 meters. The output low-pass filters are selected by means of a slide switch. A relay is included to switch around the unit during low-power operation.

A Design Exercise

Assume that a 7-W amplifier is needed for 160 or 80 meters. To minimize the chance for high levels of even-order harmonic output a push-pull circuit is chosen. Another criterion is to design for low cost, particularly with respect to the transistors and heat sinks. Available driving power is approximately 1 to 2 watts. Some measure of burn-out protection is wanted should a high output SWR occur. Finally, the amplifier should cover at least 100 kHz of either band without need to retune the collector tank.

The foregoing may seem like a tough assignment, especially if undertaken by a beginner. Actually, the chore is easier than it may seem. Nearly all of the information needed to effect such a design has been provided in the preceding text. Transistor selection, network design, and heat sinking have been tested in sufficient depth to make a simple amplifier design possible.

Transistor Choice

It was stated earlier that the transistors used in an amplifier should carry a P_D rating of approximately twice the desired rf power output. Therefore, to extract 7 W we should use a pair of transistors whose combined power-dissipation rating will be 14 W or greater. Also, the f_T should be several times the highest operating frequency (5 or 10 times as a ball-park number). This calls for an f_T of 17 to 35 MHz, or thereabouts. Maximum voltage ratings should be somewhat greater than two times the operating voltage, which suggests a safe value of 30 or more.

A search through various data showed that an RCA 2N5320 should do the job nicely. The price per unit is roughly \$1.50, f_T is 50 MHz, and maximum collector voltage is 100. Maximum dissipation at 25°C is 10 W for a 2N5320, providing a 20-W rating for two of the units. The junior version of the 2N5320 might be used for a 5-W maximum output power in the push-pull amplifier of Fig. 35. The device is a 2N2102, designed specifically for high-speed, high-voltage switching. It has an f_T of 120 MHz, which makes it suitable from 1.8 through 14 MHz. The price tag is approximately \$1, and the $P_{D(max)}$ is 5 W.

Networks

For the sake of simplicity a conventional broadband transformer, T1 of Fig. 35, is selected for the amplifier input port. It will have a turns ratio of

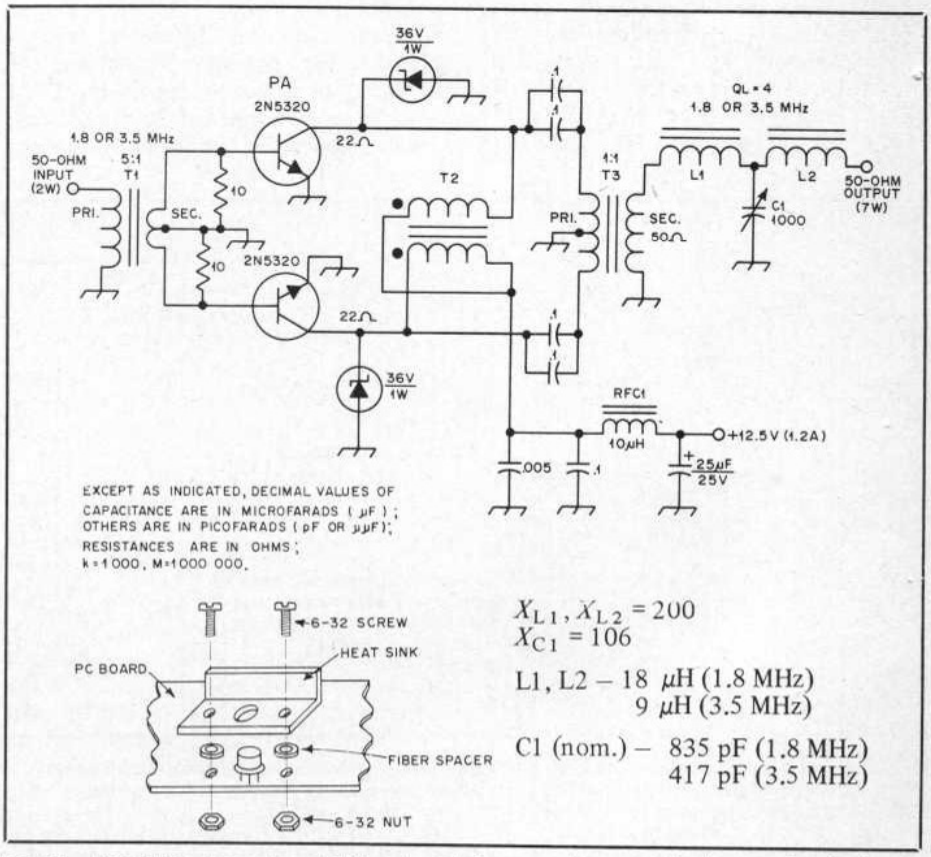


Fig. 35 – Circuit for the push-pull 7-W output design example treated in the text. Details for a practical heat sink are shown here.

approximately 2.2:1 for a Z ratio of 5:1, assuming a total secondary impedance of 10 ohms (a close approximation for a base-to-base impedance of 10 ohms). The primary inductance should be at least 17 μH for 50 ohms at 1.8 MHz, or 9 μH for 3.5 MHz (X_L of 4 times 50 ohms = 200 ohms). A 3/8-inch diameter Amidon ferrite toroid with a μ of 125 will be suitable when wound with sufficient No. 28 enamel wire to obtain the necessary inductance. The number of secondary turns is ratio-related to the 2.2:1 figure, and are set by the number of primary turns necessary to obtain an X_L of 200 ohms or greater.

A 10-ohm, 1-W resistor is connected from each transistor base to ground. This will help stabilize the amplifier by lowering the Q of T1. Final adjustment of T1 can be made with the amplifier operating at rated output power into a 50-ohm resistive load. An SWR indicator is placed between the exciter and T1; then the primary turns of T1 are reduced or increased until an SWR of approximately 1 is obtained.

A balanced-collector choke is needed for T2. Since the collector-to-collector impedance for 7 W of output is roughly 44 ohms for a 12.5-V dc supply, the choke should have an X_L of approximately 175 total, or 88 per half. That comes to 15 μH at 1.8 MHz, or 7.7 μH at 3.5 MHz. The wire size for the winding should be able to pass the collector dc current without causing an $I \times R$ drop. Each transistor will draw approximately 0.6 A at the rated dc-input power level, suggesting that No. 24 enamel wire will be suitable.

T2 can be wound bifilar fashion (8 twists per inch of wire) on a piece of ferrite rod (Amidon 0.5-inch diameter stock) about two inches in length, or on a 1-inch diameter ferrite toroid core. Q1 ferrite will be suitable (μ = 125) in either case. The phasing should be as shown in Fig. 35.

T3 is a conventional transformer wound with No. 24 enamel wire to have a primary inductance of approximately the same value used at the primary of T1. The secondary winding of T3 should have the same inductance. Although a calculated Z ratio of 1.13:1 is appropriate for T3, and 1:1 ratio (total primary to secondary) will be acceptable. A 3/4- or 1-inch diameter Q1 ferrite core will be adequate at T3.

L1 and L2 can be Amidon powdered-iron cores (T-68-2), wound with sufficient No. 24 enamel wire to provide the required inductance. A loaded Q of 4 was chosen for the T network to assure ample bandwidth and minimum chance for amplifier instability.

C1 is a large mica compression trimmer of 1000-pF maximum capacitance. A J. W. Miller No. 160-A was used in the ARRL test model. Fixed-value silver-mica capacitors can be used in place of the trimmer by combining them to obtain 835 or 417 pF, as specified on the diagram.

RFC1 is a dc decoupling choke of low inductance value. A 10-μH value will suffice for either band. It can be made by winding a 0.5-inch diameter Q1 toroid core full with No. 24 enamel wire.

Zener diodes are used at each collector for transistor protection in the event of a severe mismatch. The diodes will have no effect upon performance during normal conditions. They need not be included if it is unlikely that a high SWR will be seen.

Table 2

BAND	L1	L2	C1	C2
7 MHz	0.6 μH, 13 T No. 22 enam., 5/16" ID, no core	1.1 μH, 14 T No. 22 enam. on T-68-2 toroid core	450-pF mica trimmer	820-pF silver mica
14 MHz	0.3 μH, 8 T No. 22 enam., 5/16" ID, no core	0.55 μH, 9 T No. 22 enam. on T-68-6 toroid core	450-pF mica trimmer	220-pF silver mica
21 MHz	0.19 μH, 5 T No. 20 enam., 5/16" ID, no core	0.39 μH, 6 T No. 22 enam. on T-68-6 toroid core	450-pF mica trimmer	None

L1 coils are airwound. L2 coils are on Amidon toroid cores.

Heat Sinks

Each transistor will need its own heat sink. A simple homemade variety

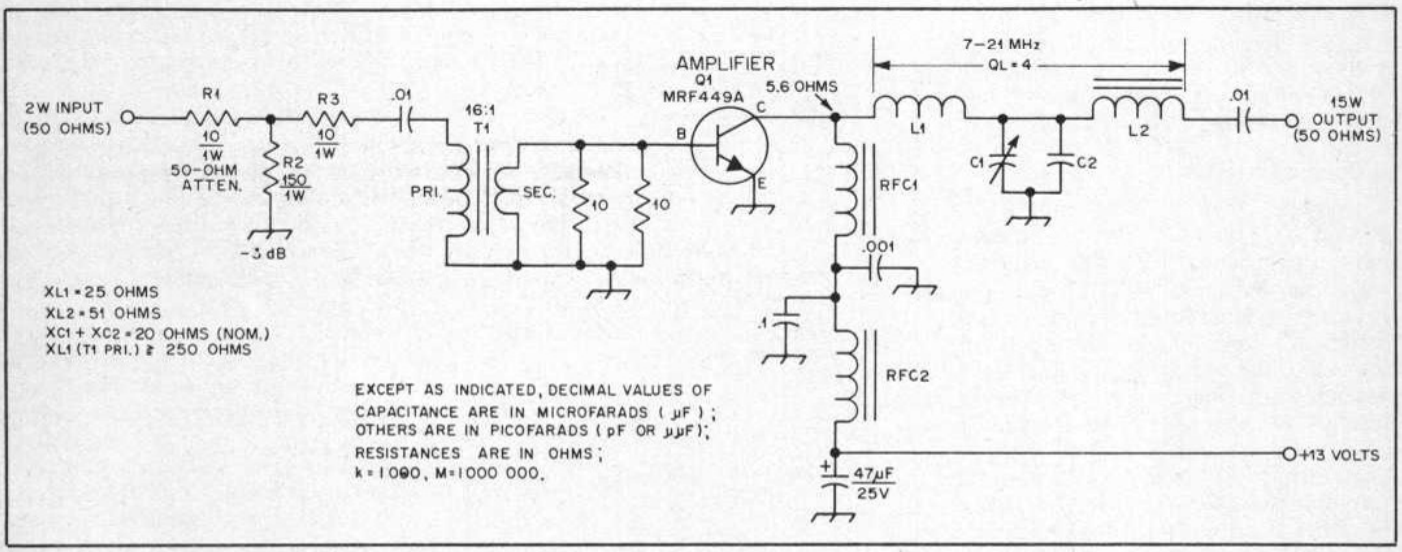


Fig. 36 — Schematic diagram of the 15-watt amplifier. Fixed-value capacitors are disk ceramic unless otherwise noted. Resistors are 1/2-watt composition unless specified differently. The 47-μF capacitor can be electrolytic or tantalum.

C1 — 450-pF mica compression trimmer (Arco-El-Menco 466 or equivalent).
C2 — See Table 2.
L1, L2 — See Table 2.
Q1 — Motorola MRF449A strip-line stud transistor.
RFC1, RFC2 — 7 turns No. 20 enam. wire on 0.5-inch OD toroid ferrite core with 125 permeability (Amidon Assoc. FT-50-61 core or equiv.), 3 μH.
T1 — Primary, 32 turns No. 24 enam. on Amidon T-68-2 core (7μH). Secondary, 8 turns No. 24 over primary winding.

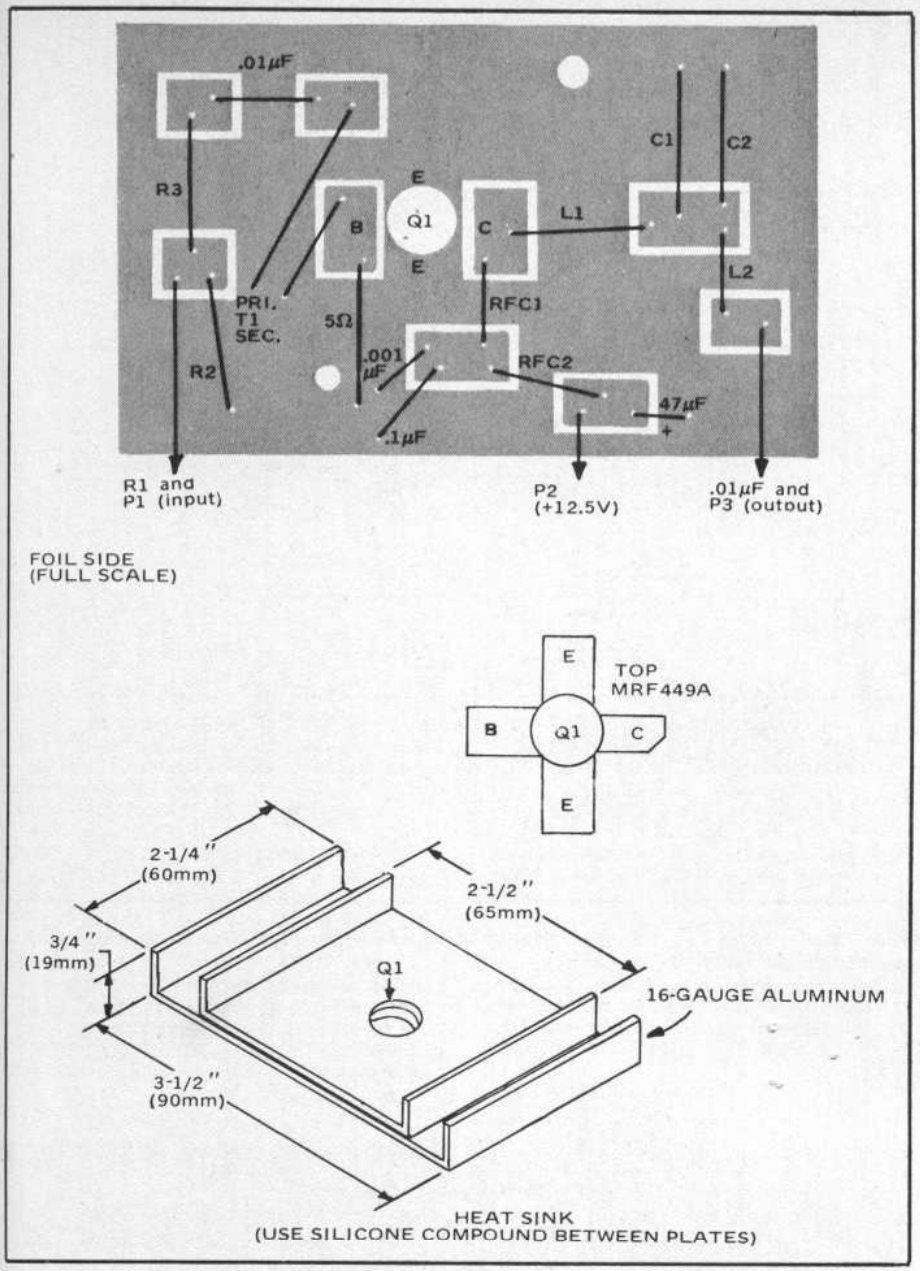


Fig. 37 — Scale layout of the 15-W amplifier pc board. Double-clad board (copper on both sides) is used, and the ground foil on the etched side is connected to the ground-plane side at several points. Details are given for the homemade heat sink.

can be fashioned from 2-inch sections of Reynolds hardware-type aluminum angle bracket (see sketch in Fig. 35). The heat sinks have clearance holes for the transistor cases, and a snug fit is necessary to assure proper heat transfer. Silicone grease should be placed on the transistor body where it mates with the sink. Each sink is isolated from ground by means of insulating-spacer washers. The foil on the bottom of the pc board should be removed so that the 6-32 mounting nuts are isolated from ground. The foil material on the top of the pc board should be removed where the 6-32 bolts pass through it. The heat sinks are snugged down against the transistor flanges by means of the mounting bolts.

Results

A laboratory breadboard of the circuit was built and tested for 1.8 and 3.5 MHz. Performance was smooth (no instability), and an output of 8 watts was obtained on 160 meters. A 7-W output was secured on 80 meters. Examination of the output waveform showed a clean sine wave on both bands. Second harmonic energy was down some 40 dB, and all other harmonics were at least 50 dB below the fundamental frequency. When the amplifier is loaded properly into 50 ohms, the tuning of C1 will be fairly broad, but a definite peak in output will occur when it is set correctly.

15-Watt HF-Band Amplifier

One advantage of high-gain transistors is that they can provide considerable output power for low-drive levels. The Motorola MRF449A is one choice a designer has among the high-gain hf-band devices. It is designed for a power output of 30 W maximum, Class C, when used below 30 MHz. A 13-V power supply is required. Power gain is rated at 13 dB at 30 MHz.

The circuit of Fig. 36 shows how it can be used in a single-band cw amplifier with an efficiency of 60 percent. The circuit was described originally in *QST* for December, 1975, where it was specified as a plug-in amplifier for the Heath HW-7 QRP transceiver. The 3-dB resistive attenuator at the amplifier input is included so that exciters having more than 1 watt of output will not overdrive the transistor. The HW-7 delivers 2 watts of output, so 1 watt is absorbed in the attenuator. Also, the attenuator provides a constant 50-ohm load for the exciter. The addition was necessary because the MRF449A requires only 3/4 to 1 watt of drive to produce full output. Those having exciters in the 1-watt class can delete the attenuator.

T1 is a conventional input transformer which is wound on a T-68-2 powdered-iron toroid. It provides a necessary 16:1 transformation ratio (50 to 3 ohms). Two 4:1 broadband transmission-line transformers were tried in cascade to replace T1, and results were identical to those with the transformer specified. The conventional transformer was used because only one toroid was required. To lower the Q of T1 a pair of 10-ohm, 1/2-W resistors have been strapped from base to ground.

Power Level

A power-output level of 15 watts was chosen to minimize power-supply

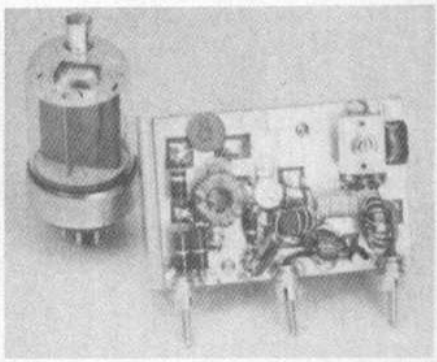


Fig. 38 — Photograph of the assembled amplifier. The circuit-board pads of Fig. 37 replace the phono plugs shown here.

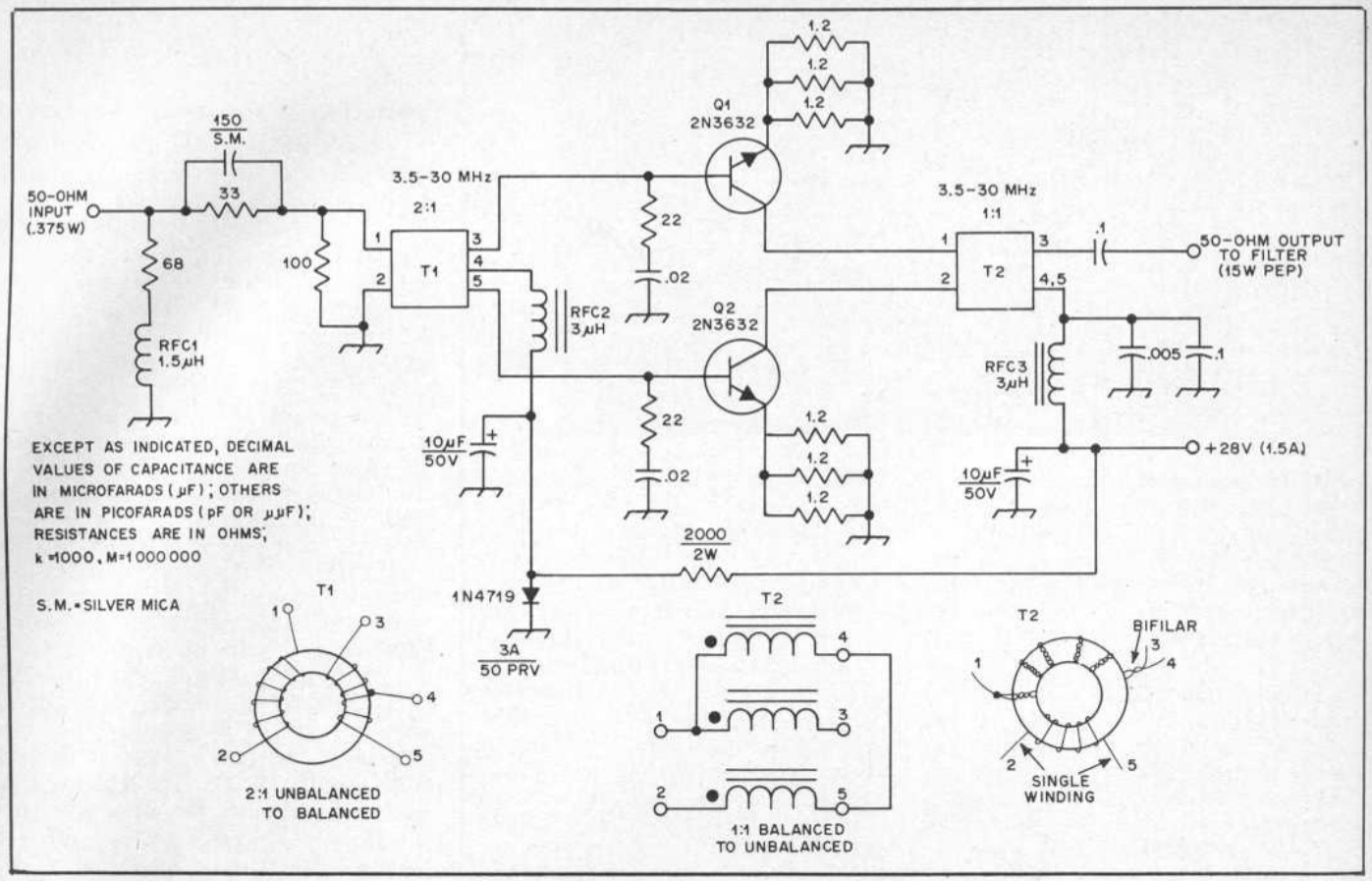


Fig. 39 – Schematic diagram of the 15-watt linear amplifier. Resistors are 1/2-W composition unless otherwise noted. Capacitors are disk ceramic unless specified differently. Polarized capacitors are electrolytic or tantalum.

RFC1 – Miniature 1.5-μH choke.

RFC2 – 15 turns No. 26 enam. wire on Amidon FT-37-61 toroid.

RFC3 – 7 turns No. 20 enam. wire on Amidon FT-50-61 toroid.

T1 – Conventional broadband transformer, 2:1 impedance ratio. 14 turns No. 28 enam. wire on Amidon FT-50-61 toroids (two cores stacked). Secondary has 10 turns of No. 28 enam. wire over primary winding.

T2 – Broadband 1:1 transformer. 15 turns No. 24 enam. wire (bifilar wound to 8 turns per inch) for winding 1/3/4. Winding 2/5 contains 15 turns of single No. 24 enam. wire. Use two Amidon FT-50-61 cores, stacked.

drain for field use. The network values are based on that power amount (Table 2), but there is no reason why the full 30-W output amount cannot be realized. The collector network would have to be revised accordingly. If that were done, a collector characteristic of 2.8 ohms would result. Therefore, a T network with a loaded Q of 4 would require an $XL1$ of 11, an $XL2$ of 7, and an $XC1$ of 12. The circuit was tested at the 30-W level and performance was good. However, a slightly larger heat sink than that shown in Fig. 37 will be necessary at the higher power amount. The dimensions for T1, RFC1, and RFC2 are suitable for either power value. A 50-W version of the '449A is available for those wanting more power. It is the MRF450A. Approximately 2 watts of drive power are needed for full output. Operating voltage is 13 for the latter also. Both transistors are stud-mount types, and each has strip-line connecting leads.

Specifications are not given for 160, 80, or 10 meters, but there is no reason why the builder could not develop suitable L and C values for the T network from the reactances listed in Fig. 25. At 80 and 160 meters there may be a tendency toward instability, owing to the higher gain of the transistor at those frequencies. An additional 10-ohm resistor from base to ground should resolve the problem. Alternatively, the negative-feedback technique shown in Fig. 25 could be applied to enhance stability.

The output waveform as viewed on a 50-MHz scope was very clean. Harmonic energy was at least 40 dB below carrier level. Fig. 38 is a photographic view of the module.

A 15-Watt Linear Amplifier

The amplifier of Fig. 39 was adapted from one which was described by Lowe (*QST* for Dec., 1971, p. 11). The basic difference is in the transformer design (T1 and T2). The Lowe transformers were similar to that of Fig. 28 in this chapter, but many amateurs had difficulty duplicating them, so the broadband transformers of Fig. 39 were developed. Performance remains essentially the same regardless of the transformer style employed. Lab tests with a spectrum analyzer show that both versions provide an IMD (3rd- and 5th-order products) of –30 dB.

A peak output power of 15 watts is available on ssb, and 15 watts of output are provided for cw work. Forward bias is supplied to the transistor bases to prevent cross-over distortion. (See chapter 8.) Idling current (no drive applied) is approximately 100 mA with 28 volts of collector supply. Peak current drain is 1.5 A.

Although the amplifier is designed for 3.5 to 30 MHz, good performance was noted on 160 meters with approximately 1 watt of drive. The original version by Lowe was not tested at 1.8 MHz, however.

The input port contains a complex RCL compensating network to level the amplifier gain by compensating the drive level. Amplifier gain is 16 dB at 15 and 10 meters, and is slightly greater on

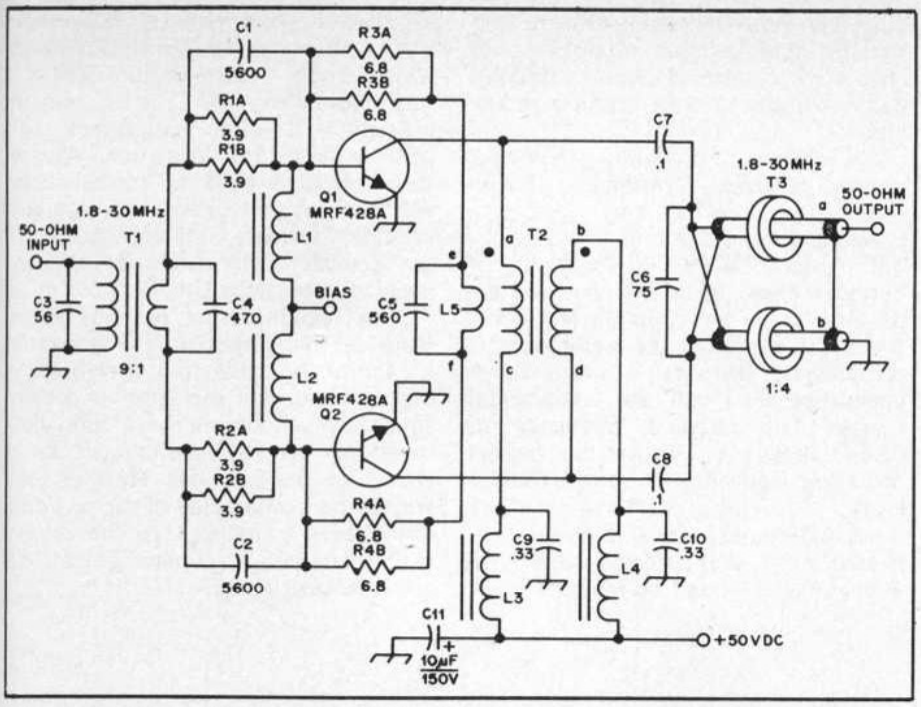


Fig. 40 — Schematic diagram of the 300-W output linear amplifier designed by Granberg. Capacitors are the ceramic chip variety except for C11, which is electrolytic. Numbered components not described here are so identified for layout purposes on the pc-board pattern offered in the *QST* series.

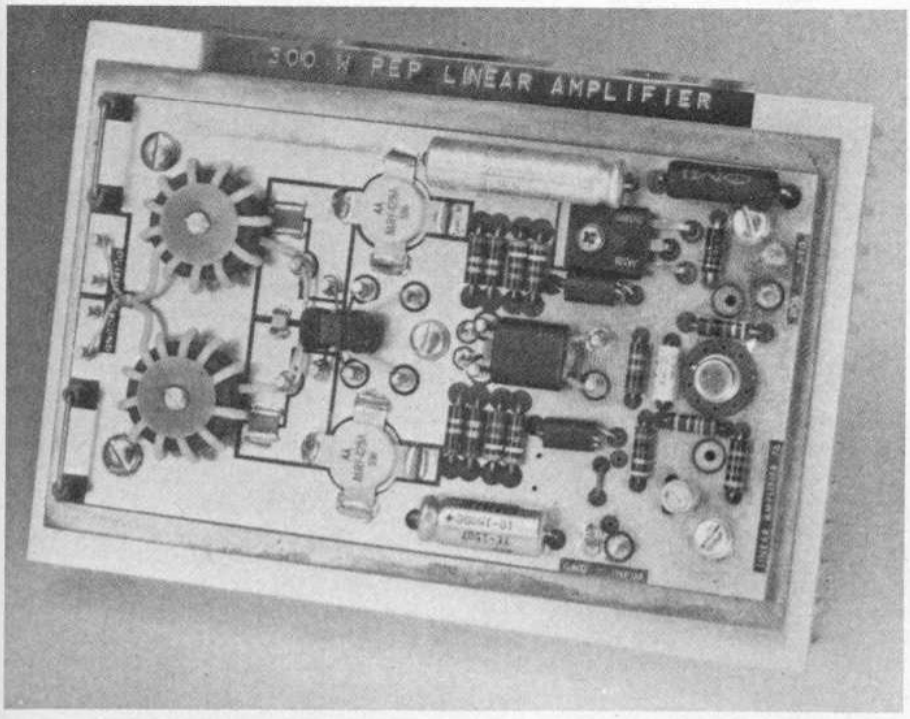
L1, L2 — Rf choke (Ferroxcube VK200-19/4B or equiv.).

L3, L4 — Rf choke (Ferroxcube 56-590-65/3B or equiv.). For these chokes and other Ferroxcube components contact Elna Ferrite Labs., Inc., 9 Pine Grove St., Woodstock, NY 12498.

T1 — Broadband 9:1 transformer on ferrite core (TV balun type Stackpole 57-1845-24B, Fair-Rite 2973000201, or Amidon equivalent of latter). Low-Z winding has one turn of 1/8-inch OD copper braid to serve as tubing. Primary contains three turns of No. 22 Teflon or enamel-coated wire.

T2 — 7 bifilar turns of No. 22 enam. wire on Stackpole 57-9322 or Indiana General F627-8Q1 toroid core. A suitable substitute core would be two Amidon FT-50-61 cores, stacked.

T3 — 14 turns Microdot 260-4118-000 25-ohm submin. coaxial line or equiv., wound on each of two Stackpole 57-9074 or Indiana General F624-19Q1 cores. A probable substitute is the Amidon FT-114-61 toroid.



Photograph of the assembled 300-W amplifier. Note 1/4-inch thick copper plate between the double-sided pc board and the aluminum heat sink.

20, 40 and 80 meters. Input SWR through the compensating network is less than 1.5:1 from 80 through 10 meters.

Q1 and Q2 are low-cost surplus vhf transistors. The 2N3632 is designed for Class A, B and C service. Maximum V_{CEO} is 65, maximum collector current is 3 A, and f_T is 400 MHz. Maximum dissipation is 23 W at 25°C.

A finned heat sink measuring 4 X 4 inches or greater is required for safe operation. Double-sided pc board is used to contain the amplifier. Output SWR should never exceed 1.5:1 if damage to the transistors is to be prevented. Although the even-order harmonics from the amplifier are at least 20 dB below the fundamental signal, filtering should be used at the output. The half-wave filters described earlier in the chapter will be suitable.

A 300-Watt-Output Linear Amplifier

This chapter would not be complete without an example of a high-power linear amplifier. The circuit of Fig. 40 shows a design by H. Granberg (WB2BHX/7) of Motorola, which is one module of a 1200-W composite amplifier (four power blocks combined). He described the latter in *QST* for April and May, 1976.

This circuit contains two Motorola MRF428A transistors. An operating voltage of 50 is required and current taken is approximately 13 A. The circuit is broadbanded for use from 1.8 to 30 MHz. Full output can be obtained with a driving power of 5 W, as observed in ARRL laboratory tests. Harmonic filtering is required at the amplifier output during on-the-air use.

The module contains a bias regulator

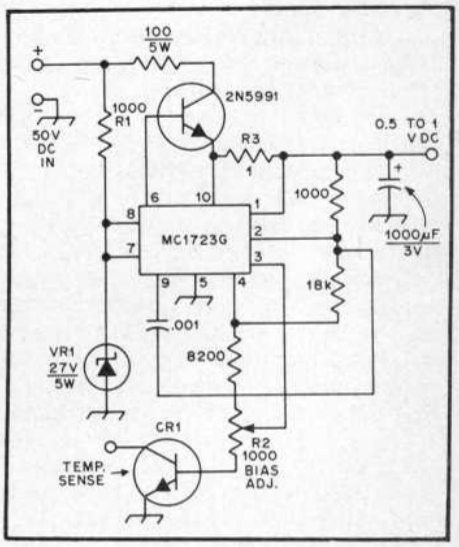


Fig. 41 — Schematic diagram of the bias regulator and temperature sensor.

to provide forward base voltage for linear operation. Fig. 41 shows the circuit. Variable bias voltage is available by means of R2, providing a range from 0.5 to 1 V, regulated. CR1 is the base-emitter junction of a 2N5190. It has a plastic case and is used as a circuit-board standoff spacer. It serves as a temperature-sensing diode. By virtue of its being coupled to the heat sink it assures automatic temperature tracking with a slight negative coefficient. When the collector idling current is set for 300 mA at 25°C, the current will decrease to a nominal 250 mA when the sink temperature rises to 60°C. The rate of change is approximately 1.15 to 1.7 mA per degree C.

In Fig. 40 a 9:1 input transformer is used, providing an impedance step-down from 50 ohms to a 5.5-ohm base-to-base characteristic. Negative feedback is employed to enhance stability and to help equalize amplifier gain. Approximately 5 to 6 dB of feedback can be utilized without impairment of linearity or stability.

In addition to providing a source for negative feedback, T2 supplies dc voltage to the collectors and serves as a center tap for output transformer T3. The currents for each half cycle are of opposite phase in *ac* and *bd*, and depending upon the coupling factor between the windings, the even-harmonic components will see a much lower impedance than will the fundamental energy. The resonant frequency of C5-L5 should be above the highest operating frequency to prevent instability.

A 4:1 transmission-line transformer is used at T3. It is a coaxial-cable type, with *a* and *b* wound on separate cores. This technique eliminates the need for having three separate transmission lines, which would be the requirement if a single core were used. The line sections consist of 25-ohm miniature coaxial cable with Teflon insulation. Alternatively, twisted pairs of enamel-coated wire can be used to form 25-ohm lines (discussed earlier in this chapter), but the coaxial cable specified is recommended strongly by Granberg.

Heat sinking is of extreme importance in this amplifier. The transistors are joined thermally to a thick block of copper plate, and the latter is coupled to a large aluminum heat sink. Chip capacitors are used throughout the rf circuit of the amplifier. Most of them are on the bottom side of the pc board. The reader is referred to the original *QST* material if duplication of this circuit is anticipated.

Chapter 5

Receiver Design Basics

The most used piece of equipment in any amateur station is the receiver. During communications with other stations the receiver accepts signals from the antenna to produce intelligible audio output. At other times, the receiver is used to "scan the band" and monitor QSOs. The station receiver is also a valuable piece of test gear.

In the early days of amateur radio, it was necessary for every ham to build his own receiver. However, by the time the 1930s arrived, it was common to find an amateur station with homemade transmitting equipment and a commercially built receiver. This was the rule rather than the exception in the early 1950s when the writers first became licensed. The onslaught of single sideband prior to the '60s brought with it the "appliance era," when few amateurs built their transmitters, much less their receivers. The complexity of each was similar, making home construction a task for only the more ambitious and enthusiastic.

The dominance of semiconductor technology has changed this. Today it is straightforward to build receivers of simple design while using transistors and ICs. Even receivers offering something approaching state-of-the-art performance are constructed easily if the builder is willing to invest in a bit of time and experimentation.

In spite of the relative ease of construction, some amateurs are not willing to build a receiver. This is unfortunate, for one of the most exciting experiences available to the ham is the thrill that results from using a receiver he has constructed himself.

In this chapter we will discuss some basic ideas associated with cw and ssb receivers. For the most part, the emphasis will be on straightforward and simple approaches to design. Several practical examples are presented.

In chapter 6 we will consider some refined details of receiver design. The emphasis will be on designing for wide dynamic range. The reader is referred to the transceiver section of the book for additional construction information.

Fundamental Considerations

Certain criteria must be met in the design of a receiver of even the simplest kind. These include meeting specifications for gain, selectivity, sensitivity and stability, to mention only a few.

The first requirement for a receiver is to provide considerable gain. The signal levels from the antenna are often quite low, while enough power output to drive a speaker or a pair of headphones is ultimately desired. If we assume a weak cw signal as being 0.1 μV available from a 50-ohm antenna, the power available to the receiver is

$$P = \frac{V^2}{R} = \frac{(1 \times 10^{-7})^2}{50}$$

$$= 2 \times 10^{-16} \text{ watts} \qquad \text{(Eq. 1)}$$

If we would like this signal to produce an output of 1 volt across a pair of 2000-ohm headphones, the output power is 5×10^{-4} watts, or half a milliwatt. The necessary power gain is then the ratio of these powers

$$G = \frac{5 \times 10^{-4}}{2 \times 10^{-16}} = 2.5 \times 10^{12} \qquad \text{(Eq. 2)}$$

This is 124 dB and is typical of the net gain in many receivers. Since the signals of well under the 1-volt output mentioned above are copied easily in 2-kΩ headphones, less gain is often satisfactory. Around 80 to 90 dB is probably an absolute minimum for communications applications.

A second requirement for a receiver is that it process the incoming signal to cause an audio voltage to appear at the output. The process is called *detection*. Circuits to perform this function will vary considerably, depending upon the nature of the information contained on the incoming signal.

In all of the receivers described in this chapter, product detection is employed. A product detector is really nothing other than a mixer (chapter 3). However, the two signals to be mixed are that of a beat-frequency oscillator (BFO) and a second signal closely spaced. The output of the mixer is at audio frequencies. The term "product detection" results from the characteristic of a mixer that the amplitude of the output signal is proportional to the product of the two incoming signal voltages. In most situations, the BFO level is very much higher than the incoming signals to be detected, often by 100 dB or more. Under these conditions, the detector is essentially a linear device in that the output of the detector is directly proportional to the amplitude of the input. This is not the case for a-m detectors where a threshold exists, or for fm detectors where the output is independent of incoming amplitude once a suitable threshold is overcome. The linearity of a product detector is of profound significance, for it allows us to achieve tremendous simplification in designing simple cw and ssb receivers.

Another characteristic which a receiver must possess is selectivity. That is, it must be capable of isolating two signals which are closely spaced in frequency. This is realized with filters of various kinds, either at radio or audio

frequencies. Both filter types are discussed later in this chapter.

Along with selectivity, a receiver must exhibit stability. The stability required will depend primarily upon the selectivity of the receiver, with the general criterion that the drift in the tuning should be small in comparison with the bandwidth of the receiver. The problems of long-term oscillator stability were outlined in the discussion of VFOs in chapter 3.

Another receiver parameter is sensitivity. This is usually specified by noting the signal power (available at the input to the receiver) required to yield a given output signal-to-noise ratio. The gain calculations outlined earlier might imply that the sensitivity of a receiver can be made arbitrarily low by providing more and more gain. Such is not the case. The culprit, in this case, is noise. Any amplifying device will have some noise generated in it. This noise will add to the signals in the output to cause a degradation in the output signal-to-noise ratio.

A measure of the degradation of signal-to-noise ratio caused by an amplifier or receiver is the *noise figure* or *noise factor*. The formal definition of the noise factor of an amplifier is given as

$$NF = \frac{\frac{S_{in}}{N_{in}}}{\frac{S_{out}}{N_{out}}} \qquad \text{(Eq. 3)}$$

where the input and output signals and noises are powers in watts. If the ratio is calculated as shown above, the term is usually called noise factor. If the power ratio is, however, expressed in dB, the term noise figure applies.

The output signal and noise powers are, in principle, easily measured. Similarly, the input signal power available from a quality signal generator is well defined. However, the input noise power is not as well defined. As a standard, the input noise power is usually assumed to be the power available at the terminals of a resistor at a temperature of 290 degrees Kelvin. The power, P_n, is given by

$$P_n = kTB \qquad \text{(Eq. 4)}$$

where T is the temperature in degrees Kelvin, B is the bandwidth in Hz and k is Boltzman's constant, 1.38×10^{-23} watts/deg.-Hz.

Consider a simple receiver, as an example, to illustrate the noise-figure concept. Assume that the gain of the receiver is 100 dB and that the bandwidth is 500 Hz. If a 50-ohm resistor is attached to the receiver, the noise power available at the antenna terminal is $kTB = 1.38 \times 10^{-23} \times 290 \times 500 = 2 \times 10^{-18}$ watts. If this receiver were perfect, with no internally generated noise, the output noise power would be 10^{10} (100 dB) times this value, or 2×10^{-8} watts. However, the receiver, being a real system, does generate some noise of its own. Hence, the output noise power will be somewhat higher. Assume that the output noise is 1×10^{-7} watts.

If we note the equation for noise factor, we see that it may be rewritten as a ratio of "noise gain" divided by "signal gain."

$$NF = \frac{G_n}{G_s} \qquad \text{(Eq. 5)}$$

Substituting the above noise powers in for the noise gain, that is, the noise output divided by the noise input, we see that $G_n = 5 \times 10^{10}$ while the signal gain was only 1×10^{10}, or 100 dB. Hence, the noise factor is 5: The noise figure is merely 10 log (noise factor), or 7 dB. This value is quite typical for the better communications receivers operating in the 3- to 30-MHz region.

The foregoing arithmetic can be worked backward to tell us what the minimum signal level is that may be detected with this receiver. The noise output of the receiver was 10^{-7} watts and the gain was 100 dB. Hence, a signal at the input which was 100 dB below 10^{-7} watt, or 10^{-17} watts would yield a unity output signal-to-noise ratio. This signal corresponds to about .02 microvolt across a 50-ohm resistor. A signal of about 0.2 microvolt would yield a 20 dB signal-to-noise ratio at the output.

There are a number of factors to be learned from this analysis. First, the lower the noise figure, the more sensitive the receiver will be. Of equal significance, the narrower the bandwidth, the less noise will get through the receiver and the more sensitive it will be. However, the bandwidth of a receiver can be decreased only to the point where it is the same as the bandwidth of the information to be recovered by the receiver. This explains why cw is so much more effective during weak-signal conditions than is any form of phone, including ssb.

There is another factor which does not drop immediately from this analysis. Often, with experienced and capable cw operators, it is found that signals can be copied which are much lower than a measurement of receiver sensitivity would suggest being possible. This is demonstrated easily with a good receiver with switchable bandwidths and a signal generator. The receiver is first set at the narrowest bandwidth available and the signal generator is adjusted to deliver the weakest possible cw signal which the operator can perceive. Then, the bandwidth of the receiver is increased to the widest available setting. More often than not, the operator can still hear the signal. The reason for this apparent discrepancy is that the operator, or listener, is part of the receiving system. His mental process essentially forms a very narrow bandwidth, adaptive (i.e., learning) filter.

This rather subtle effect is not merely a curiosity of nature. It can be used effectively to copy amazingly weak signals from simple receivers. Alternatively, it can be used for the copy of extremely weak signals which might never yield usable output on a meter. The most profound examples of this are the day-to-day moonbounce contacts which are made by means of advanced vhf and uhf amateur stations. The receivers used at such stations have bandwidths of 2 kHz down to perhaps 100 Hz, and exhibit noise figures of 1 to 2 dB.

Rarely on the hf bands is a low noise figure needed in a receiver. The reason for this is that the man-made and atmospheric noise levels found in most locations are so high that they mask any noise generated within the receiver. This factor can be used to advantage by the experimenter. It doesn't matter what the ultimate numerical value for receiver sensitivity is. There is one experiment which is more significant: Disconnect the antenna of the receiver and listen to the noise output of the receiver. Then, connect the antenna and listen to the background noise. If the noise increases dramatically, the sensitivity of the receiver is as good as it needs to be. That's all that counts! (Strictly speaking, the antenna should be replaced with a 50-ohm resistor for comparison, although this is rarely of importance with hf receivers.)

Even though low-noise-figure receivers are rarely needed for the hf bands, the concept is quite important in the design of high-performance receivers. This is especially true if it is desired to design a wide-dynamic-range receiver. An overview of the noise-figure concept has been presented here. Further information is given in chapter 6.

Block Diagrams

There are essentially two forms which the block diagram of an hf receiver can take. They are the classic superheterodyne and the direct-conversion receiver or synchrodyne. Shown in Fig. 1 is a block diagram for the latter, a design which has been popular in this country since 1968. The signals from the antenna are applied to the input of the receiver through a

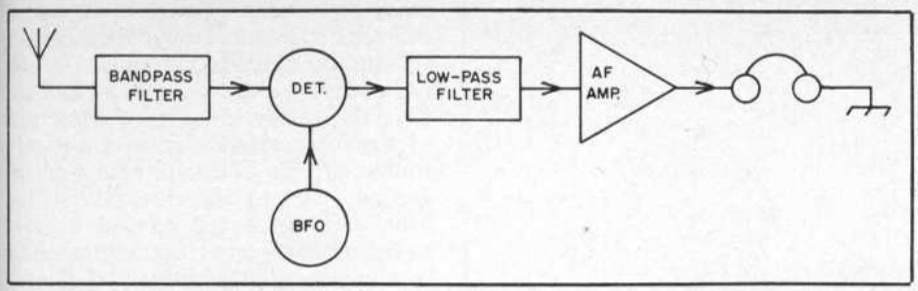


Fig. 1 — Block diagram of a direct-conversion receiver.

simple bandpass filter. The output of this filter is routed to a product detector which is driven by a BFO voltage which is very near the frequency to be received. The output of the detector is applied to a low-pass filter, then routed to a high-gain audio amplifier, thus completing the receiver. The advantage of this approach to receiver design is the extreme simplicity afforded. The number of stages is minimized. Most of the gain is obtained at audio frequencies, where construction is simple. Finally, the BFO operating at virtually the same frequency as that of the received signal leads to the design of simple transceivers.

There are other advantages to the direct-conversion concept which will be described later. However, there is a price to pay for all of this simplicity — the receiver is not a panacea. Consider, as an example, a signal to be received at 7049 kHz. The BFO might be set to 7050 kHz, resulting in a 1-kHz beat note from the detector. This signal is amplified in the audio stages of the receiver and applied to the headphones.

Consider the response to signals at other frequencies. For example, a signal at 7040 kHz would not be attenuated by the front-end bandpass filter. Hence, it would also be applied to the input to the product detector and would result in an output beat note of 10 kHz (the BFO is still at 7050). The low-pass filter will prevent most of the 10-kHz energy from arriving at the audio amplifier, so this signal causes no significant problem. Consider now, a signal at 7051 kHz. This signal will reach the input of the detector and heterodyne with the BFO output at 7050 kHz to produce a 1-kHz beat note, which is exactly the same response as obtained from the desired signal at 7049 kHz. Hence, no amount of audio filtering will eliminate this response. This undesired response is called an *audio image*, and it is a major disadvantage with direct-conversion designs. In spite of this, thousands of amateurs have built "dc" receivers and use them daily. The simplicity of design is worth the few practical problems which arise from the audio image during routine communications. Although the existence of the image would have the effect of doubling the equivalent noise bandwidth of the receiver, this effect is largely negated by the filtering nature of the human ear. There is virtually no fundamental sensitivity penalty to be paid for the use of direct-conversion receivers.

Shown in Fig. 2 is a block diagram for a classic superhet receiver. Here, the incoming signal is applied to a preselector bandpass filter and is then routed to a mixer. The mixer is also driven by a local oscillator which is separated from the incoming frequency. The output of the mixer is at a frequency which is the difference (or the sum) of the incoming signal and the local oscillator (LO). This frequency is called the *intermediate frequency*, or i-f. The i-f output from the mixer is applied to a filter which usually has a bandwidth compatible with the signals being received. The i-f signal is amplified further before it is applied to a product detector. The detector output is amplified and then applied to headphones or a speaker where the user should perceive some intelligent information.

Consider a receiver with an i-f of 1 MHz. Assume that the i-f filter has a bandwidth of 500 Hz and suppose that this receiver is tuned to the same signal at 7049 kHz that was used in the "dc" receiver example. For the 7049-kHz signal to be received, the LO will be tuned to 6049 kHz, resulting in a 1000-kHz output i-f signal. This signal moves readily through the 500-Hz-wide filter, is amplified and detected. If the detector is driven by a BFO at 999 kHz, a 1-kHz receiver output will result.

Now consider that same bothersome signal at 7051 kHz. This signal will beat in the mixer with the local-oscillator energy at 6049 kHz to produce an i-f output at 1002 kHz. However, the i-f filter is only 500 Hz wide. Hence, the filter will have significant attenuation at 1002 kHz, and no receiver output will result. The superhet has eliminated the troublesome audio image which plagued the dc receiver. This asset of a superhet is called *single-signal response*.

Image responses will still be present, but now they are associated with the intermediate frequency rather than with audio. For example, our receiver has a 1-MHz i-f and an LO at 6 MHz, for a desired input near 7 MHz. However, signals at 5 MHz will also beat with the LO to produce 1-MHz i-f signals. Hence, everything possible should be done to prevent 5-MHz signals from reaching the mixer input. This is easily realized with the 7-MHz preselector filter.

The following sections will consider design details of the various sections of direct conversion and superheterodyne receivers. Examples are presented for duplication. Emphasis will be on simple designs.

Product Detectors

The product detector is the basis of the direct-conversion receiver, and it is an integral part of a "superhet" receiver designed for cw or ssb reception. As mentioned earlier, a product detector is essentially a mixer. As such, it is a three-port circuit with two radio-frequency inputs and an intermediate-frequency output. When a mixer is used as a product detector, the i-f is at *audio*. A product detector is shown in block-diagram form in Fig. 3.

When used as the front end of a direct-conversion receiver, a product detector has a number of necessary specifications. First, it must have a

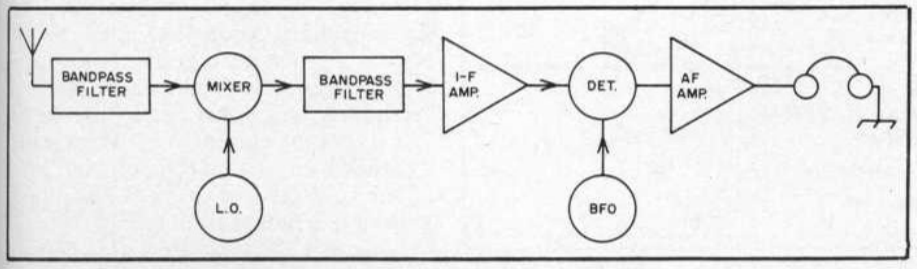


Fig. 2 — Configuration for a basic superheterodyne receiver.

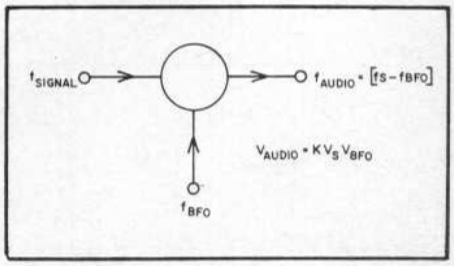


Fig. 3 — Representation of a product detector.

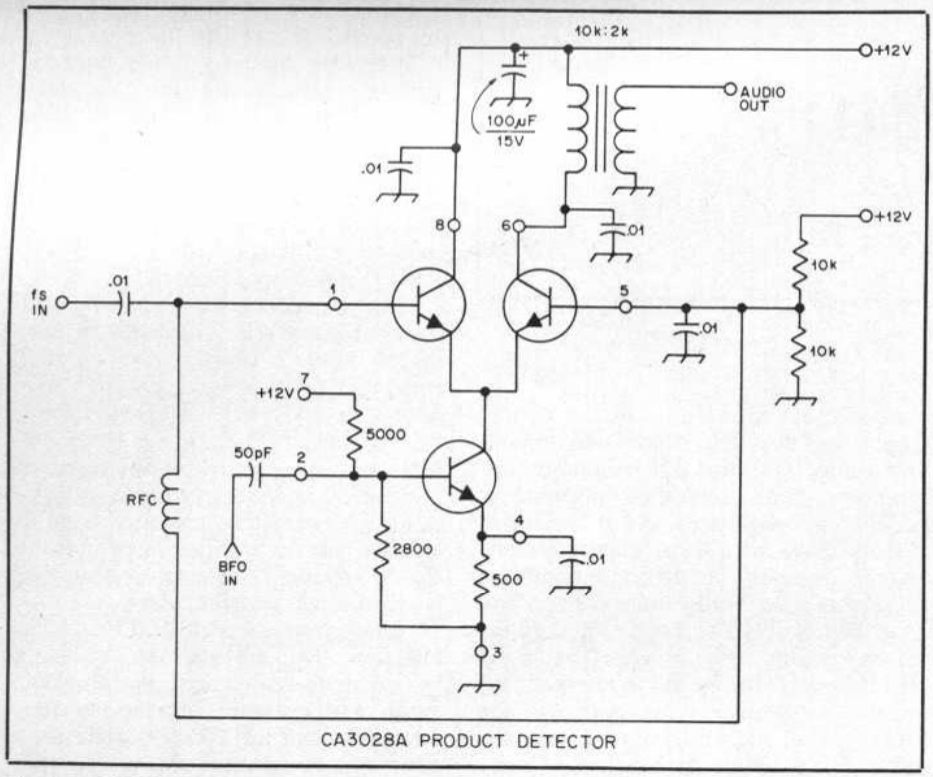


Fig. 4 – CA3028A product detector.

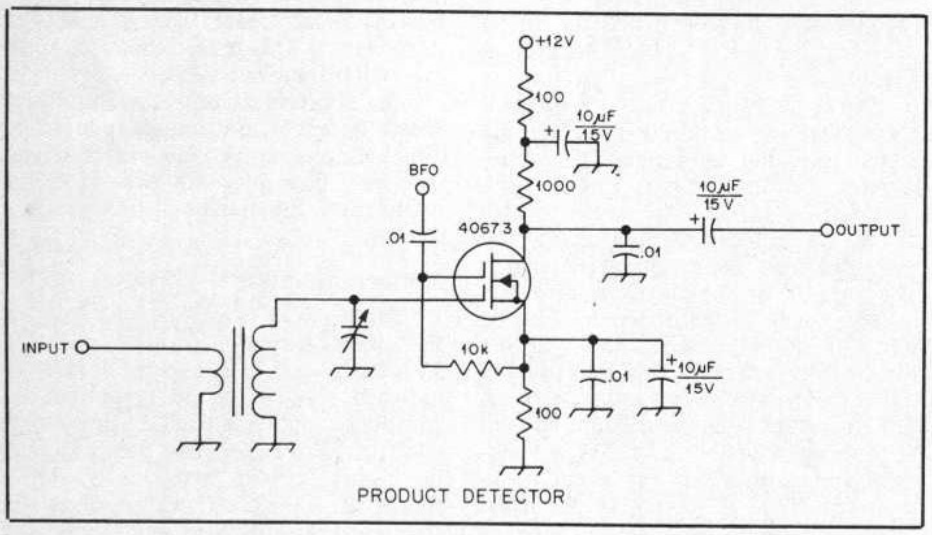


Fig. 5 – Example of a dual-gate MOSFET product detector.

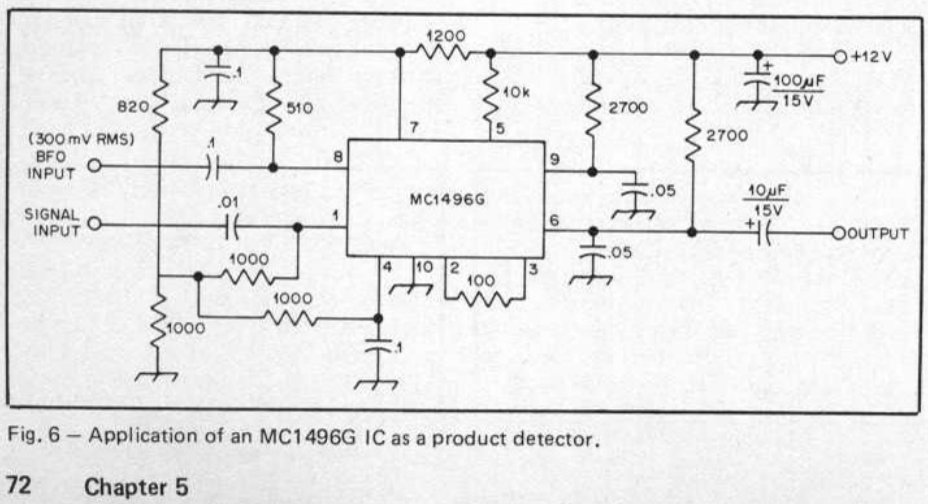


Fig. 6 – Application of an MC1496G IC as a product detector.

fairly low noise figure (low noise at rf *and* audio frequencies). Some gain is sometimes desired, although certainly not necessary. The detector should also have the ability to handle a wide range of signal-input levels without the undesirable effects of intermodulation distortion, blocking and cross modulation. Finally, there should be essentially no audio output except that which results from mixing with the BFO.

When used as a detector in a superhet, the circuit requirements are somewhat relaxed. Noise figure is no longer of major concern, since the detector is usually preceded by circuits with considerable gain. Often the dynamic-range requirements can be relaxed since the detector is protected by an automatic gain-control (agc) system. However, intermodulation distortion is still of concern, since two signals within the passband of the i-f amplifier can produce spurious outputs.

There are a number of circuits which offer satisfactory performance as product detectors. It is difficult to say categorically which of these is best, for all have assets as well as problems. A variety of circuits is presented for the experimenter to consider.

Shown in Fig. 4 is a detector popularized in 1969. It uses an RCA CA3028A differential amplifier IC. Other similar "pills" could be used. These include differential amplifiers such as the Motorola MFC8030, and transistor arrays such as the RCA CA3046. The CA3028 detector is perhaps one of the easiest circuits to use, since it has a reasonable noise figure and considerable gain. For example, direct-conversion receivers have been described using such a detector, followed by a single transistor or IC as the total audio amplifier. If maximum gain is to be realized with this circuit, the output should be terminated in a fairly high impedance. This is usually realized with an audio transformer with a 10-kΩ primary.

Several volts of BFO injection are often used with this circuit, resulting in a switching type of current waveform at the collector of the common current-source transistor of the IC. To optimize performance, it is advisable to bypass the emitter (pin 4) of this transistor.

If large-signal problems are encountered with this detector, such as blocking or cross modulation, the signal-handling properties may be improved by decreasing the output collector termination impedance and by "standing" additional current in the IC. The quiescent current may be increased by adding a 330-ohm resistor from pin 4 of the CA3028A to ground. The output termination impedance can be lowered by changing the transformer ratio, or by using low-value collector resistors in

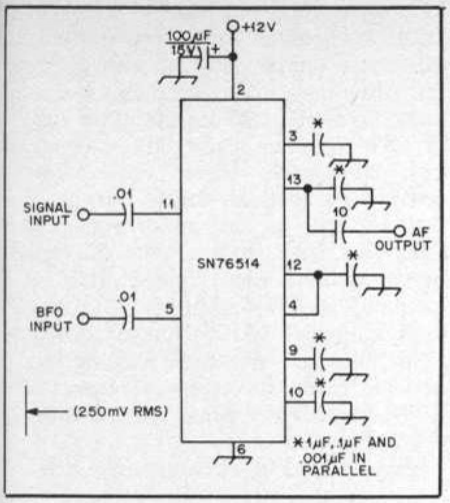


Fig. 7—A product detector can be built from an SN76514 IC. The SM-76514 mixer IC has been reidentified as TL-442-CM by Texas Instruments. It may be procured under either part number.

parallel with the transformer primary. The decreased collector load will, however, decrease the detector gain.

The CA3028A, as shown, is a singly balanced product detector. The input signals are applied differentially, while the BFO drive is applied in a single-ended fashion. This tends to minimize the BFO energy present at the antenna terminals of a direct-conversion receiver. In one case where measurements were performed, the power at the antenna terminal was –47 dBm (.02 microwatt into a 50-ohm load).

Another popular and easily applied product detector for use in direct-conversion receivers is a dual-gate MOSFET (Fig. 5). The circuit is essentially the same as the mixer circuits used with this device, except that the output is designed for audio frequencies, with rf signals being bypassed. With detectors of this kind, the BFO injection at gate 2 should be approximately 5 volts pk-pk. Additional gain can be realized by increasing the output load impedance. This, however, requires the use of transformer coupling or a supply voltage well above 12.

The dual-gate MOSFET has good immunity to blocking, IMD and cross modulation. However, the audio-frequency noise figure of MOSFETs is often not as good as those expected from diodes or JFETs, yielding a degraded receiver sensitivity in direct-conversion applications. The major deficiency of this detector is its behavior with a-m signals. The MOSFET is substantially a square-law device and will operate as a square-law detector of a-m signals. This causes severe problems in Europe and on the East Coast of the USA on 7 MHz, where large signals from international broadcast stations are present. Proper use of balance in the detector should minimize this problem.

Several ICs other than simple differential amplifiers function well as product detectors. Notable examples are the Motorola MC1496G and the Texas Instruments SN-76514. The reader is referred to chapter 3, where these devices were applied as transmitting mixers. The MC1496 is used as a product detector by merely replacing the rf collector load with a pair of 2.7-kΩ resistors to pins 6 and 9. A circuit is shown in Fig. 6. Audio is extracted from one of the output terminals through a 10-μF capacitor. Each of the output pins should be bypassed for rf via a .05-μF capacitor. Additional conversion gain can be had by using a center tapped transformer at the output.

The TI SN-76514 has built-in 600-ohm collector resistors. Hence, this chip is used as a detector by bypassing the two output pins (3 and 13) for rf, and by taking audio from one of the pins through an electrolytic capacitor (see Fig. 7). The relatively low collector load resistors in the TI balanced-modulator IC will limit the conversion gain to roughly 14 dB, while much more gain can be realized from the MC1496.

If the internal circuit of the MC1496 is studied, it can be seen that the input signal is injected differentially to a pair of transistors with externally applied emitter degeneration. The level of this negative feedback is controlled by the value of resistance between pins 2 and 3. In the interest of signal-handling capability, this resistor should be as high in ohmic value as reasonable, perhaps as much as 1000 ohms. On the other hand, the resistance should be zero if maximum conversion gain and optimum noise figure are desired. Hence, the value will probably be much different for applications in superhets than it would be for use as the input to a direct-conversion receiver.

The Motorola applications literature of the '1496 shows the chip biased so that about 1 mA flows in each of the collector output pairs. However, the signal-handling properties of the chip can be improved significantly by increasing the current to approximately 3 mA in each collector. This is effected by changing the usual 10-kΩ resistor between the 12-volt supply and pin 5 to a 3.3-kΩ unit. This biasing scheme is useful also when the chip is employed as the mixer in a ssb transmitter, where linearity is of major importance.

Another IC which functions well as a product detector is the RCA CA3102E. This IC is a dual differential amplifier and is wired externally much like the MC1496 detectors discussed above. A circuit is shown in Fig. 8. Good noise figure (as well as fine signal-handling ability) was observed with this circuit. These traits probably result from a lack of feedback in the signal input, and the switching nature of the circuit, respectively. The detector circuit shown is a doubly balanced format, requiring push-pull inputs at the signal and BFO ports. A single-ended BFO is converted

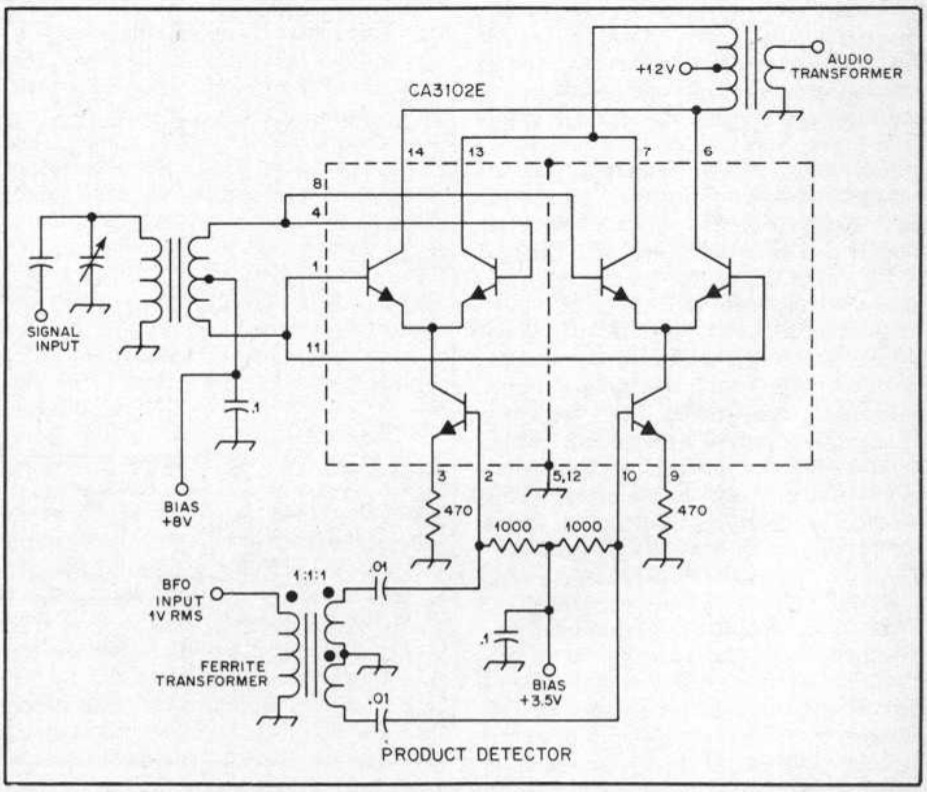


Fig. 8 — A CA3102E can be used as a doubly balanced detector as shown here.

to a balanced drive with a ferrite transformer much like those used for balanced frequency multipliers. The signal input is by means of a bifilar link around an input tuned circuit.

Diode Detectors

There is a class of product detectors which have not been described in this book. All use diodes as the nonlinear element. The experimenter may view diode detectors as being useful only in special cases where simplicity or a minimum parts count are special criteria, thinking, "Such detectors are obviously inferior to those using FETs or ICs." Nothing could be farther from the truth! Detectors (and mixers) using diodes are among the best available if they are constructed and used properly, with good transformers and adequate BFO (or LO) injection.

Shown in Fig. 9 are three detectors which use diodes. These circuits contain broadband transformers made with trifilar windings on a ferrite toroid core. (The reader is referenced to chapter 4 for details on the construction of this type of transformer.)

The simplest of these detectors is that of Fig. 9A. This is a singly balanced circuit with the BFO applied at point C. Note that a signal at C drives the two secondary windings of the transformer in opposite directions. Hence, no magnetic field is established in the core. As point C swings positive, the upper diode is driven into conduction, placing a charge on the 0.1-μF capacitor. But, on negative swings of the BFO, the lower diode conducts, and a similar charge is removed from the capacitor. The overall result is that the average voltage across the capacitor is zero. However, when a signal appears across the transformer, one diode goes into conduction slightly sooner (or later) than it would have otherwise, causing an unbalance in the net current flowing into the capacitor. Over a period of time, this net transfer of charge is observed as an audio voltage at the output. The diodes are assumed to be virtually identical.

In the detector at Fig. 9B, two diodes have been added. These diodes have the effect of presenting a more symmetrical load to the BFO, resulting in slightly improved balance and better isolation of the BFO from the signal circuit. The circuit is still singly balanced. The circuit shown in Fig. 9C is doubly balanced, resulting in good isolation between all three ports of the mixer.

Detector balance is of minimal significance when the detector is at the front of a direct-conversion receiver. However, balance can be of considerable consequence when used at the detector in an advanced superhet. In such a design, it is mandatory that the energy from the BFO be confined to the detector, and not be allowed to find its way into earlier parts of the circuit. If extraneous BFO energy gets into preceding i-f amplifiers, noise modulation may occur, which has the effect of creating a "mushy" sounding output from the receiver. Having no i-f stage preceding the detector in a direct-conversion receiver will lead to an exceptional signal "cleanliness" and "presence" that is characteristic of such a design.

Detectors using diodes have no gain. Indeed, they exhibit a loss. Measurements with mixers of the type shown in Fig. 9A (using two diodes) frequently show a low of 5 to 6 dB. The circuits using four diodes typically have a 6- or 7-dB conversion loss. In the high-frequency region, and usually throughout vhf, the diodes contribute essentially no noise, making the noise figure of such a mixer merely its conversion loss. The noise figure of a direct-conversion receiver using this as the detector will be the mixer conversion loss, plus filter losses, plus the noise figure of the audio amplifier. It is easy to build audio amplifiers with noise figures under 3 dB. Hence, receivers using direct conversion can be constructed easily to display a respectable 13-dB noise figure when using diodes as the detector.

Shown in Fig. 10 is a simple direct-

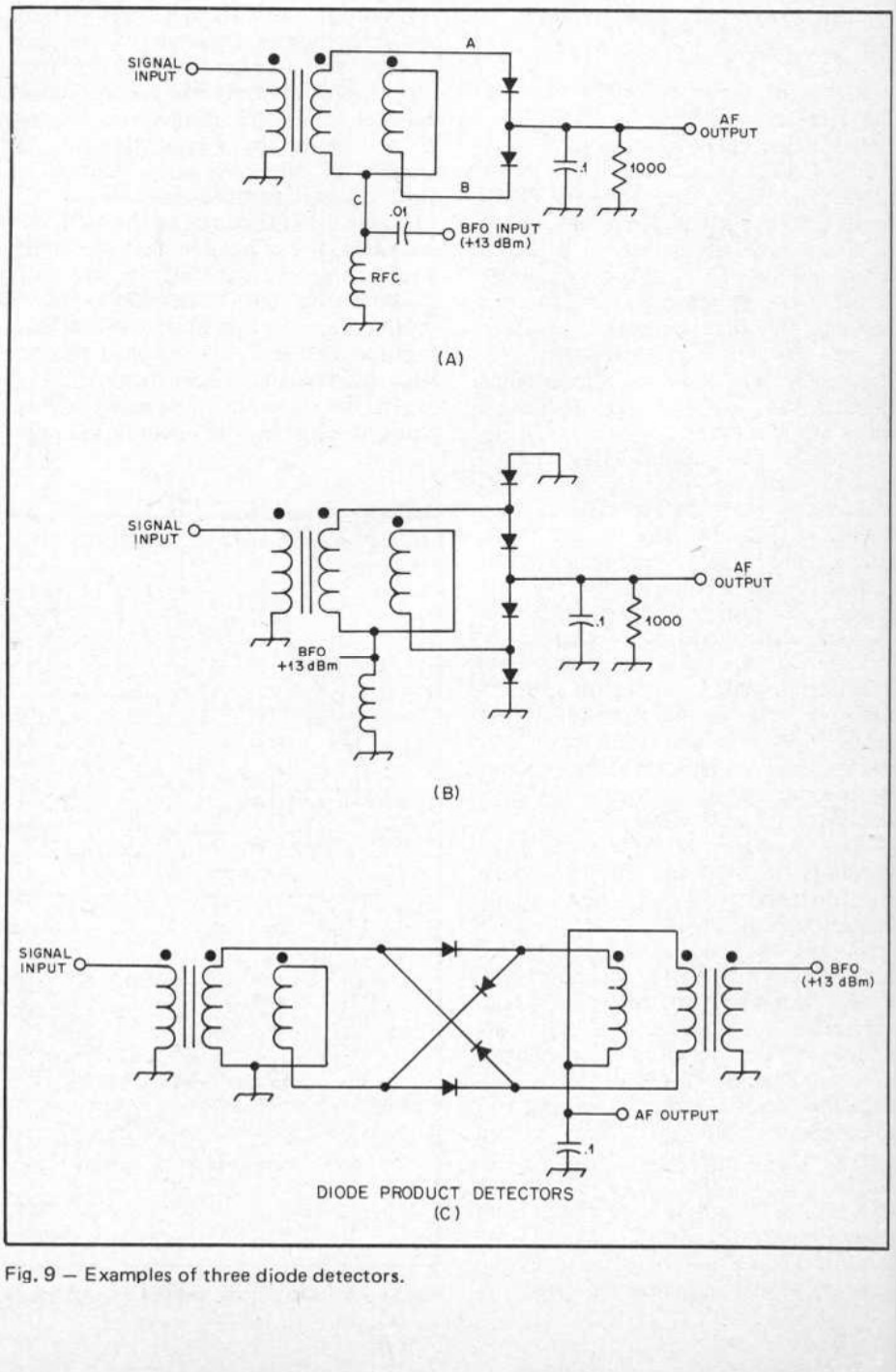


Fig. 9 — Examples of three diode detectors.

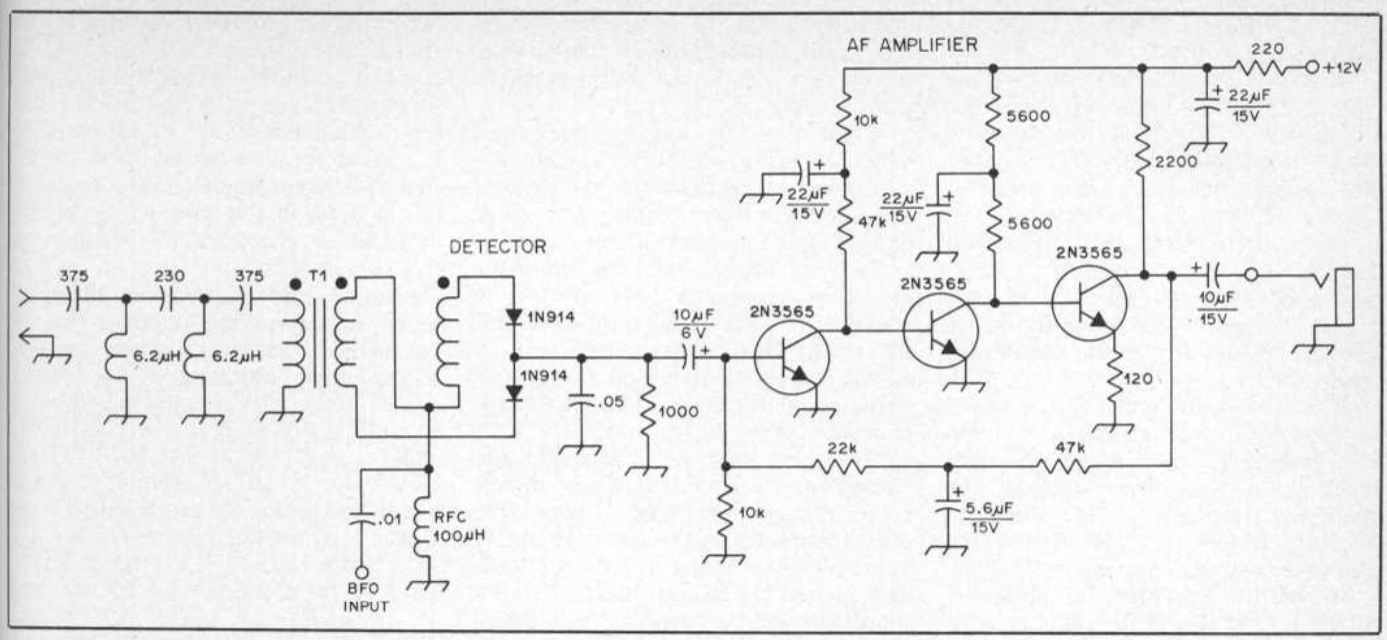


Fig. 10 — A basic direct-conversion receiver using a 5-pole high-pass network at the input port.

conversion receiver which was assembled in order to perform some detector measurements. The input filter is a 5-pole high-pass type with a cutoff at 3 MHz. This filter was inserted in order to eliminate a trace of broadcast-band rectification which was present. However, this was the only selectivity element which was used in the receiver. The detector was the simple two-diode type discussed above. It was followed by a high-gain audio amplifier, using three inexpensive (but "hot") transistors. The diodes were silicon switching types (1N914 or equivalent) which were matched for forward resistance with a VOM. A BFO energy of +13 dBm (20 milliwatts) was supplied from a homemade general-purpose signal generator.

The first experiment performed was to evaluate the sensitivity. Since a minimum of audio filtering was included in the system, a careful sensitivity measurement would not have been very enlightening because of the wide bandwidth of the system. However, a signal of 0.1 µV was easily detected at 7 MHz, and a 1-µV signal was plainly audible. An audio output of 1-volt rms was measured for an input of 6 µV, indicating a net receiver gain of 88.4 dB.

The next measurement was to evaluate receiver blocking. This was done with two signal generators and a hybrid combiner. The desired signal was from a low-level crystal-controlled generator which was well shielded. It was set for an output of 1 µV, and the BFO was adjusted for copy of this signal. The other generator was a URM-25D — another well-shielded instrument. This was set initially at 50 kHz away from the desired signal, and the level was increased until blocking occurred. However, the measurements were misleading, for there is essentially no selectivity following the detector except for a capacitor which provides low-pass filtering. The audio amplifier was overloading, so the second generator was set to 8 MHz, and the experiment was repeated. Note that the input was broadband in nature. That is, there was no selectivity ahead of the detector. Nonetheless, the detector was able to provide solid copy of the 1-µV desired signal, with no desensitization from an undesired input signal of 0.1 volt. There are many well respected commercial receivers which cannot pass this test!

In spite of the good response to the weak and strong signals described, diode detectors have deficiencies which make them difficult to use: Diode mixers, in general, should be terminated carefully if optimum signal-handling ability is to be retained. Specifically, the "signal" port of the mixer should look back at a source impedance of around 50 ohms. Further treatment of termination is presented during the mixer discussion in chapter 6.

Another characteristic which can present a problem, but can be an asset, is a tendency toward harmonic mixing. Even if the BFO energy supplied to a mixer is free of harmonics, the nonlinear nature of the diodes will create large harmonic currents. The result is that input signals at other frequencies will also cause major outputs. Diode balanced mixers are known for their high response to odd-order harmonics. The receiver of Fig. 10 was used to evalute the harmonic mixing traits of a simple two-diode product detector. The BFO was set at 7 MHz, and the signal generator was adjusted to various harmonic frequencies, with the audio output always adjusted for 1-volt rms. The results are presented in Table 1. The dominance of odd-order responses is clear from the data.

Table 1

N	F_{in}	V_{in}	*Ratio*
1	7 MHz	6 µV	0 dB
2	14	700	41.3
3	21	20	10.5
5	35	70	21.3
7	49	100	24.4

The harmonic-mixing phenomenon could be used to advantage. For example, it might be possible to construct a receiver which used both the N = 1 and N = 3 responses to cover the 7- and 21-MHz bands. More often, however, harmonic mixing is a problem. This is especially true if the user lives close to commercial TV and fm stations. As the receiver is tuned, "birdies" may appear across the band.

The answer to the harmonic-mixing problem is, of course, preselection. A good low-pass filter ahead of the receiver will attenuate harmonic inputs to the point that all spurious responses are eliminated. This can be more difficult to do than might be suspected, for it is required that the filters ahead of the receiver have the desired attenuation not only near the cutoff frequency, but in the vhf stop-band. This means that vhf layout and shielding methods should be used even in a 40-meter filter! Filtering of the BFO will do little, for

the mixing harmonics are created in the detector.

A partial solution is to replace the silicon switching diodes with hot-carrier diodes. These units differ from the usual PN semiconductor diodes. They consist of a junction between a semiconductor (usually N type) and a metal. These diodes switch fast and work well through the microwave frequencies. Furthermore, they lack the charge-storage effects which partially cause junction diodes to create high-order harmonics.

While harmonic mixing is a major problem with diode product detectors, it is present to some extent in other detectors as well. For example, the square-law response of the dual-gate MOSFET makes this device prone to even-order harmonic mixing.

A useful attribute of harmonic mixing is that it aids the calibration of direct-conversion receivers. For example, if a 100-kHz oscillator is used to calibrate a 7-MHz direct-conversion receiver, it is often possible to hear the 2nd and 4th harmonics. They can be used as markers (for free) at 50- and 25-kHz intervals.

The need for preselection filtering is significant for the reasons outlined above and in the preceding section (e.g., image rejection in superhets). Harmonic responses can be suppressed with the half-wave low-pass filters described in chapter 4. A number of narrow-band, multi section band-pass receiver filters are presented in the appendix.

Audio Amplifiers for Direct-Conversion Receivers

Direct-conversion receivers differ in a number of ways from the "superhet." Most significant is where the incoming signal is detected immediately with no intermediate heterodyning processes. Another difference is the gain distribution. The typical superhet will have most of the gain concentrated in the i-f section, with only 30 to 60 dB being achieved at audio frequencies. On the other hand, the direct-conversion receiver has nearly all of the gain concentrated in the audio section. Indeed, when a diode type of product detector is used without an rf amplifier (as described in the previous section), the only gain in the receiver is in the audio stages.

The high gain requirement of the audio section of a direct-conversion receiver places more stringent requirements on the amplifier design than would be the case with a superhet. Not only must the gain be high, there should be no instability in the amplifier. While oscillations rarely occur in the low-gain amplifiers used in superhets, they can take place when the amplifier has up to 100 dB or more of gain. Finally, the noise figure of the audio amplifier is significant, especially when low-gain detectors are employed (such as those using diodes).

Shown in Fig. 11 is a three-stage amplifier using 2N3565s. These transistors are inexpensive, have high beta and low noise figure. Using an amplifier design, we will present a fairly detailed analysis of this circuit. The transistor model is simple. A beta of 100 is assumed for each of the transistors, and the emitter-base offset is 0.7 volt. A 10-volt supply is used, and the output termination is a set of 2000-ohm headphones.

The first step is to evaluate the biasing of the amplifier. Three direct-coupled stages are used. Hence, the overall amplifier will be inverting, thereby allowing us to use negative dc feedback to bias the circuit. Since all of the transistors will be operating in an active condition, the voltage on the base of Q1 will be 0.7 volt. This voltage can originate only from the bias resistors from the collector of Q3. Noting the values used, we see that 0.7 volt occurs at the base of Q1 only when the dc potential at the collector of Q3 is 6 volts.

Knowing the dc output voltage, we can evaluate all of the dc voltages in the amplifier. The collector current for Q3 must be 2 mA [(10 − 6V)2 kΩ], leading to $V_e3 = 0.2$ volt and $V_b3 = 0.9$ volt. Continuation of this analysis gives us the voltages and collector currents for Q1 and Q2. These are shown in squares in the figure.

The next step is to evaluate the input resistances for each stage. For Q1 and Q2, the input resistance of each is given by $R_{in} = 25\beta \div I_e(\text{mA})$ leading to input resistances of 5 kΩ and 3 kΩ for Q1 and Q2, respectively. The input resistance of the overall amplifier will be the input resistance of Q1 shunted by the bias resistors, leading to an overall input resistance of roughly 3 kΩ. The input resistance of Q3 is not given by the same formula as was used for the first two stages, since emitter degeneration is used. In this case, $R_{in} = \beta R_e$ is a suitable approximation, leading to $R_{in-3} = 10$ k Ω.

Having this information, the small signal ac gain of the amplifier may be calculated. These calculations are presented next, assuming a 1-μV input signal: $V_{in} = 1\ \mu V, I_{b1} = V_{in} \div R_{in1} = 2 \times 10^{-10}$ A, $I_{c1} = \beta I_{b1} = 2 \times 10^{-8}$ A, and $V_{c1} = I_{c1}R_{L1} = 2 \times 10^{-8} \times 2.3 \times 10^{3} = 4.6 \times 10^{-5}$ V. (Note: The collector load is R_{in-2} paralleled with the 10-kΩ load resistor.) Next, $I_{b2} = V_{c1} \div R_{in2} = 4.6 \times 10^{-5} \div 3 \times 10^{3} = 1.53 \times 10^{-8}$ A, $I_{c2} = \beta I_{b2} = 1.53 \times 10^{-6}$ A, $V_{c2} = I_{c2}R_{L2} = 7.67 \times 10^{-3}$ V, and $V_{c3} = G_{V3}V_{c2} = 7.67 \times 10^{-2}$ V. Note that the emitter degeneration in the last stage leads to a voltage gain of 10 in that stage.

The overall voltage gain of the amplifier is 7.67×10^4. Taking 20 Log G_v, we arrive at 97.7 dB, a value quite close to that measured. These methods may be used to evaluate any of the simpler audio amplifiers which are used in similar applications.

There are a few subtleties to the design of the amplifier of Fig. 11. First is the 100-ohm emitter-degeneration resistor in the last stage. This serves a number of functions. First, it decreases the gain to a level which is compatible with the desired overall gain. Additionally, since the output signals from Q3 may be large, it adds linearity to this stage in order to minimize distortion. Finally, it increases the bandwidth of

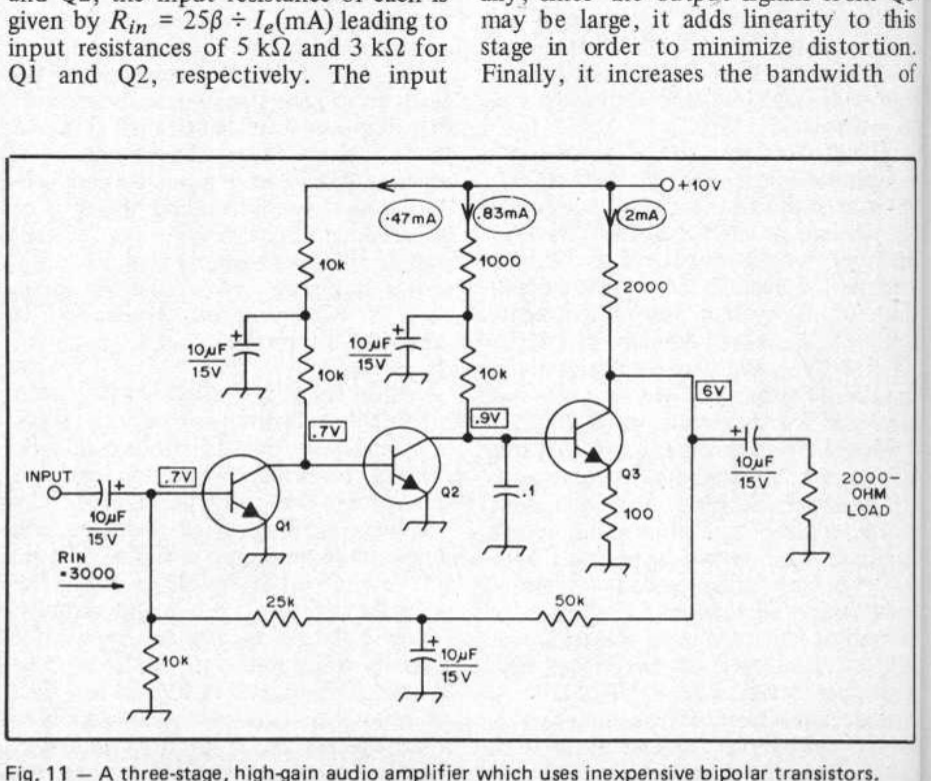


Fig. 11 — A three-stage, high-gain audio amplifier which uses inexpensive bipolar transistors.

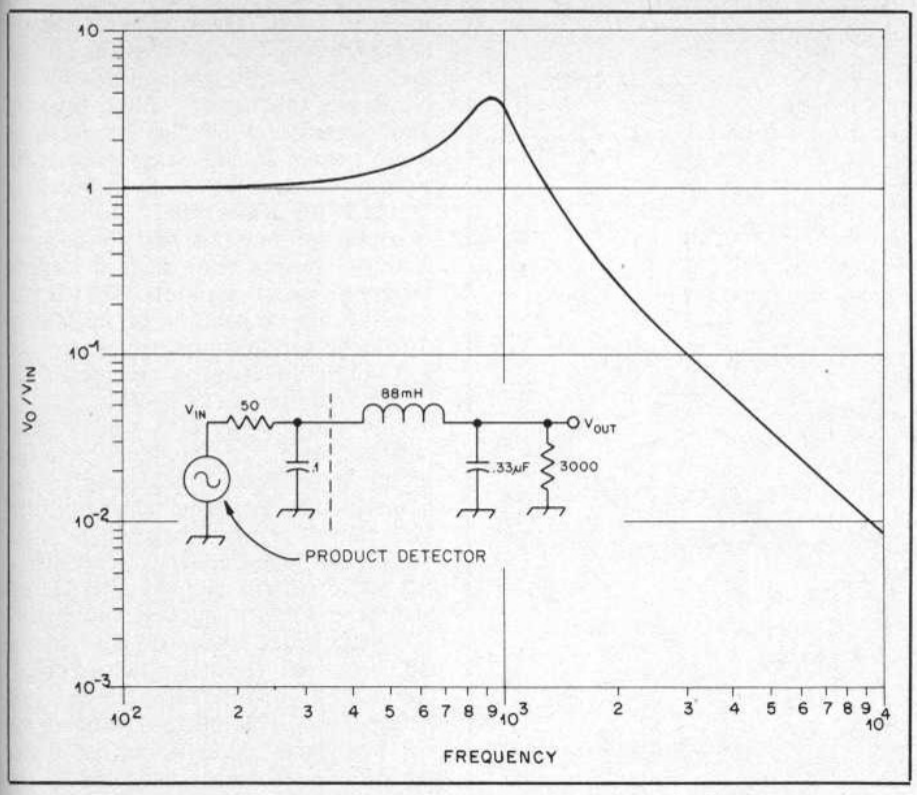


Fig. 12 — Circuit and frequency-response characteristics of a passive audio filter.

the overall amplifier. This is of significance in stabilizing the operation of the gain block, because of the dc feedback method of biasing.

A 0.1-μF capacitor is shown from the base of Q3 to ground. This capacitor will have an impedance of about 1.6 kΩ at 1 kHz, leading to a low-pass characteristic for the overall response. Note that this impedance is much less than the collector load of 5 kΩ on Q2.

The input impedance of the overall amplifier is about 3 kΩ. Hence, if the input were driven directly from the low-impedance output of a diode type of product detector (typically around 50 ohms) very little of the output energy would be transferred. To realize the full gain of the amplifier, an impedance transformation is required at the input. This could be a simple audio transformer with a turns ratio of, say, 1:5. Transformers at the input to a high-gain block of this kind are often difficult to use owing to their tendency to pick up 60-Hz energy. Shown in Fig. 12 is an alternative solution. Here, an 88-mH toroid is used as the inductor in an L network. A pot core could be used if a toroid was not available. This network has a peak response at 940 Hz, where the impedance transformation is well over 10. As an additional bonus, the L network serves as a low-pass filter, offering protection to the audio amplifier from out-of-passband signals. The figure also shows a computer-calculated response of this filter when the input is driven from a 50-ohm source, and the output is terminated in 3000 ohms. Note that over 40 dB of attenuation is present at 10 kHz.

Another convenient means for achieving high gain at audio frequencies is through the use of IC op amps. Most of the commercially available op-amp ICs have extremely high open-loop gain at dc, and are applied easily in audio circuits. Considerable care must be used if optimum results are to be obtained.

Shown in Fig. 13 is an audio amplifier using a bipolar transistor and a 741 op-amp. The advantage of this circuit is that it is decoupled easily from the supply while still providing high gain: In this case about 78 dB (assuming the output is terminated in a resistance equaling the input resistance of the transistor). The gain of the op amp is determined by the feedback resistors, in this case the 47-kΩ and 1-kΩ units. It would be possible to increase the gain considerably by shorting the 1-kΩ resistor, thus biasing the op amp to operate at its open-loop gain value. However, the noise would probably be intolerable. If op amps are used in high-gain applications, it would be wise to use low-noise types. The LM-301A is preferred over the 741, and the LM-308N is probably one of the best low-noise units available.

While op amps have appeared frequently as audio amplifiers in the ham literature, they have often been misused. The advantage of using an op amp over other kinds of circuitry is that the performance of the ultimate circuit is controllable through the use of feedback. Generally, an op amp should not be used in an open-loop manner. Furthermore, potentiometers should never be necessary to bias an op amp in an audio application!

The two amplifiers described in the foregoing text are suitable as the major gain blocks in many direct-conversion receivers. There are other ways to obtain the needed gain, leaving plenty of room for experimentation. The amplifiers described are merely examples.

If a loud speaker is driven instead of 2000-ohm headphones, other circuits must be used, ones which are capable of driving lower impedances.

Practical Audio Amplifiers

Integrated circuits have come to the fore in recent years, filling a need for compact low-power audio amplifiers of the transformerless variety. For most amateur applications a chip in the 0.25- to 2-watt class is suitable. The majority

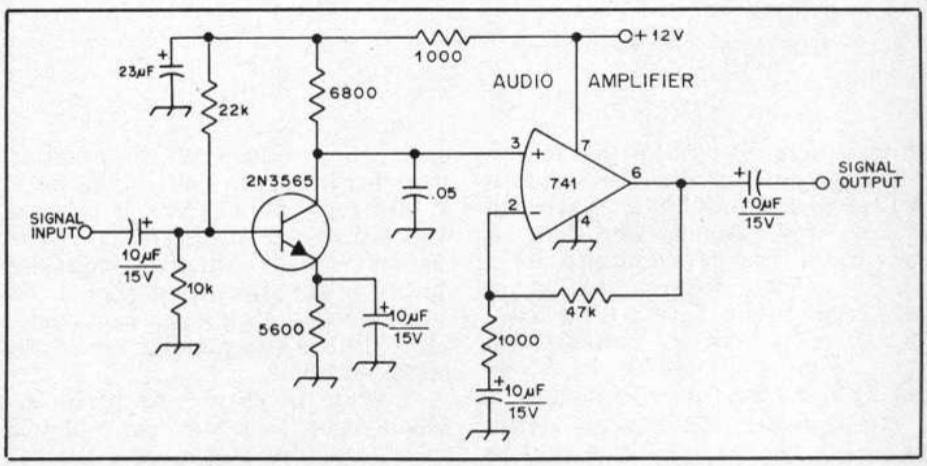


Fig. 13 — An audio amplifier capable of 78 dB of gain. It combines a bipolar transistor and an op amp.

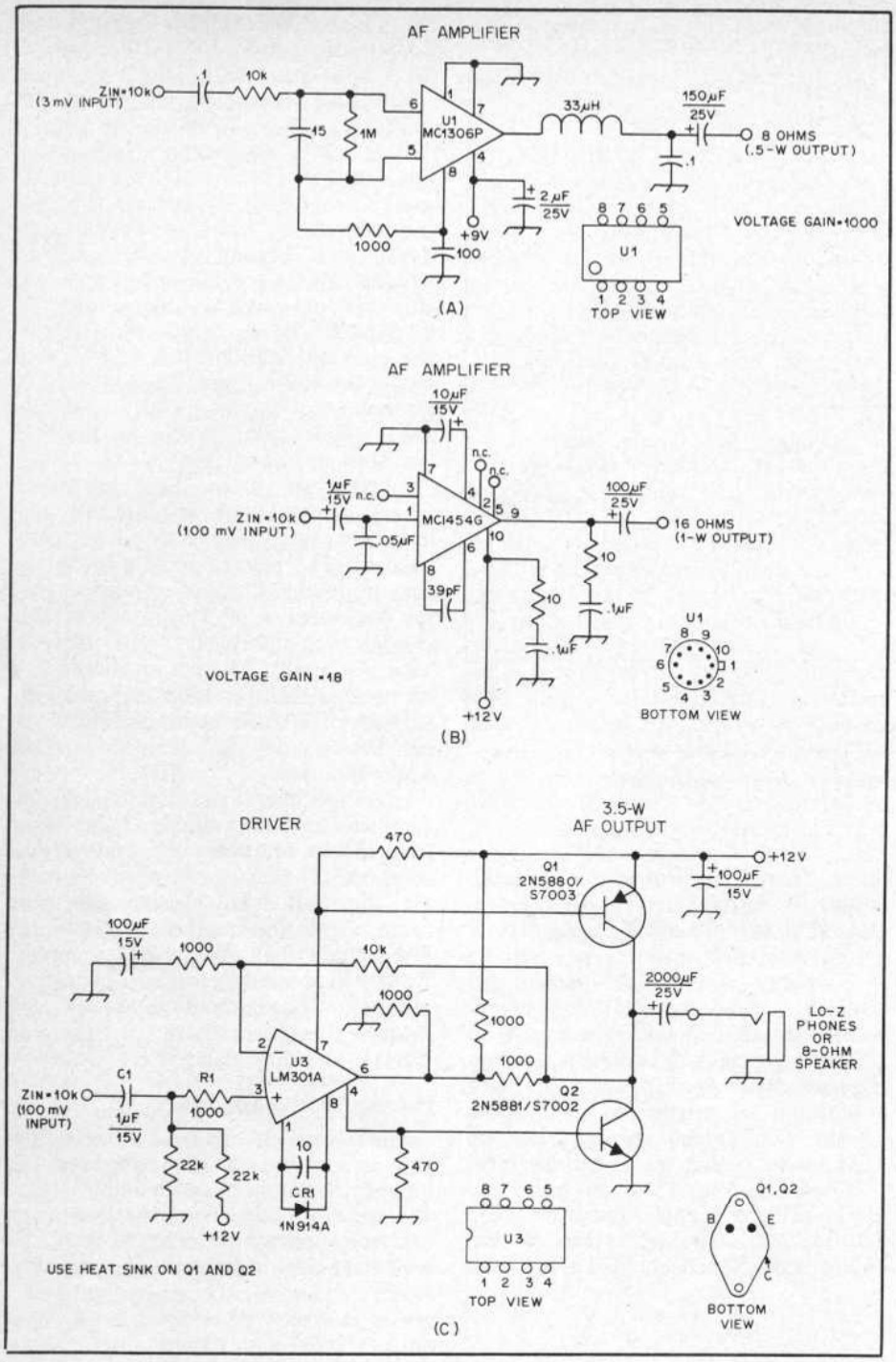


Fig. 14 — Examples of audio amplifiers.

of these ICs are designed to operate into a nominal load impedance of 8 ohms at the rated harmonic distortion characteristic set by the manufacturer. However, headphones can be substituted for a speaker in most instances, regardless of the headset impedance (4 to 2000 ohms), and satisfactory operation will result without damage to the IC resulting from a mismatched condition.

One problem exists when certain audio ICs are used: Biasing is done internally, thereby preventing the builder from improving the cross-over distortion characteristics. Distortion of that kind is not especially troublesome at high audio-output levels, but during weak-signal reception, and at moderately low audio-output amounts, the distortion will affect the quality of the received signal. A cw note, for example, will exhibit a fuzzy sound which can impair readability.

The use of discrete devices in an audio-output stage (at power levels above, say, 100 mW) permits the designer to tailor the circuit for minimum cross-over distortion. It would be wasteful in a serious design effort to have a high-performance rf/i-f receiver section, then degrade the signal quality by employing a substandard audio channel. The linearity of all the stages in an audio system should be as good as the art will permit. At least, the designer should strive to meet that criterion.

Attention must be paid to the audio voltage levels entering each af stage at maximum signal amounts. That is, the amplification capability of each stage should be set so that a successive stage will not be driven into a nonlinear state. Gain distribution is as significant as it is in the early stages of a well-designed receiver. Also, the frequency response of the stages should be shaped for the desired audio passband characteristics. This subject is treated elsewhere in the book. The high-frequency response of the audio system should roll off at the highest desired frequency – typically 1000 Hz for cw work, and 2500 Hz for ssb reception. The net effect is one of minimizing high-frequency noise and heterodynes. This aids in reducing the QRM problem and enhances the overall signal-to-noise characteristics of the receiver. Some cw operators prefer an even lower roll-off point for the audio system – 600 or 700 Hz. Similarly, one may desire to cause a low-frequency roll-off in the 100- to 300-Hz region. The exact frequency is a matter of subjectivity, depending on the operator's choice of receiver fidelity. A good low-frequency roll-off will improve reception by eliminating much of the low-frequency rumble caused by QRN and sideband energy from ssb stations operating near the chosen frequency. Furthermore, 60-Hz hum problems are minimized if shaping of that kind is used. Low-impedance hi-fi headphones are not recommended for use with receivers which do not have audio systems that have been shaped for communications bandwidths. The wide frequency response of such headsets will degrade the readability of weak signals by allowing noise and high-pitched heterodynes to pass, to say nothing about 60- and 120-Hz hum that may be present.

In the interest of reducing the harmonic distortion level of an audio-output amplifier, it is useful to have more audio power capability than is required. When the maximum rated power of an audio IC or discrete-device amplifier is depended upon for adequate sound level, the system is operating in its maximum harmonic-distortion region. Hi-fi designers rely on the general concept of having more audio than is needed, thereby permitting the amplifier to operate over a portion of its curve where minimum distortion will occur.

A 0.5-W IC audio amplifier is shown

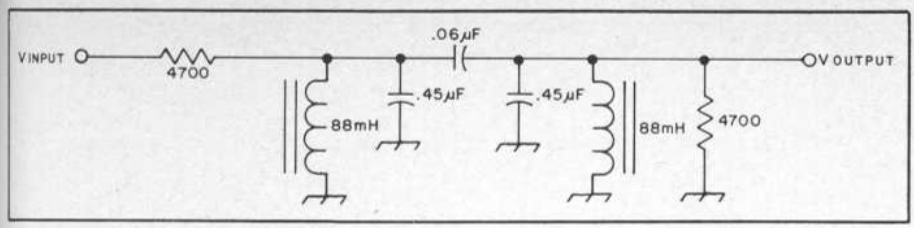


Fig. 15 — Example of a two-pole passive audio filter which contains an 88-mH toroidal-wound inductor in each resonator.

in Fig. 14A. A Motorola plastic 8-pin dual inline device is used. The chip contains a preamplifier and audio-output section for driving an 8-ohm load. The preamplifier voltage gain is nominally 100, and the audio power amplifier has a gain of 10. The combination provides a voltage gain of 1000. With 3 mV of input signal, 0.5 W of audio output will occur.

No-signal resting current is approximately 4 mA at 9 V. The IC works nicely with headphones in the 8- to 2000-ohm impedance class and is quite suitable for use in small portable receivers. The 33-μH rf choke seen at the output port is used to suppress hf parasitic oscillations which can occur. Such unwanted energy can radiate from the circuit board and speaker leads, causing interference to the front end and i-f sections of a receiver. For operation from a 12- or 13-V power supply, it is a simple matter to drop the IC operating voltage to 8 or 9 volts by means of a three-terminal regulator. If the IC is operated from a 9-V battery, a 300-μF capacitor should be placed in parallel with the battery to prevent distortion caused by increased battery resistance as the battery becomes depleted. Under normal operating conditions the harmonic distortion is rated at 0.5 percent at 250 mW of output to an 8-ohm load.

A 1-Watt Amplifier

A Motorola MC1454G can be used when a power output of 1 watt is desired. The IC has ten leads and is contained in a 602B style case (similar to a TO-5 case). Total harmonic distortion is rated at approximately 0.8 percent at 1 kHz while using a 16-ohm load. A practical circuit is given in Fig. 14B. Zero-signal current is approximately 11 mA.

The diagram shows the IC configured for an A_V (voltage gain) of 18, but by making minor changes in the pin connections one can set the gain at 10 or 36, depending on the operator's requirements. Details are given in the Motorola data sheet.

Networks consisting of a 10-ohm resistor and a 0.1-μF capacitor are connected to ground from pins 9 and 10. They help to prevent unwanted rf oscillations. The *R-C* networks and all other circuit connections to the chip should be kept as short as possible to ensure stability. A .05-μF capacitor is employed between pin 1 and ground to decrease the amplifier bandwidth — another aid to stability. This IC can be used safely with headphones which exhibit impedances from 4 to 2000 ohms. Similarly, a 4- or 8-ohm speaker can be used in place of a 16-ohm one, but the lower the voice-coil impedance below 16 ohms, the greater the percentage of harmonic distortion.

A 3.5-Watt Amplifier

In applications where maximum current drain is not a matter of prime importance, the circuit of Fig. 14C is worthy of consideration. A complementary-symmetry Class B audio pair, Q1 and Q2, is driven by U3, a noninverting voltage amplifier which serves as a phase splitter.

This circuit is designed to deliver approximately 3.5 watts to a 4-ohm load. Supply voltage can range from 12 to 14. THD (total harmonic distortion) will be roughly 0.25 percent at 3.5 watts output. Most of the voltage gain is effected at U3, with Q1 and Q2 representing a voltage gain of 3.

An audio preamplifier is necessary ahead of U3 if the system is to be used directly after a product detector. A single-stage Class A amplifier, such as a 2N2222A, will suffice. R1 functions as a protective circuit for the input of U3 during discharge periods of C1. CR1 serves as an antisaturation clamp to prevent latchup of U3. This circuit is patterned after one described by Jung (*IC Op Amp Cookbook*). Idling current is practically zero because Q1 and Q2 are biased off during no-signal periods. Additional audio amplifiers for driving a speaker are presented in the *ARRL Electronics Data Book* and in the *Handbook*.

Audio Filters

When overall selectivity in a receiver is lacking, especially for cw use, a significant improvement can be realized with the addition of an audio filter. There are two common situations. One is when a superhet receiver is designed primarily for ssb and has an i-f bandwidth of approximately 2 kHz. If this receiver is used for cw, an audio bandpass filter can do wonders in reducing the effects of QRM. The other case is when the receiver follows the direct-conversion concept, where all adjacent-channel selectivity must, by necessity, originate at audio frequencies.

Audio filters may be synthesized through two methods. The first is where inductors and capacitors are used to form resonant circuits. These resonators are coupled in order to obtain multipole responses. The other technique (more popular) is the use of *R-C* active-filter sections. Here, capacitors and resistors are used in conjunction with feedback

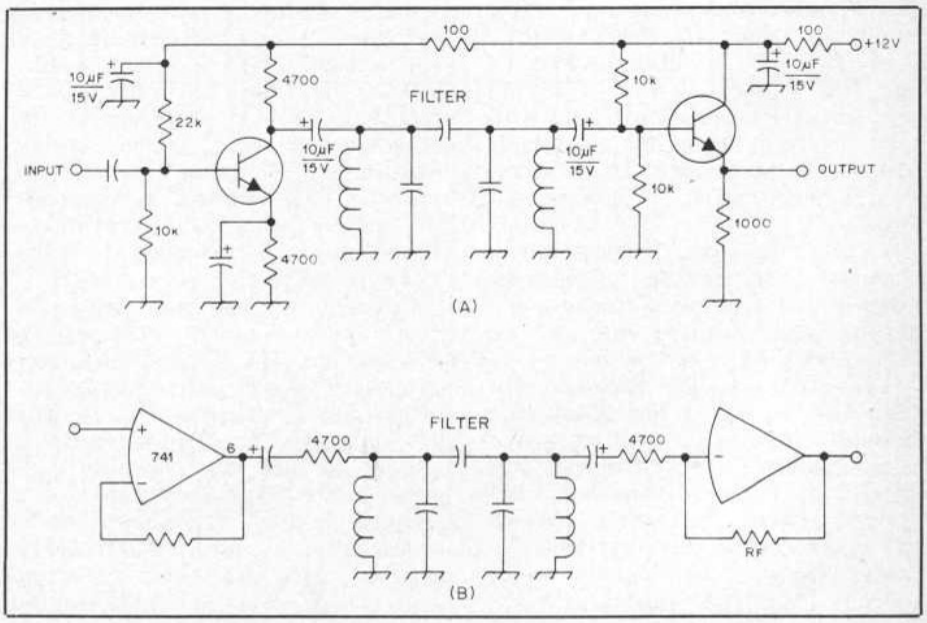
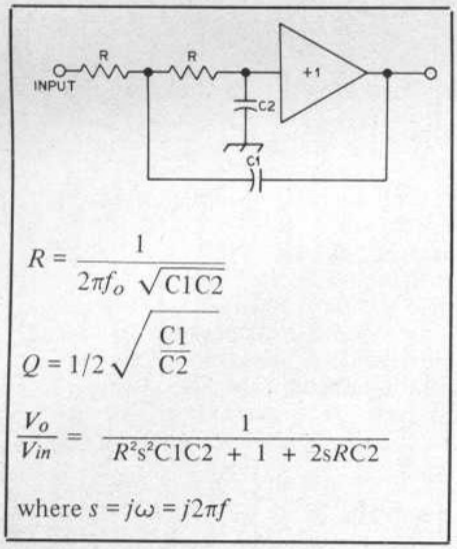


Fig. 16 — Examples A and B show methods for terminating an *LC* filter.



$$R = \frac{1}{2\pi f_o \sqrt{C1C2}}$$

$$Q = 1/2\sqrt{\frac{C1}{C2}}$$

$$\frac{V_o}{V_{in}} = \frac{1}{R^2s^2C1C2 + 1 + 2sRC2}$$

where $s = j\omega = j2\pi f$

Fig. 17 — A simple low-pass filter section using an active device.

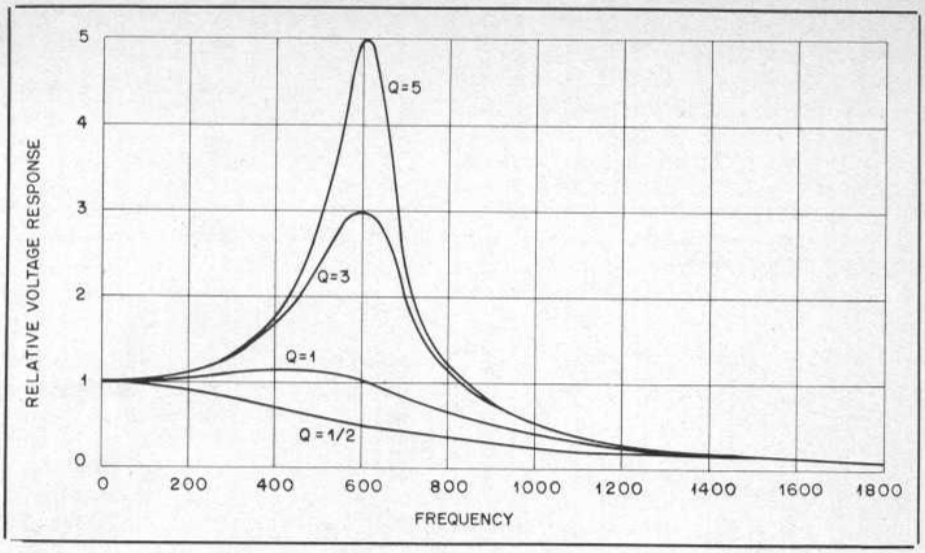


Fig. 18 — Curves for output voltage versus input frequency, illustrating the effects of *Q*.

amplifiers in order to synthesize the same effect that could be obtained with a passive combination of inductors and capacitors. The advantages of the latter are many. First, inductors for the audio frequencies are bulky, heavy and expensive. Their losses are often high. Conversely, resistors and capacitors are lightweight and compact, and are inexpensive. If desired, gain can be obtained from an active filter.

Shown in Fig. 15 is a simple two-pole band-pass filter which is designed around an 88-mH toroidal inductor of the kind used by RTTY enthusiasts. This filter was designed (using predistorted Butterworth tables) for a center frequency of 800 Hz and a 3-dB bandwidth of 150 Hz. The measured unloaded *Q* of the inductors was approximately 25 at 1 kHz.

The operation of any *LC* selective filter is critically dependent upon the resistive terminations at each end of the filter. The unit described in Fig. 14 must have a termination of 4.7-kΩ on each side if the proper passband is to result. Shown in Fig. 16 are two suitable methods for terminating the *LC* filter. Both of these systems can provide considerable gain. In the case where op amps are used, the designer should remember that the use of feedback causes both the output impedance and the impedance looking into the inverting port to be essentially zero.

The more exciting technique for audio filter design is the *R-C* active approach. Virtually all of the response types of interest can be handled. This includes the low-pass, high-pass, and bandpass responses as well as assorted band-reject and all-pass functions. An example of an all-pass response would be seen in the phase-shifting networks of the kind used in phasing-type ssb transmitters or receivers. Only simple low-pass and band-pass responses will be considered in this section.

Shown in Fig. 17 is a simple low-pass filter section. This circuit should be driven from a low-impedance source — one with an output resistance much less than the *R* used in the filter. At dc this circuit will have a voltage gain of unity. However, at well above the cutoff frequency there will be significant attenuation. The response near the center frequency will depend upon the design *Q* of the network, which is determined by the ratio of the two capacitors used. The output voltage will be *Q* times the input voltage at the center frequency, f_o. Fig. 18 presents curves of output voltage versus input frequency for cases where *Q* is 1/2, 1, 3 and 5.

The amplifier used for filters of this kind is quite simple. The voltage gain should be unity and the amplifier should be noninverting. A simple emitter follower using a high-beta transistor such as the 2N3565 is often suitable. Shown in Fig. 19 are two other circuits which may be used. One is a 741 or similar op amp, wired in the follower configuration. The other uses a pair of transistors in a feedback arrangement. Both amplifiers should be biased so the dc voltage is approximately half the supply voltage.

Useful filters are built using the circuits just discussed by cascading many sections. The fact that this circuit has unity gain at dc makes biasing easy. An example is shown in Fig. 20. The first unity gain amplifier is used as a follower to bias the following stages properly. The 10-μF input capacitor is large enough to allow response down to low frequencies. A 0.1-μF unit would be desirable since this would cause the input section to act as a single-section high-pass filter. This would ensure considerable attenuation at 60 and 120 Hz. Earphones can be driven directly from the outputs through an electrolytic capacitor.

In principle, any number of filter sections may be cascaded to obtain the response desired. For most amateur applications identical filter sections are used, resulting in a Bessel type of transfer response, while simplifying the design procedure. It is not necessary that the sections be identical. If the cutoff frequencies and individual section *Q*s are chosen properly, Butterworth and Chebyshev response filters may be synthesized.

Shown in Fig. 21 is a single-section band-pass filter. This circuit differs from

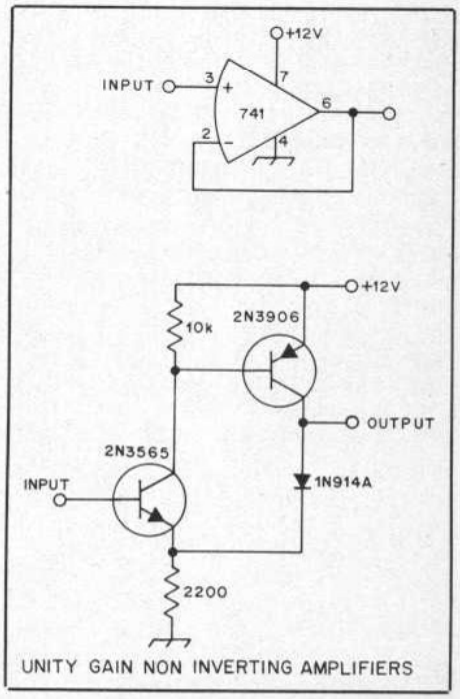


Fig. 19 — Unity-gain noninverting amplifiers.

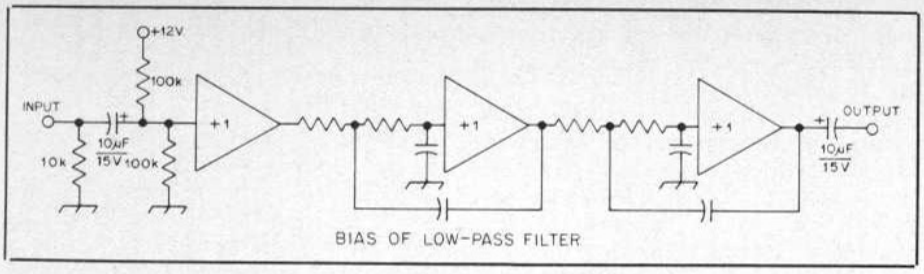


Fig. 20 — Biasing of cascaded filter sections is simple, as shown here.

the low-pass one because there is no response at dc, and the attenuation at high frequencies is not as pronounced as with the low-pass filter. The filter offers some simplification because the capacitors are equal in value. Furthermore, this circuit is capable of yielding considerable voltage gain at the center frequency.

Shown in Fig. 22 are normalized voltage responses for this circuit, as a function of frequency, for design Qs of 1, 3 and 5. The voltage gain at the center frequency can be as high as $2Q^2$.

While high voltage gain is sometimes an advantage, it can cause a problem if the filter is used with an existing receiver. In such cases, it is more desirable to operate a filter with a gain close to unity, or just slightly above. The band-pass circuit of Fig. 21 is modified easily by including an attenuator section at the input, which causes the overall voltage gain to be H_o. This is any desired value less than or equal to the maximum available value of $2Q^2$.

Since the filter section of Fig. 23 has no output response for a dc input signal, it requires a different approach to biasing if a single power supply is used. A circuit using several bandpass sections with a single power supply is shown in Fig. 24A. Multisection filters of this kind may be built with op amps, such as the 741, 747, '5558 duals or the LM-301A. For critical low-noise applications the LM-308N would be ideal, but it is more expensive.

Other circuits may be employed to obtain a bandpass response. However, the results would be essentially the same. The simple band-pass section discussed has the advantage that it is not as sensitive to component variations as some other circuits. This general approach is used commercially for some ready-built filters offered to the radio amateur.

Both of the *R-C* active filters presented allow latitude to the designer in the choice of components. In each case the capacitors may be picked on an arbitrary basis. The design frequency and the Q are then chosen. For the low-pass filter the Q will place a constraint upon the ratio of the capacitors, while the center-frequency gain must be chosen for the band-pass case. After these parameters are pinned down, the resistor values can be calculated. For low-Q situations (Qs less than 6 or 8), the nearest 10-percent resistors can be used. It is advisable to select the larger capacitance values, for this leads to lower resistance values, and keeps the impedances low enough to maintain a low-noise output. Miniaturization would lead one in the opposite direction. For the low-pass filter, a value of 0.1 μF for C1 is a good starting point, with C2 being picked to yield the desired section Q. A value of .022 μF is suitable for the band-pass circuit.

Care must be used when applying these ideas to the design of a direct-conversion receiver. Ideally, for best dynamic range, the place for selectivity in any receiver is at as low a signal level as possible. However, noise considerations may not allow this route to be followed. For example, the active band-pass filter discussed has a resistive attenuator at its input if it is designed for anything less than maximum possible gain. This attenuator, along with the noise in the op amp used for the first filter section, would severely compromise the noise figure and sensitivity of a receiver which used a diode type of product detector — if the filter were to follow the detector. On the other hand, if all of the selectivity of a direct-conversion receiver was concentrated at the output of the audio amplifier, one would have an acceptable noise figure, but the audio amplifier would severely overload from adjacent-channel signals. The best approach would be a combination of the two methods. That is, some passive low-pass filtering should be used between the product detector and the first audio amplifier in order to protect the audio amplifier, with the major close-in selectivity achieved *after* some amplification. It is worthwhile to include selected capacitors within the audio amplifier to attenuate the higher audio frequencies.

A question often posed is whether to use a low-pass or a band-pass filter. This query is difficult to answer, for it will depend to a large extent upon the personal preferences of the user. Certainly, the sharp band-pass filter built with four or five sections, each having a

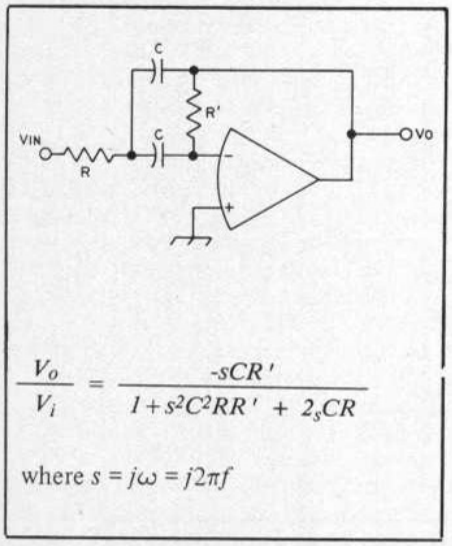


Fig. 21 — Representation of a single-section active band-pass filter.

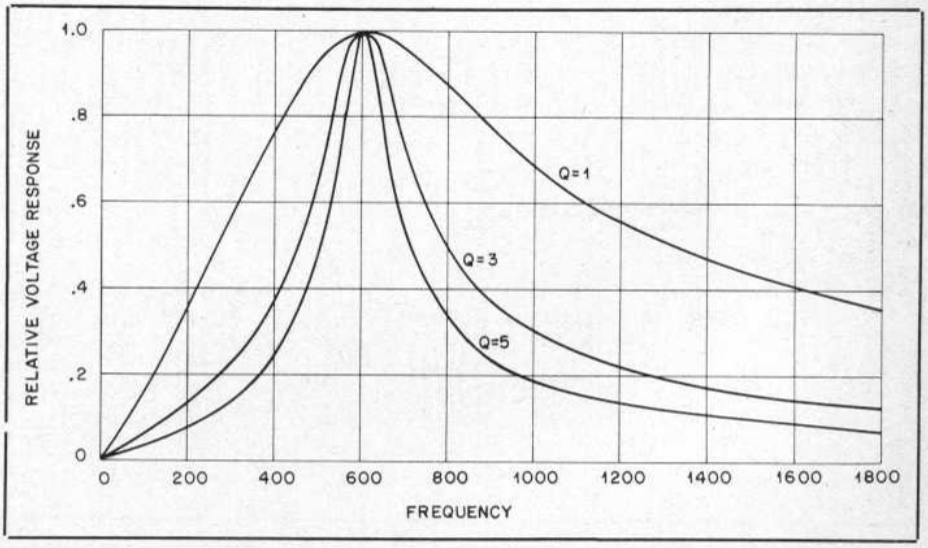
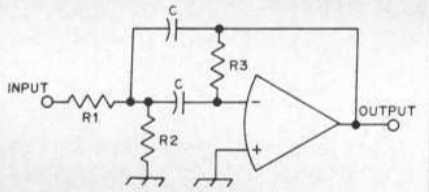


Fig. 22 — Curves for output voltage versus input frequency of the single-section band-pass filter.



Pick H_o, Q, $\omega_o = 2\pi f_c$

where f_c = center freq.

Choose C

Then $R_1 = \dfrac{Q}{H_o \omega_o C}$

$$R_2 = \frac{Q}{(2Q^2 - H_o)\omega_o C}$$

$$R_3 = \frac{2Q}{\omega_o C}$$

If $H_o = 2$, $f_o = 800$ Hz, $Q = 5$

and $C = .022\ \mu F$

$R_1 = 22.6\ k\Omega$ (use 22k)

$R_2 = 942\ \Omega$ (use 1000)

$R_3 = 90.4\ k\Omega$ (use 91k, or 100k)

Fig. 23 – Band-pass filter with suitable design equations.

Q of 5, will be impressive. However, such a filter can cause mental fatigue if it is used for long periods, such as during contest operation.

The writers feel that a low-pass filter with a cutoff frequency of roughly 1 kHz, but with several sections to ensure attenuation at high frequencies, is superior for use with most direct-conversion receivers. Such a filter is shown in Fig. 24B. The constants for a ssb unit are also included. Each section is designed with a Q of unity. However, two low-value coupling capacitors are used at the input and between the last filter section and the low-gain output amplifier in order to attenuate low frequencies and hum. The latter can be troublesome with direct-conversion receivers. This filter has been used with a number of the direct-conversion receivers and transceivers described in this book. Pleasing results were had.

An ideal solution would be to include both filter types in a receiver. The low-pass filter of Fig. 24 could be followed by a bandpass unit with a center frequency of 800 Hz and a narrow band-width. This filter, probably containing only two or three sections, could be used when necessary.

Superhet Basics – I-F System and Filter Design

In the first section of this chapter, the basic ideas governing the design of a superhet receiver were presented and were contrasted to direct-conversion designs. Now, some design information is presented concerning the general methods to be used in designing the i-f section of a superhet. This includes a discussion of crystal filters and other methods for obtaining selectivity. In the next section, the details of some different approaches for building and analyzing suitable amplifiers will be presented.

Envision a superhet receiver which was typical of those used in the late 1940s and early 1950s. This unit was a single-conversion variety – the incoming signal was applied to a mixer, then converted to an i-f where the main selectivity and gain of the receiver was obtained. Then, the signal was detected, yielding audio which was further amplified and applied to headphones or a

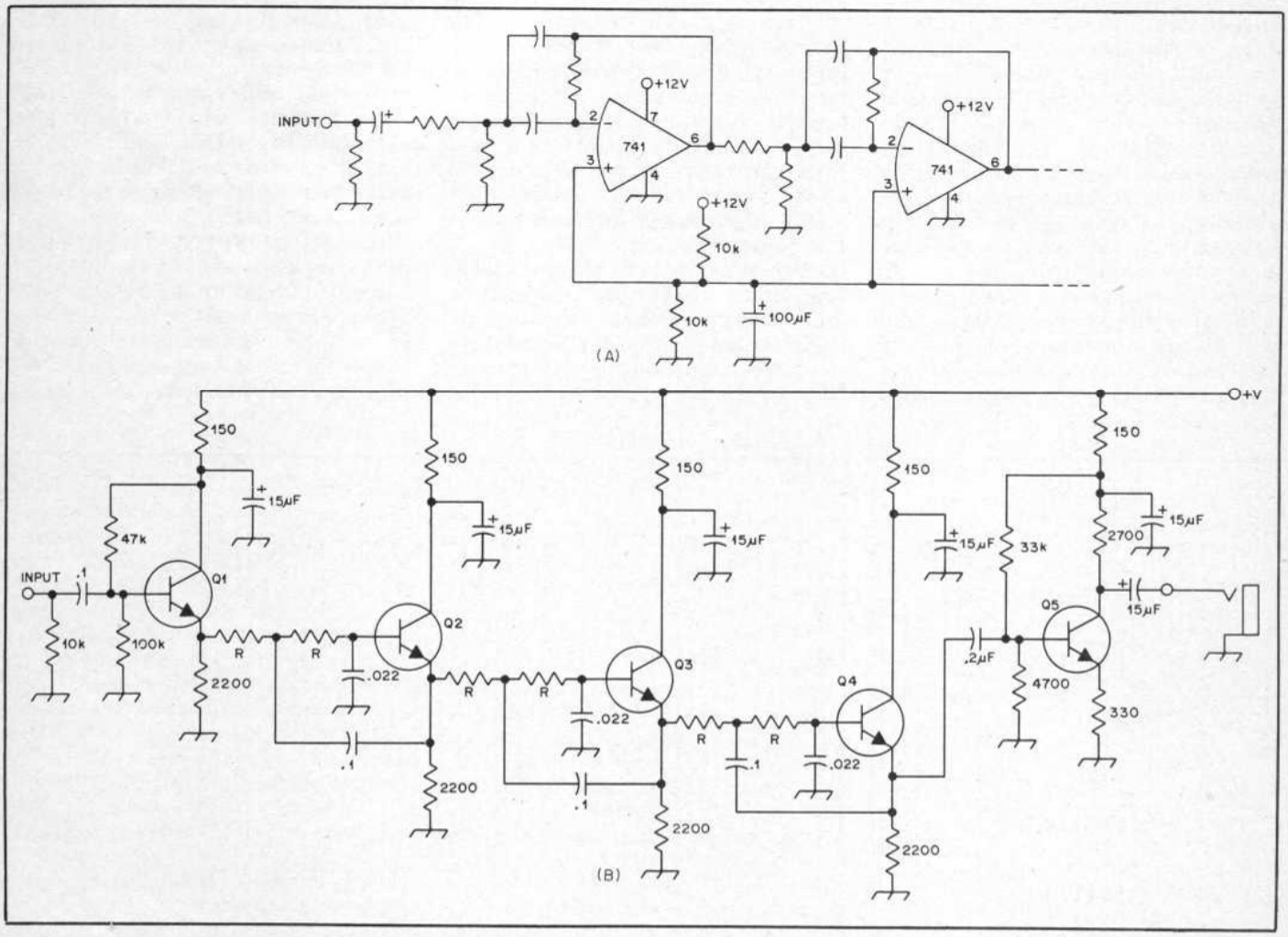


Fig. 24 – Multisection active filters with single power-supply voltage. Q1 to Q5 are 2N3565 (or equiv.). R is 3300 Ω for 1-kHz cutoff or 1500 Ω for 2.3-kHz cutoff.

speaker. The usual i-f was 455 kHz. Such a receiver, set for reception at 14.0 MHz, is seen in Fig. 25.

Note that the local oscillator in this receiver is operating at 14.455 MHz in order to produce a 455-kHz i-f from an arriving 14-MHz signal at the antenna terminals of the receiver. However, the i-f image in such a receiver is the other incoming signal at the mixer input which would also provide a 455-kHz output: in this case, 14.910 MHz. To keep the receiver from being dominated by these image responses, extensive front-end filtering is required. The filtering ahead of the mixer should be so selective that the 14-MHz signal is passed with minimal attenuation while offering considerable attenuation to signals at 14.91 MHz. Such filters can be designed easily, but they are not easily realized in a receiver which must tune over large frequency ranges. Many receivers of the early 1950s had two tuned circuits which were separated by an rf amplifier, yielding 40 to 50 dB of image rejection during 20-meter operation. On the lower amateur bands, the image rejection was better, although up on 10 meters, the image rejection was as little as 10 or 20 dB.

This image-rejection problem led to the popularity of dual-conversion receivers. The early units were similar to that shown in Fig. 26 where the incoming signal was converted first to an i-f of roughly 2 MHz, then was converted to a lower second i-f. The latter was often at 455 kHz, although many units used lower frequencies where selective transformers were more easily constructed. Triple-conversion receivers were used also. A third i-f of 50 kHz was popular.

A second form of dual-conversion receiver was built by Collins Radio (Fig. 27). The first local oscillator was crystal controlled. The first i-f, typically around 2.5 MHz in amateur receivers for the 3- to 30-MHz region, was a broadly tuned affair, often with a bandwidth from 200 to 500 kHz. This broad first i-f was converted to a selective second i-f. The advantage of this scheme was that the stability of the receiver was excellent because of the crystal-controlled first-conversion oscillator. Good frequency accuracy resulted from the high precision which could be used in designing the second oscillator. This was possible since only one tunable oscillator was required.

The image-rejection ratio of dual-conversion receivers of this vintage was often 60 to 80 dB, although this was rarely reflected in the conservative specifications offered by the manufacturers. Moreover, this image rejection was usually as good on the 10-meter band as it was on 80 or 40 meters.

In spite of improved image rejection

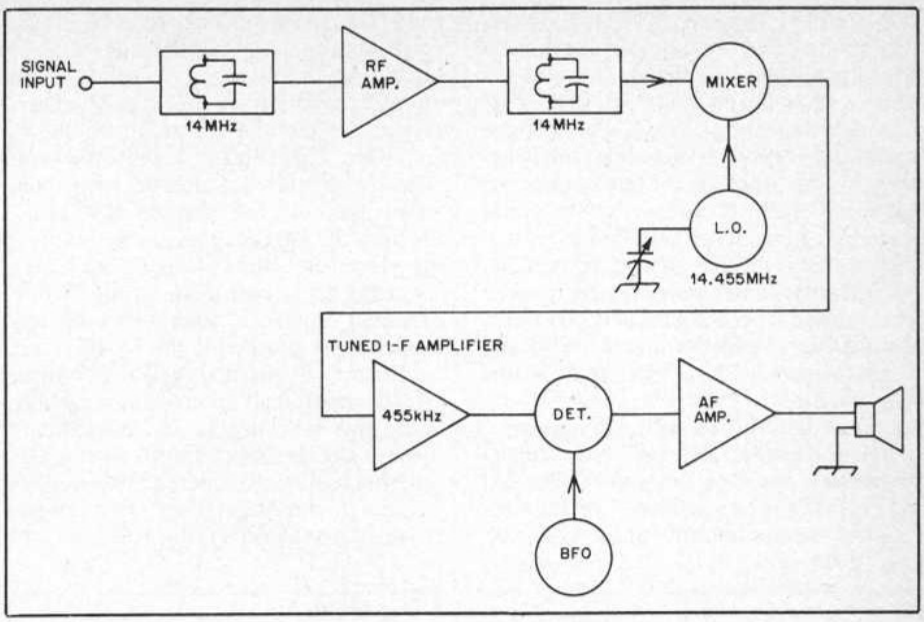


Fig. 25 — Typical receiver format for the late '40s and early '50s.

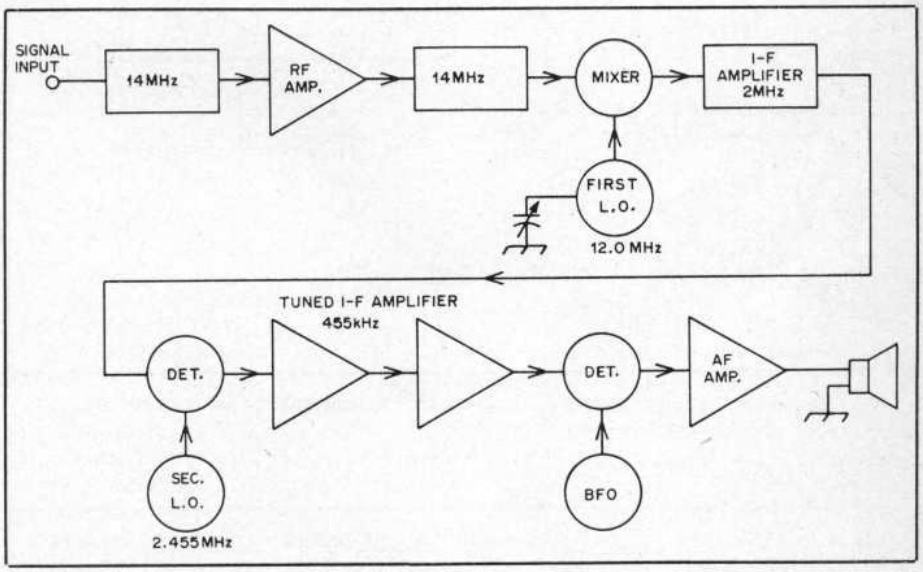


Fig. 26 — Representation of a dual-conversion superheterodyne receiver.

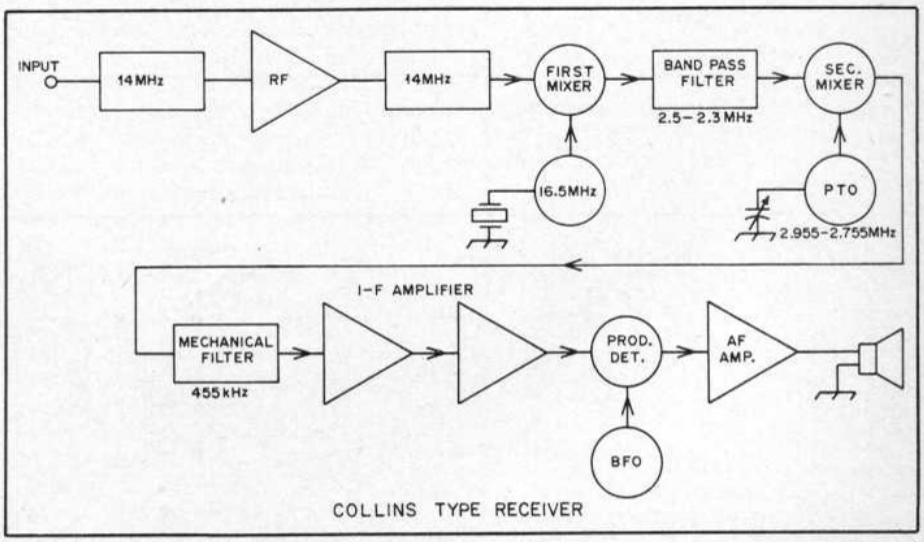


Fig. 27 — Dual-conversion receiver format used by Collins Radio Co.

and stability, the dual-conversion receivers outlined often have problems. These are related to the incoming signal being subjected to several stages of amplification prior to "seeing" the highly selective filters which would appear in the final i-f system. When an incoming signal is subjected to several stages of gain, it grows to fairly high levels. This means that effects from nonlinearity can become significant. These include cross modulation, intermodulation distortion, and blocking. These effects will be discussed in the next chapter.

A partial solution to the nonlinearity problem lies in the use of a single-conversion receiver design, as depicted in Fig. 28. This receiver, which represents most modern units used by today's amateur, differs from the classic single-conversion receiver in that a highly selective filter, usually based upon a multiplicity of high-Q quartz crystals, is used at the input to the i-f amplifier. This filter is usually the most selective circuit in the receiver, and serves not only the purpose of defining the overall adjacent-signal selectivity of the receiver, but of protecting the following i-f circuit from strong out-of-passband signals. In such a design, only those stages preceding the i-f filter are significant in producing the nonlinear effects which lead to cross modulation, IMD, and blocking by out-of-passband signals. The design of the front end of a superhet will be considered later.

The image rejection of a single-conversion receiver of this sort may still be excellent. For example, the receiver shown in Fig. 28 is for reception of the 28-MHz band. The i-f is 9 MHz and the local oscillator is at 19 MHz. In this case, the image frequency is 10 MHz. Building a front-end preselector filter which will offer significant attenuation to 10 MHz (when tuned to 28 MHz) is routine. Image-rejection ratios to 60 to 100 dB are obtained easily. With a 9-MHz i-f system the ultimate image rejection is often limited not by the design of the preselector filter itself, but by shielding and isolation practices.

This brings us to the meat of this section: the design of high-frequency crystal filters. The commercial filters which are popular among amateurs are manufactured in West Germany by KVG and marketed in the USA by Spectrum International. The reader should consult the advertisements in *QST* and *Ham Radio* for information on these filters. KVG filters are offered with center frequencies of 9 or 10.7 MHz. Filters with a center frequency of 3.395 MHz are available from Heath Co. Various crystal filters are offered on the surplus market, many with low prices and superb specifications. Some surplus filters have deficiencies which may degrade their usefulness. Beware!

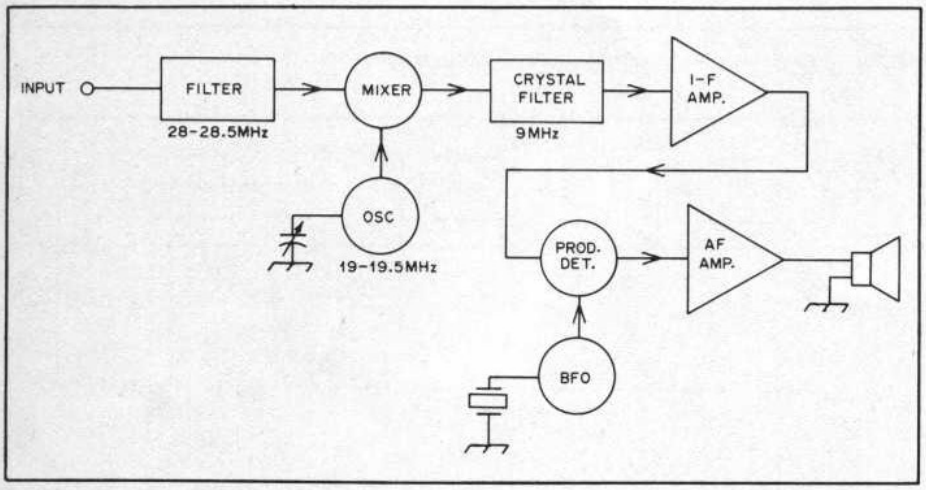


Fig. 28 — Representation of modern-day single-conversion superheterodyne receiver.

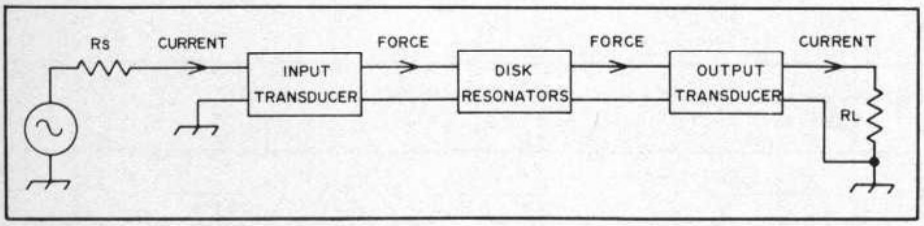


Fig. 29 — Illustration of how a mechanical filter operates.

Electromechanical Filters

A component which is useful for maintaining the required i-f selectivity of a receiver is the *mechanical filter*. Collins Radio Company introduced the first production models of this filter in 1952, and the Japanese followed with a similar unit in the mid 1960s (Kokusai).

Perhaps the most significant feature of a mechanical filter is the high Q of the resonant metallic disks it contains. A Q figure of 10,000 is the nominal value obtained with this kind of resonator. If L and C constants were employed to acquire a bandwidth equivalent to that possible with a mechanical filter, the i-f would have to be below 50 kHz.

Mechanical filters have excellent frequency-stability characteristics. This makes it possible to fabricate them for fractional bandwidths of a few hundred

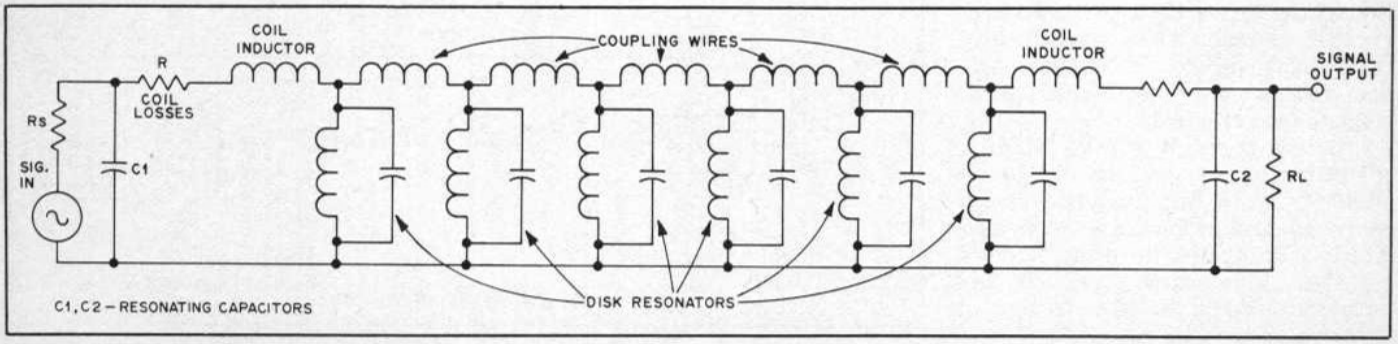


Fig. 30 — Analogous representation of a mechanical filter.

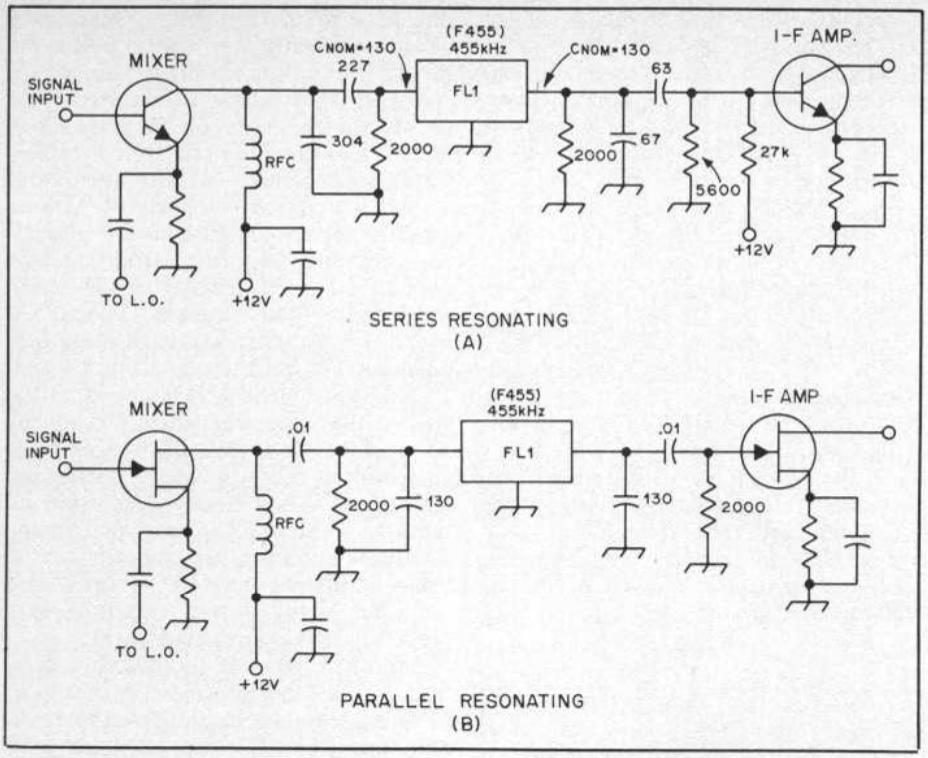


Fig. 31 — Examples of series and parallel resonating when using mechanical filters.

Hz. Bandwidths down to 0.1 percent can be obtained with these filters. This means that a filter having a center frequency of 455 kHz could have a bandwidth as small as 45.5 Hz. By inserting a wire through the centers of several resonator disks, thereby coupling them, the fractional bandwidth can be made as great as 10 percent of the center frequency. The upper limit is governed primarily by occurrence of unwanted spurious filter responses adjacent to the desired passband.

Mechanical filters can be built for center frequencies from 60 to 600 kHz. The main limiting factor is disk size. At the low end of the range the disks become prohibitively large, and at the high limit of the range the disks become too small to be practical.

An illustration of how a mechanical filter works is given in Fig. 29. As the incoming i-f signal passes through the input transducer it is converted to mechanical energy. This energy is passed through the disk resonators to filter out the undesired frequencies, then through the output transducer where the mechanical energy is converted back to the original electrical form.

The transducers serve a second function: They reflect the source and load impedances into the mechanical portion of the circuit, thereby providing a termination for the filter. An analogous representation of a mechanical filter is given in Fig. 30.

Mechanical filters require external resonating capacitors which are used across the transducers. If the filters are not resonated, there will be an increase in insertion loss, plus a degradation of the passband characteristics. Concerning the latter, there will be various unwanted dips in the nose response (ripple), which can lead to undesirable effects. The exact amount of shunt capacitance will depend on the filter model used. The manufacturer's data sheet specifies the proper capacitor values.

Most bipolar transistor i-f amplifiers have an input impedance of 1000 ohms or less. There are situations where the output impedance of the stage preceding the filter is similarly low. In circuits of this variety it is best to use series resonating capacitors in preference to parallel ones. Examples of both methods are shown in Fig. 31. Stray circuit capacitance, including the input and output capacitances of the stages before and after the filter, should be subtracted from the value specified by the manufacturer.

Collins mechanical filters are available with center frequencies from 64 to 500 kHz and in a variety of bandwidths. Insertion loss ranges from 2 dB to as much as 12 dB, depending on the style of filter used. Of greatest interest to amateurs are the 455-kHz mechanical filters specified as F455. They are available in bandwidths of 375 Hz, 1.2 kHz, 1.9 kHz, 2.5 kHz, 2.9 kHz, 3.8 kHz and 5.8 kHz. Maximum insertion loss is 10 dB, and the characteristic impedance is 2000 ohms. Different values of resonating capacitance are required for the various models, spreading from 350 to 1100 pF. Although some mechanical filters are terminated internally, this series requires external source and load terminations of 2000 ohms. The F455 filters are the least expensive of the Collins line.

Crystal Filters

Although a complete theoretical understanding of crystal filters is complicated, it is possible for the advanced amateur to build his own filters. This possibility should not be dismissed as a viable approach. We will not describe the design procedure from a formal point of view: Some basic concepts will be presented which should allow some filters to be built empirically.

Shown in Fig. 32 is the equivalent circuit for a crystal. It is used as the basis for filter synthesis. This circuit shows the normal series-resonant circuit consisting of the motional inductance and motional capacitance which are inherent in the piezoelectric crystal. The parallel capacitance, C_p, is predominantly a result of the metallic plating which is used to provide electrical connection to the quartz plate. Also shown is a series resistance, R_s, which represents the losses in a crystal.

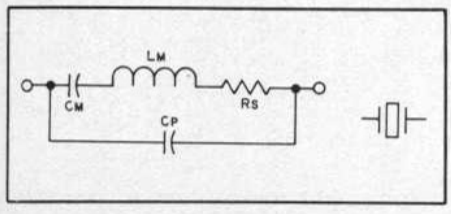


Fig. 32 — Electrical equivalent of a quartz crystal.

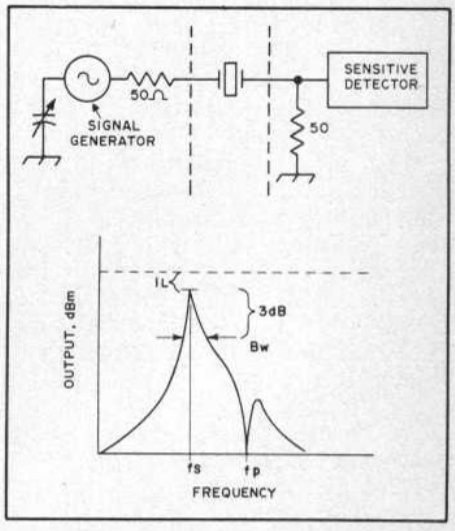


Fig. 33 — Test setup for evaluating a quartz crystal.

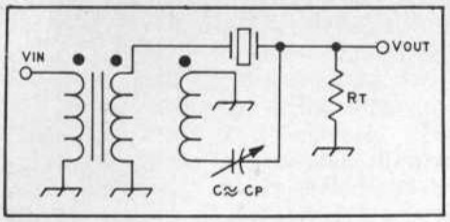


Fig. 34 — Simple form of crystal filter with phasing trimmer.

A test circuit to evaluate a crystal is shown in Fig. 33. Also shown is the response which might be seen if the signal generator was swept slowly through the frequency range of interest. The highest response is measured at the series-resonant frequency, where the motional capacitance and inductance resonate with each other. The amplitude of this response is slightly below the dotted line which represents the signal seen if the crystal is short-circuited. The difference in dB between the series response and the response without the crystal may be used to calculate the value of R_s, the series loss resistance.

The loaded 3-dB bandwidth is also shown. This value may be used to calculate a loaded Q for the crystal. If this is used in combination with the insertion loss associated with R_s, the unloaded Q of the crystal may be calculated. Alternatively, the unloaded Q of the crystal may be measured directly by placing low-value resistors (typically just a few ohms) from each side of the crystal to ground. Extreme signal-generator stability is required for this measurement.

Also shown in Fig. 33 is a parallel-resonant frequency, f_p. This resonance arises from the series combination of the motional inductance and capacitance, which appears to be an inductor at frequencies above the series-resonant frequency. This inductance, when combined with the parallel capacitance, C_p, forms a "trap" circuit, causing a null in the test output at f_p. The difference between the series- and parallel-resonant frequencies is called the *pole-zero spacing* of the crystal.

The parallel capacitance of the crystal, C_p, may be measured directly while using a bridge operating at frequencies far removed from the resonant frequencies of the crystals. Audio frequencies are used for this measurement.

The values which one obtains from these measurements are much different than those encountered with classic LC tuned circuits. For example, an 80-meter crystal was studied while using homemade test equipment, leading to a motional inductance of 69 mH, a motional capacitance of .029 pF, a parallel capacitance of about 8 pF, and a series resistance of 21 ohms. The unloaded Q was 76.000 and the pole-zero spacing was approximately 3 kHz.

The simplest form of crystal filter which can be built by the amateur uses one crystal, and is shown schematically in Fig. 34. A trifilar transformer is used (wound on a ferrite toroid core) in order to provide push-pull drive. One of the outputs drives the crystal directly. The other (out-of-phase) is applied to a variable capacitor. This variable is adjusted for about the same capacitance as the crystal parallel capacitance, and has the effect of canceling the parallel resonance of the crystal, leaving a series-resonant circuit. The value of the terminating resistance, R_t, will determine the loaded bandwidth (BWL) of the circuit. The greater the resistance, the wider the filter will be. This circuit is essentially the same as that which was used in the simple crystal filters in receivers built before 1960.

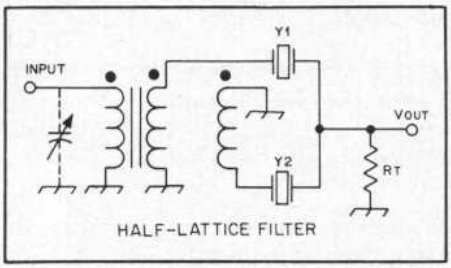


Fig. 35 — Circuit for a half-lattice crystal filter.

Shown in Fig. 35 is another common circuit, the half-lattice filter. The parallel capacitances of the two crystals tend to cancel each other, leaving the response of the filter dominated by the series resonances of the crystals. The transformer consists, usually, of a bifilar output winding on a tuned circuit which is in the output of a mixer. The crystals are on *different frequencies*. The overall bandwidth of the resulting filter is approximately 1 to 1.5 times the frequency separation of the crystals. The spacing in frequencies should not exceed the pole-zero spacing of the crystals, and the crystals should be identical except for the slightly different frequencies. This kind of filter is used in a simple superhet to be described later. In building a filter of this kind, it will be necessary to experiment with the terminating resistance. Generally, with a high-value terminating resistor, there will be passband ripple. As the resistance is decreased, the ripple will disappear, leaving a fairly flat response over a bandwidth determined by the separation in crystal frequency.

Shown in Fig. 36 is a modified version of the filter just described. Four crystals are used. This filter is called a *cascade half lattice*. The transformer balances the drive to the crystals, although the input and output are single ended. The balancing transformer may be built with a few bifilar turns on a ferrite toroid. Alternatively, a bifilar winding can be used on a powdered-iron core. The circuit is resonated with a variable capacitor. Y1 and Y4 should have the same frequency within a tolerance of 10 or 20 percent of the bandwidth of the filter. Similarly, Y2 and Y3 should be matched, although these frequencies will be different from Y1 and Y4. The bandwidth will be a little greater than the frequency difference. As was the case with the simple half-lattice filter, the terminating resistances are critical. They must be adjusted in order to minimize the passband ripple. This type of filter, and variations of it using additional crystals, is the form used for many filters currently employed for ssb and fm equipment.

Another form of filter is shown in Fig. 37. In this example a four-pole filter is presented. In principle this filter may use from two up to dozens of crystals. This filter is called the "lower-sideband ladder" configuration, since when it is built for wide bandwidths, it has an asymmetrical response which tends to pass the lower sideband. Filters of this kind are attractive to the amateur experimenter, for a filter is generally built with all of the crystals cut for the *same* frequency. The empirical approach is to choose the values of the coupling capacitors and terminating resistances in order to arrive at the desired bandwidth. This can be done by the advanced amateur who is willing to build some swept oscillators in order to perform the alignment.

Generally, filters using the lower-sideband ladder configuration are limited to bandwidths which are much narrower (50 percent or less) than the pole-zero spacing of the crystals. The ultimate passband attenuation of such a filter will be limited by the ratio of the parallel capacitance of the crystals to the coupling capacitors. This makes the

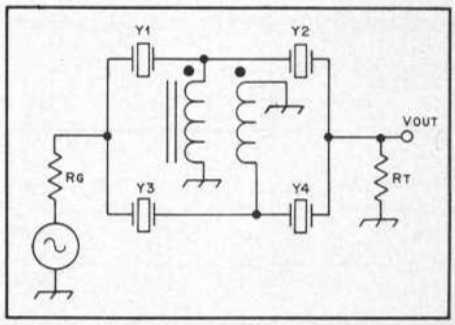


Fig. 36 — Example of a cascaded half-lattice crystal filter.

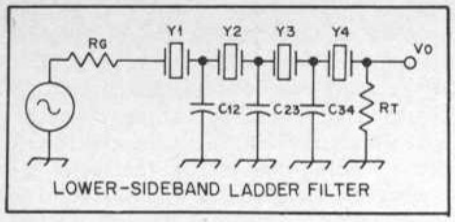


Fig. 37 — Details of a 4-pole lower-sideband ladder filter.

configuration more applicable for cw bandwidths. As a starting point the amateur should consider coupling capacitors up to a few hundred pF and terminating resistances of 50 to 500 ohms. Practical examples of this filter are not given here, since the filter components are highly dependent upon the exact characteristics of the crystals used.

These comments should be kept in mind by the home designer. Many of these statements apply also to *LC* filters.

1) The terminating resistances of a crystal filter will critically affect the response shape and bandwidth.

2) The bandwidth of a multisection filter is determined predominantly by the loaded *Q* of the resonators used and is not a strong function of the number of resonators used.

3) The shape factor of the filter (bandwidth at 60 dB down, divided by the bandwidth at 6 dB) is a function of the number of resonators used and tends to be invariant with filter bandwidth.

4) Extreme care should be used in mounting a crystal filter in order to preserve the ultimate attenuation which the filter is capable of exhibiting. Great care should be taken to ensure that the input of the filter is well isolated from the output. The filter, if built in a metallic can, should be mounted directly against a metallic ground plane.

Intermediate-Frequency Amplifiers

The intermediate-frequency (i-f) amplifier is a critical section of a superheterodyne receiver. Not only must this system provide a large part of the overall gain, but it is the place where most, if not all, of the gain control of the receiver occurs. Both of these functions must be kept in mind when a design is formulated. The noise figure of the i-f amplifier is also of some concern, although it is certainly not as critical as in the front-end part of a receiver.

Consider a modern superhet as shown in Fig. 38. The major selectivity is provided by a multisection crystal filter at the input of the i-f section. The stages that follow will have individual bandwidths which are much greater than that of the preceding filter. Assume that the output of the i-f amplifier was applied to a product detector which was followed by an audio amplifier with a bandwidth of 4 kHz. Since both of the noise sidebands present in the i-f amplifier will be processed by the detector, the effective noise bandwidth of the i-f is 8 kHz. All of the noise generated in the i-f amplifier (within this 8-kHz bandwidth) will appear at the audio output of the receiver.

On the other hand, if the main crystal filter had a 500-Hz bandwidth, the only information arriving, be it signals, antenna noise, or front-end noise, will be confined to this much narrower spectrum. If the receiver front end is designed for wide dynamic range, the net front-end gain may be only a few dB. Thus, the overall noise response of the receiver would be dominated by the noise generated in the 8-kHz effective width of the i-f amplifier.

There are two ways to minimize the i-f noise appearing at the detector. One is to keep the noise figure of the i-f amplifier reasonably low. This is a partial solution. The main need is to restrict the bandwidth of the noise reaching the audio output. This means that additional selectivity is required somewhere in the receiver.

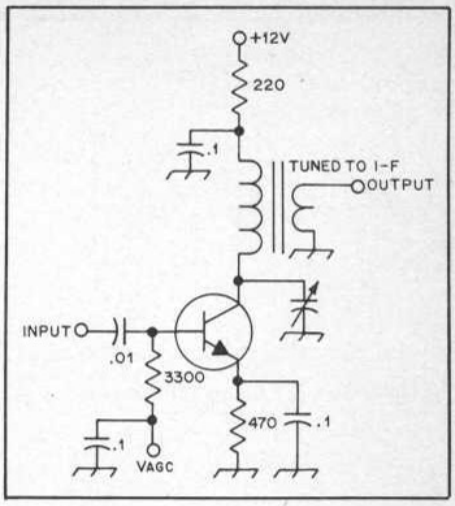


Fig. 39 — A single stage of i-f amplification, utilizing a bipolar transistor.

A partial solution would be the addition of an audio filter within the audio amplifier. If this filter had a 500-Hz bandwidth (matching that of the crystal filter in the beginning of the i-f system), the effective noise bandwidth of the i-f would be 1 kHz. The factor of 2 again results: Both noise sidebands of i-f noise are detected while only one contains useful information.

The ultimate solution is to use proper i-f selectivity just preceding the product detector. If a high frequency is chosen for the amplifier, such as 9 MHz, the only useful approach is to use an additional crystal filter. An *LC* tuned circuit will not add enough selectivity to change the overall bandwidth. The filter in this position need not be as exotic as that used "up front." A filter with one or two crystals is sufficient.

A second approach is the use of multiple conversion. The signal from a 9-MHz i-f crystal filter might be amplified by a low-noise amplifier, then applied to a second mixer with an output of 50 kHz. The rest of the gain is obtained at this frequency, and an *LC* filter is used at the system output to maintain the bandwidth the same as that of the original crystal filter. The best means for building narrow i-f filters in the 50-kHz region is probably to use ferrite pot cores. The major signal selectivity is still obtained best with the initial crystal filter.

If a multiplicity of crystal filters is used without double conversion, the two filters should be well matched in frequency. Some filter suppliers will

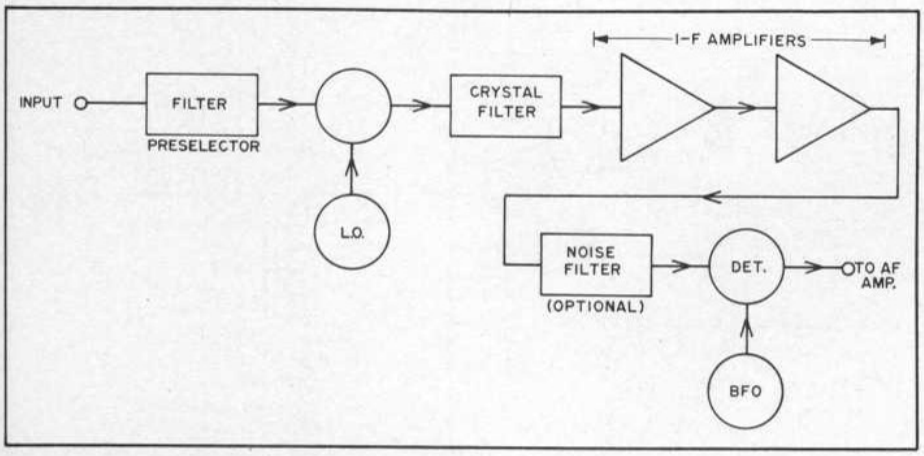


Fig. 38 — Block diagram of a modern superheterodyne receiver.

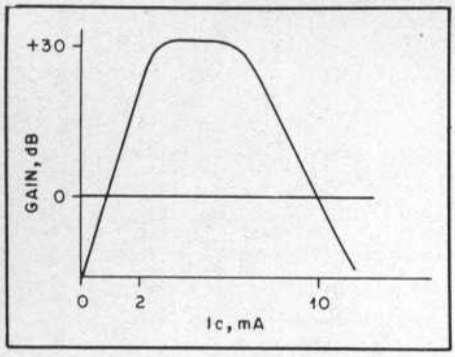


Fig. 40 — Gain as a function of current.

provide matched sets of filters for a nominal charge. If multiple conversion is employed, the system is more complicated. However, the additional advantage gained is that effective decoupling and shielding are much easier to achieve at the lower frequencies. This may be an asset when the noise-modulation effects from the BFO are considered. This phenomenon was outlined in the section on product detectors.

When choosing devices for the active stages in an i-f amplifier, there are a number of points to consider. Mentioned above were overall gain and the ability to easily change the gain over a wide range. Additional problems are presented to the first stage. This amplifier follows the main crystal filter, directly. Hence, it should have an appropriate input impedance to terminate the filter properly. Also, this stage should have a low noise figure.

Bipolar Amplifiers

Bipolar transistors have been used traditionally in the i-f sections of solid-state receivers. If designed properly they may provide excellent performance Shown in Fig. 39 is an example of such an amplifier. The gain is highly dependent upon the transistor chosen. Values of up to 30 dB are not uncommon. If the amplifier is used to follow a crystal filter directly, it should be designed to have a constant, well-defined input impedance. This is realized through proper biasing of the stage and by the application of feedback. The fundamental details of the application of emitter-degeneration feedback were presented in chapter 2 in the discussion of Class A buffer amplifiers for transmitter applications. Additional information on the use of shunt feedback is presented in the later discussion of ssb amplifiers.

Depending upon the transistor used, there are two ways that the gain of a bipolar transistor amplifier may be changed. The more common one is the application of *reverse agc* (automatic gain control). This is realized by decreasing the current flowing in the amplifier. The decrease in current leads to a decrease in the gain of the amplifier. This technique will work with almost any transistor that might be used.

The use of reverse agc in an amplifier has some disadvantages. First, as the current decreases, the input impedance of the amplifier will increase. This can cause the selectivity characteristics of the receiver to change dramatically if the amplifier follows a crystal filter directly. Another problem relates to the signal-handling ability of the amplifier. As the signal being received becomes stronger, the gain of the amplifier is reduced. However, as the current in the stage is dropped in order to reduce the gain, the ability of the amplifier to handle the signal without distortion is impaired severely.

The signal-handling ability problem may be circumvented by the application of *forward agc*. Special transistors are required for such operation. However, since these methods are used commonly for i-f amplifiers in TV receivers, the transistors are available and inexpensive.

Forward agc implies that as the current in a stage is increased, the gain decreases. A curve of gain as a function of current is shown in Fig. 40. The advantage of forward agc is that the transistor is operating with the highest currents when it is asked to amplify the largest signals. This tends to diminish distortion effects. Examples of forward-agc transistors are the Motorola MPS-H30, MPS-H32, MPS-H01, and MPS6568. A number of similar devices are available from Fairchild Semiconductor.

Negative feedback should not be applied to a bipolar amplifier that is used for gain control. The effect of negative feedback is to make the stage gain relatively independent of the transistor characteristics. This is opposite the effect desired.

Shown in Fig. 41 is a circuit of a two-stage bipolar amplifier which utilizes both reverse and forward agc. The dc biasing feedback is such that as current is pulled out of the agc point, the current in the first stage will decrease while that in the second will increase. The second stage uses a transistor chosen specifically for good forward-agc characteristics. This amplifier has a total gain of about 50 dB, and exhibits a gain-control range of 80 dB.

Most i-f amplifier devices will show an increase in noise figure as the gain is reduced. This can have the effect of placing an upper limit on the output signal-to-noise ratio of a receiver. This is rarely of significance in amateur

Fig. 41 — A two-stage amplifier which uses forward and reverse agc.

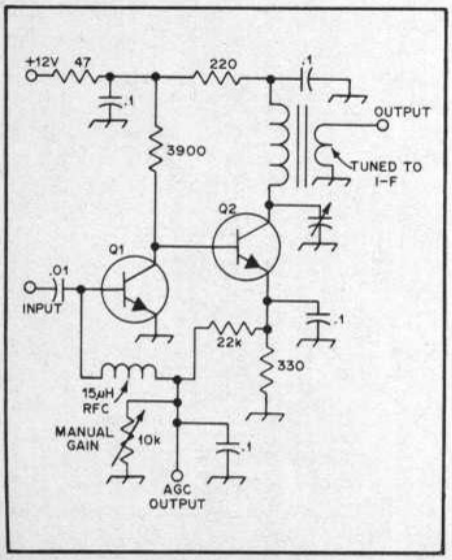
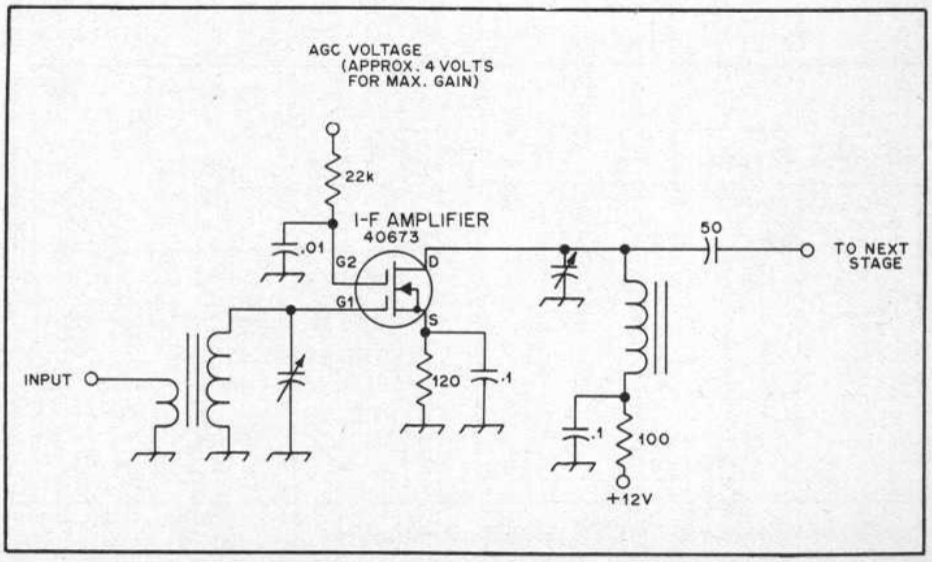


Fig. 42 — A dual-gate MOSFET i-f amplifier.



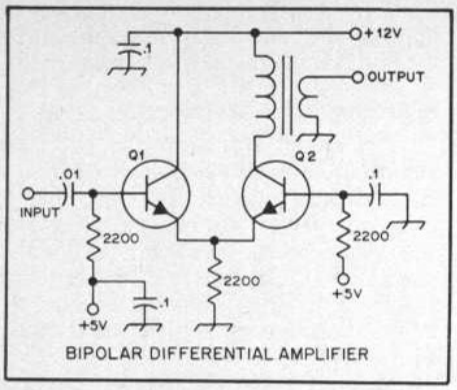


Fig. 43 – A differential pair i-f amplifier.

receivers. The main objective is to ensure that the noise figure does not increase faster than the gain decreases, as agc is applied.

MOSFET I-F Amplifiers

A popular i-f amplifier is the dual-gate MOSFET. This device has some attributes that make it attractive. First, very high stable gain can be realized. The noise figure can also be made exceptionally low. Techniques for achieving low noise figures with MOSFETs are discussed in the following chapter. Finally, by changing the bias on gate 2 of the device, considerable gain reduction can be realized. An i-f amplifier using a dual-gate MOSFET is shown in Fig. 42.

An advantage of a MOSFET amplifier is that the input impedance is relatively independent of the gain and current in the device. Furthermore, the distortion properties are relatively good, considering the low currents required. In spite of these assets, the device is not a panacea. One problem is that the noise figure of the MOSFET increases fast as the gain is decreased. Also, the distortion properties degrade markedly as reverse agc is applied to gate 2. This is evident if gate 2 is made more negative than the source. The reader is referred to the appendix for information on the analytical design of MOSFET amplifiers.

IC I-F Amplifiers

Shown in Fig. 43 is the circuit of an amplifier using a differential pair of bipolar transistors. Although it may not be obvious, the two transistors are operating essentially in push-pull. This can be seen by considering the effect of a positive-going signal at the base of Q1. This voltage causes the current in Q1 to increase. However, the emitter resistor common to the two stages supplies virtually a constant current to the pair of transistors. Hence, as the current in Q1 increases, that in Q2 decreases. Signal currents flow in both transistors with opposite phase.

The differential amplifier has its input impedance higher by a factor of 2 as contrasted to a single-stage amplifier. This can be used to advantage in terminating crystal filters.

The gain in a differential amplifier may often be lowered by decreasing the current supplied to the two emitters. While this could be achieved by lifting the grounded end of the emitter resistor and applying a positive potential, it is done more easily with an additional transistor.

This brings us to a popular IC i-f amplifier using the RCA CA3028A. A circuit is shown in Fig. 44. Q3 in this amplifier acts as a constant-current source to supply the emitters. Because of controlled techniques applied in the manufacturing of ICs, Q1 and Q2 are virtually identical. This results in good balance in the outputs. Also, since the resistors for biasing Q3 are built into the IC, circuit simplification is realized.

Reverse agc is applied to the CA3028A by decreasing the voltage on pin 7 of the chip. This causes the current in Q3 to decrease. Since the collector current of Q3 is equal to the total current in the other two transistors, their combined gain decreases. While the problems outlined for reverse agc are found in the CA3028A, the simplicity of the circuit makes this chip popular.

Fig. 45 illustrates a cascode amplifier using the CA3028A. The circuit has some interesting properties. The input signal is applied to the base of Q3, and Q1 functions as a common-base amplifier. Because the emitter voltage of Q1 remains fairly constant because of the bypass capacitor on the base of Q1, the collector voltage of the input stage, Q3, also remains constant. This results in minimal capacitive feedback in the input stage, ensuring good stability and excellent input to output isolation.

Under normal bias conditions, with no agc voltage applied to the circuit of Fig. 46, the output current of Q3 will be routed directly into the emitter of Q1. However, as current is injected into the base of Q2, this transistor will begin to conduct. As a result, part of the collector current in Q3 will be routed through Q2, causing a decrease in the signal flowing in Q1, the output. With this type of agc the operating biases on Q3 remain constant. Because of this, the input impedance of the circuit remains constant.

While the agc range available from a cascode amplifier of the type shown in Fig. 45 is limited, the technique can be applied in more complicated circuits. An IC i-f amplifier that uses this "current-robbing" method for agc is the Motorola MC1590G. A less expensive cousin is the MC1350P. A circuit using this IC is shown in Fig. 46.

The main advantages of the MC1350P amplifier come from its sophistication. Three differential amplifiers are contained in one package. The middle differential pair of transistors is paralleled with an extra pair that serves the role of current robbing from the main signal path. The MC1350P is capable of gains up to 65 dB and has agc ranges of comparable value. A curve of gain reduction versus applied agc voltage is much smoother than that of the typical differential amplifier. This

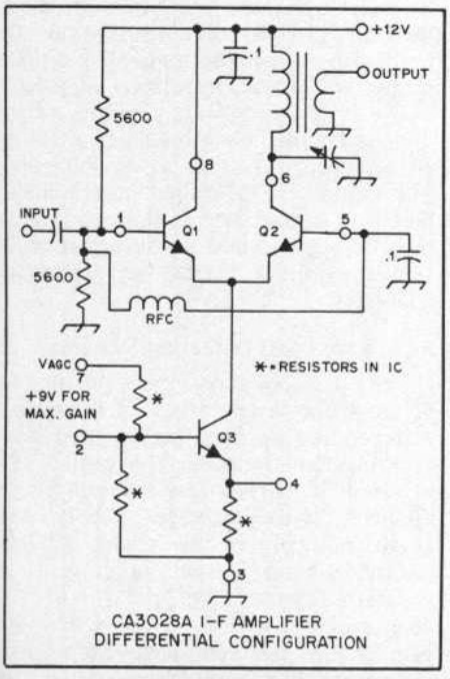


Fig. 44 – Illustration of a CA3028A differential i-f amplifier.

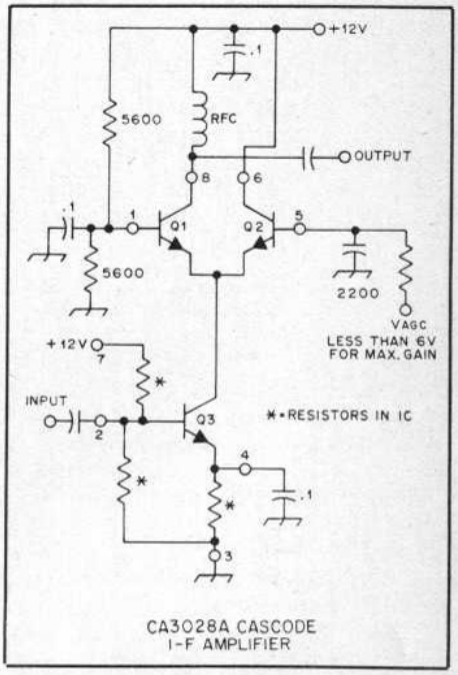


Fig. 45 – Cascode i-f amplifier using a CA3028A IC.

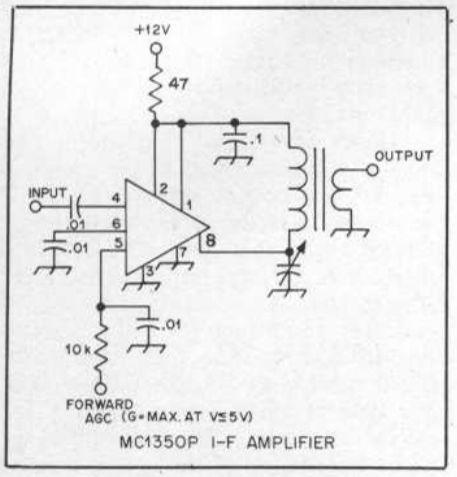


Fig. 46 — I-f amplifier in which an MC1350P IC is used.

has the effect of providing a better dynamic response in an agc system.

PIN Diodes in I-F Amplifiers

Much of the discussion has been about the agc characteristics of i-f amplifiers. Because this is an important function in the i-f system, other parameters are often compromised. These include noise figure, linearity, and impedance matching. Many of these deficiencies may be overcome through the application of PIN diodes.

The usual junction switching diode consists of adjacent layers of p- and n-doped semiconductor material. The junction between the two regions is made as small as possible in order to enhance the switching speed of the device. On the other hand, a PIN diode is made with a fairly large region of intrinsically doped semiconductor material between the p and n regions: hence the terminology of the device.

The effect of the intrinsic layer is that diode action is very slow. As a rectifier of rf most PIN diodes are nearly useless. We can take advantage of this. Because of the slow response time of the diode, it tends to behave as a resistor for rf currents, with the value of the resistance being dependent upon the dc current flowing. A common relationship would be $R_{rf} = k \div I$, with a typical value for k being around 50-ohm mA. Hence, if the diode is biased for 1 mA of dc current, the rf resistance is 50 ohms. If the dc current is increased to 2 mA, the rf resistance drops to 25 ohms. The significant characteristic of PIN diodes is that the rf current can actually be much larger than the dc current.

Hewlett-Packard is a major supplier of PIN diodes. Often it is possible in amateur applications to use high-voltage rectifier diodes in place of PINs, since the doping profile of the junction is similar. Sabin (*QST* for July, 1970) recommended the Motorola MR-990A for this application. The diodes will appear resistive for rf current so long as the rf voltage across each junction does not exceed about 20-millivolts rms. The MR990A contains four series junctions.

There are ways that PIN diodes can be used in the design of i-f amplifiers. Two are shown in the circuit of Fig. 47. In one case the diode is in series with the bypass capacitor in the emitter of the amplifier. As the dc current is increased through the diode, the gain will increase in the stage. In the other example the diode is in parallel with the collector of the amplifier. As current increases in the diode, the gain decreases.

If the designer is careful, he may construct attenuators with combinations of PIN diodes. These networks can have the virtue that the input impedance is fairly constant as the gain is varied. Such attenuators would be ideal in the front end of a receiver. The need for preserving a constant impedance comes from the requirement that front-end preselector filters need proper termination. Shown in Fig. 48 is an attenuator of the bridge-Tee variety which uses two PIN diodes. The pair is biased from a constant-current source in such a way that as the agc voltage is applied the current in one diode increases as it decreases in the other.

Switching in I-F Amplifiers

PIN diodes are useful for switching functions in receivers. One application is for switching crystal filters in order to change receiver bandwidth. A related use would be the switching required to use a crystal filter for both transmit and receive in a single-sideband transceiver. A common receiver application is shown in Fig. 49.

Many of the switching functions outlined here can be handled with high-speed switching diodes, like the 1N914A. If these diodes are used, they should be biased so the dc current flowing in them (when on) is much larger than the rf current being switched. Similarly, an "off" diode would be reverse biased by a voltage which is much larger than the peak signal amount that will appear across it. If these precautions are not followed, IMD may occur.

With the methods presented for design of i-f amplifiers, the reader may question which is best for his application. While this might be subjective, it will depend upon the application. For the typical amateur receiver where some IMD within the i-f amplifier is acceptable, the IC approach is recommended. Not only is the performance adequate, both for gain and agc capability, it is straightforward.

It is interesting to note that most commercial equipment uses IC i-f amplifiers. This includes receivers for the radio amateur as well as for TV viewers. On the other hand, professional equipment leans toward the use of PIN diodes for gain variation. Amplifiers are made sometimes from FETs or ICs, but are built also with premium-quality bipolar transistors. These transistors may have an f_T in the microwave region, and are operated with heavy feedback in order to obtain stable and repeatable gain. The equipment described here includes receivers in the several-thousand-dollar price category, and frequency-domain instrumentation, such as spectrum analyzers.

AGC Loops and Detection Systems

The previous section was devoted to i-f amplifier design. Much of the design is dependent upon obtaining good gain-control characteristics. The gain of the i-f amplifier should vary smoothly with applied control voltage. Ideally, the curve of gain in dB versus applied control voltage should be close to a constant-slope straight line. The unsuitable situation would be one where the gain change becomes large for a small change in control voltage.

Fig. 50 shows a total agc system. The main element is the variable-gain

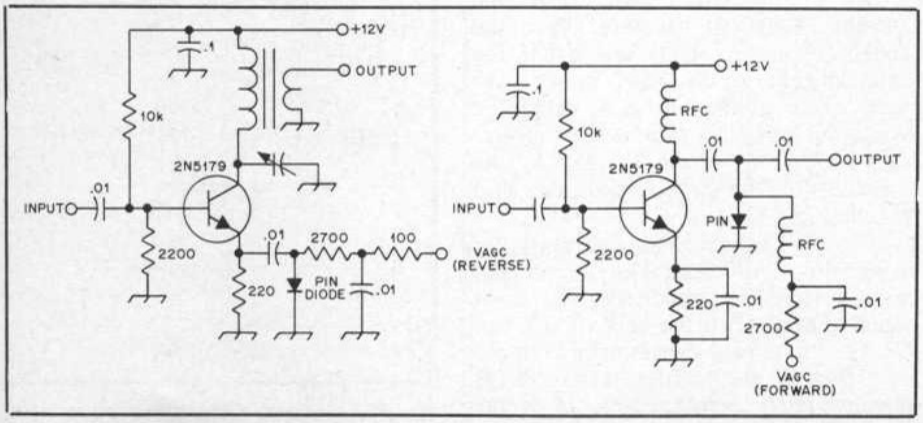


Fig. 47 — Gain control by means of PIN diodes.

amplifier. This might be followed by a mixer or a product detector which would have a different output frequency than the one at which the main amplifier operates. Eventually, a low-impedance source is used to drive a diode peak detector. This produces a dc control voltage on capacitor C1. This output is increased in a suitable dc amplifier and applied to the control line of the variable-gain amplifier. The dc amplifier may be inverting or noninverting, depending upon the nature of the desired control voltage. The choice is made so that an increased voltage on C1 leads to a decrease in gain of the controlled amplifier.

There are two schemes for detection. One detects the i-f signal while the other uses the audio that is present in the receiver. There are valid arguments for each approach. While the audio-derived agc systems are often easier to build, we will show that the i-f derived system is much better from a dynamics point of view.

Consider first the case of audio peak detection. Shown in Fig. 51 are the waveforms that will result – assuming that initially the system is operating at full gain and that the agc loop is opened at point X in Fig. 50. At some instant ($t = 0$) a strong carrier appears in the passband of the receiver. The resulting audio signal that is applied to the input of the detector is shown in Fig. 51A. The current that will flow in the detector diode is shown in part B of the figure, while the resulting voltage on C1 is displayed at Fig. 51C.

Consider now what will happen if the agc system is again turned on by removing the open circuit at point X of Fig. 50. Assume that the *desired* maximum level of audio output is V_A peak volts (Fig. 51A).

When the instantaneous voltage at the detector input reaches this level, C1 will have been charged to a level which will stabilize the gain. However, the audio cycle has just barely started. In reality it continues to grow, placing more charge into C1. Once the peak of the audio cycle has been reached, no additional diode current flows. In all likelihood, the capacitor will have charged too far, and no additional audio output will occur for several cycles of audio output. The capacitor will slowly discharge through R1 until the gain recovers to the point where current pulses again flow in the diode at the audio peaks. Because the level is now changing slowly in comparison to the rate that the current pulses are arriving from the diode output, the agc loop will now follow the strength variations of the arriving signal, holding the output fairly constant. However, the initial overshoot described not only causes a large click or thump in the receiver output, but may cause information to be lost for a short period.

The answer to stabilization of the audio-derived loop is to add some resistance in series with the diode (or to increase impedance of the diode driver). This will slow the response to the point that the capacitor C1 may not become completely charged by one cycle of audio. Unfortunately, this reduces the rate that i-f gain is reduced and leads to the initial information causing excessive receiver output.

Consider now the case of an i-f derived detection system. This is shown in the set of curves shown in Fig. 52 where the time scale is essentially the same as that used for the audio-derived case. There are a number of different features. First, the rate that current pulses from the diode detector are applied to the memory capacitor, C1, is

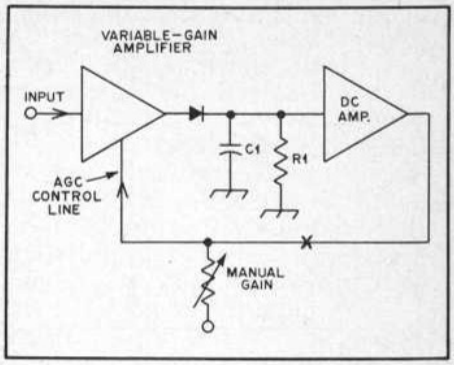


Fig. 50 – Circuit representation of a total agc loop.

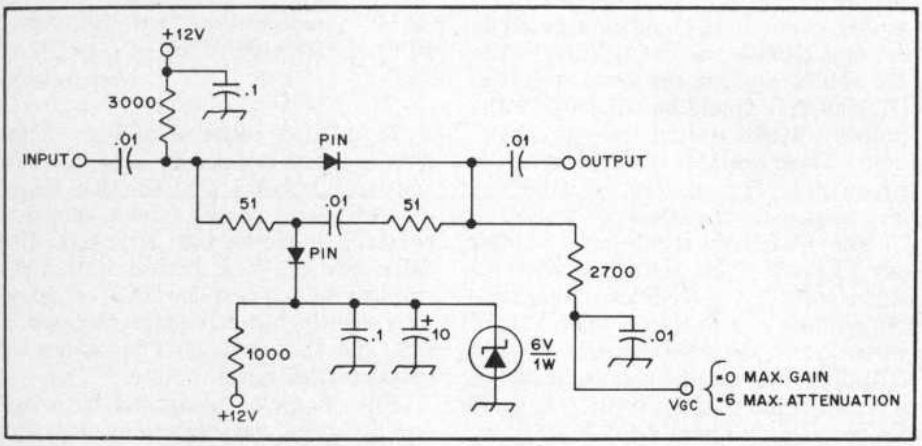


Fig. 48 – Bridge-Tee attenuator using PIN diodes.

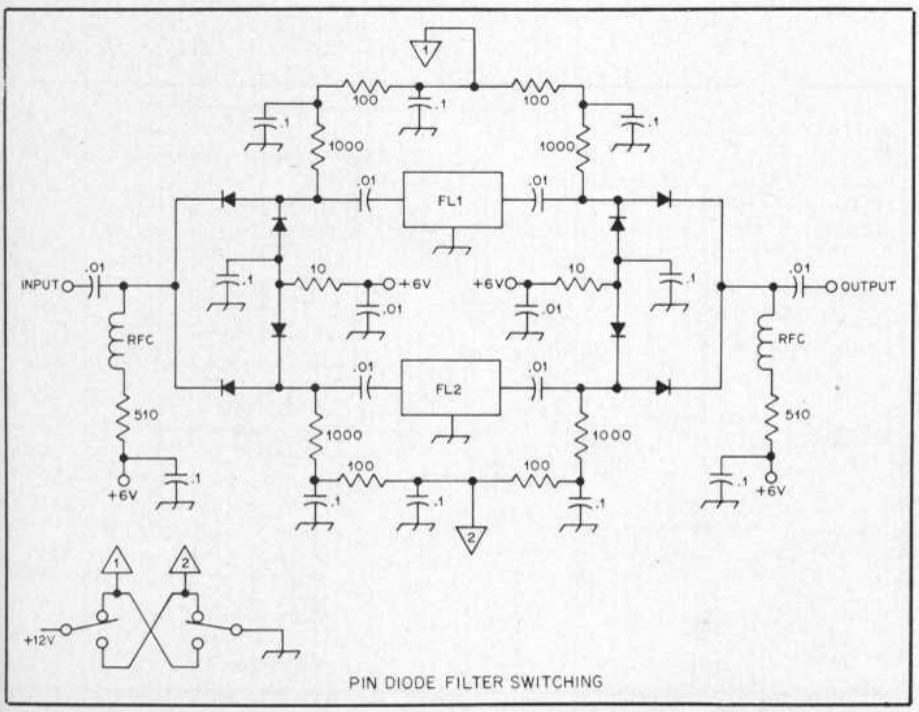


Fig. 49 – Diode switching with PIN devices in i-f filter section of a receiver.

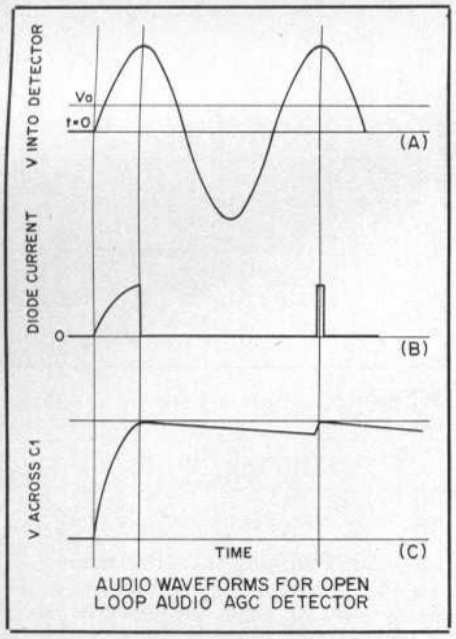


Fig. 51 – Waveforms for open-loop audio agc detector.

much higher. Hence, the impedances may be adjusted so that a single pulse does not charge the capacitor completely, without seriously slowing down the loop response time – the cause of overshoot effects. Second, even though the signals arriving at the input to the i-f filter of the receiver may all be constant in amplitude, the resulting filter output will not reach a stable amplitude immediately. This is because any filter has a rise time that is related to the filter bandwidth. The narrower the filter, the longer the rise time will be. In this situation, the agc loop is capable of responding fast enough that the gain will adjust itself so that the input signal is followed. The bandwidth of the control system should be wide in comparison with that of the information arriving from the i-f filter.

In spite of the deficiencies of audio-derived agc detectors, they may be used with satisfactory results in some receivers. The transient problems outlined here are much more severe when designing a cw receiver than they are with ssb. This is especially true if the BFO is adjusted to provide a low-pitch beat note.

A few tricks may be applied to improve the attack characteristics of audio-derived systems. The first is to employ full-wave detection instead of the half-wave type outlined in Fig. 51. Full-wave detection may be achieved with a center-tapped transformer, or with a pair of op amps. Examples are shown in Fig. 53.

Another method is the judicious application of clipping. A sample circuit is shown in Fig. 54. In this case the response time of the loop is slowed by addition of resistance in series with the diode detector. However, the audio output of the receiver is prevented from becoming excessive (thus protecting the operator's ears) by limiting the level of audio signal applied to the receiver output and to the agc detector. The control in Fig. 54 should be adjusted so the clipped peak voltage at the detector is about 3 dB above the level that the agc loop establishes eventually. If an oscilloscope with good triggering characteristics is available, the dynamics may be adjusted so stabilization will occur within about ten cycles of audio output.

Shown in Fig. 55 is a pair of agc systems that may be applied with i-f amplifiers using CA3028A or MC1350P ICs. These circuits may be used with audio or i-f detection. In each case a JFET is used as the input to the error amplifier. Suitable npn transistors are 2N3565s, 2N2222As or any equivalent silicon device. The pnp transistors are similarly uncritical. Good choices would be the 2N3906 or the 2N3638. The controls shown in the error amplifier (R2 and R3) should be adjusted for the proper voltages during full-gain conditions. These voltages are marked in the schematics. The systems also include means for manual control of the gain.

The FET type is arbitrary. Almost any FET will work, since it is used in a circuit with heavy feedback. The pinch-off should not be more than 5 or 6 volts, but other parameters are not critical if the supply voltage is 12 or more. In each circuit provision is made for muting the amplifiers. That is, by grounding the point marked "M" the gain of the i-f may be reduced to its minimum value.

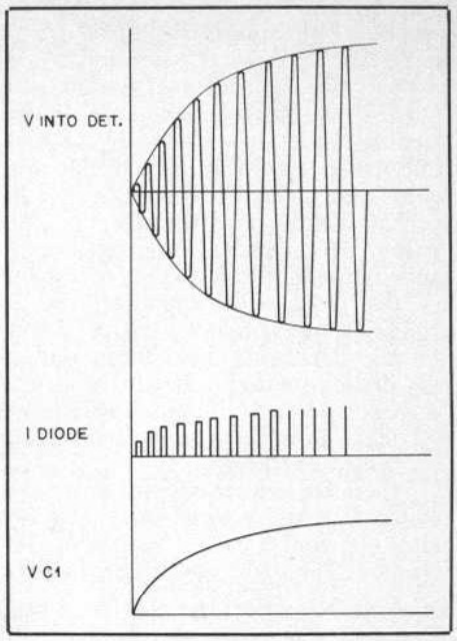


Fig. 52 – Characteristics of an i-f derived agc detection system. See text.

In the two systems of Fig. 55 the recovery time is determined by the time constant, τ = R1 C1. For the longer recovery times desired for ssb, the time constant should be 1 to 2 seconds. One deficiency of these circuits is that the stronger signals will cause C1 to charge to a slightly higher voltage. Because of this, the time will then be somewhat longer for full gain to return.

Fig. 56 shows an agc-detection system that overcomes this deficiency. This circuit may be used with i-f or audio-derived detection. A pair of detectors is utilized to produce a full "hang" action.

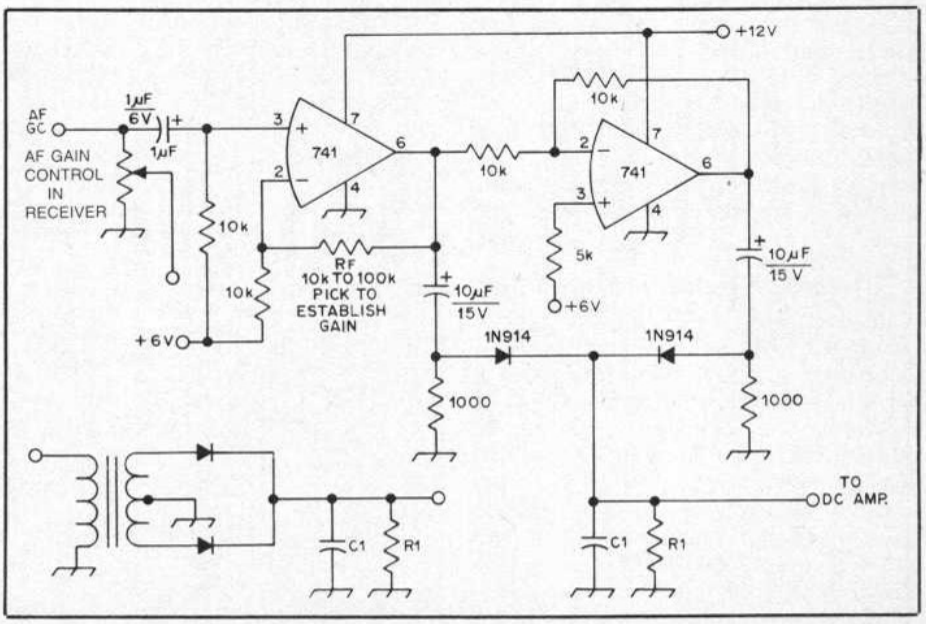


Fig. 53 – Examples of full-wave audio agc detectors.

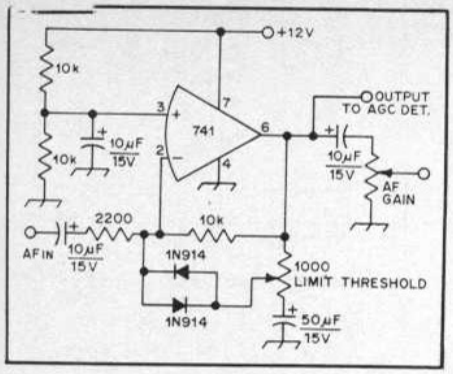


Fig. 54 — Audio limiter for use with af types of agc loops.

Diode CR1 serves as the main agc detector, with the following amplifier being adjusted to drive MC1350P or MC1590G amplifiers. The system could be adapted for the reverse agc of the CA3028A, or for virtually any i-f characteristic.

The action of the two loops is explained by considering sequentially how the circuit behaves. First, consider the effect of a short pulse of noise. This pulse will produce a lengthened response at the output of the i-f filter, which is detected ultimately by CR1 to cause a momentary reduction of the i-f gain. Audio output will result in the receiver and will also cause a signal to appear at CR2 and CR3. Because of the 100-kΩ resistor in series with CR2, C2 will acquire a small charge from this pulse. As a result, the main memory capacitor, C1, will discharge quickly through R3 and the drain of Q3.

On the other hand, consider the effect of a carrier, a string of cw characters, or a ssb signal. CR1 will again charge C1, and will lead to a gain reduction in the i-f system. The sustained audio signal that results will cause CR2 and CR3 to operate and charge capacitor C2 negatively. The gain of the op amp driving these diodes is adjusted so that normal signals cause the op-amp output to swing over its full range, from ground to the positive supply. This will cause C2 to be charged to a high negative voltage. The value will be approximately twice the supply voltage at U2. In this condition Q3 is pinched off. Because of this the only discharge path for the main memory capacitor, C1, is through R1, a 22-megohm resistor. When the signal disappears, C2 begins to discharge through R2. When the voltage at the gate of Q3 becomes close to ground, so the FET is no longer in a pinch-off condition, C1 is discharged quickly through Q3.

Listening to a system of this kind is enlightening after being accustomed to the simpler methods. With the full hang agc, the receiver is virtually silent after a strong signal disappears from the passband. However, after a timing period associated with the C2-R2 timing network, the receiver returns to full gain within roughly 50 milliseconds. The time delay is virtually independent of the strength of the incoming signal.

An audio signal is suitable for driving

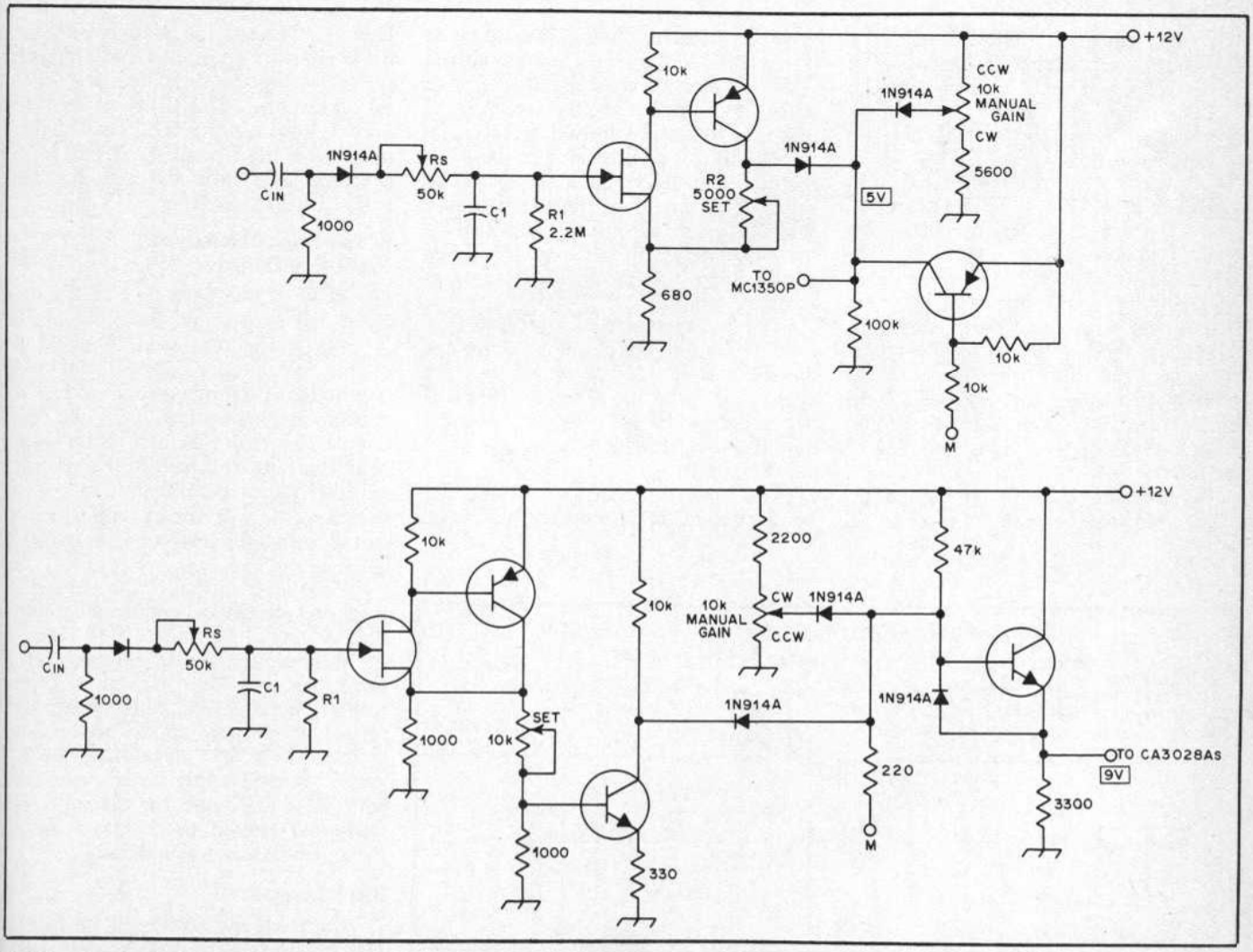


Fig. 55 — Circuits for audio or i-f derived agc which can be interfaced with IC i-f amplifiers.

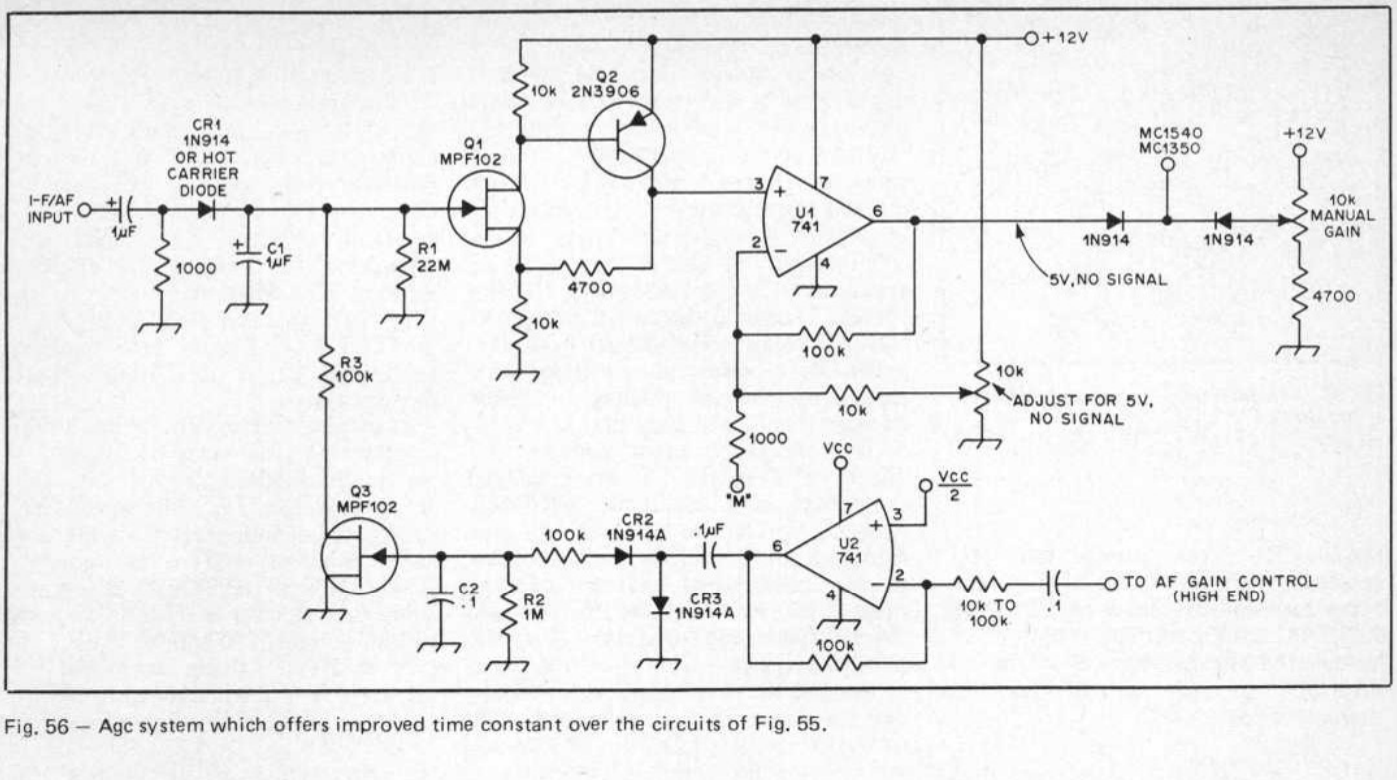


Fig. 56 – Agc system which offers improved time constant over the circuits of Fig. 55.

secondary detector, CR2, because a slow response is desired in this loop. An i-f derived signal could be used also. The 741 op amp, U2, would need to be replaced with a circuit suitable for the i-f frequency used.

An agc system of this kind is used in a receiver at W7ZOI. It will be described later. The agc characteristics have been studied extensively by means of a triggered oscilloscope. No sign of overshoot or pumping could be detected with signals ranging from the minimum detectable amount up to 50 mW at the antenna terminals. Higher levels would probably endanger the front-end components of the receiver. The signal to be detected was derived from a 9-MHz i-f amplifier.

With the agc systems outlined, additional gain may be required in order to drive the detector diode. This extra gain is usually minor with audio-derived systems, since the levels are already high when that part of the receiver is reached. With an i-f derived detector, 10 to 40 dB of additional gain is often required, depending upon the overall i-f gain. Care should be taken to ensure that the agc detector is not activated by the BFO energy. BFO energy should be confined to the product detector, as outlined earlier.

The agc threshold of a receiver (the level at the antenna terminal where agc action begins) is determined by the characteristics of the detector diode and the gain ahead of the detector. For most applications a suitable threshold is −100 to −110 dBm.

The "tightness" of an agc loop can be expressed in a number of ways. Usually, the variation in audio output in dB is given for an input variation from a few dB above threshold to a level 60 or 80 dB stronger. This figure of merit, no matter how it is defined, will depend mainly on the overall *fixed* gain in the agc loop and upon the agc characteristics of the i-f amplifier.

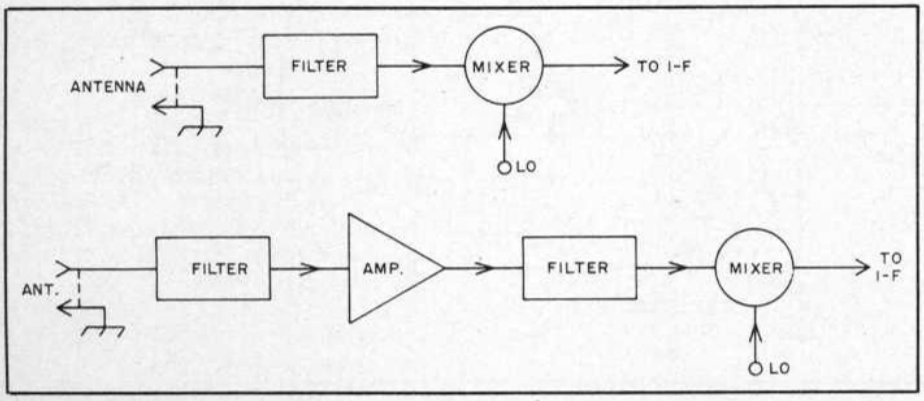


Fig. 57 – Block diagrams of receiver front end for single-conversion circuits.

Simple Superheterodyne Front-End Design

Of all of the parts in a receiver the front end is probably the most critical. A poor design can lead to disastrous results. A proper design will yield acceptable performance. This receiver section is so critical that we have devoted an entire chapter to its design. Special attention is paid to the problems of noise figure and dynamic range. The criterion for optimizing either is presented with a discussion of the tradeoffs between the two.

While not difficult, the subject of front-end design is complicated enough that it cannot be approached casually. In this section some information is presented for the beginning experimenter. Totally acceptable performance for general-purpose applications may be attained if a few precautions are followed. Some sample circuits are given with rules of thumb for their use. The reader is referred to chapter 6 and to the appendix for design details.

Block Diagram

The front-end section of a receiver is that portion containing the first mixer, preselection filters and perhaps an rf

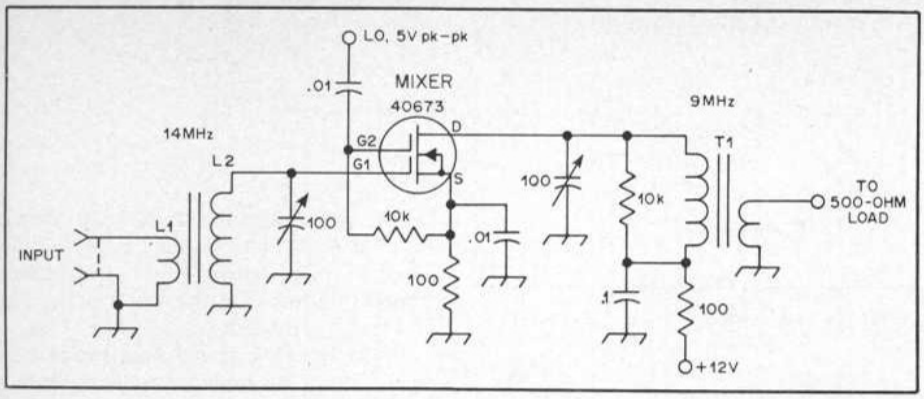


Fig. 58 — Circuit for a dual-gate MOSFET mixer.

amplifier. The standards that must be met are to provide sufficient receiver noise figure and image rejection. Gain is often desired, although not always necessary. Shown in Fig. 57 are block diagrams for the front end of single-conversion receivers.

The two systems differ only in the inclusion of an rf amplifier in the second. The first contains none. Both circuits have a preselector network and a mixer. The most tragic mistake made by the beginning experimenter is that he uses an rf amplifier when it is not really needed. The *only* purpose of an rf amplifier in a receiver front end is to reduce the overall noise figure. This will enhance the sensitivity of the receiver. However, on most of the lower frequency amateur bands an acceptable noise figure may be obtained with a mixer front end. The effect of the rf amplifier is to increase the signal levels at the mixer, causing a degradation in signal-handling ability.

A standard for evaluating a receiver for sufficiently low noise figure was presented at the beginning of this chapter. It bears repeating: When the antenna is connected to the receiver, the output noise should increase significantly. If this criterion is met there is no need to seek a lower noise figure. Generally speaking, the atmospheric and man-made noise levels from 1.8 to 21 MHz are high enough that an rf amplifier is redundant.

Image rejection must be maintained. Furthermore, it is wise to protect the receiver from signals other than those to which the receiver is tuned. In many receivers this front-end selectivity is provided with a single or a double-tuned circuit. The latter is preferred, owing to the improved skirt selectivity for a given 3-dB bandwidth. The design of simple preselector filters is covered in some of the sample circuits. The subject of loaded and unloaded Q was covered in chapter 2.

Mixer Circuits

There are a number of semiconductors that will function well as mixers. Of all that are available the simplest to use is the dual-gate MOSFET. A circuit is shown in Fig. 58. A single tuned circuit is used as the preselector. A tuned transformer at the output matches the crystal filter that follows the mixer.

The gain realized with this circuit will depend upon exact device parameters. Values of 15 dB are representative. The proper LO injection level for this mixer is 5 volts pk-pk. Lower levels will decrease gain and will compromise dynamic range. The noise figure of this front end is often 8 to 10 dB. This is low enough to ensure usable sensitivity in almost all hf applications.

The dual-gate MOSFET appears to present a very high impedance at its input (gate 1) in the hf region. Because of this, the tuned circuit is singly loaded. The loaded Q of the preselector is determined by the unloaded-Q value of the inductor and the loading presented by the 50-ohm antenna.

The values shown in Fig. 58 are for an input on the 20-meter band. The inductor has a Q_u of approximately 200 and consists of 20 turns on a toroidal form. The antenna link contains 2 turns. Because impedances transform according to the square of the turns ratio with toroidal cores, the equivalent resistance across the coil is 5000 ohms. The inductance is nominally 1.5 μH. The equivalent parallel resistance representing the unloaded Q is of the order of 27 kΩ. Since this value is large when compared to the 5000 ohms representing the antenna loading, the losses in the circuit will be small. The loaded Q will be 5000 $(2\pi fL)$ = 37.4. (See chapter 2 for details.) The 3-dB bandwidth of this circuit will be 14,000/37.4 = 374 kHz. No tuning would be required for the complete 20-meter band. It would be needed for the lower bands.

If a higher loaded Q was desired in the preselector, it could be obtained by changing the turns ratio. For example, the link could be reduced to a single turn. This would produce a Q_L value of 85. The value might be higher. This is because with only 1 turn for the antenna link, the coupling may become weak enough that the turns squared relationship no longer applies. A loaded Q of 85 would imply a bandwidth of 165 kHz. It may be shown that the insertion loss of the filter will now be much higher (nearly 10 dB), which would degrade noise figure. This is not desired.

An additional problem with the higher turns ratio configuration is the higher signal voltage appearing at the input of the MOSFET. This could compromise dynamic range. A lower voltage at the input may be realized by tapping the gate down on the tuned circuit. This will not alter the loaded Q of the preselector, nor will it reduce insertion loss. The tap may be on the coil, or it may be composed of tapped capacitors.

The method of capacitive matching is shown in Fig. 59 where it is applied to matching of the antenna. If the antenna resistance is R_a (usually 50 ohms) and the equivalent resistance presented

Fig. 59 — Method for capacitive matching at the input of a receiver.

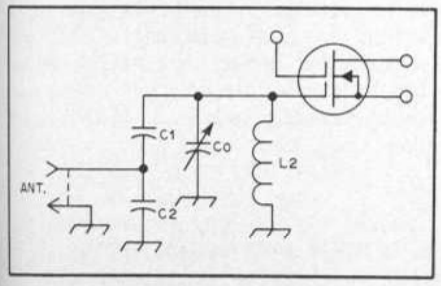


Fig. 60 — Example of capacitive-divider matching to decrease the impedance level at the gate of a MOSFET.

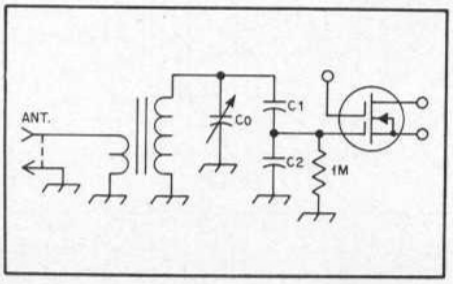
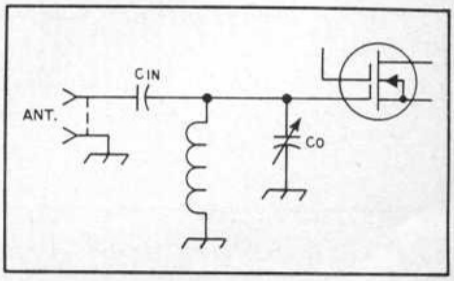


Fig. 61 — Method for using a single series capacitor at the receiver input to match a low-impedance antenna system to the input stage.



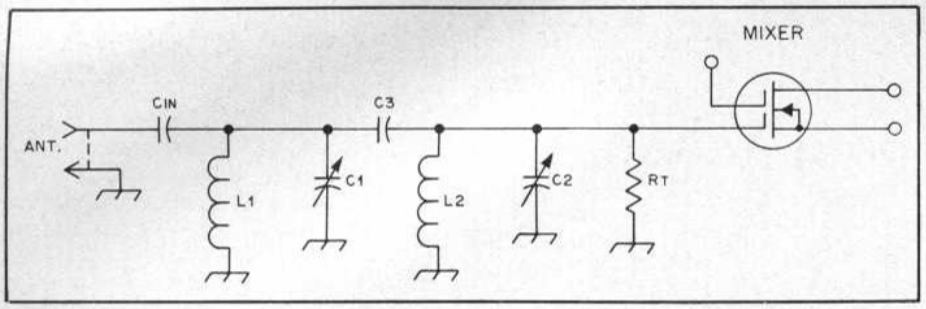


Fig. 62 – Example of a double-tuned front end circuit. See text for R_T.

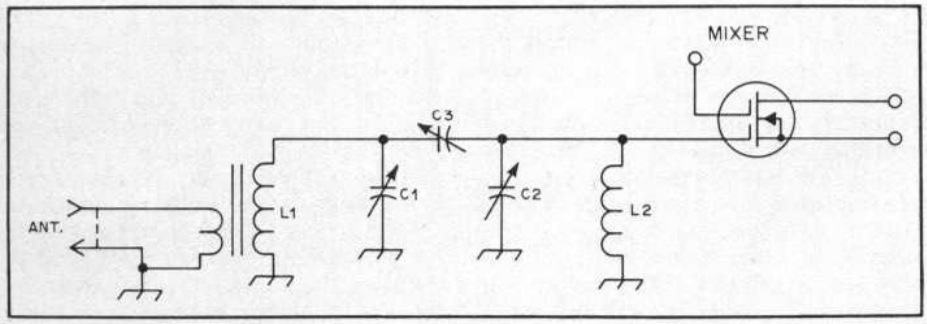


Fig. 63 – A singly terminated double-tuned input circuit.

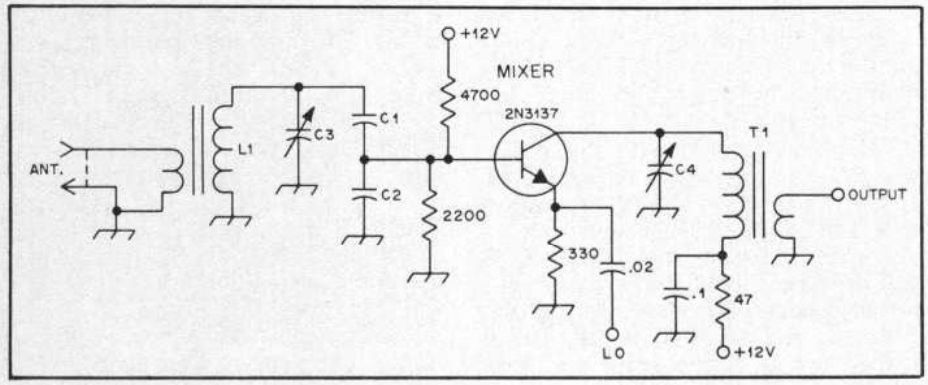


Fig. 64 – Bipolar-transistor mixer with LO-energy injected at the emitter.

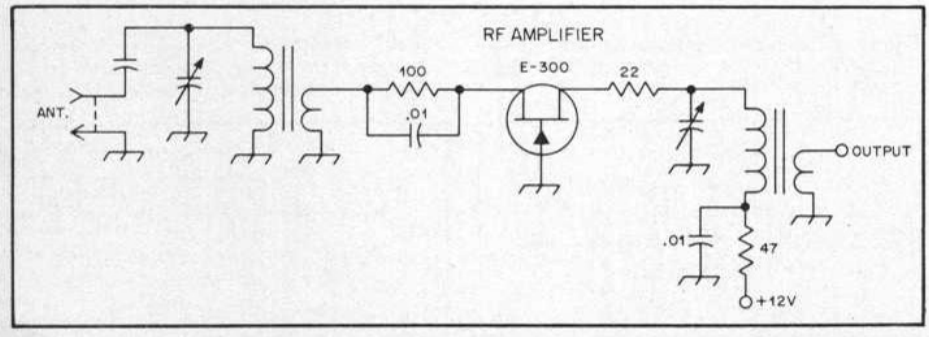


Fig. 65 – Circuit of a common-gate JFET rf amplifier.

across the coil is R_c, the two are related with

$$R_c = \frac{R_a}{(1 + \frac{C1}{C2})^2} \qquad \text{(Eq. 6)}$$

Using this equation, it may be shown that a 9:1 capacitance ratio would produce the same 100:1 impedance transformation that the link on the coil of Fig. 58 afforded.

If a capacitive transformation is used to decrease the impedance level driving the gate of the MOSFET (Fig. 60), care should be used. A resistor would be required from the gate to ground to establish a proper dc bias. This resistor should be very large in ohmic value. Otherwise, it might load the coil excessively. In a single tuned circuit, the loading should come from the antenna and not from extra resistors that are added.

A third method for matching into the resonator would be to use a low-value capacitor directly between the antenna terminal and the "hot" end of the tuned circuit. This is shown in Fig. 61. The equations for applying this method are examined in the appendix in connection with the filter tables.

In the mixer circuit of Fig. 58, a tuned transformer was used to match between the drain of the MOSFET and the crystal filter that follows. With almost all MOSFETs that are used in mixer applications, the output impedance is very high. Values of 100 kΩ or more are representative. If the transformer were designed to match between this level and the 500-ohm input to a filter (symbolic of the KVG line of 9-MHz crystal filters), the dynamic range of the mixer would be compromised severely. It is mandatory that a resistance be placed across the coil. This ohmic unit establishes a well-defined termination for the filter and limits the impedance presented to the drain of the mixer.

In the circuit of Fig. 58, the drain transformer has a 30:7 turns ratio. This causes the 10-kΩ resistor to appear as a 500-ohm termination for the filter. An equally viable (and often desirable) circuit for the output would be a pi network. It should be designed for a Q of 10.

The single tuned circuits that have been used for preselection are often lacking in skirt selectivity. This will compromise image rejection. A better circuit is a double or triple tuned one. Shown in Fig. 62 is a double-tuned front end. Again, only a mixer is used. No constants are given, since they will depend upon the band of interest. Specific designs are presented in the filter tables of the appendix.

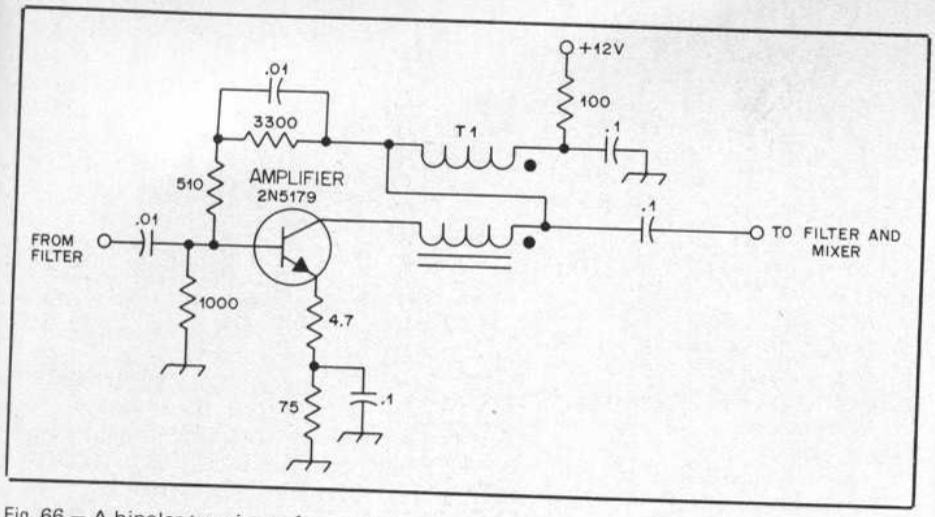


Fig. 66 — A bipolar-transistor rf amplifier.

A resistor is shown at the output of the preselectors, from the gate of the MOSFET mixer to ground. This resistor is necessary to terminate the filter properly. These filters are classed "doubly terminated," and are representative of the filters in the appendix. It is not necessary that double-tuned circuits be doubly terminated. Suitable circuits may be realized with antenna loading as the only termination. See Fig. 63. This will alter the designs from those given in the appendix. The best approach for using such filters is empirical. The coupling capacitor (C3) should be variable. Initially, it should be adjusted for minimum capacitance. The resonators are then peaked (C1, C2). The input is swept to ensure that a single response is provided. Then, coupling capacitor C3 is increased slightly, and C1 and C2 are peaked again.

This procedure is repeated until a double-humped type of response appears. The coupling-capacitor value is then decreased slightly and left in that way. If the bandwidth obtained with this course is too narrow, the loading at the antenna terminal may be increased (more turns on the link). The process is repeated until the desired bandwidth is obtained. The builder should use the filters in the appendix as a guideline for the approximate values to begin with in his (or her) empirical realization of a singly terminated filter. It is not recommended that three (or more) filter sections be attempted unless each end of the filter is terminated properly.

While we have strongly recommended the dual-gate MOSFET mixer, there are other devices that will perform suitably for such applications. These include many ICs which were discussed in the product-detector section. Bipolar transistors will also perform as mixers. A typical circuit is shown in Fig. 64. The LO is injected onto the emitter of the mixer. Best performance will be obtained from this circuit if large dc bias currents are used. Bipolar mixers are not recommended.

Some of the ICs that are used as mixers are the MC1496G and CA3028A. They have the advantage of balance. This reduces the amount of LO power that might appear at the antenna terminal. These devices are usually more subject to overload effects than the MOSFET is. A receiver described later in this chapter shows an application of a CA3028A mixer.

RF Amplifiers

It is sometimes desirable to use an rf amplifier ahead of a mixer. Special applications where inclusion could be

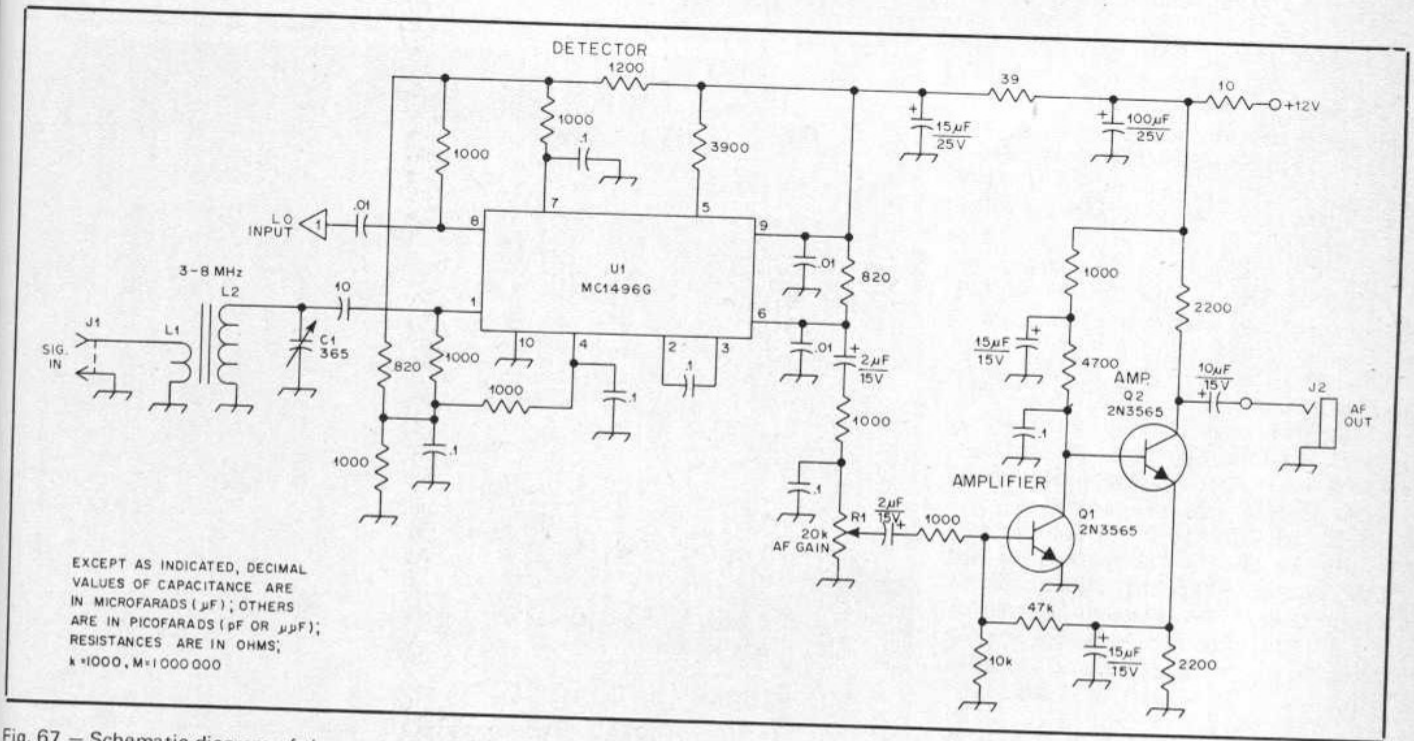


Fig. 67 — Schematic diagram of the product detector and audio amplifier. Fixed-value capacitors are disk ceramic except those with polarity marked, which are electrolytic. Fixed-value resistors are 1/2-W composition.
C1 — Miniature 365-pF variable.
J1 — Antenna receptable of builder's choice.
J2 — Two-circuit phone jack.
L1 — Three turns No. 24 enam. wire on Amidon T68-2 toroid core.
L2 — 40 turns No. 28 enam. wire on L1.
Q1, Q2 — Npn transistor, 2N3565 or equiv.
R1 — 20,000-ohm audio-taper carbon control.
U1 — Motorola MC1496G IC.



Front panel of the direct-conversion receiver

desirable would be for 10-meter or vhf reception. Alternatively, they might be included in the front end of portable receivers to be used in isolated locations which are devoid of man-made noise. These locations *do* exist.

Shown in Fig. 65 is the circuit of a simple rf amplifier that is recommended for general-purpose applications. A JFET is employed in the common-gate configuration. This circuit will provide a gain of 8 to 14 dB, depending upon the FET characteristics. The input impedance at the source will be low. Representative, values are from 100 to 300 ohms. The output should be a tuned circuit with a high *L-C* ratio. This maximizes the impedance presented to the drain, increasing the gain. The resistor in the drain suppresses uhf, vhf and parasitic oscillations. The general impression that common-gate FET amplifiers are unconditionally stable is not true.

Shown in Fig. 66 is a circuit for a bipolar transistor rf amplifier. A common thought among amateurs is that bipolar transistors are not suitable for front-end applications because of overload. This is not absolutely true. If low-noise transistors with high values of f_T are used in circuits with negative feedback, excellent performance may be obtained. The circuit shown is not subject to easy overloading. This results from the feedback and high bias current (20 mA). The input and output impedances are both close to 50 ohms. This makes the circuit easily matched to filters from the appendix. Bipolar transistors are not recommended unless these precautions are heeded.

The amplifier of Fig. 66 has a gain of nearly 20 dB. The noise figure is not low, but is reasonable. One representative sample investigated showed a 6.5-dB value. The bandwidth is over 100 MHz, making the circuit useful for all hf bands. The extensive feedback does ensure stability.

A Two-Band Direct-Conversion Receiver

There is often a need for a simple receiver which still offers good performance. An example would be a compact receiver for portable or emergency operation. Another might be a club project where a number of beginners build a "first station." The receiver described in this section is aimed at these applications.

The detector and audio circuit is shown in Fig. 67. An MC1496G IC is used as the product detector. Ample audio gain is provided by a pair of transistors. In the interest of simplicity, minimum audio selectivity is used in the system. However, an *R-C* active filter could be added at the audio output, if desired.

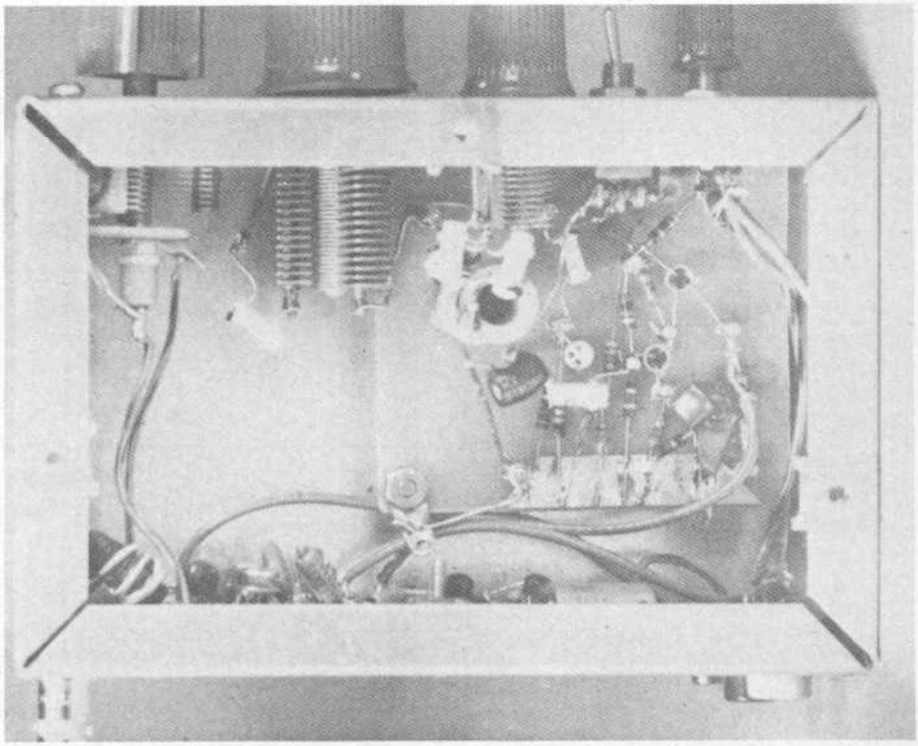
The detector differs from that normally used with this IC. First, the gain is increased significantly by placing a bypass capacitor between pins 2 and 3 of the chip. The more typical application is with a resistor (100 to 1000 ohms) in this position. The other departure from the standard circuit concerns the bias current used. This is determined by the resistor connected between pin 5 and the positive supply. The usual 10-kΩ resistor has been replaced by a 3300-ohm one. This increases the gain and signal-handling capability of the detector by about 10 dB.

The input circuit will tune from approximately 3 to 8 MHz. This allows the 80- and 40-meter amateur bands to be tuned without band switching the front end. Other tuned circuits may be substituted in order to cover additional frequencies. A 10-pF capacitor is used between the tuned circuit and pin 1 of the IC. For operation on the 160-meter band, a suitable value would be 22 pF. For operation at 14 or 21 MHz, the value should be decreased to 5 pF.

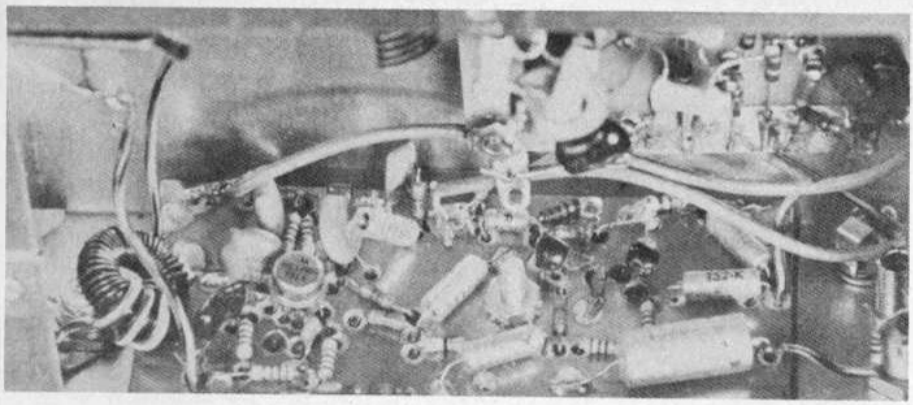
A wide range oscillator is shown in Fig. 68. A JFET is employed in a Hartley circuit. A buffer/amplifier with two bipolar transistors is used to obtain ample BFO drive voltage. A 3/8-inch diameter slug-tuned coil is used with parallel capacitors (air variable and ceramic NP0) to form the resonator. With the band switch (an inexpensive toggle-type) open, the oscillator tunes from 6 to 8 MHz. When the switch is closed, a 360-pF silver-mica capacitor is paralleled with the others, providing a tuning range of 350 kHz in the 80-meter band. The exact range desired may be obtained by adjustment of the coil slug.

An experiment was performed to move the oscillator higher in frequency. The slug was removed from the coil and all fixed-value capacitors were disconnected. In this condition, the oscillator would tune to about 15 MHz. The stability was adequate for reception of cw and ssb signals.

A pc layout is shown in Fig. 69 for the detector and audio board. The size is approximately 2 X 4 inches. The experienced builder may wish to miniaturize the circuit further. But, the begin-



Inside view of the direct-conversion receiver. The antenna trimmer is at the left. Seen at the bottom of the box is the oscillator board. A strip of flashing copper serves as a ground bus.



The main board for the receiver. The input tuned circuit is at the left, adjacent to the product-detector IC. An audio amplifier is contained on the remainder of the board.

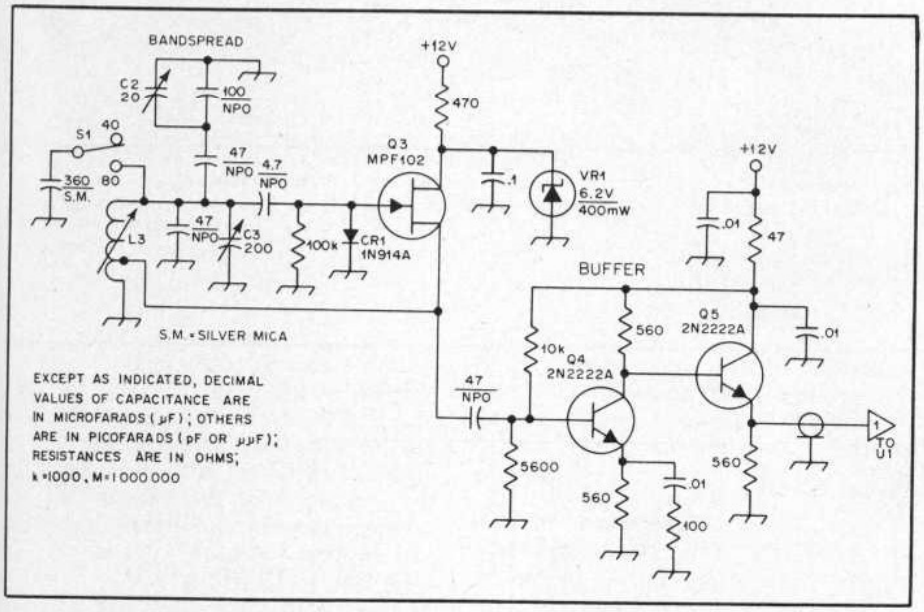


Fig. 68 — Schematic diagram of the tunable oscillator for the receiver of Fig. 67. Fixed-value capacitors are disk ceramic unless otherwise indicated. Fixed-value resistors are 1/2-W composition.

C2 — Miniature 20-pF air variable.
C3 — 200-pF mica trimmer.
CR1 — High-speed silicon diode, 1N914A or equiv.
L3 — 20 turns No. 24 enam. wire on 3/8-in. dia. ceramic slug-tuned form (Miller 4400-2 form), tapped 5 turns from ground.
Q3 — JFET, MPF102, HEP802, or TIS-88 suitable.
S1 — Spdt miniature toggle.
VR1 — 6.2-V, 400-mW Zener diode.

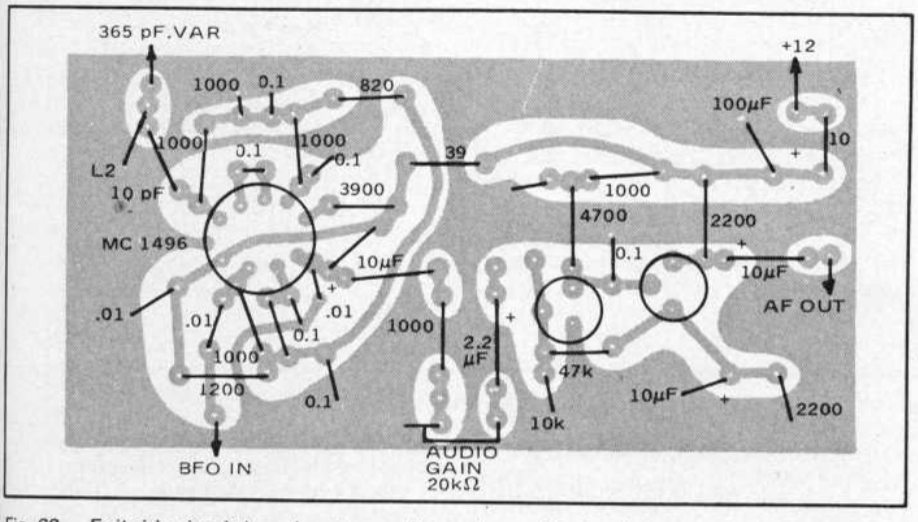


Fig. 69 — Foil-side circuit board pattern and parts layout for the detector and audio circuit of Fig. 67. Drawing is to scale.

ner may find it desirable to expand the size especially if small components are not available. The existing layout will be cramped unless rather small 0.1-μF capacitors are used.

The VFO is built on a 3 X 3 inch piece of unclad circuit board with rivet-in terminals for solder connections. (A board could be etched for this circuit.)

The two-band receiver is packaged in a 2 X 4 X 6 inch chassis. No vernier drive mechanism was used. Instead, two tuning capacitors are used in parallel. One functions as the main tuning while the other serves as a bandspread control. The advantage is one of mechanical simplicity, allowing quick completion of the project. Accurate calibration is not easily realized with this method.

The results obtained with this receiver were gratifying. Unlike some projects, this receiver functioned as designed when power was applied. Cw and ssb quality are excellent.

This receiver might serve as a step toward construction of a simple superhet. After being built as shown, a crystal filter could be added. The VFO can be moved easily to any frequency in the 3- to 15-MHz range, as outlined earlier. The addition of a dual-gate MOSFET mixer and a crystal-controlled BFO would result in a superheterodyne system (see Fig. 70).

The builder might want to add an rf amplifier, especially if the receiver is to be used on one of the higher bands. A suitable circuit using a 2N5179 is shown in Fig. 71. For dynamic-range reasons, one might scowl at the use of a bipolar transistor instead of an FET. However, this opinion is not valid.

The amplifier shown is broadband, has 50-ohm input and output impedances, and provides nearly 20 dB of gain. The use of heavy feedback ensures stability. Good signal-handling ability results from a high bias current (20 mA). The input preselector networks are in the appendix at the end of the book.

A Pocket-Size Direct-Conversion Receiver for 40 Meters

Solid-state technology permits miniaturization and low power consumption. The receiver of Fig. 72 was built to take advantage of both assets, while offering simplicity of construction.

The pocket portable uses two transistors and two ICs. Power is provided by a small battery contained in the 1 X 3-1/2 X 5-1/2-inch aluminum cabinet. The receiver is built on a 2-1/2 X 3-1/2 inch double-sided pc board (one side is all ground foil). Only 11 mA of current are required from the 9-volt battery.

The 40-meter cw band was chosen. The receiver could be adapted to any of the bands from 1.8 through 14 MHz.

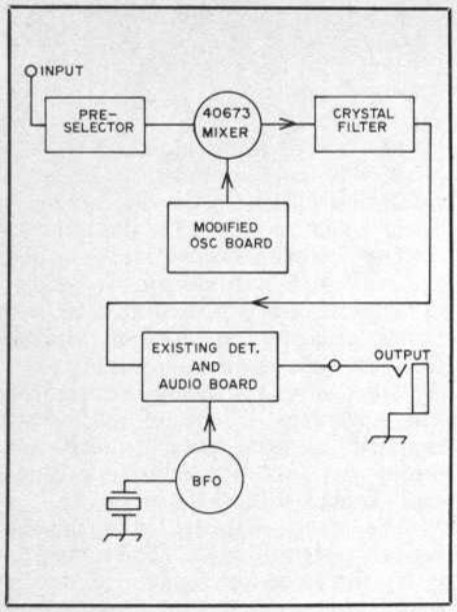


Fig. 70 — Details of how a mixer and BFO can be added to obtain a superheterodyne receiver with the circuit of Fig. 56.

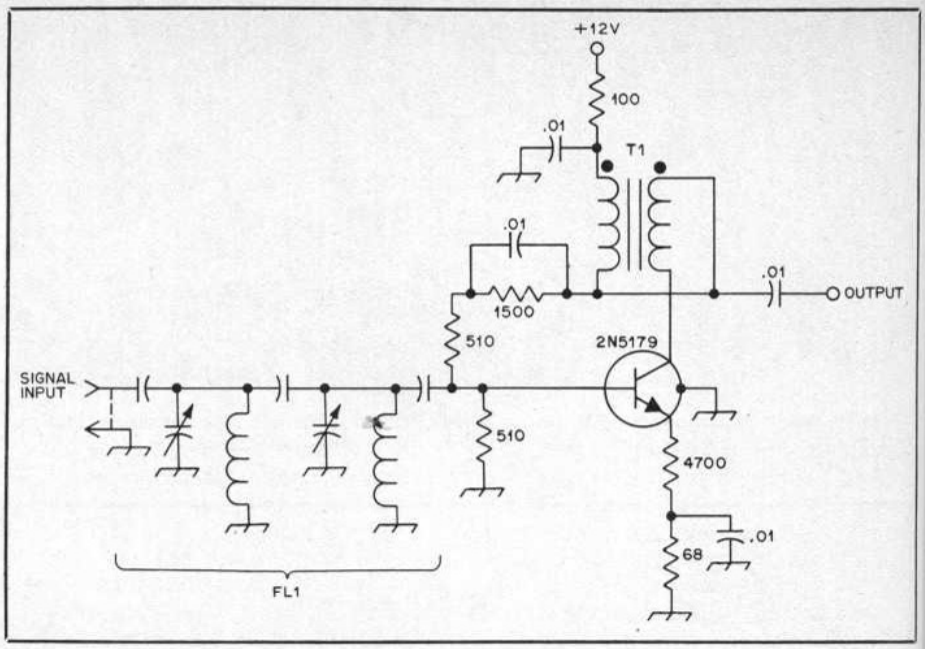


Fig. 71—Suggested r-f amplifier for use with the universal direct-conversion receiver.

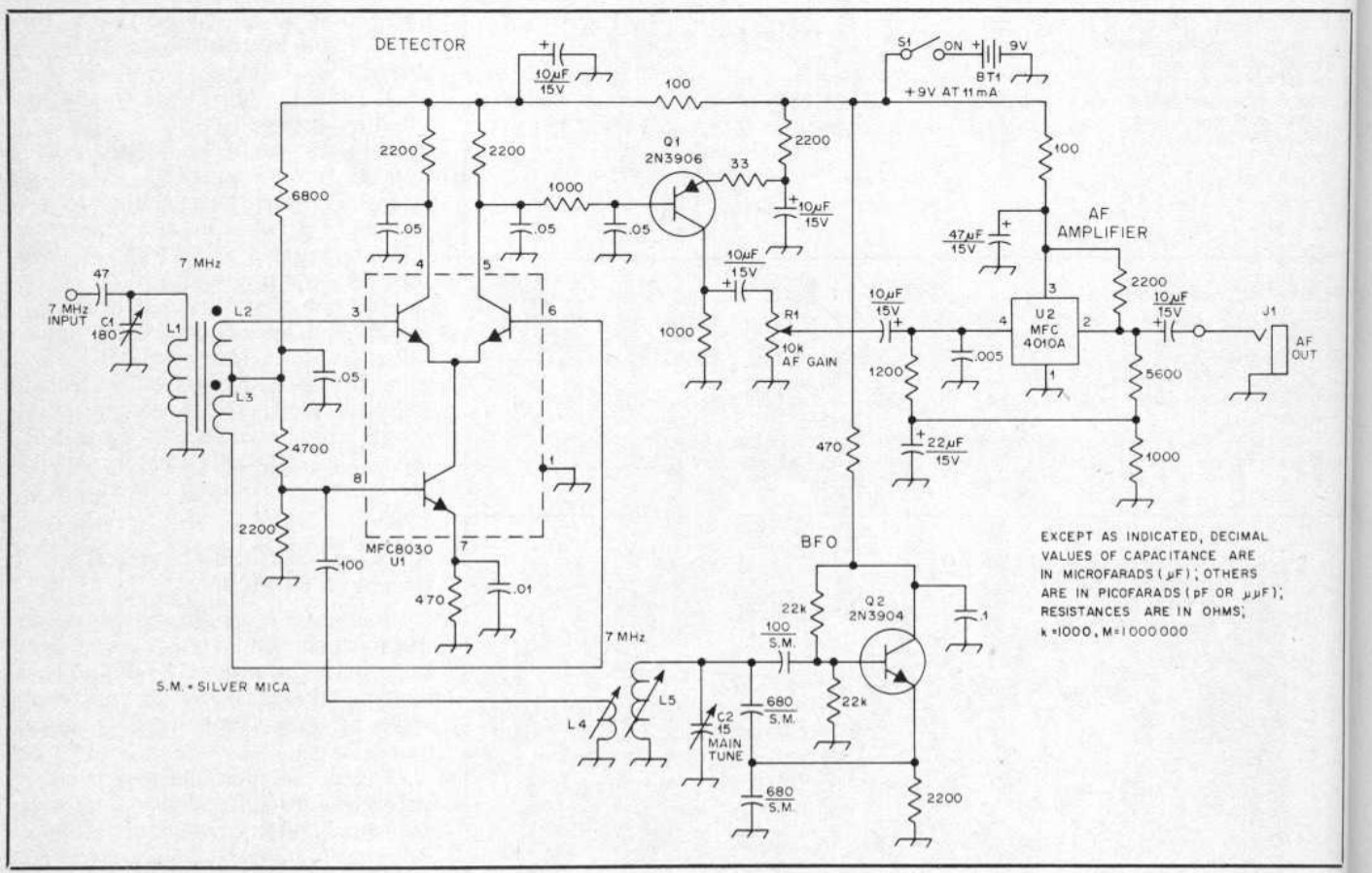
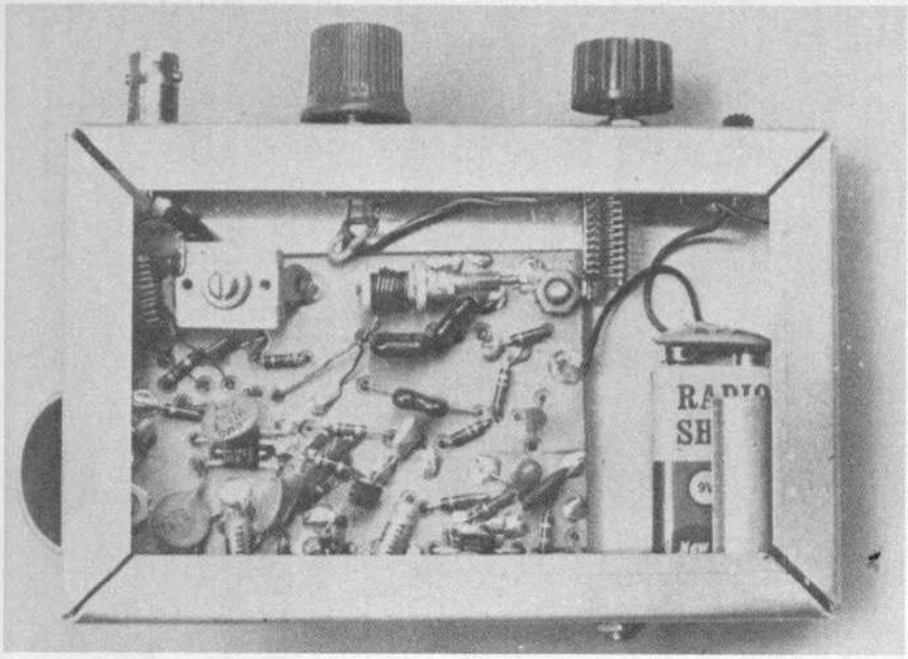


Fig. 72 — Schematic diagram of the pocket portable receiver. Fixed-value capacitors are disk ceramic unless otherwise noted. Polarized capacitors are disk ceramic. Fixed-value resistors are 1/2-Watt composition.

BT1 — Small 9-volt transistor-radio battery.
C1 — Miniature 180-pF trimmer (mica compression type).
C2 — Miniature 15-pF variable.
J1 — Two-circuit phone jack.
L1 — 30 turns No. 28 enam. wire on Amidon T50-2 toroid core.
L2, L3 — 5 turns each of No. 28 enam. wire over L1.
L4 — 4 turns No. 28 enam. wire over ground end of L5.
L5 — 15 turns No. 28 enam., closewound on 1/4-in. dia. ceramic slug-tuned form (Miller 4500-2 form). Inductance — 1.5 µH approx.
R1 — 10,000-ohm audio-taper carbon control.
U1, U2 — Motorola IC.



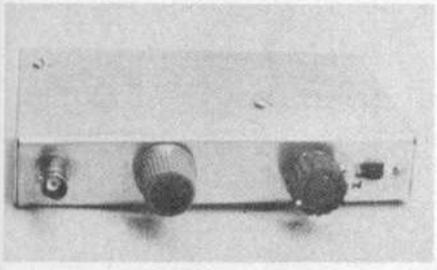
Inside layout of the receiver. All of the circuit is on a single pc board. The slug-tuned coil is part of the local oscillator.

The product detector uses a Motorola MFC8030 differential-amplifier. This IC is similar to the CA3028A, except that external biasing resistors are required. This adds to the parts count, but allows the IC to be biased for minimum current – a major design goal.

The detector output is applied to a 2N3906 pnp amplifier. This is routed through the audio-gain control to an MFC4010A. This tiny four-terminal IC is barely larger than a plastic transistor. It contains three direct-coupled stages.

The VFO uses a bipolar transistor in a Colpitts circuit. For minimum power-supply current, no Zener-diode regulation is employed. A ceramic slug-tuned coil is used with an output link to drive the detector. The stability is adequate.

In spite of simplicity the receiver performs well. Sensitivity is good. Signals from four continents were heard (on cw) during the first evening of use. Selectivity is poor, but could be improved with audio filtering. Because miniature projects like this one are dependent upon the size of the components available, no pc layout pattern is offered.



External view of the 7-MHz portable receiver. The controls are, left to right, af gain, tuning, and on-off switch.

A Simple Superhet for 80 and 40 Meters

In the 1950s nearly every issue of the *Handbook* contained a receiver which covered 80 and 40 meters. The basis of the design was a superheterodyne utilizing single conversion with an i-f of 1.7 MHz. The oscillator tuned from 5.2 to 5.7 MHz. With this set of frequencies, one band was the image of the other. This led to simplification, because band changing was realized by tuning the front-end preselector.

Shown in Fig. 73 is a solid-state version of the *Handbook* classic. This receiver was built by Jeff Damm, WA7MLH.

Only eight semiconductors are used in the receiver. Three dual-gate MOSFETs serve as the input mixer, i-f amplifier, and product detector. The *rest* of the functions are provided by means of bipolar transistors. Selectivity is obtained with a homemade two-crystal filter of the half-lattice type.

Circuit Details

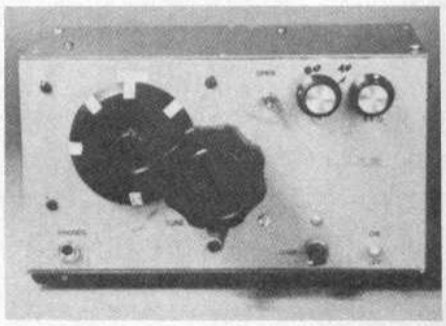
The input mixer uses a 40673 MOSFET with a single tuned circuit as the preselector. A half-wave filter is included in the antenna line to suppress spurious responses from high-order products created in the mixer. The filter is cut for a 7-MHz center frequency. The low-pass nature of the filter allows 80-meter signals to pass unattenuated. A short piece of coaxial cable is used to connect the panel-mounted variable capacitor in the preselector to the circuit board.

The drain of the mixer feeds the tuned primary of the transformer section of the crystal filter. The secondary is a center-tapped 12-turn winding. To ensure good balance, this winding is wound as six bifilar turns. The crystals were ordered for 1700.0 and 1700.3 MHZ. To keep the cost down, a .01-percent tolerance was specified. When the crystals arrived, their separation was only 200 Hz. While each crystal was within the manufacturer's specification, the bandwidth was narrower than desired. If the receiver is to be used for the reception of ssb as well as cw, a separation of 1.5 kHz is recommended. With the existing filter, cw selectivity is impressive. Single-sideband stations can be copied, but the audio sounds distorted.

A 10-kΩ resistor is used to terminate the filter. This value was arrived at experimentally. It assured minimum filter loss without passband ripple. Other values may be required, depending on the crystal characteristics.

A stage of i-f gain is provided by Q2, a dual-gate MOSFET. While the gain is not high, it is enough to overcome the loss of the crystal filter. Some variation of i-f gain is provided with a front-panel switch. In normal operation, gate 2 of Q2 is biased at about 4 volts. However, when the switch is closed, the bias on gate 2 is reduced to .0, This causes a decrease in stage gain of approximately 20 dB. In the unit built by WA7MLH, this switch is activated by pulling on the audio-gain control knob. The builder could use a separate switch.

A third 40673 MOSFET, Q5, is the product detector. This stage is typical of many using a FET, except that the bias for gate 2 (where the BFO is injected) is from a grounded resistor. The typical circuit has this resistor returned to the



Exterior of the 80- and 40-meter superheterodyne built by WA7MLH. The box measures 5 × 6 × 9 inches.

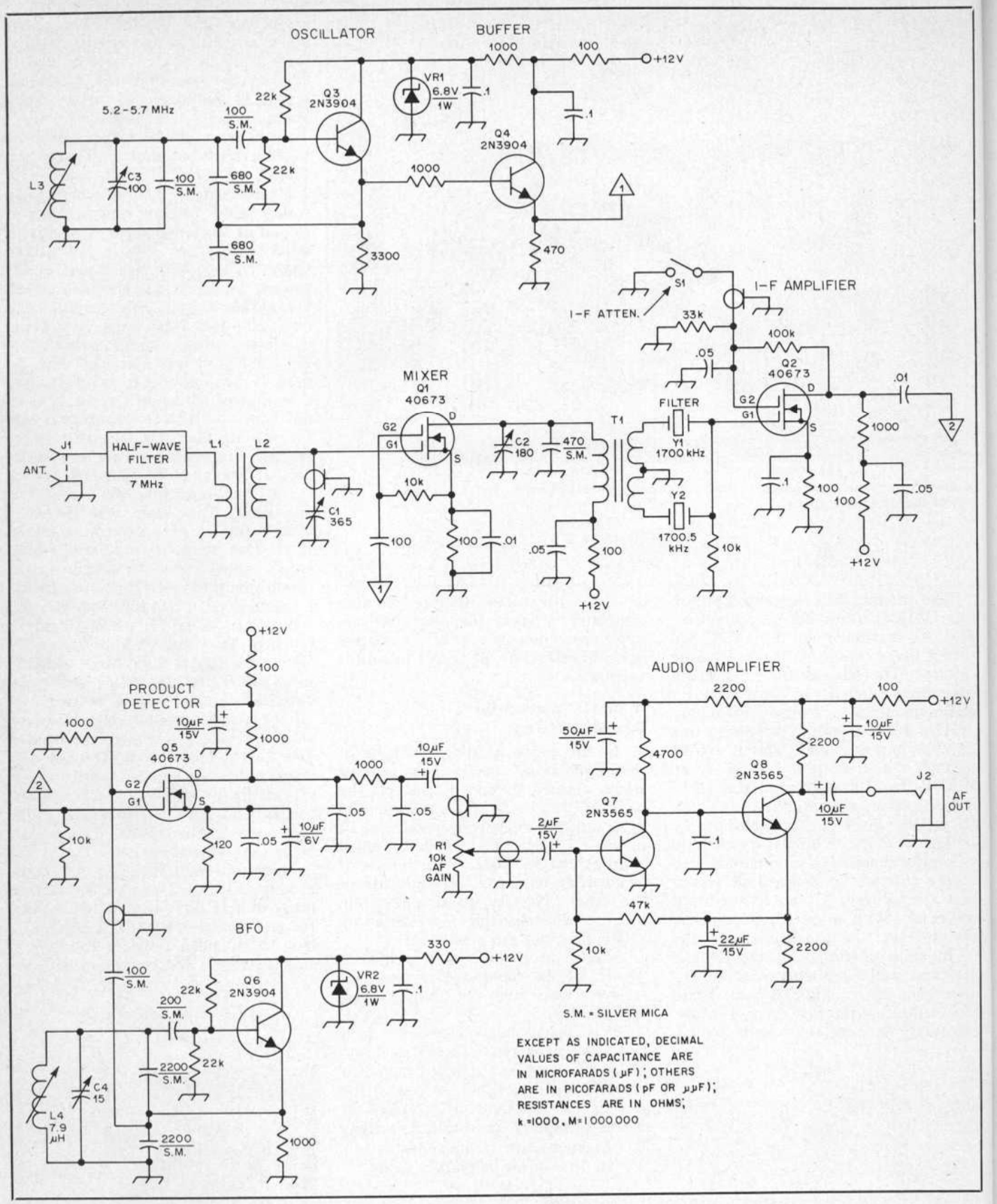
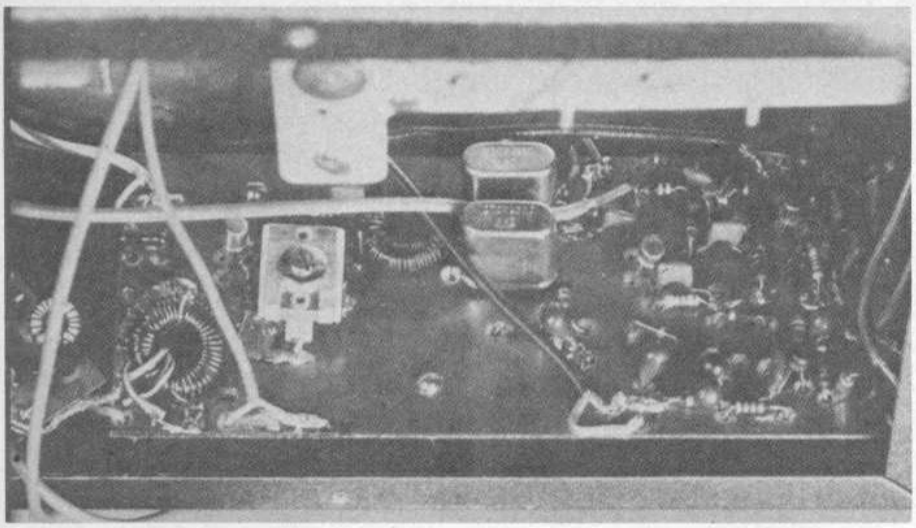
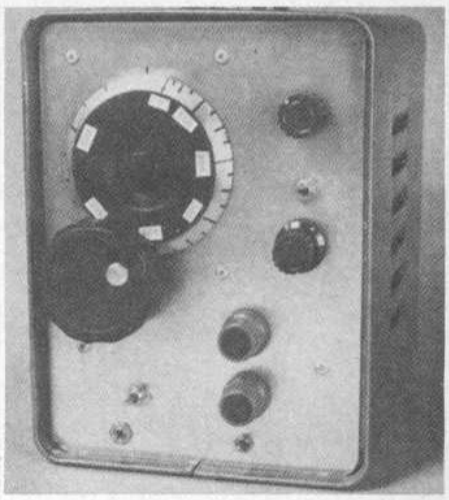


Fig. 73 — Schematic diagram of the 40- and 80-meter superheterodyne receiver. Fixed-value capacitors are disk ceramic unless otherwise noted. Fixed-value resistors are 1/2-Watt composition. Polarized capacitors are electrolytic.

C1 — Miniature 365-pF variable.
C2 — 180-pF mica trimmer.
C3 — 100-pF air variable. (See text.)
C4 — 15-pF variable. (See text.)
J1 — Antenna receptacle of builder's choice.
J2 — Two-circuit phone jack.
L1 — 3 turns No. 26 enam. wire over L2.
L2 — 36 turns No. 26 enam. wire on Amidon T68-2 toroid core.
L3 — Approximately 1.57-µH, slug-tuned coil (Miller 42A156CB1).
L4 — 7.9-µH slug-tuned coil (Miller 43105-CB1).
R1 — 10,000-ohm audio-taper carbon control.
S1 — spst toggle.
S1 — spst toggle.
T1 — Primary, 53 turns No. 28 enam. wire on an Amidon T68-2 toroid core; secondary, 12 bifilar turns.



Interior view of the WA7MLH receiver. The mixer front end is at the left, and the crystal filter is at the center of the board. At the right can be seen the product detector and audio section.



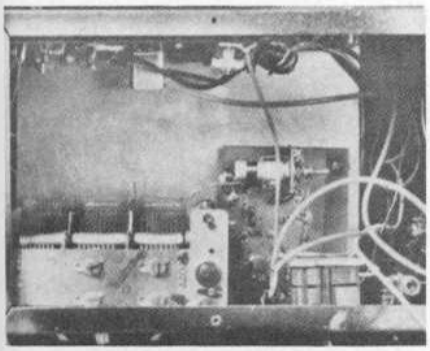
Front panel of the 80- and 20-meter superheterodyne receiver. Dial calibration is for the 20-meter band.

source of the 40673. The technique used led to a simplification.

Audio gain for the receiver is obtained from a pair of 2N3565s. Ample gain is provided for ear-shattering headphone output.

Both oscillators in the receiver use the standard Colpitts format. The main LO, which covers 5.2 to 5.7 MHz, is tuned with a single-section capacitor (C3) from a surplus BC-454. Any variable capacitor with a range of at least 100 pF will serve as well. With other capacitors, a vernier mechanism is recommended. It was not needed with the surplus capacitor since a high quality gear mechanism and dial drive are part of the capacitor unit.

While a commercially available coil was used for the LO tuned circuit, the inductor in the BFO was a junk-box item. A suitable substitute would be a J. W. Miller 43105CBI. The BFO is tunable from the front panel by means of a 15-pF variable capacitor.



Local-oscillator board of the WA7MLH receiver. The BFO is mounted on the right-hand wall of the box.

The receiver is constructed on three circuit boards. These may be seen in the photographs. The LO is built on a board that is mounted close to the tuning capacitor. The slug-tuned inductor is mounted on a scrap of pc board that is soldered to the main board. The BFO is on a second board which is located on one of the side walls of the cabinet. The remainder of the receiver is on a larger board that is affixed to the rear wall of the receiver.

All of the pc boards are double-sided, with one side serving as a ground plane. Coaxial cable (RG-174) is used for connections between boards and to the panel-mounted components.

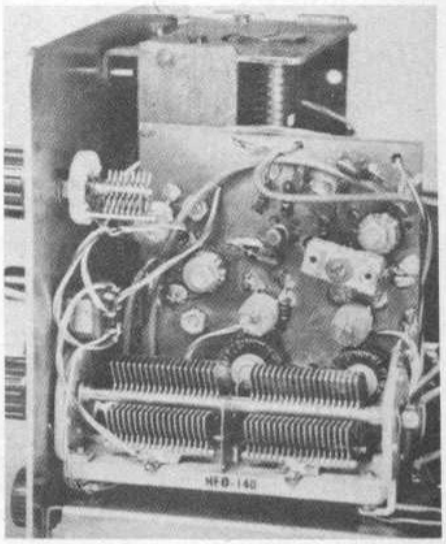
An aluminum plate is mounted to the bottom of the tuning capacitor. While this plate could serve as a chassis for some of the boards, its main function is to isolate the receiver from additional circuitry.

Considering its simplicity, this receiver performs very well. A signal of 0.1 μV from a well-shielded signal generator was copied easily, indicating more than ample sensitivity. The selectivity of the two-pole filter is quite respectable for cw operation, and the stability is compatible with the narrow bandwidth. No problems with overload or IMD products have been observed.

A Superhet for 80 and 20 Meters

There are a number of frequency schemes that lend themselves to simple two-band receivers. The previous superhet for 80 and 40 was one example. The unit shown in Fig. 74 is another. Here a 9-MHz i-f is combined with a 5- to 5.5-MHz LO in a receiver covering the 80- and 20-meter bands. Another that might be interesting would be an 80- and 15-meter design. A 12- to 13-MHz oscillator would provide full coverage of both bands with a 9-MHz i-f.

The front end of the 80/20 receiver uses a 40673 MOSFET mixer with no rf amplifier. Separate preselector networks are used for each band. A single-pole double-throw toggle switch is used to change bands at the output of the preselectors. Separate coaxial connectors are used at the input of each preselector, as the unit is used occasionally for 80-meter cw work, but was intended primarily as a tunable 14-MHz i-f system for use with vhf converters.



Front-end section of the 80- and 20-meter receiver. The circuit board is mounted on the VFO tuning capacitor. The dual-section variable capacitor tunes the 80-meter preselector. The small single-section variable is used with the 20-meter input circuit.

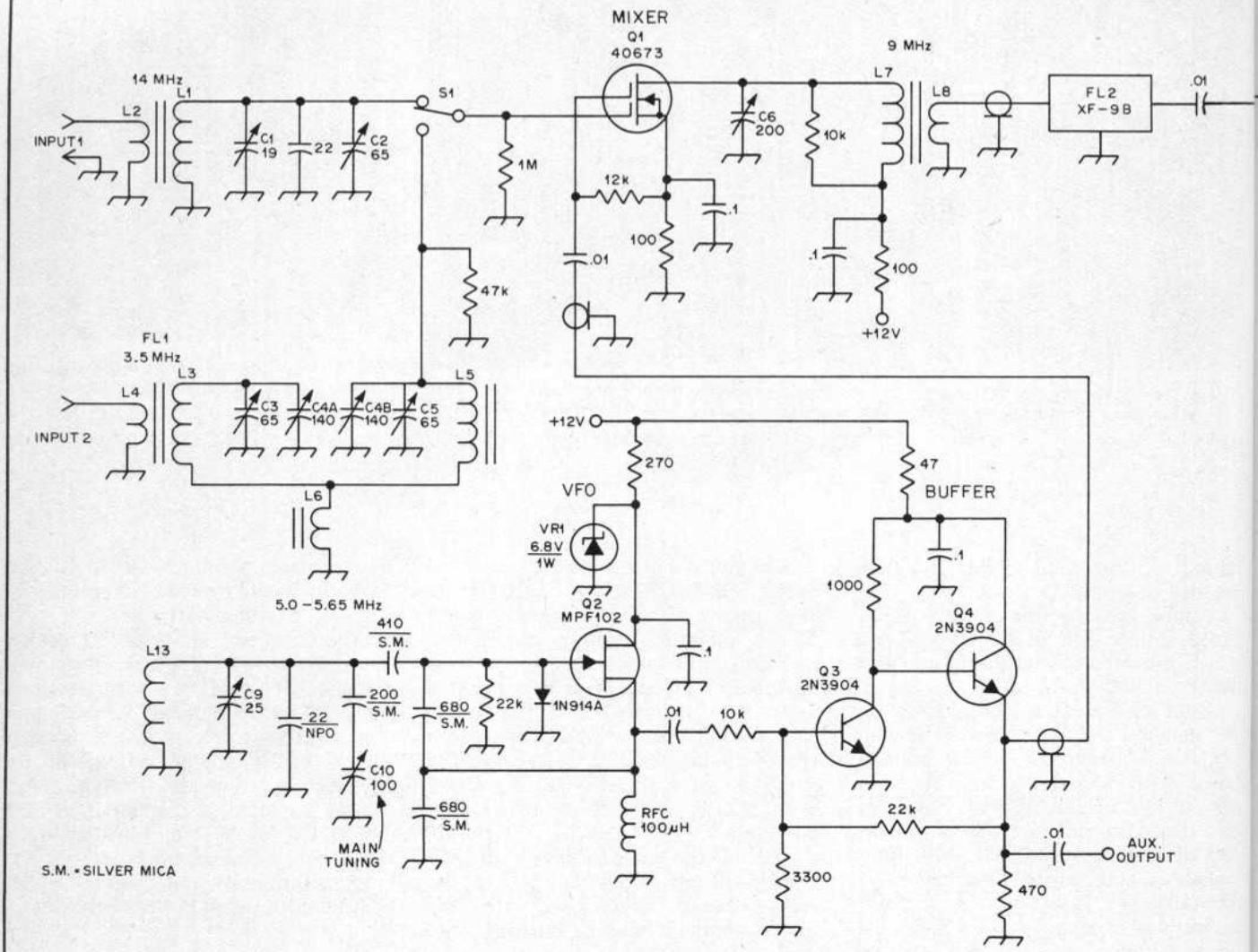
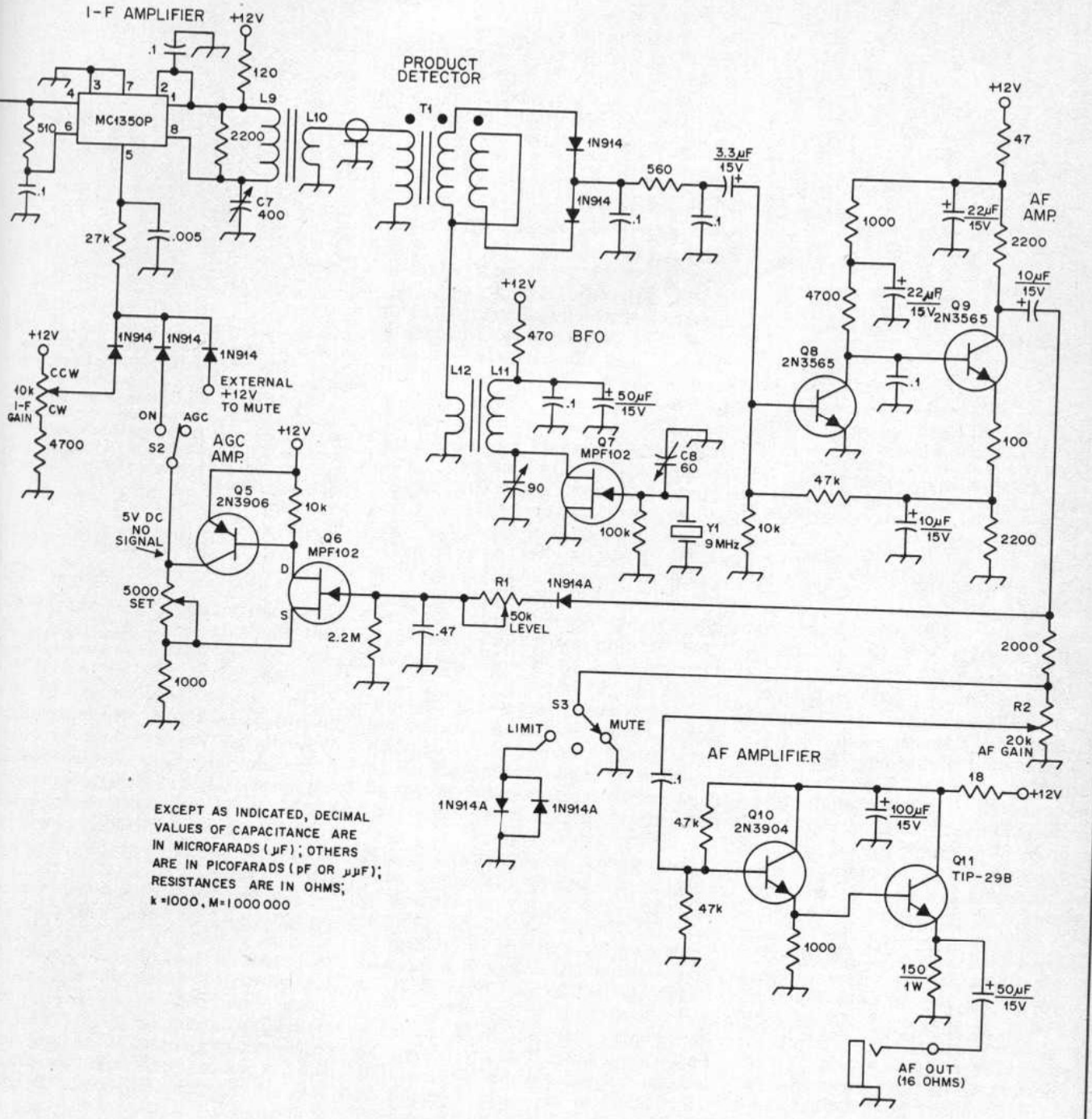


Fig. 74 — Schematic diagram of the 20- and 80-meter superheterodyne receiver. Fixed-value capacitors are disk ceramic unless noted. Fixed-value resistors are 1/2-Watt composition. Polarized capacitors are electrolytic. Numbered capacitors not listed below are trimmers.

C4 — Two-section 140-pF variable.
C10 — 100-pF variable (one section of BC-455 variable).
L1 — 25 turns No. 28 enam. on T37-6 toroid core, 1.87 μH.
L2 — 3 turns No. 28 enam. wire over L1.
L3, L5 — 44 turns No. 26 enam. wire on T68-2 toroid core, 10.8 μH.
L4 — 2 turns No. 26 enam. wire over L3.
L6 — 5 turns No. 28 enam. on T30-2 toroid core.
L7, L9 — 25 turns No. 28 enam. wire on T37-6 toroid core, 1.87 μH.
L8 — 6 turns No. 28 enam. wire over L7.
L10 — 4 turns No. 28 enam. wire over L9.
L11 — 40 turns No. 28 enam. wire on T37-6 toroid core, 4.8 μH.
L12 — 3.5-μH inductor on ceramic form (Miller 4505 coil with slug removed). Remove turns for desired tuning range.
L13 — 30 turns No. 26 enam. wire on an Amidon T50-6 toroid core, 3.5 μH.
R1 — Small 50,000-ohm carbon control.
R2 — 20,000-ohm audio-taper carbon control.
S1, S2 — Spdt toggle.
S3 — Single-pole, three-position miniature switch.
T1 — 12 trifilar turns No. 28 enam. wire on Amidon FT-37-61 ferrite toroid core, μ = 125.
VR1 — 6.8-volt, 1-watt Zener diode.
Y1 — 9-MHz crystal. International Crystal Co., type GP.



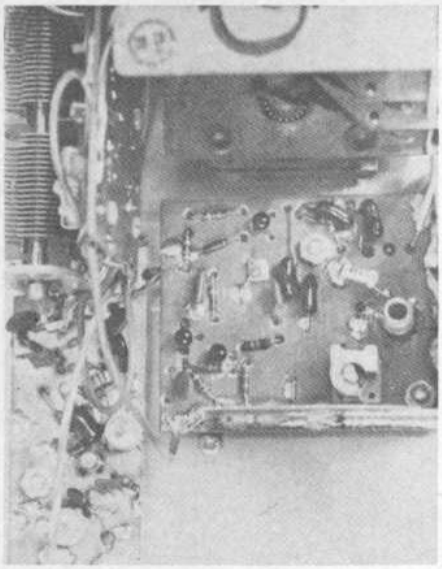
EXCEPT AS INDICATED, DECIMAL VALUES OF CAPACITANCE ARE IN MICROFARADS (µF); OTHERS ARE IN PICOFARADS (pF OR µµF); RESISTANCES ARE IN OHMS; k=1000, M=1000 000

Both preselector controls are brought to the front panel. A single tuned circuit is used at 14 MHz. For 80-meter operation, an adjustable double-tuned circuit was chosen. This filter was designed for a 50-kHz bandwidth and has a Butterworth response.

The local oscillator is a FET version of the Colpitts circuit. It is followed by a two-stage buffer amplifier using bipolar transistors. A surplus tuning capacitor from a BC455 is used for tuning. Only one section is employed.

A single-sideband type of crystal filter is used as the basis of the i-f strip. This is followed by an MC1350P IC i-f amplifier which supplies approximately 45 dB of gain, and over 65 dB of gain variation. The filter and the IC amplifier are mounted on a small double-sided pc board which is buried in the chassis.

The product detector uses two diodes. In spite of its simplicity, it performs well. The BFO employs a single transistor, and supplies +13 dBm of injection to the detector. A single crystal was used, limiting reception to upper sideband or cw. The builder might consider crystal switching if he wishes to copy lower sideband (pre-



Interior of the 80- and 20-meter receiver. The 5-MHz VFO is visible below the surplus tuning capacitor. The dual-section variable capacitor is part of the 80-meter preselector. At the lower left is a low-noise rf amplifier which is used in conjunction with some vhf converters.

dominant for the 75-meter band) and for the output of OSCAR 7 (Mode B) on 2 meters.

The major audio gain is provided by a pair of 2N3565s. The output of this amplifier is fed to the audio-gain control. The audio-output amplifier uses a Darlington emitter follower to drive low-impedance stereo headphones. This provides excellent audio quality for reception of vhf phone signals.

The high end of the audio-gain control is sampled in order to drive the agc detector. The agc amplifier consists of a JFET and a pnp transistor in a low-gain feedback pair. The control is adjusted for an output of 5 volts with no signal present at the agc detector. While the audio-derived agc system suffers from the problems typical of such circuits, it is adequate for most ssb work. If the receiver is used for other than casual cw work, the builder might consider an i-f derived agc detector.

A single-pole, double-throw, center-off type of toggle switch is mounted on the front panel. In the center position, the receiver functions normally. When thrown in one position, a pair of back-to-back silicon diodes is inserted as an audio limiter. This helps considerably in attenuating the automotive ignition noise encountered on six meters. In the other position, the switch shorts the audio output for muting purposes.

The photographs show a top view of the chassis. The LO board is mounted in back of the tuning capacitor while the front-end mixer is mounted on a vertical board which is bolted to the side of the tuning capacitor. The extra board contains a low-noise preamplifier for 14 MHz. This is used in conjunction with a diode-ring mixer for vhf reception. With the "preamp," the noise figure at 14 MHz is on the order of 2 dB. For most 20-meter operation this preamp is not necessary, since the mixer input provides a system noise figure of about 10 dB, which is adequate.

A Unitized Receiver for 40 and 20 Meters

Compactness is the key word in this superheterodyne design (Fig. 75). Coverage of 7000 to 7175 and 14,000 to 14,175 kHz is available with this mini-receiver which operates from 12 or 13 volts dc. Maximum current drain is 120 mA, and idling current is on the order of 50 mA. The dimensions (HWD) are 2-5/8 X 4-3/4 X 5 inches. A miniature speaker is built in, and a speaker-disabling jack permits the use of headphones. A minimum number of panel controls are used (tuning, band switch, and i-f gain) to make operation afield or at home as simple as possible.

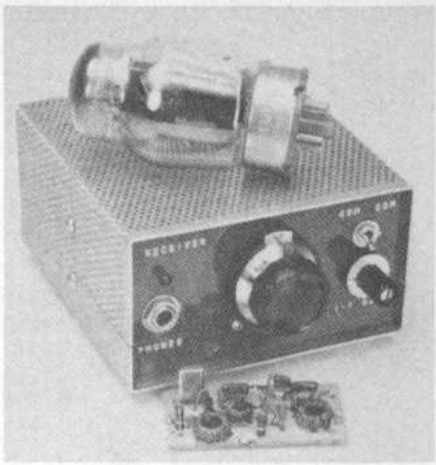
The basic receiver is a 40-meter superheterodyne. There is no agc or af gain control. A simple single-crystal i-f filter is used to minimize cost and circuit complexity. The i-f bandpass is adequate for most cw work and is wide enough for ssb reception.

Wide dynamic range was not the goal in this design. Rather, a sensitive and stable portable unit was desired, which led to some minor trading off in the performance features. However, for all but the most stringent applications, this unit is excellent.

Coverage of the 20-meter cw band is effected by means of a simple two-transistor "down converter" which is mounted inside the main cabinet. Tuning on 20 meters is the reverse of that on 40 meters, owing to the crystal frequency used in the converter. If cw and ssb coverage is desired, the VFO tuning range will need to be extended. Furthermore, two BFO crystals will be necessary, plus a switch, to permit selecting upper or lower sideband. A 0.1-μV signal is plainly audible on both bands. Since that level of sensitivity is greater than necessary for most work, an rf attenuator can be used between the antenna and receiver input to minimize mixer overloading. A simple brute-force attenuator will suffice – a 500-ohm carbon control between the mixer input link and ground, with the antenna connected to the control arm.

Circuit Details

T1 is designed to match a 50-ohm antenna to the 2000-ohm base-to-base impedance of the CA3028A balanced-mixer IC (Fig. 75). The transformer is broadband in nature (300 kHz at the 3-dB points) and has a loaded Q of 23. This eliminates the need for a front-panel peaking control – a cost-cutting aid to simplicity.



Outside view of the unitized 40- and 20-meter receiver, dwarfed by a vacuum tube. The pc board in the foreground contains the 20-meter converter. The cabinet is homemade and consists of two U-shaped pieces of aluminum stock. The front end rear panels are fashioned from double-sided pc board. Dimensions are, in inches, 2-5/8 X 4-3/4 X 5. Dymo-tape labels identify the controls. (From *QST* for September, 1976.)

The output tuned circuit, L1, is a bifilar-wound toroid which is tuned approximately to resonance by means of a mica trimmer, C2. The actual setting of C2 will depend upon the degree of i-f selectivity desired, and typically the point of resonance will not be exactly at 3300.5, the i-f center frequency.

A single crystal filter with a phasing capacitor, C3, is used. This approach provides reasonably good single-signal reception (at least 30-dB rejection of the unwanted response) and assures much better performance than is possible with the simpler direct-conversion receivers in vogue today. The latter have equal signal response each side of zero beat, which often complicates the QRM problem.

A single i-f amplifier, U2, is used to provide up to 40 dB of gain. R1 serves as a manual i-f gain control, and will completely cut off the signal output when set for minimum i-f gain. T2 is designed to transform the 8000-ohm collector-to-collector impedance of U2 to 500 ohms, and has a bandwidth of 100 kHz. The loaded Q is 33.

A two-diode product detector converts the i-f energy to audio. BFO injection voltage is obtained by means of a crystal-controlled oscillator, Q2. RFC2 and the 1-μF bypass capacitor filter the rf, keeping it out of the audio line to U3.

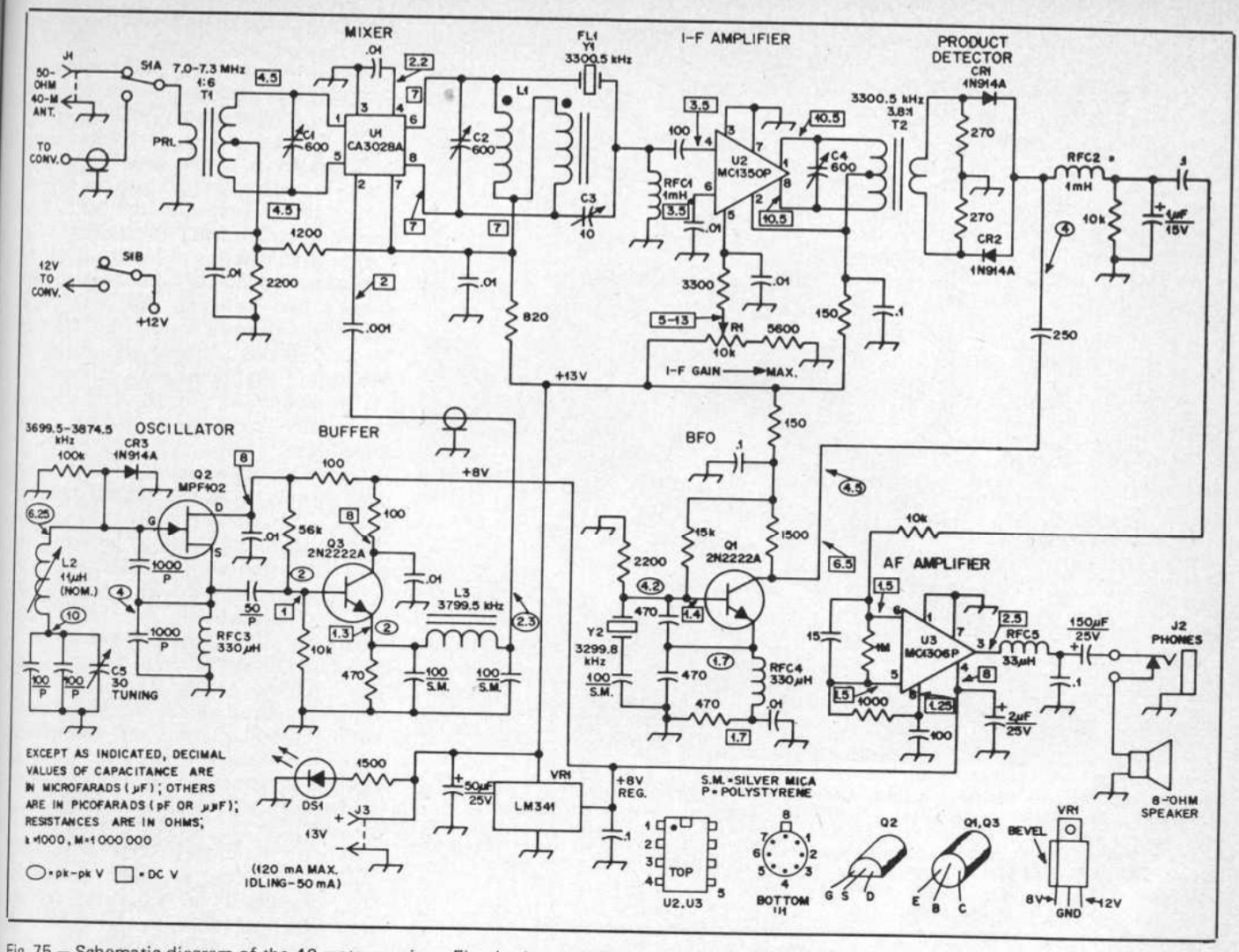


Fig. 75 — Schematic diagram of the 40-meter receiver. Fixed-value capacitors are chip or disk ceramic unless noted otherwise. Capacitors with polarity marked are electrolytic. S.M. indicates silver mica, and P is for polystyrene. Fixed-value resistors are 1/4- or 1/2-W composition.

C1, C2, C4 — 170 to 600-pF mica trimmer (Arco 4213).

C3 — 10-pF subminiature trimmer. Ceramic or pc-mount air variable suitable.

C5 — Miniature air variable, 30-pF maximum (Millen 25030E or similar).

CR1-CR3, incl. — High-speed silicon switching diode

J1, J3 — Single-hole-mount phono jack.

J2 — Closed-circuit phone jack.

L1 — Toroidal bifilar-wound inductor, Q_u = 100 at 3.3 MHz, Q_L = 33, BWL = 0.1 MHz, L = 5.8 μH. 8 turns No. 28 enam., bifilar wound on Amidon FT37-61 ferrite core. Note polarity marks.

L2 — Slug-tuned inductor (see text), 11 μH nominal. J. W. Miller 42A105CBI or equiv. Q_u = 125.

L3 — Toroidal inductor, 17 μH. 19 turns No. 26 enam. wire on Amidon FT50-61 ferrite core.

R1 — 10,000-ohm miniature composition control, linear taper.

RFC1, RFC2 — Miniature 1-mH choke (Millen J302-1000 or equiv.).

RFC3, RFC4 — Miniature 330-μH rf choke (Millen J302-330 or equiv.).

S1 — Miniature dpdt toggle.

T1 — Toroidal transformer. Primary has 2 turns No. 24 enam. wire. Secondary has 14 turns No. 24 enam. wire on Amidon T-50-2 core. Turns ratio — 6:1, Q_L of 23, BWL = 0.3 MHz, L = 1 μH.

T2 — Toroidal transformer. Primary has 9 turns No. 26 enam. wire on Amidon FT37-61 core. Q_L = 33, BWL = 0.1 MHz L = 5.8 μH, turns ratio = 3.8:1. Secondary has 3 turns No. 26 enam. wire. Primary winding has center tap.

U1 — RCA IC. Bend pins to fit 8-pin dual-inline IC socket.

U2, U3 — Motorola IC.

VR1 — Three-terminal 8-volt regulator IC (National Semiconductor).

Y1, Y2 — Surplus crystal in HC-6/U case or International Crystal Co. type GP with 32-pF load capacitance.

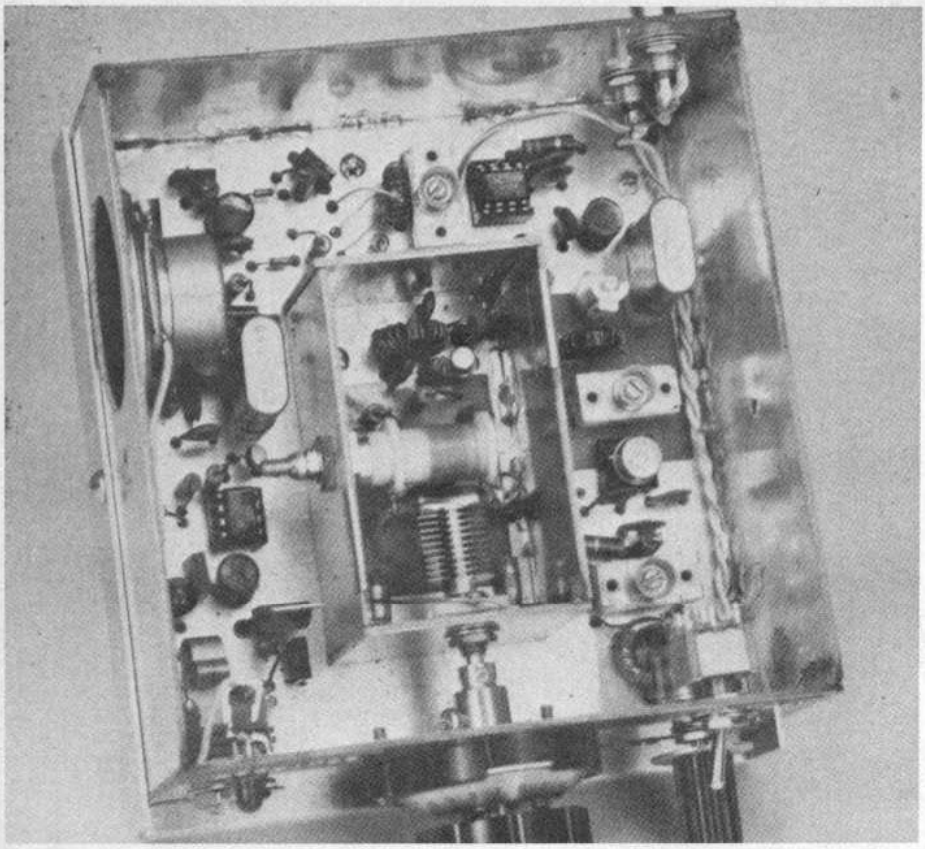
Audio-output IC U3 contains a preamplifier and power-output system. It will deliver approximately 300 mW of af energy into an 8-ohm load. RFC5 is used to prevent rf oscillations from occurring and being radiated to the front end and i-f system of the receiver. The 0.1-μF bypass at RFC5 also helps prevent oscillations.

A three-terminal voltage regulator, VR1, supplies the required operating voltage to U3. It also provides regulated voltage for the VFO and buffer stages of the local oscillator (Q2 and Q3). The latter consists of a stable series-tuned Clapp VFO and an emitter-follower buffer stage. A single-section pi network is placed between the emitter of Q3 and the injection terminal of U1. It has a loaded Q of 1, and serves as a filter for the VFO output energy. It is designed for a bilateral impedance of approximately 500 ohms. The recommended injection-voltage level for a CA3028A mixer is 1.5 rms. Good performance will result with as little as 0.5-volt rms. A 1-volt level is available with the circuit shown in Fig. 75.

A red LED is used at DS1 as an on-off indicator. Since it serves mainly as "window dressing," it need not be included in the circuit.

Construction Notes

The front panel, rear panel, side brackets, and chassis are made from double-sided circuit-board material. The chassis is an etched circuit board, the pattern for which is given in Fig. 77. There is no reason why the top and bottom covers for the receiver can not be made of the same material by sol-



Interior of the unitized receiver. The local oscillator is seen in its compartment at the center. A press-fit U-shaped cover is placed over the VFO box when the receiver is operating. The receiver front end is at the lower right. At the upper left is a miniature speaker, the rim of which is tack soldered to the box wall at four points. The 20-meter converter board mounts on the rear wall of the box (upper left).

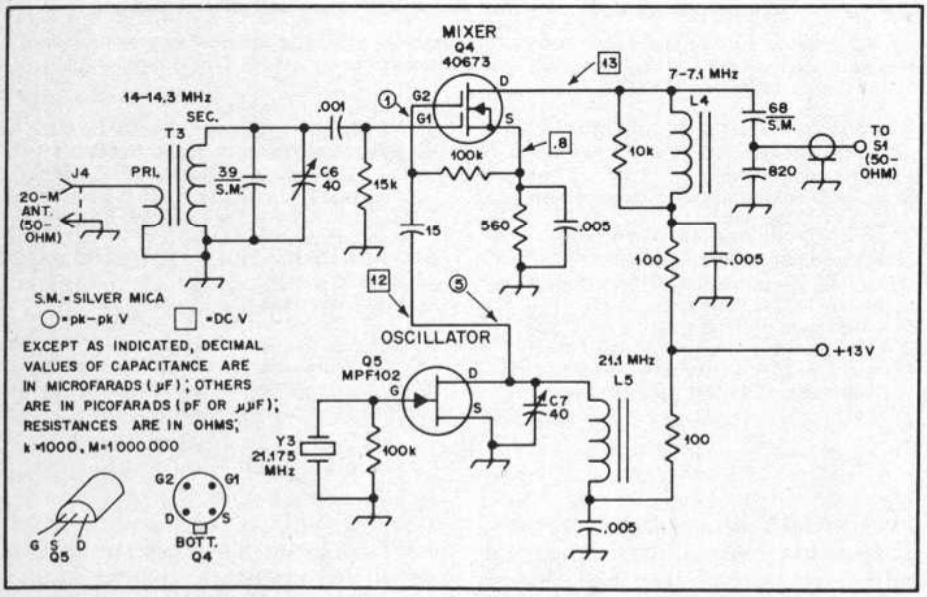


Fig. 76 — Schematic diagram of the 20-meter converter. Fixed-value capacitors are disk ceramic unless noted otherwise. Resistors are 1/4- or 1/2-W composition.

C6, C7 — 40-pF subminiature ceramic trimmer.

J4 — Single-hole-mount phono jack on rear panel of main receiver.

L4 — Toroidal inductor, 12 turns No. 26 enam. wire on Amidon FT37-61 core Q_L = 14, BWL = 0.5 MHz, L = 8 μH.

L5 — Toroidal inductor. 24 turns No. 26 enam. wire on Amidon T-50-6 core. Q_u = 200 at 7.9 MHz. L = 2.4 μH.

Q4 — RCA transistor.

Q5 — Motorola transistor, MPF102, 2N4416 or HEP802.

T3 — Toroidal transformer, 10:1 turns ratio. Q_L = 46, BWL = 0.3 MHz, L = 1.85 μH. Pri. has 2 turns No. 26 enam. wire. Sec. contains 21 turns No. 26 enam. wire on Amidon T-50-6 core.

Y3 — 21.175-MHz fundamental crystal in HC-18/U case (International Crystal Co. type GP with 32-pF load capacitance).

dering six pieces of pc board together to form two U-shaped covers.

The local oscillator is housed in a compartment made from pc-board sections. It measures (HWD) 1-3/8 X 1-5/8 X 2-3/4 inches. A 1/4-inch high pc-board fence of the same width and depth is soldered to the bottom side of the pc board (opposite the VFO top-chassis compartment) to discourage rf energy from entering or leaving the local oscillator section of the receiver (rf doesn't like to climb over right-angle barriers). Employment of the top and bottom shields stiffens the main pc board, and that helps prevent mechanical instability of the oscillator which can result from stress on the main assembly.

Silver plating has been applied to the main pc board, and to the front and rear panels. This was done to enhance the appearance and discourage tarnishing of the copper. It is not a necessary step in building the receiver. The front panel has been sprayed with green paint, then baked for 30 minutes by means of a heat lamp. A coarse grade of sandpaper was used to abrade the front panel before application of the paint. The technique will prevent the paint from coming off easily when the panel is bumped or scratched. Green Dymo tape labels are used to identify the panel controls.

There is ample room inside the cabinet, along the rear inner panel surface, to install the 20-meter crystal-controlled converter. A switch, S1, is located on the front panel to accommodate a 20-meter converter, the circuit for which is given in Fig. 76.

All of the toroidal inductors are coated several times with Q dope after they are installed in the circuit. The VFO coil is treated in a like manner. The polystyrene VFO capacitors should be cemented to the pc board after the circuit is tested and approved. This will help prevent mechanical instability. Hobby cement or epoxy glue is okay for the job. Use only a drop or two of cement at each capacitor — just enough to affix it to the pc board.

Alignment and Operation

The VFO should be aligned first. This can be done by attaching a frequency counter to pin 2 of U1. Coverage should be from 3699.5 to 3874.5 kHz for reception from 7.0 to 7.175 MHz. Actual coverage may be more or less than the spread indicated, depending on the absolute balues of the VFO capacitors and stray circuit inductance and capacitance. Greater coverage can be had by using a larger capacitance value at C5, the main tuning control. Those interested in phone-band coverage (only) can align the VFO accordingly and change Y2 to 3302.3 kHz.

Final tweaking is effected by attaching an antenna and peaking C1, C2 and C4 for maximum signal response at 7087 kHz To obtain the selectivity characteristics desired (within the capability of the circuit), adjust C2 and C3 experimentally. C2 will provide the major effect. C3 should be set for minimum response on the unwanted side of zero beat. A fairly strong signal will be needed to hear the unwanted response.

For reception of lower sideband it will be necessary to use a different BFO frequency – 3298.7 kHz. The crystal indicated in Fig. 75 was used because only cw reception was intended. Those wishing to shift the BFO frequency a few hundred Hz can place a trimmer in series with Y2 rather than use the 100-pF capacitor shown.

Because there is no agc in this receiver, the i-f gain should be set low, for comfortable listening. Too much gain will cause the audio circuit to be overdriven, and distortion will result. To prevent ear-splitting signal levels one can install a pair of 1N34A diodes (back to back) across the output jack, J2.

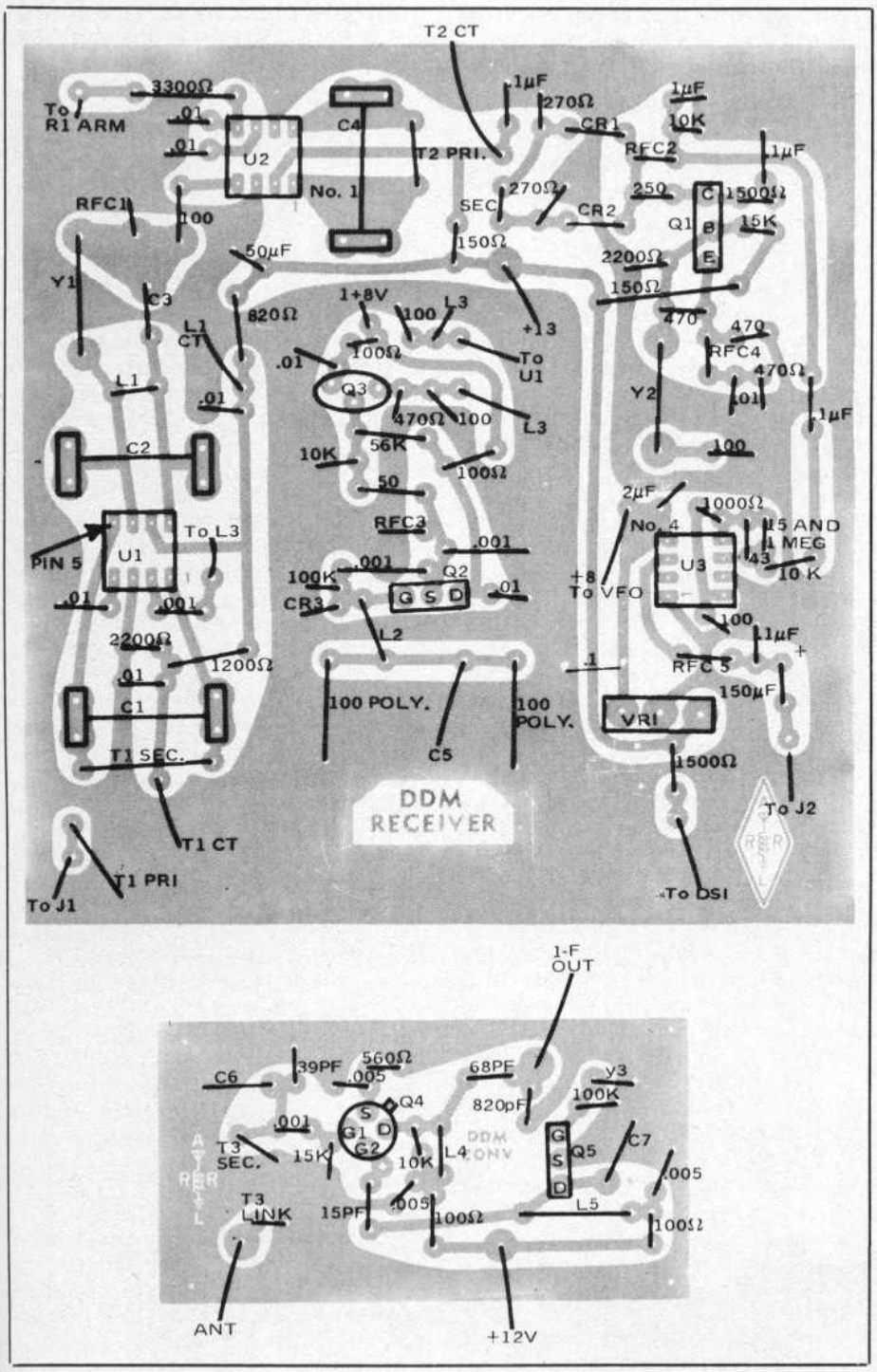


Fig. 77 — Foil-side scale pattern of the pc board. Circuit board is double-sided glass-epoxy material. Ground-plane copper should be removed directly opposite Q2 and related components (oscillator) for area of 1-1/2 inches. Remove copper in similar manner on ground-plane side of board opposite L1, C3 and Y1 (1 X 1-1/4 inch area). Removal of foil will prevent unwanted capacitive effects. The 100-kr gate 2 resistor is on etched foil of board, gate 2 to source. Ground-plane side of board should be electrically common to ground foils on opposite side of board at several points.

Bits and Pieces

The photograph shows some fancy-looking components on the circuit board. Tantalum capacitors are seen where electrolytics are indicated on the diagram. Either type will work nicely. Tantalums were found at a flea market for 10 cents each, so they were used. Similarly, the 0.1-µF capacitors used are the high-class kind (Aerovox CK05BX) which sell for roughly 70 cents each. At the flea market they sold at $1 for 44 pieces! Mylar or disk ceramic 0.1-µF units will be fine as substitutes.

The polystyrene capacitors were obtained from Radio Shack in an assortment pack. New units are made by Centralab, and they sell for less than 20 cents each in single lots. Since they are more stable than silver micas, they are recommended for the VFO circuit. All of the toroid cores were purchased by mail from Amidon Associates.

A J. W. Miller 42-series coil is used in the VFO, but any slug-tuned ceramic form can be used if it has good high-frequency core material. The unloaded *Q* of the inductor should be at least 150 at 3.5 MHz. L2 in this design has a 3/8-inch diameter body. The winding area is 5/8 inch long.

The metal cases of both crystals should be connected to ground by means of short lengths of wire. This will prevent unwanted radiation from the BFO crystal, and will help keep the filter crystal from picking up stray energy. A metal cover should be placed on the VFO compartment for reasons of isolation.

James Millen encapsulated rf chokes are used in the receiver. Any subminiature choke of the approximate inductance indicated will be suitable, and it need not be encapsulated. The VFO tuning capacitor is also a Millen part. Ample room exists between the VFO box and the front panel to allow making the box longer. That will permit use of a larger variable capacitor. A double-bearing capacitor is recommended for best mechanical stability of the VFO.

The i-f system and BFO can be tailored to frequencies other than those indicated. If crystals of other frequencies in the 2- to 3-MHz range are

chosen, the VFO, mixer, and i-f amplifier tuned circuits will have to be altered accordingly.

No hum or distortion is heard in the output of the receiver at normal listening levels. VFO drift is 45 Hz from a cold start to stabilization, and strong signals do not pull the oscillator.

Extremely strong local signals (1000 μV or greater) will cause desensitization of the receiver when they appear off frequency from where the operator is listening. Under ordinary conditions this will not be a problem. At some sacrifice in noise figure and sensitivity, those living in areas where other amateurs are nearby can modify T1 to aid the situation. C1 remains across all of the T1 secondary, and a 2200-ohm resistor is paralleled with C1. Pins 1 and 5 of U1 should be connected two turns each side of the center tap of the secondary. This will require cutting the pc-board elements to divorce pins 1 and 5 from C1. This design tradeoff is quite acceptable at 40 meters, as the atmospheric noise level will mask the reduction in receiver noise performance. With the circuit change there was no desensing evident below approximately 8000 μV.

Agc could be used in this receiver by applying an audio-derived type. If the feature were adopted, agc voltage would be applied to pin 5 of U2 and the manual gain control would be eliminated. In such a case it would be necessary to add an af-gain control between the product detector and U3. It should be remembered that minimum gain results when 13 volts are applied to pin 5 of U2. The lower the voltage at that point, the greater the gain.

Chapter 6

Advanced Receiver Concepts

Some fundamentals of receiver design were presented in chapter 5. However, there was minimal discussion of receiver front-end design. That information forms the basis for most of this chapter.

Conditions in the amateur bands are much different than they were even ten years ago. The spectrum is crowded, with demands for additional space arising daily. Furthermore, the power levels are increasing. In the past it was only the occasional amateur that ran the full legal power limit. Today kilowatt amplifiers are common.

These conditions call for better receivers than those used in the past. Not only must selectivity, sensitivity and stability be maintained, but the receiver must meet these specifications while operating in the presence of numerous strong signals. We will present information in this chapter that will help the amateur experimenter to meet these goals.

The critical portion of the receiver is the front end, that part which precedes the main selectivity-determining elements. Distortion effects in the front end will lead to blocking, intermodulation products and cross modulation. Careful design is necessary if these phenomena are to be minimized.

Dynamic Range

In the previous chapter, some of the basic specifications of receivers, including the idea of noise figure, were outlined. Implicit in the noise-figure concept was the fact that the minimum discernable signal (MDS) of a receiver is dependent not only upon the amount of noise generated by the transistors in the receiver, but upon the bandwidth of the system.

While sensitivity is of major significance to the amateur with an interest in DXing, a receiver must be able to survive in the presence of strong signals. This has a twofold meaning. First, the gain-control mechanisms in the receiver, manual or automatic, must have a range that will permit signals with wide strength variations to be received. However, this can be realized easily – in the extreme case, attenuators in the antenna line can be used to decrease the signal level to a point where intelligence can be recovered.

The second, and more subtle figure for dynamic range, is a number which provides a measure of the range of signals which may be present at the antenna terminals of a receiver while no undesired responses are created in the output. The various ways that such a range can be defined, and the way it is measured, are described in this section. Also, we will show how the concepts surrounding these measurements may be utilized in the design of a receiver.

Consider a simple amplifier in the rf or i-f portion of a receiver. For our example, we will assume that the amplifier uses a bipolar transistor and is biased for a collector current of 10 mA. The concepts are applicable to any amplifier, mixer or complete receiver.

First, we will consider the measurement of the noise figure of the amplifier. By definition, the noise factor of the amplifier is the input signal-to-noise ratio divided by the output signal-to-noise ratio

$$NF = \frac{S_{in}/N_{in}}{S_{out}/N_{out}} = \frac{S_{in}N_{out}}{S_{out}N_{in}} \quad \text{(Eq. 1)}$$

The terms in the equations are noise or signal powers, and the noise factor is an algebraic ratio. If we express that ratio in dB, as is often done with other power ratios (e.g., gains), the result is the noise figure.

As presented, the noise figure is a nebulous number, for the input (and hence, the output) noise power is dependent upon what is hooked to the input of the amplifier. In order to attach some meaning which will make a noise figure number a standard measure of the "noisiness" of an amplifier or receiver, the input noise is assumed to be the noise power available from a resistor at a temperature of 290 degrees Kelvin. Using this value for T_o, the noise power is given as $P_n = kT_oB$, where T_o = 290 degrees Kelvin, B is bandwidth in Hz, and k is Boltzman's constant, 1.38×10^{-23} watts/degree. It is convenient to use logarithmic units and to note that in a bandwidth of 1 Hz, $P_n = -174$ dBm.

Consider a receiver with a bandwidth of 500 Hz. The bandwidth is greater than one Hz by a factor of 500, or 27 dB. Hence, in a 500-Hz bandwidth, the power available from this resistor would be −174 dBm + 27 dB = −147 dBm. If the noise output from this receiver with the input terminated in a 50-ohm resistor corresponds to that output which would result from a signal of −140 dBm, the noise figure of the receiver is then the difference, or 7 dB. The MDS, or noise floor of the receiver is −140 dBm.

One might ask why noise figure is even specified. The same essential information is contained in a specification of the MDS of a receiver. However, such is not the case for an amplifier. Here, the MDS is not specified – it will depend not only upon the noise contribution of the amplifier, but on the bandwidth of the system using that amplifier. Noise figure is independent of bandwidth.

A further asset of noise figure is that it is, at least in principle, measured easily. This is a direct result of the bandwidth invariance. The measurement is performed by attaching a source to the input of a receiver (or amplifier) that has a noise output which is known

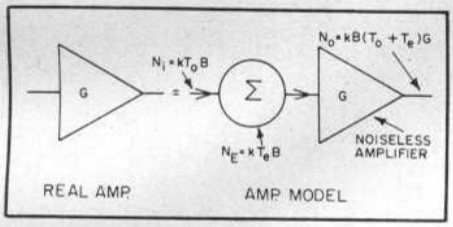


Fig. 1 — The principle of noise temperature.

to be some well-defined factor greater than that of a room-temperature resistor. As long as the noise from this source is distributed evenly (white noise) over the frequencies of interest, the device being measured will respond to this known input with exactly the same filtering bandwidth that is applied to the internally generated noise. By measuring the increase in output noise, the noise figure is easily calculated. Knowledge of the system bandwidth is not required in the calculation.

A related concept which also describes the noisiness of an amplifier or receiver is that of noise temperature. This concept is outlined in Fig. 1, where the device being evaluated is modeled by a noiseless amplifier preceded by an "ideal adder" and a noise generator. The excess-noise generator represents the noise that is contributed by the amplifier. Effective noise temperature is related to noise factor by the equation $F = 1 + T_{eff}/T_o$, where T_{eff} is the effective noise temperature of the amplifier and T_o is the reference temperature (usually 290 degrees Kelvin). This equation is derived easily if we recall that noise factor can be expressed as the ratio of noise gain to signal gain. If the available gain of the amplifier is G, the noise output will be

$$N_{out} = G(kT_oB + kT_{eff}B). \qquad \text{(Eq. 2)}$$

The input noise power is just kT_oB. Noise factor is then

$$\text{NF} = \frac{G_N}{G_S} = \frac{1}{G}\, kB\, \frac{(T_o + T_{eff})}{kBT_o} \qquad \text{(Eq. 3)}$$

$$= 1 + \frac{T_{eff}}{T_o}$$

As an example, assume that the effective noise temperature of an amplifier is 400° Kelvin. The noise factor is $F = 1 + 400 \div 290 = 2.38$. The noise figure is 3.76 dB.

The advantage of the noise-temperature concept over that of noise figure is that it is an *absolute* number. It is not dependent upon the more or less arbitrary choice of a reference temperature. It also has the advantage that it is in some cases, a more meaningful indicator of the ability of a system to detect very weak signals. This requires some elaboration.

Assume that a receiver with a noise figure of 3 dB is made more sensitive by adding a preamplifier which provides a net system-noise figure of 0.5 dB. One might assume that because the noise figure of the system has decreased by 2.5 dB, we will be able to hear signals which are 2.5 dB lower. However, this is generally not the case. It would be true *only* for the situation where the input noise to the system was originating from a 290° Kelvin source. If the noise was originating from atmospheric disturbances (causing noise in the hf spectrum), the increase in output signal-to-noise ratio would be virtually imperceptible. On the other hand, if the noise was from a large parabolic antenna pointed toward one of the quieter parts of outer space, the input noise would be nearly zero. In this case, the 2.5-dB improvement in noise figure could lead to an approximate 9-dB improvement in receiver sensitivity. This conclusion results from Eq. 3, which shows that a drop in noise figure from 3 to 0.5 dB corresponds to decreasing the effective noise temperature from 290° to 35° Kelvin.

Consider now the case where two relatively strong signals are placed simultaneously at the input to the 20-dB amplifier mentioned earlier. Assume that two input signals of −50 dBm are placed at the input of the amplifier at frequencies f_1 and f_2. Analysis of the amplifier using mathematics outlined in the appendix will show that distortion in the amplifier will give rise to outputs not only at the desired input frequencies of f_1 and f_2, but at $(2f_1 - f_2)$ and $(2f_2 - f_1)$. For example, if the input frequencies were 14,040 and 14,050 kHz, the distortion products would appear at 14,030 and 14,060 kHz. In the amplifier the desired outputs would be 20 dB above the −50-dBm input signals, or −30 dBm, and the 3rd-order distortion products would be at −130 dBm. In this case the distortion will be 100 dB down from the desired outputs.

The interesting and significant characteristic of Class A linear amplifiers is that while the desired outputs will vary

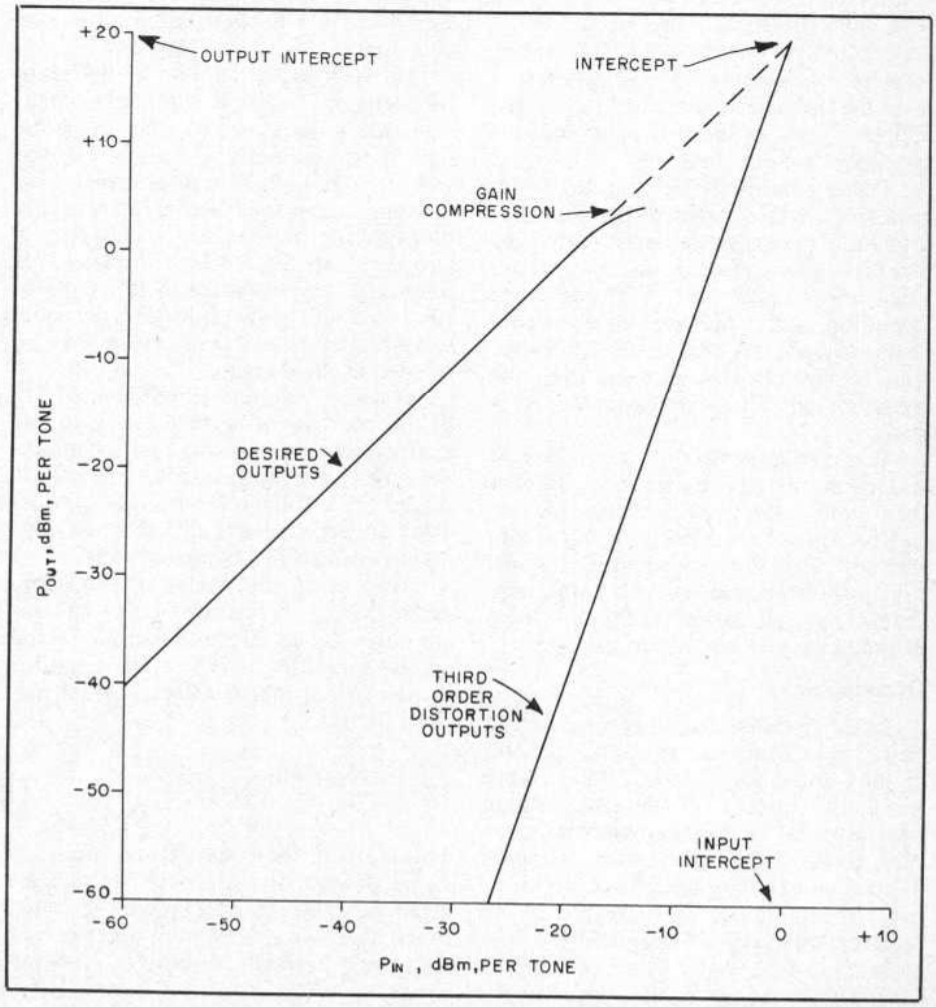


Fig. 2 — Plot example showing signal power versus distortion products as a function of input power of two identical input signals.

linearly with changes in the input signals, the dominant distortion products will vary as the cube of the input powers. Hence, if we increase the signals driving the input to −40 dBm, the output power of the desired signals will be −20 dBm for each of the desired input tones. However, while the level of the desired frequencies increased by 10 dB, the output power of the distortion products will have increased by 30 dB to −100 dBm. The distortion products are now only 80 dB below the desired results.

Shown in Fig. 2 is a plot for our hypothetical amplifier, showing the power of the desired output signals and the output power of the distortion products as a function of the level of the input power of each of the two identical input signals. Eventually, the level of the input signals will be large enough so that the desired outputs cease to follow the input power linearly. This effect is called *gain compression,* and is the phenomenon in a receiver which ultimately leads to "blocking." It is not viable to plot the data for the amplifier much beyond this compression point.

The linear portions of the curves may be extended, or extrapolated to higher powers even though the amplifier is not capable of operating at these levels. If this is done, as is shown in a dotted line in the figure, eventually the two curves will cross each other. That is at some usually unattainable output power, the level of the distortion products equals that of the desired outputs. This point is commonly referred to as the *amplifier intercept.* More specifically, the output power where the curves intersect is called the *output intercept* of the amplifier. Similarly, the input power corresponding to the point of intersection is called the *input intercept.*

It is important to distinguish between the input and the output intercepts when specifying a given device. In any useful amplifier (one with power gain) the output intercept is always greater than the input intercept by an amount corresponding to the gain of the amplifier. But with lossy circuits (such as a diode mixer) the input intercept will exceed the output intercept. In professional literature the number usually given is the output intercept. However, the input intercept is an equally important number when discussing receivers.

The value of knowing the intercept of an amplifier is that it is a general measure of the distortion properties. It can be used to describe the distortion for all operating levels. In the case just depicted the output intercept is +20 dBm. Hence, if the amplifier is operated with an output which is X dB below the intercept, the distortion will be 3X below the intercept. For example, if the amplifier is operated with outputs of 0 dBm, which is 20 dB below the intercept, the distortion products will be three times the 20-dB difference, or 60 dB below the intercept at −40 dBm. In our example amplifier, the input intercept is 0 dBm. The same relationships apply using this figure of merit.

It is generally not viable to specify the output intercept of a receiver, for this is a function of the gain setting of the unit. However, such is not always the case, with an input intercept. This number may be specified and is an extremely useful general parameter. Suppose, for example, that the input intercept of a receiver is 0 dBm. (This number is not purely arbitrary, but is representative of a well-designed communications receiver.) This means that if two signals are placed at the antenna terminals with levels of −40 dBm, the response when the receiver is tuned to the frequencies of the distortion products ($2f_1 - f_2$ or $2f_2 - f_1$) will be three times 40 dB below the input intercept, or the same as an input signal of −120 dBm.

As is usually the case with receivers, the analysis of performance is complicated by noise. If the two inputs just mentioned were dropped to −60 dBm (which is 60 dB below the input intercept) the response at the distortion-product frequencies would be 180 dB below the input intercept or at −180 dBm equivalent input signal. If this receiver had an exceptionally low noise figure and a bandwidth of a fraction of one Hz, this level of signal could be detected. However, this is not usually the case with communications receivers. If the receiver had a more typical MDS or noise floor of −140 dBm, the distortion products would not be detectable. This brings us to the concept of *dynamic range.*

The two-tone dynamic range of a receiver is defined as the ratio of the noise floor (MDS) of the receiver to the level of one of two identical input signals which will cause distortion products at the noise floor level. This concept is illustrated by considering a measurement on the receiver described in the foregoing discussion.

First, the instrumentation is gathered and interfaced with the receiver. This includes a pair of signal generators with means for combining their outputs while minimizing interaction between them, and an ac voltmeter to monitor the audio output signal.

The initial measurement uses only a single signal source. The generator is adjusted so that the output of the receiver is 3 dB above the level present when the generator is turned off. The power output (available output power in dBm) of the generator is then the MDS of the receiver.

After measuring the receiver MDS, the two generators are set up for IMD measurements. The two generators are added in a 6-dB hybrid combiner. The output is applied to a step attenuator and then to the receiver. The attenuator is adjusted until the responses at the third order IMD frequencies are the same as that produced by the MDS. The DR in dB is then the dB difference between the power in each tone available to the receiver input and the MDS.

The two-tone dynamic range of a receiver is related to the input intercept of the receiver by the relationship

$$\text{Dynamic range (in dB)} = 2/3(P_i - \text{MDS}) \qquad \text{(Eq. 4)}$$

where the input intercept, P_i, and MDS are in dBm.

At the time that the receiver is being evaluated for intermodulation distortion, blocking measurements are also performed easily. This is done by setting one of the generators to provide a medium-strength signal in the receiver. With the receiver tuned to this output, the other generator is increased in output until the desired output is reduced by 1 dB. This onset of desensitization, when compared with the noise floor of the receiver, might be referred to as a "single-tone dynamic range."

The use of blocking, and more specifically, intermodulation distortion as the mechanisms to define the strong signal performance of a receiver, might appear esoteric and restrictive. However, such is not the case. The blocking measurement will tell the user how well his receiver will survive when subjected to a strong neighbor. The two-tone dynamic range will indicate the level of signals which the receiver will tolerate while producing essentially no undesired responses.

The authors have evaluated a number of commercially built receivers. The best unit studied at this writing had a two-tone dynamic range of 88 dB with a noise figure of 5 dB. The single tone dynamic range was only 116.5 dB. This unit used tubes in the front end. An "average" performer yielded two-tone and single-tone dynamic ranges of 80 and 109 dB, respectively. On the other hand both authors have constructed solid-state receivers with two-tone dynamic ranges approaching 100 dB, single-tone ranges of over 120 dB and noise figures from 6 to 13 dB. While sophisticated instrumentation was used for evaluation, both units were built using only equipment available in many amateur shops. Both receivers are described in this book.

It is interesting to consider the effect of cascading two or more amplifiers (or a receiver with a "preamp" or converter) with respect to the effect on noise figure and dynamic range. Knowing these, we will be able to calculate the resulting dynamic range.

Consider two cascaded amplifiers. If they have noise factors F_1 and F_2, and gains G_1 and G_2 (both are algebraic ratios, not dB relationships) the net noise factor of the combination will be given by

$$F_{net} = F_1 + (F_2 - 1)/G_1 \qquad \text{(Eq. 5)}$$

For example, assume that each amplifier has a gain of 20 (13 dB), that the first one has a noise factor of 2 (3 dB) and the second has a noise factor of 5 (7 dB). The net noise factor is $F_{net} = 2 + (5 - 1) \div 20 = 2.2$, which corresponds to a noise figure of 3.42 dB. The net gain is 400, or 26 dB. Note that the net noise figure is dominated by the first stage of the amplifier if the gain of the first stage is large in comparison to the noise figure of the second stage. But, excess gain in the first stage beyond this level does little to improve the net noise figure.

Assume that the first stage has an output intercept of +15 dBm and that the second stage is stronger, with an output intercept of +20 dBm. Since the gain of the second stage is 13 dB, the input intercept of the second stage will be $+20 - 13 = +7$ dBm. Noting that this input-intercept amount is less than the output intercept of the first stage (a margin of 8 dB), the IM response of the composite amplifier will probably be dominated by the distortion in the second stage. We can estimate the output intercept of the combined amplifier to still be +20 dBm. Since the overall gain is 26 dB, the input intercept of the cascaded pair will be $+20 - 26 = -6$ dBm.

It should be mentioned that the IM distortions from two cascaded stages will add in a simple manner, with the output stage usually being the dominant contributor. However, there are some situations where the IM from one stage will add in a phase-coherent way with a following stage: The overall result is IM which is much worse than anticipated. In rare examples the opposite effect will occur, yielding better distortion properties than predicted. These cases do not lend themselves to easy analysis or duplication.

Note that in the foregoing discussion nothing has been said about dynamic range. This is because the dynamic range is defined while using an input equivalent noise floor which is dependent upon noise figure and system bandwidth. A dynamic range can be specified only when a bandwidth is given simultaneously.

As an extension of the discussion, let us consider adding a preamplifier to a receiver which is lacking in noise figure. Assume that the receiver has an exceptionally poor noise factor of 100 (20 dB), and a dynamic range of 80 dB. The bandwidth of the receiver is 500 Hz. The minimum detectable signal, or noise floor of the receiver will be

$$\begin{aligned}\text{Noise floor} &= -174 \text{ dBm} + \text{noise figure} \\ &\quad + \text{bandwidth factor} \\ &= -174 \text{ dBm} + 20 \text{ dB} \\ &\quad + 27 \text{ dB} \\ &= -127 \text{ dBm}\end{aligned} \qquad \text{(Eq. 6)}$$

If this receiver was to be used in the 10-meter band a much lower noise figure might be in order. Assume that a preamplifier with a 3-dB noise figure is added. Following the earlier argument about noise figure, a preamplifier gain of 20 dB, equal to the receiver basic noise figure, is used. The net noise figure becomes

$$F = 2 + \frac{99}{100} = 2.99 \text{ or } 4.76 \text{ dB} \qquad \text{(Eq. 7)}$$

The noise floor decreases to

$$\begin{aligned}\text{Noise floor} &= -174 \text{ dBm} + 4.76 + 27 \\ &= 142.24 \text{ dBm}\end{aligned} \qquad \text{(Eq. 8)}$$

The improvement in sensitivity is profound.

Consider now the effect of the preamplifier on the dynamic range of the receiver. Using the formula relating dynamic range to noise floor and input intercept, we deduce that the input intercept of the basic receiver is −7 dBm. If the preamplifier is even reasonable (from a distortion point of view), the distortion properties of the overall system will be dominated by the receiver basic input intercept of −7 dBm. The system input intercept will be −27 dBm. The overall system dynamic range is

$$\begin{aligned}\text{Dynamic range} &= 2/3\,(-27 + 142.24) \\ &= 76.8 \text{ dBm}\end{aligned} \qquad \text{(Eq. 9)}$$

The dynamic range has been slightly degraded from the original dynamic range of 80 dB, which is an acceptable compromise.

If, however, the gain in the preamplifier had been set at 30 dB instead of the 20-dB level chosen, the extra gain would drop the noise floor to −143.78 or merely 1.5 dB more sensitive. However, the input intercept would drop to −37 dBm, resulting in a dynamic range of 71.2 dB. Such a compromise would not be acceptable except perhaps in very rare situations such as moonbounce work on 144 or 432 MHz where noise figure is all! The price to be paid is always a severe degradation in dynamic range.

One final comment should be made about receiver dynamic range. The common "cure" that is suggested for a receiver plagued with problems of overload and excessive intermodulation distortion is the addition of an attenuator in front of the receiver. Often this is an excellent thing to do. The attenuator is adjusted until the antenna noise still determines the overall noise output but is not excessive.

The addition of a 10-dB pad in front of a receiver has the effect of increasing the system noise figure and the input intercept by 10 dB. The difference between the two, and hence the system dynamic range, remains constant. A much better solution would be to bypass the offending amplifier, allowing smaller signals to impinge upon the mixer. While the noise figure will be compromised, the dynamic range will usually be improved.

There is another technique that may be applied to regain some of the system dynamic range: the application of attenuation between the preamplifier and the main receiver. Consider the previous case where a mediocre receiver was preceded by a 30-dB-gain preamplifier, causing a net dynamic range of only 71.2 dB. If a 12-dB attenuator was inserted between the preamplifier and the receiver, the net system MDS would increase from −143.8 to −141.5 dBm. However, the dynamic range would increase to 77.6 dB. This technique could be of major significance when building a dual-conversion system with crystal-controlled converters ahead of a tunable i-f receiver.

While it is dangerous to generalize, it is clear that the optimum dynamic-range systems will be those utilizing single conversion. However, wide dynamic range is certainly possible in multiconversion designs. Great care must be applied in tailoring the gain distribution properly, in order to optimize the trade-off between dynamic range and noise figure. Careful measurements, as well as detailed calculations during the design phase, are mandatory.

In the following sections, the design of mixers, amplifiers and filters will be considered in more detail than presented in chapter 5. The major difference in this approach will be our inclu-

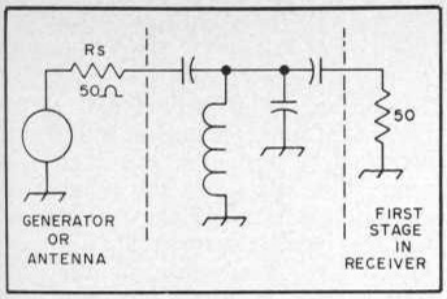


Fig. 3 — Representation of a receiver input circuit, coupled capacitively.

sion of intercept data as well as noise performance of the various devices.

Preselector Design

The previous section outlined the concepts of dynamic range and described some of the undesired effects that arise from excessively strong signals appearing at the input of a receiver. Much of the key to minimizing these effects lies in the design of the mixers and amplifiers that make up the front end of a receiver. As much as possible should be done to ensure that the front-end components are subjected to a minimum of strong signals. This is realized with careful filtering at the antenna terminal of a receiver. Such a filter is called a *preselector.*

The subject of filter synthesis is a complicated one. Sophisticated mathematics are required, making a complete discussion impractical in this book. However, some of the basic ideas can be presented. An extensive catalog of computer-designed filters for the amateur bands is given in the appendix for use in specific projects.

The Single Tuned Circuit

With most receivers in use today, the preselector consists of nothing more than a single tuned circuit preceding the rf amplifier (if one is used) or the mixer.

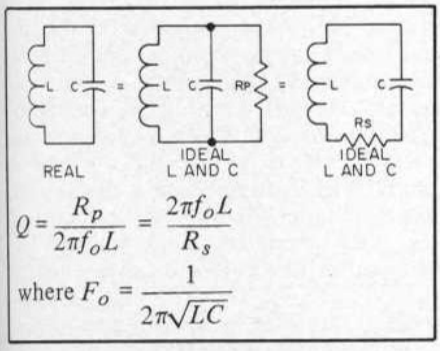


Fig. 4 — Modeling of an ideal resonator with series or parallel resistance.

While this may be adequate to provide marginally acceptable image rejection, it usually provides a minimum of protection from out-of-band signals that might lead to IMD products. We will investigate this type of preselector for two reasons. First, the inadequacy of such a circuit will be demonstrated. Of more significance, we will use the single-tuned circuit to demonstrate some fundamentals that are applicable to any preselector.

Consider a receiver with the first semiconductor device having an input impedance of 50 ohms. If a preselector is to be designed for this receiver, it must be a circuit that is terminated on both sides (input and output) with a 50-ohm load. A typical circuit is shown in Fig. 3 where capacitive coupling is used at both terminals.

The concept of Q was introduced in our discussion of tuned transmitter buffer amplifiers. Q is a number that gives us information about the losses in a resonator. (The term resonator will be used interchangeably with "tuned circuit." The concepts are applicable to microwave resonant circuits just as they are to low-frequency LC tuned circuits and even to nonelectrical oscillations.) While Q tells us the amount of energy that is lost during each cycle of oscillation, we can model a real resonator by replacing it with an ideal lossless one with either a parallel or series resistance. This is shown in Fig. 4 along with the equations which define the resistances.

If the resonator exists alone, attached to no external load, the Q is the unloaded value, designated Q_u. The associated resistances model the inherent losses within the inductor and capacitor. In the high-frequency region inductive losses are predominant in most cases. Hence, one will often see a Q_u specification for a coil at a given frequency.

If external resistances are attached to the resonator, the resulting Q is termed the loaded value and is represented by Q_L. The corresponding resistance is the equivalent of all of the loads, including that representing the inherent resonator losses.

A term that is rarely used but can occasionally be useful in calculations is Q_e, the external Q. This is merely the Q associated with the external resistances attached to the tuned circuit.

Let us now return to the filter described in Fig. 3 and consider the effect of the finite unloaded Q of the resonator. This is done by substituting the model of Fig. 4 for the tuned circuit, now shown in Fig. 5. First, there will be loss associated with this filter. If the filter was removed completely, with a direct connection between the source and load resistors (which here are equal), the power that would be delivered to R_L would be the maximum available amount that the generator could deliver. Substitution of the filter places another resistive element into the circuit. This is the loss resistance, R_u, associated with Q_u of the resonator. Since a voltage will appear across the resistor, it must dissipate power. This will be subtracted from the maximum available power from the generator.

The loaded Q of the resonator is calculated easily by performing a straightforward transformation which is detailed in the filter appendix. It may be shown that, at a single frequency, a given series R-C combination may be replaced with an equivalent parallel one. The input voltage generator is also replaced by a current generator. The resulting circuit is shown in Fig. 5B.

The resistance across the resonator is now the parallel equivalent of R_s', R_u and R_L'. If this circuit is analyzed with respect to the loaded and unloaded Q of the resonator, it may be shown that the insertion loss of the resonator is given by

$$IL = -10 \log (1 - Q_L/Q_u)^2 \qquad \text{(Eq. 10)}$$

In order to minimize the insertion loss of the filter, the loaded Q must be small in comparison with Q_u. Noting the relation between resonator Q and its 3-dB bandwidth, this means that the bandwidth should be fairly large in order to hold the insertion loss down to a reasonable level.

This characteristic is qualitatively true for much more sophisticated filters. However, the simple relationship of Eq. 10 no longer applies with filters of more than one resonator.

Fig. 6 shows a general example of a multiple-resonator filter. In this case a

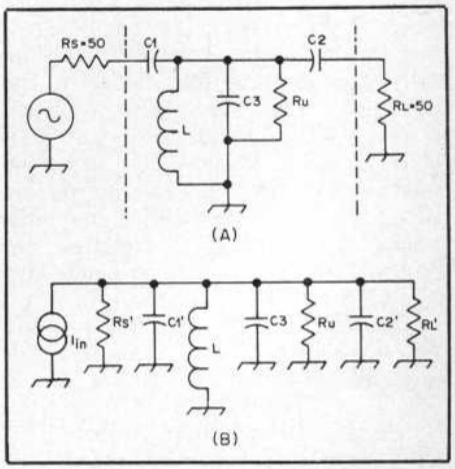


Fig. 5 — Example of a filter which has loss.

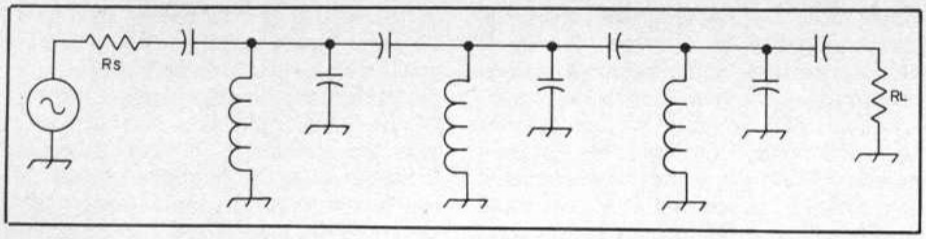


Fig. 6 – Example of a multiresonator filter.

3-section filter is shown, although the general circuit configuration may be extended arbitrarily to any number.

Capacitors are used in the 3-pole example of Fig. 6 in order to couple energy between the resonators, and to couple the source and load into and out of the filter. Inductive coupling could also be used, or a mixture of the two methods could be employed.

The techniques of modern filter synthesis tell us that a given filter may be realized with resonators of equal Q_u if we establish the coupling between sections and control the singly loaded Q of the end sections. By singly loaded Q, we mean the loaded Q of the end resonator, when terminated, but with no coupling to the rest of the filter.

Virtually any type of passband shape may be specified. Some of the common types include the Butterworth, Chebyshev and Gaussian responses. These names are ones that we often hear in connection with filters, but are rarely explained in the amateur literature. They are essentially mathematical terms naming the sometimes fairly complicated polynomials that describe the position of the poles of the filter in the complex frequency plane. In more practical terms, they also lead to different filter characteristic shapes. The Butterworth filter is one that is relatively flat across the passband. Indeed, this filter is often called a maximally flat response (mathematically, the first derivative of the transfer function vanishes at the center of the passband). The Chebyshev filter is somewhat more complicated. Some passband ripple may exist, but the skirt response close to the edges of the passband is steeper. The Gaussian response is not as flat across the passband as the Butterworth or some Chebyshev filters. However, it has the advantage that "ringing" is minimal. Hence, Gaussian transfer functions are optimal for very narrow-bandwidth crystal filters, as an example.

The filters described in the appendix are all designed for a Butterworth response. The main reason for this is that a Butterworth filter is among the easiest to align without resorting to advanced alignment techniques or extensive instrumentation. The attenuation of a Butterworth filter is given by

$$\text{Atten (dB)} = 10 \log (1 + S^{2n}) \qquad \text{(Eq. 11)}$$

where n is the number of resonators. S is the ratio given by

$$S = \frac{f - f_c}{F_{3+} - f_c} \quad \text{or} \quad S = \frac{f_c - f}{f_c - f_{3-}} \qquad \text{(Eq. 12)}$$

where f is the frequency of interest, f_c is the center frequency of the filter, and f_{3+} and f_{3-} are the upper and lower 3-dB attenuation frequencies of the filter. Which form of the equation is used will depend upon whether the frequency of interest is above or below the center frequency of the filter.

As an example of this equation see Fig. 7, where responses for a number of Butterworth filters are given. They are all designed for 3-dB attenuation frequencies of 7.0 and 7.2 MHz. Curves are plotted for one through five resonators. The difference in skirt response as the number of tuned circuits in the filter is increased is profound, but there is a price to be paid. As the number of resonators is increased, the insertion loss will also increase dramatically for filters with a fixed bandwidth, all using the same type of resonator (constant Q_u).

This is not the only effect of the loss elements in a filter. It turns out that the finite Q of the resonators complicates the design. If classic image-parameter methods were used for the filter design, we would find that the filter shape would be distorted over that predicted when it was built and measured. In order to compensate for this effect, so-called predistorted filter tables (see the reference by Zverev in the bibliography) were used for the designs. Because of the subtlety, a general equation set cannot be specified for the design. Furthermore, the filters described in the appendix *can not* be scaled to other frequencies in the simple way that image-parameter filters can.

As mentioned earlier, there is sometimes an advantage to the use of capacitive or inductive coupling over the other. When capacitive coupling is used, the skirt response tends to be a bit steeper on the low frequency side. This is because the filter tends to degenerate into a high-pass structure *away* from the passband. Similarly, inductive coupling seems to make the high-frequency skirt steeper. These effects become signifi-

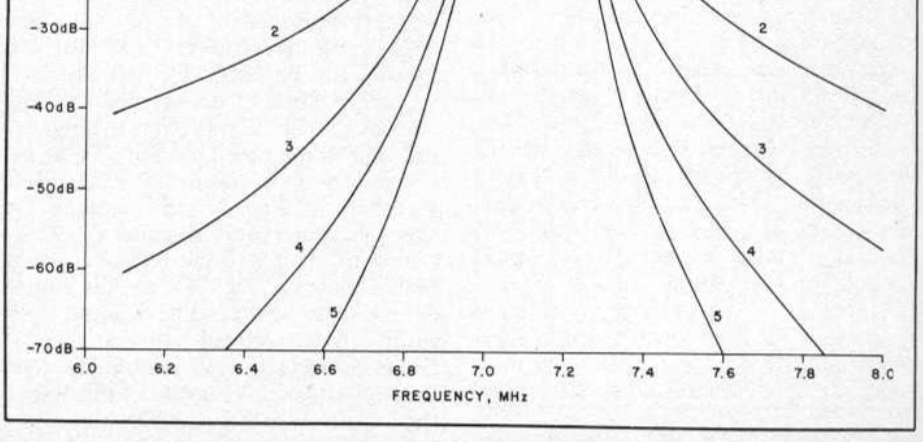


Fig. 7 – Response curves for a number of Butterworth filters.

cant well down on the response curves. For a 3-pole filter the differences become apparent when attenuations of more than 50 or 60 dB are achieved.

If a narrow filter is designed so it may be tuned over a range of frequencies from the front panel of a receiver, proper coupling techniques should be used. If a multisection variable capacitor is used, inductive coupling is preferred between resonators. On the other hand, if a number of inductors are tuned simultaneously, capacitive coupling is desired.

Although there are some exceptions, most filters using a multiplicity of resonators must be terminated properly at each end. The filters described in the appendix have components listed for termination of each end in 50 ohms. It is possible, however, to terminate them in much different impedances. The methods for achieving this are also outlined.

A preselector filter that has become popular recently is the so-called Cohn filter. This circuit is tunable from the front panel over a reasonable frequency range. The unusual characteristic of this circuit is that four resonators are used. However, only a three-section variable capacitor is required to tune it. The filter, as originally designed, was optimized for minimum loss in the passband, making it ideal for receiver applications. A representative circuit for the Cohn filter is given in Fig. 8. Generally, this circuit may be scaled to other frequencies. The 3-dB bandwidth may be increased by making the coupling inductors (1.45-μH units in Fig. 8) larger in value. The skirt response can be made steeper by increasing the value of the shunt capacitors (270-pF units of Fig. 8).

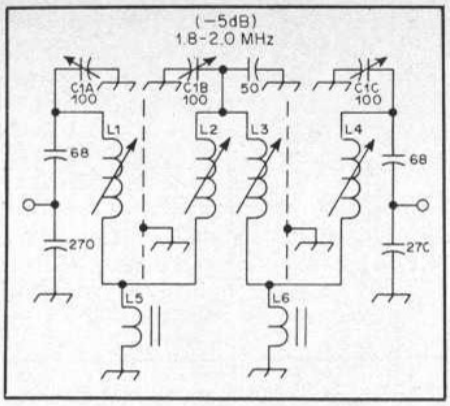


Fig. 8 — Tunable Cohn type of filter for 1.8 MHz. L5 and L6 are 1.45-μH bottom-coupling toroidal inductors.
L1, L4 — 70 μH
L2, L3 — 140 μH

Mixer Design

At the start of this chapter were concepts to define and measure the two-tone dynamic range of a receiver. The effects of adding or subtracting gain in a receiving system were discussed. However, little was said about the main origins of the IMD which limits dynamic range. This topic is treated now.

In the current state of the art we find that the design of filters and amplifiers is highly refined. By proper choice and application of transistors, low noise figure and high-intercept amplifiers are possible. The next section will present some of this information. Generally the mixer is the limiting element in a receiving system. If better mixers can be built, the amplifiers that are needed to accompany them are within reach, although still difficult to realize.

An amplifier is a device that relies upon the linear characteristics of a transistor in order to provide gain. By using devices that operate at high current levels, and by the application of feedback, this linearity can be emphasized. Similarly, filters employ passive elements which tend to be inherently linear. However, in order to achieve mixing action, nonlinear operation is desired. We must utilize square-law characteristics or the switching action in order to realize mixing. (The fundamental mathematics are outlined in the appendix.) Hence, in a device operated purposefully in a nonlinear mode, we would expect other responses, including unwanted ones, to occur.

There are a number of devices that will function well as mixers. They all have their assets and problems. Some of these will be presented with some guidelines for their use.

The Dual-Gate MOSFET

A popular mixer device in amateur equipment today, both commercially manufactured and homemade, is the dual-gate MOSFET. There are many varieties available. Unfortunately, adequate data are not provided by the venders, making it hard to say which is an optimum choice. Experiments suggest that the variations are not great.

There is good reason for the popularity of the MOSFET. It is a device that can provide considerable gain (sometimes desired). Furthermore, the noise figure is fairly low and the output intercept is rather high, especially when minimum power consumption is considered. Finally, the local-oscillator power required is low, making the device easy to apply.

A typical mixer is shown in Fig. 9. In this circuit pi networks are used to match both the input (gate 1) and the output at the drain. This is done to establish the impedances seen at the two ports of the device. A variable voltage bias source is used to establish the operating conditions at gate 2 which lead to the best performance. The output was applied to a spectrum analyzer while the input was driven from a pair of signal generators which were added in a hybrid combiner. An attenuator was used after the combiner in order to ensure proper operation of that component. (An easily made combiner will be described later for use in the amateur shop.) A third generator was used as an LO.

First, it was found that the gain of the mixer was dependent upon the terminating impedances and the level of the LO voltage applied to gate 2. There was also some variation when other similar device types were used in the circuit. Of major significance is the fact that the conversion gain was always about 12 dB lower than the gain of the same device when operated as an amplifier with the same termination impedances. This implies that the conversion transconductance is 1/4 of that displayed when the same device is operated as an amplifier. This optimum gain occurred with an LO injection of about 5 volts pk-pk at gate 2. It was also found that the optimum dc bias voltage for gate 2 was about 1 volt. This tells us that the common practice of attaching gate 2 to the source of the device through a large resistor is a good one.

The intermodulation distortion performance was good. With a 2000-ohm termination on the drain (at 9 MHz) the output intercept for third-order IM was +19 dBm. This same output intercept was obtained when the device was operated as an amplifier at 14 MHz (same termination impedances). When the MOSFET was operated as an amplifier or a mixer, gain compression occurred just a few dB below this intercept level. The 5-volt pk-pk LO injection appeared optimum for both blocking and IMD performance.

The nature of the output termination is critical with this mixer. In the experiment outlined, the output of the pi network at the drain was the 50-ohm input of a spectrum analyzer. This termination was quite flat at virtually all frequencies. This is not typical in the usual application. The more common termination for the mixer is the input of a crystal filter. While the filter may appear to be a clean resistive termination within the passband of the filter, the input impedance is usually quite different at other frequencies. The usual ladder type of filter looks something like an open circuit at frequencies near (but not exactly in) the passband of the filter. If this were applied directly to the drain of the mixer, the results could be quite compromising. The reason is that a signal which can cause undesired distortion effects is usually not the signal to which the receiver is tuned. Hence, when this signal is heterodyned

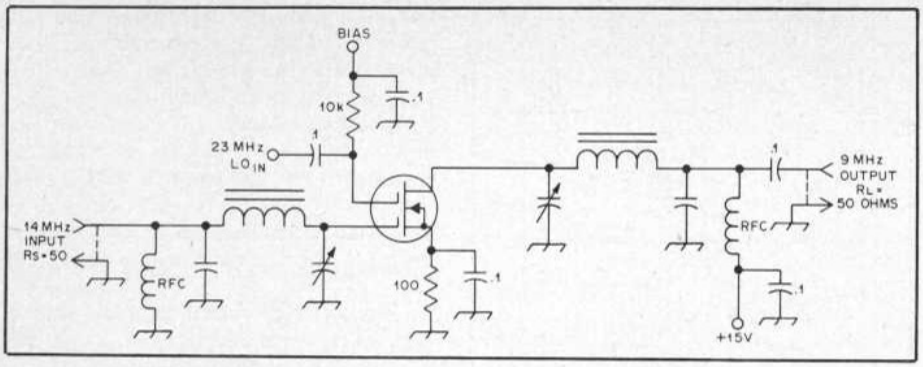


Fig. 9 – Circuit of an active mixer using a dual-gate MOSFET. The pi networks are designed to transform 50 ohms to 2000 ohms. The Q_L is 10.

in the mixer, the output will not lie within the passband of the filter. This can result in large voltage excursions at the drain, leading to blocking or IMD.

The pi network used in Fig. 9 is one of the better choices as a matching mechanism to work into a crystal filter. The reason for this is that the pi network has an impedance-inversion property. That is, if the output termination is less than that for which it was designed, the input impedance appears higher than the design center. On the other hand, if the output termination appears high in impedance value, the input seen at the drain is low. The latter situation is desired. When the input impedance of the crystal filter appears to be an open circuit (out of the passband), the load presented to the drain approaches that of a short circuit. This prevents large voltage excursions.

Sabin suggested the use of another type of impedance inverting network (*QST*, July, 1970). He used an undercoupled double-tuned circuit. This kind of network has the advantage that it acts as a bandpass filter. This protects the crystal filter and following circuits from spurious filter responses that sometimes occur.

There is another mixer output that might be investigated as a possible source of IMD – the image. In the circuit of Fig. 9, the LO frequency is 23 MHz, and the input is at 14 MHz. The desired i-f is 9 MHz. However, the mixer will produce not only difference frequencies (23 – 14) but sums also, in this case at 14 + 23 = 37 MHz. It is possible that the existence of these currents in the drain would degrade the output intercept. No experiments were performed to achieve a proper termination for this frequency.

There is a problem with pi-network matching that has not been mentioned. Although the network has the advantage of presenting a proper load to the drain of the MOSFET in order to minimize blocking, it does not provide an output that terminates a filter properly. The output impedance of the FET is much higher than the 2000-ohm value for which our network was designed. It may be as high as 100 kΩ. If the pi network was designed for a value this high, the conversion gain would be very high, but the output intercept and blocking level would be degraded severely. As a result of the need for filter termination, it is common practice to put a resistor within the output-matching section. This resistor will absorb part of the available output power, with degradation of the output intercept as well as reduced gain.

Detailed noise-figure measurements were not performed with the test circuit of Fig. 9. However, in testing a number of receivers with dual-gate MOSFET mixer front ends, with low-loss input matching, we found that noise figures of 8 to 10 dB are common. Careful design may improve this.

The detailed performance evaluations just presented may sound pessimistic. However, this is not the case. The results are quite acceptable, especially when the ease of application of the MOSFET is considered. Some single-conversion receivers using such a front end were evaluated. They displayed a two-tone dynamic range of over 90 dB, which is better than most commercially available units. Receivers with poorly applied MOSFET mixers often have a DR as low as 60 or 70 dB.

Diode Mixers

Next to the dual-gate MOSFET, the most common mixers in amateur receivers are those using diodes. This class has a number of advantages. The first one is that they are inherently broadband. Therefore, they are applied easily to multiband designs. Another advantage is the relatively low noise figure. Most diode mixers generate very little noise. As a result the noise figure is nearly the conversion loss of the mixer. Another asset is that diode mixers display high intercept points. Finally, most diode mixers are balanced. The implications here are twofold. First, the balance has the effect of preventing energy applied to the LO port of the mixer from appearing at the i-f or rf ports. Second, certain types of noise (a-m noise) that would appear at the LO port all attenuated when they reach the i-f port, even if that noise might actually be at the i-f. Balance can also improve IMD immunity.

In spite of the virtues, diode mixers have their faults. They require high LO power in order to provide optimum performance. Proper termination of the mixers is critical, especially at the i-f port. Finally, depending upon diode

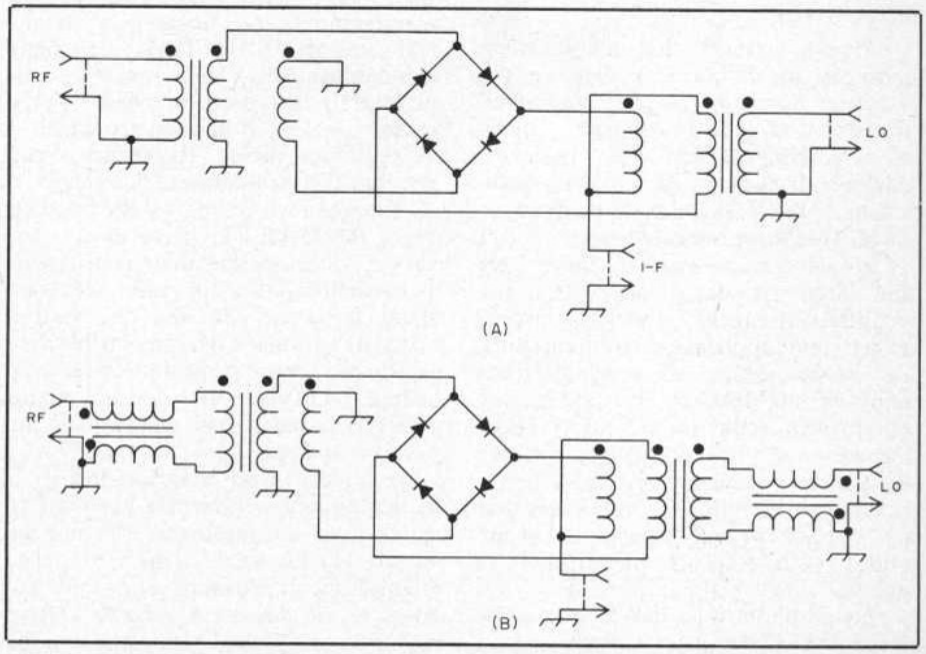


Fig. 10 – Circuit of a doubly balanced diode mixer. The diodes are HP-2800s.

type, many mixers of this kind are prone to harmonic mixing. This phenomenon was discussed in connection with diode product detectors (chapter 5).

A double-balanced diode-ring mixer is shown in Fig. 10A. The usual mixer of this type contains hot-carrier diodes, although high-speed silicon switching diodes are used sometimes. The most critical detail in building a mixer of this kind is in the winding of the transformers. The characteristics of the transformers will be the main factor that limits the bandwidth of the mixer. The balance (the ratio of the power at one port which appears at one of the others) will depend upon the transformer quality and upon the uniformity of the diodes.

If a diode-ring mixer is built to cover the hf spectrum, the transformers should be wound on high permeability ferrite toroids. A typical transformer would contain 10 trifilar turns of No. 30 enameled wire on an Amidon FT-37- 43 core. It is useful to employ wires of three different colors. If this is not possible, care should be used to ensure that the proper windings are identified: The section in chapter 4 on transformer design should be consulted.

If a mixer is built to cover the vhf or lower uhf spectrum, cores with low permeability are often used. A typical value might be 125 (Q1 material, or Amidon type-61). Toroids do not always present the optimum geometry for such applications. Excellent mixer transformers can be built using ferrite beads with multiple holes.

In applications where good balance is desired over a very wide bandwidth, it is useful to add another transformer or two. This is realized by driving each balanced port with an isolating "sorta-balun." This scheme is shown in Fig. 10B. Balance of 60 dB or more in the hf region is not unusual.

Several mixers of the simple ring configuration (Fig. 10A) have been investigated experimentally. These included homemade mixers and commercially available units. There is no significant difference between the two except in cases where extremes of balance or bandwidth are desired.

When using HP-2800 hot-carrier diodes and transformers like those just described, the typical conversion loss is 6 to 7 dB. This value is constant over most of the mixer bandwidth, reaching higher levels at very high and very low frequencies. Although the signal-handling capability of each mixer will differ, a good rule of thumb is that the output intercept of simple rings is roughly equal to the level of LO power applied. This is extremely important in the design of wide-dynamic-range receivers. Most diode mixers will achieve close to minimum conversion loss with as little as one or two milliwatts of LO power. However, for best IM performance, it is wise to increase the LO power to +10 to +13 dBm, or even more if the diodes will handle the larger currents.

The measurements of output intercept outlined in the foregoing were obtained with a test setup like that used for the evaluation of the dual-gate MOSFET, with the i-f port terminated in the 50-ohm input of a spectrum analyzer. A common result with simple ring mixers was an output intercept of +15 dBm with an LO power of +13 dBm. After the initial measurements were performed with broadband terminations at all ports, tuned circuits were inserted in various lines to the mixer. These were single-tuned *LC* circuits. The results were profound! When a single tuned circuit was put in the i-f port it had the effect of still presenting a 50-ohm termination at the desired i-f of 9 MHz. (The rf energy was at 14 MHz and the LO was at 23 MHz.) However, at frequencies other than the 9-MHz i-f, the impedance seen was highly reactive. This had the effect of decreasing the output intercept from +15 dBm to +5 dBm in several of the mixers studied. The conversion loss did not change significantly.

When a narrow-band termination was used at the rf and LO ports of the mixer, a degradation in output intercept was also observed. However, it was not nearly as severe as that seen at the i-f port.

The critical frequency that must also be terminated in the diode mixer is the image. In the case outlined, this would be the sum of the rf and LO frequencies, or 37 MHz. If this energy is not absorbed in a resistive termination, it may be reflected back into the ring where it can interact with existing signals to produce IMD.

There are two general approaches to this termination problem. One is through the use of attenuators. A 3- or 6-dB pad is often used at the output of the mixer to ensure a broadband termination. Unfortunately, this attenuation adds directly to the noise figure of the mixer. A more satisfactory solution is to terminate the i-f port in a diplexer.

A diplexer is a network of resonant circuits that is arranged so that the desired frequency is passed through the network with minimal attenuation. However, additional inductors and capacitors are arranged so that other frequencies are terminated also. That is, the network has an input impedance which is close to 50 ohms at all the frequencies of interest. Two possible configurations are presented in Fig. 11. The first is a combination of bandpass filters. C1, C2, C3 and L1 form the single-pole bandpass filter. At frequencies other than the 9-MHz design center of the filter, the input impedance will be capacitive. The out-of-band energy is handled by R1, C4 and L2. The inductor and capacitor are also resonant at 9 MHz. At the i-f frequency they appear as a high impedance. Minimal current flows in R1. When the frequency departs from 9 MHz considerably, L2 and C4 appear as a low-impedance path to ground. Now R1 is directly across the mixer output, providing a proper termination.

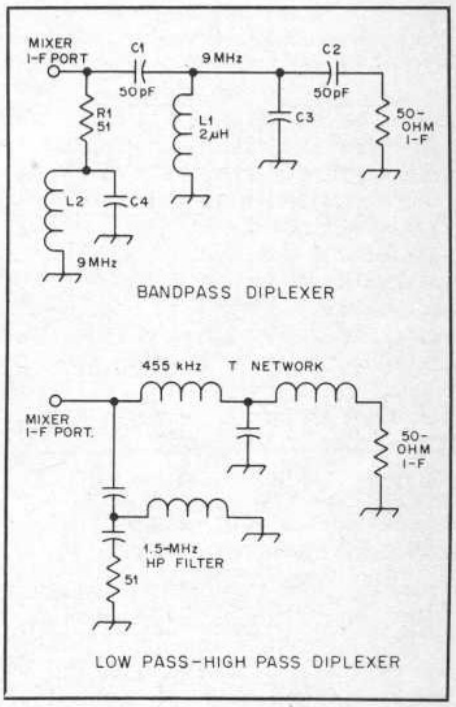


Fig. 11 — Diplexer circuits for use after a mixer.

The other diplexer shown uses a combination of a low-pass and a high-pass filter. The circuit operates in a similar fashion to the bandpass design just described and is especially useful in receivers using low intermediate frequencies (such as 455 kHz). The low-pass filter is a T network cut for the i-f of interest. The high-pass filter should be designed for a 50-ohm characteristic impedance and a cutoff frequency of about three times that of the i-f. Such a filter has reactances equal to the characteristic impedance at the cutoff frequency.

Some measurements of noise figure and IMD suggest that the termination of a diode-ring mixer at the i-f may not be as critical as the image termination. This leads to the possibility of accepting some compromise in match at the i-f in

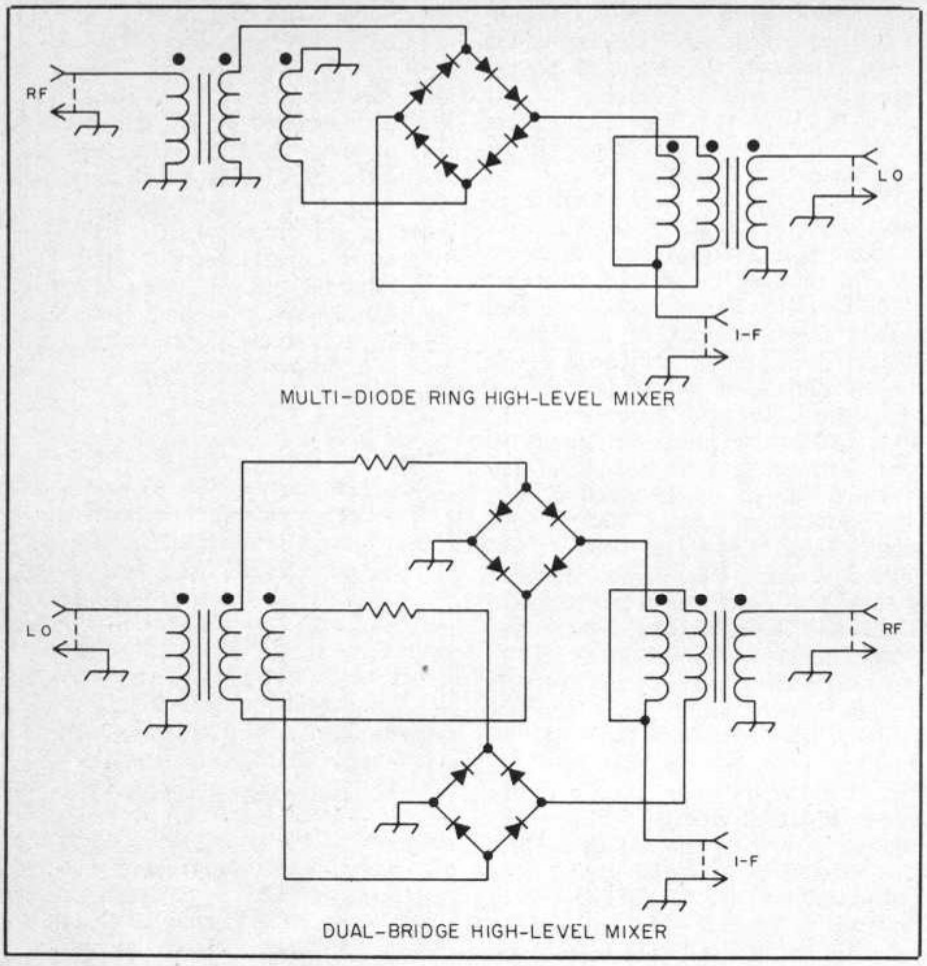


Fig. 12 — Examples of multi-diode high-level ring and dual-bridge high-level mixers.

order to obtain improved system noise figure. An example would be a dual-gate MOSFET low-noise amplifier following the mixer. In this case it would still be necessary to provide proper termination for the image energy, still making a diplexer desirable. In stringent designs, all products resulting from harmonic mixing should be terminated. Such considerations emphasize the need for doing broadband designs with good matching well into the vhf spectrum, even when the receiver is for use on the hf bands.

There are other diode mixers that offer improved intercept characteristics with virtually no compromise in noise figure. Two of these are shown in Fig. 12. In the first, the single diodes have been replaced by a series combination of two diodes. This helps the mixer to accept higher LO power without burning out the diodes. In the ring configuration, even when multiple diodes are used, the limitation is the maximum current that the diode can handle. Reverse voltage breakdown is not a problem, since each pair of conducting diodes protects the reverse-biased ones. In multidiode mixers (more than 4) designed for high intercept factors, care should be taken to ensure that the diodes are well matched. Diodes with a high junction area are desired also, since they will handle larger currents.

The second mixer (Fig. 12) departs from the ring configuration: A pair of bridge rectifiers is used. The local oscillator is applied to each bridge in parallel. However, the bridges are arranged with respect to the LO transformer so that only one is "on" at one time. Each bridge conducts on alternating half cycles of the LO waveform. The bridge that is on at any instant connects that end of the rf-port transformer to ground. The opposite side of the rf-port transformer is, in effect, connected directly to the i-f port.

One unusual characteristic of the second mixer of Fig. 12 is that resistors appear in the local oscillator lines to each bridge. These resistors cause significant effects. They allow the LO port to be driven with higher voltages than would be possible otherwise. This not only leads to high currents flowing in the diodes during their "on" half cycle, but it allows a larger reverse voltage to be established across the "off" bridge. This causes the diodes to operate in nearly a true switching mode. Note that the resistors are not in the rf to i-f path.

The classic diode ring is analyzed best if the diodes are thought of as switches that are controlled by the LO signal. In this condition, an incoming rf signal is "chopped" at the LO rate. A mathematical analysis will show that this leads to sum-and-difference frequencies. Detailed study indicates that the IMD effects which limit the intercept are a result of departures from the switching action. If a weak sine-wave drive is used at the LO port, the diodes will spend a portion of each cycle near a zero-bias condition. Because of this, strong rf signals can have a major effect in changing the conduction state of the diodes. On the other hand, if the mixer is driven with a much stronger LO, and ideally even a square wave, the diodes are allowed to spend a much shorter portion of each cycle near this zero-bias point. The stronger mixers are those that allow large LO signals to be applied, and permit larger reverse voltages to appear across nonconducting diodes.

Both of the mixer types described have been studied by the writers. The original designers of these mixers are not known to the writers. Both have been outlined in recent papers (see the bibliography: Cheadle, 1973, and Rohde, 1975) although only the multidiode ring is described in detail. Our measurement results were virtually identical for the two mixers. The insertion loss was about 6.5 dB and the output intercept was +22 to +23 dBm. The frequencies were the same as those used in the other evaluations. It was found that the dual-bridge mixer exhibited extremely good balance, up to 60 dB in the hf region.

One problem that was noted with both high-level mixers was that they are not always "well behaved." This means that the intermodulation distortion products did not always drop by 3 dB when the input tones were decreased by 1 dB. Although it is conjecture at this point, this departure could result from mismatch in the diodes at specific current levels, or from nonlinearities in the ferrite transformers. The intercepts quoted are indicative of the well-behaved range of operation with an LO

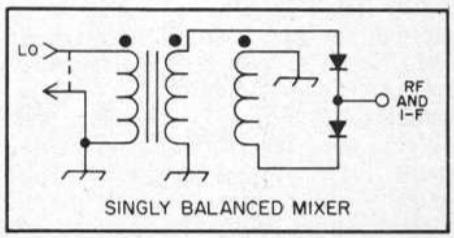


Fig. 13 — A singly balanced mixer which uses two diodes.

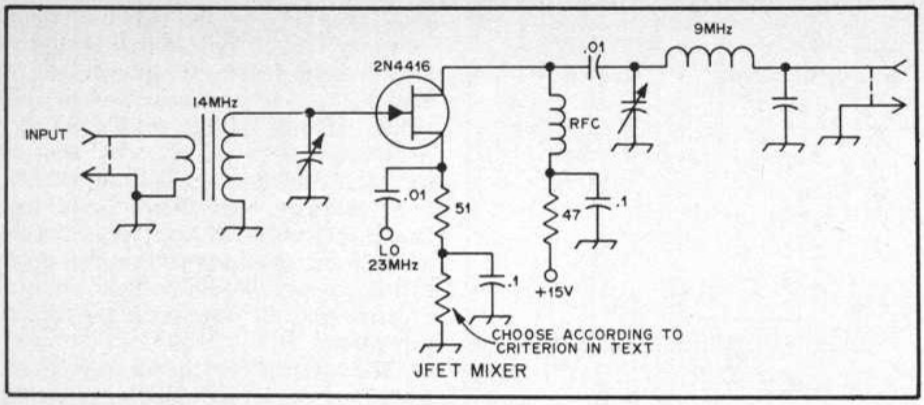


Fig. 14 – Circuit of a JFET mixer.

drive of +17 dBm.

All of the diode mixers discussed have been doubly balanced designs. That is, balanced transformers have been used at two of the three ports. However, it is not mandatory that a mixer be doubly balanced in order to assure strong performance. Shown in Fig. 13 is a singly balanced mixer using only two diodes. This design has the virtue that large voltages can be established across the diodes in the off condition. The center tap of the LO transformer is grounded. This improves balance. If this configuration is used, the i-f and rf are applied to the connection of the diodes. A diplexer is used to isolate the two frequencies. At the lower frequencies it may be acceptable to extract the i-f from the center tap.

One virtue of mixers of this kind is that they often have a lower insertion loss than is typical of the four-diode mixers. Such a mixer has an insertion loss of 5 dB with an output intercept of +15 dBm. These results were obtained with an LO drive of +15 dBm. Two-diode mixers are popular for vhf and uhf application.

Mixers Using JFETs

Some JFETs can provide exceptionally good performance as mixers. They are, however, more difficult to use than MOSFET mixers.

Shown in Fig. 14 is a 2N4416 mixer. The properties are similar to those obtained with the dual-gate MOSFET: The input impedance is high and the conversion gain is commensurate with a transconductance of 1/4 that seen with the same device operated as an amplifier. Biasing is critical. It should be chosen so that the gate-to-source voltage is equal to 1/2 of the pinchoff voltage of the device. The local oscillator signal, which is applied to the source, should be as large as possible within the constraints that the device should never go into the pinchoff region, nor should the gate diode be driven into conduction. This means that the pk-pk LO voltage should be a little below the pinchoff voltage of the FET.

The MOSFET mixer had a high output impedance. On the other hand, a JFET has a lower value, typically around 10 kΩ for the resistive portion. This makes matching to filters a bit easier. An impedance inverting network should still be used.

The major advantage of the JFET mixer over the MOSFET is that the noise figure is lower. Values as low as 4 dB have been reported (Sabin, 1970). The writers have not done intercept measurements on this mixer.

Shown in Fig. 15 is a mixer using a dual JFET (Siliconix E430) which has been designed especially for mixer applications. The input transformers are similar to those used in diode mixers. Pi networks are used at each drain to do part of the impedance matching as well as perform impedance inversion. Each pi network is designed to transform from 2000 ohms at the drains to 100 ohms. The push-pull 100-ohm outputs add to form a 200-ohm balanced source. This is transformed to a single-ended 50-ohm output with a trifilar transformer. The LO power requirement for this mixer is fairly high, since the sources are driven rather than the gates. With a +17-dBm LO drive, an output intercept of +26 dBm was measured. The gain was 2 dB. Noise figures from 6 to 8 dB are quoted as typical by the manufacturer.

Doubly balanced mixers using four JFETs have also been described. Although the writers have not investigated them (yet!), they appear to offer great promise.

Mixer Comparisons

Great care should be used when comparing mixer designs. Many workers suggest that mixer gain is an advantage. This is not necessarily true. Compare, for example, a dual-gate MOSFET mixer with a simple diode ring. The former may have a gain of 20 dB, an output intercept of +18 dBm, and a noise figure of 10 dB. If a receiver is built with this mixer as the front end, driving a filter directly, the MDS will be –137 dBm. A bandwidth of 500 Hz was assumed. The input intercept of this receiver will be +18 dBm – 20 dB, or –2 dBm. Recalling that DR = (2/3) (P_i – MDS), the dynamic range of the system will be 90 dB.

Consider now the simple diode ring with a conversion loss of 6 dB. Assume that the circuit following the ring is strong, has a noise figure of 3 dB, and that a preselector filter with a 1-dB loss is used ahead of the mixer. The overall system noise figure will again be 10 dB, leading to an identical MDS of –137 dBm. The input intercept of the mixer will be the output intercept *plus* the conversion loss. Assume that the output intercept is +15 dBm. The receiver input intercept will now be +15 dBm + 6 dB

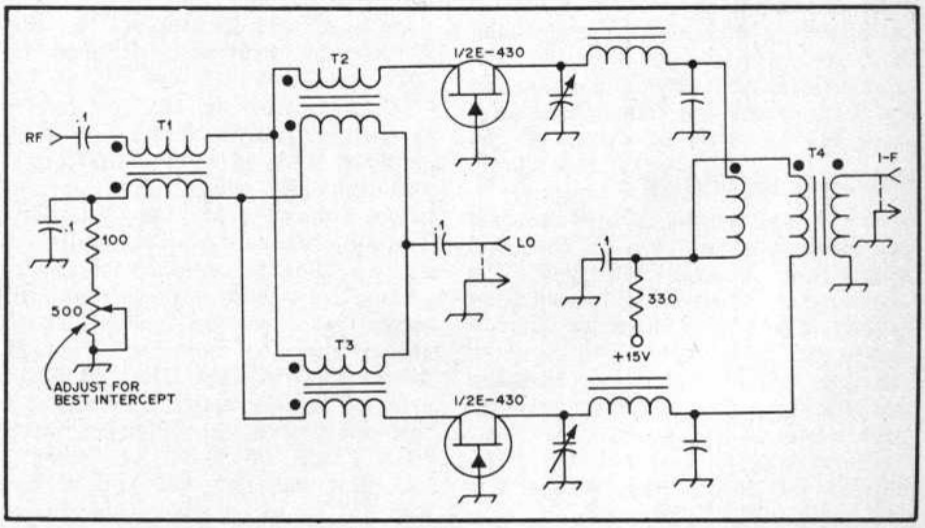


Fig. 15 – A high-level balanced JFET mixer. T1, T2 and T3 contain 10 bifilar turns of No. 28 enam. wire on FT-37-61 toroid cores. T4 has 10 trifilar turns of No. 28 enam. wire on an FT-37-61 core.

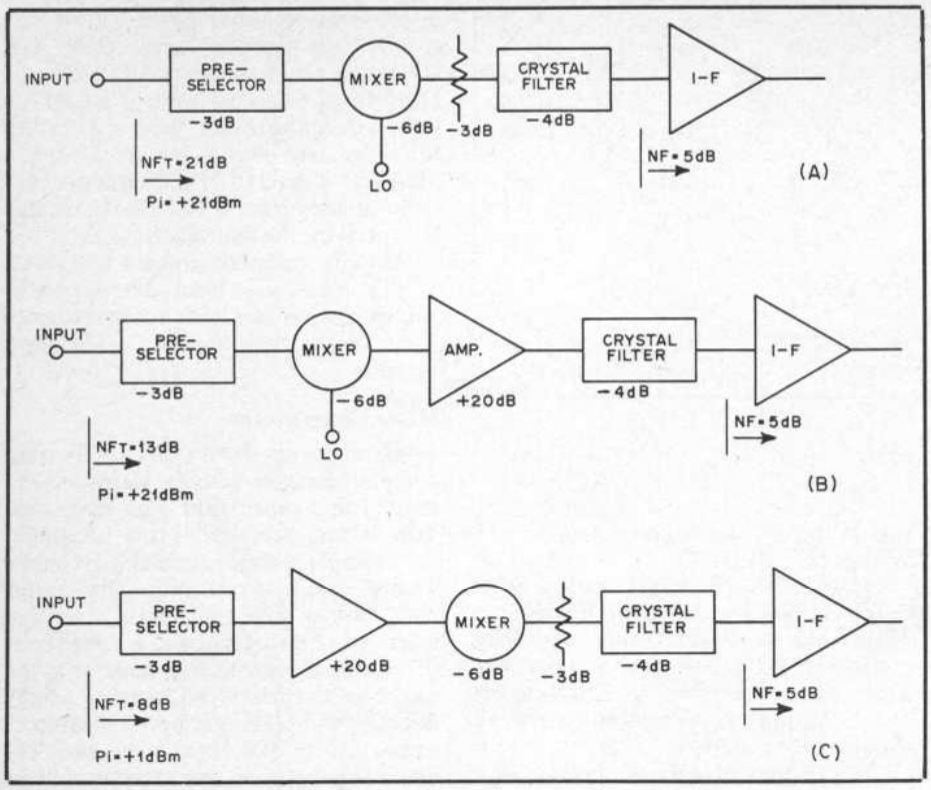


Fig. 16 — Illustration of a receiver which has no gain ahead of the filter (A). At (B) and (C), 20 dB of gain has been added.

(mixer loss) + 1 dB (preselector loss) = +22 dBm. The dynamic range of the receiver is now 106 dB! In this case, the loss of the diode mixer is a profound advantage, leading to a 16-dB increase in dynamic range.

For general applications in straightforward receivers, the dual-gate MOSFET is highly recommended. For improved performance, simple diode mixers are suggested. However, more care is required in designing the circuitry following the mixer. For the experimentally inclined amateur with instrumentation for evaluation of the circuits, high-level mixers using diodes or balanced JFETs are suggested. The advanced amateur may build equipment to do this evaluation himself (see *QST*, July, 1975).

Front-End Amplifiers

There are two major ways in which amplifiers are used in the front-end section of a superheterodyne. The classic one is as an rf preamplifier preceding the mixer. The other, which is not quite as traditional, is as an i-f amplifier following the mixer. In multiconversion systems, amplifiers are often used in between the mixers.

Consider a single-conversion receiver designed for cw operation. Assume that the bandwidth of the crystal filter is 500 Hz (27 dB above one Hz), and that a simple diode-ring mixer is used. The mixer will have a 6-dB insertion loss and a +12-dBm output intercept. Assume that the noise figure of the i-f amplifier following the crystal filter is 5 dB.

As a start, imagine a receiver that has no gain ahead of the filter. This system is shown in Fig. 16A where the preselector network is assumed to have a 3-dB insertion loss. The crystal-filter loss is 4 dB. Also, we will use a 3-dB attenuator between the mixer and the crystal filter to ensure that the output intercept of the mixer is preserved. Using the methods outlined in the earlier sections of this chapter, this system can be analyzed. The results are noise figure = 21 dB, MDS = −126 dBm, P_i (input intercept) = +21 dBm and DR = 98 dB.

Consider now the modified receiver shown in Fig. 16B. Here a very strong amplifier with an output intercept of +40 dBm and a gain of 20 dB is inserted between the mixer and the crystal filter. The input intercept of this amplifier will be +20 dBm. Since this is quite a bit greater than the output intercept of the mixer that drives the amplifier, we will assume that the amplifier is virtually free of IMD. A noise figure of 4 dB is assumed for the amplifier. Analysis of this design gives the following results: NF = 13 dB, MDS = −134 dBm, P_i = +21 dBm, and DR = 103.3 dB. We have gained 8 dB in sensitivity and about 5 dB in overall dynamic range, while leaving the input intercept constant.

The third case for consideration is shown in Fig. 16C. Here the same amplifier has been placed as an rf preamplifier between the preselector and the mixer. Analysis yields noise figure = 8 dB, MDS = −139 dBm, P_i = +1 dBm, and DR = 93.3 dB. The low noise gain has yielded an improvement in noise figure, but has brought about a dramatic decrease in input intercept and dynamic range. For most amateur applications, case B would be the optimum. Clearly, gain distribution is a vital consideration.

The criteria for the design of preamplifiers and post-mixer amplifiers differ considerably. In the case of the latter, the amplifier input intercept should exceed the output intercept of the mixer used. For the preamplifier, the output intercept should exceed the input intercept of the mixer. If these criteria are not met, IMD from the amplifiers will add to that generated within the mixer.

Post-Mixer I-F Amplifiers

Amplifiers operating at the intermediate frequency, and following the mixer directly, should have output intercepts of +30 dBm or more, In searching the literature we find that commonly available FETs, both the junction and MOS types, are not strong enough. This leaves the job to bipolar transistors. FET technology is changing rapidly, however, and there are indications that much better units will be available in the future.

Shown in Fig. 17 is an amplifier that was breadboarded for preliminary investigation. This amplifier used an Amprex BFR-94 transistor. This is a stud-mounted power device designed for cable-TV applications. No impedance matching was performed at either the input or the output. Still, at 10 MHz the transducer gain was well over 25 dB and the noise figure was about 5 dB. The output intercept of this amplifier was

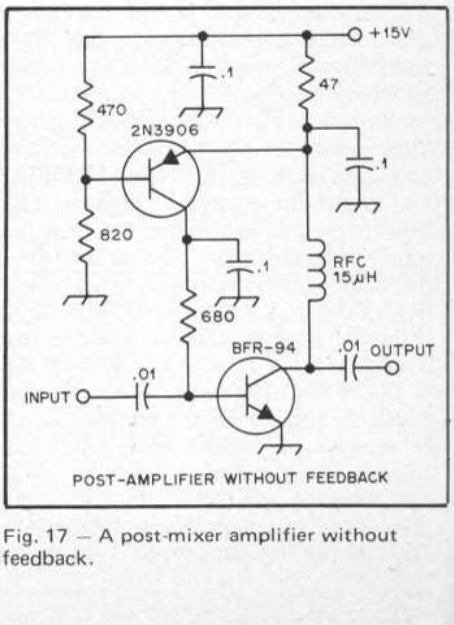


Fig. 17 — A post-mixer amplifier without feedback.

+40 dBm. The feature of this circuit is that there was 100 mA of collector current flowing in the transistor. At this level, the saturated power output of the amplifier was over one-half watt!

After the initial experiment, the BFR-94 circuit was modified. A 2:1 turns-ratio transformer was placed in the collector circuit, providing a 200-ohm collector load resistance. Also, negative shunt feedback was introduced by a 1000-ohm resistor, ac coupled from collector to base. With this modification the output intercept went up to +45 dBm. Noise figure was not measured. Input matching would be required when using the modified circuit, for shunt feedback will have the effect of depressing the input impedance well below 50 ohms.

In general, post amplifiers made from bipolar transistors will use negative feedback as well as some impedance matching at the output. An amplifier from one of the writers' receivers is shown in Fig. 18. A Nippon Electric 2SC-1252 transistor is biased to 65 mA of collector current. A 2:1 turns-ratio ferrite transformer was used at the output, presenting a load of 200 ohms to the collector. Emitter degeneration and shunt feedback were employed. This combination has the result of controlling stage gain as well as the input and output impedances. Without the 6-dB attenuator in the output, the amplifier provided 23 dB of gain and an output intercept of +41 dBm. The noise figure was 6 dB at 10 MHz, and the input match to 50 ohms was excellent (30-dB return loss over the hf spectrum). NEC transistors are available from California Eastern Labs of Burlingame, California.

A 6-dB attenuator is included in the output in actual application. This has the effect of reducing the net gain to 17 dB and dropping the output intercept to +35 dBm. However, it has the asset of keeping the input impedance of the amplifier relatively constant at all frequencies. If it were not there, variations in input impedance of the crystal filter that follows the amplifier would reflect back through the amplifier to cause variations in the input impedance. This characteristic is typical of amplifiers with heavy shunt feedback. Additional information on the design of negative-feedback Class A amplifiers is presented in connection with our discussion of ssb methods.

The NEC transistor was a convenient unit to use. It is mounted in a TO-5 package. However, unlike most TO-5 devices, the collector is not common to the case. There is good internal thermal bonding, nonetheless. In our application a suitable hole was drilled in the circuit board allowing the transistor to be soldered to the ground foil: The board serves also as a heat sink.

A general equation may be applied to bipolar transistors to estimate their output-intercept characteristics. It is assumed that the collector is terminated in a 50-ohm load. Under these conditions, the output powers in dBm for 1 dB of gain compression, and for IM intercept, are given by

$$P(\text{compression}) = -16 + 20\log_{10} I_c$$

$$P_o = 20\log_{10} I_c = \text{output intercept} \qquad \text{(Eq. 13)}$$

where I_c is the collector current in mA. These equations should be regarded as an optimistic rule of thumb rather than as an absolute definition of the performance. The intercept may often be improved by impedance matching to the collector. This was the case in the 2SC-1252 amplifier. Deriving the equation for gain compression is straightforward: The output power is that where the peak signal current is equal to the standing dc current. It is surprising to the writers that these simple relationships are so accurate in practice.

There are some general requirements for the choice of transistor types for amplifiers of this kind. From the equations we see that a high output intercept will result only from a high collector current in the transistor. The transistor must be capable of operating at high currents and of dissipating the power. However, a reasonably low noise figure is also desired. Usually, feedback needs to be applied. Because of these criteria, the transistor should have a very high f_T. In the two circuits presented the devices have gain-bandwidth products of well over 1 GHz. For applications with I_c of approximately 100 mA,

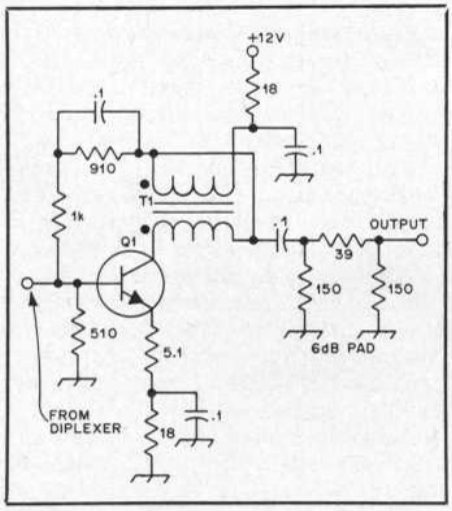


Fig. 18 — A bipolar type of post-mixer amplifier which uses feedback.

the Amperex BFR-94 and A-209 types, the NEC 2SC-1252 as well as the Motorola 2N5947, are suggested. For amplifiers with up to 50 mA, the Amperex A-210, Motorola 2N5943 or RCA 2N5109 are suggested. Vhf power transistors are worth consideration. Examples would include the 2N3553 and 2N3866. For strong bipolar amplifiers in the vhf and uhf region, the NEC V021 is recommended. With I_c = 30 mA, this device will give an 18-dB gain and 4-dB noise figure at 432 MHz, without careful matching.

Preamplifier Design

The criteria for the design of amplifiers that precede the mixer in a superheterodyne are somewhat different than those for post amplifiers. First, the intercept requirements are not as stringent. Since the usual diode-ring mixer will have an input intercept of +15 to +18 dBm, amplifiers only a bit stronger than this will suffice. Second, lower noise figures are usually desired. Both FETs and bipolar transistors may be used.

FETs have some general advantages. Less current is required in order to realize an equivalent output intercept. Their noise figures are quite low in the hf region. Finally, their output powers for gain compression are closer to the output intercept than is the case for bipolar transistors. This means they are less prone to blocking problems.

In spite of the virtues of FETs, bipolar transistors may be used quite successfully as hf preamplifiers. They come into their own in the vhf and microwave regions. The major advantage of the bipolar transistor over the FET is that it has well defined input and output impedances and is much more easily used with negative-feedback systems. This can be of profound importance if a low-loss preselector is used ahead of such an amplifier. If preselector performance is to be maintained, the filter *must* be terminated properly. In the hf region this is not possible with FETs operating in the common-source configuration. A clean input match is realized with an FET only if a resistor is added for termination. This has the effect of degrading the gain and noise figure. This compromise may be altered with the application of advanced feedback methods.

Although the theory is beyond the scope of this text, it is possible to apply advanced feedback methods to bipolar transistors to great advantage. The results are that low noise figure and a good input and output match may be obtained simultaneously. One of our colleagues (WA7TZY) has built amplifiers using bipolar transistors with equivalent noise temperature under 100°K at 432 MHz, with input and output return

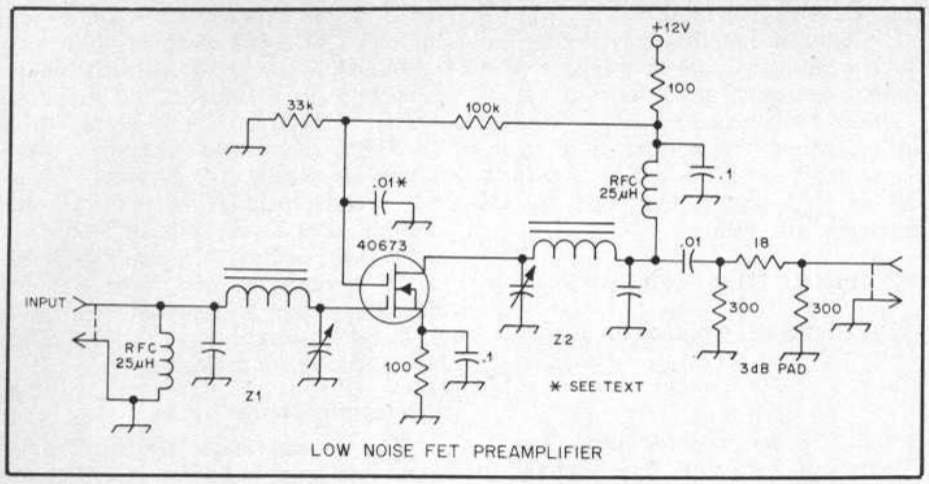


Fig. 19 — A low-noise preamplifier using a dual-gate MOSFET. Z1 and Z2 are pi networks with *Q* values of 10 or less (see text).

losses of better than 20 dB. In general, the simple resistive feedback methods shown for post amplifiers (Fig. 18) have the effect of degrading the noise figure. (See the analysis in the appendix.)

An excellent choice for general-purpose bipolar amplifiers in the hf region is the 2N5179, biased to approximately 20 mA. The Amprex BFR-91, biased between 10 and 20 mA, is excellent for the 144- and 432-MHz bands.

In spite of the input-match problem with FETs, they can have low noise figures. Shown in Fig. 19 is a preamplifier using a 40673 dual-gate MOSFET. A pi network is used for input matching, transforming the input 50-ohm source to an impedance at gate 1 between 2000 and 3000 ohms. The loaded *Q* of the network should be as low as possible if minimum noise figure is desired. Several hf amplifiers built by the writers had noise figures under 2 dB.

The MOSFET amplifier should have careful bypassing at gate 2. The capacitor should be effective up to 1 GHz. Otherwise, drain-voltage variations will couple back through gate 2 to the input. That can cause oscillations in the lower uhf spectrum. In one amplifier built for 14 MHz, an oscillation was found at 800 MHz. It was cured by placing a 470-pF capacitor in parallel with the existing .01-μF one, and by reducing the pigtails of the FET as much as possible. Reisert (W1JAA) has solved this problem by placing a ferrite bead on the gate-2 lead. He reported noise figures of under 1 dB with circuits like the one of Fig. 19, using a 40673 operated at 28 MHz (*Ham Radio*, Oct. 1975).

Shown in Fig. 20 is a pair of amplifiers using JFETs which are operated in the common-source configuration. Neutralization is used to stabilize the amplifier. Bridge neutralization has the advantage that it operates over a wide band of frequencies. The first amplifier, which uses a coil from gate to drain, provides cancellation of the effect of the gate-to-drain capacitance only at one frequency. Oscillation at frequencies outside the band of operation is still possible. Noise figures of just over 1 dB have been reported for such amplifiers in the hf region.

A common-gate JFET amplifier is shown in Fig. 21. It is claimed that such a circuit is inherently stable. This is not necessarily true, as can be demonstrated with a stability analysis using two-port network theory (see the appendix for comments on stability analysis). The spurious oscillations that might occur with the common-gate circuit are usually in the vhf or uhf region and are often cured with a small resistor in series with the drain. With clean circuit layout, instabilities in the hf region are rarely a problem. The noise figure of this circuit can be close to that of the same device operated in the common-source configuration. The available power gain is not as high, with values of 10 to 14 dB being typical. An advantage of the common-gate circuit is that the input impedance is well defined and fairly low. It is approximated by $R_{in} = 1/g_m$, where g_m is the common-source transconductance. For devices like the 2N4416 with g_m near 5000 micromho, a 200-ohm input is produced This is easily matched to 50 ohms by means of a 2:1 turns ratio ferrite transformer.

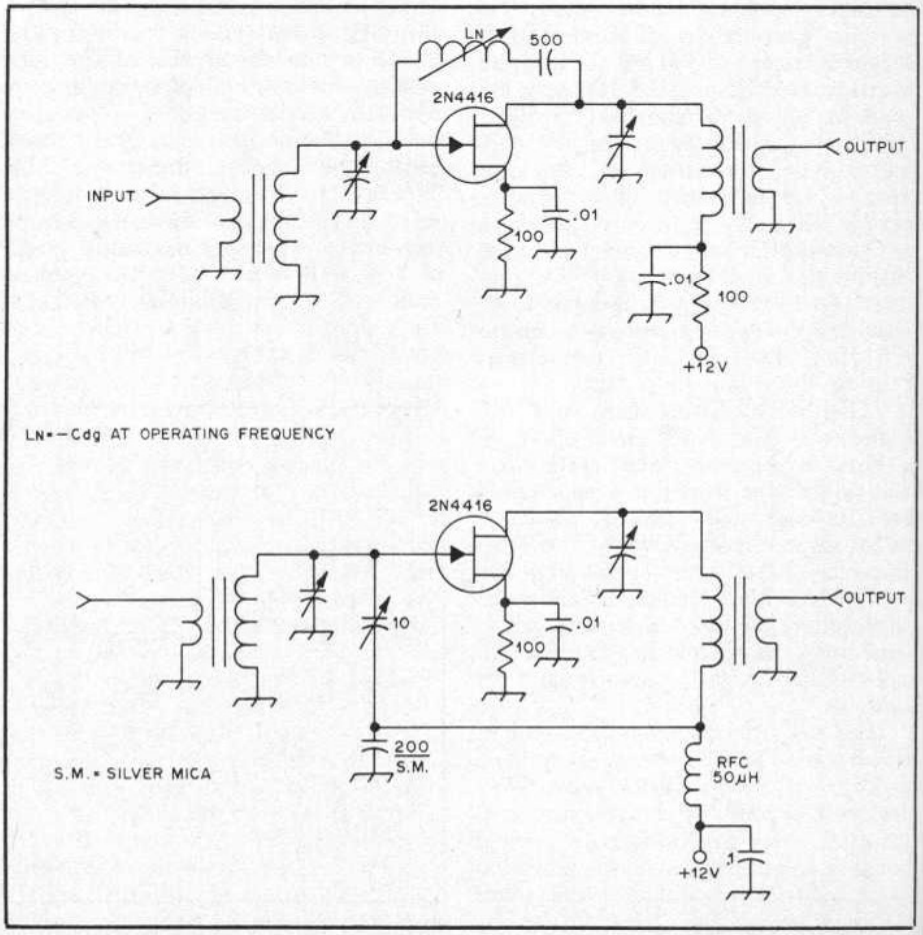


Fig. 20 — A pair of JFET amplifiers which operate in the common-source mode.

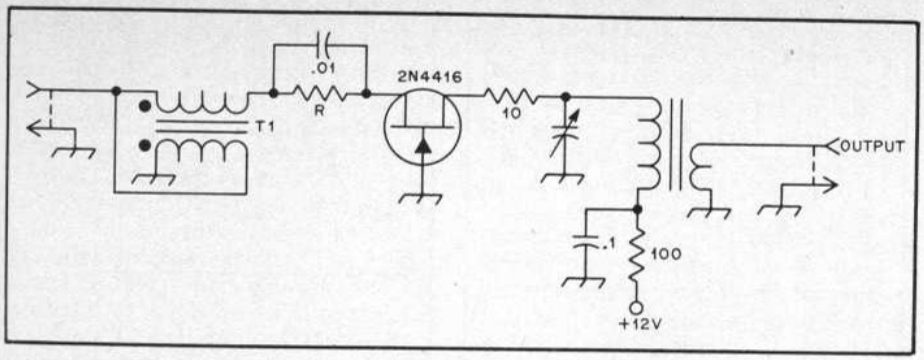


Fig. 21 — Circuit of a common-gate JFET amplifier. Typical gain is 10 dB, and output intercept is +26 dBm. Select R to provide a low-input VSWR. T1 contains 10 bifilar turns of wire on an FT-37-61 toroid core.

This would provide a good broadband termination for a preselector network. A good input match here would probably degrade noise figure.

The major point to emphasize when considering preamplifiers for hf receivers is that the gain must be chosen carefully. Excess gain will do little to improve noise figure beyond the value that is needed. However, it can have disastrous effects on the overall dynamic range of the receiver.

Oscillators for Receiver Application

The problems of oscillator stability were covered in chapter 3. A number of sample circuits were presented, many of them offering excellent long-term stability for use in transmitter applications. For the simpler receivers, these oscillators are generally adequate.

Problems appear in the design of wide-dynamic-range receivers which make the general criteria in chapter 3 (for obtaining stability) less than sufficient, and in some cases even incorrect. The performance parameter we bypassed was that of oscillator noise.

The phenomenon of noise in an oscillator output is best understood by considering how an oscillator would appear when viewed with an *ideal* spectrum analyzer. The amateur may not be familiar with this instrument. A spectrum analyzer is essentially a receiver which has been optimized for test purposes. Unlike the receivers used for communications, the output is a display on the face of a cathode-ray tube. The instrument is swept, with the tuning knob used to set the frequency of interest at the center of the CRT screen. The spectrum analyzer is a calibrated instrument, with the vertical axis representing the power delivered to the input at the frequency corresponding to the horizontal position of the display at that instant.

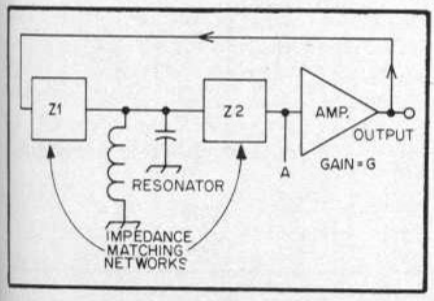


Fig. 22 — Generalized diagram of an oscillator.

When we refer to a spectrum analyzer as being ideal, we mean that it has an unlimited dynamic range and has no internally generated noise. Such instruments do not exist. We will deal with these realities later.

A generalized schematic of an oscillator is presented in Fig. 22. This circuit is the same as that given in the earlier VFO discussion and is used to examine the criteria necessary for oscillation. Reviewing the Barkhausen criteria, we recall that a signal at point A will be increased in level in the amplifier. Part of the output will be matched to the resonator by means of Z1. The signal across the resonator will be matched to the amplifier input by inclusion of Z2. A self-sustained oscillation will result if (1) the amplitude of the resulting signal at A is larger than the original, and (2) the phase of the output signal from Z2 is exactly the same as that initially impressed at point A.

Now, how would this signal appear in our hypothetical ideal spectrum analyzer? Our usual image of oscillator behavior suggests the analyzer output shown in Fig. 23. Here, there is no output at any frequency except that to which the oscillator is tuned. The shape of the response is merely the shape of the filter used in the analyzer. A more realistic picture is that shown in Fig. 24, which is much different than the one provided by the ideal oscillator.

The first difference noted is that the broadband noise is higher in level. That is, the baseline of the display is not at the bottom of the screen, but is a few dB higher. The origin of this noise can be understood if we go back to the oscillator block diagram of Fig. 22. The network, Z2, will reflect some real resistive impedance to the input of the amplifier. A noise power of kTB is thus available at the input to the amplifier. The noise power at the output of the amplifier will just be kTB multiplied by the amplifier noise factor and gain. (The details of these noise calculations were presented earlier in this chapter.) While this noise will cause problems in a receiver, it is necessary in order to begin oscillation when power is applied initially.

The second difference between the two spectrum-analyzer representations is the "noise pedestal" surrounding the carrier in Fig. 24, which was not present in Fig. 23. This noise is usually attributed to phase variations in the system. The width of the noise pedestal is equal to the loaded 3-dB bandwidth of the resonator. When the noise breaks out of the broadband noise floor, it will increase at a rate of 6 dB per octave as it approaches the carrier of the oscillator.

Consider an oscillator operating at 5 MHz with a loaded resonator Q of 10. The noise pedestal will begin at 4.75 MHz, and will drop back into the broadband noise floor at 5.25 MHz. The noise will be 6 dB above the noise floor at 4.875 MHz and 12 dB up at 4.938 MHz. Eventually, the carrier of the oscillator appears within the passband of the analyzer and dominates the display.

If a very narrow bandwidth is used in the analyzer, with some oscillators, a point may be reached where the noise increases at a 9 dB per octave rate instead of the 6-dB figure. The additional 3 dB is the result of $1/f$ noise in the amplifier.

It is interesting to study further the basic oscillator of Fig. 22. Assume that dc bias has just been applied to the amplifier. Immediately, noise will result at the output. It will be routed through the phase-shift networks and resonator where it is applied again to the input. Some filtering occurs in the resonator, so the noise spectrum is already confined somewhat. The amplified input noise is routed through the amplifier and resonator system repeatedly, always increasing in amplitude with each pass around the loop.

If we were to extend this analysis, we would predict that the positive feedback in the oscillator would cause the level of the signal in the loop to be an ever-increasing function of time. This is, of course, impossible. Something must happen to cause the amplitude of the loop signal to stop and stabilize at some finite level. There are two mechanisms that will cause this to happen: agc or limiting.

As an example of agc, consider the FET oscillator of Fig. 25. As the voltage across the tank builds up, the voltage impressed on diode CR1 will increase.

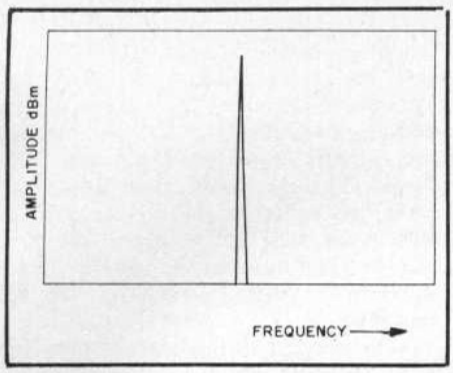


Fig. 23 — How a signal would appear on an "ideal" spectrum analyzer display.

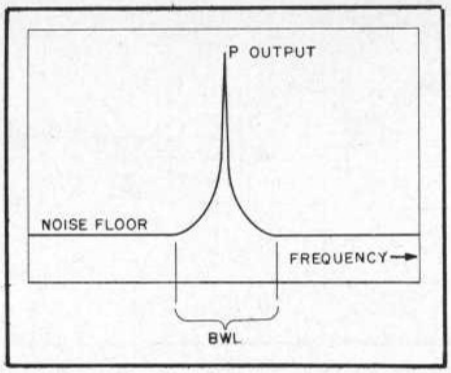


Fig. 24 — A more realistic example of that given in Fig. 23.

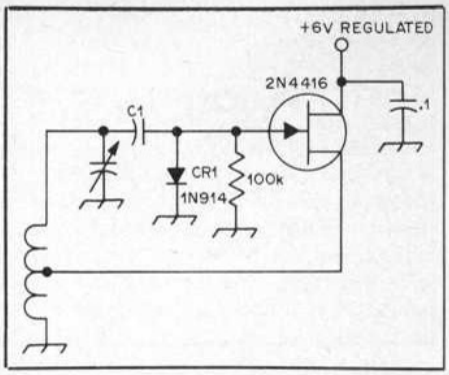


Fig. 25 — An FET oscillator.

Rectification will occur, causing a dc voltage to build up across capacitor C1. This voltage is applied to the gate of the FET and will serve as bias. As the magnitude of this bias increases, the average gate voltage becomes more negative, driving the FET toward pinchoff and thereby reducing the gain of the amplifier. Amplitude stabilization occurs when the net gain is just enough to sustain oscillation.

Limiting, as a mechanism for amplitude stabilization, is demonstrated in the circuit of Fig. 26. This oscillator was designed for low-noise performance by L. Gumm, K7HFD, and operates at 10 MHz. The voltage from the resonator, which is applied to the base, causes the collector current to change. This changing collector current is coupled back into the resonator through a link which is arranged to yield the proper phase for positive feedback. The maximum peak current that can be supplied to the link is the current standing in the transistor pair. This is defined by the emitter resistor and the inductor, which has the effect of making the current appear to originate from a constant current source. With the peak collector current well defined, the voltage across the tank is also well defined and limited.

In general, limiting is preferred over agc as an amplitude-stabilization mechanism, especially in oscillators for critical receiver applications. The reason is the same as the one which makes fm receivers immune to noise in the presence of strong signals — amplitude variations, including a-m noise, disappear from the output. This is not the case with oscillators utilizing an internal agc loop for stabilization (Fig. 25). When considering the broadband noise floor of an oscillator (Fig. 24), half of the noise is associated with random-phase variations, with the other half being attributed to amplitude variations. By the use of limiting, the amplitude noise is virtually eliminated, yielding a 3-dB decrease in the noise floor.

Additional comments about the K7HFD circuit will illustrate other features of low-noise oscillators. The collector link consists of two turns, while the base is tapped only one turn up from the cold end. Hence, the signal voltage at the base is quite large — a few volts pk-pk. This is highly desirable.

The reader will recall from our discussion of noise in amplifiers, that the degradation in output signal-to-noise ratio resulting from internally generated noise decreases as the input signal-to-noise ratio increases. The goal in an oscillator design is to maximize the output signal-to-noise ratio. Hence, a general rule of thumb emerges: The drive at the input to the amplifier should be as high as possible. In the case of bipolar- transistor oscillators, such as the K7HFD example, the only limit imposed is that the emitter-base breakdown of the transistor should not be exceeded. Not only will this lead to a degradation of transistor beta in time, but will cause extreme amounts of noise to be generated from the Zener-diode action.

The same argument with regard to emitter-base breakdown can be applied to buffer amplifiers following an oscillator. Class C operation is quite acceptable and will preserve low-noise performance as long as emitter-base breakdown does not occur.

It is important in the K7HFD oscillator that the resonator energy be re-

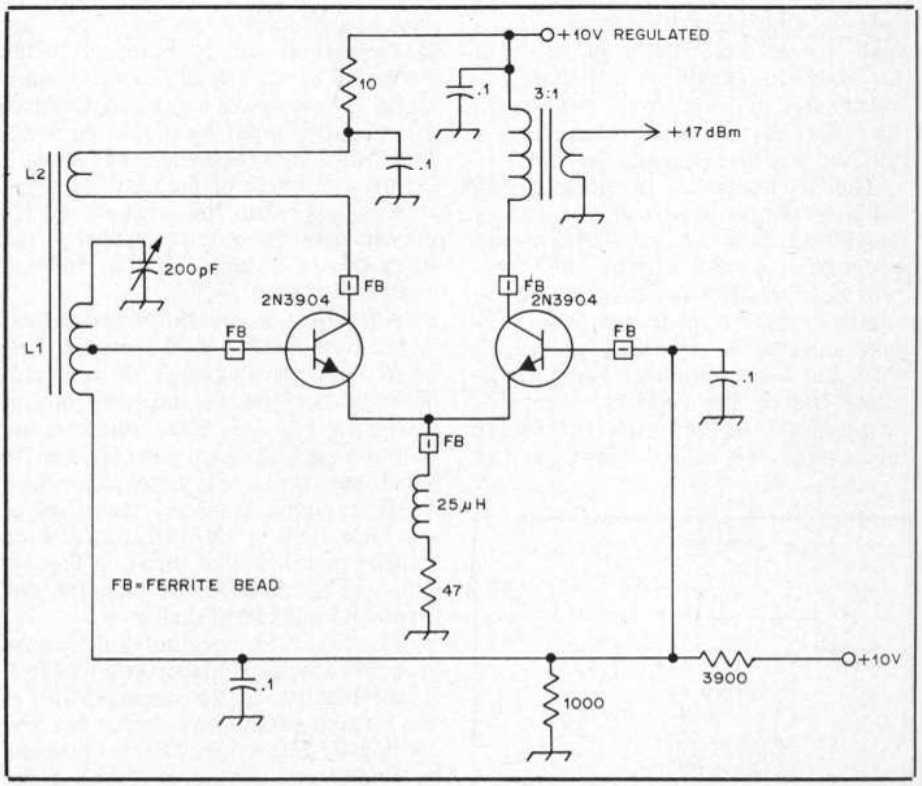


Fig. 26 — Circuit of the K7HFD low-noise oscillator. L1 is 1.2 μH and uses 17 turns of wire on a T68-6 toroid core. The tap is at 1 turn. *Q* at 10 MHz is 250. L2 is a 2-turn link over L1.

stricted by current limiting in Q1, and not by voltage clipping. Should the transistor go into saturation, the tank would be loaded severely by the saturation resistance of Q1, and would increase the width of the noise pedestal. In the configuration shown, the resonator has minimal external loading. This is due to the high output resistance presented by the collector. The loading at the base is also minimal, resulting from the extreme turns ratio used and the Class C operation of Q1. Class C operation implies that the base of Q1 extracts energy from the resonator only during a small fraction of the oscillation cycle.

The presence of saturation in oscillators using limiting is detected easily with simple equipment. If the transistor is going into saturation, the output power will change significantly as the operating voltage is varied. This does not occur with the K7HFD circuit.

Measurement of Noise in Local Oscillators

It would be straightforward to measure the level of noise from oscillators if the "ideal" spectrum analyzer were available. Unfortunately, such instruments do not exist. The better spectrum analyzers have dynamic ranges of 80 to 100 dB and are priced well beyond the reach of an amateur. Any good oscillator will have a noise floor which is over 100 dB below the carrier in a communications bandwidth. Hence, if the sensitivity of the analyzer were increased to the point that the noise could be seen, the analyzer would be overloaded. The answer to the problem is to use an existing analyzer in conjunction with a crystal filter which has a center frequency near the oscillator output frequency.

Shown in Fig. 27 is the system used for evaluation of the K7HFD oscillator. A 10-MHz filter with a 3-kHz bandwidth (6 poles) was used in conjunction with a Tektronix 7L12 Spectrum Analyzer and a frequency counter. The crystal filter had a skirt response which caused the attenuation 10 kHz away from the center to be over 80 dB. The counter was used to set the oscillator to 10.010 MHz and the output at 10.000 MHz was observed in the analyzer. Because of the attenuation of the filter, the carrier of the LO was not overloading the analyzer and the noise could be measured. The result was that the noise was over 120 dB below the output of 50 mW (+17 dBm) in a 3-kHz bandwidth, 10 kHz away from the carrier.

The results of LO noise can be observed readily in some receivers. This results from the multiplier nature of mixers. That is, a mixer is a device with an output voltage which is proportional to the product of the two input signals. If a receiver with a very steep-skirted filter is tuned to a strong carrier, a clean-sounding tone is usually heard. However, as the receiver is tuned slowly away from the carrier, a point will be reached where there is no longer a clean tone coming from the receiver. Here, the attenuation of the crystal filter has suppressed the carrier signal. A noise output is, sometimes, still present. This will be the result of the strong carrier at the mixer rf port, mixing with the noise from the LO.

It should be emphasized that the foregoing observation is based upon the assumption that the input strong carrier applied to the receiver is virtually noise-free. In a laboratory experiment this cleanliness is obtained by using a high-quality signal generator in conjunction with a narrow bandwidth (50 Hz or less) multipole crystal filter. This will ensure that the observed noise is a result of the local oscillator and not the noise output of the signal generator.

On-the-air listening experiments can be enlightening. In one series of tests at W7ZOI, a receiver using an FET oscillator was used. With a 4-pole, 500-Hz-wide crystal filter as the main selectivity element, the receiver sounded exceptionally clean. However, when a 10-pole filter with the same bandwidth was substituted, the effects of noise modulation were observed readily.

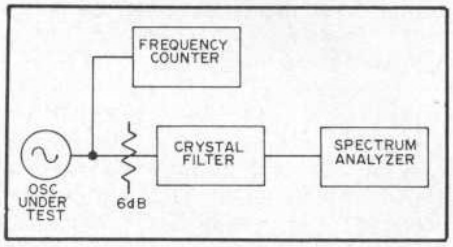


Fig. 27 – System for evaluating the oscillator of Fig. 26.

Just as signal-generator noise was critical in a laboratory evaluation, the character of strong received signals is observable. As the receiver becomes more sophisticated, it is possible to detect subtleties in signal quality that would not be noticed in a more mundane receiver.

There is one final experiment that can detect the presence of phase or frequency modulation in a receiver LO. This involves the use of a triggered audio-frequency oscilloscope, an instrument found in some amateur shops. A clean signal, such as might come from a crystal oscillator, is tuned with the receiver, and the audio output is monitored with the oscilloscope. The left side of the trace will always be clean – that's the point where the sweep in the 'scope is triggered. However, if fm noise is present in the receiver LO, the right-hand end of the trace will appear fuzzy. The time base of the oscilloscope should be set to display several cycles of audio. (Audio discriminators could be used for more exacting measurements.)

General Design Criteria

Using the above analysis it is possible to formulate a number of general rules for the design of quiet oscillators for critical receiver applications.

1) Use as high a loaded resonator Q as can be obtained. This means not only that the unloaded Q should be high, but that the external loading by the oscillator be minimal. Also, the high Q_u requirement often dictates the use of toroids which might have compromised temperature properties.

2) Drive the input to the amplifying device as hard as possible without exceeding any breakdown specifications. This also implies that the resonator should operate with high amounts of stored energy and the attendant large circulating currents. The high currents along with the first criterion will probably compromise the long-term stability, making temperature compensation necessary.

3) The transistor or FET should have capabilities to operate at frequencies very much higher than the operating frequency. This ensures not only that the device will have adequate gain, but will exhibit minimum undesired phase shift. This keeps the phase shift in the resonator and impedance matching networks (Fig. 22A) where they belong. For the same reason, single transistor or FET oscillators are preferred over those using a multiplicity of devices. (This does not preclude buffer amplifiers.)

4) While good output buffering is desirable, it is not generally necessary for receiver applications that the output be a pure sine wave as was advocated for transmitter VFOs. The reason for this is that most good mixers – that is mixers with low IMD – will create harmonics anyway. The undesired effects of these harmonics must be eliminated with proper choice of receiver i-f amplifier frequency and proper front-end preselection. With some diode mixers a square-wave LO is desired for least distortion. The LO waveform should be symmetrical, however, since an asymmetry can destroy the balance of an otherwise well-balanced mixer.

Practical Examples

There are a number of oscillators which will fulfill the foregoing criteria. How well they need to perform will depend upon the nature of the receiver being designed. Many of the simpler receivers in this book use straightforward LOs. In no case has the receiver

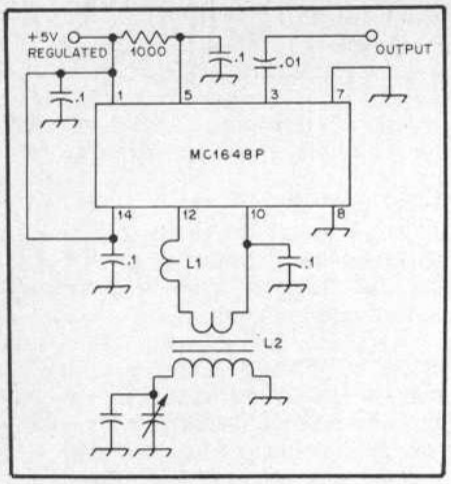


Fig. 28 — An oscillator which employs an MC1648P IC. L1 is a ferrite bead with two turns of wire. L2 is a high Q_u toroid tuned circuit. The link should contain only the number of turns necessary to sustain oscillation. A typical turns ratio is 4:1.

dynamic range been compromised by the oscillator.

In several cases shown in the text, FET oscillators have been used. When operated at low frequencies (in the 2- to 3-MHz range), they are quite suitable. If moved to higher frequencies, where it becomes harder to obtain a small loaded bandwidth, they may not be as appropriate.

Many of the FET oscillators in this book use the Clapp circuit in place of the simpler Colpitts one. This is desirable from a noise standpoint. A detailed analysis of the Clapp network shows that the stored energy in the resonator is much larger than with the usual Colpitts design.

The K7HFD oscillator used in the preceding discussion (Fig. 26) is one of the best that we have investigated. This oscillator operates with from 50 to 65 volts pk-pk across the resonator, so some temperature compensation will probably be needed. This can be done with N750 ceramic capacitors as part of the tank capacitance. Also, several ferrite beads were used in the original design in order to suppress vhf parasitic oscillations.

Shown in Fig. 28 is an oscillator using a Motorola MC1648P integrated circuit. This chip was designed specifically for oscillator applications and offers fair performance. In order to obtain the lowest noise output from this device, it is necessary that link coupling to the tank be employed. This is because the internal circuitry of the IC is such that the maximum pk-pk voltage that can be obtained across the tank terminals is about 1.4. This would severely limit the stored energy in the tank.

Link coupling between the tank and the MC1648P presents a problem that can make the chip difficult to use. This is the very high-frequency capability of the device. Because of this, the circuit is very prone to oscillate at a frequency determined by the inductance of the link and the stray capacitances. These vhf oscillations are usually killed with judicious use of a ferrite bead in series with the link. Often two turns of wire through the bead are required. A short lead length is mandatory, also.

The MC1648P has a built-in agc loop. For best spectral purity this should be defeated, and is accomplished by connecting a 1000-ohm resistor between the +5-volt supply and pin 5. If a sine-wave output is desired, a resistor connected between this pin and ground can be used. Experimentation will be required to determine the proper value.

If pin 5 is shorted to ground, oscillation will cease. This characteristic can be useful in a multiband design where several oscillators might be used, one for each band (see Fig. 29). All of the outputs may be connected directly together. Then, all of the oscillators except the one being used may be inhibited. This is easily done with a saturated-transistor switch.

The output of the MC1648P is only a little more than 1 mW, which is too low for most diode mixers. The output may be increased through the use of a broadband amplifier. This approach is used in a transceiving system described later in the book. Alternatively, the output stage of the IC may be operated at a higher supply potential. The reader should consult the Motorola literature for this application.

Since the MC1648P is capable of operation well into the vhf spectrum, careful bypassing and grounding techniques should be used. If high quality 0.1-μF capacitors are not available, the builder should use a .001-μF capacitor in parallel with the larger value shown in the figure. Double-sided pc board is recommended.

In any of the receiver LOs discussed, good power-supply regulation is needed. It is highly preferred if a separate voltage regulator be used on the pc board containing the oscillator. Special attention should be devoted to the rejection of power-supply hum. A high-gain active voltage regulator circuit is preferred over a simple Zener diode. If a Zener diode is used, it should be bypassed with a large electrolytic capacitor.

Ideally, a receiver local oscillator should be well shielded in an rf-tight box. It does little good to carefully preselect and shield a receiver front-end, only to end up with spurious responses resulting from vhf signals finding their way to the mixer along the LO line.

All of the arguments outlined here apply equally to BFOs used to drive a product detector. Good noise characteristics can be achieved easily with a BFO by using crystal control. The high unloaded Q of a crystal eases the design considerably. Generally, any of the crystal-oscillator circuits described in chapter 2 are suitable, although shielding and decoupling requirements still apply. Examples of tunable BFOs are given in several of the construction projects in the book.

Crystal-Controlled Converters

Often it is desired to extend the tuning range of a receiver to bands other than those covered by an existing receiver. This is done easily by the addition of a crystal-controlled converter ahead of the receiver. All of the basic concepts outlined in previous examples will be presented and some philosophy will be added on our approach to the design of high-performance vhf converters.

Shown in Fig. 30 is the block diagram of a typical converter. A preselector network is used at the input, tuned to the band of interest. The output of this is applied to an rf amplifier and then to another filter. The second filter is important in order to keep noise at the image frequency from reaching the mixer. Consequently, this filter is often called an "image-stripping filter." The resulting signal is applied to a mixer. The mixer is driven by a crystal-controlled oscillator in order to provide stability and frequency accuracy. If a tuned output is used for the mixer, a multipole bandpass filter is a good choice if a wide tuning range is to be covered.

In many situations the rf amplifier is not needed. This will depend upon the noise figure desired. Furthermore, in some converters it is desirable to dispense with the rf amplifier, but to include a post-mixer amplifier. This is done to preserve dynamic range of the overall system.

Shown in Fig. 31 is the schematic of a simple converter for the 160-meter band. At 1.8 MHz the noise levels are

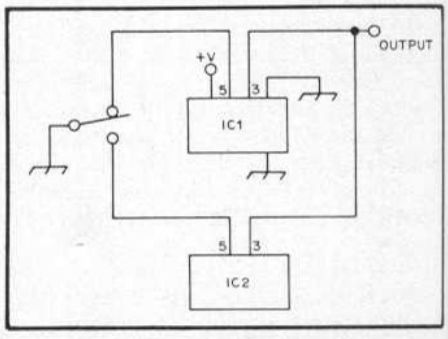


Fig. 29 — Method for easy band switching of MC1648P oscillators.

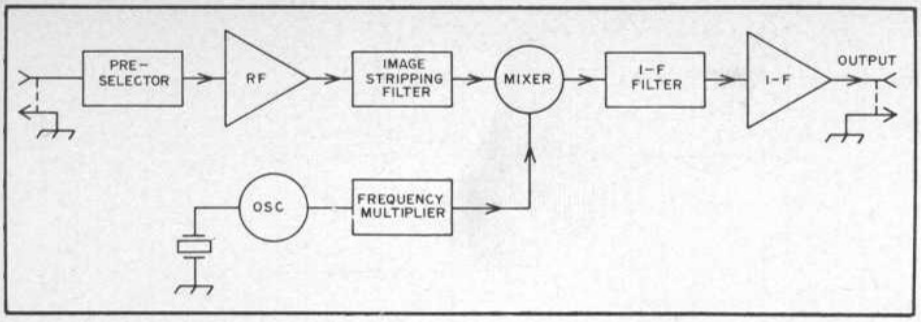


Fig. 30 — Block diagram of a typical crystal-controlled converter.

extremely high. As a result, it is pointless to strive for a low noise figure. Because of this, no rf amplifier is used, and the preselector is adjusted for a loaded *Q* of near 200. The output of the dual-gate MOSFET mixer is at 7 MHz. Although simple, this converter has performed well on "top-band." No spurious responses from broadcast stations have been detected, and the dynamic range has been adequate for some contest operations. All continents except Africa have been received with this unit from Oregon, indicating adequate sensitivity.

Shown in Fig. 32 is a simple converter for the 6-meter band. In this case, a diode-ring mixer is preceded by a two-pole bandpass filter. The preselector was adjusted for a bandwidth of 1 MHz and had an insertion loss of 1 dB. The output of the diode ring is applied to a low-noise 14-MHz amplifier (see Fig. 19), and then to the receiver used as the tunable i-f. The oscillator operates with a 36-MHz third overtone crystal and delivers +13 dBm to the diode ring. Careful measurements have not been performed on this converter. However, the noise figure appears to be about 10 dB. The sensitivity is adequate to hear background noise when using a 2-element Yagi antenna. Of major significance is that there are no spurious outputs from channel 2 TV, even though the converter is used in a strong signal area. The usual level of channel 2 on the 2-element Yagi is 0 dBm. One spurious response resulted from a local fm broadcast station. Its signal was converted to the 14-MHz band as a result of third-harmonic conversion in the diode-ring mixer. This response was eliminated by adding a low-pass filter.

A similar approach to converter design is presented in a later example. This family of converters is used to extend the coverage of a high-performance 160-meter receiver to the high-frequency bands.

VHF Converters

A popular application of the crystal-controlled converter is for reception in the vhf and uhf bands. Most converters used one to three stages of rf amplification, an active mixer, and often a post-mixer amplifier. The local-oscillator injection voltage was developed with a low-frequency crystal and a frequency-multiplier chain. Usually, the circuitry was contained on an open chassis.

While converters of that type were satisfactory once, times have changed. The vhf spectrum has become more heavily used. As a result, dynamic-range considerations are more important today than before. Furthermore, current interest in the reception of very weak signals, such as those encountered in moonbounce communications, places a severe constraint on noise figures. The following guidelines are offered for the design of high-performance vhf converters. While each point will not be justified, the reader will see that they are all consistent with the design information presented for hf receivers.

1) Use the highest frequency crystal in the LO that can be purchased. For example, if a 2-meter converter is built for 28-MHz output, use a 116-MHz crystal. If a frequency-multiplier chain is necessary (for example, a 432 converter), use balanced multipliers and extensive output filtering. All subharmonic components should be attenuated at least 60 dB.

2) Use diode mixers. Make sure that they are performing as desired. Provide diplexers at the i-f port to ensure image termination.

3) Use low-noise methods at the i-f to provide a reasonably low system-noise figure at the mixer input.

4) All rf amplifiers should be in separate, well-shielded containers with coax cables for interconnection. This will allow each stage to be matched and optimized individually. Use broadband techniques so that system stability is maintained at all frequencies.

5) Use as much preselection as possible. The input filter should have at least two poles, and the insertion loss should not exceed 1 dB. The image-stripping filter can have higher loss, but should have good stopband rejection. Helical resonators are recommended for the 2-meter band, while interdigital filters are suggested for frequencies above 432 MHz.

6) Use extensive interstage shielding and decoupling of power supplies. Each stage should be packaged in its own container. High-quality feedthrough capacitors should be used for power supply connections.

The techniques outlined are typical of those used in the communications industry. This is especially true for the construction of receivers for deep-space work, or for high-performance vhf and microwave instrumentation. Many of

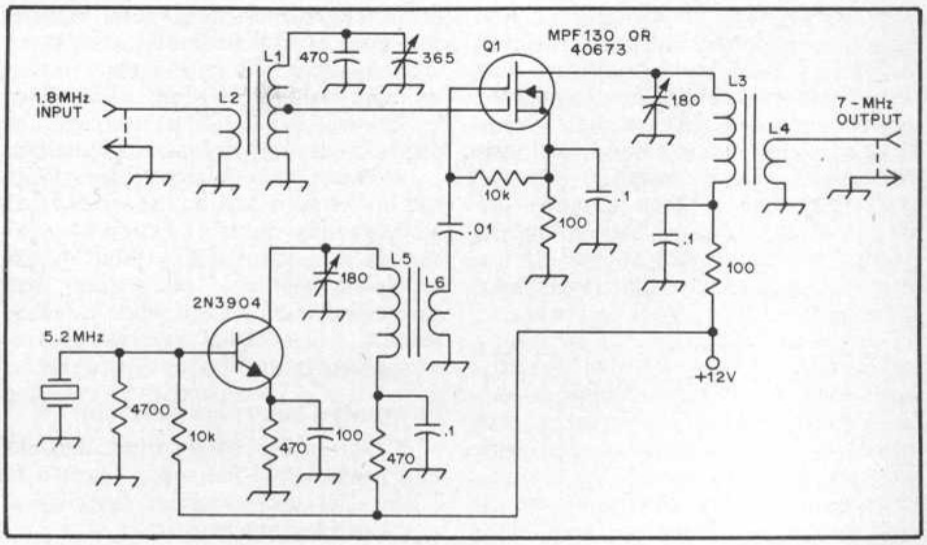


Fig. 31 — Circuit for a simple 160-meter converter. L1 has 33 turns of No. 22 wire, center tapped, on a T106-2 toroid. L2 is a 1-turn link. L3 has 30 turns of wire on a T50-2 core. L4 is a 5-turn link. L5 has 40 turns of wire on a T50-2 toroid, and L6 is a 7-turn link. Q1 can be an MPF130 or a 40673.

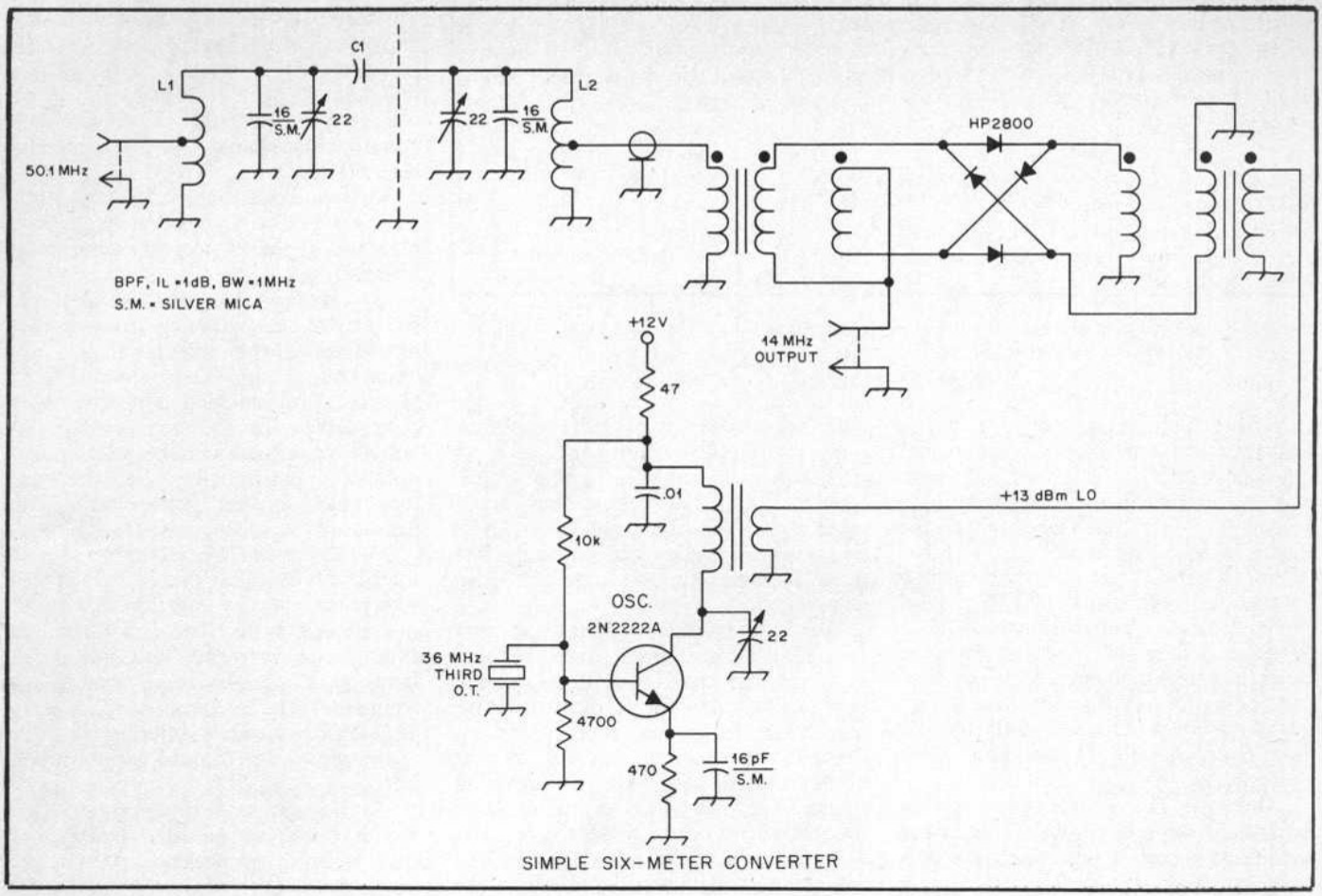


Fig. 32 — Circuit of a simple 6-meter converter. L1 and L2 have 8 turns of No. 18 wire, have an ID of 3/8 inch, and are 1 inch long. Tap at 1 turn. C1 is 0.3 inch of RG-174 coaxial line (*C* = approx. 0.5 pF).

the suggestions can be ignored for casual applications. However, spurious responses may result.

Digital Frequency Readout

A problem that has plagued the receiver builder was the construction of a frequency-readout mechanism. Not only were accurate and attractive dials difficult to build in the home shop, but they often caused the circuit design to be compromised. For example, some builders elected to build a dual-conversion receiver instead of a single-conversion one — they regarded the virtues of a linear tuning scale with good resolution and accuracy to be worth the resultant degraded dynamic range. Such a compromise is no longer necessary.

A modern approach to frequency readout is the use of digital circuitry with electronic display. Additional circuits are required. However, mechanical construction problems are avoided. With a digital readout there is no need to couple a dial to the main tuning capacitor. Linearity of tuning is of little consequence. Long-term stability requirements may even be relaxed. While a moderate amount of circuitry is needed to realize a digital readout, the design is straightforward and construction is elementary.

The virtues of digital readout do not come without a penalty. High-speed digital logic can create a large amount of rf noise. Some of this noise is broadband in nature, while some is related to the discrete clock frequencies used in counters. Special precautions must be taken to keep this noise from creating spurious responses within the receiver.

We will not attempt to cover in depth the theory of digital-logic design. There have been innumerable articles and books published on the subject (see bibliography). In this section we will confine our discussion to those details which are applicable to receivers. The barest fundamentals will be reviewed. A receiver using digital readout is presented later in the book.

Frequency-Counter Fundamentals

Shown in Fig. 33 is a block diagram of a fundamental frequency counter. It consists of two sections: the signal counter and a time base.

A time base consists of a crystal-controlled oscillator (often at 1 MHz) and a frequency divider. The circuit of Fig. 33 employs a division ratio of 1000. This produces an output of 1 kHz. The output of this divider is divided again by 2, yielding a string of pulses which are 1 ms wide. This signal occurs at point A in the figure.

The 1-ms-wide pulse is applied to an AND gate. The other input to the gate is the signal to be counted. Assume that the incoming frequency to be counted was 1.2 MHz. In a 1-ms period this signal will undergo 1200 complete transistions. If the counters that follow the gate are set to 0 prior to application of the output of the gate, they will count up to 1200 during the 1-ms "timing window." One decade counter is labelled LSD, standing for least significant digit. The last counter in the string is the most significant digit (MSD).

The outputs of the decade counters are in a binary-coded decimal (BCD) format. There are four lines which can each take on a digital 0 to 1. The BCD outputs are applied to elements termed "latches." These are memory elements. Each IC package is actually a quad latch with one memory element for each BCD line from the counters. When a "strobe" line on the latches is activated with a positive voltage, the logic state present at the latch input is connected to the

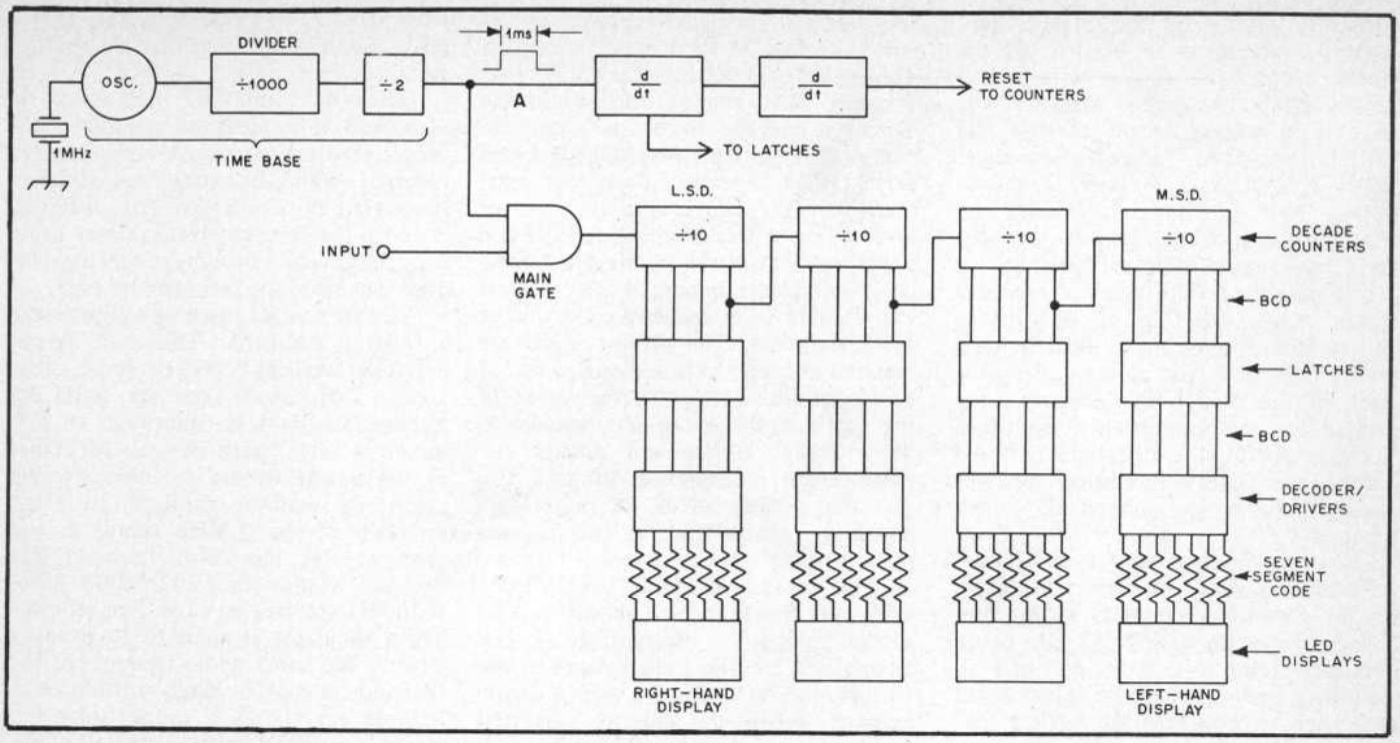


Fig. 33 — Block diagram of a fundamental frequency counter.

output. When the strobe input again goes low, the information in the latch at that instant is retained. A signal to strobe the latches is derived from the 1-ms time-base pulse. The trailing edge of the gate timing window is differentiated. This leads to a short pulse that follows the gate-control pulse.

The output of the strobe pulse is also differentiated. This leads to another short pulse which follows the strobe action. This pulse is applied to the counters to reset them to 0, making them ready for the next burst of input data.

The latches "remember" the state of the counters at the end of the counting period. The latch outputs are applied to ICs called decoder/drivers. They serve a dual function. First, they convert the BCD information supplied from the latches to the appropriate format to drive 7-segment light-emitting diode (LED) displays. Second, they provide enough output power to drive the LED displays.

In the example described in Fig. 33, four digits were displayed, and a 1-ms timing window was used. The display was updated once in each 2-ms period. If the 1.2-MHz input was measured, the display would read "1200." The readout is in kHz.

If the timing window was extended to 1 second, the results would be quite different. (This is realized by adding another divide-by-1000 chain to the time base.) The counter would then read out in Hz. The display would read "0000." What has occurred is that the MSD counter has changed state 1000 times during the period, ending up at 0. If the input frequency departed from 1.2 MHz by, say, 2 Hz positive, the output would read "0002."

Assume that the time base is 1 ms, as shown, and that the input frequency is increased to 16.15 MHz. In this instance the output would read 6150. The leading 1, signifying the 10-MHz part, would have overrun the counter. This in no way decreases its utility. If it were desirable to read out the 10 MHz and higher frequencies, an additional counter, latch, decoder and LED could be added. Alternatively, the time base could be changed to 100 microseconds.

One major problem occurs with the counter shown in Fig. 33. The display is updated once each 2 milliseconds. The human eye can only respond to changes that occur within about 100 ms. Because of this, the display will appear to flicker in the LSD position. This will occur even if the stability of all signals was uncompromised in stability, so long as they were not coherent. Additional circuitry will allow the display update period to be extended.

Receiver Applications

The counter just described is suitable for general-purpose applications. However, it is not sufficient for receiver use. There are a number of reasons for this. The main one is that the frequency to be counted is not the incoming frequency but that of the local oscillator. The two frequencies differ by the i-f. Sometimes this is of no consequence. For example, if the i-f is exactly at a frequency that is divisible by 1 MHz, the LO can be counted directly. The digits that represent the MHz part are not displayed. This is especially effective for a cw receiver.

Even if the i-f lies at an exact multiple of 1 MHz, difficulties arise where ssb receivers are concerned. This is because the frequency of interest in ssb is not that at the center of the information being transmitted, but that of the suppressed carrier. This corresponds to the sum or the difference of the receiver LO and BFO. In principle these two oscillators could be mixed appropriately, and the resultant information counted. This method can work well if excessive shielding is used, which is possible. If the shielding and isolation are not nearly perfect, the receiver will respond to the mixed product which is precisely at the frequency being received.

A cleaner approach is through the application of additional gates. Assume, for example, that the frequency to be counted corresponded to the *sum* of the BFO and the LO. A suitable display could be achieved by first counting one oscillator and then the other. The counters would not be reset after switching between the first and the second. The result would correspond to the sum of the frequencies.

If the desired output was the *difference* of the BFO and the LO, additional

difficulty would be enountered. This may be circumvented by use of up-down counters as well as with appropriate gating. As pulses arrive at the input to a normal decade counter, the output follows the following sequence: 0, 1, 2, 3, 4, 5, 6, 7, 8, 9, 0 and so forth. That is the unit *counts up*, starting at 0. Some more-elegant ICs have two inputs. One is the count-up input described. The other is a count-down input. Starting at 0, arriving pulses would cause it to read sequentially: 0, 9, 8, 7 and so forth. By using each of the inputs in a properly controlled way, one can obtain a result that corresponds to the difference of two frequencies. Multi-conversion systems may also be accommodated with these methods.

Another method that may be used to read frequency more accurately is by use of presettable counters. In the fundamental system of Fig. 33, the counters were reset to 0 at the end of each counting period, after the information had been strobed into the latches. Presettable counters are more flexible. With the application of the proper programming signals, the "reset" pulse will set them to any desired output. Counting then commences from that point. By choosing the proper preset input, the offsets resulting from the i-f may be accommodated.

The use of presettable counters is generally more direct. However, it is subject to any errors that may occur in the BFO frequency. The up-down counter method automatically accommodates these drift and aging effects.

Counter Noise Considerations

If a frequency counter is to be used with a receiver, there are several precautions that must be taken. If they are not, the noise from the counter may dominate the receiver output. Some of the problems are outlined below.

The interface between the oscillators being counted and the digital circuits should be exceptionally clean. FET buffers are suggested. The oscillators may be attenuated significantly and then reamplified to further enhance the isolation.

Extensive shielding should be used. Ideally, the counter circuitry should be in an rf-tight box. High quality feedthrough capacitors should be used for power supply lines. The 5-volt power supply often used for the digital circuits should be decoupled well from the receiver power supply. Often, some of the shielding recommendations may be relaxed if the rest of the receiver is well shielded. This would be required for other reasons in a high-performance receiver.

Multiplexed displays should be avoided. This requires some explanation. The decoder/drivers used in the circuit of Fig. 33 all operate in parallel. The signals sent to the LED displays are dc ones that change only when the display is updated. In contrast, there are many displays and matching decoder/drivers that operate in a sequential manner. This allows the outputs of several sets of latches to be applied to a single decoder/driver at one time. Similarly, a fewer number of output lines are required to attach to a collection of LED segments. The various digits are scanned at a high rate and pulsed on for short periods. The eye perceives all of the digits as being on, simultaneously. Most digital clocks and pocket calculators use multiplexed displays. The fact that high-speed digital circuits are changing state continually leads to large noise outputs.

The crystal oscillator used as the clock for the time base should not be related directly to the receiver i-f. For example, a receiver built by one of the writers uses a 9-MHz i-f and a digital readout. When the counter was first constructed, a 1-MHz clock was used. The ninth harmonic could be heard faintly in the i-f (at a very low level corresponding to an input signal of −138 dBm). The clock was moved to 2 MHz, thereby solving the problem. The seventh harmonic can be heard at 14 MHz only when an antenna is connected to the receiver.

A final precaution is to time-sequence the time base. This is realized in the counter of Fig. 33 by placing a gate between the crystal oscillator and the divide-by-1000 counter. The oscillator is allowed to run continuously. However, the divider circuit is on only when it is needed. If the display up-date rate is made slow (1/2 second), there is no digital circuitry operating during most of the listening time. In the extreme, provision could be made to completely shut the counter circuits off by means of a front-panel switch.

High-Resolution Frequency Readout

The use of a counter as the frequency display in a receiver has a number of advantages. Many have been outlined. One is the high resolution of the counter, which allows the receiver to be reset precisely to a previously logged frequency. The limit is the interval used for the time-base and the short-term stability of the oscillators.

While reset ability may be high, similar accuracy in readout is not implicit. First, there may be some drift in the clock oscillator used in the time base. Of greater significance is the bandwidth of the receiver. For example, if a 500-Hz-bandwidth receiver is used with a digital readout, the accuracy of a received signal is, at best, 500 Hz. The receiver may be tuned over a 500-Hz range, leading to a 500-Hz change in the readout while still copying an arriving signal.

There is a method that may be employed to extend the accuracy of a digital readout. Auxiliary equipment is required, which is constructed easily or integrated into an existing receiver. Assume that the receiver counter has a time base with a 1-second counting window. The resulting resolution is 1 Hz.

The first extra piece of equipment is a 1-MHz standard. This unit is set carefully against WWV or some other standard of known accuracy. After the transfer standard is calibrated, an harmonic is tuned with the receiver. Once in the passband, the receiver is tuned until the readout displays an exact multiple of the 1-MHz standard. For example, on the 20-meter band, the readout would read 14.000000 MHz. With the receiver so tuned, an external audio oscillator is adjusted to produce exactly the same audio frequency. The comparison may be done with an oscilloscope (in the X-Y mode using Lissajous patterns), with a digital phase-frequency detector, with the counter, or even by ear.

Once the pitch calibration is performed, an unknown signal may be tuned to produce exactly the same pitch. When this is realized, the precise frequency is read directly. On several occasions one of the writers achieved 1-Hz accuracy in W1AW Frequency Measuring Tests with this technique.

It should be mentioned that this method appears to be more accurate than those using a "zero-beat" comparison. Also, the receiver used for these tests had sufficient i-f selectivity that zero beat could not be detected. The ultimate limitation of this approach to frequency measurement is the short-term stability of the oscillators used and the inaccuracies related to Doppler shift during WWV calibration.

A 1-Hz frequency accuracy is rarely needed for an amateur receiver. A question of more practical nature concerned the general usefulness of a digital readout during routine communications. Would an analogue dial be missed? The writers' answer to this query is an uncategorical no! The digital readout was found remarkably easy to adapt to. The ability to set the receiver on a known frequency for monitoring purposes has been immensely useful.

A High-Performance Receiver for 160 Meters

A high order of dynamic range is important to good reception in areas of high signal density. Operation on 160 meters requires a better than average communications receiver, particularly in situations where commercial a-m broadcast stations are nearby, and when the

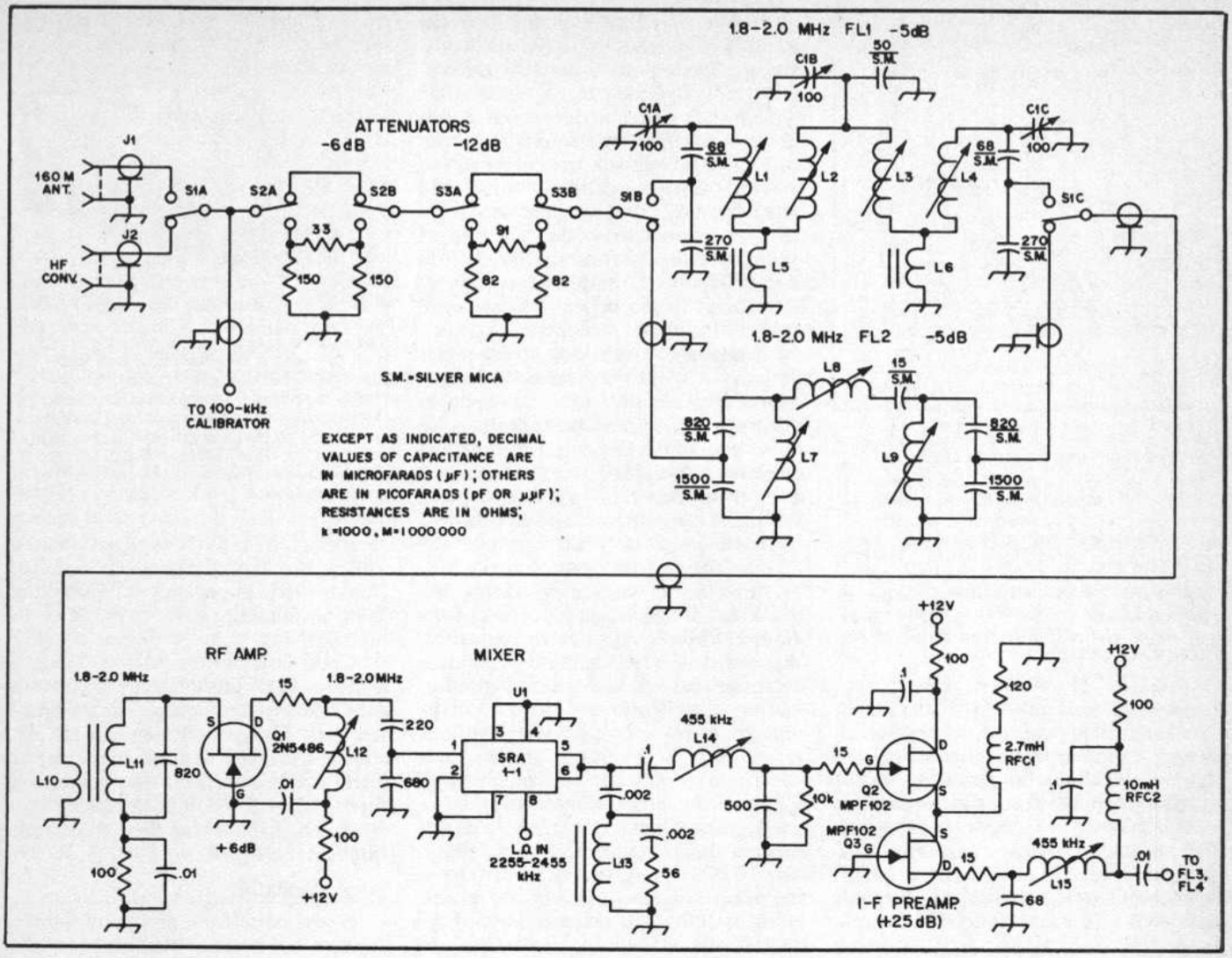


Fig. 34 — Schematic diagram of the receiver front end. Fixed-value capacitors are disk ceramic unless otherwise noted. Resistors are 1/2-W composition. All slug-tuned inductors are contained in individual shield cans which are grounded.

C1 — Three-section variable, 100 pF per section. Model used here obtained as surplus.
J1 — SO-239.
J2 — Phono jack.
L1, L4 — 38 to 68 μH, Q_u of 175 at 1.8 MHz, slug-tuned (J. W. Miller 43A685CBI in Miller S-74 shield can).
L2, L3 — 95 to 187 μH, Q_u of 175 at 1.8 MHz, slug tuned (J. W. Miller 43A154CBI in S-74 shield can).
L5, L6 — 1.45-μH toroid inductor, Q_u of 250 at 1.8 MHz. 15 turns No. 26 enam. wire on Amidon T-50-2 toroid.
L7, L9 — 13-μH slug-tuned inductor (J. W. Miller 9052).
L8 — 380-μH slug-tuned inductor (J. W. Miller 9057).
L10 — 16 turns No. 30 enam. wire over L11 winding.
L11 — 45 turns No. 30 enam. wire on Amidon T-50-2 toroid, 8.5 μH.
L12 — 42-μH slug-tuned inductor, Q_u of 50 at 1.8 MHz (J. W. Miller 9054).
L13 — 8.7-μH toroidal inductor. 12 turns No. 26 enam. wire on Amidon FT-37-61 ferrite core.
L14 — 120- to 280-μH, slug-tuned inductor (J. W. Miller 9056).
L15 — 1.3- to 3.0-mH, slug-tuned inductor (J. W. Miller 9059).
Q1, Q2, Q3 — Motorola JFET.
RFC1 — 2.7-mH miniature choke (J. W. Miller 70F273AI).
RFC2 — 10-mH miniature choke (J. W. Miller 70F102AI).
S1 — Three-pole, two-position phenolic wafer switch.
S2, S3 — Two-pole, double-throw miniature toggle.
U1 — Mini-Circuits Labs. SRA-1-1 doubly balanced diode mixer (2913 Quentin Rd., Brooklyn, NY 11229).

operator lives near other 160-meter enthusiasts who are active on the band. The effects of blocking, cross modulation, and IMD can render a poorly designed receiver useless in the foregoing situation, making weak-signal work an impossible task.

Some ordinary design procedures can be followed when building a receiver with above average dynamic-range parameters, and the construction job is not a difficult one. Special care must go into the front-end design and gain distribution of the receiver circuitry to assure the performance specified here, but construction of such a receiver should be no more exacting than would be the case when building a mediocre one.

Although this is a single-band receiver, coverage of 80 through 15 meters can be accomplished with good dynamic-range traits by employing the converters described later in this chapter. They were designed for high performance also, and the desired chacteristics were based on the dynamic-range profile of this receiver. That is, the two systems are compatible by design intent. IMD of the main-frame receiver (tested at 1.9 MHz) is −95 dB. Noise floor is −135 dBm, and blocking of 1 dB occurs at some point in excess of 123 dB above the noise floor. With the mating 20-meter converter attached the IMD = 88 dB, noise floor is −133 dBm, and blocking is in excess of 123 dB. The 20-meter tests were performed with the fixed-tuned 160-meter front-end filter in the circuit. Tests for dynamic range

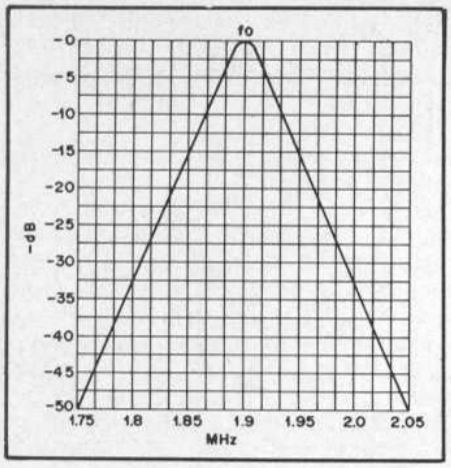


Fig. 35 — Response curve of the tunable front-end filter, centered on 1.9 MHz.

on 160 meters were performed with the tunable Cohn filter in the circuit. This receiver was described first in *QST* for June and July, 1976.

Front-End Circuit

Fig. 34 shows the rf amplifier, mixer, and post-mixer amplifier. What may seem like excessive elaboration in design is a matter of personal whim, but the features are useful, nevertheless. For example, the two front-end attenuators aren't essential to good performance, but are useful in making accurate measurements (6, 12 of 18 dB) of signal levels during on-the-air experiments with other stations (antennas, amplifiers and such). Also, FL2, a fixed-tuned 1.8- to 2-MHz bandpass filter, need not be included if the operator is willing to repeak the three-pole tracking filter (FL1) when tuning about in the band. The fixed-tuned filter is convenient when the down converters are in use.

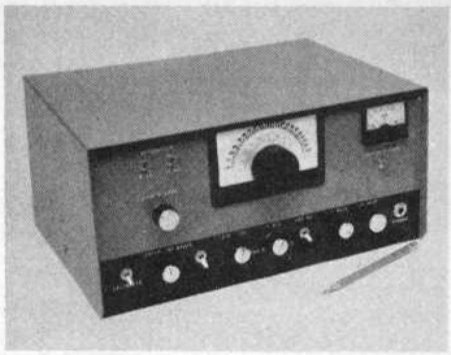
The benefits obtained from a highly selective tunable filter like FL1 are seen when strong signals are in or near the 160-meter band. The rejection characteristics can be seen in Fig. 35. Insertion loss was set at 5 dB in order to narrow the filter response. In this example the high-Q slug-tuned inductors are isolated in aluminum shields, and the three-section variable capacitor which tunes them is enclosed in a shield made from pc-board sections. Bottom coupling is accomplished with small toroidal coils.

Rf amplifier Q1 was added to compensate for the filter loss. It is mismatched intentionally by means of L10 and L11 to restrict the gain to 6 dB maximum. Some additional mismatching is seen at L12, and the mixer is overcoupled to the FET tuned output tank to broaden the response (1.8 to 2 MHz). The design tradeoffs do not impair performance.

The doubly balanced diode-ring mixer (U1) was chosen for its excellent reputation in handling high signal levels, having superb port-to-port signal isolation, and because of its good IMD performance. The module used in this design is a commercial one which contains two broadband transformers and four hot-carrier diodes with matched characteristics. The amateur can build his own mixer assembly in the interest of reduced expense. At the frequencies involved in this example, it should not be difficult to obtain performance equal to that of a commercial mixer.

A diplexer is included at the mixer output (L13 and the related .002 capacitors). The addition was worthwhile, as it provided an improvement in the noise floor and IMD characteristics of the receiver. The diplexer works in combination with matching network L14, a low-pass L-type circuit. The diplexer is a high-pass network which permits the 56-ohm terminating resistor to be seen by the mixer without degrading the 455-kHz i-f. The low-pass portion of the diplexer helps reject all frequencies above 455 kHz so that the post-mixer amplifier receives only the desired information. The high-pass section of the diplexer starts rolling off at 1.2 MHz. A reactance of 66 ohms (X_c and X_L) was chosen to permit use of standard-value capacitors in the low-Q network.

A pair of source-coupled JFETs is used in the post-mixer i-f preamplifier. The 10,000-ohm gate resistor of Q2 sets the transformation ratio of the L network at 200:1 (50 ohms to 10 kΩ). An L network is used to couple the pre-amplifier to a diode-switched pair of Collins mechanical filters which have a characteristic impedance of 2000 ohms. The terminations are built into the filters.



The receiver is built in a homemade aluminum cabinet. A two-tone gray and flat-black paint job has been applied. Black Dymo tape labels are used for identifying the controls in the black area, and gray labels are affixed to the gray portion of the front panel. A cut-down Jackson Brothers vernier dial mechanism (two-speed) is used for frequency readout.

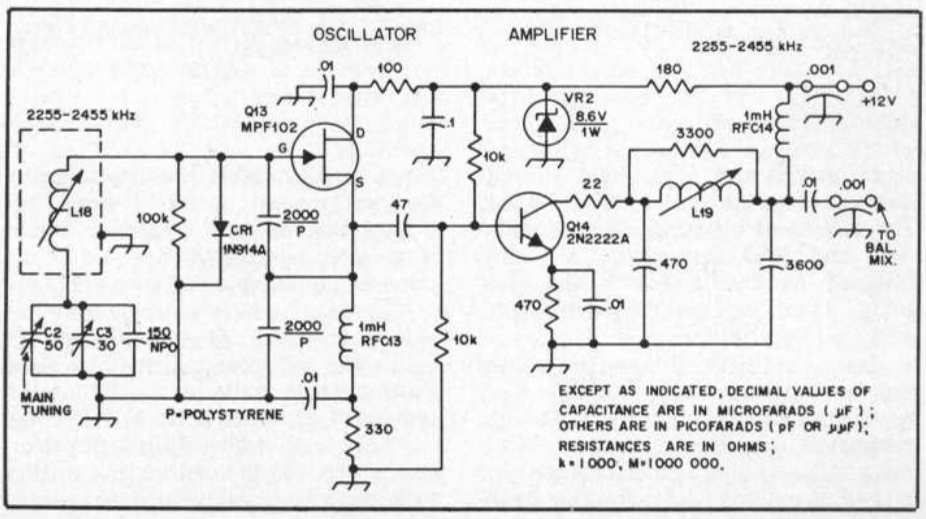
Gain distribution to the mixer is held to near unity in the interest of good IMD performance. The preamplifier gain is approximately 25 dB. The choice was made to compensate for the high insertion loss of the mechanical filters — 10 dB. Without the high gain of Q2 and Q3 there would be a deterioration in noise figure.

Local Oscillator

A low noise floor and good stability are essential traits of the local oscillator

Fig. 36 — Circuit diagram of the local oscillator. Capacitors are disk ceramic unless specified differently. Resistors are 1/2-W composition. Entire assembly is enclosed in a shield box made from pc-board sections.

C2 — Double-bearing variable capacitor, 50 pF.
C3 — Miniature 30-pF air variable.
CR1 — High-speed switching diode, silicon type 1N914A.
L18 — 17- to 41-μH slug-tuned inductor, Q_u of 175 (J. W. Miller 43A335CBI in Miller S-74 shield can).
L19 — 10- to 18.7-μH slug-tuned pc-board inductor (J. W. Miller 23A155RPC).
RFC13, RFC14 — Miniature 1-mH rf choke (J. W. Miller 70F103AI).
VR2 — 8.6-V, 1-W Zener diode.



in a quality receiver. The requirements are met by the circuit of Fig. 36. Within the capabilities of the ARRL lab measuring procedures, it was determined that VFO noise was at least 90 dB below fundamental output. Furthermore, stability at 25°C ambient temperature was such that no drift could be measured from a cold start to a period three hours later. Mechanical stability is excellent: Several sharp blows to the VFO shield box caused no discernible shift in a cw beat note while the 400-Hz i-f filter was actuated. VFO amplifier Q14 is designed to provide the recommended +7 dBm mixer injection voltage. Furthermore, the output pi tank of Q14 is of 50 ohms characteristic impedance. Though not of special significance in this application, the measured harmonic output across 50 ohms is −36 dB at the second order, and −47 dB at the third order.

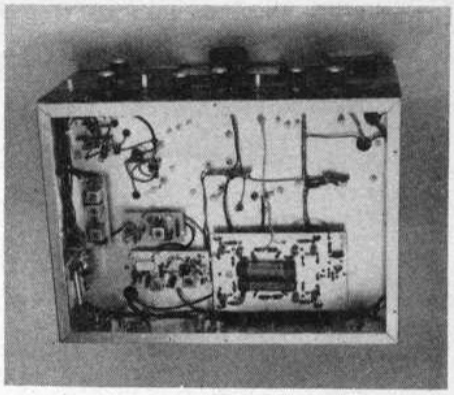
Filter Module

In the interest of minimizing leakage between the filter input to output ports (Fig. 37), diode switching is used. The advantage of this method is that only dc switching is required, thereby avoiding the occasion for unwanted rf coupling across the contacts and wafers of a mechanical switch. 1N914 diodes are used to select FL3 (400-Hz bandwidth) or FL4 (2.5-kHz bandwidth). Reverse bias is applied to the nonconducting diodes. This lessens the possibility of leakage through them. Because the Collins filters have a characteristic impedance of 2000 ohms, the output coupling capacitors from each are 120 pF rather than the low-reactance .01-μF units, as used at the filter inputs. Without the smaller value of capacitance the filters would see the low base impedance of Q4, the post-filter i-f amplifier. The result would be one of double termination in this case, leading to a loss in signal level. Additionally, the 120-pF capacitors help to divorce the input capacitance of the amplifier stage. The added capacitance would have to be subtracted from the 350- and 510-pF resonating capacitors at the output ends of the filters.

The apparent overall receiver gain is greatest during cw reception, owing to the selectivity of cw filter FL3. To keep the S-meter readings constant for a given signal level in the ssb and cw modes, R7 has been included in the filter/amplifier module. In the cw mode, R7 is adjusted to bias Q4 for an S-meter reading equal to that obtained in the ssb mode. Voltage for the biasing is obtained from the diode-switching line during cw reception.

I-F Amplifier

A receiver i-f system should be capable of providing a specific gain, have an acceptable noise figure, and respond satisfactorily to the applied agc. This almost bromidic judgment is not as trite as it may seem, for some designers use a haphazard approach to this part of a receiving system. Two of the more serious shortcomings in some designs are



Considerable space remains beneath the chassis for the addition of accessory circuits or a set of down converters. At the upper left are the adjustment screws for the tunable filter, plus the bottom-coupling toroids. At the left center is the fixed-tuned front-end filter. To the right is the rf-amplifier module. A 100-kHz MFJ Enterprises calibrator is seen at the far lower left. Immediately to its right is the mixer/amplifier assembly. The large board at the lower center contains the i-f filters and post-filter amplifier. Most of the amplifier components have been tacked beneath the pc board because of design changes which occurred during development.

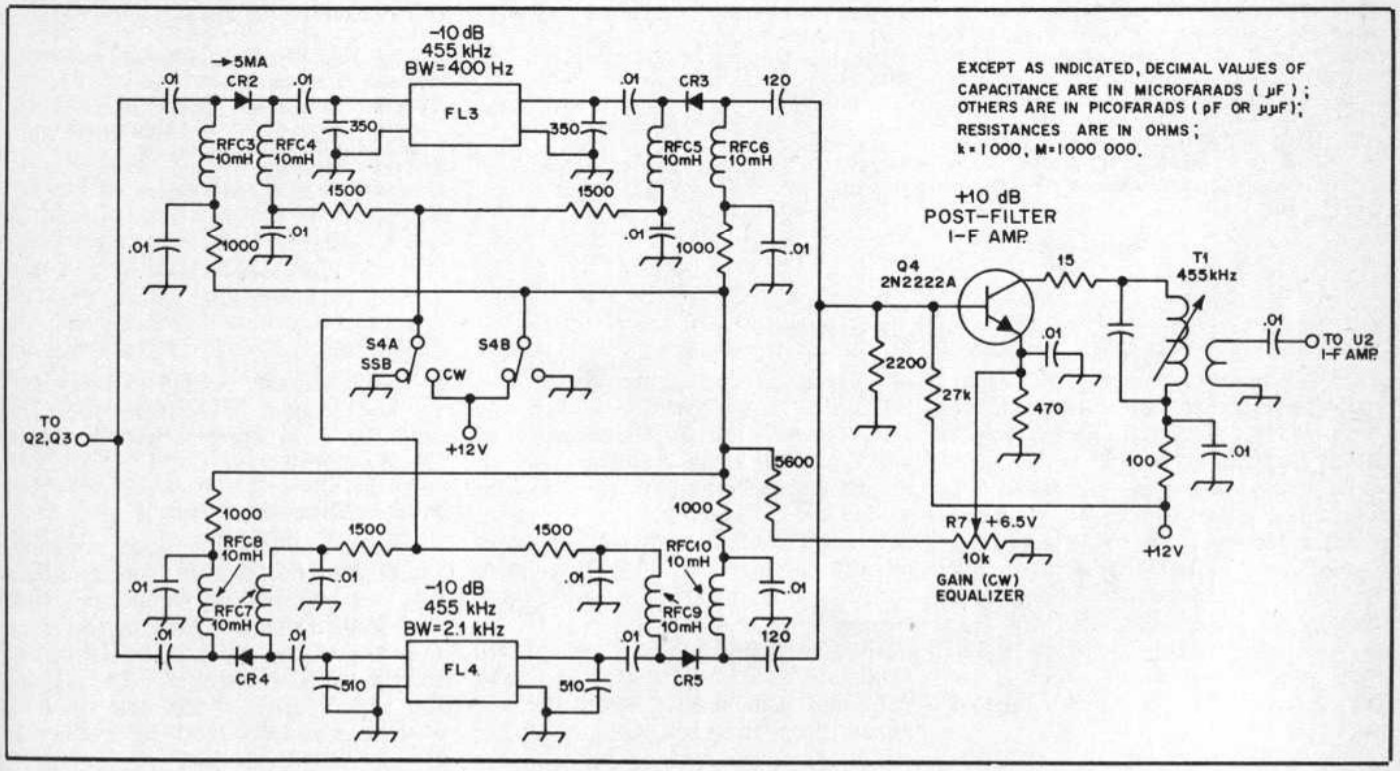


Fig. 37 — Schematic diagram of the filter and i-f post-filter amplifier. Capacitors are disk ceramic. Resistors are 1/2-W composition.

CR2-CR5, incl. — High-speed silicon switching diode, 1N914A.
FL3 — Collins mechanical filter F455FD-04.
FL4 — Collins mechanical filter F455FD-25.
RFC3-RFC10, incl. — 10-mH miniature rf choke (J. W. Miller 70F102A1).
R7 — Pc-board control, 10,000 ohms, linear taper.
S4 — Double-pole, double-throw toggle or wafer.
T1 — Miniature 455-kHz i-f transformer (J. W. Miller 2067, 30,000 to 500 ohms).

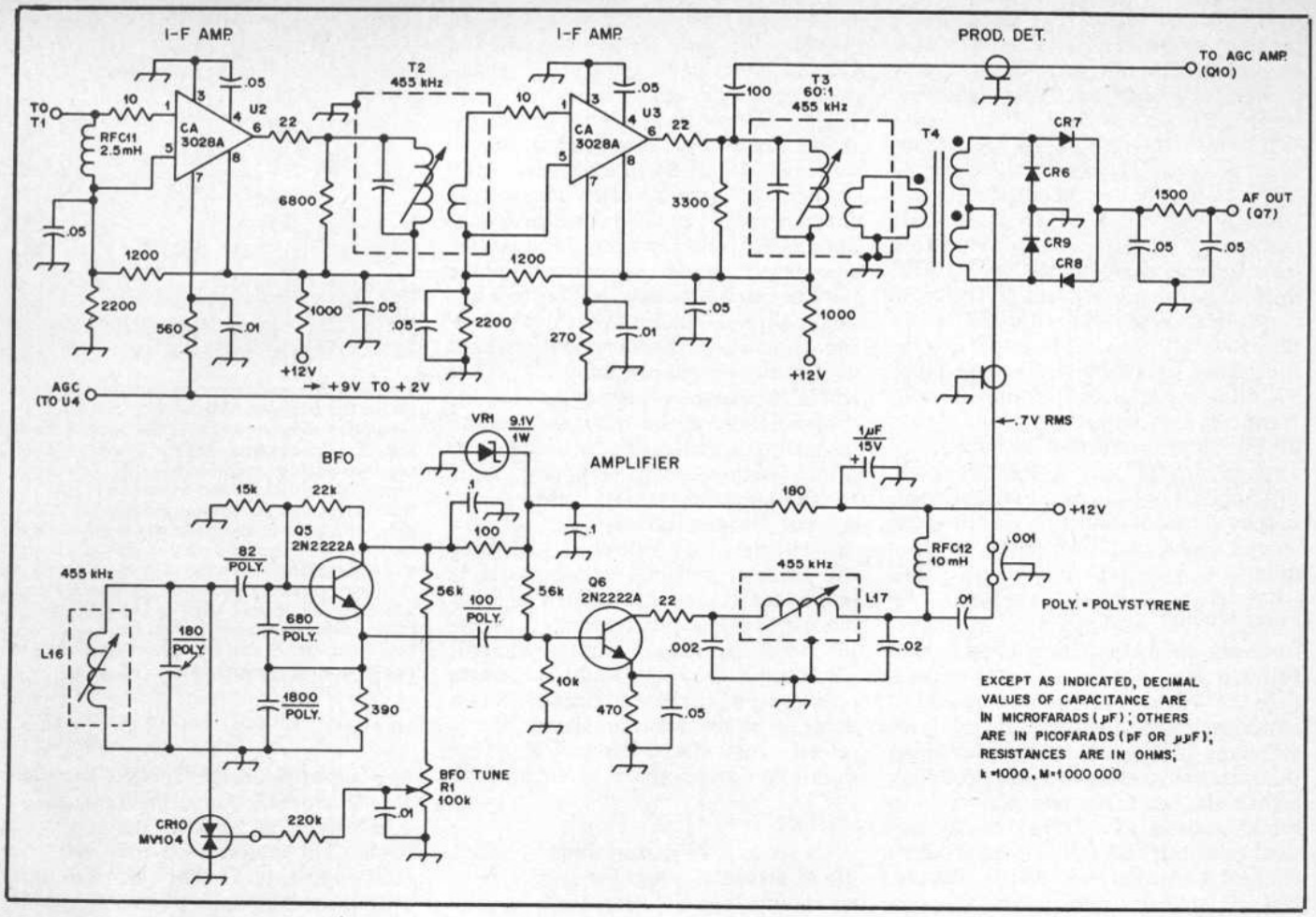


Fig. 38 — Circuit of the i-f amplifier, BFO, and product detector. Capacitors are disk ceramic unless noted differently. Fixed-value resistors are 1/2-W composition. Dashed lines show shield enclosures. The BFO and i-f circuits are installed in separate shield boxes. The R-C active filter and af preamplifier are on a common circuit board, which is not shielded.

CR6-CR9, incl. — High-speed silicon, 1N914A or equiv.
CR10 — Motorola MV-104 Varicap tuning diode.
L16 — Nominal 640-μH slug-tuned inductor (J. W. Miller 9057).
L17 — Nominal 60-μH slug-tuned inductor (J. W. Miller 9054).
R1 — 100,000-ohm linear-taper composition control (panel mount).
RFC11 — 2.5-mH miniature choke (J. W. Miller 70F253A1).
RFC12 — 10-mH miniature choke (J. W. Miller 70F102A1).
T2, T3 — 455-kHz i-f transformer. See text. (J. W. Miller 2067).
T4 — Trifilar broadband transformer. 15 trifilar turns of No. 26 enam. wire on Amidon T-50-61 toroid core.
U2, U3 — RCA IC.
VR1 — 9.1-V, 1-W Zener diode.

poor agc (clicky, pumping, or inadequate range) and insufficient i-f gain.

A pair of RCA CA3028A ICs is used in the i-f strip. Somewhat greater gain and agc range is possible with MC1590G ICs, and they are the choice of many builders. However, the CA3028As, configured as differential amplifiers, will provide approximately 70 dB of gain per pair when operated at 455 kHz. This gives an agc characteristic from maximum gain to full cutoff which is entirely acceptable for most amateur work.

Fig. 38 shows the i-f amplifiers, product detector, and Varicap-tuned BFO. Transformer coupling is used between U2 and U3, and also between U3 and the product detector. The 6800-ohm resistors used across the primaries of T2 and T3 were chosen to force an impedance transformation which the transformers can't by themselves provide: Available Miller transformers with a 30,000-ohm primary to 500-ohm secondary characteristic are used. U2 and U3 have 10- and 22-ohm series resistors in the signal lines. These were added to discourage vhf parasitic oscillations.

Agc is applied to pin 7 of each IC. Maximum gain occurs at +9 V, and minimum gain results when the agc voltage drops to its low value, +2 V. The agc is rf-derived, with i-f sampling for the agc amplifier being done at pin 6 of U3 through a 100-pF blocking capacitor.

The 1000-ohm decoupling resistors in the 12-V feed to U2 and U3 drop the operating voltage to +9. This aids stability and reduces i-f system noise. The amplifier strip operates with unconditional stability.

Product Detector

A quad of 1N914A diodes is used in the product detector. Hot-carrier diodes may be preferred by some, and they may lead to slightly better performance than the silicon units. A trifilar broadband toroidal transformer, T4, couples the i-f amplifier to the detector at a 50-ohm impedance level. BFO injection is supplied at 0.7 V rms.

BFO Circuit

In the interest of lowering the cost of this project, a Varicap (CR10 of Fig. 38) is used to control the BFO fre-

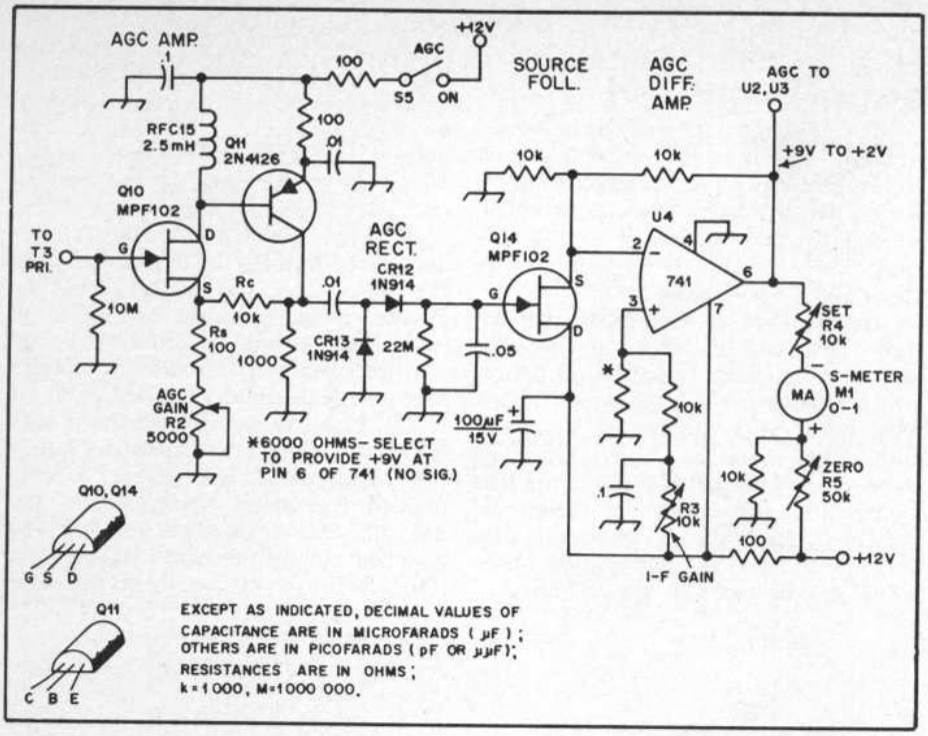


Fig. 39 — Schematic diagram of the agc system. Capacitors are disk ceramic except when polarity is indicated, which signifies electrolytic. Fixed-value resistors are 1/2-W composition. This module is not enclosed in a shield compartment.

CR12, CR13 — High-speed silicon. 1N914A or equiv.
Q10, Q11, Q14 — Motorola transistor.
R2, R4, R5 — Linear-taper composition pc-board mount control.
R3 — 10,000-ohm linear-taper control, panel mounted.
RFC15 — 2.5-mH miniature choke (J. W. Miller 70F253A1).
S5 — Single-pole, single-throw toggle.
U4 — Dual-in-line 8 pin 741 op amp.
M1 — 0- to 1-mA meter.

the Q10/Q11 gain is determined as: Gain (dB) = 20 log $R_c \div R_s$ Control R2 has been included as part of R_s to permit adjustment of the agc loop gain. Each operator may have a preference in this regard. The agc is set so it is fully actuated at a signal-input level of 10 μV. Agc action commences at 0.2 μV (1 dB of gain compression).

Agc disabling is effected by removing the operating voltage from Q10 and Q11 by means of S5. Manual i-f gain control is made possible by adjusting R3 of Fig. 39. Agc delay is approximately 1 second. Longer or shorter delay periods can be established by altering the values of the Q14 gate resistor and capacitor. Agc amplifier gain is variable from 6 to 40 dB by adjusting R2. Agc action is smooth, and there is no evidence of clicks on the attack during strong-signal periods. At no time has agc "pumping" been observed.

Audio System

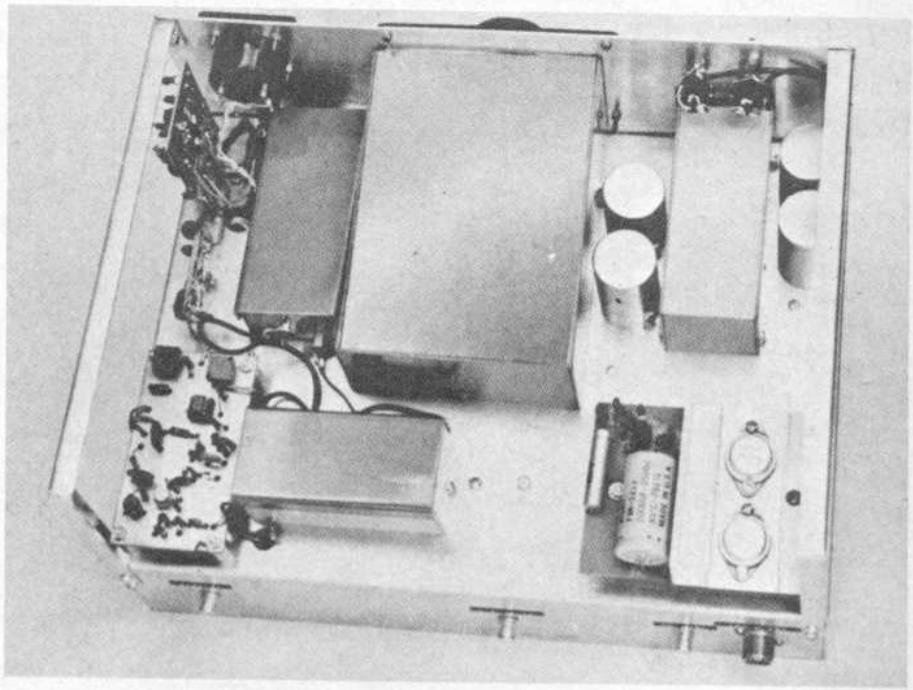
A major failing of many receivers is poor-quality audio. For the most part this malady is manifest as cross-over distortion in the af-output amplifier. Moreover, some receivers have marginal audio-power capability for normal room volume when a loudspeaker is used. Some transformerless single-chip audio ICs (0.25- to 2-W class) exhibit a prohibitive distortion characteristic, and this is

quency. Had a conventional system been utilized, three expensive crystals would have been needed to handle upper sideband, lower sideband, and cw. The voltage-variable capacitor tuning method shown in Fig. 38 is satisfactory if the operator is willing to change the operating frequency of the BFO when changing receive modes. Adjustment is done by means of front-panel control R1. Maximum drift with this circuit was measured as 5 Hz from a cold start to a time three hours later. A Motorola MV-104 tuning diode is used at CR10.

Q6 functions as a Class A BFO amplifier/buffer. It contains a pi-network output circuit and has a 50-ohm output characteristic. The main purpose of the amplifier stage is to increase the BFO injection power without loading down the oscillator.

AGC Circuit

Fig. 39 shows the agc amplifier, rectifier, dc source follower, and op-amp difference amplifier. An FET is used at Q10 because it exhibits a high input impedance and will not, therefore, load down the primary of T3 in Fig. 38. Q1 is direct coupled to a pnp transistor, Q11. Assuming that R_s and R2 are treated as a single resistance, R_s,



Top-chassis view of the receiver. The R-C active filter and audio preamplifier are built on the pc board at the upper left. To the right is the BFO module in a shield box. The agc circuit is seen at the lower left, and to its right is the i-f strip in a shield enclosure. The large shield box at the upper center contains the VFO. To its right is the tunable front-end filter. The three-section variable capacitor is inside the rectangular shield box. The audio amplifier module is seen at the lower right. The small board (mounted vertically) at the left center contains the product detector. Homemade end brackets add mechanical stability between the panel and chassis and serve as a support for the receiver top cover.

especially prominent at low signal levels. The unpleasant effect is one of "fuzziness" when listening to low-level signals. Unfortunately, external access to the biasing circuit of such ICs is not typical, owing to the unitized construction of the chips.

Since undistorted audio is an important feature of a quality communications receiver, discrete devices have been employed in this circuit. The complementary-symmetry output transistors and the op-amp driver are configured in a manner similar to that used by Jung in his *Op Amp Cookbook*. Maximum output capability is 3.5 W into an 8-ohm load. An LM-301A driver was chosen because of its low-noise profile. There has been no aural evidence of distortion at any signal level while using the circuit of Fig. 40. The rationale in this situation is one of having considerably more audio power available than is needed – a practice used in hi-fi work.

R-C Active CW Filters

A worthwhile improvement in signal-to-noise ratio can be realized during weak-signal reception by employing an R-C active bandpass filter. A two-pole version (FL5) is shown in Fig. 40. A peak frequency of 800 Hz results from the *R* and *C* values given.

The benefits of FL5 are similar to those described elsewhere in this volume, where a second i-f filter (at the i-f strip output) is used to reduce wideband noise from the system. The R-C active filter serves in a similar manner, but performs the signal "laundering" at audio rather than at rf. The technique has one limitation – monotony in listening to a fixed-frequency beat note, which is dictated by the center frequency of the audio filter. The R-C filter should be designed to have a peak frequency which matches the cw beat-note frequency preferred by the operator. That is, if the BFO is adjusted to provide an 800-Hz cw note, the center frequency of FL5 should also be 800 Hz.

Experience with FL5 in this receiver has proved in many instances that weak DX signals on 160 meters could be elevated above the noise to a Q5 copy level, while without the filter solid copy was impossible. It should be stressed that high-*Q* capacitors be used from C4 to C7, inclusive, to assure a sharp peak response. Polystyrene capacitors satisfy the requirement. To ensure a well-defined (minimum ripple) center frequency, the capacitors should be matched closely in value (5 percent or less). Resistors of 5-percent tolerance

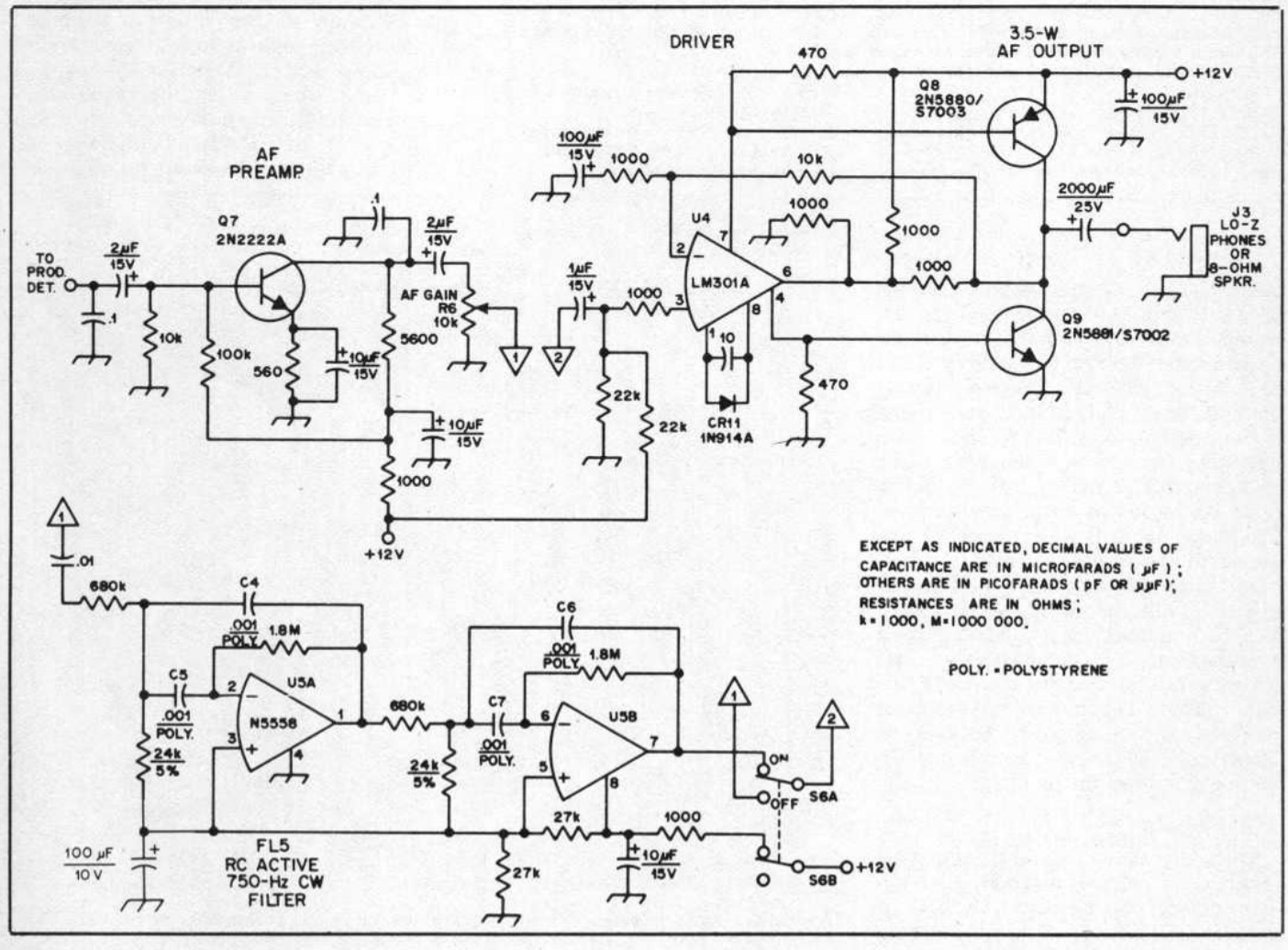
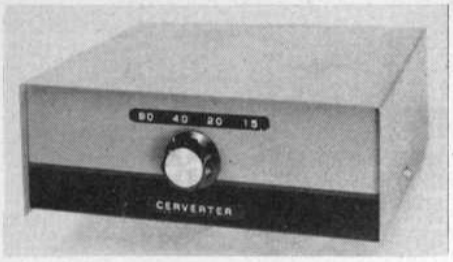


Fig. 40 – Diagram of the audio amplifier and R-C active filter. Capacitors are disk ceramic unless otherwise noted. Polarized capacitors are electrolytic or tantalum. Fixed-value resistors are 1/2-W composition. This circuit is not contained in a shield box. Heat sinks are used with Q8 and Q9.

C4-C7, incl. – See text.
CR11 – High-speed silicon, 1N914A or equiv.
J3 – Phone jack.
R6 – 10,000-ohm audio-taper composition control, panel mounted.
S6 – Double-throw, double-pole toggle.
U4 – National Semiconductor LM-301A IC.
U5 – Signetics N5558 dual op-amp IC.



Exterior view of the high-performance converter assembly. A gray and black spray-paint finish is applied to the homemade aluminum cabinet. Lettering is by means of a Dymo tape labeler.

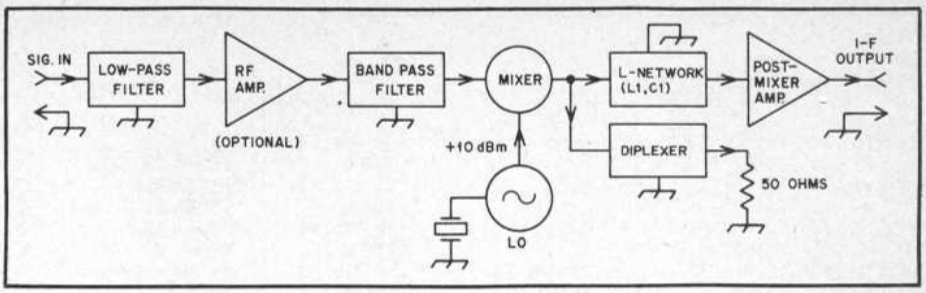


Fig. 41 — Block diagram of the CER-verters.

should be employed in the circuit, where indicated in Fig. 40.

Summary Comments

The photographs illustrate a modular construction technique. All rf-circuit assemblies are isolated from one another, and from outside energy influences, by means of shield compartments. Signal points are joined (module to module) with RG-174/U subminiature coaxial cable, the shield braids being grounded to the chassis at each end. Feedthrough-type .001-μF capacitors are used at the 12-V entry points of the modules. The foregoing measures help to prevent birdies and unwanted stray rf pickup.

The tuning range of the receiver is 200 kHz. This means that for use with converters the builder will have to satisfy himself with the cw or ssb band segments. The alternatives are to increase the local oscillator tuning range to 500 kHz, or use a multiplicity of converters to cover the cw and ssb portions of each band.

High-Performance Converters

This section provides circuits for a group of converters (80 through 15 meters) for use with the high-performance 160-meter receiver described in this chapter. These units were described originally in *QST* for June, 1976.

Converter Designs

After a bit of number crunching it was concluded that the converters should have a net gain of about 10 dB and an output intercept of approximately +17 dBm or higher. For work on the bands up through 14 MHz, a noise figure of 13 to 16 dB was deemed acceptable. On the higher bands some compromise in dynamic range would be tolerable in order to achieve lower noise figures. In studying the available circuit combinations it was decided to base the front end of the converters on a diode-ring mixer. The mixer would be

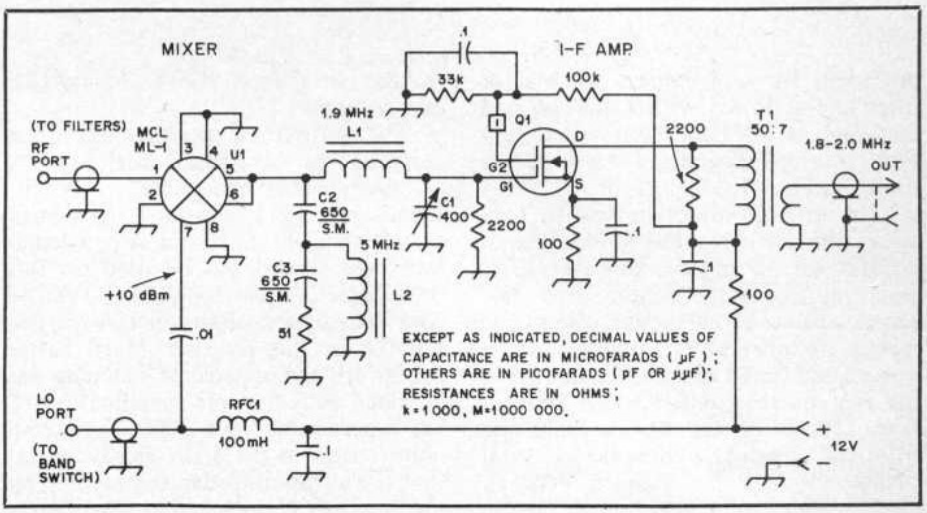


Fig. 42 — Diagram of the mixer and amplifier. Fixed-value capacitors are disk ceramic unless noted otherwise. Resistors are 1/2-W composition. See tables for component values not marked. U1 is a ML-1 or SRA-1 doubly balanced diode-ring mixer assembly. L2 (1.62 μH) has 18 turns of No. 22 wire on a T50-2 toroid core. T1 primary has 50 turns of No. 22 wire on an FT-50-72 toroid core. The secondary contains 7 turns of No. 22 wire. L1 has 65 turns No. 26 enam. wire on a T68-2 toroid core.

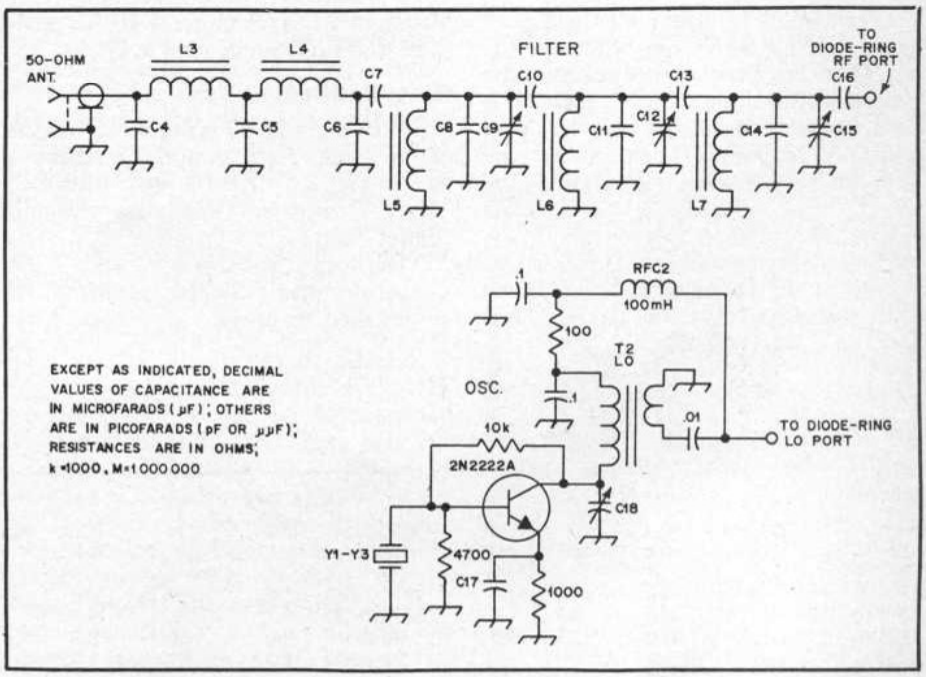


Fig. 43 — Diagram of the filter and crystal oscillator used on 20, 40 and 80 meters. Numbered fixed-value capacitors are silver micas. Resistors are 1/2-W composition. See Tables 1 and 2 for parts values.

Table 1

BAND (MHz)	L3, L4, L8 (TURNS-CORE)	L9 (TURNS-CORE)	L5, L6, L7 L10, L11, L12 (TURNS-CORE)	T2, T3 (TURNS-CORE)
3.5 to 3.7	19, No. 22 T50-2	none	35, No. 24 T68-2	25, No. 24 T50-2, 2-t. link
7.0 to 7.2	15, No. 22 T50-2	none	20, No. 22 T68-6	25, No. 24 T50-2, 2-t. link
14.0 to 14.2	12, No. 22 T50-6	none	12, No. 22 T68-6	28, No. 24 T50-6, 3-t. link
21.0 to 21.2	10, No. 22 T50-6	21, No. 22 T50-6	10, No. 22 T50-6	19, No. 24 T50-6, 2-t. link

Coil and transformer data. Toroid cores are Amidon Assoc. powdered-iron type. Y1, Y2, Y3 and Y4 for 3.5 through 21 MHz, respectively, are 5.5, 5.2, 12.2 and 19.2 MHz (International Crystal Co. type GP, 30-pF load capacitance).

preceded by a bandpass preselector filter and followed with a diplexer and dual-gate MOSFET amplifier at 1.9 MHz. A block diagram of the system is shown in Fig. 41.

The original intention was to construct separate converters for each band, 80 through 10 meters. However, after reviewing the design requirements, this was found to be redundant. Diode-ring mixers are inherently broadband and do not require tuned circuits. Furthermore, the post-mixer amplifier would be identical for all of the bands. Only the front-end preselector networks and local oscillators need be changed between bands. The final configuration chosen was to use a master board which contained the diode-ring mixer and the post amp. A family of boards was then constructed, each containing a suitable local oscillator and the preselector network for the band of interest.

Mixer and Post-Amplifier Board

The circuit for the mixer and dual-gate MOSFET amplifier is shown in Fig. 42. There are a few departures from the typical in this design. First, a diplexer is used between the mixer and the "post amp." A 2200-ohm resistor at the gate provides a termination, causing the mixer to see 50 ohms in the 1.9-MHz frequency range.

In order to simplify the band switching, +12 volts dc is supplied through the local oscillator port of the mixer. This is realized with an rf choke and suitable capacitors.

The output of the amplifier was designed for broadband performance. To obtain a large bandwidth, the output transformer (T1) is wound on a high-permeability ferrite toroid. A powdered-iron core should not be used for this transformer. It was found that a ferrite core with a permeability of 125 was not suitable in this position. Much better bandwidth and impedance matching was obtained with the core specified, which has a permeability of 2000. The 2200-ohm resistor in the drain circuit ensures that the output impedance presented by the amplifier is close to 50 ohms. This is important in order to assure that the input filters of the 160-meter receiver are properly terminated.

A ferrite bead is used on gate 2 of the amplifier. This may not be necessary in some cases. However, it was included to lessen the possibility of uhf oscillations occurring within the amplifier. A Fairchild FT-0601 or RCA 40673 dual-gate MOSFET can be used at Q1.

Front-End Sections

Shown in Fig. 43 is the circuit used as the front end for each of the lower-input bands (3.5-3.7, 7.0-7.2 and 14.0-14.2 MHz). Component values are given in Tables 1 and 2.

The local oscillator for each of the converters uses a bipolar transistor and is designed to provide an output from +10 to +13 dBm. This level of LO injection was found to be near optimum for the diode-ring mixer.

The preselector filters are fairly elaborate. However, the results are well worth the extra expense and effort. Predistorted filter-synthesis methods were used when designing the bandpass filters. They were designed for a three-pole Butterworth response.

One problem with multisection filters using capacitors as coupling elements between the resonators is that the stop-band attenuation may degrade in the vhf spectrum. This is due to slight amounts of lead inductance in the tuning capacitors, and the fact that the capacitive-intersection coupling method degenerates toward a high-pass filter response away from the passband. In order to suppress these responses, should they occur, a 5-pole low-pass filter is included at the antenna terminal.

Two methods were used for evaluation of the filter designs. First, after initial calculation of the component values, a computer program was used to determine the frequency response of the filters over a wide range. In this analysis, resistors were placed in the circuit to stimulate the distortion effects caused by the losses in the cores.

After the filters were built and aligned in the home shop, they were checked with laboratory instrumentation. In that case a Tektronix 7L13 spectrum analyzer and TR-502 tracking generator were used. The measured results around the passband corresponded with the computer simulation. The stop-band attenuation was measured, with one exception, to be over 100 dB for all three filters evaluated. The exception was for the 80-meter filter. At about 70 MHz the attenuation degraded to about 95 dB, but returned to the better values at frequencies up through 200 MHz.

A Butterworth response was chosen because that filter shape is aligned easily with simple test equipment. Alignment is performed by driving the filter with a 50-ohm signal generator and terminating the output in a sensitive 50-ohm detector. The generator is set at the center

Table 2

BAND (MHz)	C4, C6 C19 (pF)	C5, C20 (pF)	C7 (pF)	C8 (pF)	C9, C12 C15 (pF)	C10 (pF)	C11 (pF)	C13 (pF)	C14 (pF)	C16 (pF)	C17, C31 (pF)	C18, C32 (pF)	C21 (pF)
3.5 to 3.7	790	1580	130	–	90 to 400	12	–	10	–	91	100	400	–
7.0 to 7.2	450	890	43	–	90 to 400	4.7	–	4.7	–	62	100	400	–
14.0 to 14.2	220	450	33	90	20 to 90	3.3	90	3.3	90	22	47	20 to 90	–
21.0 to 21.2	150	300, 345	–	51	20 to 90	1.2	51	1.2	51	12	47	20 to 90	20 to 90

Fixed-value and trimmer capacitors. Fixed-value capacitors are silver-mica or similar high-*Q*, stable types. Trimmers are mica compression type. See text for obtaining precise non-standard fixed-capacitance values.

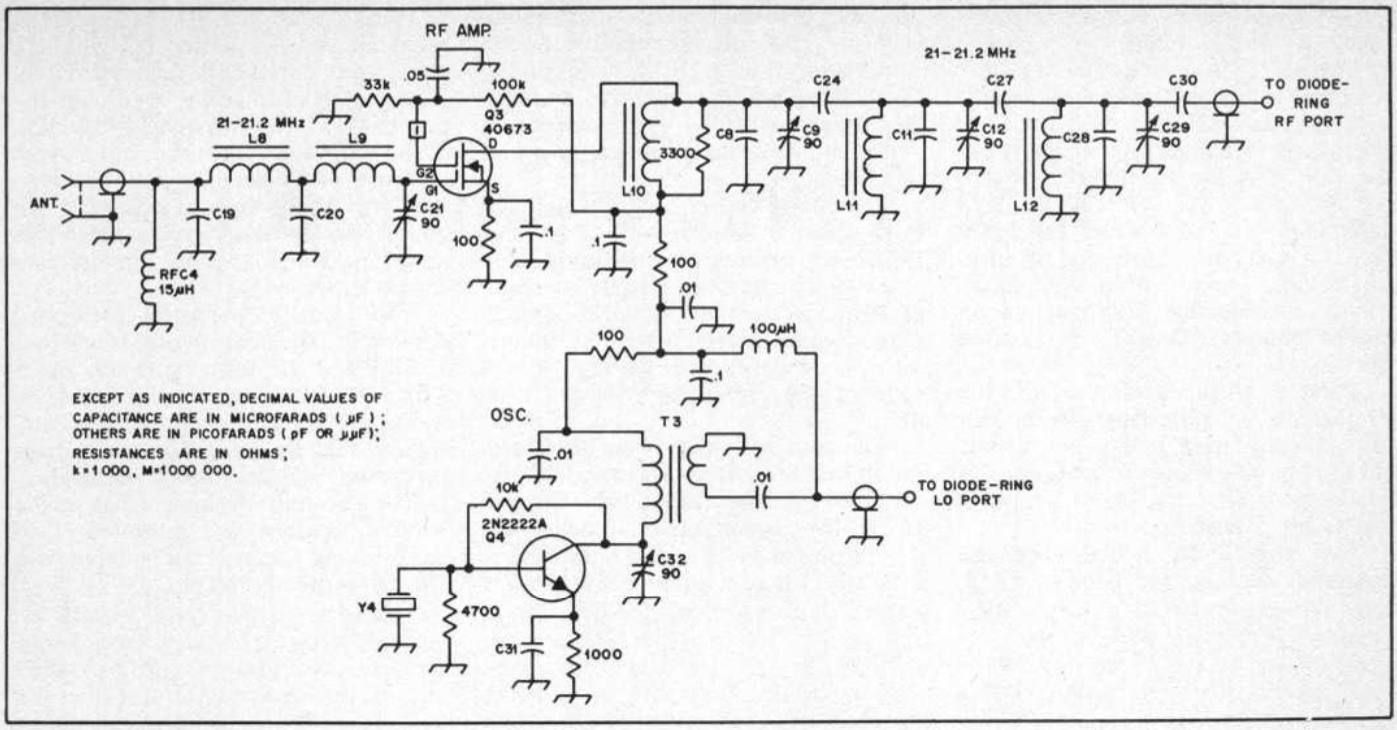


Fig. 44 — Diagram of the 15-meter front-end circuit. Numbered fixed-value capacitors are silver micas. Resistors are 1/2-W composition. See Tables 1 and 2 for other parts values.

frequency of the filter and the variable capacitors are adjusted for a maximum response. Experimentally, it was not found necessary to readjust the filters when the swept instrumentation was available.

The converter for the 15-meter band was built using the circuit in Fig. 44. On this band it was felt that a better noise figure might be useful. This was provided by inserting an rf amplifier between the low-pass filter and the bandpass circuit. The low-pass circuit was modified. The input section is a symmetrical pi network with a Q of 1. This is followed by a pi network with a Q of 10 and an impedance transformation from 50 to 2000 ohms. A 3300-ohm resistor is used in the drain circuit to ensure proper termination of the bandpass filter. In the unit built, the drain was attached directly to the "hot" end of the resonator (L10). However, it would be desirable to reduce the gain somewhat. This would be realized easily by tapping the drain down on the tuned circuit as shown. The terminating resistor should remain across L10.

Those building the converter for 80 meters may wish to also cover the 75-meter phone band. While the filter shown could probably be realigned for a range about 100 kHz higher, the shape of the filter would no doubt deteriorate if it were moved farther. A better approach would be to change the value of the inductors. Proper results should be obtained by reducing the coils from 35 to 32 turns, keeping all capacitor values the same. A 5.8-MHz crystal would be required for tuning the range from 4.0 to 3.8 MHz.

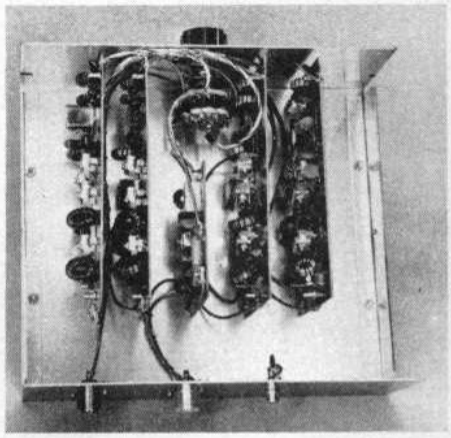
Additional Design Notes

The reader should note that the tuning will be "backward" for the 80-meter band. This was done because a strong 1.7-MHz local-oscillator signal would have appeared at the input to the post-mixer amplifier. This could have resulted in IMD products. Furthermore, for the 75-meter band the crystal would have been at 2.0 MHz if low-side injection were used. This would have placed a strong signal within the tuning range of the main receiver. If it is desirable that all hf bands tune in the same direction, the builder should pick high-side crystals for all of the bands.

The approach used for the 15-meter converter in order to obtain low-noise performance could also be applied to the 10- and 6-meter bands. Filter designs for these bands can be extracted from the appendix. The image rejection might be a little poor with such a low i-f frequency in the 6-meter case.

Another revision would be the construction of a high-performance *80-meter* receiver with converters for the higher bands. The converters described would be suitable for this situation. The crystal frequencies would change accordingly. The diplexer between the diode mixer and the "post amp" should be redesigned. This could be done easily by halving the inductance and capacitance values used in the diplexer circuit. The broadband output circuit in the drain of Q1 should work equally well at 3.5 MHz. The 15- and 20-meter bandpass filters were designed with enough bandwidth to cover the total band. This was done in order to keep the insertion losses at a reasonable level. A slightly wider filter would be required for the total 40-meter band.

The converters are built on large circuit boards. This was done in order to



Interior view of the converter unit. The boards are mounted edgewise. The mixer module is seen at dead center. A multisection wafer switch, with shield partitions between wafers, should be used in place of the one seen in this photograph (see text).

ensure a reasonable level of stopband rejection in the filters and to ease construction. Those interested in a more compact format should consider the inclusion of shields between the sections of the input bandpass filter and between the filter circuitry and the corresponding oscillators. It is useful to build miniature equipment when there is a need for small size. However, for high-performance home-station equipment, where considerable experimentation may be required, a larger format is often desirable.

Because the pc boards shown in the photograph are quite large, the builder will probably elect to lay the circuits out for a more compact format. For this reason there are no pc-board templates and layouts available.

Care should be taken when the front-end sections are band-switched. Shielding between switch wafers should have over 100 dB of isolation. Diode switching is not recommended unless the builder has equipment to evaluate the effects on IMD. The single-wafer switch shown in the photographs is not recommended.

The only converter evaluated for IMD was the 14-MHz unit. Two-tone IMD measurements were performed and it was found that the output intercept of the converter was +22 dBm. This is more than sufficient for the application, since it greatly exceeds the input intercept of the 160-meter receiver, +7.5 dBm.

The gain and MDS were measured for all four converters. The signal generator used was an HP-8640B. On the three lower bands, the noise figure was 12 dB plus the loss of the input filters. Similarly, the gain of the converter was 12.5 dB, minus the loss of the input filters. It was found that the gain and noise figure could both be improved by removing the 2200-ohm resistor at the gate of Q1. There was a slight reduction in the output intercept, but not enough to cause problems. However, the low-pass part of the diplexer became much sharper in frequency response. This would make a front-panel trimmer control necessary.

The 15-meter converter performed differently. The net gain of this unit was 32.5 dB and the noise figure was about 3 dB. This is too much sensitivity to be usable at this frequency. It is recommended that the builder move the drain tap on the bandpass filter as outlined.

The two-tone dynamic range of the complete receiver was measured at 88 dB. Blocking occurred for an input over 120 dB above the MDS.

Chapter 7

Test Equipment and Accessories

Measurements are the key to obtaining good results in amateur experimentation. This form of test procedure will help assure proper equipment performance while enabling the builder to establish a log of normal operating voltages and parameters. A laboratory logbook which contains such data will be useful when it becomes necessary to troubleshoot the homemade equipment. The information will be valuable when designing new circuits which employ some of the stages and devices used in earlier assemblies.

Some amateurs have concluded that sophisticated and costly test equipment is needed to obtain high quality results. Certainly, this can be true if experimentation is taking place well within the state of the art. But, a lot of good work can be done with only a VOM. A great deal more can be achieved if the amateur is willing to construct some simple test equipment for his personal laboratory: A less than optimum measurement is still better than no measurement at all!

From the foregoing commentary emerges a primary rule which the writers have adopted: *Keep the test equipment simple!* Another principle they have embraced is that of not planning so far ahead that every application conceivable shall be handled by the assortment of homemade test equipment. The more esoteric pieces of laboratory gear can be built on an as-needed basis.

Some Basic Recommendations

The number of power supplies needed in the workshop always seems to exceed the quantity available. For this reason it is best to utilize power supplies which are outboard from the test equipment. The exception might be in the case of weak-signal sources which require superb isolation to minimize unwanted leakage.

Dry-battery packs of various voltage levels are useful to the experimenter. They are beneficial when it is necessary to effect a high degree of power supply isolation. Also, a variable-voltage regulated dc supply is extremely useful in the amateur laboratory. The circuit which illustrates the use of an LM317K IC (Fig. 48) is suggested.

Those who desire a high-current, ripple-free dc power supply may wish to consider inclusion of a 12-volt automotive battery in the shop. It can be "topped off" by means of a trickle charger when it is not being used. The life span of such a battery can be increased by periodic high-current loading and recharging, say, two or three times a week.

DC Voltage Measurements

Ordinary VOMs (volt-ohm-milliammeter) are suitable for much of the routine work done in the amateur lab. Some of the small imported instruments can be purchased for less money than one would spend to build a comparable tester from scratch. The primary limitation of most VOMs is, however, that of loading the circuit under test. A typical VOM will exhibit a characteristic of 1000 to perhaps 5000 ohms per volt when applied to a circuit test point. Loading of this variety will sometimes cause incorrect readings (lower than normal). A more practical voltmeter is one which has a high input resistance, such as a VTVM (vacuum-tube voltmeter) or a solid-state equivalent. The latter can often be built at a cost lower than that of a factory-assembled unit or commercial kit. The complexity of a homemade instrument will depend upon the accuracy desired. Some practical examples follow.

Low-Cost FET Voltmeter

Fig. 1 shows a simple voltmeter which uses one active device – a JFET. It is designed to accommodate two

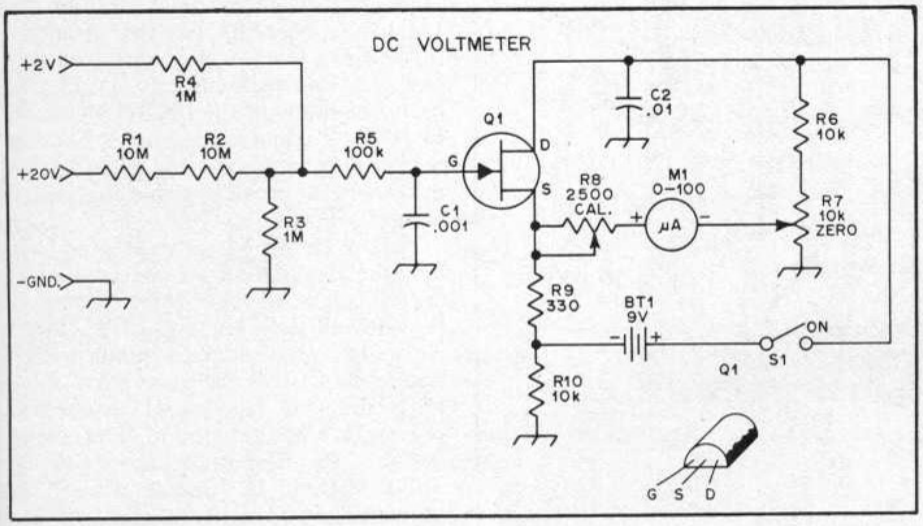


Fig. 1 – Circuit for the FET voltmeter. Fixed-value resistors are 1/4- or 1/2-W composition. C1 and C2 are disk ceramic. M1 is a 100-μA dc meter. Q1 is a Motorola MPF102 or HEP802. R7 and R8 are pc-board-mount composition controls.

dc-voltage ranges, 0 to 2 and 0 to 20 volts. For most amateur solid-state experimentation it will not be necessary to measure dc levels greater than 20. The accuracy of this instrument is ample for all but the most exacting applications (±10 percent).

As the dc voltage at the gate of Q1 is increased, the FET current rises, causing an elevation in the voltage drop across source resistance R9. The level change is indicated at M1, a 100-μA meter. Some current will flow in Q1 even when no dc voltage is applied to the gate. Therefore, control R7 is adjusted to provide a zero reading on M1. R8 is tweaked to provide a full-scale meter reading when two volts of dc are applied through R4. It may be necessary to readjust R7 and R8 a couple of times to effect final calibration.

When the voltmeter is first turned on by means of S1, there may be a short stabilization period caused by internal changes in the FET (junction heating). For this reason it is best to calibrate the voltmeter after it has been turned on for approximately one minute. When it is used for voltage measurements later on, allow a one-minute warm-up period to assure proper zeroing of the meter. Fig. 2 shows a circuit-board layout for the meter. Isolated pads have been formed by means of a Moto Tool and cutting bit. The builder may choose to mount R7 and R8 on the front panel of the tester case. This will permit recalibration of the circuit as the battery depletes. For greatest accuracy, R1 through R4, inclusive, should be 1-percent units. However, 5-percent resistors will suffice for most amateur work.

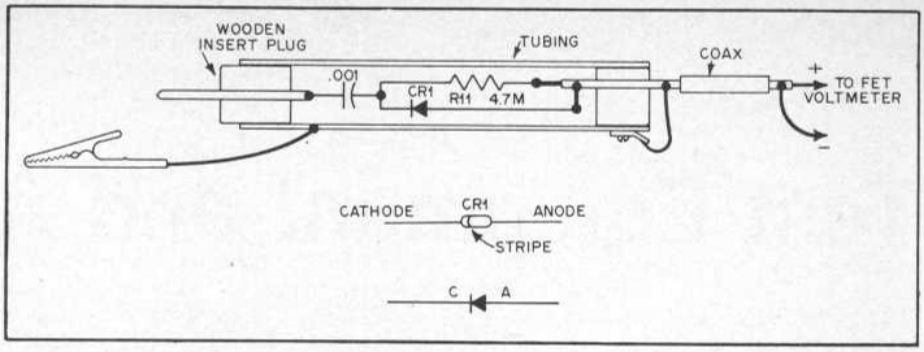


Fig. 3 – Details of the rf probe for use with VTVMs or the circuit of Fig. 1 (see text). CR1 is a 1N34A or equivalent. A 1N914A silicon diode is suitable also.

Readout on M1 will be linear. That is, full-scale deflection will represent 2 or 20 volts, depending on the range in use. Midscale readings will equal one and ten volts, respectively, and so on. The builder may find it helpful to draw a new meter scale, having two ranges represented – 0 to 2, and 0 to 20 volts.

Building an RF Probe

Fig. 3 shows how an rf probe can be built for use with the voltmeter of Fig. 1. It will be useful when determining relative rms values of rf voltage from 50 kHz to at least 148 MHz. It can be used with numerous commercial VTVMs to provide *accurate* rms voltage measurements, provided the voltmeter with which it is used has a 10-megohm input characteristic. However, when employed with the circuit of Fig. 1 the readings will not be perfectly coincident with the calibration of the meter at M1. The internal 4.7 megohm at Fig. 3 is chosen to change the peak rf voltage response of the probe to an rms value compatible with voltmeters which have the 10-megohm characteristic.

Despite the lack of accuracy resulting from utilizing the probe with the circuit at Fig. 1, signal tracing and relative rf voltage readings can be taken during circuit development or troubleshooting. When used with a 10-megohm instrument, best accuracy will result when the waveform under test is a pure sine wave. Distorted waveforms will change the voltage readings significantly.

The probe is made from a short length of copper tubing (3/8 or 1/2 inch in diameter). Wooden end plugs are installed to fit snugly inside the tubing. The probe tip can be made from a small nail or a piece of brazing rod which has been sharpened to a point on one end.

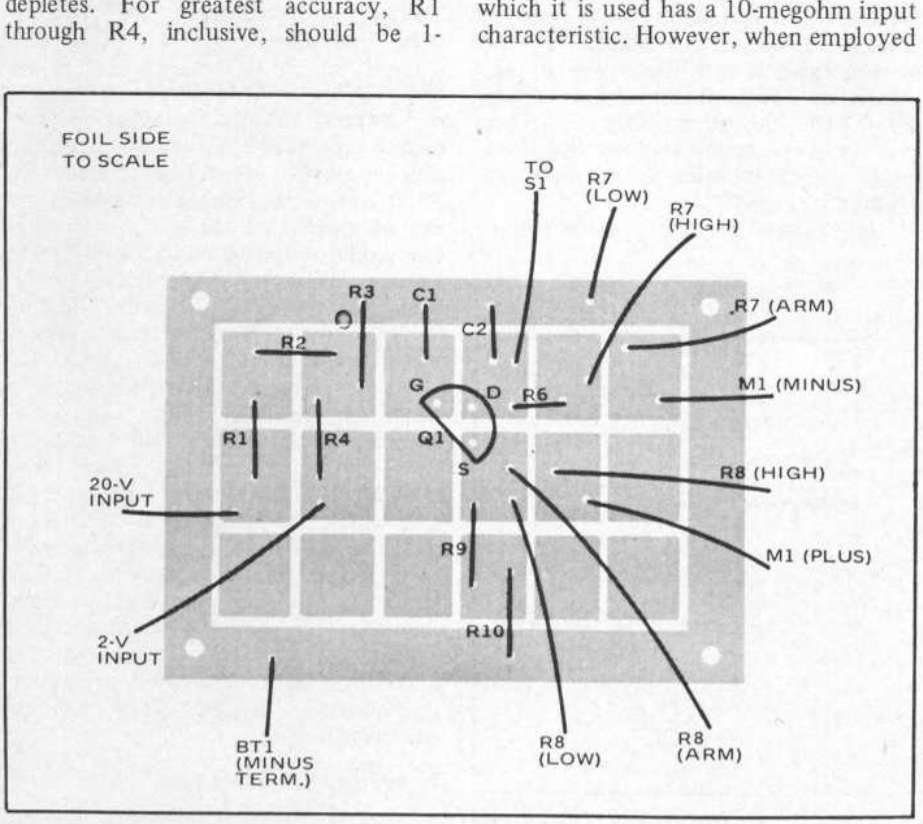


Fig. 2 – Circuit-board pattern for the circuit of Fig. 1. The metal between the copper pads can be removed by means of a hobby tool and cutting bit.

Op-Amp Voltmeter

Shown in Fig. 4 is a simple voltmeter that uses a pair of op-amp ICs and a 0-1 mA meter. Type 741 op amps may be used. A better choice would be the LM-308N. This unit has the advantage of requiring low power from the battery and has low bias currents, leading to better accuracy. If the LM-308N is used, a 1000-pF capacitor should be connected between pins 1 and 8 of the chip in order to provide stable frequency compensation.

In this circuit U1 serves as a fed-back current amplifier. Two input resistors are selected with a slide switch to provide full-scale readings of 2 and 20 volts. The gain of the circuit is 0.5, leading to a 1-volt change at pin 6 of U1 for a full-scale reading. U2 is used to provide a synthetic ground. This allows the circuit to be powered from a single, 9-volt battery of the kind used in transistorized bc-band receivers.

A pair of diodes is provided at the input to protect the semiconductors from excessive input voltages. The two

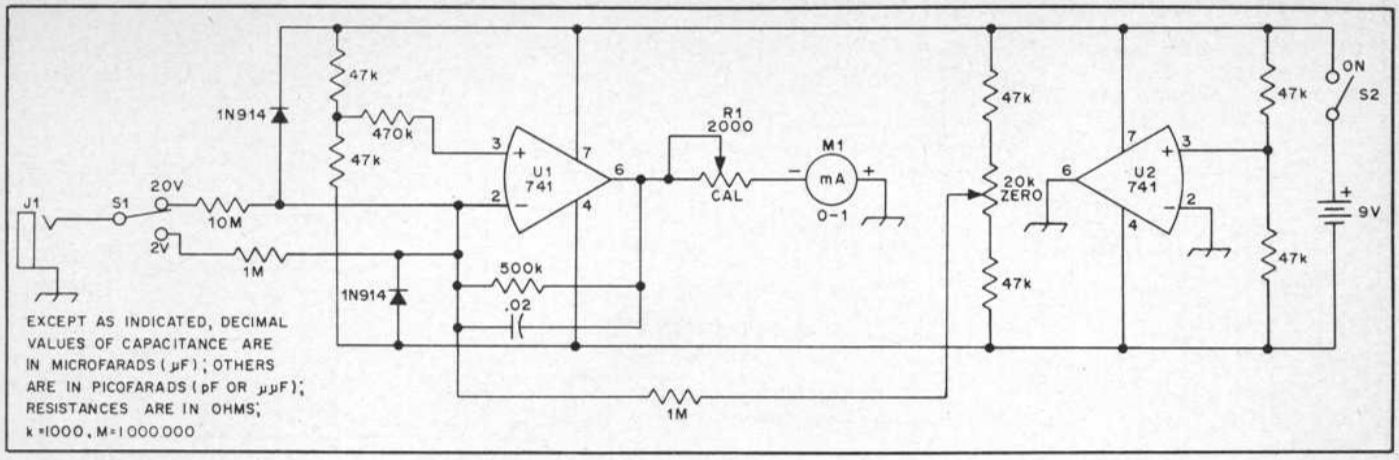


Fig. 4 — Circuit of the op-amp voltmeter.

controls in the circuit serve to calibrate the meter movement and to zero the output when there is no input signal.

This meter functions like a VOM with a sensitivity of 500 kΩ per volt. The input resistance changes for the different ranges. Because of this, the circuit cannot be used with the usual rf probe. Most rf probes are built to work with a VTVM or FET voltmeter that has a constant input resistance of 10 megohms. As in Fig. 3, they usually contain a 4.7 megohm resistor. Such a probe could be used with good accuracy on the 20-volt range of the meter in Fig. 4, but errors would occur on the 2-volt scale.

Shown in Fig. 5 is another FET voltmeter. This circuit is the semiconductor equivalent of some popular VTVMs. A dual FET is used in this circuit, resulting in exceptionally low drift characteristics with temperature changes. Also, the FET chosen is a unit with a low pinchoff voltage. This has the asset that the meter may be powered from a low-voltage supply. The required 6 volts are provided by means of four D-type dry cells. Since the current consumption is only a few mA, Pen-light cells would serve as well. The circuit is a full differential amplifier. Each side consists of the FET and a pnp transistor arranged as a noninverting amplifier with feedback to produce a voltage gain of 2. The output of this amplifier is applied to an emitter follower to drive the meter.

The dual JFET used in the schematic may be a difficult item to obtain. However, if the voltage is increased in the circuit, almost any dual FET will work. If a dual FET cannot be located, the modified amplifier shown in Fig. 6 is recommended, where individual FETs of the same type are used. Two units of similar characteristics should be chosen. They should be matched for I_{dss} and pinchoff voltage.

The unit utilizes a meter with a 0-1 mA movement, but with three scales labeled 0-70, 0-140 and 0-350. The resistive divider was designed specifically to be compatible with these scales, with a circuit sensitivity of 0.35 volt full scale. In the circuit shown in Fig. 5, the basic sensitivity is assumed to be 0.5-volt full scale, and the resistive divider has been designed to yield full-scale sensitivities of 0.5, 1, 2, 5, 10, 20, 50, 100, 200 and 500 volts. The sensitivity is controlled with the range switch, S1. A double-pole, double-throw slide switch, S2, is used for polarity reversal, while S3 serves to switch power to the meter. Although not shown in the schematic, a second set of contacts on S3 is arranged to short out the meter movement when the unit is off. This is a good practice with high quality meter movements to prevent damage during transit.

In the modified circuit of Fig. 6, the pair of FETs are used as source followers to drive a pair of 741 op amps. The 741s then drive the meter. This circuit could use either a 747 or a 5558 dual op amp in place of the two 741s. While the

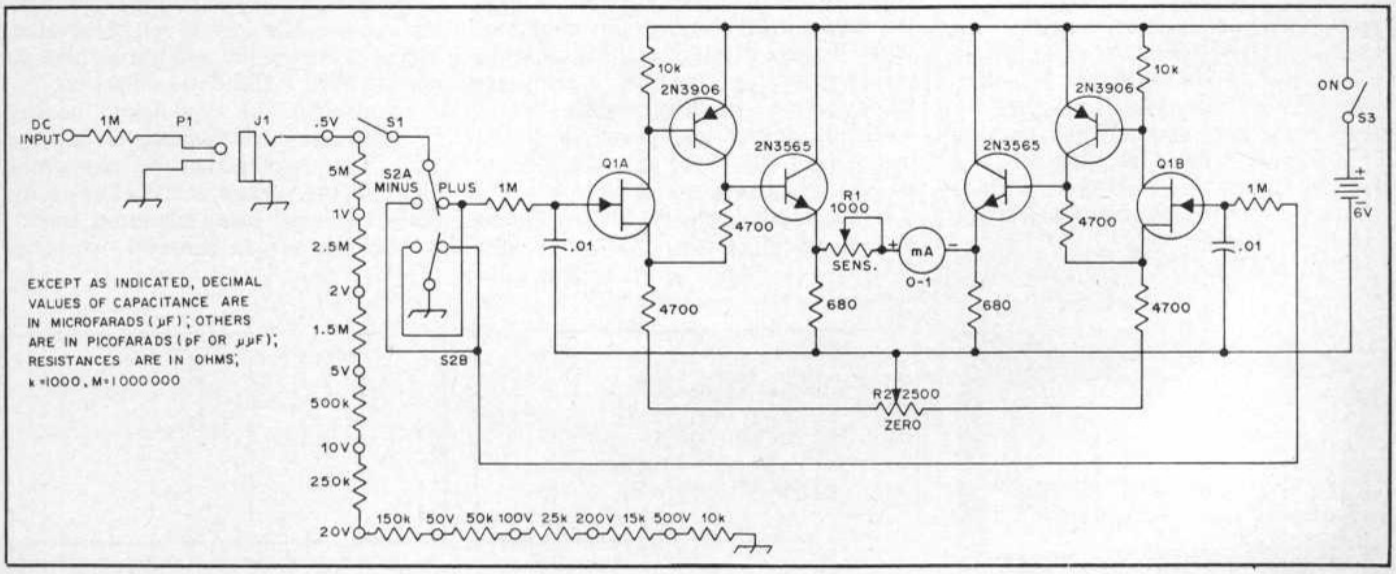


Fig. 5 — A semiconductor equivalent of some of the popular VTVMs used by amateurs. The circuit exhibits low-drift characteristics respective to temperature changes. Q1 = Dual N-channel JFET, $V_p \approx 1.5V$.

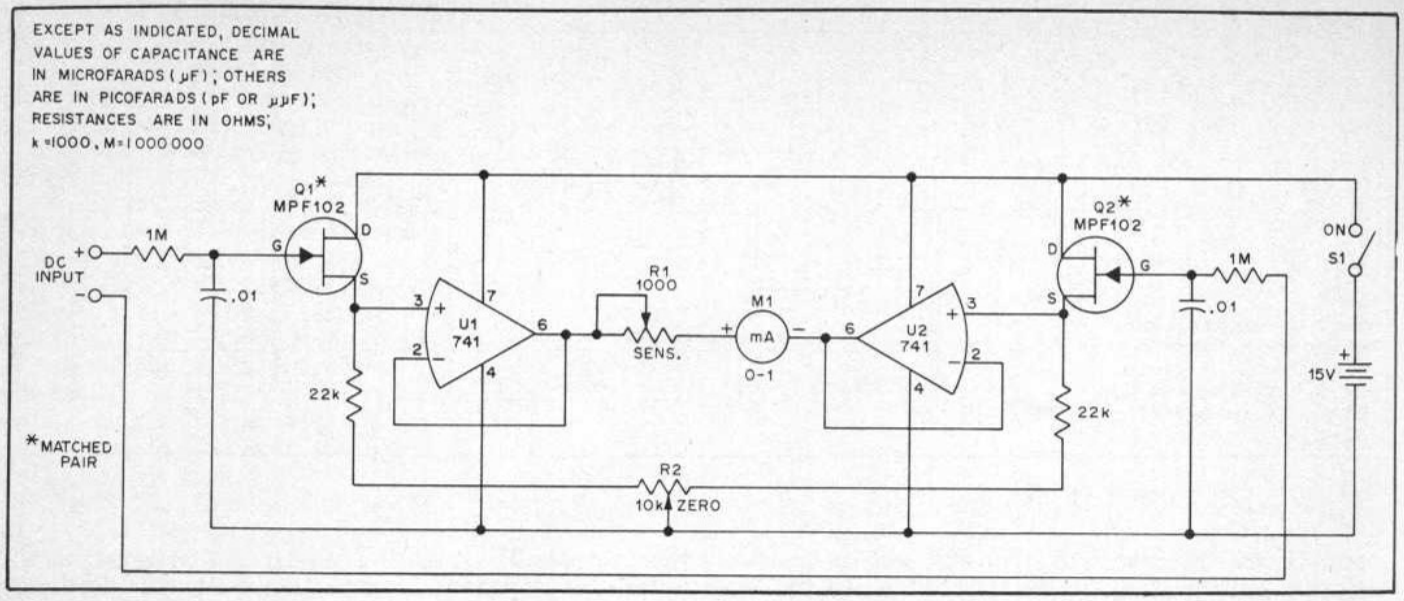


Fig. 6 — Alternative to the circuit of Fig. 5 for those who cannot obtain a dual FET.

drift of this circuit is certain to be greater than when a dual JFET is used, it should still be better than those circuits which contain a single FET.

RF Power Measurement

One of the most frequent measurements performed by the amateur experimenter is that of rf power. The most common application is during the testing of transmitters. The receiver builder needs to know the power available from his LO and BFO. Also, if he is to evaluate the dynamic range of his receiver, he must have signal generators with known output powers. These are obtained with low-power oscillators followed by a step attenuator. The output power must be measured before application of the attenuator.

For hf transmitter work, rf power is most easily measured with a high-level diode detector and a dummy load or termination. A circuit suitable for powers of 10 or 15 watts for short time periods is shown in Fig. 7. Six 300-ohm, 2-W resistors have been paralleled to serve as the termination, R1. Detection is performed with a 1N914 diode, and the dc voltage is monitored with a voltmeter. Any VOM is suitable at the higher power levels.

The diode serves as a peak detector. That is, the largest positive voltage appearing across the 50-ohm termination is the value that the capacitor attains, and is measured by the voltmeter. For a sine-wave input, which is the usual waveform of interest, the power is given as $P = V^2_{dc} \div 2R$ where R is the termination, in this case equal to 50 ohms.

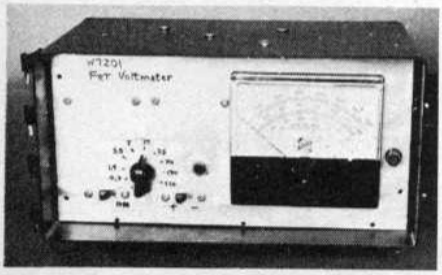
As higher powers are to be measured, simple techniques like those shown in Fig. 7 may not be suitable. The reason is that the peak reverse voltage appearing across the diode may exceed the diode breakdown specification. One simple way of circumventing this problem is shown in Fig. 8 where a voltage divider is placed across the termination. The net termination should still equal 50 ohms. The measured voltage must be multiplied by the appropriate division factor in order to calculate the power with the previous equation. With voltage-divider techniques, the power-measuring capability is easily extended to the 1-kW level.

Significant errors appear when the methods of Fig. 7 are extended to low powers. The major source of error is the V-I characteristic of the diode. Recall that a silicon diode like the 1N914 requires about 0.6 to 0.7 volt across it before significant current flows. Hence, with rf powers corresponding to a peak voltage of 0.6 volt, no detected output will appear. (Actually, there may be some, but the accuracy of the measurement will be poor.)

The first step toward better sensitivity is to substitute a more sensitive diode type. Either a germanium or a hot-carrier silicon diode would be a much better choice, since they turn on at much lower voltages. Values for diode turn-on voltage down to 0.1 to 0.2 volts are common. For best accuracy the voltmeter should draw minimal current from the detector. Hence, a VTVM or FET voltmeter is preferred over a simple voltmeter.

Shown in Fig. 9 is a power meter that is built on the back of a 500-μA meter. This unit uses a hot-carrier diode detector and will yield an indication for input powers as low as +1 or +2 dBm. The resistor was chosen for a full-scale reading of +17 dBm (50 milliwatts).

A meter of this type cannot be used to determine power with a simple formula. The reason is that the value of the diode offset voltage is too close to the peak rf voltages being measured, leading to excessive errors. However, meters of



Solid-state voltmeter which uses FETs.

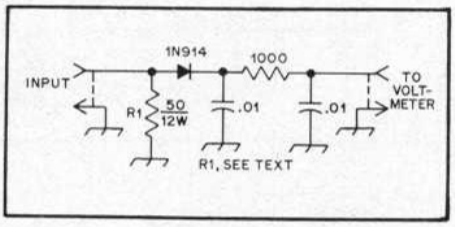


Fig. 7 — A high-level diode detector for rf power measurements into a dummy load. See text for information concerning R1.

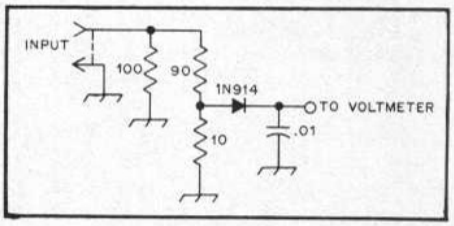
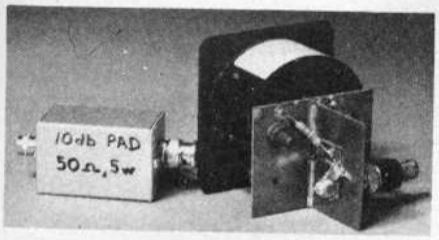
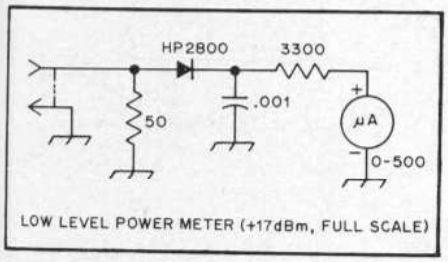


Fig. 8 — A voltage divider can be employed to increase the power measuring capability of the circuit shown in Fig. 7.



Rf power meter seen assembled on the back of a meter.

Fig. 9 — Circuit of an rf power meter which can be built on the back of a 500-μA meter.



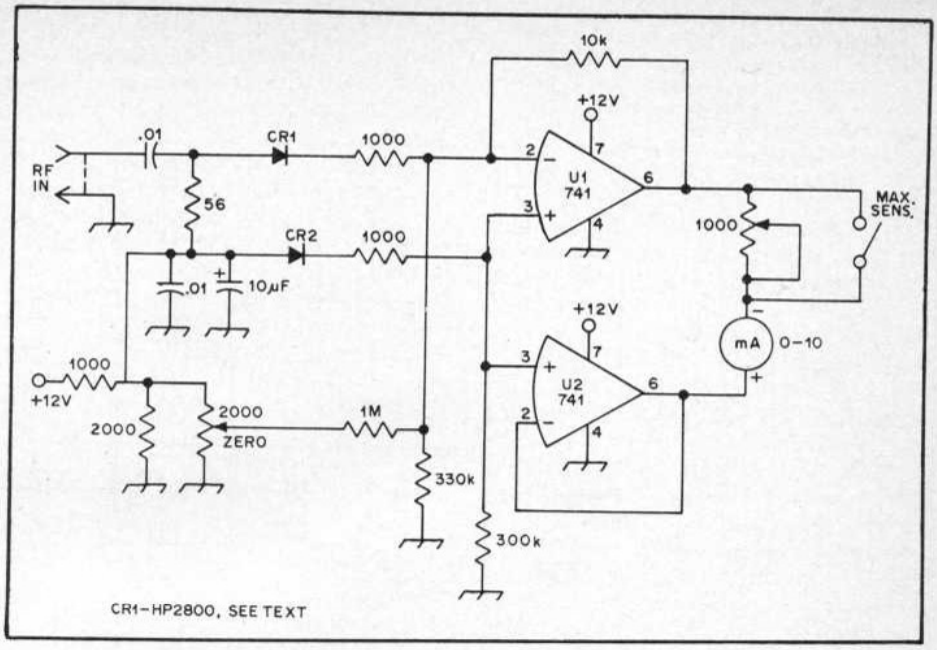


Fig. 11 — Circuit for proper biasing to obtain square-law detection.

this kind are easily calibrated by noting that the circuit is still a peak-reading detector. This allows a dc calibration to be done.

Imagine that a power of 10 mW was to be measured. This power would correspond to 1-volt peak across a 50-ohm resistor. To calibrate the meter for 10 mW, place 1-volt dc across the termination and note the meter response. Similarly, 2-volts dc would correspond to 40 mW. Using this method, a calibration curve can be generated for the power meter. In the unit shown, such a calibration was found to correspond within 1 dB of that from industrial instrumentation.

While a sensitivity near 1 mW is adequate for most situations, it is often useful to be able to measure powers which are much lower. One approach to this would be to precede the diode detector with a broadband amplifier. A better approach, however, is to increase the basic detector sensitivity before adding amplifiers. The simplest way to do this is by biasing the diode detector with dc.

Shown in Fig. 10 is a small-signal waveform applied to a diode detector and the resulting output. Note that an input voltage as small as that shown (about 0.1-volt peak) would produce no current in a diode with zero bias. However, when the voltage is applied to the biased diode, we see a definite current flow. The current that flows is not what we would expect if the diode were replaced with a resistor. Instead, we see that the positive-going half of the input voltage yields a much larger current flow than the negative part. The result is that if the diode current is monitored, a dc component is present. This form of detection is usually referred to as "square law" detection. The mathematics are outlined in the appendix under a discussion of distortion phenomena.

In order to achieve square-law action, a diode must be biased carefully. Specifically, it should be biased at a constant current level from a low impedance dc source. While this could be achieved with a battery and a variable resistor, a much better method is to use an operational amplifier.

Shown in Fig. 11 is a circuit to accomplish this biasing. A pair of identical diodes are used. However, only one (CR1) has rf applied. The other serves as a reference for properly biasing the detector. With this circuit, input powers as low as −26 dBm (3 microwatts) can be detected.

The calibration is straightforward. An oscillator is built to deliver about +10 dBm output. This power is easily measured with the peak detector described earlier. The oscillator output is applied to a step attenuator with up to a 40-dB range. The available output powers are now suitable for the square-law detector, and are well defined within the errors of the collection of instruments.

The diode square-law detector is quite flat from about 1 MHz up through the vhf spectrum. Either hot-carrier diodes or small-signal silicon switching diodes can be used. If better op amps were used with lower drift specification,

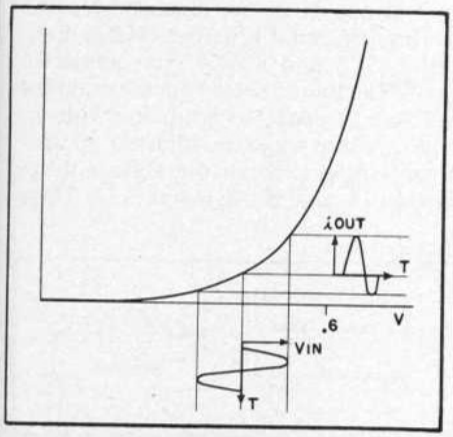


Fig. 10 — Small-signal waveform applied to a diode detector and the resultant output.

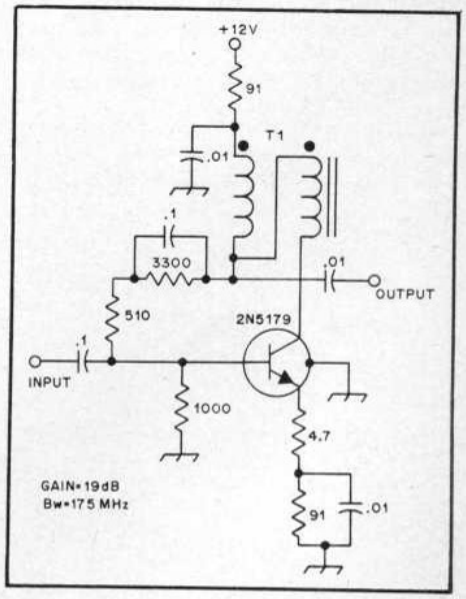


Fig. 12 — Diagram of a broadband amplifier which can be used to extend power-meter sensitivity to lower power levels. T1 contains 7 bifilar turns of enameled wire on an Amidon FT-23-43 toroid core. Circuit gain is 19 dB and the bandwidth is 175 MHz.

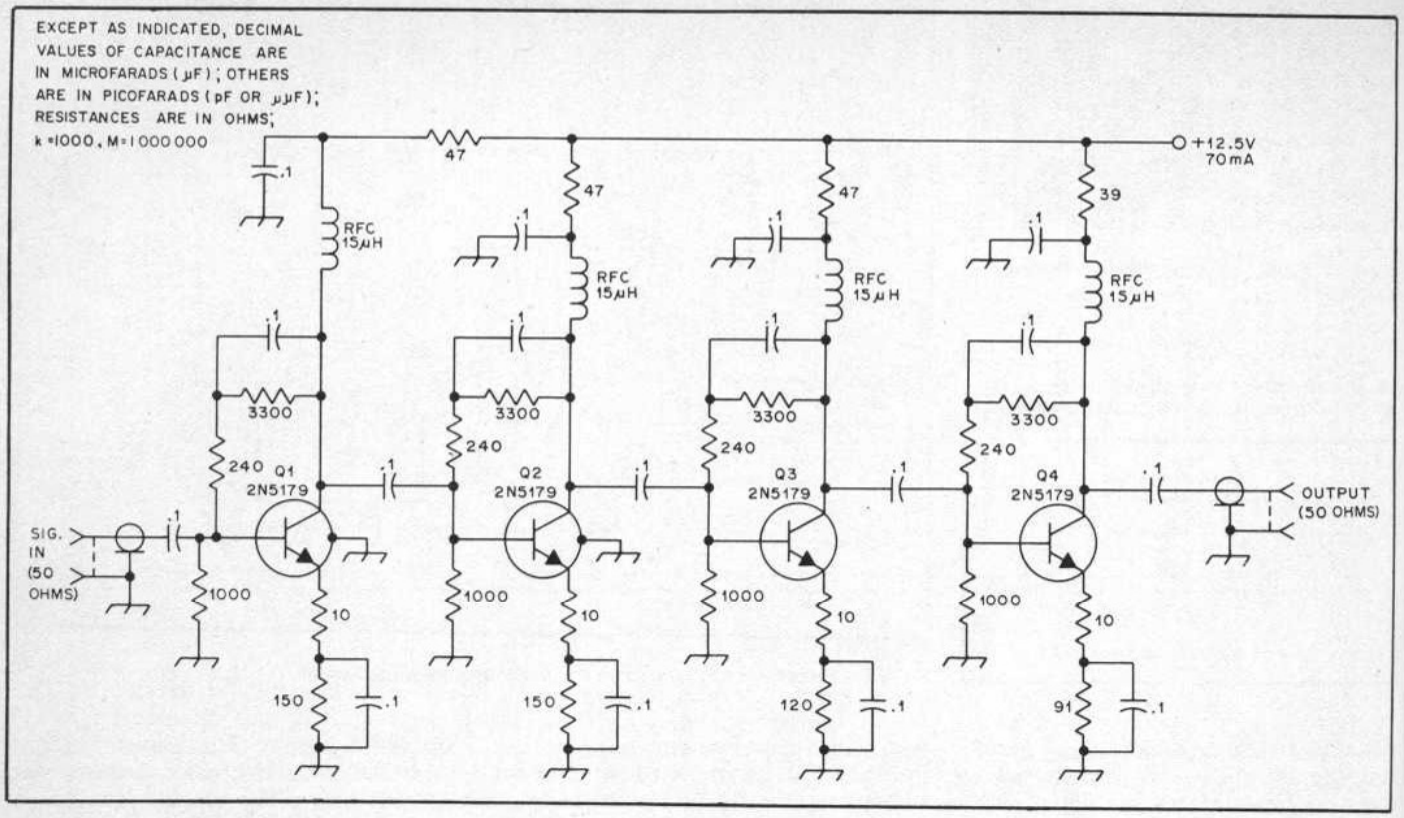


Fig. 13 — A four-stage broadband rf amplifier. Gain = 40 dB and the upper 3-dB point of the amplifier is 65 MHz.

the system could be operated with higher dc gain, yielding even better sensitivity. Some manufacturers make diodes which will detect signals down to −50 dBm.

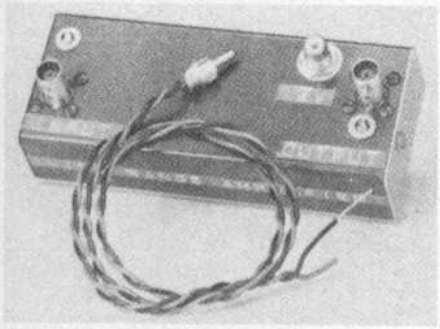
The best way to extend sensitivity to lower power levels is with a broadband amplifier. Shown in Fig. 12 is a single-stage amplifier using a 2N5179. Heavy feedback is used to stabilize gain and to provide 50-ohm input and output impedances. The 3-dB points in this circuit were about 2 MHz and 175 MHz.

The 50-ohm transducer gain was 19 dB, the noise figure 6.5 dB (at 10 MHz), and the output intercept +24 dBm. Gain compression starts near +10 dBm.

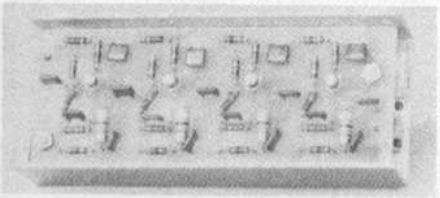
Shown in Fig. 13 is a four-stage amplifier. The upper 3-dB point in this amplifier was about 65 MHz and the gain was 40 dB. Noise figure was not measured.

These amplifiers are useful accessories for applications other than power measurements. For example, they may be used as preamplifiers for a frequency counter, or even a receiver.

Shown in Fig. 14 is a block diagram of a useful general-purpose instrument. An attenuator, amplifier and sensitive detector are combined for a wide sensitivity range. If the input is driven from an outboard tuned circuit, a wave meter of spectacular sensitivity would result.



Exterior of the broadband, 50-ohm amplifier.



Interior layout of the broadband amplifier.

In-Line RF Power Measurement

Rf power measurements can be made accurately at specified impedance levels by using an rf bridge circuit of the type illustrated in Fig. 15. The basic circuit was described by Bruene in *QST* for April, 1959. The concept was treated in a practical manner by DeMaw in *QST* for Dec., 1969.

The principle of operation is that the inner conductor of a coaxial transmission line passes through the center of toroidal transformer T1 to function as the transformer primary. A multiturn secondary winding is placed on the core. Rf current through the primary induces a voltage in the secondary, causing current to flow through R1 and R2. The voltage drops across these resistors are equal in amplitude, but are 180 degrees out of phase with respect to common, or ground. Practically speaking, they are in and out of phase, respectively, with the line current. Capacitive voltage dividers, C1/C3 and C2/C4, are connected across the line to secure equal-amplitude voltages *in phase* with the line voltage. The division ratio is adjusted so that these voltages match the voltage drops across R1 and R2 in amplitude. These

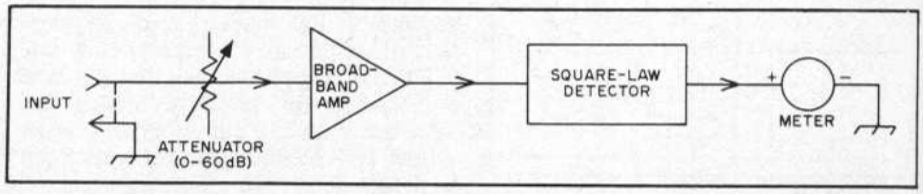


Fig. 14 — This block diagram illustrates a test instrument which contains an attenuator, amplifier and sensitive detector.

conditions exist at only a specified load impedance – usually 50 or 75 ohms to match the characteristics of the transmission line. Initial adjustment of the bridge is done while using a resistive load standard of the value desired.

Under the foregoing conditions, the voltages rectified by CR1 and CR2 represent, in one case, vector *sum* of the voltages caused by the line current and voltage. In the other case, the vector *difference* is represented. With respect to the resistance for which the circuit has been adjusted, the sum is proportional to the forward component of a traveling wave of the variety that occurs on a transmission line, and the difference is proportional to the reflected component.

Fig. 15A shows the main portion of the power bridge as being contained in a shielded enclosure, as indicated by the dashed lines. External to the shield are the components needed to meter the forward and reflected components. In the example at A, a single potentiometer is used to set the full-scale power indication of M1. In this case R3 can be calibrated for various full-scale power levels by observing the rms output voltage from the bridge with an rf probe, or the pk-pk value by means of a scope. The voltage is measured across a resistive termination which matches the characteristic impedance of the bridge unit. A 10-turn Helipot and mating dial mechanism will allow greater reset accuracy than will a simple control-and-knob arrangement.

Fig. 15B shows an alternative technique for presetting the instrument for a specific full-scale power level. Trimpots can be mounted inside the instrument case and adjusted for a particular power sensitivity; e.g., 10, 50, 100, 500 or 1000 watts. If more than one power range is desired, an assortment of controls can be used, then switch-selected for the power ranges required.

It is important to maintain good isolation between the through-line ports, and between the line and the remainder of the bridge circuit. It is good practice to use an isolating divider such as that seen in the photograph of Fig. 16. Some manufacturers who follow this general design, utilize a Faraday screen between the primary and secondary windings of T1. This helps prevent unwanted capacitive coupling, thereby aiding the nulling of the bridge circuit.

The bridge is balanced by connecting a 50-ohm signal source to the input port, and terminating the output port in 50 ohms, resistive. With the instrument set to read reflected energy, C1 is adjusted for a zero reading at M1. The load and source cables are reversed next, and the procedure repeated while adjusting C2 for a zero meter reading. Following the null adjustments the builder can calibrate the instrument for a specific full-scale power level, as discussed earlier in this treatment. Bridges of this general type are suitable for use

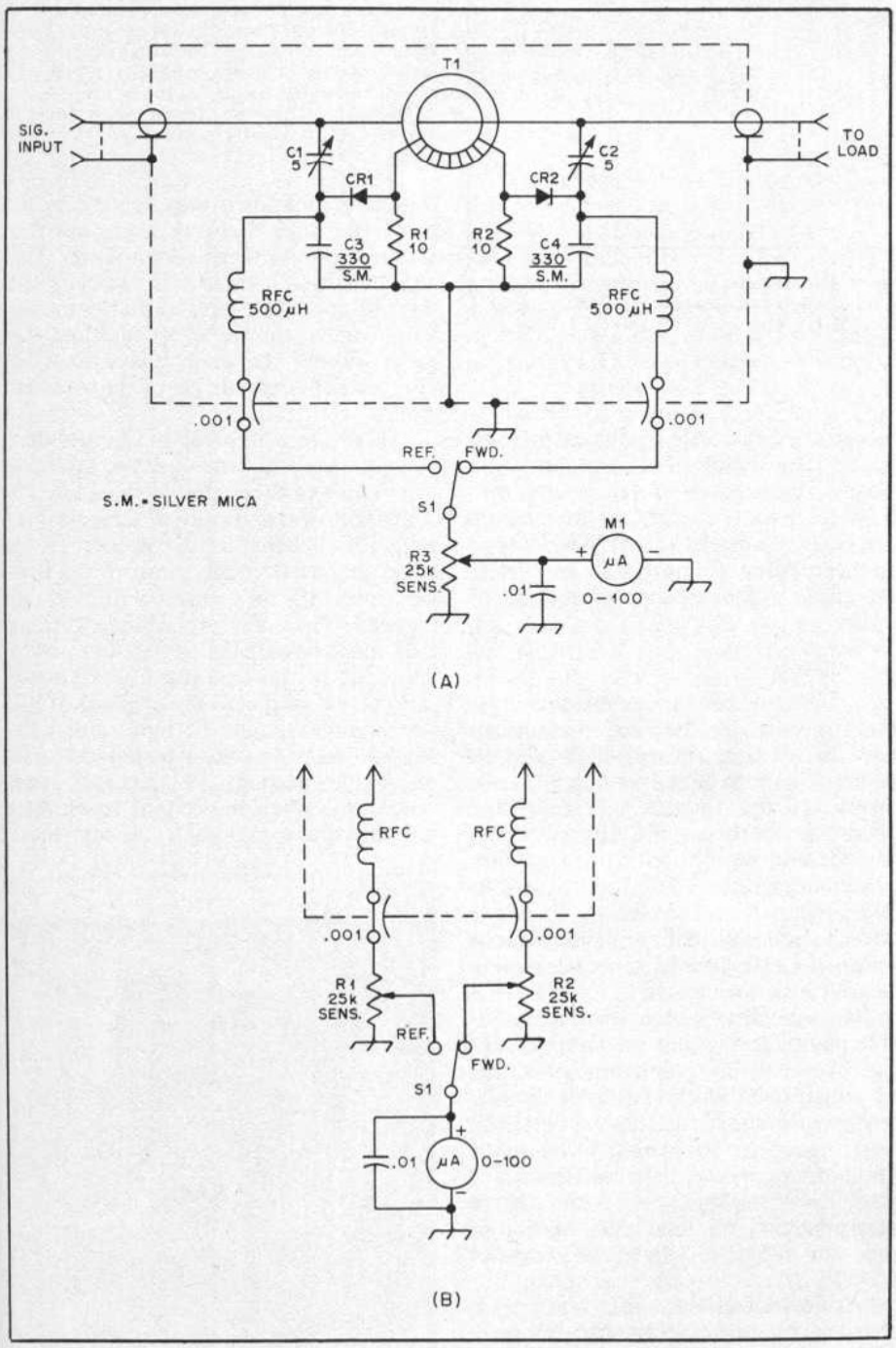


Fig. 15 – Schematic diagram of an rf power bridge. T1 has 60 turns of no. 30 enameled wire and uses an Amidon T68-2 toroid core. C1 and C2 should be piston or air trimmers to assure a low minimum capacitance. CR1 and CR2 can be 1N34A or 1N914A diodes (matched pair recommended). See text for a discussion of the circuits at A and B.

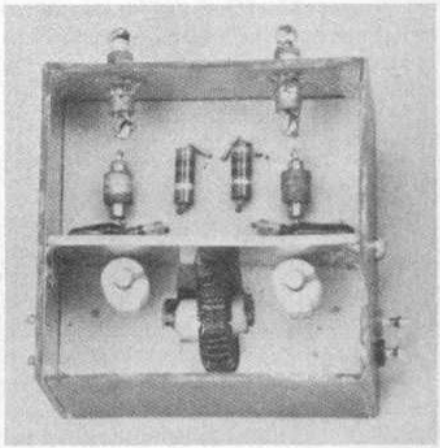


Fig. 16 – Photograph which shows a shield divider between the rf and dc portions of the bridge (double-sided pc-board strip across center of box).

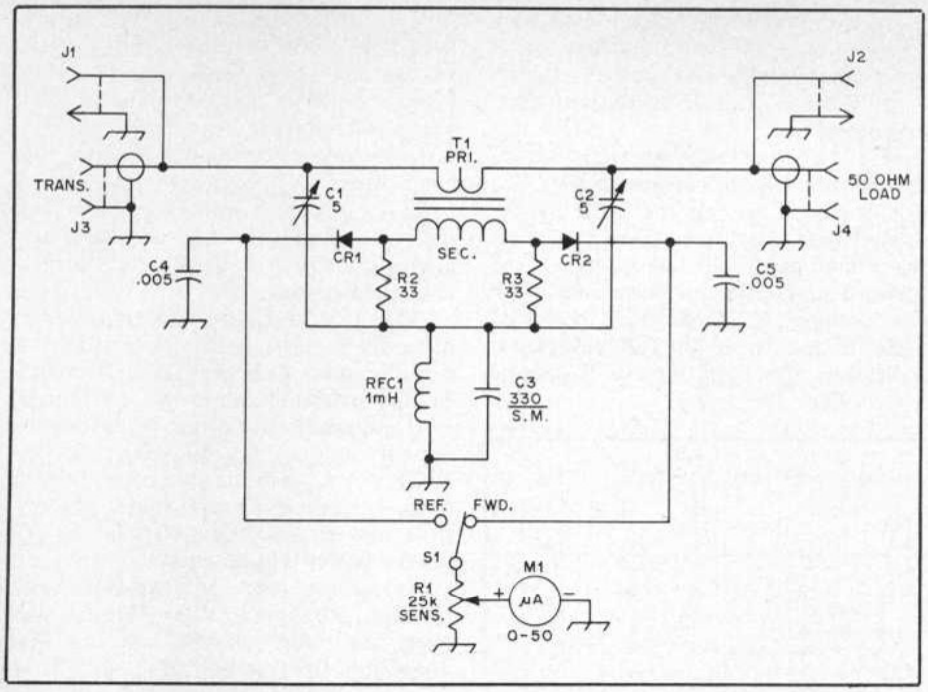


Fig. 17 — Schematic diagram of a QRP rf power meter. It is suitable for levels from 1 to 100 watts, 1.8 to 30 MHz. T1 contains 60 turns of No. 30 enameled wire and uses a T68-2 toroid core. The primary of T1 consists of two turns of No. 20 insulated wire. C1 and C2 follow the rule set forth for the circuit of Fig. 15.

up to 30 MHz. The lower frequency limit, with the component values given, is approximately 1.8 MHz. If a pc-board format is used, the constructor may elect to employ pc-board strip-line techniques to assure a relatively constant 50-ohm line characteristic between the input and output ports. The value of such an approach will be seen at 21 MHz and higher, where the composite bridge can cause a slight line-impedance discontinuity (a line "bump") if the through-line is not close to 50 ohms.

If a 50-μA meter is used at M1, maximum forward-power sensitivity for this circuit will be on the order of 10 watts. This type of bridge is not "frequency conscious," as is the Monimatch circuit popularized in *QST*. That is, it will respond uniformly to a given power level from 1.8 to 30 MHz. Nulling adjustments should be done at the highest frequency of use (30 MHz in this example).

A QRP Power Meter

Fig. 17 illustrates a suitable bridge for use in measuring power levels from 1 to 100 watts. The circuit is a variation of that shown in Fig. 15. To increase the sensitivity, a two-turn link is used for the primary. This represents a slight tradeoff in through-line impedance at the higher end of the hf spectrum, but the line discontinuity is not great enough in magnitude to spoil the utility of the instrument.

Figs. 18 and 19 show the construction technique used. R1 has been calibrated for a full-scale reading at M1 of 5 watts. The calibration chart atop the bridge case shows power levels from 0.25 to 5 watts, versus the meter-scale markings. Phono jacks and SO-239 type connectors are connected in parallel at the input and output ports, purely for utility.

Attenuators

An attenuator is one of the most useful accessories that the amateur can have in his shop. It will allow a given power source to be reduced by a known factor. If the amount is variable, as would be the case with a step attenuator, the unit can be used with a sensitive power-measuring meter in order to determine gain and to evaluate linearity. Attenuators may be used to extend the range of existing sensitive power meters to arbitrarily high levels.

High quality attenuators are available commercially and are fairly expensive. Alternatively, step attenuators may be constructed from resistors and slide switches. While the accuracy is certainly not as good as one would realize with better units, it is usually sufficient for amateur work. Again, we offer that a measurement of less than optimum precision is better than *no* measurement!

Attenuators have assets other than reducing the power in a controlled way. Since they are made from resistors, they will change a source or load that may be highly reactive into one that is known and resistive. Similarly, a source or load

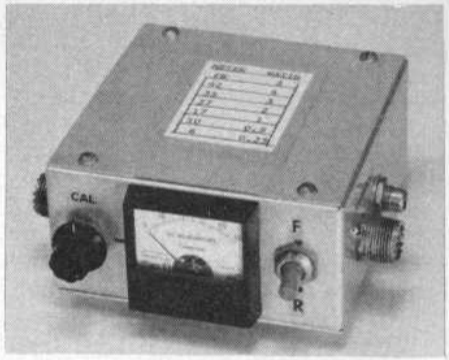


Fig. 18 — Exterior view of the QRP power meter. A 4 × 4 × 2-inch aluminum utility box serves as a case. Phono jacks are placed in parallel with uhf connectors at the input and output ports of the unit to permit use of a greater variety of cable connectors. A calibration mark can be seen at the left center of the meter. The mark represents the control setting for 5 watts full scale.

that is of unknown impedance may be made to appear as a clean, resistive termination with an attenuator. The extent to which these effects occur will depend upon the amount of attenuation employed — the more attenuation, the more closely the load approaches the characteristic impedance of the attenuator.

There are a number of circuits that may be used to form resistive attenuators. Three of these are shown in Fig. 20 along with the appropriate design equations for choosing resistor values. These equations are derived easily from first principles if the experimenter is so inclined. There are two vital conditions that must be satisfied. First, the power delivered to the load must be a known ratio of that supplied to the input of the attenuator. Second, the input resistance seen at one end of the attenuator should equal the desired characteristic resistance, R_o, when the output is terminated in the same value. Using these

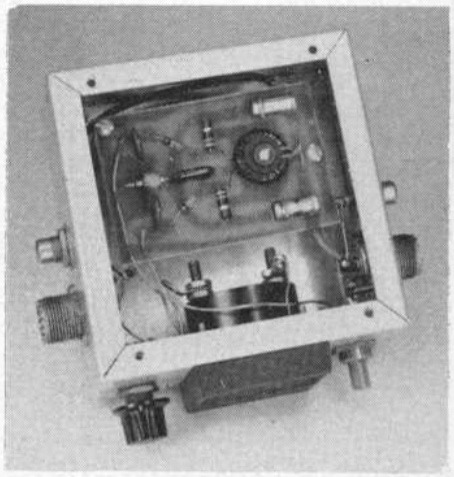


Fig. 19 — Interior view of the power meter of Fig. 17.

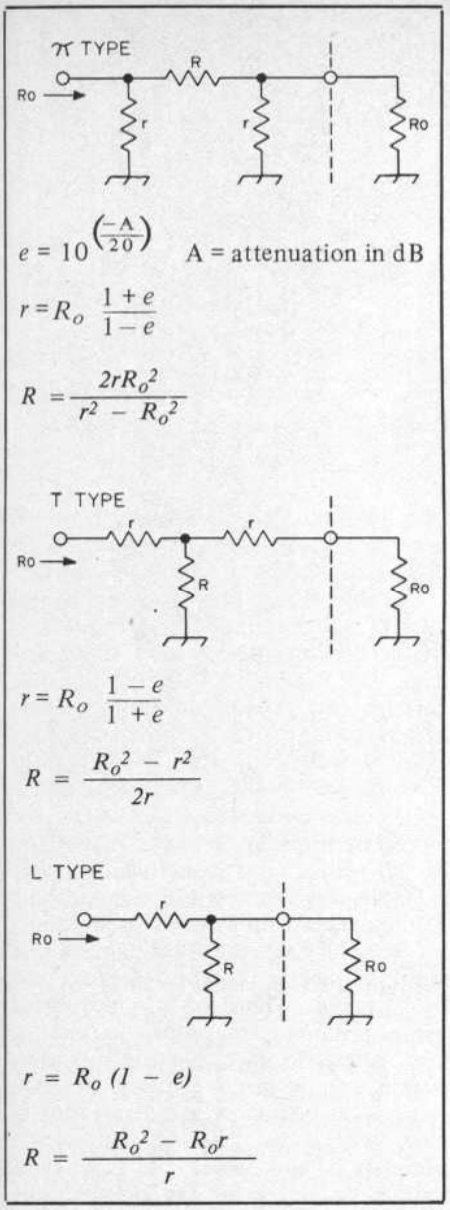


Fig. 20 — Three circuits for forming resistive attenuators.

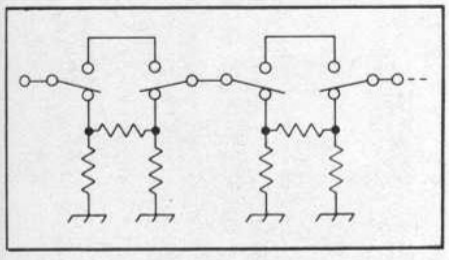
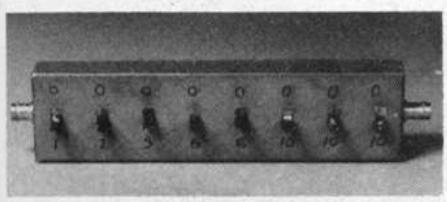


Fig. 21 — Circuit for a step attenuator which is useful into the vhf spectrum.



Outside view of the step attenuator.

Table 1

5-Percent Resistor Values for Simple Attenuators

	π		T		L	
A, dB	r	R	r	R	r	R
1	910	6.2	2.7	390	5.6	390
2	430	12	5.6	220	10	200
3	300	18	9.1	150	15	120
6	150	39	16	62	24	51
10	91	68	27	36	33	24
20	62	240	39	10	43	5.1

conditions, the equations may be set up so that, when solved, they yield the design equations shown.

When using the equations in Fig. 20, A is the attenuation ratio in dB. The voltage attenuation ratio, "*e*," is related to A with the equation given in the figure.

Care should be used in the construction of attenuators with slide switches. If 1-percent tolerance resistors are available, they should be used. However, the results are often quite suitable with 5-percent resistors. Every effort should be made to keep the lead lengths as short as possible. This will help to extend the upper frequency of usefulness. Shields are beneficial if the unit is to be used at vhf. This is especially significant for single sections of 20 dB or more.

Three types of attenuator are shown: a pi, a T and an L circuit. The pi and the T are symmetrical, and are, thus, the more useful types. The L circuit has the problem that the output resistance of the section may be much different than the input resistance of the circuit. In some cases, this presents no obstacle. For switchable attenuators, the pi circuit offers the best compatibility with the slide switches. A circuit for a step attenuator is shown in Fig. 21. The photograph shows a unit that offers good accuracy up through the vhf spectrum.

Shown in Table 1 is a list of values of common 5-percent resistors that may be used for various amounts of attenuation. Half- or quarter-watt resistors are suitable for small-signal work. For higher power units, the specific circuit must be evaluated carefully to ascertain the power specifications of the resistors. As an example, consider the 10-dB pi attenuator shown in Fig. 22, and assume that it is to be designed for a resistance of 50 ohms. Assume that the maximum input power, when properly terminated, will be 10 watts, which corresponds to a voltage of 22.4 across 50 ohms.

Solving the equations given earlier, the resistor values are 96.3 ohms at the ends and 71.2 ohms for the connecting arm. If we solve for the voltages, which are shown in circles in Fig. 22, we may calculate the powers dissipated in the three resistors. The input resistor dissipates 5.19 watts, the 71.2-ohm resistor consumes 3.29 watts, while the output resistor consumes only 0.52 watt. A good choice for the input resistor would be a parallel combination of two 300-ohm ones and a 270-ohm unit, all with a 2-watt dissipation rating. The connecting arm could be another parallel pair of 2-watt units with resistances of 150 and 130 ohms. The output could be a 1-watt, 91-ohm resistor. If such an attenuator was built for rf power measurement, the input should be clearly marked.

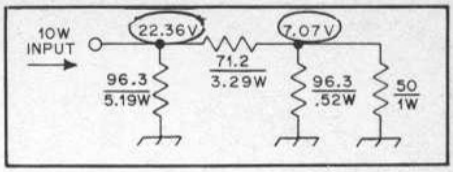


Fig. 22 — A 10-dB pi type of attenuator.

The attenuators discussed here have been dissipative devices, with some of the input power applied to them being absorbed within the circuit. However, other methods are useful for measurement applications that are not dissipative. Shown in Fig. 23 is one example, a 20-dB coupler. This is a high-permeability ferrite toroid core set up as a current transformer. The primary of the transformer is a single wire passing through the core while the secondary is a 10-turn winding. If the secondary is terminated in a 50-ohm load, such as a low-level power meter, this termination will reflect back through the transformer according to the square of the turns ratio. Hence, the core will appear as a 0.5-ohm resistor in series with the line. If the main line is also terminated in 50 ohms, the net resistance presented to the source is 50.5 ohms (essentially unchanged). Noting that the ratio of the two resistances is 100, or 20 dB, the power delivered to the power meter will be attenuated from that delivered to the main load by 20 dB. Techniques of this kind can be applied to the evaluation of higher power sources (such as trans-

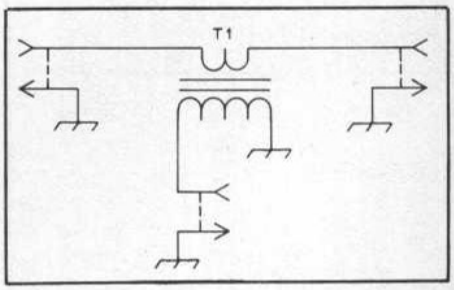


Fig. 23 — An example of a 20-dB coupler. T1 uses a single wire through the toroid core as the primary. The secondary is a 10-turn winding of enameled wire. An FT-23-43 core is used.

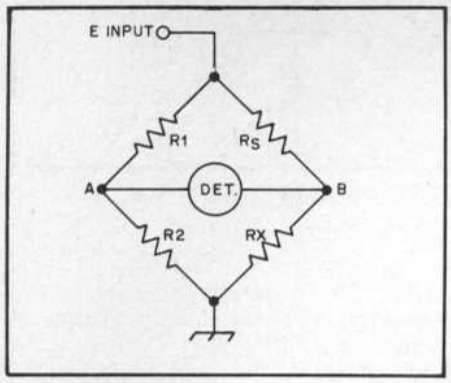


Fig. 24 – A Wheatstone bridge for measuring dc resistance.

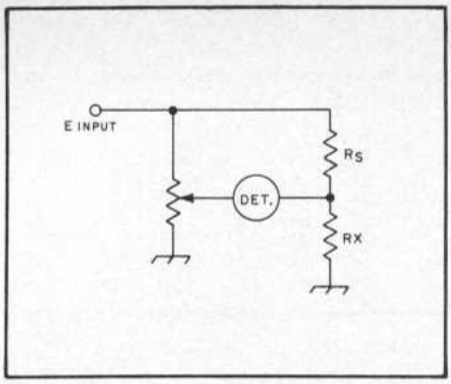


Fig. 25 – An alternative to the circuit of Fig. 24.

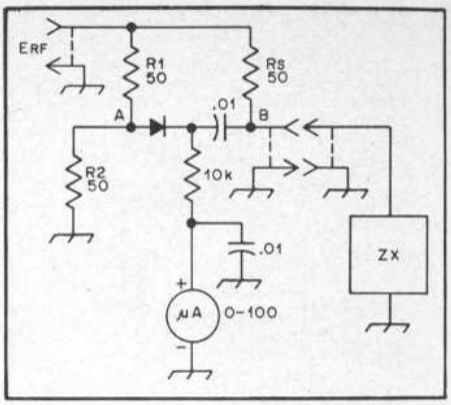


Fig. 26 – A bridge circuit for rf sine waves.

mitters) when being evaluated with low-power instrumentation. Note that this unit is not a directional coupler – it makes no difference which way the current is flowing.

Bridges for RF Measurements

A useful instrument is an rf bridge. While the classic application of such a device is for antenna and transmitter evaluation and tuning, there are a number of other applications. Most of the measurements done with bridges occur at relatively high-power levels. However, often one wants to determine the impedance of low-power active circuits. If the usual high-level bridges were used in measuring such circuits, the results would be inaccurate. In the extreme, the circuit being studied could be damaged.

Consider the Wheatstone bridge that is used for dc resistance measurements. This is shown in Fig. 24. Assume that voltage E is applied to the bridge, and that resistors R1 and R2 are equal in value. This being true, the voltage at point A will be $E/2$. The other two resistors in the bridge are R_s, a "standard," and R_x, the unknown resistance. The voltage at point B will be determined by the ratio of the two resistors. If R_s and R_x are equal, the voltage at point B will also be $E/2$. The bridge is now balanced and there is no voltage difference between point A and B. Thus, there will be no indication in the detector.

What will happen in the more typical case where R_s and R_x are not equal? Since the voltage at point B is no longer $E/2$, a potential difference exists between points A and B and a current will flow in the detector. We could calibrate the meter to tell us the level of unbalance, and thus infer the value of the unknown resistance, R_x. However, a better approach would be to make the standard, R_s, a calibrated variable resistor. It could then be varied until a null is indicated with no response in the detector. Then knowing R_s, and observing that the bridge is balanced, we know the value of R_x.

Shown in Fig. 25 is another approach. Here, we have replaced R1 and R2 with a potentiometer. R_s now has a fixed value. The control is varied until a null is again achieved. A bridge of this kind is calibrated by placing various known values in the R_x position. The dial on the control is then marked accordingly.

The foregoing examples occurred at dc, so the detector would be a meter with a capability for deflection in either direction (zero center). However, the same principles will apply if a different kind of detector is used and the input driving voltage, E, is an rf sine wave. Such a bridge is shown in Fig. 26. The resistors are all 50 ohms. However, for the bridge to operate properly, this is not necessary. The only requirement is that R1 and R2 be equal, and R_s is the same as the load the bridge is designed to measure. The typical values for R_s are 50 or 75 ohms.

The detector in the rf bridge is a diode in series with a capacitor. Assume that the unknown impedance is a 50-ohm resistor. In this case the bridge will be balanced because the rf voltages at points A and B are equal. There is no potential difference across the detector.

Consider now the case where a 100-ohm resistor is placed across the R_x port. The voltage at B will be higher than that at A. This voltage difference will appear across the detector diode and will charge the capacitor to some dc voltage. This will cause a current to flow through the 10-kΩ resistor and the meter, giving an indication. A similar result would occur if a 25-ohm resistor were placed on the unknown terminal.

Consider now the case where the unknown impedance had a magnitude of 50 ohms, but was reactive. For example, the unknown load could be a 35-ohm resistor in series with an inductor that had 35 ohms of reactance at the input frequency. The bridge *would not* be balanced. While the magnitudes of the impedances are proper to balance the bridge, the fact that the unknown termination is reactive means that the voltages at points A and B are not in phase with each other. An analysis will show that this leads to a detector output. In order for the bridge to be balanced, the unknown load must be 50 ohms and be purely resistive.

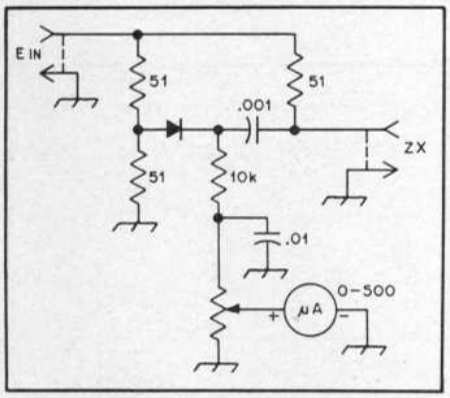
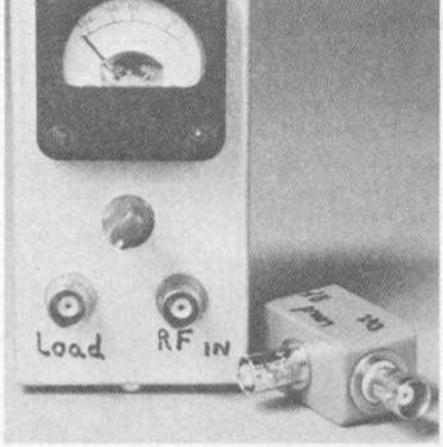
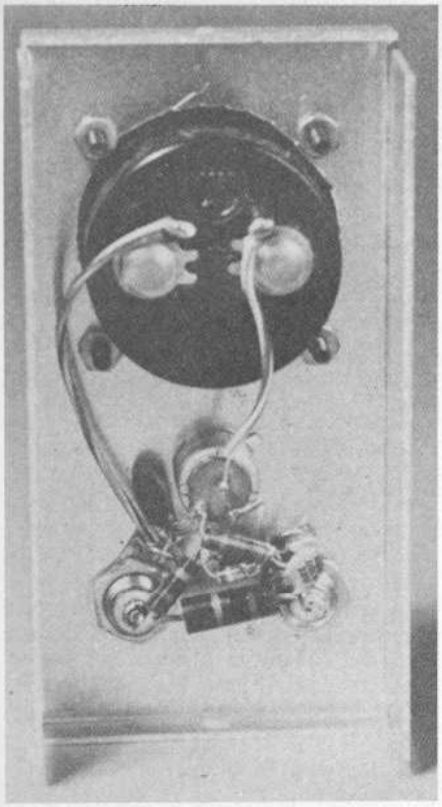


Fig. 27 – A bridge circuit which has a sensitivity control.



Exterior view of the bridge. The small unit is the return-loss bridge of Fig. 36.



Inside view of the bridge. Note short leads in the signal path.

The bridge just described is useful in spite of its simplicity. Shown in Fig. 27 is the circuit of a similar unit that has a potentiometer added as a sensitivity control. The unit is shown in a photograph. By keeping the leads short, and by using a germanium diode, the bridge is reasonably accurate through the 2-meter band. It can be driven with as little as 100 mW of rf power. The small size makes it convenient for rooftop adjustment of antennas.

Shown in Fig. 28 is a similar unit using a control for the ratio arm of the bridge. This unit is useful for experimental work since a wide variety of resistances can be measured, ranging from, say, 10 to 1000 ohms. In a bridge of this kind the exact value of the "standard" resistor is not critical, for this will merely determine the R_x value for which the control will be in the center. The bridge is calibrated by substitution of known resistances at the R_x port. The major limitation of this instrument is its upper frequency limit. This arises from the capacitance of the arm of the control to ground. The reactance will be constant (more or less), but the resistance above the arm of the control will vary, leading to a variable phase for the reference voltage of the bridge.

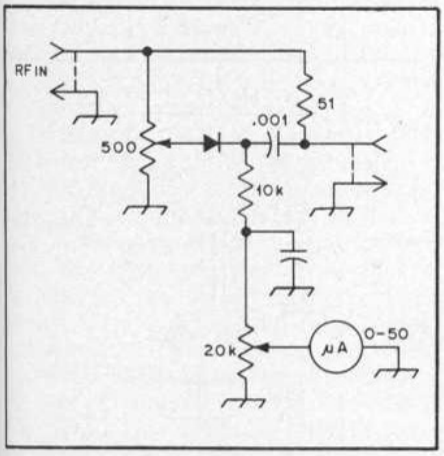


Fig. 28 — Here a control is used as the ratio arm of a bridge.

The problem of errors from stray capacitances can be circumvented by replacing the variable resistance arm with a variable capacitance voltage divider (Fig. 29). It may be shown that such a divider produces a voltage that is in phase with the driving signal. Sevick, W2FMI, has described several bridges of this kind (see the bibliography). The advantage is that stray capacitances are absorbed in the variable element and do not lead to frequency-dependent errors.

All of the bridges described have the capability of measuring only resistances. If a reactive termination is present, a complete null cannot be obtained. However, reactive impedances may be measured by using an outboard adaptor as shown in Fig. 30. This unit is a series-tuned circuit. The inductor is chosen so the bridge will see a null when a resistive termination is placed on the output and the variable capacitor is at midrange. In practice, the capacitor and the resistance-measuring arm in the basic bridge are adjusted repeatedly until a complete null is obtained. The position of the variable capacitor in the reactance-canceling arm will tell the user if the termination is inductive or capacitive. The system may be calibrated if desired.

Bridges for Antenna Tuners

Consider now a bridge that might be used to tune a Transmatch. Such a unit is shown in Fig. 31. This bridge differs slightly from the others we have considered: A resistor has been added at the input, and the values of the resistors in the divider arm have been reduced from 50 to 15 ohms. These changes are significant. Consider the impedance extremes that can appear at the output termination. One is a short circuit, while the other is an open circuit. For these two extremes, the resistance seen at the input of the bridge will vary only from 46 to 57 ohms. Both values are close to 50 ohms. As a result, the transmitter will always see something close to a proper termination. This can be a profound advantage if the transmitter being used to drive the bridge is prone to

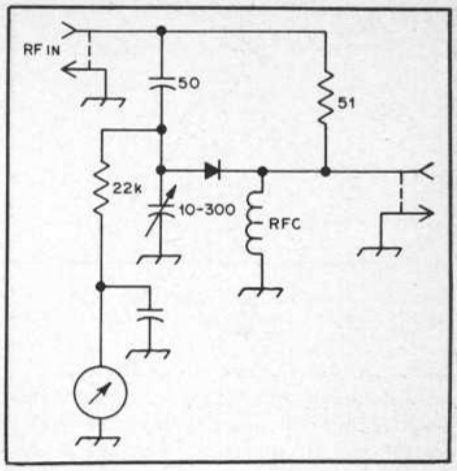


Fig. 29 — A variable-capacitance voltage divider is used in this circuit to replace a resistive divider.

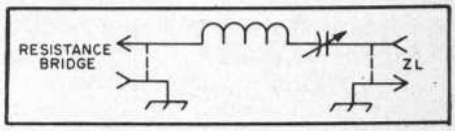


Fig. 30 — An outboard adapter for use in measuring reactive impedances.

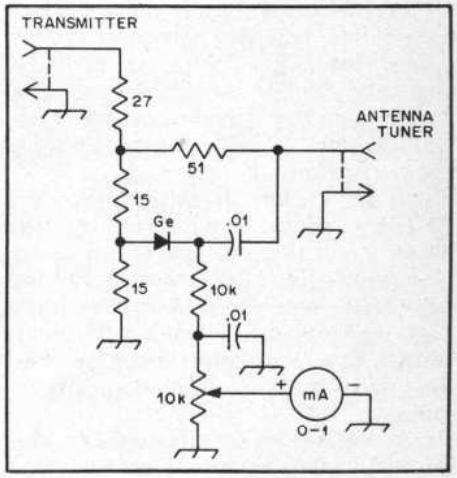


Fig. 31 — A bridge circuit suitable for use when adjusting a Transmatch.

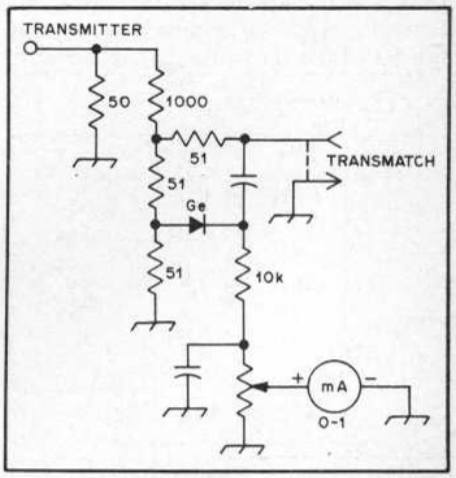


Fig. 32 — A high-power adaptation of the circuit shown in Fig. 31.

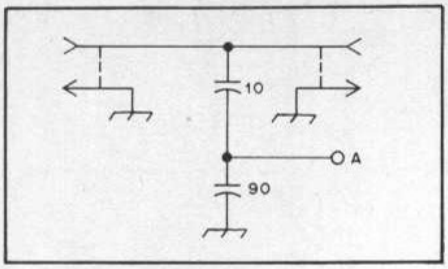


Fig. 33 — A capacitive voltage divider in parallel with a transmission line.

self-destruction when a mismatch occurs. Severe mismatches *can* occur during the tuning of a Transmatch. An additional advantage of the bridge shown is that, when matched, the output applied to the antenna is down 12.8 dB from the full transmitter output that is applied to the input. Use of bridges of this type would help eliminate carriers during tune-up periods.

This absorptive-bridge technique is by no means limited to low power applications even though the unit of Fig. 31 can be driven with less than a watt. Shown in Fig. 32 is a high power adaptation of this method. One of the writers has used this technique when tuning the station Transmatch for several years. It's comforting to know that only 50 mW of rf is reaching the antenna during tune-up periods even though 25 or 30 watts is available from the transmitter.

In many cases, a bridge of the kind described above is not sufficient. Instead, a unit that operates at full power is desired. Such units are useful for monitoring antenna VSWR on a continuous basis, or for measuring the input VSWR of a high-power amplifier. The latter could vary as a function of drive power.

In the section on attenuators earlier in this chapter, a ferrite transformer was used as a 20-dB coupler. In this application, the voltage appearing across the coil secondary was proportional to the current flowing in the line. Consider now the effect of a capacitive voltage divider across the transmission line (Fig. 33). The voltage at point A will be in phase with the voltage on the line. However, the magnitude of the voltage will be one-tenth the value on the line.

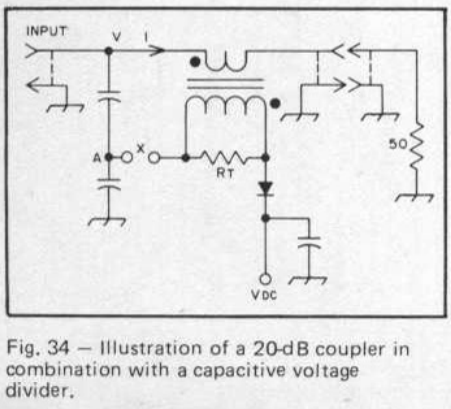


Fig. 34 — Illustration of a 20-dB coupler in combination with a capacitive voltage divider.

Consider the result of combining the two effects. This is shown in Fig. 34. The voltage appearing across the terminating resistor, R_t, is proportional to the current flowing in the transmission line. The voltage appearing from the capacitive divider is proportional to the voltage on the line. The ratio of these two quantities, $E \div I$, is indicative of an impedance. Assume that the capacitors are adjusted such that the voltage from A is the same magnitude as the voltage across R_t. Then, when the connection is made at point X in the circuit, the two voltages will add in phase. The resultant will be detected by the diode, producing a dc output.

Consider now the effect of reversing the in-line bridge. That is, the port that was terminated with the 50-ohm load is now driven by the transmitter, and the original input is terminated in 50 ohms. The voltage at point A will be virtually the same. However, the current is now flowing in the opposite direction from the earlier case. Because of this, the voltage appearing across R_t will be out of phase by 180 degrees from the original case. The two rf voltages will now cancel each other. No detected output will occur. Units of this type are appropriately called directional bridges.

In the typical unit, a double secondary is used on the transformer in order to allow both forward and reverse powers to be monitored simultaneously. Some examples are seen in Figs. 15 and 17.

The Return-Loss Bridge

Let us return now to a simple resistive bridge. Shown in Fig. 35 is a bridge that departs slightly from those described earlier. First, it is driven from a 50-ohm source. This was not necessarily the case when a transmitter was used. The output impedance of a transmitter could look like something very much different than 50 ohms, even though it may have been designed to be terminated in a 50-ohm load. The second difference is that a 50-ohm resistor is connected between points A and B. Clearly, if R_x is 50 ohms, the bridge is balanced and there is no voltage difference between points A and B. There will be no power dissipated in the detector resistance, R_d.

Assume that the unknown port is now either open or short circuited. It may be shown that in either of these cases an identical voltage difference will appear across R_d. If the bridge is not driven from a 50-ohm source, the voltage across R_d will not be the same when one goes from a short to an open circuit.

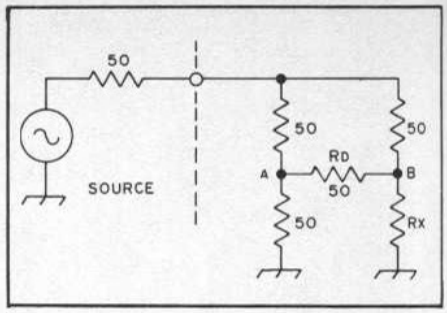


Fig. 35 — Another version of a simple resistive bridge.

The circuit has a drawback. Most 50-ohm detectors (like those described earlier in this chapter) are single-ended. This deficiency may be solved with the circuit of Fig. 36, where a "sortabalun" has been inserted from the floating detector port to a single-ended port. This allows the voltage difference between points A and B to appear across a single-ended output. Also, the impedance presented to the single-ended detector port is now impressed between points A and B. The transformer has approximately 10 bifilar turns of No. 30 enameled wire on an FT-23-43 ferrite toroid. Ferrite should be used instead of powdered iron.

When using the bridge, the unknown port is either short or open circuited, and the power in the detector is noted. Then, the unknown termination is attached to the unknown Z port and the detector power is again noted. The ratio, expressed in dB, is known as the return loss. The higher the return loss, the closer the unknown termination is to 50 ohms. It may be shown that the return loss (R-L) is related to the magnitude of the reflection coefficient r, by $R\text{-}L = 20 \log_{10} r$. The reflection coefficient is related to the voltage standing wave ratio by $r = (\text{VSWR} - 1) \div (\text{VSWR} + 1)$. Table 2 compares return loss, reflection coefficient and VSWR for a wide range of values. If phase angle is to be included, a more complete representation would be $r =$

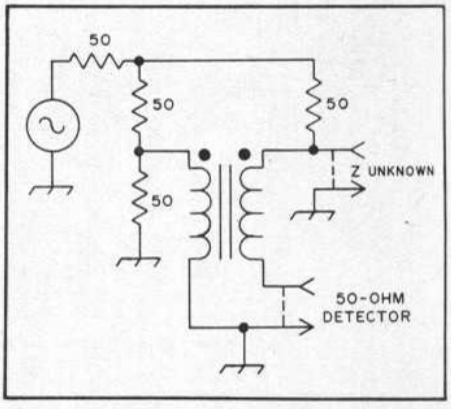


Fig. 36 — A return-loss bridge for impedance measurements. See text.

Table 2

$$YSWR = \frac{1 + r}{1 - r}$$

RETURN LOSS, dB	r, REFLECTION COEF.	VSWR
1	0.891	17.4
2	0.794	8.72
3	0.707	5.85
4	0.631	4.42
5	0.562	3.57
6	0.501	3.01
7	0.447	2.61
8	0.398	2.32
9	0.355	2.10
10	0.316	1.92
12	0.251	1.67
14	0.199	1.50
16	0.158	1.38
18	0.126	1.29
20	0.100	1.22
25	0.056	1.12
30	0.032	1.07
35	0.018	1.04
40	0.01	1.02
45	5.6×10^{-3}	1.011
50	3.16×10^{-3}	1.006
60	1.0×10^{-3}	1.002

$(Z - Z_o) \div (Z + Z_o)$. All of these parameters are of significance when using a Smith chart for impedance representations.

One major advantage of a return-loss bridge is that the measurement of impedance can be done at low-power levels. For example, a low-level signal generator could be used as the rf source, and one of the sensitive rf detector systems described earlier could be used as the detector. In fact, a receiver could be used in conjunction with a step attenuator as the detector. The simple detectors described will provide only information about the magnitude of the reflection coefficient. To measure the angle, a vector voltmeter would be needed.

Another application of the return-loss bridge would be as a simple 6-dB hybrid combiner. A typical application would be to combine the outputs of two signal generators for the purpose of measuring intermodulation distortion and gain compression in, for example, a receiver. One generator is applied to the source port while the other is connected to the detector port. Shown in Fig. 37 is such an application. Assuming that each generator is set to deliver 10 mV to a 50-ohm load, the resulting voltages are shown. Note that generator A delivers 5 mV to the output load, hence the 6-dB loss. However, note that 5 mV appears at both of the detector points in the bridge as a result of drive from generator A. There is no voltage difference, and none of the signal from generator A appears at generator B. The converse is also true. This is needed in IMD measurements. If one generator is allowed to "talk to the other," the result may be that one generator will phase modulate the other. This modulation leads to sidebands at the same frequencies where IMD products appear and can cause errors in the IMD measurements.

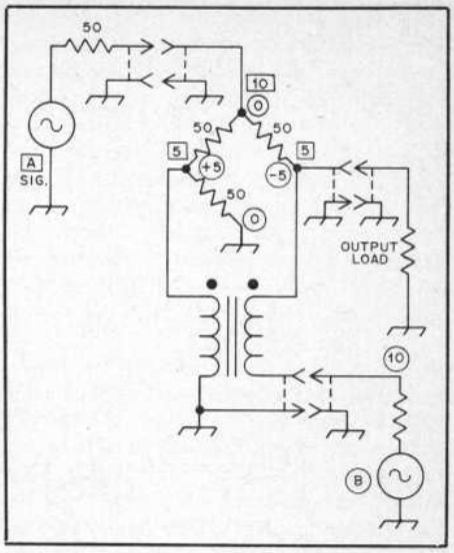


Fig. 37 — A 6-dB hybrid combiner can be used to connect two signal generators to a test circuit for measuring, as one example, receiver dynamic range.

Solid-State Power Supplies

Nearly all of the equipment in this book requires an external dc power source. Although some battery-powered gear is described for field use, the subject of batteries shall not be treated here. Rather, we will focus attention on power supplies and voltage regulators which operate from the ac power line. Some rules of thumb are offered for those who wish to design and build their own power supplies and regulators. A more concise treatment of the general subject can be found in *The Radio Amateur's Handbook*, and in the references given in the bibliography section.

A Basic Power Supply

Fig. 38 shows a typical unregulated dc power supply. A quad of silicon rectifier diodes is used in a full-wave hookup. Since full-wave bridge rectification is the most efficient of the common types, we shall deal with that circuit in this chapter.

An advantage of a bridge rectifier is that it delivers full-wave output without the need for a transformer with a secondary center tap. Another feature of the full-wave rectifier is that the ripple frequency at the output is twice the line frequency, thereby making filtering less difficult. Thus, the capacitance of the filter capacitor for a specified percentage of output ripple will be considerably lower than with a half-wave rectifier.

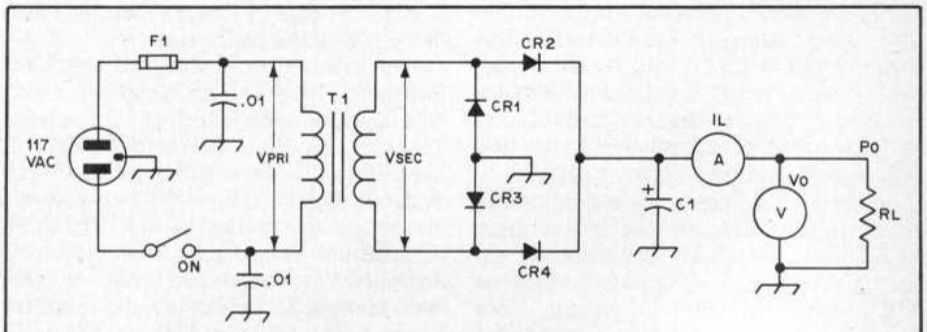
A Design Example

Let's assume we need a simple power supply that is able to provide a voltage output of 13. Maximum current taken by the external load will be 500 mA (0.5 A). Maximum ripple will be 3 percent, and the load regulation will be 5 percent.

The rms secondary voltage for T1 of Fig. 38 must be the desired V_o plus the voltage drops across CR2 and CR4 ($\cong$ 1.4V) divided by 1.41. Thus, T1 V_{sec} = 13 + 1.4/1.41 = 10.2 volts. The nearest standard transformer would be a 10-volt one, which would be close enough in value. Alternatively, the builder could wind his own transformer, or remove secondary turns from a 12-volt transformer to obtain the desired rms secondary voltage.

A 3-percent ripple referenced to 13 volts is 0.39 V rms. Therefore, the pk-pk value is found from: V_{rip} = 0.39 $\times$ 2.82 = 1.09 V. This figure is necessary to calculate the required capacitance for C1.

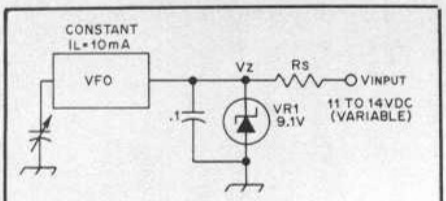
Also needed for determining the value of C1 is the time interval (t) between the full-wave rectifier pulses,



V_o (no load) $\cong V_{sec} \times 1.41$
$P_o = V_o \times I_L$
$R_L = V_o \div I_L$

C1 (V_{min}) = $Vsec \times 1.41$
F1 (A) = 2I/N (N = turns ratio)
$V_{sec} \cong V_o \div 1.41$

Fig. 38 — A circuit which illustrates the configuration of a basic unregulated dc power supply.



$$R_s = \frac{V_{in(min)} - V_Z}{I_L + 0.1\ I_L}$$

$$R_s = \frac{11 - 9.1}{.01 + .001} = \frac{1.9}{.011} = 173 \text{ ohms} \quad \text{(A)}$$

$$P_{D(VR1)} = \left[\frac{V_{in(max)} - V_Z}{R_s} - I_L\right] V_Z$$

$$P_{D(VR1)} = \left[\frac{14 - 9.1}{173} - .01\right] 9.1$$

$$= .028 - .01 \times 9.1 = 0.167\text{W} \quad \text{(B)}$$

$$P_{D(RS)} = (V_{in(max)} - V_Z)^2 \div R_{s(ohms)}$$

$$= (14 - 9.1)^2 \div 173 = 0.138\text{W} \quad \text{(C)}$$

Fig. 39 — Zener diodes are effective as simple voltage regulators.

which for that circuit is 8.3 milliseconds (ms). Therefore, C1 is calculated from

$$\text{C1}(\mu\text{F}) = \frac{I_L t}{V_{rip}}$$

$$= \left[\frac{0.5\text{A} \times 8.3 \times 10^{-3}}{1.09}\right] 10^6$$

$$= .0038 \times 10^6 = 3800\ \mu\text{F} \quad \text{(Eq. 1)}$$

where I_L = the current taken by the circuit which is powered by the supply V_o. The nearest standard capacitor value is 4000 μF. It will be an acceptable one to use, but since the tolerance of electrolytic capacitors is rather loose, a 5000-μF unit will probably assure that the design requirements are met.

Diodes CR1-CR4, inclusive, should have a PRV rating of at least two times V_{sec} peak, which means with our example we have 14.4 volts. Therefore the PRV should be 28.8 or greater. Four 50-V diodes will work nicely. Similarly, the forward current of the diodes (I_{fwd}) should be at least twice the load current, I_L. So for a 500-mA I_L the diodes should be rated at 1 A or greater.

The load resistance, R_L, is determined by V_o/I_L, which in this example is 13/0.5 = 26 ohms. This factor must be known in order to find the necessary series resistance for the target 5-percent regulation. $R_{s(max)}$ = Load reg. × $R_L/10$ = .05[26/10] = 0.13 ohm. Therefore, the transformer secondary dc resistance should be no greater than 0.13 ohm. The secondary current rating should be equal to or greater than the I_L of 0.5 A. A transformer of that type will usually have a secondary resistance of less than our maximum acceptable amount for a 5-percent regulation trait.

Information on calculating the value of the fuse, F1, is given in Fig. 38. C1 should have a minimum working voltage of 18.33 in accordance with the formula in Fig. 38. The next standard value is suggested — a 25-volt capacitor.

Regulated Voltages

When the need arises to regulate small amounts of current, say, up to 100 mA, Zener diodes offer a low-cost approach. Even though higher amounts of current are handled sometimes by Zener diodes, the practice is not a common one in amateur work. Our treatment will be confined to the lower current amounts.

Most Zener diodes are known also as *avalanche* diodes. They are similar in construction to junction rectifiers, but the primary characteristic for their intended purpose is the *reverse-breakdown* profile. In simple terms, positive voltage is applied to the cathode of the diode rather than to the anode. As this reverse voltage is made higher the leakage current in the diode stays fairly constant until a critical plateau is reached. This point is known as the *breakdown voltage*. There is a marked contrast between the end result of the breakdown point of a Zener diode and a conventional rectifier diode. With the latter it is essential to operate the diode well below the breakdown or PRV (peak reverse voltage) to avoid damaging it. When the breakdown point of a diode is reached, copious amounts of current flow through the junction, and in the case of Zener diodes this area is known as the *Zener current*.

At breakdown, the normal high back resistance of the diode drops to a very low amount and, therefore, the current increases rapidly. The amount of current is, however, limited by the series resistance (R_s of Fig. 39) between the diode and the voltage source. The rated breakdown value of a Zener diode is that level for which the semiconductor was designed. Typically, the plateaus range from 3.9 to as high as 200 volts. The amount of safe sustained Zener current is determined by the wattage rating of the component. These values run from 150 mW to 50 watts at present.

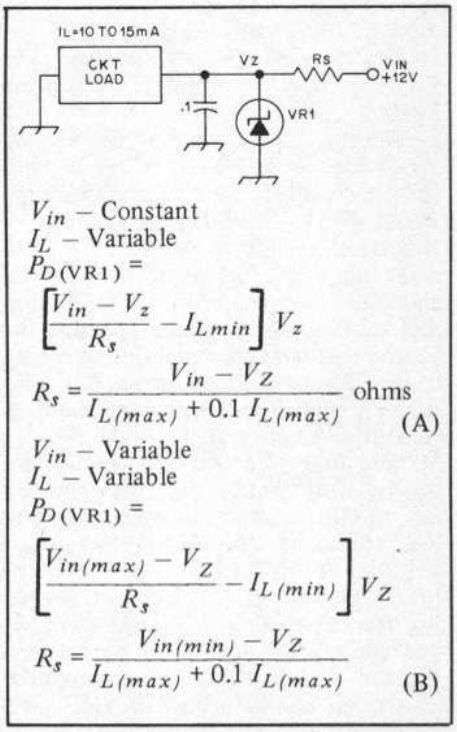
Because of the characteristics we have just described it can be seen that a Zener diode will serve nicely as a voltage regulator, sine-wave clipper, or as a series-gate element. Voltage regulation is made possible by virtue of the high current which flows at conduction. The regulator current must always be considerably higher than that which is drawn by the I_L (circuit to which the regulated voltage is applied). Under that rule the *significant current* which flows through the series dropping resistor is that of the diode: Small changes in input voltage or circuit load current are disguised by the diode current and R_s by means of the $E = I \times R$ rule.

Designing with Zener Diodes

There are three sets of conditions common to regulator circuits: *variable* load current and *constant* supply voltage, *constant* load current and *variable* supply voltage, and *variable* load current and *variable* supply voltage. A slightly different equation applies in each case, Figs. 39 and 40.

A rule of thumb can be used with respect to the ratio of minimum Zener-diode current (I_{Zmin}) and the load current (I_L). For best regulation the ratio should be 10:1. That is, the load current should be roughly 10 percent of the Zener diode current.

Fig. 39 shows a shunt type of Zener-diode regulator. It provides 9.1 volts regulated to a VFO which has a constant load current of 10 mA (.01 A). The 10:1 current ratio does not result from the values given, but the figure is close enough for most amateur work. Had a lower value of V_z been chosen,



V_{in} — Constant
I_L — Variable

$$P_{D(VR1)} = \left[\frac{V_{in} - V_z}{R_s} - I_{Lmin}\right] V_z$$

$$R_s = \frac{V_{in} - V_Z}{I_{L(max)} + 0.1\ I_{L(max)}} \text{ ohms} \quad \text{(A)}$$

V_{in} — Variable
I_L — Variable

$$P_{D(VR1)} = \left[\frac{V_{in(max)} - V_Z}{R_s} - I_{L(min)}\right] V_Z$$

$$R_s = \frac{V_{in(min)} - V_Z}{I_{L(max)} + 0.1\ I_{L(max)}} \quad \text{(B)}$$

Fig. 40 — Zener-diode application for circuits which have changes in load current.

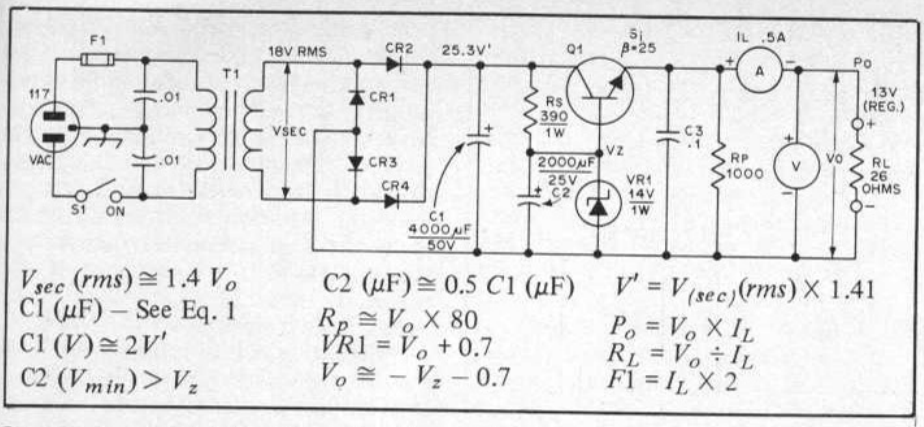


Fig. 41 — Illustration of a power supply with regulation. A pass transistor, Q1, is used to extend the range of the Zener-diode regulator.

the ratio would have been closer to the suggested one.

In the equation of Fig 39A a series resistance of 173 ohms is called for. The nearest standard value is 180 ohms. That will be entirely suitable for R_s.

The equation at Fig. 39B determines that the maximum Zener-diode power dissipation is 0.167 W. A good rule of thumb for choosing a wattage rating for the diode is a times-5 factor. This will allow ample safety margin for diode internal heating. Since we determined that VR1 will dissipate 0.167 W, a 5-times value will be 0.8 W. The nearest standard power value is 1 W, so a diode of that type will suffice.

Fig. 39C gives an equation for computing the wattage dissipated in R_s at $V_{in(max)}$, which is 0.138 W. To stay on the safe side of things we will again use the 5-times rule. This gives us a wattage rating for R_s of 0.69. In practice, a 1/2-watt resistor will suffice – that being the nearest standard value.

When high-wattage Zener diodes must be used (10- to 50-W types, in general), they will be of the stud-mount variety. Heat sinking is done in the same manner as with power transistors and power-type rectifier diodes. The general rules for this have been given earlier in the book. A more complete discussion of Zener-diode applications was given in *QST* for April, 1976.

Extending Zener-Diode Range

The foregoing section outlines some of the limitations when using Zener diodes as regulators. Greater current amounts can be accommodated if the Zener diode is used as a *reference* at low current, permitting the bulk of the I_L to flow through a series *pass transistor* (Q1 of Fig. 41). An additional benefit in using a pass transistor is that of reduced V_o ripple. This technique is sometimes referred to as "electronic filtering."

Q1 of Fig. 41 can be thought of as a simple emitter-follower dc amplifier. It increases the load resistance seen by the Zener diode by a factor of beta (β). In this situation VR1 is required to supply only the base current of Q1. The net result is that the load regulation and ripple characteristics are improved by a factor of beta. Addition of C2 reduces the ripple even more, although many simple regulated power supplies of the type seen in Fig. 41 do not have C2 as a part of the circuit.

The primary limitation of this type of circuit is that Q1 can be destroyed almost immediately if a severe overload occurs at R_L. The fuse, F1, cannot blow fast enough to protect Q1. Furthermore, if a low-current fuse was used at V_o it would be subject to the same limitations. In order to assure longevity of Q1 it is necessary to include a current-limiting circuit of the kind shown in Fig. 42. Modern three-terminal regulators have replaced the circuit of Fig. 42, and that subject will be discussed later in the chapter.

It should be mentioned that the greater the value of V_{sec} at T1, the higher the power dissipation in Q1. This not only reduces the overall efficiency of the power supply, but requires stringent heat sinking at Q1. The circuit of Fig. 41 could be made to operate with a V_{sec} as great as 25 volts, but a more suitable voltage level for a 13-volt output at V_o would be 18 volts rms. In this regard it is not difficult to remove the required number of secondary turns from a 24-volt transformer.

A Design Example

We desire a regulated, well-filtered dc voltage of 13. I_L maximum shall be 0.5 A. The circuit of Fig. 41 will be the one used in this example. The ratings for T1, CR1-CR4, and C1 can be determined by using the formulas given for the circuit of Fig. 38. V_{sec} shall be 18 V rms.

In order to calculate the value of R_s in Fig. 41 we must learn what I_b (base current) for Q1 will be. The base current is approximately equal to the emitter current of Q1 in amperes divided by beta: I_b = 0.5/25 = .02A, or 20 mA. The transistor beta can be found in the manufacturer's data sheet, or measured with simple test equipment (beta = I_c/I_b). Since the beta spread for a particular type of transistor – 2N3055 for example, where it is specified as 20 to 70 – is a fairly unknown quantity, more precise calculations for Fig. 41 will result if the transistor beta is tested before the calculations are done. A suitable, conservative approach is to design for *beta minimum* of the transistor used.

As we learned earlier, in order for VR1 to regulate properly it is necessary that a fair portion of the current flowing through R_s should be drawn by VR1. Therefore, let us set a rough rule of 30 mA for I_{Rs}. Knowing this figure, plus the I_b of .02A just computed, the Zener-diode current (I_z) will be .03 A – .02 A = .01 A, or 10 mA. From this we can learn that: R_s(ohms) = $(V' - V_z)/I_{Rs}$ = (25.3 – 14)/.03 = 376 ohms. The nearest standard ohmic value for R_s is 390, so it shall be used. The wattage ratings of R_s and VR1 can be obtained from the formulas given earlier for Zener-diode regulators.

A safe power rating must be provided for Q1. In this context it should be known that the dissipation in Q1 will be equal to the emitter current times the collector-to-emitter voltage. Thus, for our circuit of Fig. 41 $P_{Q1} = I_E \times V_{CE}$, where V_{CE} equals the desired $V' - (V_z - V_{BE})$. Therefore, P_{Q1} = 0.5 A × 12 V = 6 watts. V_{BE} for a silicon transistor is approximately 0.7 V. A good rule of thumb in this example is to choose a transistor at Q1 which has a $P_{D(max)}$ of at least twice P_{Q1}. Therefore, Q1 should be rated at 12 watts or more. Since the cost of power transistors is quite low, a 25-, 50-, or 100-watt unit will allow considerable safety factor if heat-sinked properly, and would represent a good choice.

Load regulation with the power supply of Fig. 41 will be approximately 2 percent, and the output ripple will be low. Line regulation will be on the order of 7 percent, assuming the 117-V line has variations.

The .01-μF capacitors at the primary of T1 serve two functions. They act as transient suppressors and help prevent rf energy from entering the power-supply regulator. C3 serves in a similar manner. R_p is used as a minimum-load resistance for periods when the power supply has no external load.

Current Limiting

Damage to Q1 of Fig. 41 can occur when the I_L exceeds the safe amount, or when V' becomes excessive. Fig. 42 illustrates a simple current-limiter circuit which will protect Q1. All of the I_L passes through R2. Therefore a voltage difference will exist across R2, the

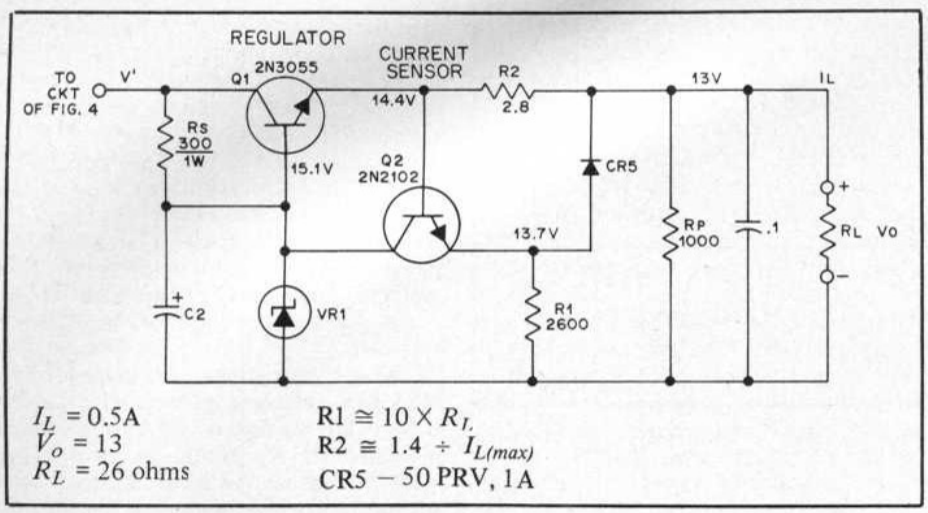


Fig. 42 — Overload protection for a regulated dc supply can be effected by addition of a current-overload protective circuit contrasted to that of Fig. 43.

precise amount being dependent upon the exact I_L value at a given time. When the load current exceeds a predetermined safe value, the voltage drop across R2 forward biases Q2 and causes it to conduct. Since CR5 is a silicon diode, and because Q2 is a silicon transistor, the combined voltage drops through them (roughly 0.7 V each) will be 1.4 V. Therefore, the drop across R2 must exceed 1.4 V before Q2 can turn on. This being the case, R2 is chosen for a value that provides a drop of 1.4 V when $I_{L(max)}$ occurs. In this instance 1.4 volts will be seen when I_L reaches 0.5 A.

When Q2 turns on, some of the current through R_s flows through Q2, thereby depriving Q1 of some of its base current. This action, depending upon the amount of Q1 base current at a precise moment, cuts off Q1 conduction to some degree, thus limiting the flow of current through it.

Specifications

Addition of the current limiter will cause a loss of roughly 1.4 volts over that obtained from the circuit of Fig. 41, owing to the inclusion of R2. Therefore, if a V_o of 13 is desired, the output from Q1 should be 14.4 V.

Q2 can be a medium-beta, low-power device. It must be able to sustain the full V_o. In this example a 25-V V_{CE} will be ample, and a P_D of 1 W will be suitable for Q2. A 2N2102 would be a good choice.

R1 will be approximately 100 times the R_L value. Since R_L in this example is 26 ohms, $V_o/I_{L(max)}$, R1 will be 2,600 ohms. The value of R1 can be trimmed to provide Q1 cutoff when I_L exceeds the safe amount.

R2 is chosen from R2 = 1.4 V/0.5 A = 2.8 ohms. The closest standard resistor value is 3 ohms, which should be acceptable. R2 must handle $I_{L(max)}$ without overheating. Therefore, its dissipation will be $0.5^2 \times 3 = 0.75$ W. A 2-watt resistor should allow sufficient safety margin. Magnet wire of small cross-sectional area can be used to wind R2. This practice will enable the builder to obtain the precise ohmic value needed.

Refinements in Discrete Regulators

In the example of Fig. 42, suitable performance was obtained for the case where a constant load current was to be supplied. The ripple of the power supply was fairly low and the output voltage was reasonably stable. However, there are some inexpensive refinements that may be applied to simple regulators which will improve performance significantly.

The first thing that can be done to improve regulation is to decrease the resistance value of R1 (Fig. 42). In the circuit shown the design was tailored such that a 1.4 volt drop would occur across R2 when the output current was 0.5 A. However, if the load was removed from the output, the voltage would go from the desired output level of 13 volts up to 14.4 volts. Q2 is not turned on until the power supply goes into current limiting.

The diode in the regulator circuit (CR5) provides a well-defined current where limiting occurs. However, if the desire is mainly to protect the power supply from self-destruction, this diode may be eliminated, as may R1. The result is shown in Fig. 43. This circuit has better load regulation. At full current (0.5 A) the output voltage is 13. When the load is removed, the voltage goes up to 13.7. Note that it was necessary to decrease the value of the Zener diode from 15.1 to 14.4 volts.

While this is a 2:1 improvement in regulation over that originally obtained, it is still less than desired for many situations. Another problem is that the exact value of the Zener diode has a direct bearing on the output voltage obtained. The typical voltage tolerance of inexpensive Zener diodes is ±5 percent. A 5 percent variation in the 14.4-volt diode required could allow the output to range from 13.7 down to 12.3 volts. A more desirable situation would be a power supply that used a lower voltage Zener diode and an additional transistor. The exact output voltage could then be set with a variable resistor. Such a power supply regulator is shown in Fig. 44. We will assume that the Zener diode chosen has a rating of 6.2 volts.

When power is initially applied to this circuit, the series pass transistor is turned on with the 300-ohm bias resistor. This causes the voltage at the output to increase in value. The output voltage is attenuated by resistors R1 and R2, and causes a voltage to appear at the base of Q2. This turns transistor Q2 on, and charges capacitor C1. C1 will charge until it reaches the Zener-diode voltage of VR1. The Zener diode then clamps the voltage at the emitter of Q2 at 6.2 volts. The base voltage on Q2 will be 0.7 volt greater, or 6.9 volts.

What will the output voltage be? Assume that the two resistors are equal in value and that their ohmic value is

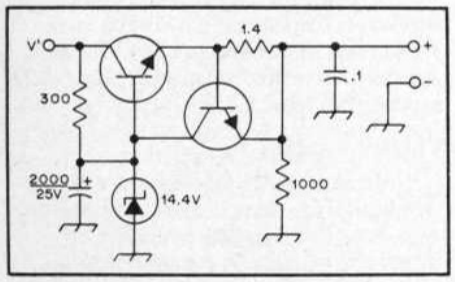


Fig. 43 – An improved current-overload protective circuit contrasted to that of Fig. 42.

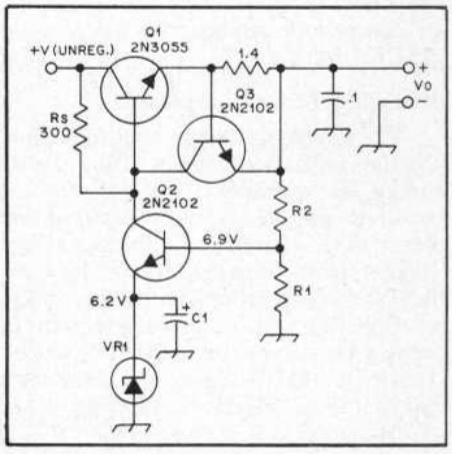


Fig. 44 – This regulator circuit is more precise than that of Fig. 43, permitting the builder to obtain a specific output voltage.

reasonably low. The low value ensures that the current flowing in the resistors is large in comparison with the base current in Q2. Since the resistors are of equal value, the voltage at the junction of the two resistors must equal 0.5 the output voltage. But, because of the Zener diode and the e-b drop of Q2, the base voltage at Q2 must be 6.9. Hence, the output must be equal to twice this value, or 13.8 volts.

The foregoing analysis was carried out for no external load on the power supply. What happens if a resistive load is now placed on the supply? This would tend to drop the output voltage. However, when this begins to occur, the voltage on the base of Q2 will decrease. As this happens, the collector current in Q2 will decrease also. This will cause a reduced voltage drop across R_s. This means that the voltage on the base of Q1 will increase, causing the output voltage to again increase until it reaches 13.8. The voltage drop across the 1.4-ohm current-limit sensing resistor has no effect upon the output voltage.

This voltage regulator utilizes an amplifier in a negative feedback loop. The fact that the output voltage was not affected by the drop across the 1.4-ohm sensing resistor was the result of the feedback signal being obtained after the current limiting circuitry. The limiting circuit (Q3) was within the feedback loop.

If the desired output was not 13.8 volts, but 13 volts as before, it could be obtained by changing the ratio of R1 to R2: If R1 were 470 ohms, the output voltage would be 13.0 volts when R2 was 415.5 ohms. The best way to design this power supply would be to make R2 a 500-ohm variable resistor. Then the output could be adjusted from 6.9 to 14.2 volts.

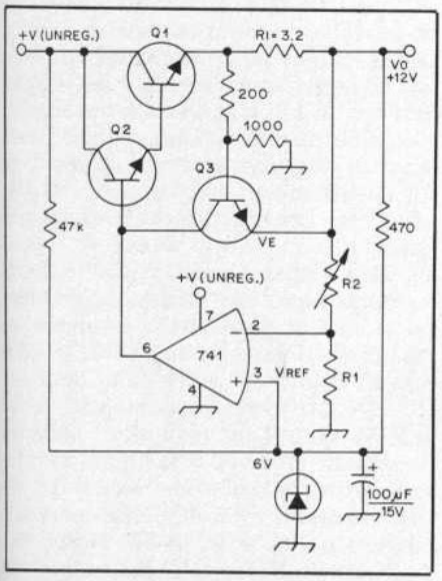


Fig. 45 — Example of some refined techniques for use in a regulated power supply.

The current limiting still functions. If the output current becomes high enough that 0.7 volt is developed across the sensing resistor, Q3 will turn on. This will then rob base current from Q1, the pass transistor. The output voltage will then decrease accordingly, with no more than 0.5 A flowing in the external load. When the power supply is short circuited (crowbarred), the current will remain at 0.5 A.

Shown in Fig. 45 is a regulator that demonstrates some refined techniques that might be used in a regulated power supply. In looking back at the regulator of Fig. 44, we see that Q2 functioned as an inverting amplifier. It can be shown that the regulation of the circuit is directly dependent upon the gain in this amplifier. In the supply of Fig. 45, we have replaced the single, discrete transistor amplifier with a high-gain operational amplifier. A 741 will function well in circuits of this kind. However, a 741 has a maximum output current of around 10 mA. This would not have been enough to drive the base of Q1 directly if high output currents were desired. Hence, another transistor, Q2, is added to form a Darlington-connected pass transistor. The effective beta of such a configuration is approximately the square of the beta of a single transistor. It is reasonable to assume an effective beta for the combination of 500 to 1000. Because of this high beta value, the op amp needs to deliver only a few mA of current to the base of Q2 for an emitter current in Q1 of 1 ampere.

The current limiting is different in this circuit than it was in Fig. 44. Note that the emitter of Q3 is tied to the output directly. However, the base of Q3 is biased from a voltage divider from the current sensing resistor. This divider has a ratio of 5/6. That is, the voltage on the base of Q3 is $(5/6)V_{e1}$, where V_{e1} is the voltage on the emitter of Q1. Let's assume that the regulator is to go into current limiting when the load current reaches 1 A. With the emitter of Q3 at the output voltage of 12, the base voltage must be equal to 12.7 at this instant. Due to the divider action, the voltage on the emitter of Q1, the pass transistor must be (6/5)12.7 = 15.2 volt: We choose a sensing resistor of 3.2 ohms.

This circuit has tremendous implications when we consider the behavior of the supply under a crowbar condition. With the output shorted, the emitter of Q3 is at 0 volts, and the base will be at 0.7 volt. Following the earlier analysis, the emitter of the pass transistor will be at 1.2 times this level, or 0.84 volt. The current in the supply is then 0.84/3.2 ohms = 0.26 amperes. This is much less than the current that the supply will deliver prior to going into limiting. This technique is called fold-back current limiting. The advantage is that the supply components need not be capable of handling such high currents during short-circuit conditions.

The price to be paid for this extreme protection is that the unregulated voltage must be higher. This is because there will be higher voltage drop across the sampling resistor, R1 prior to the point where limiting occurs.

Another feature of the regulator of Fig. 45 is the nature of the reference diode biasing. The reference is a 6-volt Zener diode which is biased to a current of about 13 mA. The diode establishes the bias on the noninverting input of the error amplifier. The output voltage is established by adjusting R2. The asset of biasing the Zener diode as shown is that virtually all of the current in the Zener comes from the *regulated* output. In earlier supplies, such as that shown in Fig. 43, the Zener diode is biased from the unregulated supply which has high ripple. Measures of this kind will help immensely in removing the last traces of hum from a power supply output.

If a builder is constructing power supplies using the techniques outlined in Fig. 45, care must be exercised to ensure that device specifications are not exceeded. Specifically, the maximum supply voltage rating of a 741 op-amp is 30 volts between pins 7 and 4. Since pin 7 is connected to the unregulated supply, this value should not exceed 30 volts.

Sometimes it is desirable to build variable voltage supplies that will go all the way down to 0 volts. This can be done with a modification of the circuit of Fig. 45. A negative power supply is first built and is well regulated. A typical value might be −6 volts. This supply is used to provide operating voltage for pin 4 of the 741. Pin 3 of the 741 is grounded directly. The end of R1, which is presently grounded, is returned to the negative supply.

Three-Terminal Regulators

Power supply design has been simplified in recent years by the appearance of the three-terminal regulator IC. These units contain all of the essential components for voltage regulation and current limiting. These include a high-gain error amplifier, sensing resistors and transistors for current limiting, a temperature-compensated voltage reference, and suitable pass transistors. These ICs are available in a number of fixed-voltage ratings from 5 to 24. They may be obtained for load currents up to 3 amperes, and come in various package styles.

These ICs have a number of advantages. The main one is the simplicity of application. The three terminals are for a ground reference, an input for the

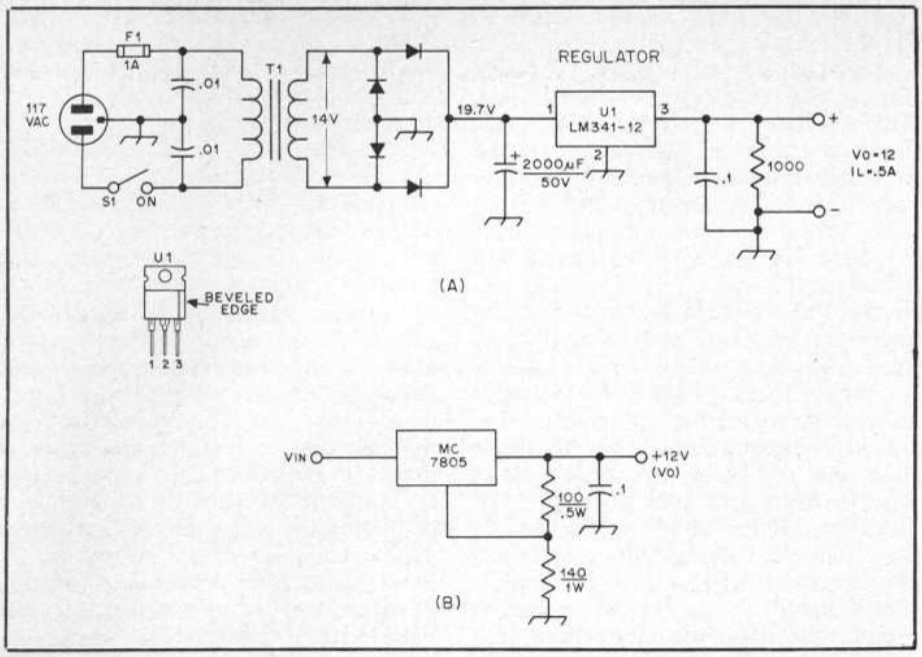


Fig. 46 — The illustration at A is that of a 12-V, 0.5-A supply which employs an LM341-12 regulator IC. Shown at B is a resistive divider which permits elevating the IC output voltage above the value it is designed to handle (see text).

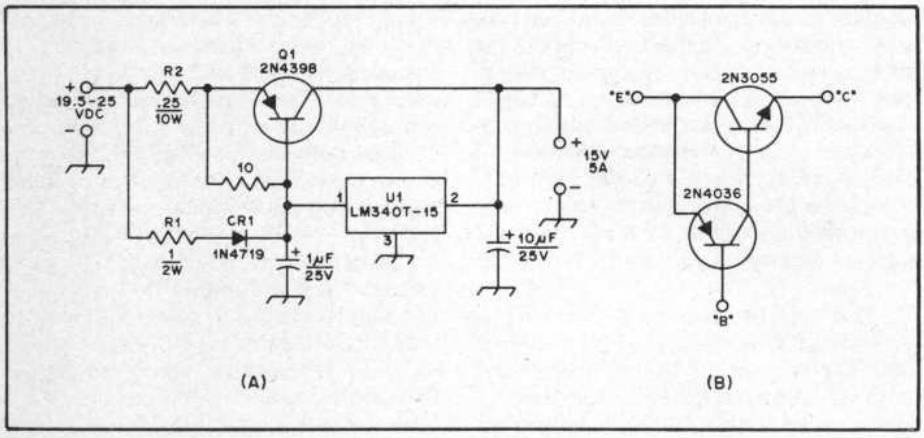


Fig. 47 — Method of extending the current range of a regulator IC. Here we see Q1, a pass transistor, "wrapped around" U1 to increase the current capability of the power supply.

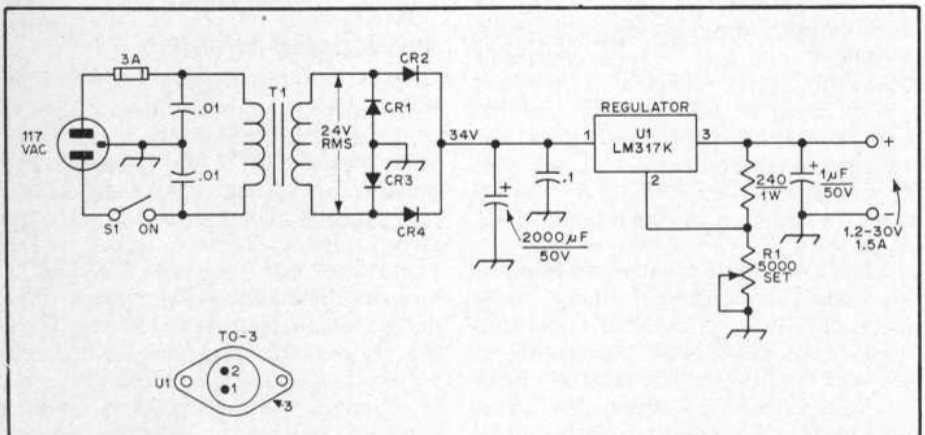


Fig. 48 — Circuit of a continuously variable regulated supply which utilizes the LM317K regulator IC. CR1 through CR4 are 100-PRV, 3-A diodes. Line regulation is .01 percent/V and load regulation is 0.1 percent.

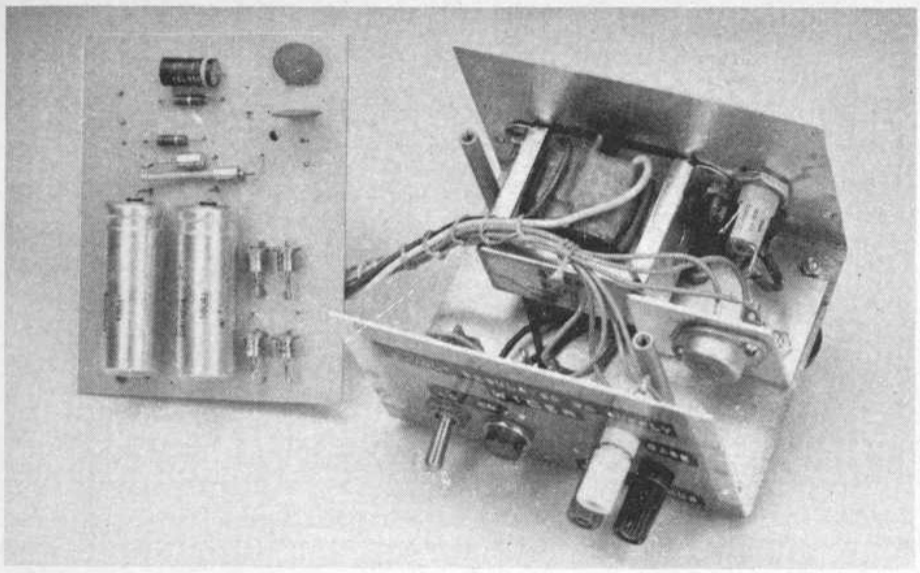
unregulated voltage and an output. The ground reference is usually connected to the mounting surface. Because of this, it is not necessary that the IC be electrically insulated from ground. This eases the heat-sinking problem. Another typical feature is that of "thermal shutdown." If the chip should become excessively warm due to insufficient heat sinking, the temperature rise that accompanies the excessive power causes the current to decrease. Some of the newer three-terminal regulators even have a rather "heroic," fail-safe mode built into them. They are designed such that should excessive power dissipation occur (which would cause destruction of the IC) they fail as a short circuit. The result is a blown fuse farther back in the power supply. However, the circuit that is powered by the IC is never subjected to excessive, potentially destructive voltage.

Since most of the design work is done by the manufacturer, our discussion will deal mainly with practical applications of these components. The first consideration is to ensure that sufficient heat sinking is provided. The power dissipation will be determined by the current in the output and the voltage difference between the regulated output and the unregulated input. Another precaution that should be followed is proper bypassing. Under normal power supply construction this is of minimal significance. Only a 0.1-μF capacitor is required at the output. If, however, the regulator is to be located some distance from the unregulated supply, it is recommended that an electrolytic capacitor be placed across the input port. Usually, a value of 5 μF is sufficient.

Fig. 46A illustrates a 12-volt, 0.5-A regulated power supply which employs a National Semiconductor LM-341-12 IC. U1 should be affixed to a heat sink if heavy continuous currents are anticipated. If only intermittent current loads are expected such as might be encountered with a low power cw transmitter, the chassis will usually offer adequate heat sinking. Available also for the type of circuit shown are 3-A regulator ICs. They are contained in a TO-3 type of case.

One virtue of most of the three-terminal regulators available is that very little current flows in the ground leg of the devices. Assume that an MC-7805 is available. This IC provides an output of 5 volts, but is otherwise similar to the LM-341-12. This regulator could be employed in the 12-volt supply by using a resistive divider connected to the common pin of the IC. This variation is shown in Fig. 46B. In this application, the case of the MC-7805 must be insulated electrically from ground.

When it is desirable to extend the



Inside layout of the 500-mA supply. The pass transistor is seen on an L-shaped homemade heat sink.

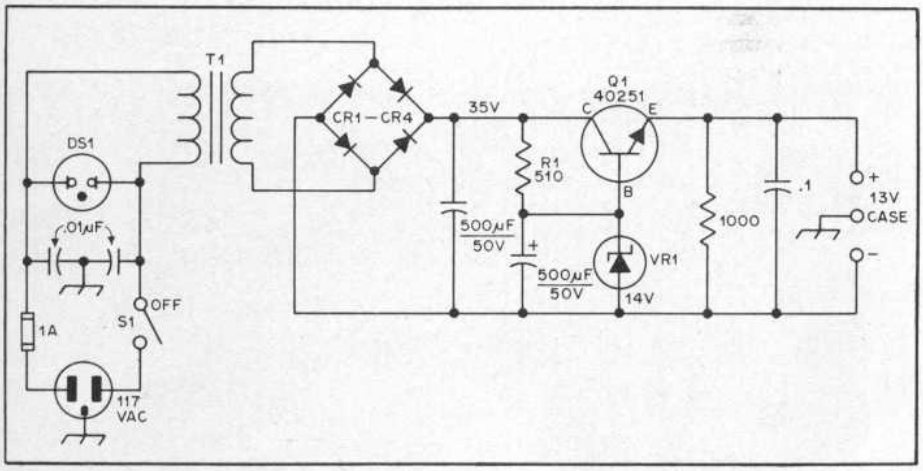


Fig. 49 — Schematic diagram of a 13-V, 0.5-A regulated supply. No overload protection is included, making it mandatory that the operator avoid dead shorts or heavy overloading at the output of the supply. T1 is rated at 1 A and has a 25-V secondary. CR1 through CR4 are 50-PRV, 1-A diodes. VR1 is a 14-V, 1-W Zener diode. Q1 should be mounted on a large heat sink, at least 3 X 3 inches in size.

current range of a regulated power supply beyond that of the regulator IC, the circuit of Fig. 47A can be used. In this example a series pass transistor, Q1, is "wrapped around" the IC to boost the current capability of the circuit. The operation of this circuit can be understood by noting the values of R1 and R2. Assume that the beta of Q1 is high. Most of the three-terminal regulator current will flow through the 1-ohm resistor and the diode, CR1. The offset voltage in CR1 is approximately the same as the emitter-base voltage of Q1. Because of this, the voltage drop across the 1-ohm resistor, R1, will be the same as that across R2. Since the ohmic value of R2 is 0.25 of R1, four times as much current will flow in Q1 as appears in the input terminal of U1. The net result is that the current capability of the overall circuit is increased by a factor of 5. The current limiting characteristics of the IC are transferred directly to the composite circuit.

Sometimes a power pnp transistor is not available in the home stock. Npn power transistors are much more common. Fig. 47B shows a scheme for building a "synthetic" pnp power transistor. This variation uses an npn power device with a smaller pnp transistor.

A continuously variable 1.5-A regulated supply can be built as shown in Fig. 48. The LM317K IC can be used at any fixed output-voltage level by setting R1 to provide the desired output, V_o. Alternatively, R1 can be panel mounted to enable the builder to have a supply which can be varied from 1.2 to as much as 37 volts output. U1 of Fig. 48 has built-in current and temperature limiting (thermal shutdown). A ripple rejection ratio of 80 dB is possible with this circuit. Output current limiting occurs at approximately 2.3 A. This much current could not be obtained at the higher output voltages. This is because of the relatively small value of filter capacitance used. The design rules for the unregulated power supplies which feed the regulators in Figs. 46 through 48 are as given earlier in this section.

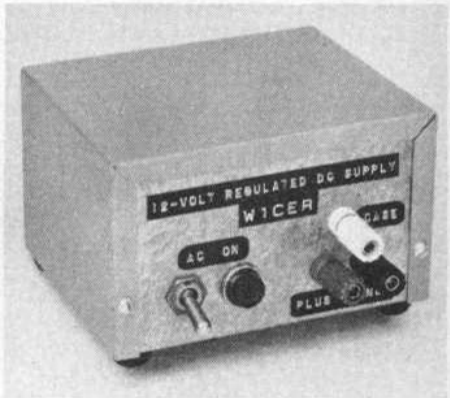
A Low-Cost 13-V Supply

Fig. 49 shows a practical circuit for a 13-volt, 0.5-A regulated dc supply. It is housed in an aluminum Minibox, and some of the components are mounted on a homemade pc board in an effort to enhance compactness.

There is no temperature compensation or short-circuit protection circuitry included. The operator should exercise care by preventing crowbar conditions to exist at the power-supply output. Short-term overloads other than a dead short can be withstood for a few seconds without damage occurring to Q1, the pass transistor. Loads in excess of 500 mA will degrade regulation and cause excessive ripple in the output voltage.

Output voltage amounts other than 13 can be obtained by substituting suitable component values at R1 and VR1 (Fig. 12). Necessary information for the design changes was given earlier in this chapter.

The 510-ohm value listed for R1 was based on a minimum dc beta of 15 for Q1 — the value given in RCA's data sheet for the 40251. The calculated value was 488 ohms, so the nearest, higher, standard resistance value was used, 510 ohms. The photographs show the general layout of the power supply. The container measured 3 X 4 X 5 inches. The positive and negative terminals at the output are above chassis ground, thereby permitting the operator a choice of power-supply polarity. A third terminal is common to the case. It can be wired to the polarized terminal which will be employed as the common



External view of the 12-V, 500-mA regulated dc supply.



Exterior view of the 12-V, 2-A regulated supply.

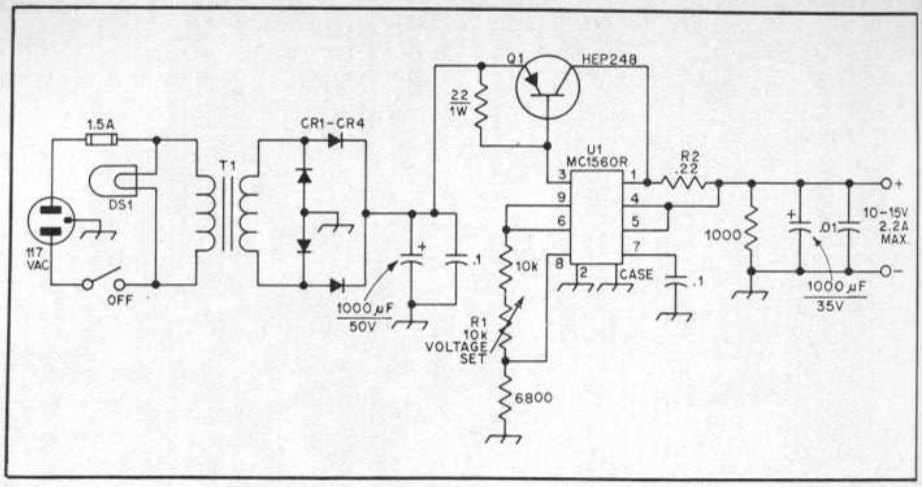


Fig. 50 — Circuit details for a variable-voltage (10 to 15) 2-A regulated power supply which has overload protection. Resistors are 1/2-W composition unless noted differently. CR1 through CR4 are 100-PRV, 6-A diodes. DS1 is a 117-V neon lamp assembly. Q1 should be affixed to a large heat sink (3 X 4 inches or greater), and is a Motorola HEP248 or equivalent. R1 is a pc-mount control. R2 can be formed by winding a suitable amount of magnet wire on a short length of 1/4-inch diameter insulating rod (see wire table in *Handbook* or *ARRL electronics data book* for wire resistance per foot). T1 should have an 18-V secondary with a 3-A or greater rating. A 24-V transformer can be used by removing a few secondary turns. Noise output is 10 mV under a 2-A load. U1 is a Motorola regulator IC.

bus for the equipment used with the regulated supply.

A 2-A Regulated Power Supply

Shown in Fig. 50 is a 2-A regulated dc power supply which can be adjusted to deliver 10 to 15 V. It is protected against overloads and short circuits. Output ripple is low, amounting to 10 mV when a 2-A resistive load is connected across the output terminals. Regulation and filtering remain good up to load conditions of 2.2 A.

An 18-V, 3-A transformer was used for T1 in the example shown. It was obtained as a surplus item — brand and number unknown. However, it should be a simple matter to modify a 24-V transformer of suitable current rating, thereby obtaining an rms secondary voltage of 18. At the expense of overall efficiency, a 24-V transformer can be used with this circuit.

The power supply is contained in a homemade aluminum case (see photographs), which measures 3 X 5 X 6 inches (HWD). A perforated top cover is used to permit the egress of heat from the transformer, regulator IC, and pass transistor.

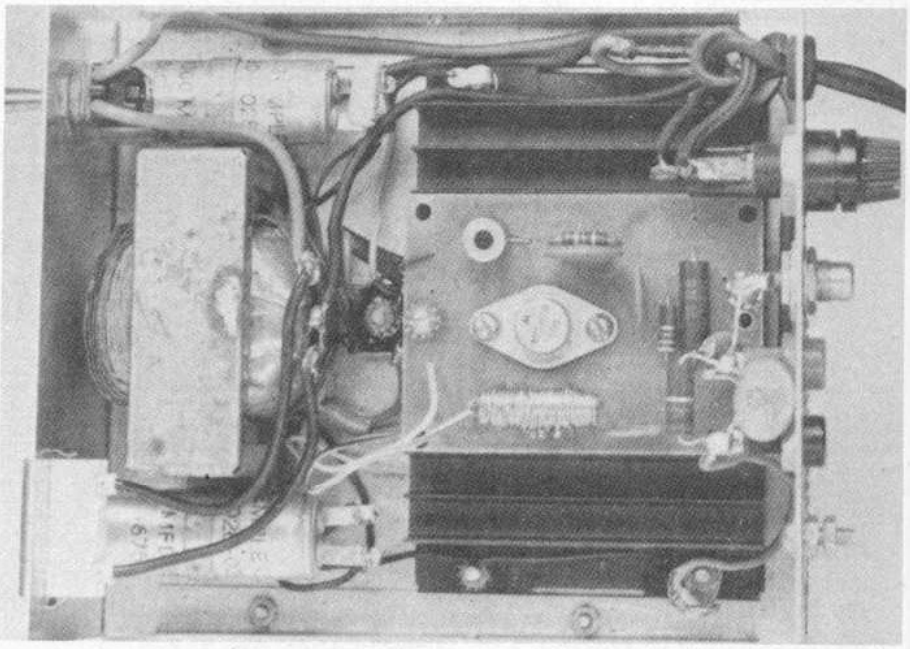
Q1 is mounted on a 3 X 4-inch heat sink. The latter is affixed to insulating hardware, as all three terminals of Q1 must be above ground.

R1 is adjusted for the desired dc output voltage. R2 is fashioned from No. 30 enameled wire. The required number of wire inches to provide 0.22 ohm of resistance are scramble wound on a 10,000-ohm, 2-W resistor body. The resistor pigtails are used as terminals for the winding. U1 and some of the small components are installed on a homemade pc board.

A Husky 12-V Power Supply

Fig. 51 shows the schematic diagram of a 10-A regulated power supply which can deliver 11 to 14 volts of output. It was designed and built by W1GQO.

Three 6.3-V, 10-A filament transformers are used with their primary windings in parallel. The secondaries are series-connected in the proper phase to provide a combined rms output of 18.9 V. This causes a dc output potential from the bridge rectifier of 26.6 volts.



Interior view of the 2-A power supply. The pass-transistor heat sink is below the regulator board.

There is nothing critical about the packaging format of this power supply. The important consideration is, however, one of using heavy-gauge conductors for point-to-point wiring in those circuits which carry the full voltage and current of the unit: No. 14 or heavier hookup wire is recommended. The accompanying photograph shows how the power supply can be assembled to assure reasonable compactness. The case is homemade, and measures 3-3/4 X 6 X 10 inches (HWD).

Q2 is mounted on a home-built heat sink which was fashioned from 1/32-inch thick aluminum. It measures 3 X 1-1/2 inches. Similarly, the rectifier diodes (stud mount) are located on a homemade sink, 2 X 3 inches. Both handmade sinks have mounting feet formed by bending the stock at 90 degrees to form an L bracket. The small part of each L is 3/4 inch deep. Q3, the main pass transistor, is placed on a finned heat sink purchased from Radio

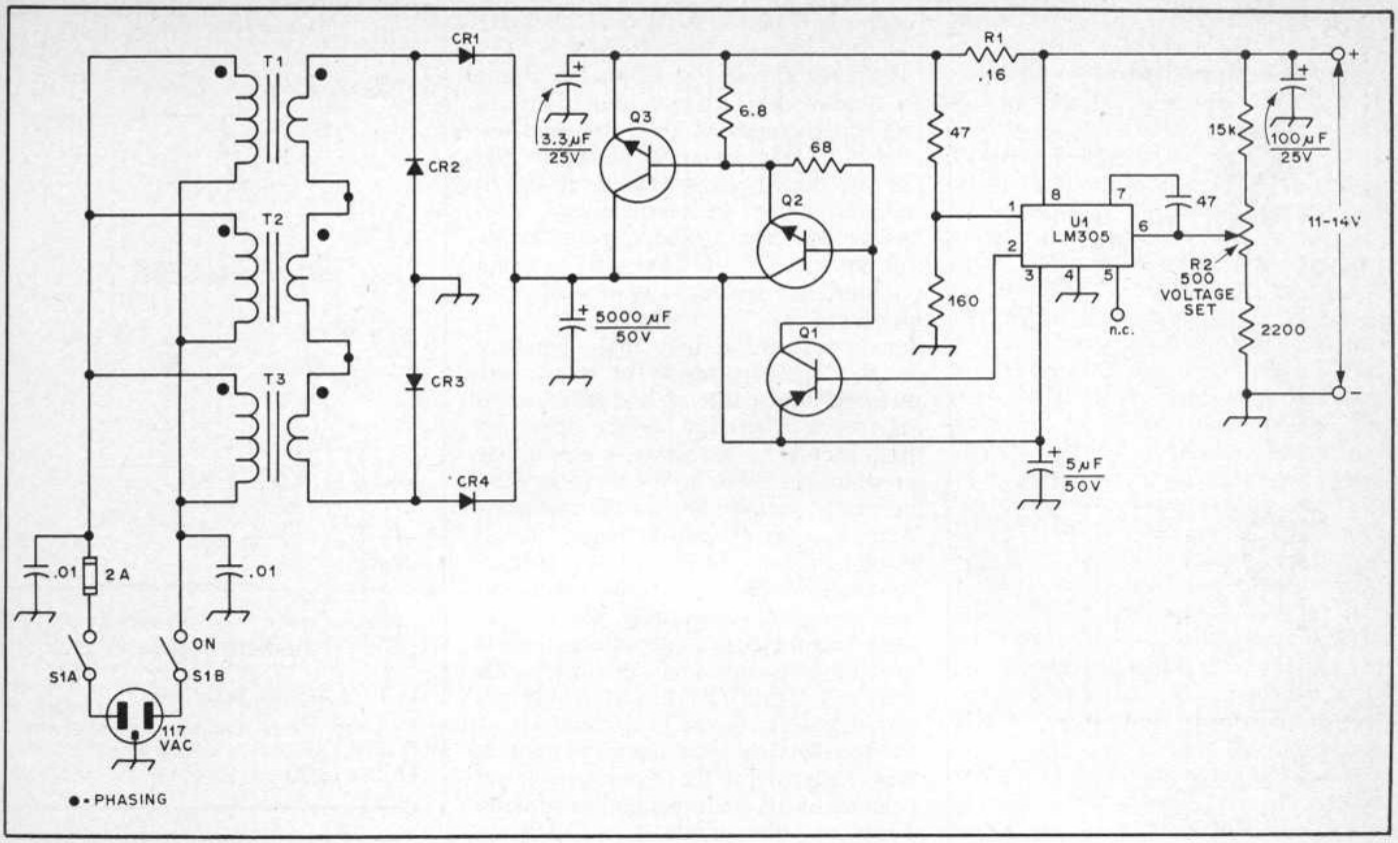


Fig. 51 – Schematic diagram of an 11- to 14-volt power supply with regulation, overload protection, and a 10-A rating. This circuit appeared first in *QST* for August, 1976, p. 26. CR1 through CR4 are 50-PRV, 12-A diodes. Q1 is a 2N2905, Q2 is a 2N3445, and Q3 is a 2N3772. U1 is a National Semiconductor LM305 IC. T1 through T3 are 6.3-V, 10-A filament transformers connected so that the secondaries are in series (observe proper phasing). See text for data on R1 and R2.

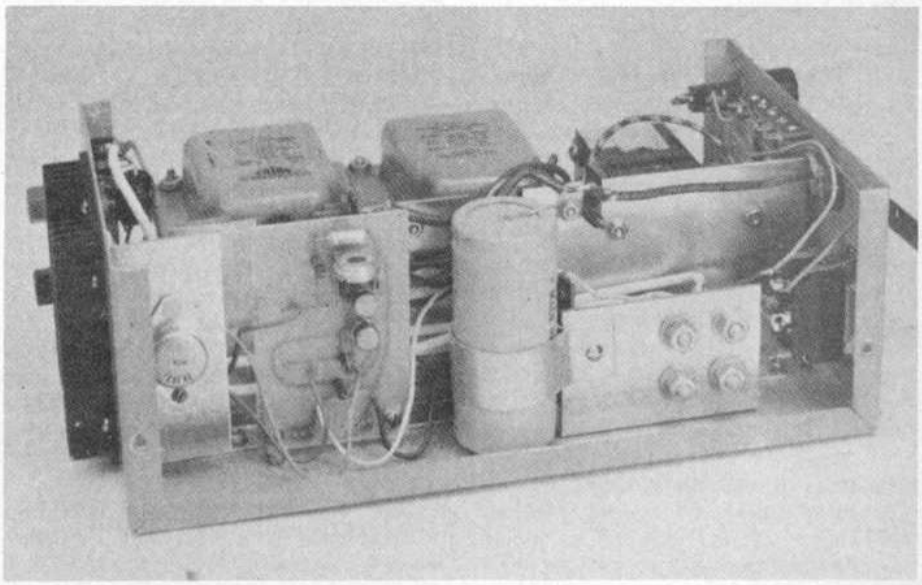
Shack. It is 3-inches long and 2-inches wide. All of the heat sinks are bolted to the main chassis as an aid to heat transfer. Silicone grease is used between the sinks and the chassis, and between the transistor bodies and the heat sinks. Diodes CR1 through CR4 are treated in a similar manner.

R1 is made by winding 9.7 feet of No. 22 enameled wire on the body of a 10-kΩ, 2-W resistor. The desired output voltage is set by means of R2. The power supply has low ripple and is protected against overloads and short circuiting.

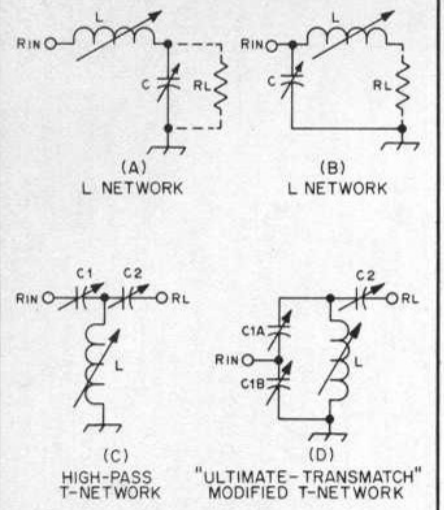
Antenna Matching Techniques

Most solid-state transmitting and receiving equipment is designed to interface with a specific load impedance, respective to the antenna system. In most applications that impedance is between 50 and 75 ohms, assuming that unbalanced coaxial feed lines are used. Generally, coaxial feeders are used with single-band dipoles or gain types of antennas (beams). Multiband trap dipoles, beams, and verticals also dictate the use of coaxial feeders in most examples, although it is possible and practical to employ balanced two-wire feed systems with most of the antennas just mentioned.

Because amateur transmitters and receivers are designed to operate at a particular antenna-impedance level, a matching network is used sometimes to effect maximum power transfer between the antenna and the equipment – the purpose for creating a matched condition. Although many antennas can be matched at the feed point to the type of transmission line used, thereby eliminating the need for a matching network at the equipment end of the circuit, a matcher at the shack end of the system has some virtues. (1) A Transmatch (transmission-line matcher) enables the operator to maintain an



View showing the interior of the 10-A regulated supply.



$$R_{in} < R_L$$

$$X_c = R_L \sqrt{\frac{R_{in}}{R_L - R_{in}}}$$

$$X_L = \frac{R_L R_{in}}{X_c} \qquad (A)$$

$$R_{in} > R_L$$

$$X_L = \sqrt{R_L R_{in} - R_L^2}$$

$$X_c = \frac{R_L R_{in}}{X_c} \qquad (B)$$

$$X_{C1} = R_{in} \sqrt{\frac{R_L (Q_L^2 + 1)}{R_{in}} - 1} \qquad (C)$$

$$X_{C2} = Q_L R_L$$

$$X_L = \frac{R_L (Q_L^2 + 1)}{Q_L} \times \frac{1}{1 + \left[\frac{X_{C1}}{Q_L R_{in}}\right]} \qquad (D)$$

Fig. 52 — Examples of L and T types of matching networks.

SWR of 1, or nearly so, over an entire amateur band without a need to readjust the match at the antenna feed point. Having a Transmatch at the equipment end of the circuit does not, of course, correct the mismatch *at* the antenna: It merely disguises the condition so that the equipment sees the *desired* load impedance. (2) Depending upon the kind of Transmatch used, harmonic energy from the transmitter can be attenuated by 30 dB or more as the signal passes through the matching system. This requires a low-pass or bandpass type of network. High-pass networks of the kind found in the Ultimate Transmatch, popularized by W1ICP in *QST* for July, 1970, are of less value in this regard, despite the wide range of impedances they are capable of matching. Fig. 52C shows the basic high-pass T network. At D is the modified configuration described by W1ICP.

Shown also in Fig. 52 are two forms of L network which are useful in matching the equipment to a transmission line. All of the equations shown in Fig. 52 are based on matching loads to sources which are, respectively, pure resistances. The L and C components for the circuits are illustrated as being variable. In a practical situation the load presented by the transmission line is purely resistive at only that frequency in the amateur band for which the antenna is constructed and matched to its feeder. Therefore, as the operating frequency is moved above or below that at which an SWR of 1 exists, the load becomes reactive. Should the reactance become great enough in magnitude to result in a high SWR, say, 2:1 or greater, the transmitter may not load into the antenna system effectively, thereby endangering the output transistors (if SWR protection is not included in the PA stage). A high SWR will also reduce the power transfer to the load. Similarly, if the receiver front end has a filter which was designed for the characteristic impedance of the transmission line (usually 50 ohms), the mismatch will degrade the filter performance. Because of the foregoing considerations it is necessary to make the L and C elements of the network variable to permit matching to loads which exhibit unknown reactances. These reactances are reflected to the equipment end of the feed line by the antenna when a mismatch is present at the feed point. As the mismatch at the antenna increases so does the loss in the feeder. The higher the operating frequency, the more pronounced the loss condition becomes. In situations where a high SWR must be accepted, as may be the case in some portable or emergency operations, high-quality (low-loss) feed line should be used. If the feeder length is less than 50 feet at frequencies in the hf and mf spectrum, RG-58/U and RG-59/U should be suitable with respect to losses versus SWR. Subminiature coaxial cable (RG-174/U type) is not recommended except when other types of cable are too heavy. For feed-line runs greater than approximately 50 feet, RG-8/U or RG-11/U cable is a better choice, even when the SWR is not high. Open-wire feeders will have the lowest loss factor of the numerous kinds of transmission lines because the dielectric material is air, principally. Feeder losses and impedance matching are especially significant when QRP equipment is being used — every dB counts!

The T networks at C and D of Fig. 52 are capable of accommodating a much greater range of impedances than would be possible with L or pi networks. For field work this is an important consideration, for makeshift antennas are often used during portable operations. The equations given are based on a loaded Q of 5, which is an arbitrary figure picked by the writers. Other values of Q would be acceptable, but the low figure of 5 has proved to be practical in the interest of matching-network bandwidth. More specifically, the higher Qs require that the Transmatch be readjusted even when small changes in operating frequency are made. The higher the network Q, the more critical the adjustment procedure — another consideration. A Q of 5 is a practical ball-park figure, and yields practical L and C values for a wide range of impedance conditions.

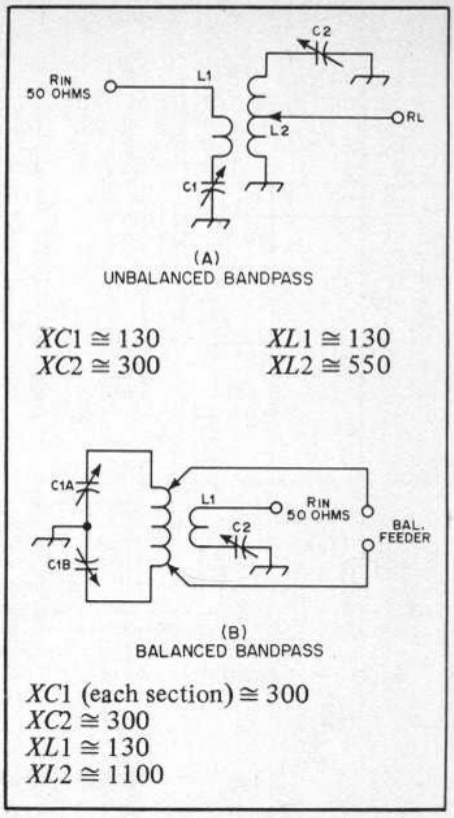


Fig. 53 — Bandpass types of matching networks. These are used frequently in Transmatches. They offer harmonic rejection.

T-network Transmatches of the type shown have the advantage of rejecting frequencies below the one to which they are tuned. Therefore, the high-pass characteristic can be used to advantage in rejecting bc-band energy which could affect the performance of a receiver. Those who live near bc stations often experience problems with receiver overloading and IMD when operating on 160 or 80 meters.

Other Matching Networks

Operators who wish to take advantage of the harmonic-suppression characteristics of a bandpass type of Transmatch may elect to use one of the circuits shown in Fig. 53. A bandpass

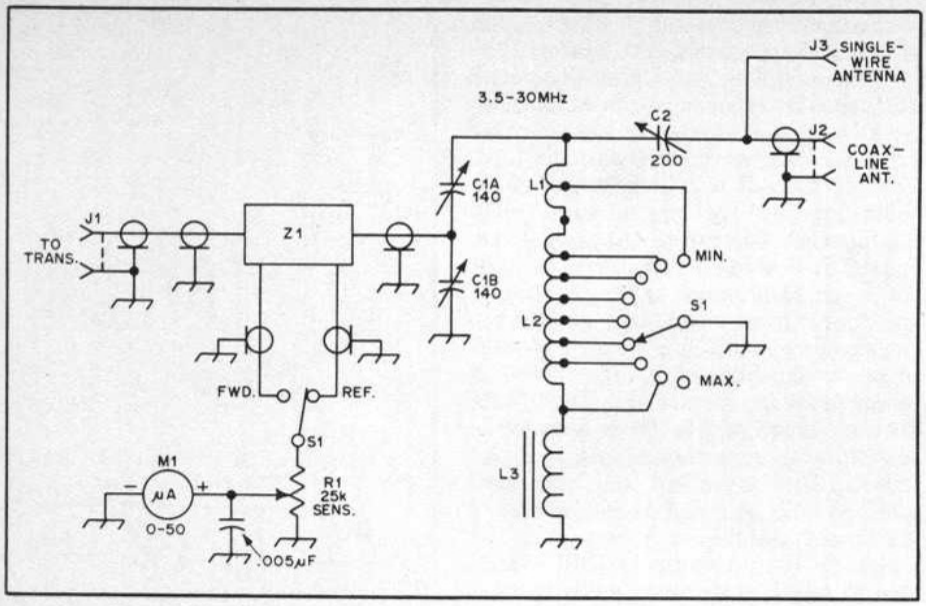


Fig. 54 — Transmatch which features a modified T network. C1 is a ganged pair of Millen 19140 variable capacitors. C2 is a 200-pF variable taken from a surplus Command transmitter. L1 has a 1/2-inch diameter, is 2 inches long, and contains 8 turns of No. 18 wire 3002 Miniductor). L2 is 4 inches long, has a diameter of 1-3/4 inches, has 32 turns of No. 14 wire, and is tapped every 4 turns (3022 Miniductor). L3 is a toroid inductor with 35 turns of No. 20 enam. wire on an Amidon T130-2 core. S1 is a single-pole, 10-position rotary ceramic wafer switch with the shaft and collar insulated from ground. Z1 is the circuit of Fig. 15.

network will also aid reception through rejection of frequencies above and below the one to which the network is tuned.

At Fig. 53A is an unbalanced bandpass matching network that can be used between the station equipment and the coaxial feeder. Alternatively, it can be placed between a single-wire antenna (resonant or random length) and a coaxial feed cable to the amateur station. Reactance values are given to permit calculation of the *L* and *C* values for a given band of operation. For multiband use, *C* and *L* should be chosen for the lowest operating frequency anticipated. In such an event, taps should be placed on L1 to permit matching at the high end of the Transmatch frequency range. L1 and C1 must be able to form a resonant circuit at the operating frequency. Likewise with L2 and C2. The tap on L2 is moved experimentally, along with adjustment of C1 and C2, to obtain an SWR of 1.

The operating principle and adjustment procedures are the same for the circuit of Fig. 53B. In this example the Transmatch is designed to accommodate balanced feeders, such as would be used with an end- or center-fed Zepp antenna. *The ARRL Antenna Book* contains in-depth descriptions of various antennas that can be used with these Transmatch circuits.

For multiband use of the network in Fig. 53B, it will be necessary to tap C1 toward the center of L2 as the operating frequency is increased. Similarly, taps should be placed on L1. Respective to all of the matching circuits shown here, the wire size of the inductors and the plate spacing of the variable capacitors must be adequate for the power level employed. The wire size should be great enough to minimize *IR* losses and heating. Capacitor plate spacing should be such that arcing does not occur during periods of high SWR – as encountered during system adjustment.

Transmatch Adjustment

Precise adjustment of a Transmatch is done best by applying transmitter power and observing an SWR indicator while adjusting the network. Tuning should be done at the lowest power-output level practicable, thereby minimizing damage to the PA stage and lessening the chance of causing QRM to those who may be using the frequency. Various kinds of SWR indicators are suitable for use with Transmatches, but for on-the-nose adjustments the instrument should have high sensitivity: Full-scale deflection of the indicating meter should be possible at the low-power level used during initial setup of the Transmatch. In this regard the circuit treated by Bruene in *QST* for April, 1959 is excellent. He described the design features of a directional wattmeter which used a toroidal current-sampling transformer in an rf bridge circuit. Practical examples of that type of instrument were given earlier in this chapter and in *QST* for December, 1969. Circuits were described for power levels from 5 to 1000 watts. There are two distinct advantages offered by the Bruene circuit over that of the so-called Monimatch SWR meter described by McCoy in the 1950s (*QST*). The latter exhibits extreme frequency sensitivity, with declining sensitivity as the operating frequency is lowered. Instruments of that kind are not suitable for QRP work unless a meter amplifier is used. Additionally, it is difficult to employ the Monimatch circuit as a calibrated wattmeter because of its frequency sensitivity. The Bruene circuit, however, is suited to the purpose in an ideal manner. An SWR indicator of this variety can be used for Transmatch adjustment and for measuring rf power. Fig. 15 shows a practical circuit for a 10- to 1000-W version of the bridge. Fig. 17 shows the schematic diagram of another version of the instrument – capable of full-scale deflection at 1 watt. Each of the examples are suitable for use when adjusting Transmatches.

In a practical situation, the SWR indicator is placed between the transmitter and the Transmatch. The indicator is set for maximum sensitivity in the reflected-power position. Transmitter power is advanced to obtain a few divisions of meter deflection. The Transmatch controls are adjusted to cause a meter reading of zero. The transmitter is retuned for maximum PA output without increasing the drive. Next, the SWR indicator is set for a forward-power reading and the sensitivity control is adjusted for a full-scale meter reading. Then, the operator returns the bridge to the reflected-power mode and makes final adjustments with the Transmatch to secure zero meter deflection. Normal operating power can be established now, setting the sensitivity control of the bridge for full-scale indication on the meter (forward-power mode). Bridges which are intended for rf-power reading do not necessarily have sensitivity controls on the instrument panel. There-



Fig. 55 — Exterior view of the Transmatch as seen in its homemade aluminum case. The control at the upper right is not used.

fore, adjustments of the Transmatch must be made while utilizing whatever amount of meter-scale deflection is available.

In situations where the meter will not drop to zero, no matter how carefully the Transmatch is adjusted, it will be likely that the transmitter is putting out considerable harmonic energy. Even though a perfect match has been effected at the desired operating frequency, the harmonic energy is being reflected back to the bridge, causing a false indication that high SWR exists. A remaining cause of imperfect meter zeroing can be brought about by a bridge that was not nulled properly at the operating frequency. That is, although it had a characteristic impedance of 50 ohms at some frequencies in the hf spectrum, internal unwanted reactances in the bridge circuit could make the instrument other than 50 ohms at some specified frequency. The effect is one of not getting a reading of zero when an SWR of 1 exists in a 50-ohm feeder system.

Fig. 54 shows the circuit of a modified T-network Transmatch of the kind illustrated in Fig. 52D. It is designed to operate from 80 through 10 meters at power levels up to 150 watts continuous. Although C1 is a dual-section capacitor assembly, configured as a dual-differential variable, a single capacitor can be used to form the circuit of Fig. 52C. The dual-differential capacitor arrangement of this circuit was employed for experimental purposes. In practice there is little difference in the matching ranges of the three circuits. A rotary inductor can be used in place of the tapped coil and switch shown, and will ensure a greater impedance-matching range than the tapped coil will. Transmatches of this type should always be adjusted so that the maximum practical amount of C is used at C2 during a matched condition. The tighter coupling will provide greater Transmatch efficiency (lower insertion loss), and will lower the circuit *Q* by virtue of tighter coupling to the load. The latter will lessen the need of readjusting the Transmatch when small changes in operating frequency are made. It should be noted that an SWR of 1 can be obtained at *various* settings of the controls, but always use as much capacitance at C2 as is possible, consistent with an SWR of 1. Figs. 55 and 56 show how the Transmatch is built. A Bruene type of rf bridge is included in the box to permit monitoring of the SWR. The assembled unit measures (HWD) 4-1/2 X 8 X 7 inches, and has a homemade aluminum cabinet.

Fig. 57A illustrates a QRP Transmatch which is suitable for power levels up to 25 watts. Because of its small size it is ideal for field applications. An external SWR indicator is needed with this unit. A homemade variable inductor, designed and built by K1KLO, is the heart of the matcher. It contains one half of a powdered-iron toroid core (1-inch diameter core, No. 2 iron mix, wall height and thickness of 3/16 inch). The core material moves in and out of a hand-wound coil which contains 32 turns of No. 22 enamel wire, 7/16-inch OD. A detailed description of this Transmatch was published in *QST* for February, 1976. Fig 57B shows a method for adding 80-meter coverage.

A slug-tuned coil (L1 of Fig. 57A) is switched in parallel with the half-toroid one (L2) to lower the inductance during operation on 20, 15, and 10 meters. The former has an inductance range of 3 to 9 μH. The slug-tuned inductor has a 3.1- to 4.8-μH range.

Simplification of the circuit will result if C1 is replaced by a single 365-pF unit of the type used at C2. The resulting circuit would be similar to that of Fig. 52C. This Transmatch is housed in a 1-1/2 X 2-3/4 X 4-inch plastic meter case. Phono connectors are used for the input, output and ground terminals. Alligator clips have been soldered to phono plugs to facilitate connections to earth ground and a single-wire antenna, if the latter is used. Figs. 58 and 59 show how the unit is built.

A 40-Meter Transmatch

Fig. 60 shows the circuit of a QRP Transmatch for use on 40 meters. The input circuit is arranged for switching a resistive bridge in series with the matching network during adjustment for an SWR of 1. C1, C2 and L1 comprise a high-pass network for matching a wide range of impedances to a 50-ohm source.

During normal operation S1 is placed in the operate mode, bypassing

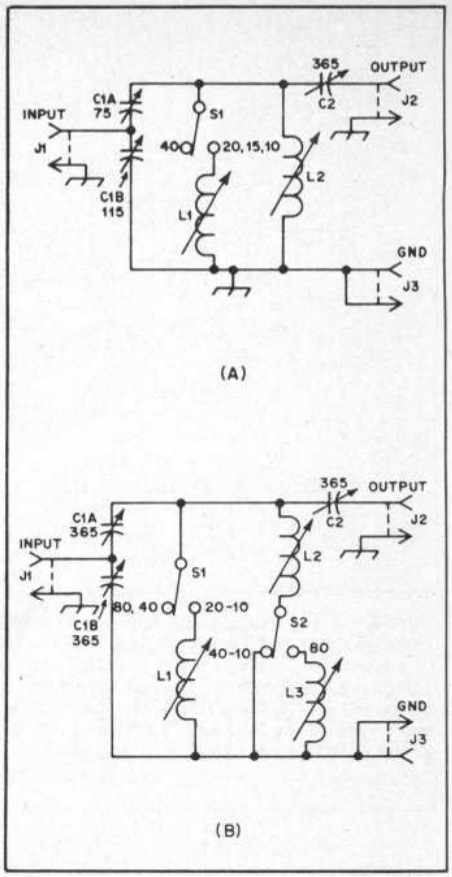


Fig. 57 — The diagram at A is for the 40- through 10-meter Transmatch. At B, a suggested circuit for coverage from 80 through 10 meters.
C1 — Dual-section air variable (Miller 2109, J. W. Miller Co., 19070 Reyes Ave., Compton, CA 90224). See text.
C2 — Calectro or Archer single-section miniature 365-pF variable.
J1-J3, incl. — Phono jack.
L1 — 3.1- to 4.8-μH slug-tuned inductor (Miller 4504 with red core).
L2 — See text. Contains 32 turns of no. 22 enam. wire, air wound, 7/16-inch OD.
L3 — 5.5- to 8.6-μH slug-tuned inductor (Miller 4504 with red core).
S1, S2 — Spdt slide or toggle switch.

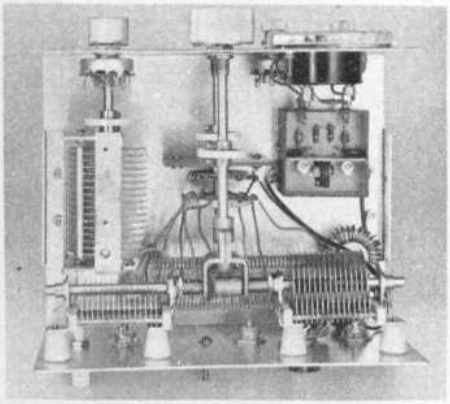


Fig. 56 — Interior of the Transmatch. C1A and C1B are joined by means of a right-angle drive. Insulated shaft couplings are used at S1, C1 and C2 (Millen 39016). S1 is mounted on a phenolic plate (center of picture). An unused coaxial connector is visible at the lower center. Z1 is at the upper right.

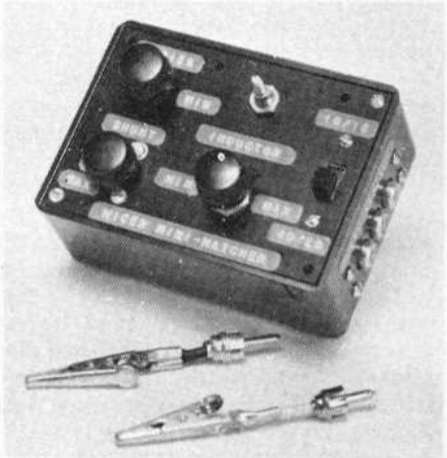


Fig. 58 — Exterior view of the QRP Transmatch. J1, J2 and J3 are seen at the far right.

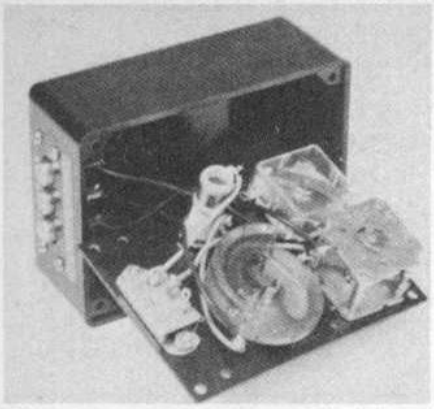


Fig. 59 — Interior view of the Transmatch showing the K1KLO variable inductor at the lower center.

the bridge. Any meter with a sensitivity of 50 to 500 μA will be suitable at M1. The instrument used in this example was borrowed from a junked tape recorder. CR1 is a germanium diode of the IN34A variety.

J3, a single-terminal binding post, is connected in parallel with coax connector J2 to permit attachment of a single-wire antenna. The assembled unit is contained in a small aluminum chassis (5 X 3 X 1 inches). A smaller case can be used if a more compact assembly is desired. Fig. 61 shows how the components are arranged in the box.

Assorted Test Equipment

This section contains a collection of circuits that have been built by the writers for their own use. Many of the details are not presented, because they will depend upon the characteristics of the parts used by the builder. The junk box and surplus market can provide many of the needed components.

Noise Generator

Shown in Fig. 62 is a circuit for a noise generator. This unit was inspired by an investigation of the effects of Zener doides on the noise performance of amplifiers. The experiments suggested that Zener diodes were not optimum for biasing very low-noise amplifiers. This was due primarily to noise-modulation effects when strong signals were present, rather than actual degradation of noise figure.

There is an expression among design engineers when a problem is encountered: "If you can't lick the problem, feature it." This was the policy that was followed in the noise generator shown. The major noise source is CR1, a 5.1-volt Zener diode that is used to bias the first amplifier. Since no bypassing of the Zener is used at the base, and the current in the diode is small, the excessive noise currents in the diode will flow through the base of the transistor. The amplified output is applied to a second stage of gain. The second amplifier has a 51-ohm resistor in the collector to provide a controlled output impedance.

The noise output of this circuit has been measured on a spectrum analyzer. The detailed distribution of noise with frequency will not be presented since it will vary considerably with Zener diode and transistor characteristics. Generally, the noise in the hf region was quite

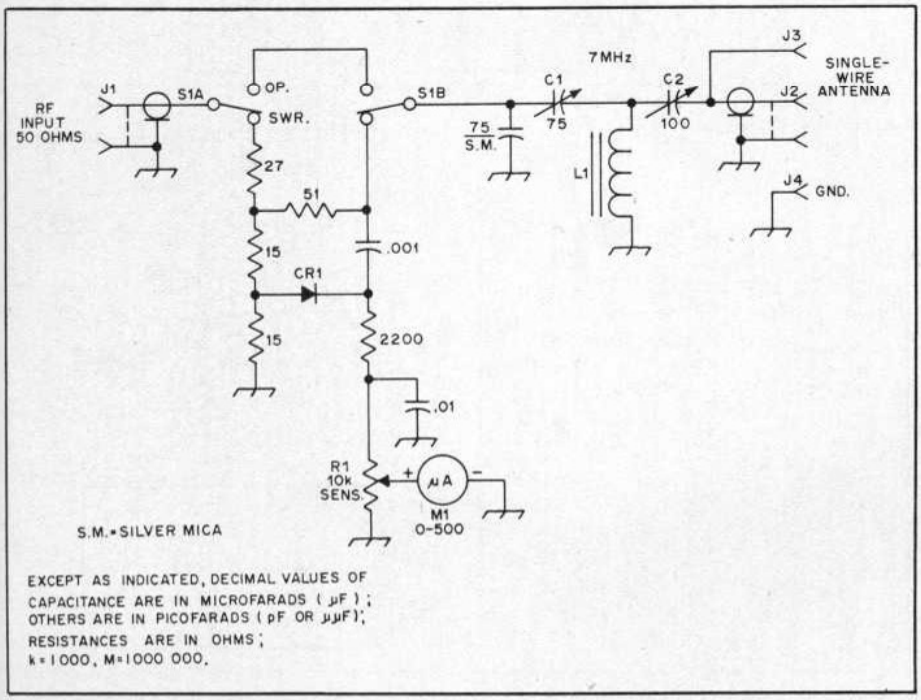


Fig. 60 — Schematic diagram of the 40-meter Transmatch. Resistors are 1/2-W composition. L1 contains 30 turns of No. 22 enam. wire on a T68-2 toroid core. S1 is a dpdt slide switch.

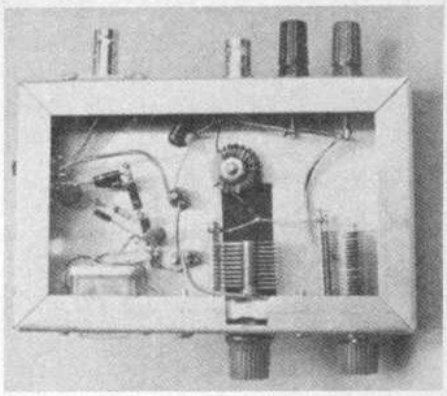


Fig. 61 — Interior view of the 40-meter Transmatch.

robust, reaching levels of 80 dB higher than the noise output from a room-temperature resistor. The noise output is still 20 dB above a 290° K resistor at 432 MHz.

The builder should not attempt to estimate noise figure with a device as crude as this one. It may be used, however, as a source for tuning receivers or amplifiers. If one were to build a free-running multivibrator, using a 555 timer, with a total period of 1 to 2 seconds, it could be used to automatically turn the generator on and off. The system could then be used in conjunction with a step attenuator to adjust a vhf preamplifier for low noise figure. The output detector would be the operator's ears, although refined circuitry could be built for the purpose.

Audio Voltmeter

Shown in Fig. 63 is a circuit for an uncalibrated audio voltmeter. Two 741 operational amplifiers are used. The first one is an amplifier with a voltage gain of 11. The output of this stage has a pair of attenuators that may be switched into the system. The second amplifier contains a meter within a bridge rectifier. Since the rectifier is in the feedback loop of the op amp, diode characteristics are not critical. The diodes should all be of the same type, though.

Calibration of the attenuators is straightforward, although unusual for audio applications. First, a 50-ohm resistor is placed temporarily across the input. Then, an audio generator is obtained and set for a sine-wave output of several volts. A 50-ohm resistor is placed in series with the audio generator output, if the output impedance is as low as would be the case with an op amp output. Then, a 50-ohm step attenuator is set for 30 dB of attenuation and placed between the two units. Power is applied to each, and the input control is set for a full-scale meter reading. The attenuator controlled by S1 is set to the −3 dB position, and the 50-ohm step

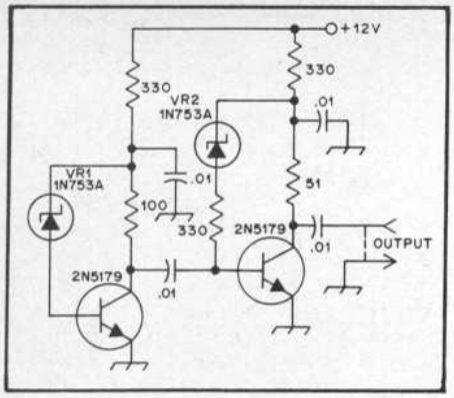


Fig. 62 — Circuit details of a noise generator. Excess-noise output is greater than 70 dB at 14 MHz, and is detectable at 432 MHz.

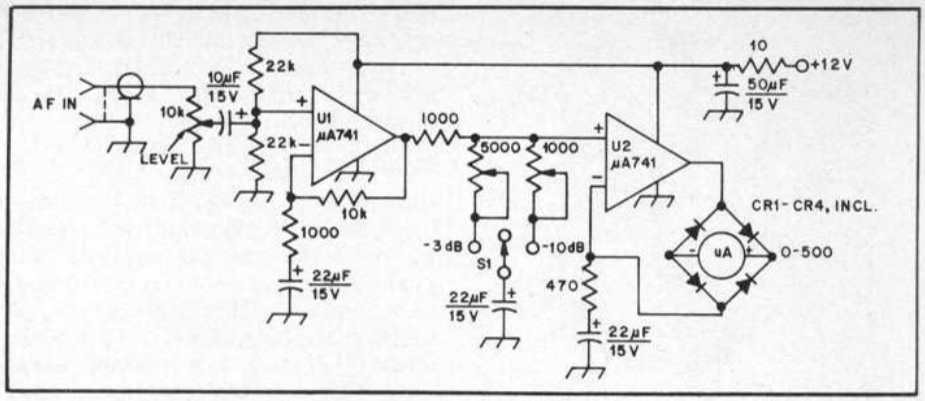


Fig. 63 — Details of the audio voltmeter. CR1 through CR4 are 1N914s. M1 is a 500-µA meter, and S1 is a center-off spst toggle switch.

attenuator is adjusted to increase the power to the audio voltmeter by 3 dB. The 5000-ohm control is then adjusted for a full-scale reading on the meter. The same procedure is used for calibrating the −10 dB position of S1.

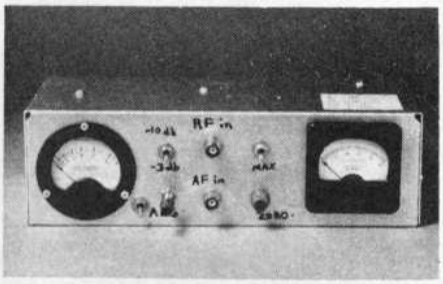
Typically, this meter is used for receiver testing. It can be used without the attenuators for alignment. If a measurement of MDS is to be performed, the gain in the receiver and the level control in the voltmeter are set for a full-scale reading from the noise output of the receiver. Then, a signal generator is applied to the input of the receiver, and S1 is thrown to the −3 dB position. The rf signal generator is set to again yield a full-scale reading. The output power available from the rf generator is then the MDS of the receiver. The signal required to obtain a 10-dB signal plus noise-to-noise ratio can be evaluated in a similar way by using S1 in the −10 dB position.

One refinement that the builder should consider is to add a 47-kΩ resistor between the output of U1 and the electrolytic capacitor connected to S1. This will keep the dc potential on the capacitor equal to that at the output of U1, preventing a large transient when S1 is thrown into one of the attenuation positions.

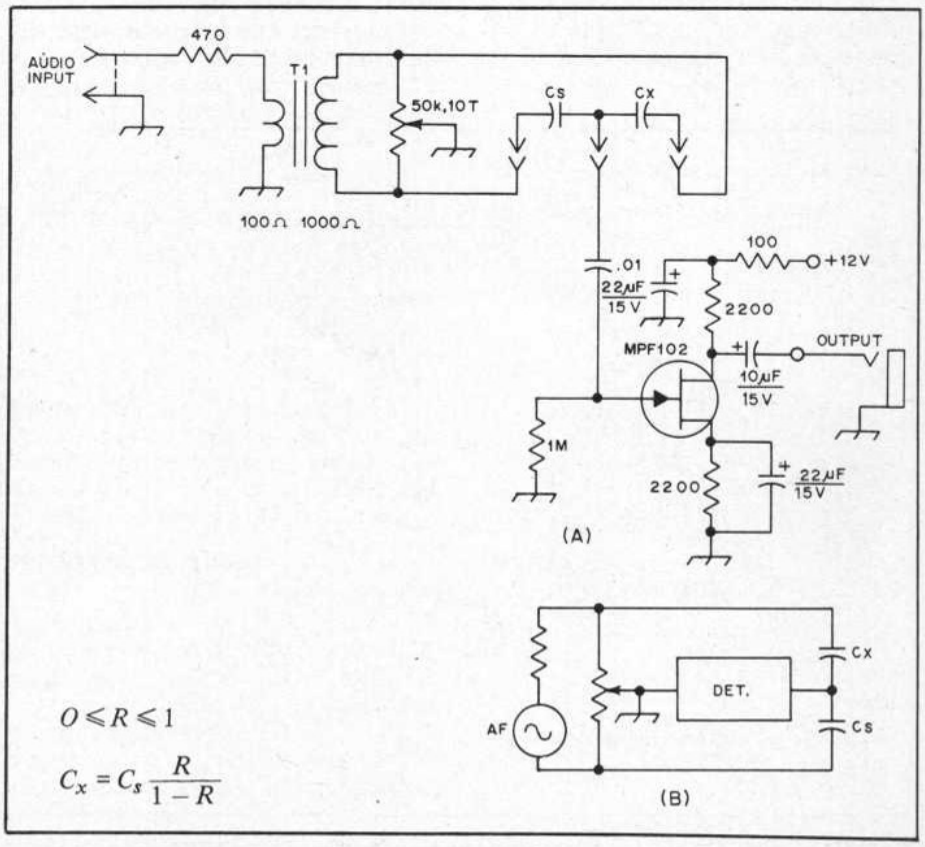
Capacitance Bridge

A simple capacitance bridge is shown in Fig. 64A. This unit is useful for determining the value of unmarked capacitors, such as those of the "dog bone" ceramic variety. The audio source may be from an audio oscillator, a square-wave oscillator, or even the station receiver which is tuned to a steady carrier. The audio input signal is applied to a transformer, T1. The secondary of this transformer is allowed float with respect to ground. However a 50-kΩ control is placed across the secondary with the arm attached to ground. The "unknown" capacitor is placed in series with a capacitor of known value to form a bridge configuration. This is emphasized in the circuit of Fig. 64B.

The junction of the two capacitors is connected to a high input inpedance JFET audio amplifier. Nearly any available FET should be suitable for this application. In use, the control is tuned until minimum output is noted in the



A general-purpose test instrument. The meter at the left is for readout of an audio voltmeter. The instrument is not calibrated, but there are calibrated attenuations of 3 and 10 dB. The meter at the right is an indicator for a broadband rf detector. A broadband amplifier is contained in the case. It allows sensitivities of −65 dBm up through 50 MHz. The rf detector is used in conjunction with a step attenuator to produce 1-dB accurate measurements of gain, return loss and related parameters.



$$0 \leqslant R \leqslant 1$$

$$C_x = C_s \frac{R}{1-R}$$

Fig. 64 — Schematic diagram of the capacitance bridge.

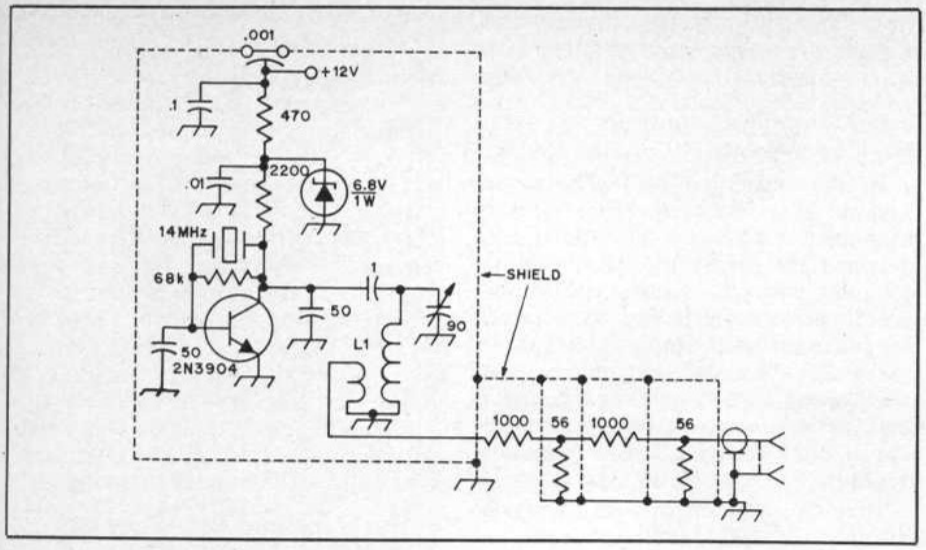


Fig. 65 — Circuit of the weak-signal 14-MHz generator. L1 has 24 turns of No. 26 enam. wire on an Amidon T50-6 toroid core. The link consists of a single turn of wire.

headphones. The depth of the null is quite large in our unit.

In order for this bridge to be useful, it is necessary that the 50-kΩ control be linear and calibrated. In our unit, a 10-turn control is used with a turns-counting dial. If the output of the dial is interpreted as a ratio between 0 and 1, the unknown capacitor is related to the standard capacitor with the equations shown in Fig. 64. If the two capacitors are equal, the bridge will be balanced when the control is set at midrange, where R = 0.5. If the builder does not have a 10-turn control with a turns-counting dial, a more mundane system could be calibrated with a handful of capacitors of known value.

The bridge will operate over a wide range of capacitance. Using a 10-pF standard, very small values are easily determined. An example would be the parallel capacitance of a quartz crystal. Values of up to 0.1-μF have been measured as well. The best accuracy will always be obtained when the standard capacitor is close in value to the capacitor being measured. A group of known-value capacitors with 1-percent tolerance are kept on hand. Three binding posts are provided on the instrument for easy insertion.

Low-Level RF Source

When working on receivers, one of the most useful pieces of test gear one can use is a low-level signal source. While there are signal generators available that will do the job nicely, they are expensive. The less expensive kit generators are not too suitable for precise receiver work, since they have too much leakage to allow the measurement of weak signals. If one ever has the chance to observe the level of shielding and decoupling that is used in a high-quality signal generator, he will realize why inexpensive generators are so leaky.

All is not lost — meaningful weak-signal measurements can be made in the home shop. Shown in Fig. 65 is the circuit of a 14-MHz source. The key to good performance is the shielding. The generator is built in a box made from double-sided pc-board material. A high quality feedthrough capacitor is used to get power into the box, and thorough power supply decoupling is applied within the unit. Extensive attenuation is used within the oscillator housing with shield partitions between the sections of the attenuator. A battery is used to power the unit, thereby avoiding signals that could leak along signal ground paths in the power lines.

The best way to calibrate this source is by using a better generator in conjunction with a receiver. The agc in the receiver is defeated and an audio voltmeter is used to monitor the receiver output. The resistors in the attenuator are picked to provide an output that corresponds to a reasonably weak signal (S5 or thereabouts). The box is soldered shut with the crystal inside the shielded enclosure. The level in the receiver is carefully noted on the audio voltmeter. Then, the signal generator is substituted in place of the crystal-controlled source, and is adjusted for an identical output response. The output is noted, then marked on the outside of the box.

This source is now usable in the shop in conjunction with step attenuators for the measurement of receiver MDS. We have been able to duplicate laboratory results within 1 dB with these methods. It should be mentioned that even if calibration is not possible, a source of the type described can be useful for comparative measurements. Furthermore, since the calibration may be done with a generator that might be too leaky to be useful at really low levels, the techniques may be applied to extend the measurement capabilities of a moderately equipped home shop.

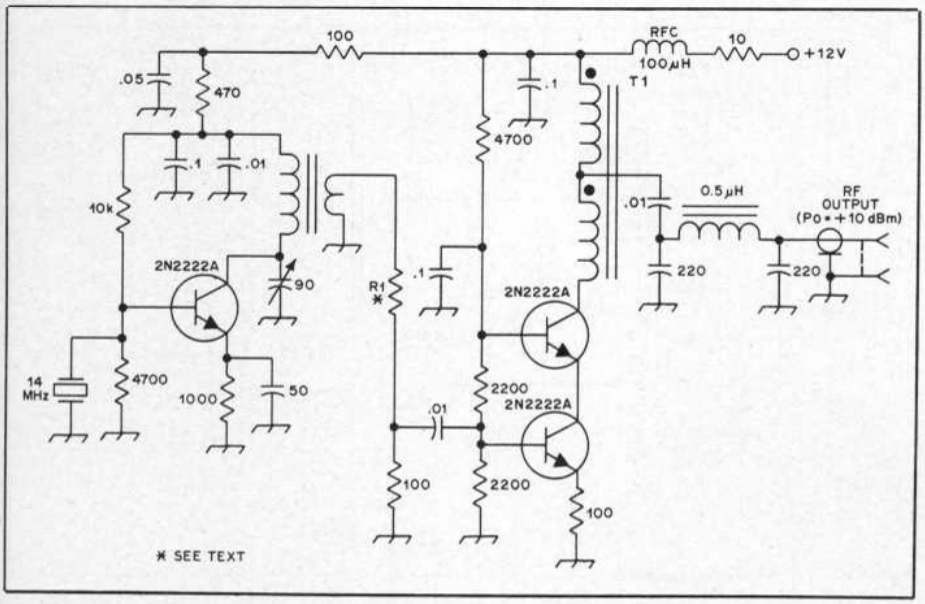
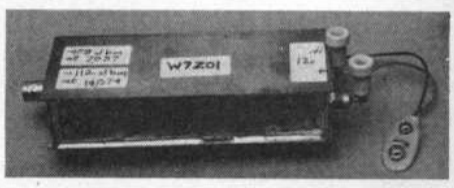


Fig. 66 — Diagram of a signal generator with greater power-output capability than the example in Fig. 65. T1 contains 10 bifilar turns of No. 28 enam. wire on an Amidon FT-37-61 toroid core.



Exterior of the signal generator. It provides low-level output for 7 and 14 MHz.

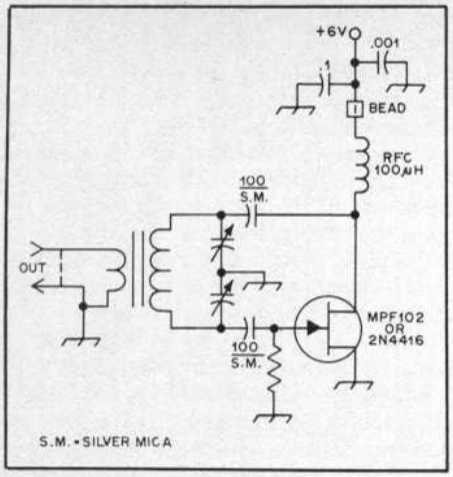


Fig. 67 — Circuit details for a wide-range rf oscillator (see text).

It is not necessary that the units be confined to a single band. One source was built which used a 7-MHz crystal in the circuit shown, but had the tuned circuit peaked at 14 MHz. Both of the outputs were calibrated, resulting in a two-band source.

Crystal-Controlled Sources for IMD Measurements

In the evaluation of the two-tone dynamic range of a receiver, the two parameters needed are the input intercept and the MDS. The MDS can be measured with the weak-signal source just described, and a step attenuator. For evaluation of the input intercept, or for evaluating the dynamic range directly, and then calculating the equivalent input intercept, a pair of stronger sources are needed. The frequencies should be separated by 20 kHz.

A suitable circuit for such sources is shown in Fig. 66. This circuit should be well shielded, although the requirements are certainly not as severe as with the weak-signal source. Of greater significance is that the sources be well decoupled from the power supply lines and that the buffering be effective. In the source shown, output buffering is achieved with a cascode amplifier. This circuit was chosen because of the low feedback capacitance. Because of this, the impedance seen at the input of the buffer is virtually independent of the load or signals present at the output. A dual-gate MOSFET would probably do an excellent job as an output buffer as well, and is certainly capable of delivering 10 mW of output power. In the circuit shown in Fig. 66, R1 is picked for an output power of +10 dBm. A low-pass filter is used in the output to ensure that the power measurements indicate the power available at 14 MHz and not be influenced by harmonic content. Also, harmonics could, in some cases, confuse the IMD results.

The nature of the measurements were described in chapter 6 in connection with our discussion of dynamic range and the intercept concept. Two of the generators of the type shown in Fig. 66 are required, and with equal output powers. The two outputs are added in a

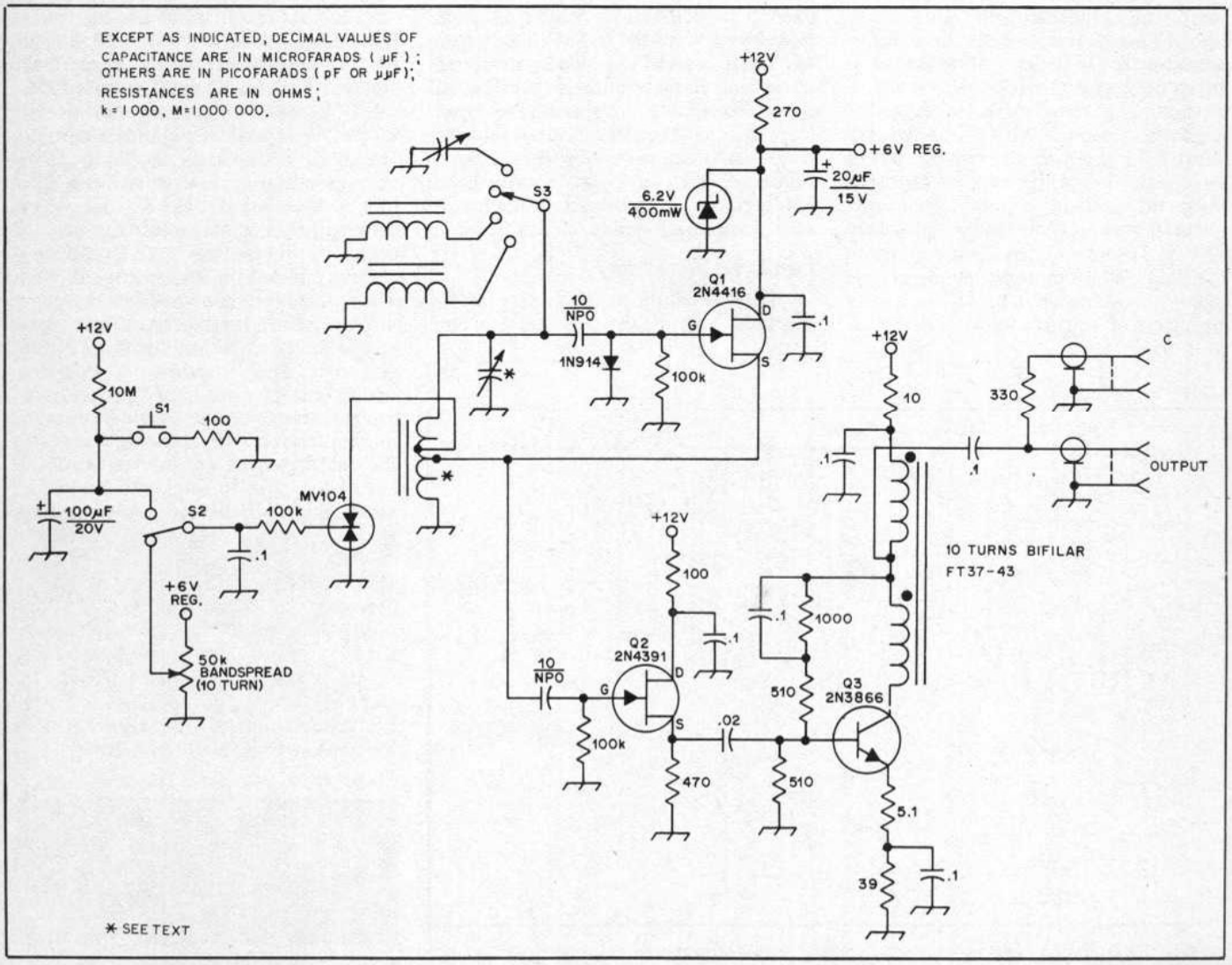
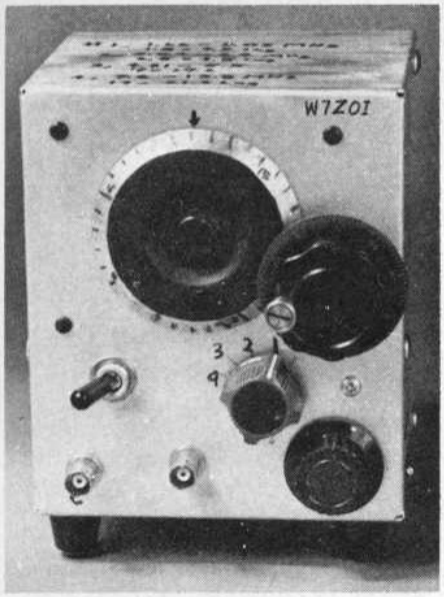


Fig. 68 — An elaborate version of the circuit shown in Fig. 67.



Outside view of the wide-range test oscillator.

6-dB hybrid combiner (described in this chapter as a return-loss bridge). The output of the "hybrid" is applied to a 50-ohm step attenuator and then to the receiver being tested. The reason for using the "hybrid" and the extremes of buffering is to prevent one generator from being phase modulated by the other. This effect is detected easily as a difference in the IMD levels at the two distortion frequencies. A good precaution (besides those outlined) would be to use separate battery packs for power of each of the generators.

Tunable RF Generators

For many measurements a *tunable* source of rf is desired. Applications would range from antenna evaluation and impedance measurement with a return-loss bridge, to measurement of the resonant frequencies of tuned circuits. Shown in Fig. 67 is an FET oscillator that is capable of operation over a wide range of frequencies. The Colpitts configuration is used with a split stator variable capacitor. With most capacitors used for tuning, a frequency range of over 3:1 may be covered with a single toroid coil. If the capacitor has a reasonable low minimum capacitance (10 pF or so, including strays) the oscillator will operate at frequencies up to about 250 MHz. Toroidal coils or air-wound inductors may be used. A 6-volt lantern battery is suitable for power.

This oscillator may be used for evaluating tuned circuits by placing a 6- or 10-dB attenuator at the output. The output of the pad is then applied to a link on the unknown resonator. A sensitive rf detector is loosely coupled to the resonator, and the oscillator is tuned for a peak response.

Band-switching versions of this oscillator may be built. However, it is important that all three of the hot leads of coils be switched. If they are not the stray resonances in the larger coils used for the lower frequencies may cause the output level to vary on the higher ranges.

A more elaborate oscillator is shown in Fig. 68. This unit is band-switched to cover a range of 1.7 to 15 MHz, hitting the four lowest amateur bands. Q1 is a FET that serves as a simple Hartley oscillator. It is normally tuned over the range of 3 to 6.5 MHz, using a single section of a BC-455 surplus receiver capacitor. The oscillator is moved to lower frequencies by switching in the parallel combination of the other two capacitor sections. To move the oscillator to higher frequencies, additional inductors are paralleled with the main one.



Fig. 70 — Exterior view of the test unit. A small Minibox serves as a case.

The gear-drive mechanism built into the surplus tuning capacitor provides more than adequate bandspread. However, for special situations, even finer tuning is desirable. This is realized with a back-to-back pair of varactor diodes. In the circuit shown, a Motorola MV104 dual is used, with both diodes in the same package. The diodes are tapped well down on the tuned circuit in order to provide high tuning resolution. The varactors may be controlled from one of two separate sources, which are selected by a switch. One is a 10-turn 50-kΩ control that is biased with the 6-volt regulated supply used for the oscillator. The other is a swept voltage source consisting of a large electrolytic capacitor, a charging resistor to the 12-volt supply, and a push-button to initiate the sweep. The tuning range of the Varicaps is very restricted, covering only about 7 kHz on the 80-meter band. The main application for this absurd level of resolution was for the evaluation of homemade crystal filters.

Output buffering is handled with a two-stage amplifier. Q2 serves as a source follower to drive Q3 which is a fed-back power stage. A separate attenuated output is provided on the panel of the generator to drive a frequency counter.

Exact component values are not given for the tuned circuits. They will depend upon the parts the builder has on hand. All of the coils are wound on Amidon toroids. The main resonator is wound on a T68-2, with T50-6 cores being used for the high-frequency coils. The tap on the main resonator coil

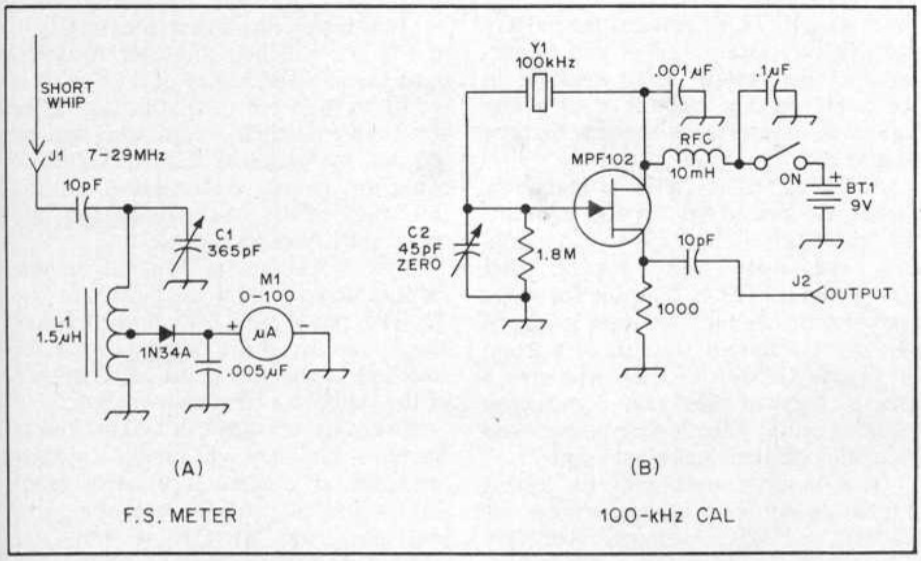


Fig. 69 — The circuit at A is the field-strength meter. L1 has 20 turns of No. 26 enam. wire on an Amidon T50-6 toroid core. The tap is located 5 turns above ground. C1 is a subminiature transistor-radio type of variable. At B is the 100-kHz standard. C2 is a 45-pF mica or ceramic trimmer. Y1 is an International Crystal Mfg. Co. type GP crystal.

should be about 0.25 up from the ground end.

A Handy Field Tester

The matter of including a 100-kHz secondary frequency standard in each receiver built can be costly. A good alternative is to have a separate assembly that can be used with any receiver, thereby reducing the cost which would result from purchasing several crystals. Fig. 69B shows a 100-kHz FET oscillator which operates from 9 volts. A short length of wire can be connected between J2 and the input terminal of the station receiver to provide 100-kHz markers. C2, a 45-pF mica trimmer, is used to zero beat the oscillator with WWV.

Contained in the same 1-1/4 X 2-1/4 X 4-1/4-inch Minibox is the circuit of Fig. 69A. It is a tunable field-strength meter with a range of 7 to 29 MHz. No provisions have been included for calibration of the instrument. It functions only as a relative-indicating meter, but is useful in the field for "sniffing" rf in equipment, and for determining if antennas are functioning properly. It also enables the user to get a reasonable idea of what a near-field antenna pattern looks like. This assembly was built especially for QRP DXpeditions, where lightweight test gear is desirable.

L1 in Fig. 69A consists of 19 turns of No. 24 enameled wire on an Amidon T50-6 toroid core. The diode tap is placed 5 turns up from the ground end of the coil. This prevents the rectifier diode from loading the tuned circuit. A short piece of hookup wire, or a whip made from brazing rod or piano wire, is inserted into J1 for sampling rf. C1 is tuned for a peak response at the operating frequency, as indicated at M1. C1 is a miniature 365-pF variable of the variety used in transistorized a-m band receivers.

Y1 of Fig. 69B is an International Crystal Co. unit of the general-purpose type. Load capacitance is 30 pF. Any crystal with similar characteristics should work satisfactorily in the circuit. Fig. 70 shows an exterior view of the assembled tester.

Transistor and Crystal Testers

Fig. 71 contains the circuit of a "go-no-go" type of transistor tester which can be used to determine whether transistors are defective, npn or pnp varieties, or FETs. A fundamental type of crystal is used at 20 MHz to permit the devices under test to function as oscillators. Output from the oscillator is rectified by a voltage doubler (CR1 and CR2). The dc voltage is routed to a 50-μA meter, M1, to provide a visual indication of performance. S3 is used to apply forward bias to bipolar transistors. It is switched to the open position

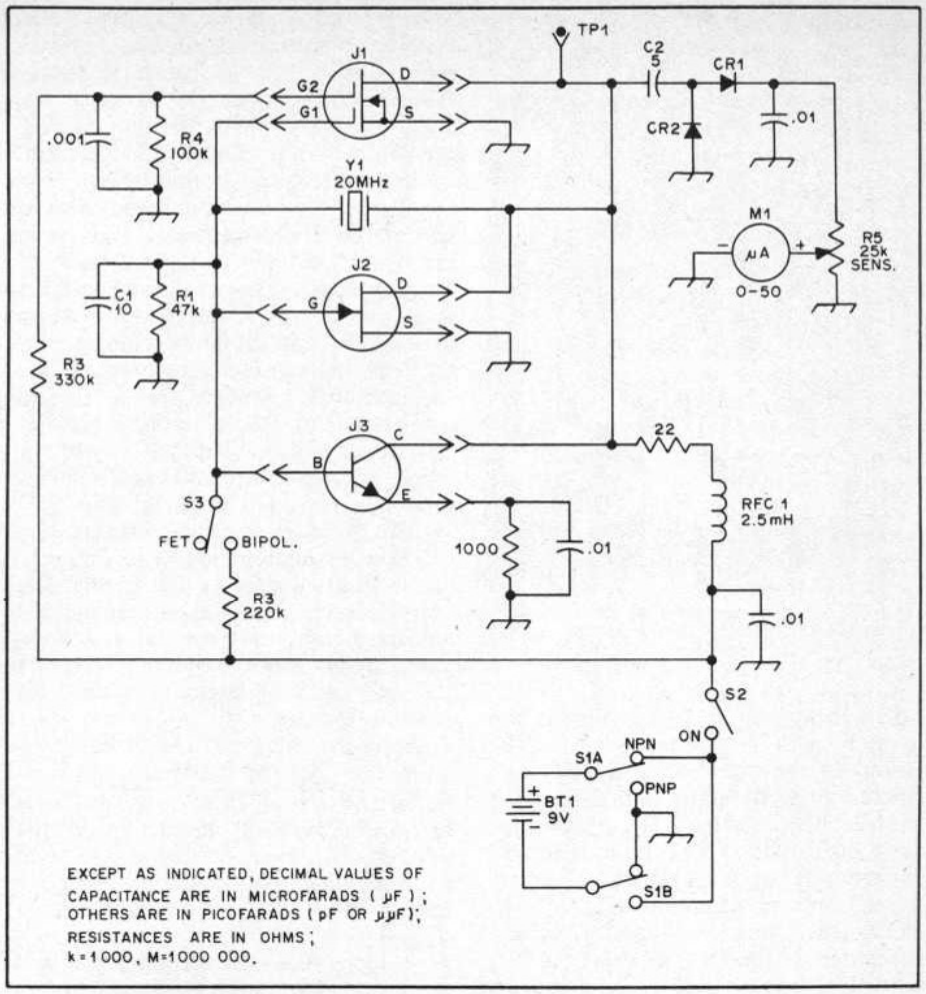


Fig. 71 — Diagram of the FET and bipolar-transistor tester. Resistors are 1/4- or 1/2-W composition. Capacitors are disk ceramic.
BT1 — Small 9-V transistor radio battery.
CR1, CR2 — 1N34A germanium diode or equiv.
J1 — Four-terminal transistor socket.
J2, J3 — Three-terminal transistor socket.
M1 — Microampere meter. Catectro D1 910 or similar.
R5 — 25,000-ohm linear-taper composition control with switch.
RFC1 — 2.5-mH rf choke.
S1 — Two-pole double-throw miniature toggle.
S2 — Part of R5.
S3 — Spst miniature toggle.
Y1 — surplus crystal.

for testing FETs. S1 reverses the battery polarity for testing npn or pnp transistors. At the voltage levels available in the tester, damage will not occur to any transistor, regardless of the positions of S1 and S3.

Three different styles of transistor socket are placed on the top panel of the tester (J1, J2 and J3) to accommodate the three most popular lead arrangements. TP1 is available for scope attachment, should the user wish to measure the output voltages of a group of similar transistors. This will give a general idea of the gain comparison between units — the higher pk-pk levels indicating greater small-signal gain.

This tester is useful only for testing transistors whose f_T characteristics are 50 MHz or higher. Although most transistors will function as oscillators at some frequency lower than the rated f_T, the test results with the circuit of Fig. 71 will not be of value.

The tester illustrated schematically in Fig. 72 will help the user to determine the relative quality of crystals. It is set up as a Pierce oscillator, and three fixed-value feedback capacitors can be selected by means of S1. The feedback capacitor chosen will depend on the frequency of the crystal under test, and on its activity characteristic.

Visual readout is handled in the manner described for the circuit of Fig. 71. TP1 can be used for connection to a scope, or a short antenna can be attached to the test point to permit use of the tester as a frequency marker.

Overtone crystals can be checked in this unit, but they will oscillate at their fundamental modes. A polarity reversing switch, S3, permits use of npn or pnp transistors at Q1. A transistor socket is located on the top panel of the tester, thereby making the tester useful for checking transistors of unknown characteristics. J1 through J4, inclusive,

are crystal sockets with different hole sizes and spacings. This feature makes the unit more versatile with respect to checking crystals in various holder styles. Both testers are housed in homemade aluminum cases. Fig. 73 shows how the testers are laid out.

Timing and Control Circuits

There are a number of places in the design of amateur equipment where timing circuits must be used. These include circuits for the control of transmitters, receivers or transceivers, audio side-tone oscillators, antenna switching circuits, sweep and control systems for SSTV, and even systems for the control of repeaters. There are literally dozens of ways to design these circuits. Some samples are presented in this section.

Sidetone Oscillators

One need during the transmission of cw is that the operator have a means for monitoring his fist. One method is to listen to the transmitted signal in the station receiver. It allows the operator to know the frequency that he is transmitting on, if he is using a separate transmitter and receiver. However, it places some constraints upon the receiver. The muting system must allow the receiver gain to be reduced by 80 to 100 dB, while still delivering a clean tone. Alternatively, the operator must be willing to accept receiver deficiencies, such as clicks generated within the receiver, and even a possible frequency shift in the receiver local oscillator, caused by strong rf fields.

A superior approach to cw monitoring is to use a sidetone oscillator. This is an audio oscillator that is keyed simultaneously with the transmitter. It may be activated by rf detection, by the dc voltage changes that occur within the transmitter, or by a signal from an electronic keyer. Many electronic keyer circuits have sidetone oscillators built into them, along with small speakers. The writers prefer systems that inject the sidetone signal directly into the audio chain of the receiver. This is more compatible with headphone operation.

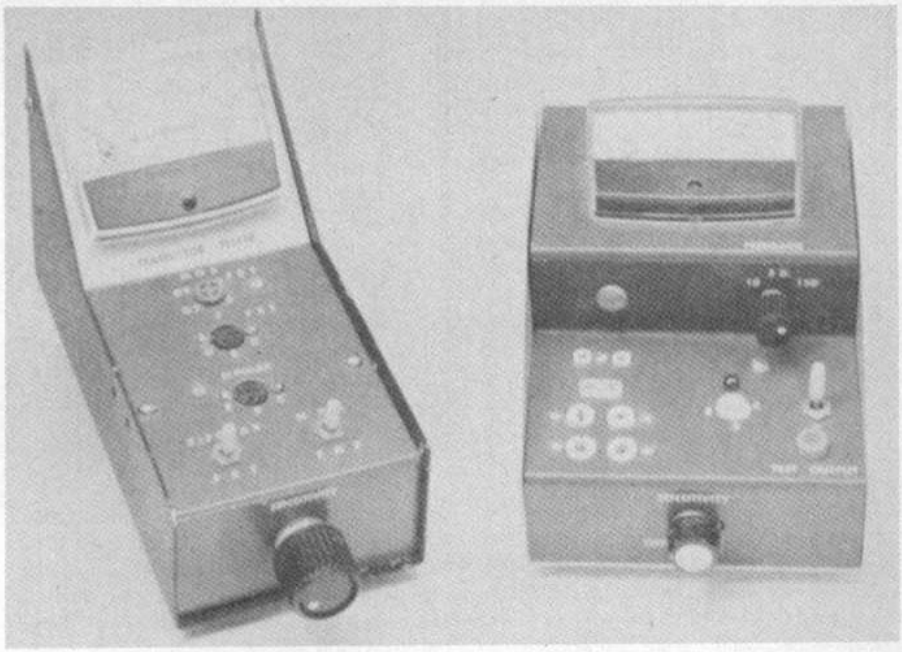


Fig. 73 — Photograph of the two testers. They are housed in homemade aluminum cases. The unit at the left is the FET and bipolar-transistor tester. At the right is seen the crystal and bipolar-transistor checker. Various sizes of crystal sockets are installed in order to accommodate the popular pin sizes and spacings.

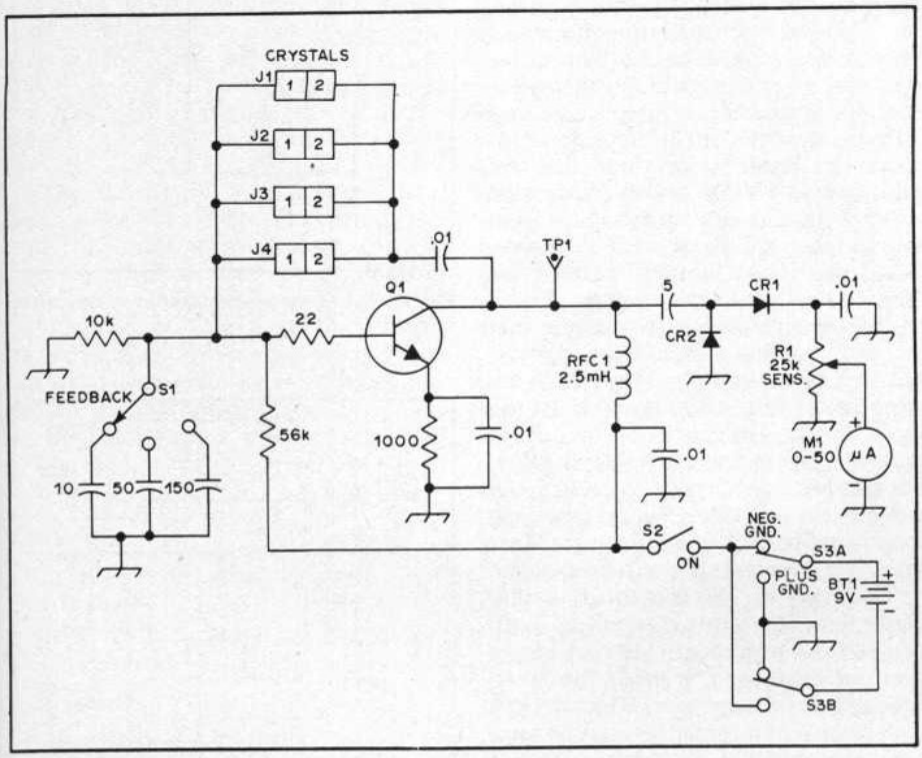


Fig. 72 — Schematic diagram of the crystal and bipolar-transistor tester. S2 is part of R1. Q1 is a vhf or uhf npn transistor (2N2222A or equivalent). CR1 and CR2 are 1N34A diodes. S1 is a single-pole three-position phenolic rotary wafer switch.

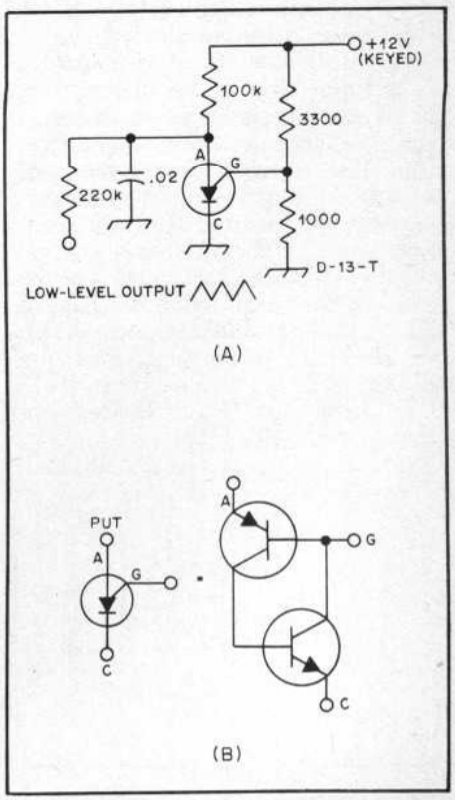


Fig. 74 — Circuit of a PUT audio oscillator.

Shown in Fig. 74A is an audio oscillator using a General Electric D-13T programmable unijunction transistor (PUT). The output frequency is about 1 kHz, and may be changed by replacing the capacitor value shown. The output is at low levels, suitable for injection into the input of a medium-gain audio amplifier.

The PUT is a device, similar to a silicon-controlled rectifier (SCR), that can be used for many applications. Shown in Fig. 74B is a model for a PUT. The three-terminal device may be thought of as being composed of a combination of an npn and a pnp transistor in the form shown. While this circuit is used primarily as a model to explain the operation of PUTs, one can also use this circuit in order to build PUT circuits when suitable devices are not available. Good choices for the devices are the 2N3904 and 2N3906 for the npn and pnp, respectively.

Shown in Fig. 75 is another sidetone oscillator consisting of a free-running multivibrator which uses two bipolar transistors. This circuit will operate with virtually any common silicon transistor type, and does a good job of generating a sidetone. The output is a square wave at approximately 1 kHz. The diodes in the base are necessary in order to prevent damage to the emitter-base junctions of the transistors from breakdown. If the oscillator is run from lower voltage supplies (5 volts or less on the collector resistors) the diodes may be eliminated.

A variation of this circuit is shown in Fig. 76. The circuit uses a voltage that is derived from the rf output of a QRP transmitter in order to provide part of the operating voltage for the circuit. This circuit has the characteristic that when the transmitter is keyed, the output tone occurs. This tells the operator that the transmitter is delivering rf. Moreover, the pitch of the oscillator is proportional to the rf output voltage. This means that the transmitter may be tuned in the field without having a meter built into the equipment. This can be handy when ruggedness and minimal weight are design criteria.

Shown in Figs. 77 and 78 are a pair of oscillators using 741 operational amplifiers. The circuit shown in Fig. 77 is an oscillating variation of a type of low-pass audio filter. If the resistor values were chosen carefully, it might be possible to obtain a fairly clean sine wave from the circuit, although it might then be sluggish in starting. The circuit of Fig. 78 utilized the 741 as a differential comparitor with positive feedback. It is generally more predictable than the other circuit.

The oscillator of Fig. 79 uses a 555 timer IC. These ICs are useful in timing applications, and will be discussed later.

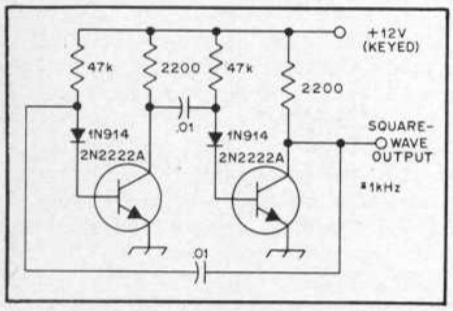


Fig. 75 — Sidetone oscillator using a multivibrator.

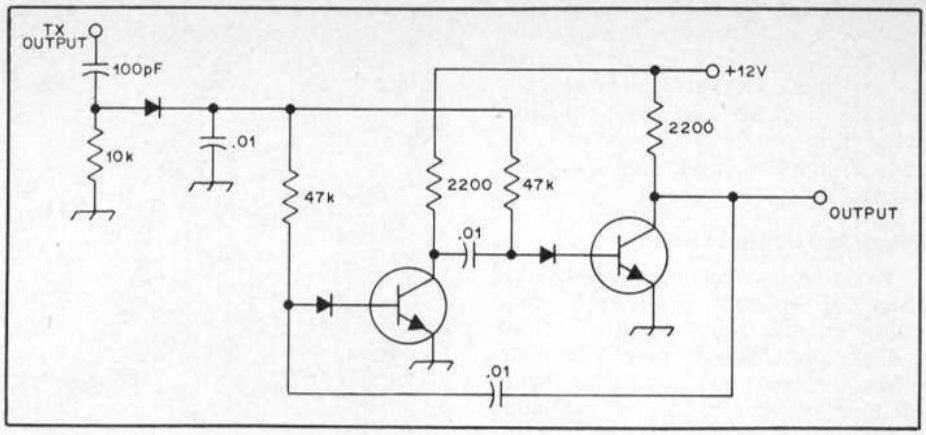


Fig. 76 — Another type of sidetone oscillator (see text).

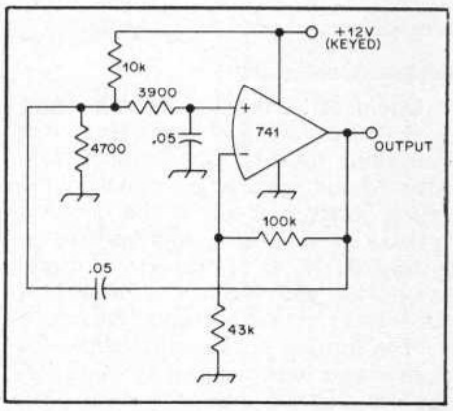


Fig. 77 — An op-amp sidetone oscillator.

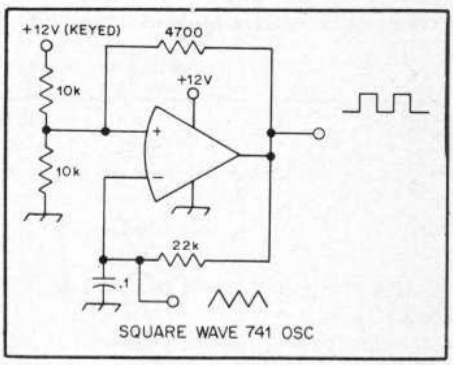


Fig. 78 — An improved version of the circuit in Fig. 77.

T-R Relay-Control Systems

In the construction of cw and ssb transmitters (or transceivers), one useful accessory is a key (or VOX) controlled transmit-receive system. In cw service, this means that when the key is pressed, the relay used for transferring the antenna from the receiver to the transmitter is activated automatically. Furthermore, the transmitter circuit is activated and the receiver is muted. In the case of ssb operation, these same functions are realized with a VOX, or "voice-operated switch." In this case, some audio from the speech amplifier is rectified to provide a dc voltage that will activate the relay-control circuitry. In either the cw or the ssb situation, it is desirable that the antenna changeover occur quickly, and that after the key is released, or the voice ceases, the relay stay closed for a short period. The length of the hold-in time will depend upon the application. For contest work, periods around 0.5 second are suitable. For ragchewing, longer periods may be desired. More than 1 to 2 seconds is generally avoided.

Shown in Fig. 80 is a circuit used in many stations. This system is compatible with a keying mode that keys a positive voltage to ground, the usual case with solid-state gear. When the key is depressed, the pnp transistor is saturated immediately. This sends current through the base of the npn, which activates the relay. The usual practice is

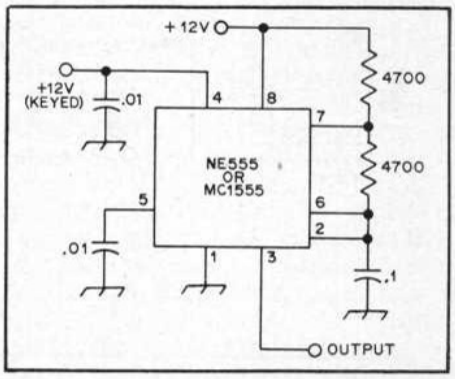


Fig. 79 — An audio oscillator which employs an NE555 timer IC.

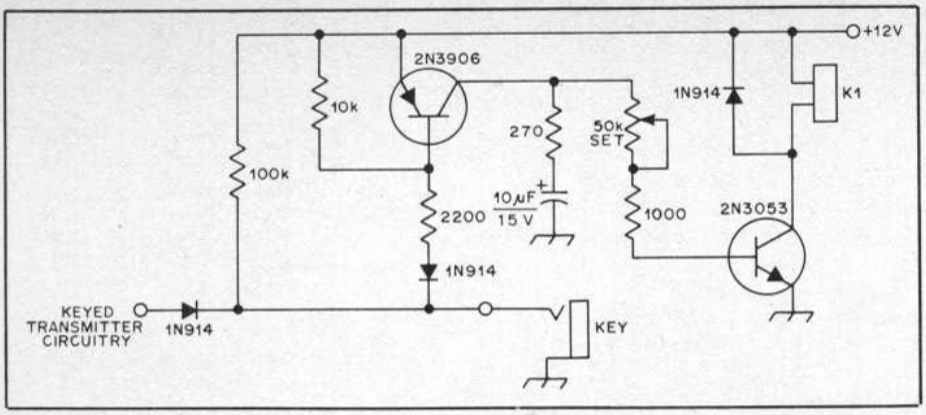


Fig. 80 — A relay-driver circuit for T-R applications.

to use a multipole relay. One set of contacts transfers the antenna while the remaining contacts apply dc voltages to the transmitter circuits. When the pnp transistor comes on, part of the output current flows into the capacitor through the 270-ohm resistor. This causes the capacitor to charge quickly to +12 volts. When the key is released, the pnp device is immediately cut off. However, the timing capacitor is now charged to a high potential. The capacitor will discharge through the potentiometer, determining the hold-in time. A diode is placed across the relay coil. It protects the npn transistor. If the diode was not there, a high positive collector-voltage spike would occur at the instant the relay turned off. Depending upon the inductance and resistance of the relay coil, and the stray capacitances, this potential could reach several hundred volts. The diode clamps this positive voltage spike to the positive power supply line. The current that flows in the diode will have the effect of extending the hold-in time of the circuit slightly.

The circuit of Fig. 80 has some deficiencies. The main one is that the capacitor must be almost completely discharged before the relay will drop out. The exact time of relay dropout will depend on the beta of the npn transistor. Beta variations among transistors of a given type are often large, and may be temperature-dependent.

The deficiencies outlined may be circumvented by using more precise timing circuits. One of the easier approaches to such design is through the use of a differential comparitor circuit. Such a circuit is shown using a 741 op-amp in Fig. 81. V_r is a reference voltage that is derived from the power supply through a voltage divider. Typically V_r will be about 0.5 V_{cc}. The input voltage of the comparitor is increased from zero toward the positive supply. As it approaches the reference voltage, the output of the 741 will start to increase. Since the dc gain of the 741 is high, the transistion from the low to the high state will occur over a range of input voltage of a millivolt, or thereabouts. A curve of this response is presented in Fig. 81.

In the example shown, the reference voltage is applied to the inverting input of the op amp while the control voltage is placed on the noninverting input. If the reverse circuit was used with the input signal applied to the noninverting input, the output would be high for low values of input. With 741 op amps the output voltage will approach the positive supply within a volt or two. Similarly, in the low state, the output can drop down to about 2 volts. The characteristic that the output does not come closer to ground is sometimes a problem that makes additional components necessary. Some of the newer op amps will allow their outputs to approach the supply voltages more closely. An excellent choice for circuits of this kind would be the LM-324, which is a quad op amp (four op amps in a single package, each with characteristics similar to the 741).

A simple T-R control system using a 741 as a differential comparitor is shown schematically in Fig. 82. The reference is obtained from a divider, and hold-in time is determined by the 100-kΩ and the 5-μF capacitor. A 5-volt Zener diode in the output of the op amp assures an output that drops to ground potential. If one section of an LM-324 was used, the Zener diode could be eliminated. A multipole relay is used with one set of contacts shifting the voltages to the transmitter, as required. This ensures that the transmitter does not come on until the antenna is connected to the transmitter.

It is important in many cases that the antenna relay be in the transmit position before rf is applied. If this is not done, the relay is required to switch when large rf voltages are present. This places severe requirements on the relay. Furthermore, the transmitter final amplifier may be operating for a short period with *no* termination on the output. This can lead to instabilities and can, in some cases, destroy the transmitter output transistor. Receiver front-end damage is also common.

Shown in Fig. 83 is a modified system that is designed to circumvent these problems. The main relay control circuit is identical with that shown in the previous schematic. However, when the key is depressed and the output of U1 goes high, a current will flow through the 220-kΩ resistor that connects to U2. The 0.1-μF capacitor at the input to U2 will cause a delay of about 20 milliseconds before the output of U2 goes high. The high output at U2 can be used to turn on a switch (Q1) that grounds the oscillator control. Alternatively, the output of the switch can be used to control a pnp switch (Q2) that applies a positive voltage to the oscillator in the transmitter.

At the end of a timing cycle when U1 returns to an off condition, the oscillator voltage is terminated quickly. This is realized with the diode across the 220-kΩ resistor. Receiver muting signals should be derived from the output of U1. This will ensure that the receiver is muted *before* any rf is generated.

The 20-ms delay introduced by the U2 timing circuit presents a minor problem: The first dot of a cw transmission is a bit shorter than the signal generated by the key. This problem is

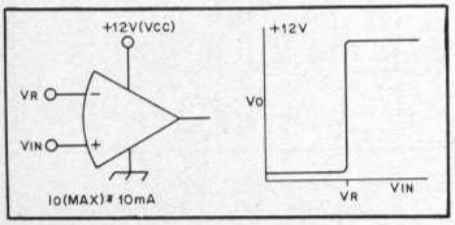


Fig. 81 — A differential comparator which uses a 741 op amp.

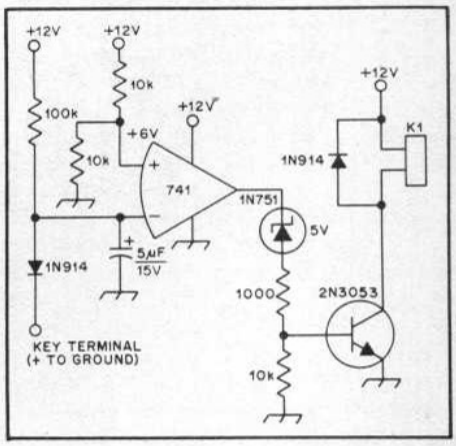


Fig. 82 — A T-R circuit which uses an op-amp differential comparator.

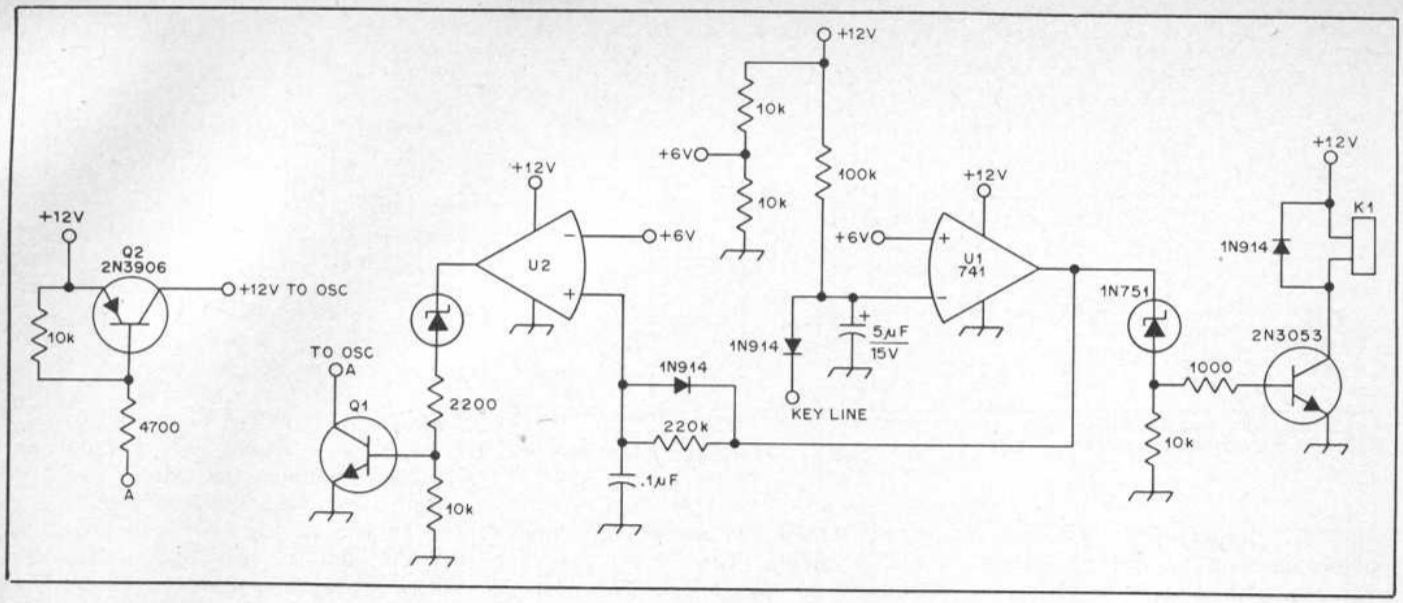


Fig. 83 – Recommended T-R circuit (see text explanation).

not severe, however, since the length of a dot at 20 wpm is about 50 ms. It is better to suffer this slight inconvenience than it is to burn out a final amplifier, or to create a tremendous key click on the air when the relay switches while "hot" with rf. This characteristic is noticeable with some commercial transmitters. The 20-ms period was chosen because most relays take approximately 10 ms to pull in. This includes dc-controlled coaxial relays. The cautious experimenter should measure the pull-in characteristics of his relay with a triggered oscilloscope, then tailor the time constants accordingly.

While op-amp ICs have been used in the previous circuits, they are not the only way to handle the relay driver problem. Shown in Fig. 84 is a simple comparator type of switching scheme that offers good timing accuracy. This circuit uses two transistors and a Zener diode as the main elements.

Often it is desired to run an outboard power amplifier as an accessory to a low-power transmitter. The best way to switch the antenna would be to run appropriate dc control voltages to the outboard final. However, this would make the accessory less convenient to use. An alternative approach is that of using detected rf energy from the exciter to control a suitable relay in the outboard amplifier. Shown in Fig. 85 is a circuit that was developed for this purpose. The user should be sure his exciter is capable of operation (temporarily) without a load without self-destruction. Ideally, a set of contacts on the relay would be used to apply dc voltage to the outboard amplifier.

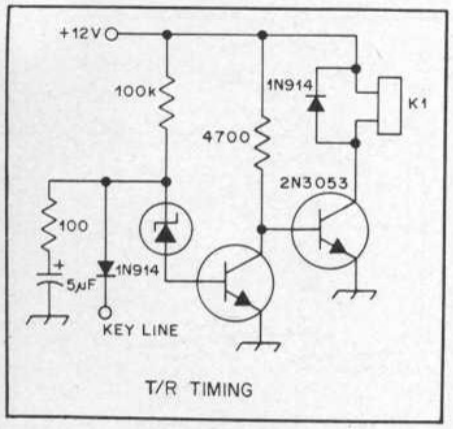


Fig. 84 – Example of a relay driver which uses two transistors and a Zener diode as the main elements.

Circuit Description

R1 of Fig. 85 serves as an rf voltage divider to permit the circuit to be used with transmitters of various power-output amounts. Rf energy is routed through C1 to the base of broadband amplifier Q1. The amplified hf-band energy is supplied to a voltage-doubler (CR1 and CR2) through a broadband toroidal step-down transformer, T1. The rectified rf voltage at the output of CR1 and CR2 is filtered by means of RFC2.

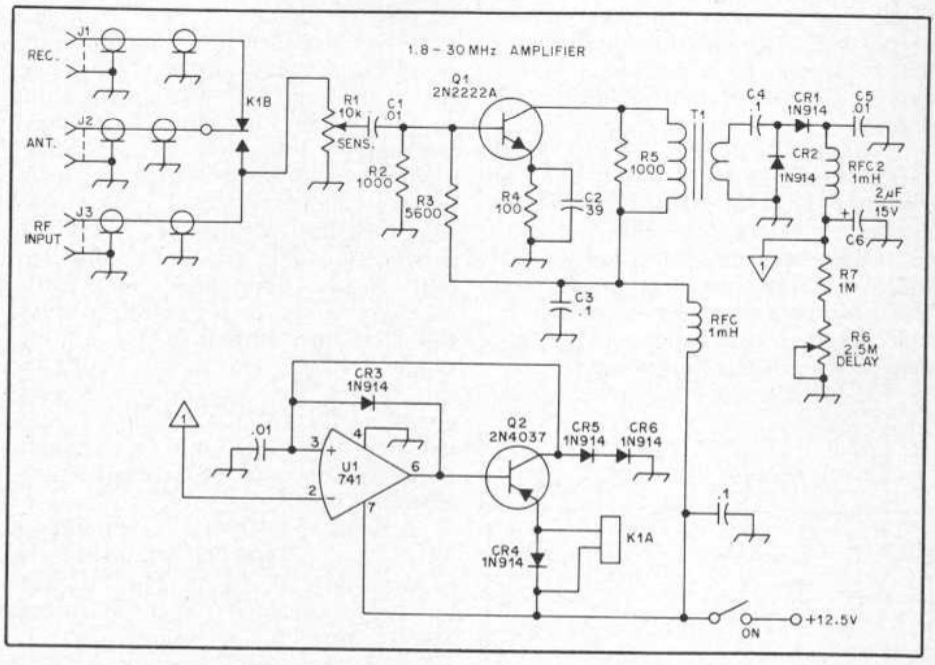


Fig. 85 – Circuit of an rf-actuated relay driver. This unit was first described in *QST* for Aug., 1976, p. 21, inclusive of a pc-board layout. K1 is a 12-V relay with a field coil dc resistance between 500 to 1,000 ohms. T1 primary has 25 turns of No. 28 enam. wire on an Amidon FT-50-43 toroid core. The secondary consists of 5 turns of No. 28 wire wound over the primary. RFC1 and RFC2 contain 42 turns of No. 28 enam. wire on FT-50-43 toroid cores.

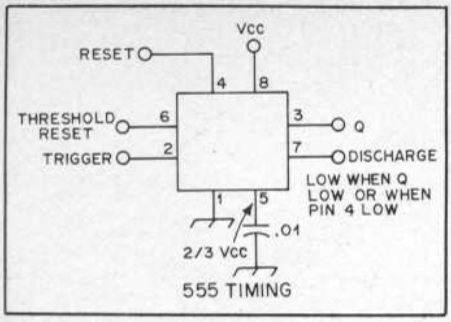


Fig. 86 — Block presentation of an NE555 timer IC.

C5, and C6. This prevents unwanted rf from reaching U1 and affecting its performance.

C6, R7, and R6 comprise a timing network (variable) which governs the hold-in time of the relay, K1. The smaller the resistance amount at R6, the shorter will be the time delay.

U1 functions as an inverting amplifier. When the input dc voltage at pin 2 increases, the output dc voltage at pin 6 decreases. The output voltage causes the base of relay driver Q2 to be forward biased negatively when it drops below approximately 1.4 volts. Diodes CR5 and CR6, by virtue of their combined barrier voltages (0.7 V each), established the 1.4-V fixed bias level. Without the diodes, Q2 would conduct sufficiently to prevent the relay from dropping out during no-signal periods, CR4 is used to suppress transients caused by the field coil of K1. When no rectified rf reaches U1, Q2 is cut off because of the high positive base voltage it receives from U1, and the relay contacts to the transmitter are open.

The NE555 Timer

An IC that is useful for timer applications is the NE-555. Several companies manufacture versions of this chip. The Motorola part number is the MC-1555. The principles that are applied in this chip are similar to those described. The chip contains a set-reset flip-flop (RSFF), an output buffer that will supply or sink up to 200 mA of current, two differential comparitors for control of the RSFF, as well as some other control functions. The typical package is an 8-pin mini-DIP. The circuit also has a built-in resistive divider that provides two reference voltages at 1/3 and 2/3 of the supply voltage. The chip will operate with supplies from 5 to 18 volts.

Shown in Fig. 86 is a block presentation of the 555 timer chip. The output appears at pin 3. Pin 7 can also be used as an output. It is an open collector of a transistor with a grounded emitter. Under most conditions this transistor is in an "on" condition when the output, Q, at pin 3, is low. The chip is triggered into an on condition (Q high) by pulling pin 2 below 1/3 of the supply voltage. Pin 4, which is labeled in the literature as a reset, serves the function of turning on the transistor with output at pin 7. If this reset is not to be used, it should be tied to the positive supply. The other

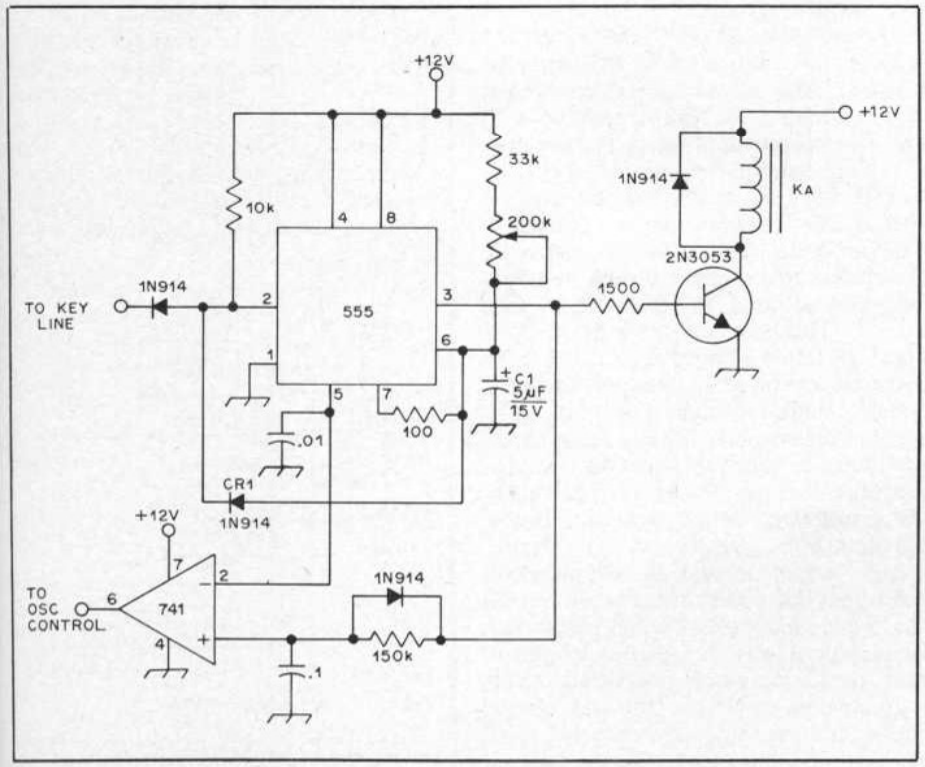


Fig. 87 — Break-in delay circuit which uses an NE555 timer IC.

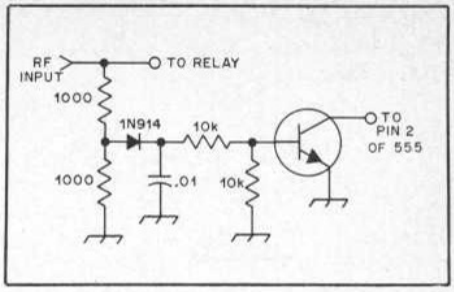


Fig. 88 — An rf-derived circuit for operating the system shown in Fig. 87.

reset, which is labeled "threshold," resets the RSFF when the terminal becomes more positive than 2/3 of the supply voltage. Pin 5 is the 2/3 V_{cc} reference voltage and should normally be bypassed for high frequencies. In situations where several 555 timers are linked for complex timing functions, all of these reference voltages may be tied together to ensure accuracy.

Shown in Fig. 87 is a break-in delay circuit using the 555 timer. Under normal key-up conditions, the FF will have been reset and Q (pin 3) will be low. When the key is pressed, the RSFF is set into a high condition. This results from pin 2 going low. The circuit is inhibited from "timing out" by the clamping action of CR1. If this diode were not present, timing would begin as soon as the RSFF was set. The timing is prevented so long as the key is depressed. When the key is lifted, the timing capacitor, C1, begins to charge. If the key is depressed before the timing has finished, the capacitor is discharged through the key and CR1. If the key is left open for a period of time, C1 will eventually charge to 2/3 V_{cc}. This action applied to the threshold terminal (pin 6) causes the flip-flop to be reset.

While the circuit may be used as described for simple relay control, a simple modification may be made to obtain a delayed control for the transmitter. This is the circuit associated with the 741 op-amp. The internal reference of the 555 timer is used as the reference for the 741 comparitor. The delay operation is virtually identical with that described for Fig. 83.

This circuit (Fig. 87) may be modified easily to operate from an rf-derived signal for use with an outboard amplifier. This application is shown in Fig. 88, and the relay is now used in a manner identical to that of Fig. 85.

A somewhat more complex application of the 555 timer is the electronic keyer shown in Fig. 89. This circuit is straightforward, as keyers go, and the performance is good. It has advantages over some of the circuits that are popular. One is that when a character is started (a dot or a dash), no more information may be entered into the

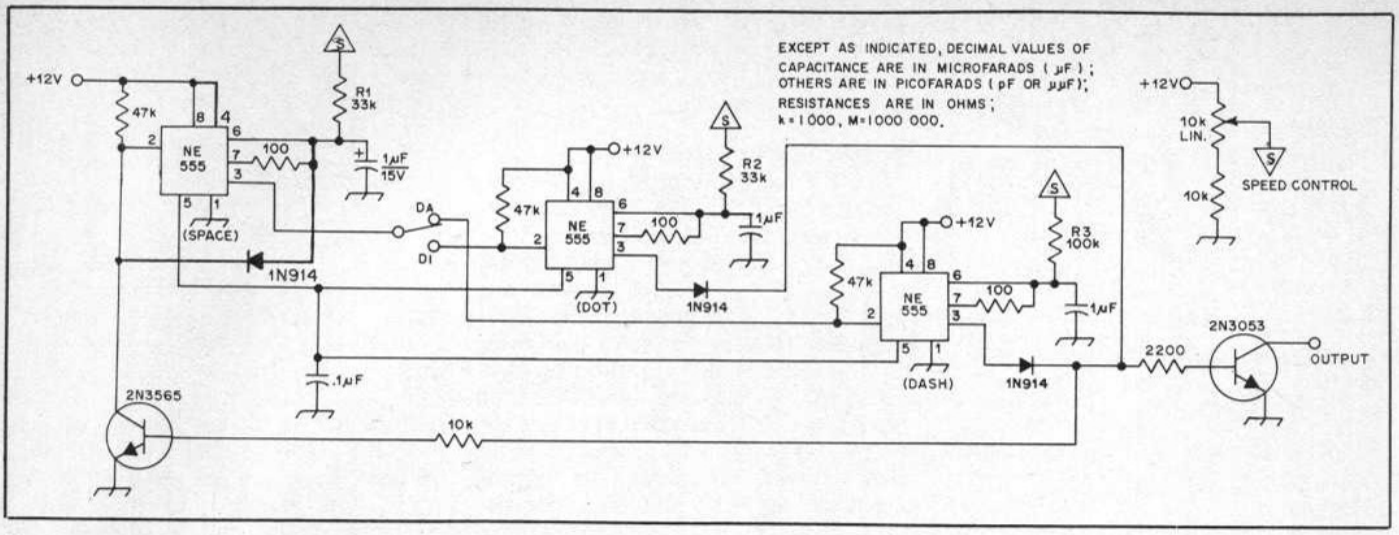


Fig. 89 — An electronic keyer which employs three NE555 timer ICs.

circuit, irrespective of paddle position, until the end of the following space. In many circuits it is necessary that the user be "off the paddle" before the end of the dot or dash. Otherwise, another character will be generated. Another advantage of this circuit is that the capacitors start a timing cycle in a completely discharged condition. Because of this, it is not necessary to discharge the capacitors quickly through the paddle and additional circuits. This phenomenon led to timing errors when poor quality components were used in an earlier circuit described by one of the writers (*QST* for Nov., 1971). A final advantage of this circuit is that all three of the functions (dot, dash, and space) are timed with separate circuits. As a result, the timing resistors (R1, R2, and R3) may be changed in order to adapt to any individual taste.

The purpose of the foregoing keyer description was to demonstrate the versatility of the 555 timer. While the keyer functions quite nicely and is presently in use, there are dozens of keyer circuits available that will function as well. Undoubtedly, the optimum route to follow in such a design would be to use CMOS ICs. The power consumption is very low with such devices, and so is the cost.

The writers have taken a slightly different approach to the keyer design problem than is perhaps typical. The usual approach is very pragmatic, that of finding a design of an acceptable level of complexity that will provide the best performance available. On the other hand, keyers offer another profound advantage from an educational point of view. The function that is to be designed is fairly straightforward, and yet certainly not trivial. As a result, designing keyers is an excellent mechanism for learning about new circuit techniques. Even if the circuits are never built, it can be enlightening to go through the learning exercise of designing them.

Electronic T-R Switching

All of the techniques outlined above for T-R switching have utilized a relay. However, the function can be handled completely with electronic switching methods. There are two general approaches to the electronic T-R switch problem. The one that is most common is one of attaching the antenna directly to the transmitter. Then, the receiver is paralleled with the transmitter output with suitable circuitry to prevent damage to the receiver during transmit periods. The other approach is to actually switch the transmitted power directly. Clearly, this is the more difficult of the two.

For most work on the hf bands the simpler method of T-R switching is suitable. The advantage of electronic T-R switching is that it allows full break-in operation on cw, a feature that is convenient for the contest operator, traffic handler or vhf meteor-scatter enthusiast. There are some constraints that must be applied to the design of the system. First, there should be minimal degradation in receiver performance. This can originate from two considerations. One is that distortion products can be generated in the switch, which would degrade the dynamic range. Furthermore, losses in the switch can degrade receiver noise figure. The second and major source of problems is the transmitter output network. If the signal is not extracted from the transmitter output in the proper manner, there may be significant attenuation of the signal. Examples of this effect will be presented later. Another constraint is that the T-R switch not create extra harmonic output from the transmitter that could cause interference to other services. This problem is usually handled easily.

Shown in Fig. 90A is a simple T-R switch. The receiver is attached to the collector terminal of the matching pi network. In this example for 7 MHz, a 50-pF capacitor is used for coupling into the low-impedance port of the receiver. The receiver is protected from strong rf signals by the back-to-back

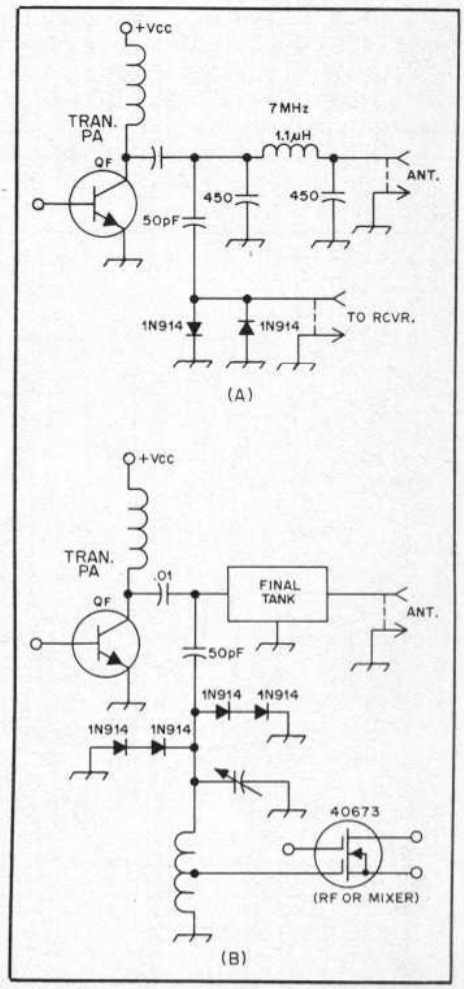


Fig. 90 — Circuit of a simple T-R switch.

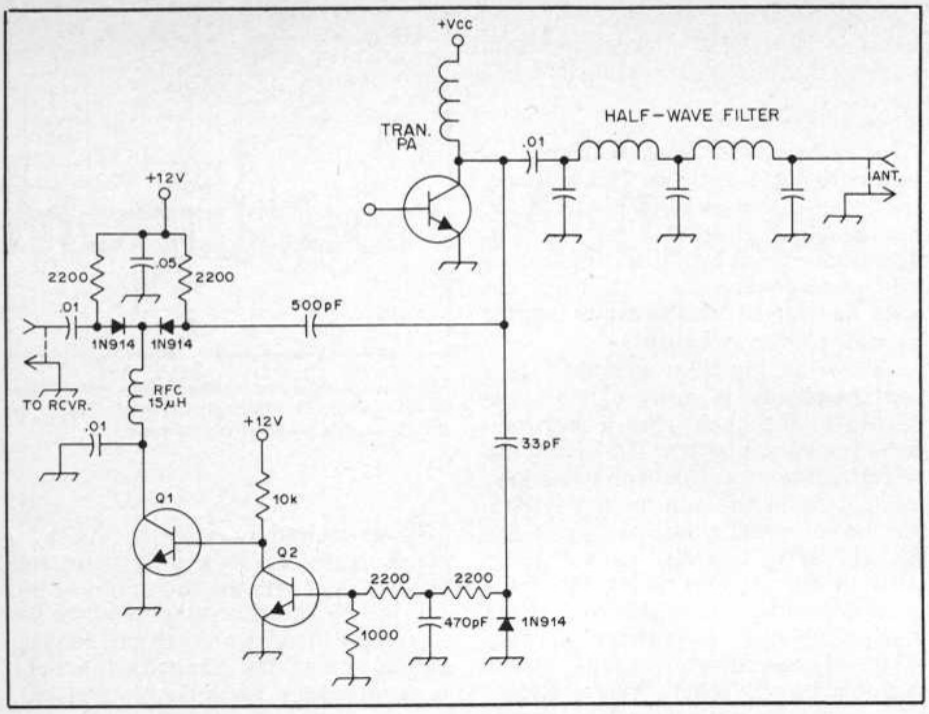


Fig. 91 — T-R circuit in which diodes are used in series.

silicon switching diodes. The 50-pF capacitor will become part of the transmitter tank and reasonable rf currents will flow here. The diodes must be capable of handling this current. Because of the reactance in the 50-pF capacitor (about 500 ohms at 7 MHz), there may be some attenuation of signals. This may be avoided with the system shown in Fig. 90B. Again, back-to-back diodes are used. However, there are now *two* diodes per leg. The coupling is into a higher impedance point in the receiver input. Because of this, the switch presents virtually no loss during receive periods.

There are two major observations that should be mentioned about the circuits described. First, measurements indicate that the back-to-back diodes do not cause IMD at the receiver input as long as the signal is not large enough to turn the diodes on. In a 50-ohm system, levels at the antenna terminal of up to nearly a volt pk-pk could probably be tolerated without compromise in dynamic range. The second item of significance is the point of attachment to the transmitter. The antenna appears as a 50-ohm load at the frequency of interest, presumably. On the Q_f side of the pi network, a 50-ohm load is also seen. However, if the receiver was attached on the antenna side of the pi network, the receiver would see the 50-ohm resistance of the antenna in parallel with a series-tuned circuit at the operating frequency. This series-tuned circuit could lead to significant attenuation. This effect would not be quite as pronounced in the system of Fig. 90B, for some of the reactive effects of the series-tuned circuit could probably be tuned out by readjustment of the receiver input tuned circuit. In either case, the diodes will create some harmonic currents. It is best that the low-pass filtering action of the pi network be used to ensure that the harmonics never reach the antenna, where they may be radiated.

In all of the schemes described, *silicon* switching diodes should be used. The relatively low turn-on voltage of germanium or hot-carrier diodes would cause them to create IMD in the received path.

The examples of Fig. 90 used silicon diodes as shunt clamp elements. The diodes can also be used in a series configuration. An example is shown in Fig. 91 where a pair of back-to-back series diodes are biased on to a current of about 6 mA in each diode. When rf is generated by the transmitter, some of the output is sampled and rectified. The resulting dc is used to saturate Q2 which has the effect of turning Q1 off. In one system of this kind that was investigated with a 2-watt QRP transmitter, it was found that the receiving insertion loss of the switch was about 1 dB, completely insignificant at 7 MHz. The attenuation of rf from the transmitter was over 40 dB, which was enough to prevent damage to the receiver front end. Because the receiver being used had a fast, wide-range agc system, complete break-in operation was possible with no clicks or thumps. It's an unusual experience to hear signals between the dots in a 30-word-per-minute string. A system of T-R switching of this kind is used in a superhet transceiver described later in the book. That system was not set up for QSK operation, however.

Measurements have not been performed to evaluate the IMD levels created by the series-biased diodes. They would probably not be detectable unless the receiver had an input intercept greater than 0 dBm.

In many cases it may be desirable to provide additional protection at the input of a receiver from the effects of a transmitter. A system shown in Fig. 92 will do this. It is assumed that some sort of a control voltage is available, providing +12 volts when the transmitter is on. This signal is used to bias a switching diode on to about 6 mA. Since this diode is across the hot end of the tuned circuit (a high-impedance point), it must be reverse biased during receive periods. The additional diodes are used to prevent damage to the FET during switching transients. They may not be necessary. The user should not rely upon the Zener diodes that are built into the MOSFET front end of the receiver. These diodes are typically very small and will only handle small currents before burnout. They are useful mainly

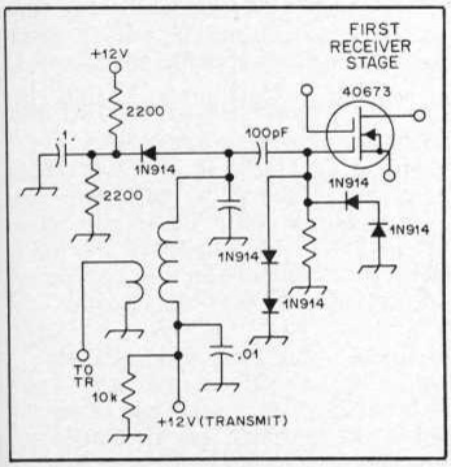


Fig. 92 — Receiver input protection circuit which uses silicon diodes.

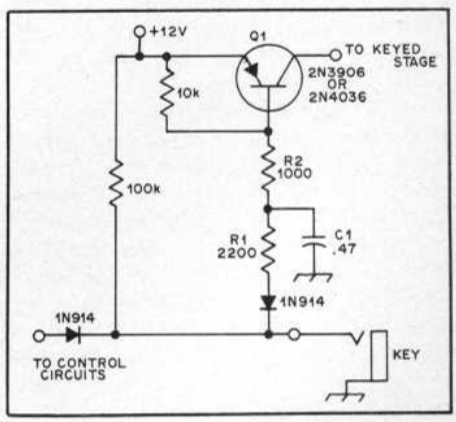


Fig. 93 — Circuit for using a pnp transistor switch for shaped keying of a stage in a transmitter.

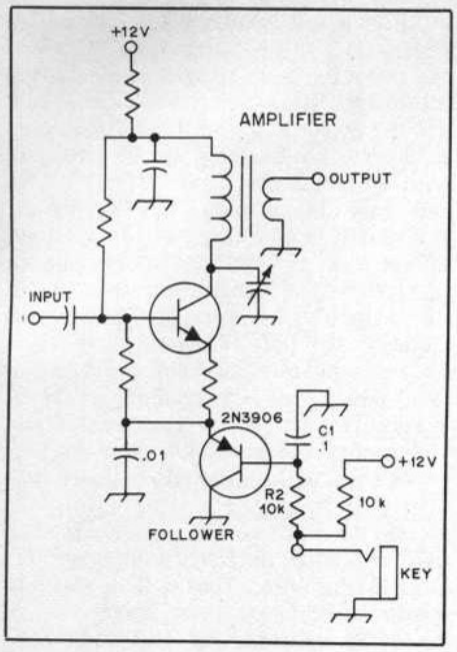


Fig. 94 — A pnp keying transistor functioning as an emitter follower.

for the protection of the MOSFET from damage during handling.

While additional measurements are required, it appears that methods of the kind shown offer great promise for the QSK enthusiast. The better results will probably come from combinations of the methods outlined. The single largest factor, other than the obvious effectiveness of protecting the receiver, is to ascertain the IMD effects. Spectrum analyzer measurements are required.

Shaped Keying

A problem with many solid-state cw transmitters is key clicks. This is usually the result of oversight by the designer. So much effort is devoted to the rf details of the design that the shaping can be forgotten. There are many circuits that can be used to assure that the cw note is clean and crisp.

Shown in Fig. 93 is a circuit that is used frequently in many of the transmitters in the book. Here a pnp transistor is used as a switch. This circuit has several advantages. One is that the keying is done in the positive supply line, but the key is still grounded. This allows the builder to carefully ground the rf parts of the circuit without regard for extra dc control wires. The other virtue is that the switch provides an easy means of controlling the timing. This is performed with the network in the base. C1 and R1 determine the rise time of the waveform while C1 and R2 control the fall time. This circuit is suitable for keying stages requiring up to about 50 mA of current. Greater current amounts may be keyed if larger switching transistors are used. The base resistors must be decreased in ohmic value though.

A Class A amplifier may be keyed with the circuit of Fig. 94. Again, a pnp transistor is used. However, in this application it functions as an emitter follower instead of a switch. As such, the dc waveform is slightly more predictable than with the circuit of Fig. 93.

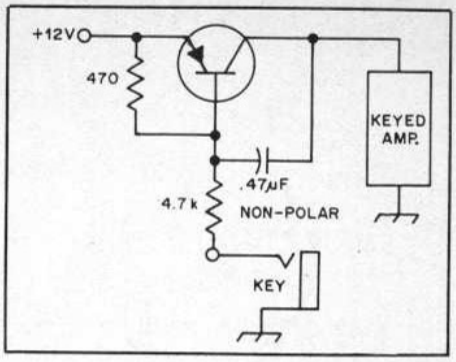


Fig. 95 — A pnp keying transistor operated as an integrator. The timing capacitor, C1, should not be an electrolytic type.

Fig. 95 shows a third method for applying a transistor to shaped keying. In this circuit, the transistor functions as an integrator. When the key is closed, base current starts to flow. However, this causes the voltage on the keyed amplifier to begin to rise. The increasing voltage is coupled back to the base, decreasing the base current. The final result is that the collector voltage ramps up linearly. A similar action occurs during the fall period. While the waveforms are trapezoidal instead of the more classic exponentials, they have low sideband energy. This results in a clean keying characteristic.

Chapter 8

Modulation Methods

The theory presented in the preceding chapters has been general, applying to cw and ssb systems. The construction projects have been predominantly for the cw enthusiast. However, phone is the principal mode of operation for most amateurs. On the hf bands single sideband is predominant. At vhf and uhf, there is a split. There is an increasing number of stations using ssb at vhf and uhf. The most common mode is fm.

In this volume we will treat the details of single and double-sideband phone transmitters. Frequency modulation methods are omitted because they are covered in detail in many other books.

Our treatment of sideband methods will include many problems. The text will deal with some introductory information on the design of the component parts of a phone transmitter, the design of high-power amplifiers, and the various methods that are available for realizing these ends. We will attempt to fill in some gaps that have appeared in the amateur literature. Specifically, the design of low- and medium-power Class A broadband amplifiers will be covered.

The Nature of an A-M Phone Signal

If a cw signal was to be expressed mathematically, it would be a simple sine wave. That is, the voltage appearing at the transmitter output would be $V_o = A \sin \omega t$ where $\omega = 2\pi f$ is the angular frequency. F is the frequency in hertz. The term A is the peak amplitude of the signal.

Modulation is a term that implies a controlled change. The terms in the simple cw signal that may be changed or modulated are the amplitude, A, or the frequency, f. The amplitude modulation (a-m) that is used for standard broadcast, and at one time was the dominant phone mode for the amateur, is described by

$$V_o = A_o(1 + k\sin 2\pi f_m t)\sin 2\pi f_c t$$

$$= A_o \sin 2\pi f_c + A_o k(\sin 2\pi f_m t)(\sin 2\pi f_c t). \quad \text{(Eq. 1)}$$

There are two frequencies represented. F_c is the carrier frequency. This is evident from the expanded form of the equation where we see that there is a steady output at the carrier frequency. A_o is the peak carrier amplitude, and the term k is called the modulation index.

The other term in the expanded form of Eq. 1 is a product of two sine waves. The two sine terms have two frequencies, f_c and f_m. The second frequency, f_m, is the modulation frequency. If we refer back to the discussion of mixers and product detectors in the receiver chapters, we will recall that a device which has an output that is a product of two input voltages (a multiplier circuit) has as its output a pair of signals at the sum and difference frequencies. Hence, the total output of the a-m transmitter will contain three frequencies. One is the carrier, f_c. The other two are called the sidebands, and are at frequencies $f_c + f_m$ and $f_c - f_m$. Shown in Fig. 1 are oscilloscope presentations of an a-m signal. In Fig. 2 are spectrum-analyzer presentations of the same signal. The carrier and the sidebands are clearly evident. In the case shown, the carrier frequency is 432 MHz. The modulation frequency is 1 kHz. If the value of k is multiplied by 100, the result is the percentage of modulation. The signal shown in the photographs is modulated 100 percent.

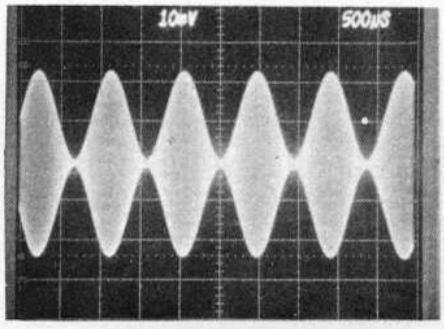


Fig. 1 — Time-domain oscilloscope presentation of an a-m phone signal. 100 percent modulation is present. The modulating frequency is 1 kHz.

It is interesting to note the powers that are associated with the various frequencies in a 100-percent modulated carrier when a single sine wave is used as the modulating signal. From Eq. 1 we see that the carrier power is a constant. The voltage of the carrier is A_o volts, peak. The rms value is $A_o \div \sqrt{2}$. Since the power is delivered to a resistive load, R, the power is just $V^2 \div R$, or ${A_o}^2 \div 2R$.

If Eq. 1 is expanded, using trigonometric identities, we see that the amplitude of each of the side bands is $A_o \div 2$. Hence, the average power in each of these sidebands is 0.25 of that in the carrier. A spectrum-analyzer display of an a-m signal which is modulated 100 percent by a single audio frequency will show accordingly that the average power in each of the sidebands is 6 dB below the carrier power.

If we go back to Eq. 1, we see that the normal cw signal with an amplitude of A_o is replaced by one with a variable amplitude. At some parts of the audio

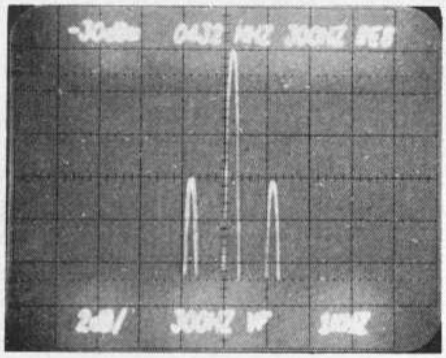


Fig. 2 — Frequency-domain presentation of the a-m signal of Fig. 1. The spectrum analyzer was set for a vertical display of 2 dB per division. Note that the sidebands are 6 dB below the level of the carrier. The modulating frequency is 1 kHz and the carrier is at 432 MHz.

cycle, the instantaneous amplitude is zero. At other parts, where the audio oscillation is at its peak value, the amplitude of the sine wave is twice as high as that of the carrier alone. This factor of 2 in voltage results in a factor of 4 in power. This power is called the peak-envelope power (PEP) and is 6 dB greater than the carrier power.

In the foregoing discussion, it has been assumed that the modulation signal is a single-frequency sine wave. This makes the mathematics easy. In the real situation, the audio signal would be the voice of the operator. This is composed of a number of sine waves added together to form a composite voice waveform. The transmitter will behave essentially as if each of the component sine-wave modulating signals were present alone. Then, the net output would be the addition of each of the individual components.

The Double-Sideband Signal

If the a-m signal is studied with a spectrum analyzer or mathematically, we find that the carrier at f_c has a constant level. As the audio signal is applied to the transmitter, the levels of the two sidebands vary, but the level of the carrier remains constant and unaltered. Hence, it contains no information. It is necessary if the signal is to be detected in a receiver using a simple rectifier detector, but it serves no other purpose. On the other hand, when we examined the average power in the carrier and in the two sidebands of an a-m signal, we found that most of the power was in the carrier. It would be much more efficient if we could concentrate all of the transmitted power in the sidebands where voice information is contained, and dispense with the carrier. This can be done: The result is a double-sideband signal.

If the spectrum-analyzer photographs of an a-m signal are studied, a double-sideband signal can be envisioned: It is the same presentation without the carrier present. This is done with a balanced modulator in practice. This circuit, which will be discussed in more detail, is essentially a balanced mixer. One of the inputs is at audio while the other accepts the carrier frequency. The output is balanced so that the carrier does not appear in the output.

The output of a double-sideband transmitter differs from the a-m one, in that there is virtually no rf output present until an audio tone (or voice waveform) is applied to the modulator. Then, rf output will occur. The average power in each of the individual side bands is always equal.

The Single-Sideband Signal

If the double-sideband signal is investigated, we see that each of the two sidebands is of equal amplitude and each contains the same information. Because of this, an improvement in efficiency can be obtained by removing one of the sidebands. With only one sideband being transmitted, all of the available power can be concentrated in the remaining one. Shown in Fig. 3 is a collection of spectrum sketches. A combination of three audio tones (Fig. 3A) is impressed simultaneously on a carrier. Fig. 3B shows the result with an a-m transmitter. Fig. 3C shows the result when a double-sideband transmitter is used. The spectrum that would result from these tones being transmitted on a single-sideband (ssb) transmitter is presented in Fig. 3D.

It is interesting to consider the power related to a single side-band transmission. Consider the usual case that is used for the testing of an ssb transmitter where two equal audio tones are applied to the audio input of the transmitter. The resulting output is shown in Fig. 4. Each of the tones has a given power associated with it. The average total power is merely the sum of these two, or twice the value of the individual signals. The peak-envelope power, however, is four times the value of each of the individual tones. The reason for this difference is because the two audio tones are not related to each other (they are incoherent). Because of this, there will be instants during the transmission when the individual equal voltages from each tone are both at their peak values simultaneously. The net voltage at the output at the instant is twice the value of one of the tones, and the resulting peak-envelope power (PEP) is 6 dB above the power in each of the two tones.

In a practical case it is much more difficult to relate the average power of an ssb signal to the PEP value. This will depend upon the characteristics of the

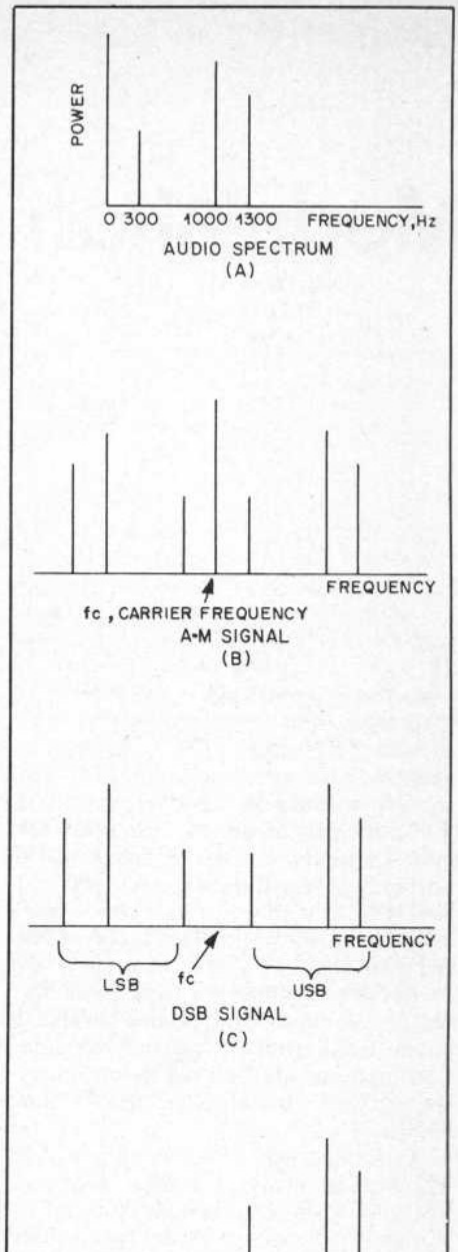


Fig. 3 — Representative spectrum displays for various modulation forms. The audio spectrum of three tones is shown at A. B shows the result when this audio signal is applied to an a-m phone transmitter. C shows the output spectrum when the three audio tones are applied to a suppressed-carrier double-sideband transmitter. At D, the output spectrum of an ssb transmitter is presented. Note that the ssb signal is exactly the same as the audio input except that it is translated in frequency.

voice that is being transmitted and upon the nature of the transmitter. Some transmitters use speech clipping or processing in order to limit the peak value of the waveforms while increasing the average power. In most cases where such methods are not employed, the

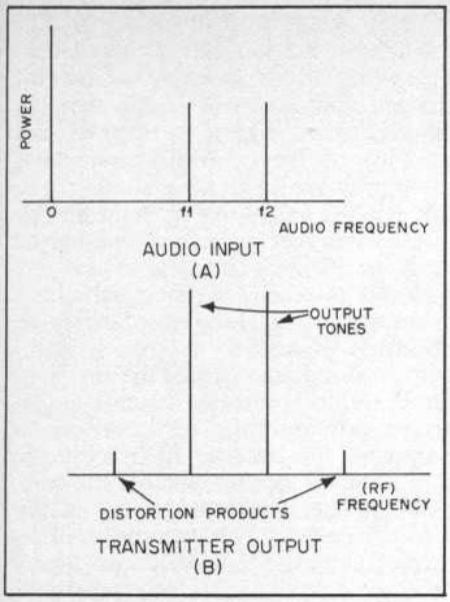


Fig. 4 — Spectrum obtained during two-tone testing of an ssb transmitter. At A the audio input is shown, consisting of two equal audio tones of identical amplitude. At B is shown the resulting ssb output including third-order intermodulation distortion products. Note the frequency spacings of the IMD products.

PEP value will be much greater than the average power of an ssb signal.

Single-Sideband Generation

There are two general methods that are commonly used for the generation of ssb. One is the filter method and the other is the phasing method. A block diagram of a filter type of ssb generator is shown in Fig. 5. This technique is virtually identical to that used in a superheterodyne receiver, except that the signal direction is different through the transmitter than it is in the receiver. An audio signal is obtained from a microphone and is amplified in a speech amplifier. It is then applied to the input of a balanced modulator. The output of the balanced modulator is a double-sideband signal. The carrier for the balanced modulator is most often obtained from a crystal-controlled oscillator.

The dsb signal from the balanced modulator is applied to a crystal or mechanical filter. This filter is designed such that one of the sidebands from the modulator is within the passband while the other is not. The result is an ssb signal. For high-quality ssb signals to be generated it is not necessary that the filter response have a symmetrical shape. It is only necessary that the suppression be quite good for the unwanted side band. If a symmetrical filter is used, as is usually the case, the crystal in the carrier generator (used to drive the balanced modulator) may be switched, allowing the operator to change the sideband that is being transmitted. If the sideband from the filter is higher in frequency than the carrier, it is called the upper sideband (usb). The lower sideband (lsb) is similarly defined.

Since fixed-frequency filters are usually employed for the generation of ssb, it is necessary that the intermediate frequency ssb output be heterodyned to the frequency of interest. This is done with a mixer and LO system. Again, we emphasize that the filter method is an exact analogy to the superheterodyne receiver. Either single or multiple conversion may be employed.

The second method used for the generation of ssb is called the phasing method. This is shown in Fig. 6. The basis of such a ssb generator is a pair of balanced modulators. Each is driven with identical carrier frequencies and audio signals of identical amplitude. However, the phase of the signals is different. The carrier signal driving one modulator is 90 degrees out of phase with that driving the other. Similarly, the audio driving one balanced modulator is 90 degrees out of phase with that driving the other. The outputs of the two balanced modulators are added together with the result that only one of the sidebands remains. It is not immediately obvious that such a collection of circuits will lead to a single sideband. However, the mathematics used to show that this does occur are straightforward and are presented in the appendix.

In the early days of *amateur* ssb the phasing method of generation was popular. The reason for this was that the technology for filter construction was not as advanced as it is today. Furthermore, the phasing method may be applied directly at the band of interest. A superheterodyne approach to design is not mandatory, although it may certainly be used.

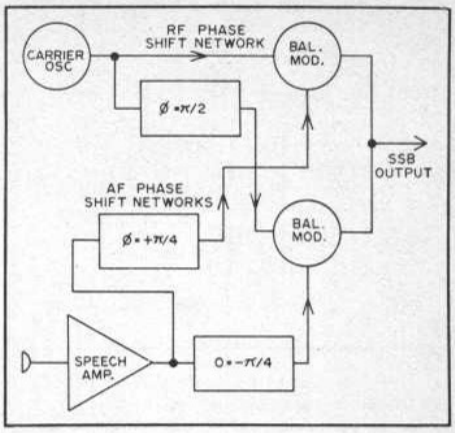


Fig. 6 — Block diagram showing the phasing method of ssb generation. Two balanced modulators are used, each being driven with rf and audio signals of identical amplitudes. The rf and audio signals to the two modulators are in phase quadrature. A mathematical analysis is presented in the appendix.

Today, the situation is reversed. The filter method is predominant for sideband generation. This is largely a result of the nature of the filters that are available, along with the transceive concept where the same filter may be used for sideband generation and to obtain receiver selectivity. The other reason is that the filter method does not exhibit the fundamental disadvantages that are typical of the phasing method. This requires some elaboration.

If a single audio tone was to be transmitted at a single frequency with the phasing method of ssb, the design would be straightforward. Building networks that provide 90 degrees of phase shift at a single frequency is generally easy. This is not what is needed for sideband generation, though. The rf phase-shift network must operate accurately over a small range, equal in the worst case to the width of a phone segment of an amateur band. This is not difficult to realize in practice. What *is* difficult is the construction of the audio phase-shift network. The voice spectrum is generally considered to be from 300 to 3,000 Hz. This is a ratio of 10 in frequency. It is difficult to build phase-shift networks that will maintain a 90-degree phase difference with constant output amplitude over this large range. It can be shown that as little as one degree of error in the audio phasing will lead to an ultimate suppression of the undesired sideband of only 41 dB.

Technology is changing and modern methods may inspire a renewed interest in the phasing types of ssb generators.

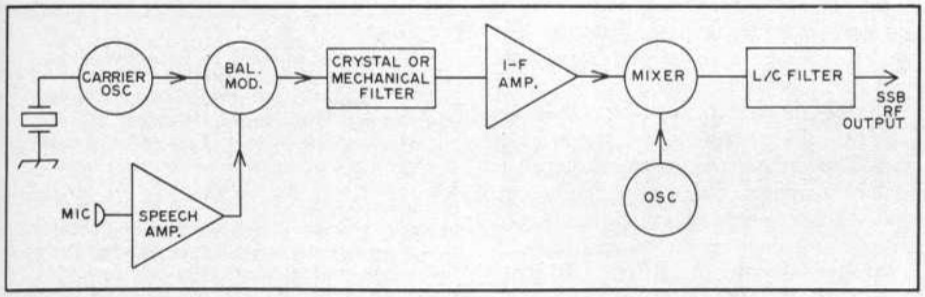


Fig. 5 — Block diagram showing the filter method of ssb generation. The carrier oscillator frequency is adjusted so that it coincides with a point that is 20 dB down on one of the sides of the response of the bandpass filter. After an ssb signal is obtained at an intermediate frequency it must be hetrodyned to the desired output frequency.

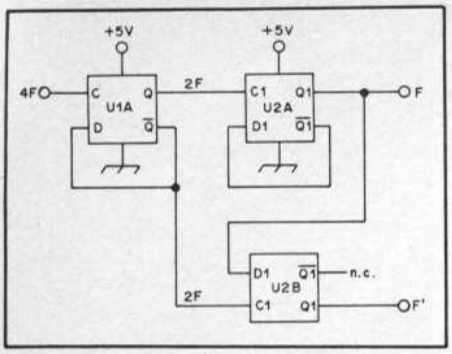


Fig. 7 — A circuit using digital ICs for generation of quadrature rf signals as the drive for the balanced modulators in a phasing ssb generator. Nominally, the ICs would be TTL D-type flip flops such as the SN7474. For higher frequency operation, suitable MECL equivalents could be used.

Radio-frequency phase shift can be achieved easily using digital methods which are inherently broadband. Specifically, if two quadrature (90-degree phase difference) outputs are desired at a frequency f_c, one starts with an oscillator at $4f_c$. This signal is then applied to a digital divider using a flip-flop with complimentary outputs. The result will be two output signals at a frequency of $2f_c$ which are 180 degrees out of phase with each other. Each of these signals is applied to flip-flop dividers. The resulting outputs will be at the desired f_c and will be in quadrature. A slightly more elaborate interconnection of digital ICs will be required than that described, in order to ensure that the proper sideband will result every time power is applied. This is shown in Fig. 7.

The other phase-shift problem which is being changed by modern technology is the one occurring at audio frequencies. The classic circuits that were used for audio phase shifting contained resistors and capacitors in a complex network. The newer approach embodies an active phase-shift network. A circuit is shown in Fig. 8, where resistors and capacitors have been combined with an operational amplifier to obtain a phase-shifted output. High-performance versions of this method will use a multiplicity of these active networks (cascaded) in order to obtain accurate phase shifts over a wide range of audio frequencies. No component values are given in Fig. 8 since they will depend upon the accuracy desired. The reader who is interested in studying this design technique should investigate the 1970 paper by F. R. Shirley (see the bibliography). Using quad op amps like the LM324, builders should be able to make the phase-shift networks compact and low in power comsumption. If a phasing ssb exciter is to be built, it is important that the audio signals reaching the phase-shift networks be carefully confined to the spectrum of interest. Because of this requirement, the speech amplifier should include extensive filtering. The *RC* active filters discussed in connection with receiver design may be used to realize this end.

The writers have not used any of the technology outlined for the construction of phasing exciters. Our work has been confined to the filter approach and to double-sideband transmission methods.

While the filter and the phasing methods of sideband generation are familiar to many amateurs, there are others that may be used. One is known as "the third method," or Weaver method, named after its originator. A reference is given in the bibliography for this technique. Also, it may be shown mathematically that a carrier which is amplitude modulated properly and frequency modulated simultaneously will yield a single-sideband output.

Single- and Double-Sideband Receivers

Receivers for single-side-band are usually "superhets." That is, they employ the filter method for reception. However, the phasing method or the Weaver approach may be applied to ssb reception. There has been some recent work with both of these, which are essentially extensions of direct-conversion receivers. While each method works, both have limitations. The main problem with the phasing method for receivers is the limited sideband suppression available. A sideband suppression of 40 dB is acceptable in the ssb transmitter. However, this level would be intolerable in any but the simplest of receivers. Furthermore, the complexity of the filter method is so much less than a phasing approach to receiver design that we do not recommend the phasing technique. One exception would be those cases where extremes of sideband suppression are not needed. For example, one might use the phasing method in a receiver as a technique for filtering the i-f amplifier for noise. This would replace the matched noise filter that might be used between the i-f amplifier and the product detector in an advanced receiver. The main selectivity of the receiver is still provided by a high-performance crystal filter at the input to the i-f amplifier. References are given in the bibliography.

Direct-conversion receivers may be used for the reception of single sideband. The only problem encountered is the audio image. This image frequency may contain another station that would cause interference to the desired one.

Double-sideband reception is straightforward with the typical superhet receiver. This is because the filter in the receiver removes one sideband, converting the signal arriving at the product detector to an ssb signal. For this reason, dsb transmitters are compatible with stations operating ssb. Indeed, the operator may not realize that the other sideband is present. However, the presence of the other sideband could cause interference to other stations. For this reason, we do not recommend dsb transmitters for use in crowded bands.

A problem is incurred when one attempts to receive a dsb signal with a direct-conversion receiver. Since the operation of a "d-c" receiver is essentially that of heterodyning the energy in the rf spectrum directly down to audio, proper dsb operation detection occurs only when the receiver BFO is *exactly* at the same frequency, and has the same phase as the suppressed carrier of the dsb transmitter. This can be realized through advanced detection methods, but is generally not recommended. Again, the reason is the circuit complication and the extra spectrum occupied by the dsb signal.

Balanced Modulators

All of the techniques used for the generation of dsb and ssb use a balanced modulator. There are numerous methods for realizing such a circuit. Some of them will be presented in this section.

A balanced modulator is nothing but a balanced mixer. These circuits have been discussed in detail in connection with receiver applications. The difference between the ordinary receiver balanced mixer and a balanced modulator is in the frequencies presented to the input of the circuits. The receiver mixer has two radio frequencies at its input, with an intermediate-frequency output. The balanced modulator has one radio

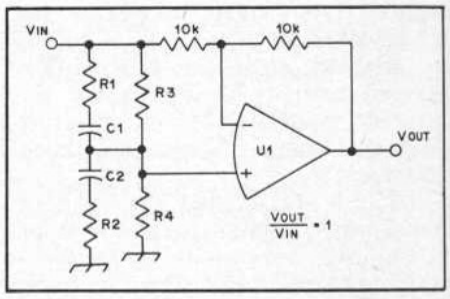


Fig. 8 — Circuit showing an RC active audio phase-shifting circuit. This is an "all pass" network, with the output-voltage amplitude equal to that at the input. However, the output will be phase shifted. In practice, a pair of chains of such circuits will be employed. Each chain will contain from two to four cascaded circuits of the type shown. The inputs of the two chains are driven in parallel. The two resulting outputs are applied to the balanced modulators. For calculation of the values of R1, R2, R3, R4, C1 and C2, the reader is referred to the engineering literature (F.R. Shirley, *Electronic Design*, Sept. 1, 1970). The op amps may be a 741, one half of a 747 or 5558, or one quarter of an LM324.

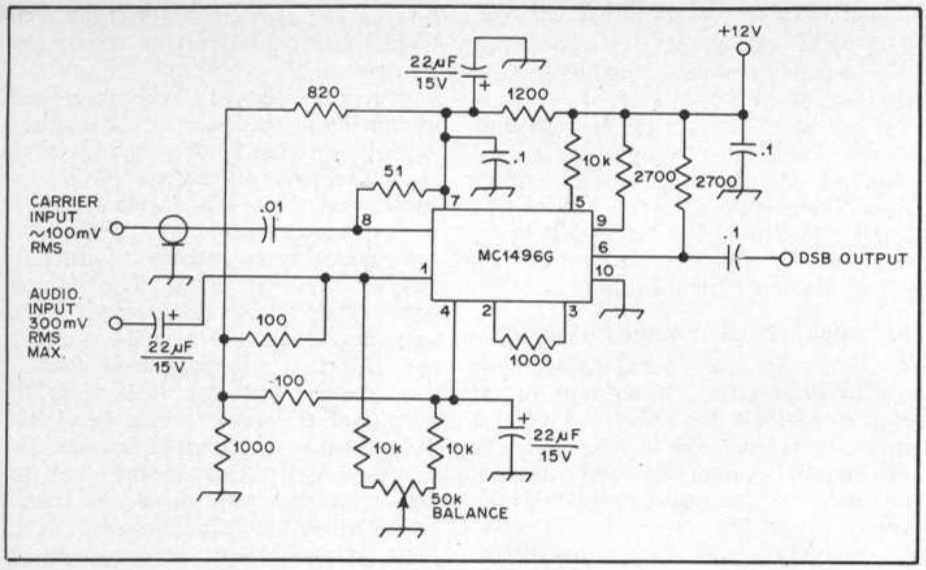


Fig. 9 — A balanced modulator using the MC1496G. The 50-kΩ control is adjusted for optimum carrier balance.

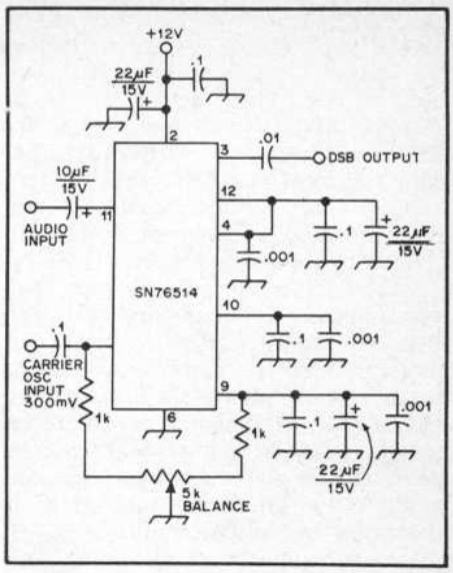


Fig. 10—The SN76514 mixer IC used as a balanced modulator. The SN76514 has been reidentified as TL-442-CN by Texas Instruments. It may be procured under either part number.

frequency and an audio input. The outputs are the sum and difference frequencies, or the two sidebands. The balance in the circuit ensures that a minimum amount of carrier energy feeds through to the output. Representative values of carrier balance or suppression are from 30 to 70 dB. For an ssb transmitter using the filter method, carrier suppressions of 50 dB or greater are sufficient. This is because the filter will often add another 20 dB of carrier suppression.

The operating power level of a balanced modulator is somewhat critical. As outlined, a voice waveform can be analyzed as a composite number of sine waves. If the balanced modulator is operated at levels that are too high, intermodulation distortion will occur between these components to make the voice sound distorted. If the balanced modulator is used in a filter type of ssb exciter, all of the resulting distortion products reaching the antenna will be within the voice spectrum. This is because the filter will remove the undesired ones. However, in a phasing ssb rig or in a dsb transmitter, some of the distortion products could lie outside of the desired voice sideband.

Shown in Fig. 9 is the circuit for a balanced modulator using the MC1496G IC. A potentiometer is used for adjustment of the balance. With careful setting of this control, a carrier suppression of 60 dB is achieved easily up through 10 MHz. An easy way to adjust this control is to listen to the mixer output in a receiver, then set the control for minimum output (no audio applied). Using this circuit, the recommended output level is around −10 dBm. If it is desired to operate at high output levels, the current standing in the IC should be increased. This is done by changing the 10-kΩ resistor leading to pin 5 to a 3,300-ohm unit. In this case the maximum output should be around 0 dBm or less. The recommended output levels are for each tone during a two-tone test, where a single audio signal is placed at the input. With the levels suggested, IMD products should be below the output by 20 dB or more. This is probably adequate for ssb exciters using the filter method. Phasing ssb exciters and simple dsb transmitters using this circuit should be run at lower output levels.

A similar circuit is shown in Fig. 10, where an SN76514 is used. Although not shown in the literature for this device, the carrier suppression may be improved with the addition of a control, as shown. The recommended output levels for this circuit are about the same as with the MC1496G.

A number of balanced-modulator circuits are available to the builder who uses diodes. Shown in Fig. 11A is one of the simplest of these. It has but two diodes. In this circuit the balance will vary with frequency and is dependent primarily upon the match in the diodes and the symmetry of the transformer. The recommended carrier-oscillator injection power for all of the diode circuits shown is +13 dBm. At this injection level, the circuit may be operated at output powers up to 0 dBm per tone in a two-tone output (which results from a single audio input tone).

Some variations of this circuit are also shown in Fig. 11. One uses a variable resistor in series with the diode pair, with the output being obtained from the arm of the control. This circuit is recommended for use on the lower hf bands and is capable of providing a

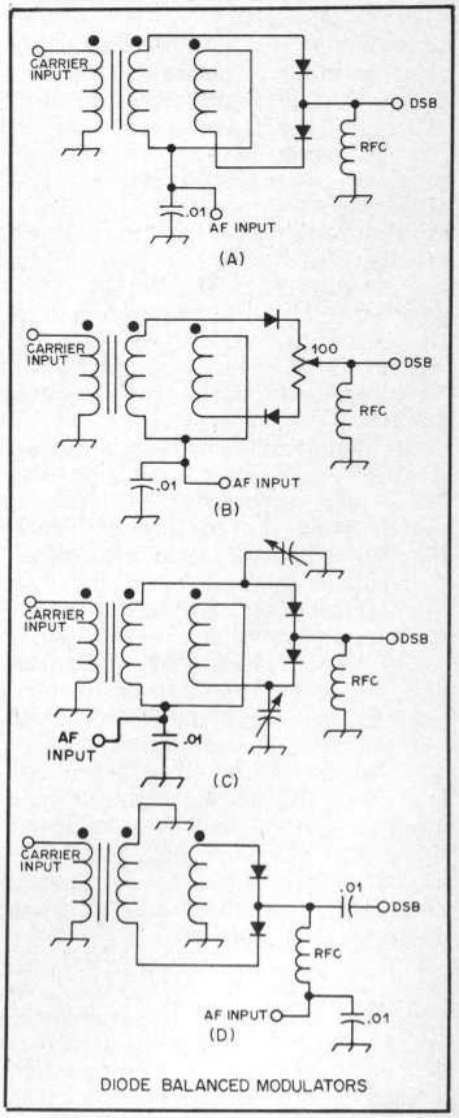


Fig. 11 — Balanced modulators using two diodes. These circuits are ideal for the construction of simple dsb transmitters (see text for a discussion of components).

carrier suppression of up to 50 dB, if carefully adjusted. The other method for balance adjustment (Fig. 11C) uses a pair of variable capacitors. This technique is best for vhf applications. We have measured over 50 dB of carrier suppression at 144 MHz with this circuit. The two methods could be combined for an improvement in suppression at the lower frequencies.

The choice of diode type will depend upon the frequency of operation. For vhf applications, a hot-carrier diode is suggested. However, for the hf bands suitable results could be realized with 1N914 or similar types of silicon switching diodes. In most of the circuits presented the audio signal is introduced at the center tap of the transformer. It is possible to apply the audio directly at the connection between the diodes. This is realized with an rf choke to isolate the rf output from the audio system (see Fig. 11D). This may lead to slightly improved balance at uhf and could be the recommended circuit for building a 432-MHz dsb exciter.

Shown in Fig. 12 are two other diode balanced modulators. Those circuits with four diodes are doubly balanced, although it is not a necessity in this application. With any of the diode balanced modulators the output should be terminated in 50 ohms on a broadband basis. It may be useful to employ a low-pass filter at the output of the modulator to reduce the harmonic content, especially when dsb transmitters are being built. Prepackaged diode-ring mixers are not recommended, since there is no way to adjust carrier suppression.

If careful design work is intended, the data presented in connection with mixers for receivers should be consulted. Specifically, the intercept at the output port should be studied in order to determine the level for proper operation of the mixer. The higher output levels have the advantage that less gain is needed in the following stages. This can be a major advantage in a double-sideband transmitter. On the other hand, in a filter type of ssb exciter, gain is achieved in an i-f amplifier. This means that the balanced modulator can be operated at a low level to make distortion effects inconsequential. The output should not be reduced too far though. This could raise the broadband noise output of the transmitter.

In all of the balanced modulators shown, the audio port is dc coupled. As a result, a cw output can be produced by injecting a dc voltage to unbalance the modulator. If the carrier suppression is good, the transmitter may be keyed by shaping the dc that is applied. In most situations it will be desirable to key an additional stage in the transmitter. Examples are presented later.

An additional advantage of the dc-coupled nature of the audio-input port is that a-m phone operation may be realized. A slight amount of carrier is inserted by injecting a dc component of current until the proper levels are obtained at the output. The output should be monitored on an oscilloscope until 100-percent modulation is obtained. Methods are outlined in *The ARRL Radio Amateur's Handbook*.

I-F Amplifier and Transmit-Mixer Design

With a few exceptions, the design of the i-f system for a filter type of ssb exciter parallels the same section of a superhet receiver. The differences are in the output level of the mixer desired, and the level that may be applied to the crystal filter.

As mentioned in the previous discussion of balanced modulators, in a filter ssb system the output of the modulator may be kept to a low level. This minimizes distortion in that circuit. The additional gain is then obtained in the i-f system. There are upper limits on the signal level that should be reached within the i-f. First, it is sometimes dangerous to crystal filters if the power level impressed at their input is excessive. This will, to some extent, depend upon the nature of the filter. With most units designed for ssb bandwidths, levels as high as 10 to 100 mW will not cause damage. The real problem comes with narrow-bandwidth crystal filters, as might be used for some cw applications. This only becomes of significance in the present discussion of ssb methods if the builder is considering a multimode transceiver.

The main constraint on power levels within the i-f section of an ssb exciter is in the level used to drive the mixer. In our discussion of receiver mixers, we found that there was a wide variety of performance available. Specifically, various mixers were capable of different output intercept values. The transmit mixer that follows the transmitter i-f amplifier should be operated such that the distortion is minimized. Generally, this implies that the IMD from the output of the mixer should be at least 40 dB below the desired outputs in a two-tone test. This means that the output of each tone should be at least 20 dB below the output intercept of the mixer. On the basis of measurements that we have done, this suggests that the mixer output should be around −5 dBm for diode-ring mixers, and should be −10 dBm or less for an MC1496 mixer. This assumes that the MC1496 is biased for optimum signal-handling capability.

Shown in Fig. 13 is an i-f amplifier using bipolar transistors. It is followed by a MC1496 mixer. This circuit is designed around a KVG crystal filter which has an input and output termination requirement of 500 ohms. The first stage in the amplifier has a variable gain. This is realized with a variable resistor in the emitter circuit of the stage. Note that the current in the transistor is kept constant to maintain a high signal-handling capability. The second stage

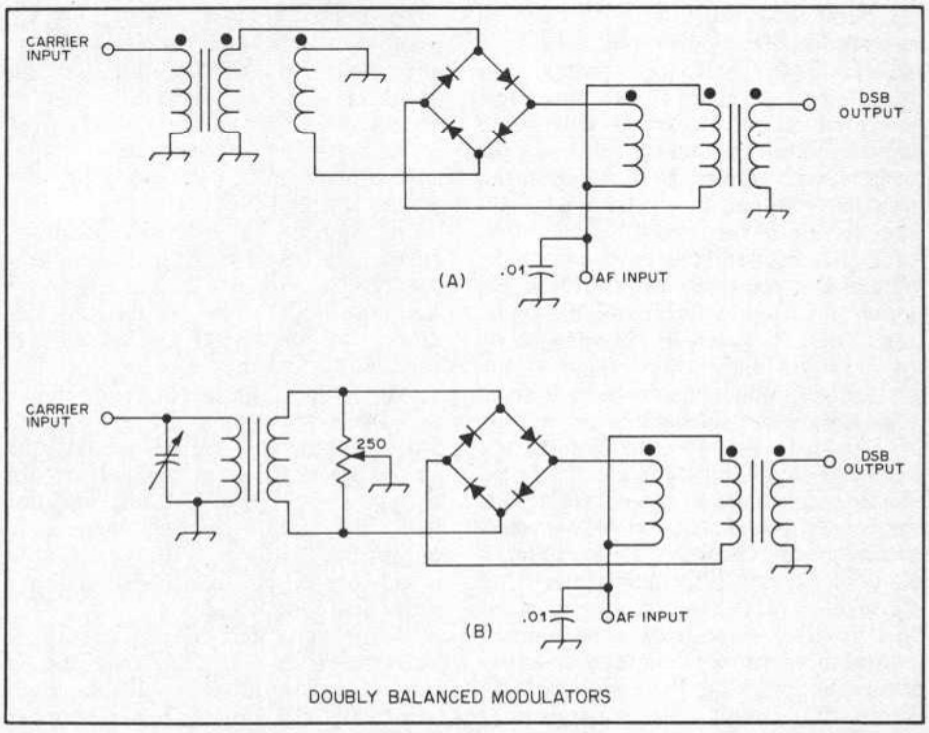


Fig. 12 — Balanced modulators using diode rings. The 250-ohm control in B is adjusted for optimum carrier balance (see text). Also see the previous discussion of product detectors and mixers using diodes (chapters 5 and 6).

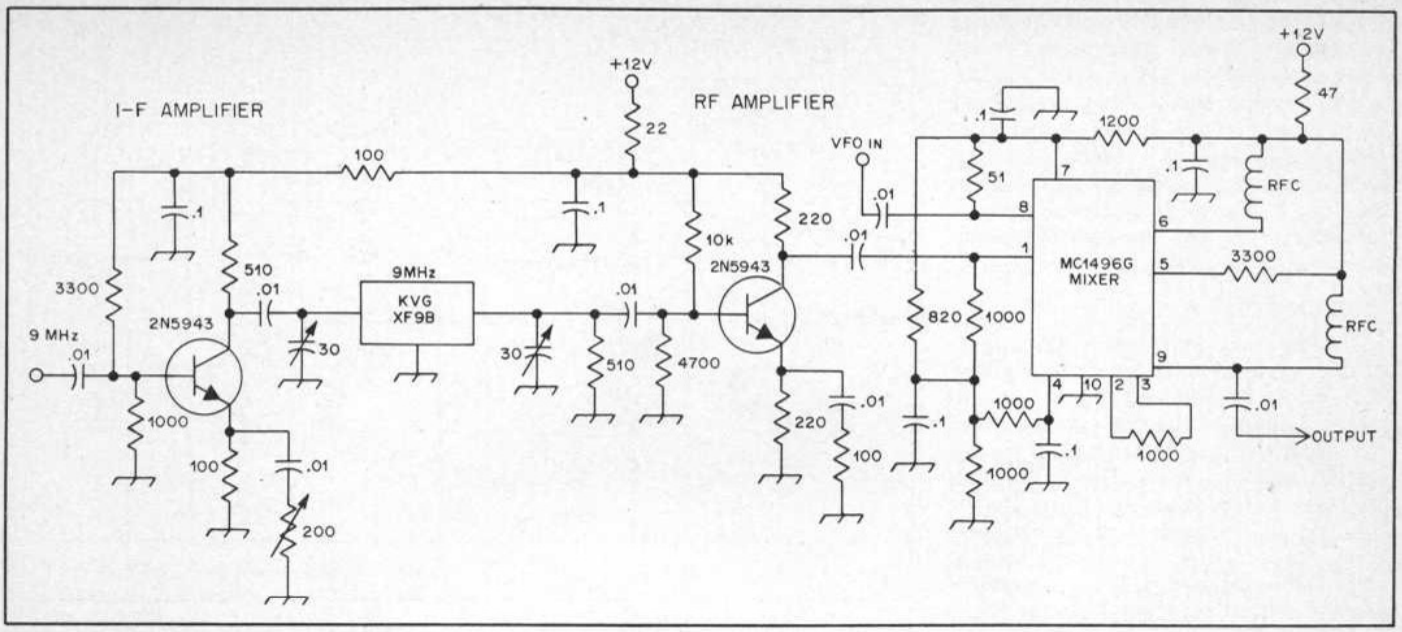


Fig. 13 — Representative i-f amplifier and transmit mixer for use in a filter type of ssb exciter (see text).

also employs emitter degeneration. The main need for this is to maintain a high input impedance to the amplifier. Because of the light loading that the amplifier presents, the termination on the crystal filter is determined by the external 510-ohm resistor. The output of the amplifier is applied directly to the mixer input, while the LO port of the mixer is driven by a suitable VFO.

Field-effect transistors may also be used in the i-f amplifier. A transmitter presented later in the chapter uses dual-gate MOSFETs in the i-f section.

While there are a large number of mixer devices that may be used in high-level transmit applications, it is highly recommended that a doubly balanced design be chosen. The section on the discussion of transmit mixers given in an earlier chapter should be consulted. Generally, we would suggest that a MC1496 be used for single-band designs up through 30 MHz. This IC is easy to apply and offers suitable, if not spectacular signal-handling capability. For use into the vhf spectrum, a diode type of doubly balanced mixer is recommended. This would also be ideal for a multiband hf design because of the broadband capability of the circuit. However, it is important that the proper levels be maintained throughout the system. The measurement of low levels of rf power was discussed in chapter 7. It is recommend that the designer use a low-level detector (such as the square-law detector described earlier) in conjunction with a step attenuator in these projects.

When using a diode-ring mixer, all of the precautions about termination detailed in the receiver chapter should be followed. In a single-band design it is often possible to use diplexer circuits, as were presented. However, a much simpler approach is to use a 6-dB attenuator with a characteristic impedance of 50 ohms at the output. This was not desirable for the receiver because of noise-figure degradation. However, noise figure is of less significance in transmitter applications. The 6-dB pad should ensure that all mixer products are terminated properly. Correct LO injection should also be employed for the mixer. For diode rings, this is from +10 to +13 dBm to a 50-ohm load.

The mixer should be followed with a bandpass filter. The complexity of this filter will depend upon the exact frequencies involved. The main spurious response to guard against is the image. For example, if a 9-MHz i-f were used in a single-conversion transmitter for the 50-MHz band, the required LO frequency would be 41 MHz. The image frequency would be 32 MHz. A double-tuned circuit would provide more than sufficient rejection of this component. Three-pole filters might be more desirable for most of the hf bands. The filters listed in the appendix are suitable for this application.

In some cases, a low-pass filter might suffice. For example, if a transmitter was built for the 75-meter band, with an i-f of 9 MHz, the LO would probably be at 5 MHz. If the balance of the mixer is reasonable, the 5-MHz output component will be attenuated considerably prior to filtering. The main spur would be the image at 14 MHz. A low-pass filter with a 4-MHz cutoff frequency would provide more than sufficient suppression. The better circuit would include a trap or two with frequencies of high attenuation near 5 MHz. This would provide additional attenuation of the LO than might result from less than optimal mixer balance.

Dual-conversion systems should be avoided. The high signal levels that are often present can lead to distortion effects. These are complicated with extra conversions of the signal. A better approach would be to premix a low-

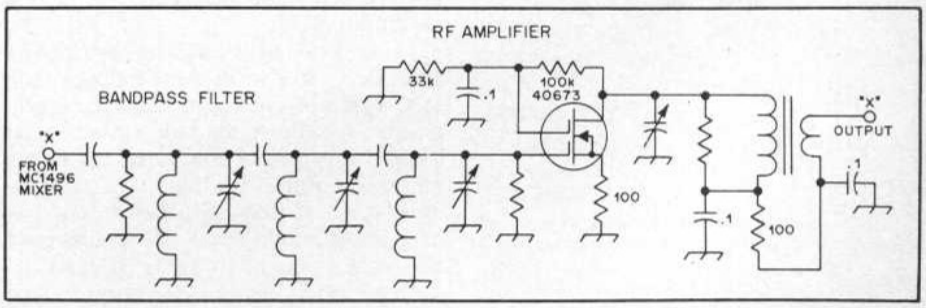


Fig. 14 — Circuit of an rf amplifier that might follow the mixer in Fig. 13. Band switching is simplified by multiplexing the dc voltage for the amplifier on the output signal line. A typical gain for this circuit would be 20 dB, with an output intercept of +20 dBm.

frequency local oscillator with a crystal-controlled source. This output would serve as a suitable LO injection for the single transmit mixer. This method can be followed on all amateur bands up through 432 MHz if a 9-MHz i-f is used and advanced filter design methods are employed. These filters are difficult to fabricate in the vhf and uhf region, but are certainly possible.

Broadband Class A Amplifiers

In the previous section limits were placed on the maximum output power that should be obtained from a transmit mixer. These contraints resulted from the need to keep intermodulation distortion to a minimum. The level for an MC1496 was around −10 dBm. By the time we add in the loss of the bandpass filter, levels as low as −15 dBm might be available. While diode-ring mixers can provide output powers which are somewhat higher, much of this extra power is absorbed in the 6-dB attenuator recommended for proper mixer termination.

On the other hand, most of the higher level Class AB amplifiers that are used for ssb service require a drive power of 1 to 5 watts. To reach this level, 45 or 50 dB of gain are required following the mixer. While this is not difficult to obtain, the problems become more severe when distortion requirements are considered.

One solution is to use narrow-band circuitry. This would not be out of line for part of the output chain of a band-switched exciter. An amplifier could be imbedded within each of the filters in order to provide gain. Shown in Fig. 14 is a filter of this kind, with a dual-gate MOSFET amplifier included.

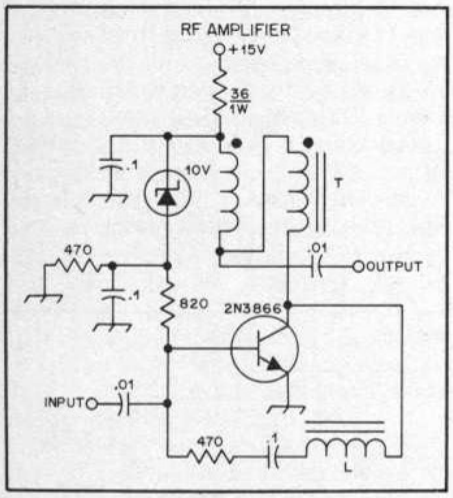


Fig. 15 — Example of shunt feedback in a broadband Class A medium-power amplifier. This circuit will provide a gain of 20 dB, outpower power of +23 dBm PEP and an output intercept of +37 dBm.

T = 5 bifilar turns, 0.2-in. ferrite core, u_i = 850.

L = 16 turns on T-37-6 (0.77 μH).

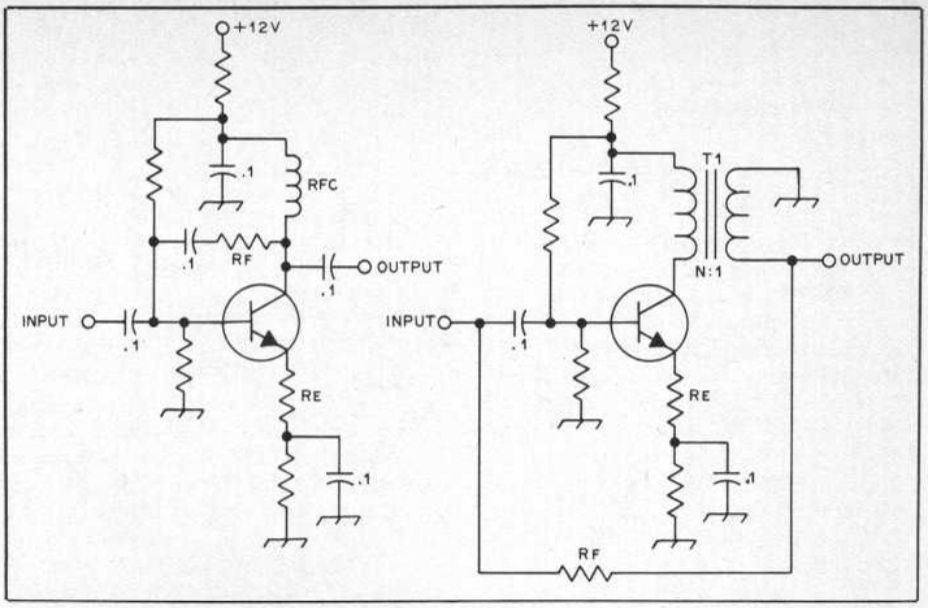


Fig. 16 — Examples of amplifiers with shunt feedback and emitter degeneration. T1 is a broadband transformer on a ferrite core with an N:1 turns ratio.

Note that no additional band-switch wafer would be required for this circuit, since the power supply is multiplexed onto the output of the circuit. The input triple-tuned circuit is one from the catalog of filters in the appendix and the output is a single, broadly tuned circuit. This stage should provide up to 20 dB of gain on all of the hf bands with an output intercept power of approximately +20 dBm. To ensure that the intermodulation distortion contribution from this stage is kept to a level of −40 dB or better, the output power should not exceed 0 dBm.

One could continue with a narrow-band amplifier design. This would be ideal in the case of a single-band transmitter. However, if the transmitter were to be used on several bands, a better solution would be to use broadband circuitry. The spectrum of signals arriving at the input to such an amplifier is now well defined. The distortion in a broadband amplifier may cause intermodulation products and harmonics to be created. The distortion products can be minimized with proper design of the amplifiers, while the harmonics are well attenuated with a low-pass filter at the output. The band switching is held to a minimum.

The key to designing broadband amplifiers is *feedback*. Feedback can take a number of forms. Emitter degeneration has been used in a number of designs throughout the book, and is one common form of feedback. Alone, however, it is not sufficient in the design of broadband amplifiers. While it does have the effect of establishing a constant voltage gain where the output load resistance is established, it has the additional effect of increasing the input impedance of the transistor. This increase is roughly proportional to the beta of the transistor. Since transistor beta is well approximated as $f_T \div f$ in the high-frequency region, the increased beta at lower frequencies leads to an increasing input resistance as frequency is lowered. In a multistage amplifier, this leads to increasing gain with decreasing frequency.

One other form is shunt feedback. This usually takes the form of a resistor between the collector and the base of the transistor. This has two advantages. First, it stabilizes the current gain of the amplifier, an effect similar to the virtues of emitter degeneration. However, it also decreases the input and output resistances of the stages.

Many examples have appeared where we have applied emitter degeneration alone. Shunt feedback may also be applied alone. Shown in Fig. 15 is an amplifier that uses shunt feedback. The transformer allows the 50-ohm load to appear as 200 ohms at the collector. This is adequate for a maximum power output on the order of 1/4 watt. The feedback path from the collector to the base contains a blocking capacitor, a small inductor and a 470-ohm resistor. The inductor has the effect of decreasing the feedback at high frequencies while the 470-ohm resistor is the dominant element at low frequencies.

Measurements were performed with this amplifier. The transducer gain was measured in a 50-ohm system as 19 dB. The points where the gain was down by 3 dB were 1 and 50 MHz. The upper limitation was the result of decreasing transistor gain — the f_T of the transistor was approximately 500 MHz. The low-frequency limit was a result of the transformer running out of inductive reactance. Only five bifilar turns were

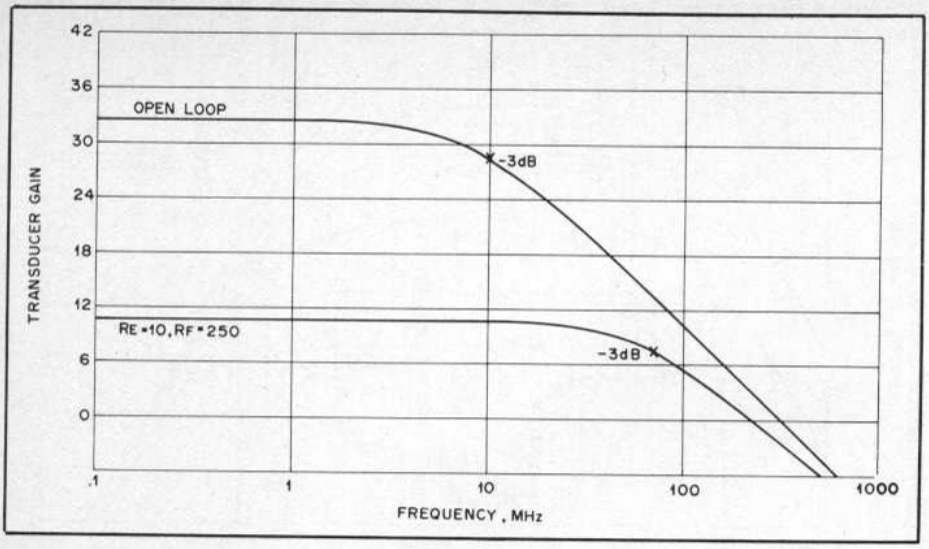


Fig. 17 — Transducer gain as a function of frequency for an amplifier with and without feedback. The hybrid pi model of a bipolar transistor was used for this calculation. A dc beta of 100 was assumed with f_T = 500 MHz, C_{cb} = 3 pF and R_b' = 50 ohms. Note that the gain with feedback is always lower than the open-loop gain (with no feedback) and that the bandwidth is always extended by application of negative feedback.

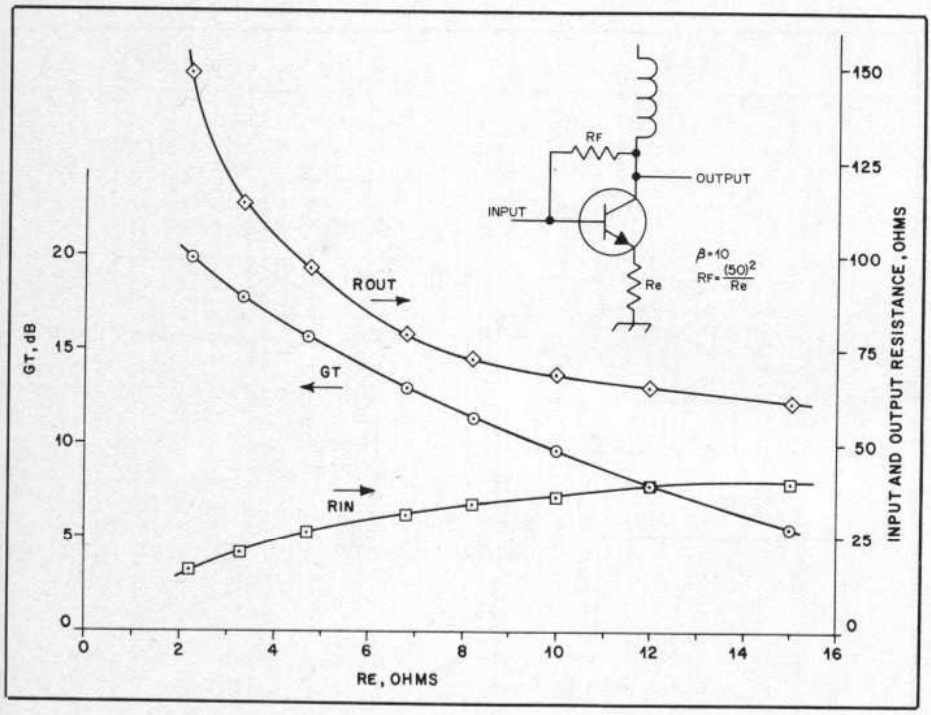


Fig. 18 — G_t, R_{in} and R_{out} as a function of feedback components. A simple model was assumed for this calculation with a beta of 10. No account was taken for phase shifts in beta. Nonetheless, the calculations agree well with measured results. A profound advantage of feedback is predictability in design.

used on a small ferrite core (Amidon FT-23-43).

The transistor was biased for about 120 mA of collector current. With this much current it would be expected that the output intercept might be fairly high. It was measured with two outputs of +17 dBm each, or +23 dBm PEP (200 mW) output. The intermodulation distortion products were 40 dB down from each tone, indicating an output intercept of +37 dBm. The measurements were done at 10 MHz.

Another approach to broadband design is to utilize a combination of emitter degeneration and resistive shunt feedback (see Fig. 16). This scheme has a number of advantages. First, it provides two "handles" on controlling feedback, which leads to greater flexibility. Second, the effect of feedback on impedance can be exploited. Since emitter degeneration has the effect of increasing input impedance, while shunt feedback decreases it, the combination effect causes the input impedance to be approximately constant. Also, the shunt feedback decreases the output impedance, leading to better interstage matching. Finally, emitter degeneration often has the effect of making an amplifier self-oscillate at some frequencies. This is especially true if the transistor has a very high f_T. On the other hand, shunt resistive feedback almost always has the effect of making an amplifier unconditionally stable. This can be of significance in a high-gain amplifier chain.

Shown in Fig. 17 is the effect of feedback upon transducer gain. This is a calculation based upon a transistor with a dc beta of 100, an f_T of 500 MHz and a 3-pF collector-to-base capacitance. As shown, without feedback, the gain at low frequencies was over 32 dB. However, the 3-dB bandwidth was only 8 MHz. When a 10-ohm emitter resistor and a 250-ohm shunt feedback resistance were added, the gain dropped to a little over 10 dB. However, the 3-dB bandwidth is now extended to 65 MHz. The transistor model used in this analysis is the so-called hybrid-π model, and is covered in the appendix.

If the amplifiers are to be cascaded, it is desirable that their input and output resistances be equal. Analysis shows that a rule of thumb may be applied. If the desired characteristic impedance is Z_o, then the emitter resistance and the shunt feedback resistance should be chosen such that $R_e R_f = Z_o{}^2$.

Fig. 18 has a curve showing the effect of emitter resistance upon stage gain, plus input and output resistance. The amplifier was designed for a 50-ohm characteristic resistance. *Hence*, for a given emitter resistance, R_e, the feedback resistor used was chosen according to the rule given above. That is, $R_f = (50)^2 \div R_e$. In this calculation, a simpler model was assumed for the transistor, with no account taken for a phase change of beta. The value of beta assumed was 10. In spite of the simple model, the results agree with the measurements we have done on amplifiers of this variety. It is interesting to note that the rule is a little away from the stated design center. That is, the input resistances are a little under 50 ohms, while the output resistances tend to be a little higher. Measurements confirm this calculation, also.

The gain of a single stage may be increased over those values given in Fig. 18 by the inclusion of a transformer in the output. The turns ratios are from 1:1 to 4:1. It is not necessary that transmission-line transformers be used, although this may enhance performance in the vhf spectrum.

Shown in Fig. 19 is a curve of gain vs. frequency for four different cases

where transformers are used. The high-frequency rolloff is determined by transistor characteristics, while the low-frequency drop in gain is a result of the transformer model used. These calculations were performed using the hybrid-π model which includes the effect of beta changes at high frequency, including a phase change.

The information presented so far has dealt with small signal models. We have used the data to predict gain and input and output resistances for the amplifiers. Shown in Fig. 20 is a practical circuit where these ideas are applied. Assume that an output power of 1/2 watt is desired from this amplifier. If this power is to be realized, the output load resistance presented to the collector must be reasonably low. A 2:1 transformer could not be used in the output, since this would place a 200-ohm load at the collector. This load would be too high unless a supply voltage greater than 12 were used. For simplicity, we will terminate this stage in 50 ohms, and ask for a gain of 10 dB. Looking at Fig. 18 we see that this level of gain can be achieved with an emitter resistance of 10 ohms and a shunt feedback resistance of 250 ohms.

For this stage to deliver 1/2 watt of output, the dc input power must be at least one watt. In chapter 2 we found that the maximum efficiency which could be obtained from a Class A amplifier is 50 percent. Part of the supply voltage will be taken by a voltage drop across the 10-ohm emitter resistance. Hence, assume that the net supply available is 10 volts, to be placed across the transistor. This means that the current in the transistor will need to be at least 100 mA. To be on the safe side, we will bias it to 135 mA. Using the equation which relates output intercept to standing current in the transistor (presented in chapter 6 in connection with receiver front-end amplifiers), we would expect this circuit to have an output intercept of +40 to +43 dBm. If the number was +40 dBm, and the output power was 1/2-watt PEP (+27 dBm), the output power in each tone would be +21 dBm, yielding IM products that were 38 dB down. Such performance is reasonable to expect.

Note that an rf choke is used to feed dc to the collector, and that another is used in the base-bias circuit. The choke is helpful in the latter case to prevent the input impedance of the output stage from being suppressed by the 100-ohm resistor in the bias divider. Also, since a 500-ohm resistance is needed in the bias divider, but only 250 ohms were required for rf feedback, part of the bias divider is bypassed.

Assume that a net gain of 30 dB was required from the amplifier. A 10-dB gain is provided by the output stage,

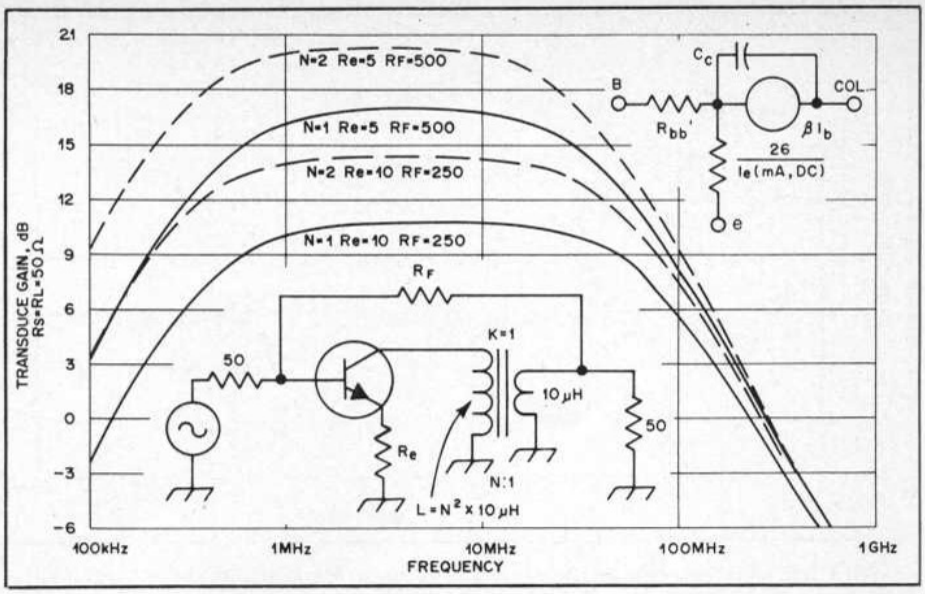


Fig. 19 — Transducer gain vs.frequency for amplifiers using transformers in the collector circuit. The hybrid pi model for a bipolar transistor was used in these calculations. The transistor specifications were the same as those in connection with Fig. 17. The low-frequency decrease in gain results from transformer characteristics.

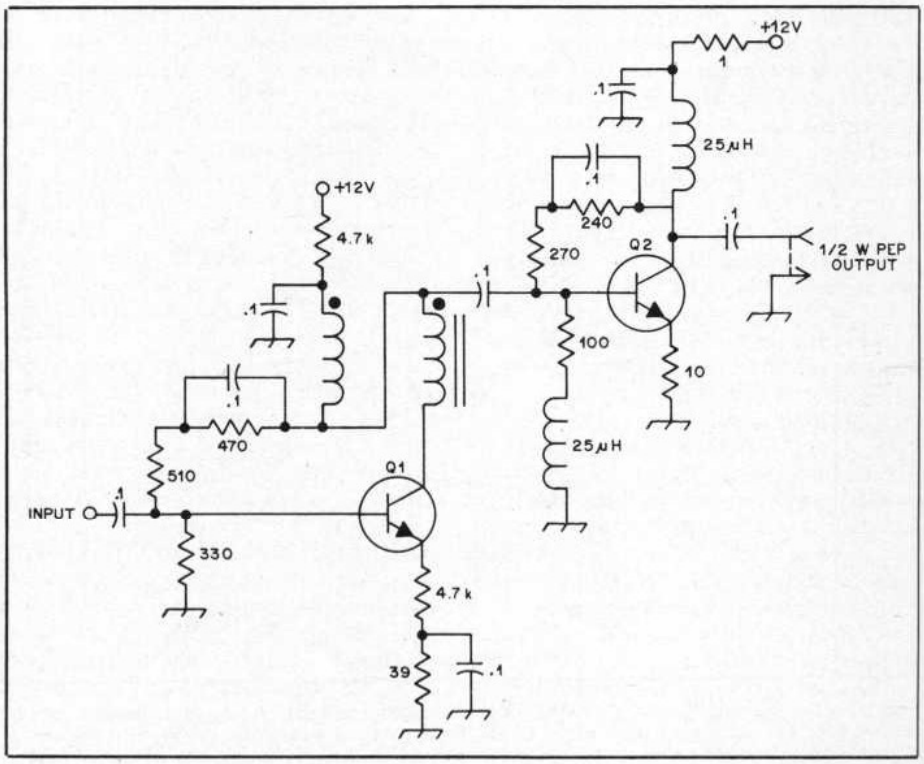


Fig. 20 — Example of a 30-dB gain broadband amplifier with 0.5 watt of PEP output (see text).

leaving 20 dB required from the driver. The output power required from the driver is only 50-mW PEP, or +11 dBm per tone in a two-tone test. If the IMD ratio for this stage alone is to be 40 dB, the output intercept should be +31 dBm. An amplifier with a standing current of 50 mA should provide this performance. Looking at the curves, we see that the needed gain can be provided with a 2:1-turns-ratio transformer in the collector, with a 5-ohm emitter resistance, and a 500-ohm feedback resistor. The other resistors in the circuit are chosen to provide the proper bias current for the transistor.

An almost identical amplifier is described later as a construction project. The major difference is that the construction project amplifier delivers 1-W PEP of sideband or 1 W of cw output.

Various transistors may be used in amplifiers of this sort. Because of the heavy feedback employed, detailed transistor characteristics are not of great importance. The f_T of the devices

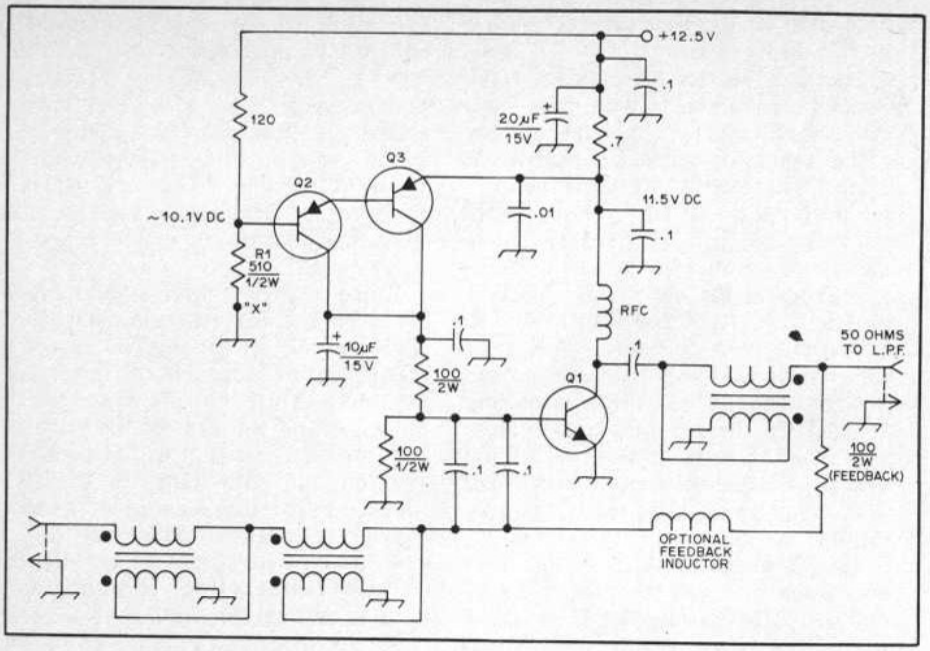


Fig. 21 — Suggested circuit for a 5-watt output Class A power amplifier. Q1 is an MRF449 or similar rf power transistor. Q2 is a 2N4037, and Q3 is an MJE105. Similar devices may be substituted for Q2 and Q3 (see text).

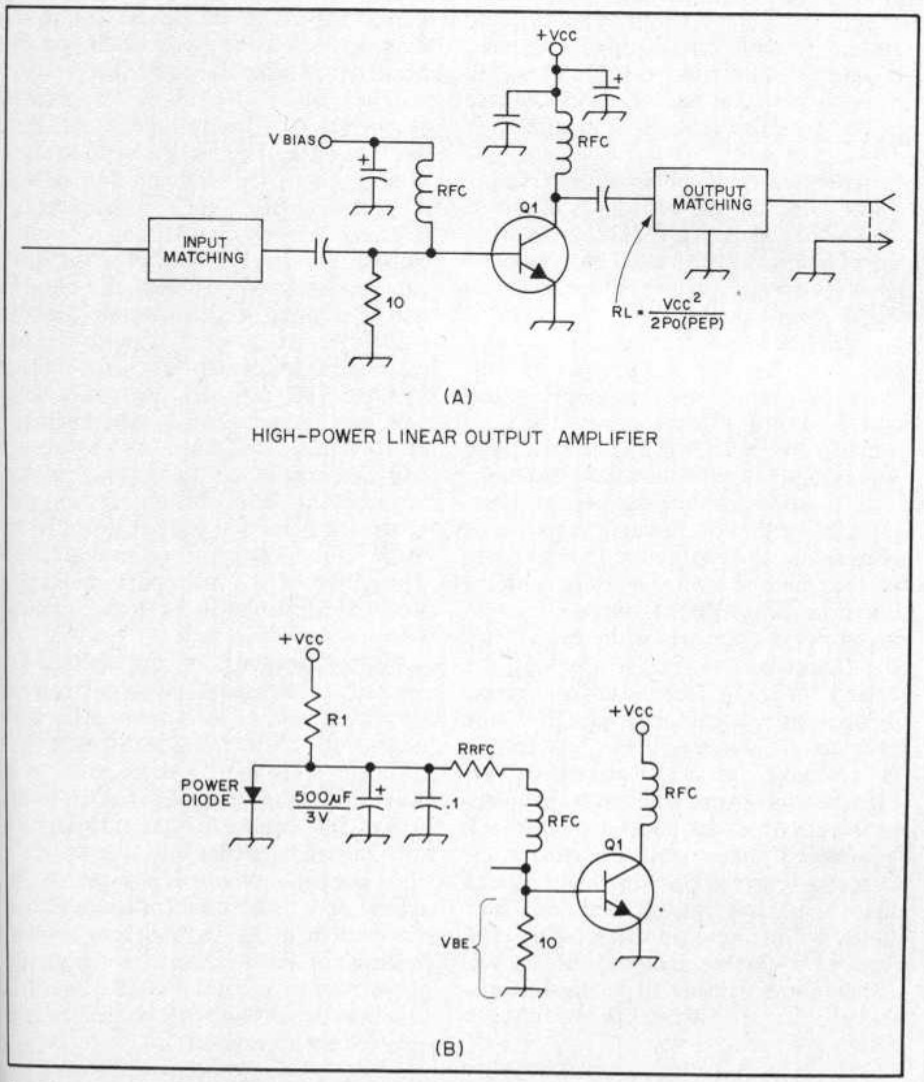


Fig. 22 — Generalized schematic of a single-ended high-power Class AB rf amplifier. The rf circuitry is presented at A, while B emphasizes the details of the biasing circuit.

should be at least 10 times greater than the highest frequency of operation. Also, the transistors should have sufficient power dissipation. Amplifiers of this kind are much different than the Class C amplifiers used for cw. The current in a Class A amplifier is constant, independent of the power output. Hence, the designer does not have the advantage of a low duty cycle that helps him when building cw rigs of similar power output. The writers have used the 2N3553 for output stages at this power level in the hf bands. Although they have not been investigated experimentally, some of the transistors designed for the output of citizens band transceivers should be ideal for these applications. Devices worth consideration would be the Motorola MPS-U31 and MRF472. These parts are relatively inexpensive. In any case, careful heat sinking is required because of the high power dissipation.

If it is desired to extend the bandwidth of amplifiers of this kind to higher frequencies, there are a few tricks that may be employed. From the curves it is evident that the widest bandwidths occur with the lower gain numbers. Because of this, a lower gain per stage will lead to increased bandwidth. Another trick that works well is to place a small inductor in series with the collector. This will increase the voltage swing on the collector at the upper frequencies while leaving the lower frequency gain unaltered. Values as low as 50 to 100 nanohenrys are suitable for vhf work. Similarly, some inductance in series with the shunt feedback resistor will peak the high frequency gain. Finally, some impedance matching can be done. This would take the form of a pi or L type of network as an interstage coupling element. It should have a Q near unity, and should be tuned at the upper frequency of operation. It will then appear virtually "transparent" at the lower frequencies.

It is sometimes desirable to run a Class A amplifier at even higher powers, although the power dissipations encountered may make the thermal designs difficult. Also, the high collector currents may make it difficult to use much emitter degeneration. This places the burden of feedback on the shunt element. Without a large emitter resistance, biasing will also be more cumbersome. A sample circuit is shown in Fig. 21. This amplifier is biased for a current of 1 A. With a 12.5-volt supply the input power will be a little over 10 watts. The value of V_{cc} is less than 12.5 owing to the voltage drop across the collector resistor that is used as a current-sensing element for biasing. A 2:1-turns-ratio transformer is used at the output, transforming a 50-ohm termination to a 12.5-ohm load at the

collector. Because of the lack of emitter degeneration, the input resistance will be quite low. The base is matched with a composite 16:1 impedance ratio transformer made from two "sortabaluns." Although the writers have not built this amplifier, it should be capable of providing about 5 watts of output throughout the hf spectrum with excellent IMD and high gain. For standby periods, or even keying, the circuit may be shut down by breaking the circuit at the point marked "X." It may be necessary to adjust R1 slightly to obtain 1 A of collector current. A large heat sink should be used at Q1. It would be advisable to provide some heat sinking at Q3 as well.

High-Power Linear SSB Amplifiers – The Biasing Problem

For output powers exceeding 1 or 2 watts, the Class A amplifiers outlined are not generally desired. The efficiency is too low, considering that the power must be dissipated on a continuous basis during the total transmit period. For the higher powers the more typical approach is to use a Class AB amplifier.

Shown in Fig. 22 is a circuit for a typical linear amplifier for ssb service. No details are presented as to component value, for these will vary greatly with the frequencies of operation and the power levels desired. However, all of the circuits for this purpose follow the general form shown.

In most ways the rf part of the design is exactly the same as was presented for cw amplifiers in chapters 3 and 4. The output network should be designed for the peak-envelope output power and not the average power. That is, under two-tone testing conditions at a given PEP level, the average power will be half the PEP. The output load presented to the collector is well approximated by $R_L = V_{cc}^2 \div 2P_{out}$. However, the power to use in the calculation is the PEP. If the network were designed for average power, the amplifier would be voltage-limiting, leading to severe distortion of the flat-topping variety.

The input resistance, input capacitance and output capacitance are well specified for most transistors designed for ssb power service. The networks are designed accordingly. The methods outlined in earlier chapters may be used, with narrow-band or broadband transformers being suitable.

The major difference between the cw power amplifiers and the ssb amplifier is in the biasing. If a cw amplifier were to be used for ssb service, severe distortion would occur. This would be most apparent at low levels. This is because the output transistor is cut off when there is no drive. The drive must be large enough to turn on the emitter-base junction to about 0.7 volt before any rf output occurs.

The usual corrective method is the application of some forward bias. This establishes a quiescent operating current in the transistor when no rf drive is present. The base is already turned on, and the application of rf drive merely increases the base current. The dc collector current will increase accordingly. Unlike the case with Class A amplifiers, the transistor is not biased to full current on a dc basis. The level of quiescent current will depend upon the specific transistor used and is usually specified by the manufacturer. Values range from 15 to 100 mA. Probably the most informative reference is by Hejhall (*QST* for March, 1972 and Motorola *AN-546*).

Fig. 22 shows a sketch of the usual biasing scheme used for this class of amplifier. The basis of the biasing is a diode: High-current type is the common choice. The transistor base bias should be chosen to deliver the desired quiescent current in Q1 under no-drive conditions. However, the bias should not vary more than 0.1 volt for all rf drive conditions. This means that the dc current standing in the diode (supplied through R1) should be larger than the peak current that will occur in the base of the transistor at times of maximum rf drive.

The biasing of the amplifier is sometimes aided by the resistance of the rf choke that isolates the bias diode from the rf energy at the base. This resistance allows a voltage divider action to occur which allows the bias diode to be run at a larger standing current than it would if the rf choke had no resistance. This extra current is used to supply base current during rf input peaks. The large bypass capacitor (500 μF) also helps to supply base current on a transient basis.

The problems outlined here are complicated further by the thermal-runaway phenomenon. If Q1 were in a virtually perfect thermal environment (a constant junction temperature), there would be no problem. This is not the case. When the transistor has rf drive applied for a period of time, the resultant power dissipation will cause the junction temperature to increase. If the bias voltage is constant (as was advocated) the higher temperature will cause the quiescent current to be larger when drive is removed. If the increase is excessive, the collector current will be high enough that high-power dissipation will continue within the transistor. This will lead to a further increase in junction temperature, causing an increase in quiescent current. Thermal runaway is the ultimate result.

There are solutions to this problem that are partially effective. One is to thermally bond the bias diode to the transistor. This causes the increase in transistor temperature to be communicated to the diode. Most silicon diodes which are fed from a constant-current source will show a voltage change with temperature of about –2 millivolt per degree C change. The negative sign indicates a decreasing voltage with increasing temperature, which is just the effect needed.

Unfortunately, thermal bonding of the reference diode to the transistor is only partially effective. The reason is that the diode is capable of sensing only the temperature of the case of the transistor and not that of the junction. The thermal resistance between the junction and the transistor case will allow the junction to run at a much higher temperature than that of the case. It is the junction temperature that controls current flow and ultimately leads to thermal runaway.

One protective method is to include a diode within the transistor body for temperature sensing. The anode of the diode is brought out to a separate pin on the transistor and is used as a reference for a dc amplifier that provides bias for the transistor. The reader is referred to the work of Chang and Locke (RCA note, *AN-4591*).

The other technique is emitter degeneration. This can be external to the transistor. The more common situation is where the degeneration is built into the device. Such transistors are referred to as containing "emitter ballasting." The advantage of the internal degeneration is that the emitter resistance may be distributed over the entire transistor structure with a separate resistance element for each emitter section. The resistors are made from nichrome, which has a high-temperature coefficient of resistance. As a result, if a given section of the transistor begins to increase in temperature faster than others, that section is shut down faster. Such "hot spots" lead to second breakdown, one of the main phenomena that leads to destruction of power transistors.

The experience of the writers suggests that transistors without internal ballasting must use external emitter degeneration if thermal runaway is to be avoided. This may not be vital in an amplifier to be used only for ssb service where the average power dissipation is low (because of the low duty cycle of the human voice). However, if the amplifier is to be used for cw operation, or even if it is to undergo two-tone testing for linear service, some emitter degeneration must be used. Usually a fraction of an ohm will be sufficient to protect the transistor.

In any case where emitter degeneration is used, either in the form of ballasting or as external degeneration,

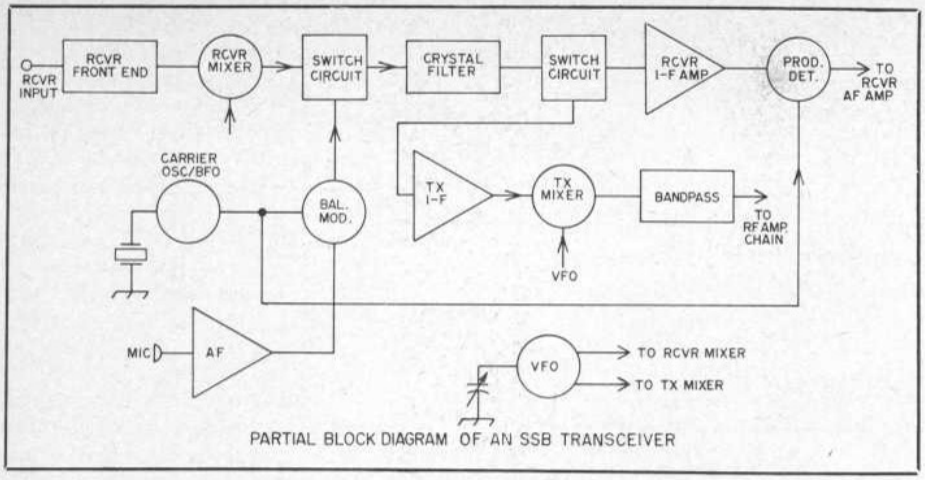


Fig. 23 — Partial block diagram of an ssb transceiver. The system differs from an ssb transmitter in the inclusion of switching circuits and the multiple use of the carrier oscillator — BFO and VFO.

the resistance will cause the efficiency of the amplifier to be degraded. Also, it can have the effect of degrading the stability of the amplifier. Unconditional stability can sometimes be regained through the application of shunt feedback, at the cost of reduced stage gain.

Modern transistors designed for high-power linear rf applications have excellent IMD specifications. Typically, third-order distortion products are 30 dB or greater below each tone during a two-tone IMD measurement at full output power. The distortion does not behave as nicely with such amplifiers as it does with a Class A design. With the Class A amplifier, an output intercept can usually be specified for a given circuit. This defines the IMD performance of the amplifier at all power levels. Specifically, if the output power is decreased by X dB, the IMD ratio will improve by $2X$ dB. Class AB amplifiers are not as well behaved. When the output power is dropped from the specified maximum, the IMD ratio can degrade. For this reason, the best mode of operation is at full rated power. If a low-level output is desired (for QRP experiments or driving vhf transverters), an attenuator should be used. Alternatively, the high-power final amplifier should be bypassed, with the output signal being obtained from an earlier Class A stage in the amplifier chain.

One problem with the diode biasing scheme is the high current required to bias the diodes properly. This current is often obtained from the same power supply that is used for the collector bias. Most of the power used to derive the bias current is dissipated in the large resistor (R1 of Fig. 22). This will degrade the system efficiency considerably from that value given by the manufacturer. There are at least a couple of solutions to this problem. One would be to use a separate power supply for biasing the diode. The cost of a 5-volt supply would be small. Another solution was suggested to the writers by W7UDM: Use the current that is standing in a previous Class A amplifier to also bias the diode. The power is then used more effectively. Careful decoupling would be required.

No construction examples of high-power linear amplifiers are given in this chapter. However, some were presented earlier. They were designed for ssb service.

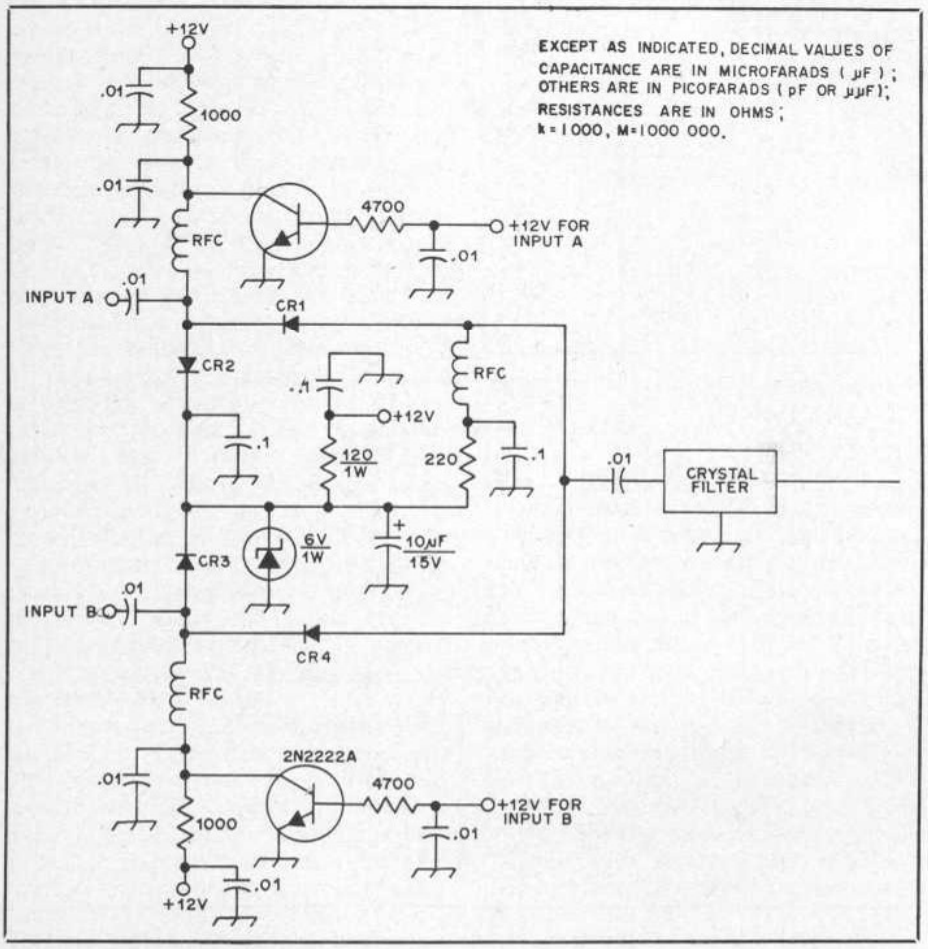


Fig. 24 — A method for diode switching a crystal filter. Only the input is switched in this example. A similar circuit could be used at the output.

Transceivers for SSB

Although some operators use separate transmitters and receivers for ssb, the trend is toward transceivers. The major reason is convenience of operation. With an ssb transceiver, once a station is tuned in so that it sounds proper in the receiver, the transmitter is automatically on the proper frequency. Another reason is that much of the transmitter and receiver circuitry can be shared, leading to economy in construction.

Shown in Fig. 23 is a partial block diagram of a single-conversion ssb transceiver. The carrier oscillator used for ssb generation at the i-f is used also as the BFO for the receiver. It is not mandatory that this signal be switched. It may be applied to both inputs simultaneously. However, great care should be taken to ensure that minimal energy from the carrier oscillator finds its way into the receiver i-f amplifier. This avoids the noise-modulation problems which were reviewed in the receiver chapters.

The VFO is also shared. Again, this signal may be applied to each of the

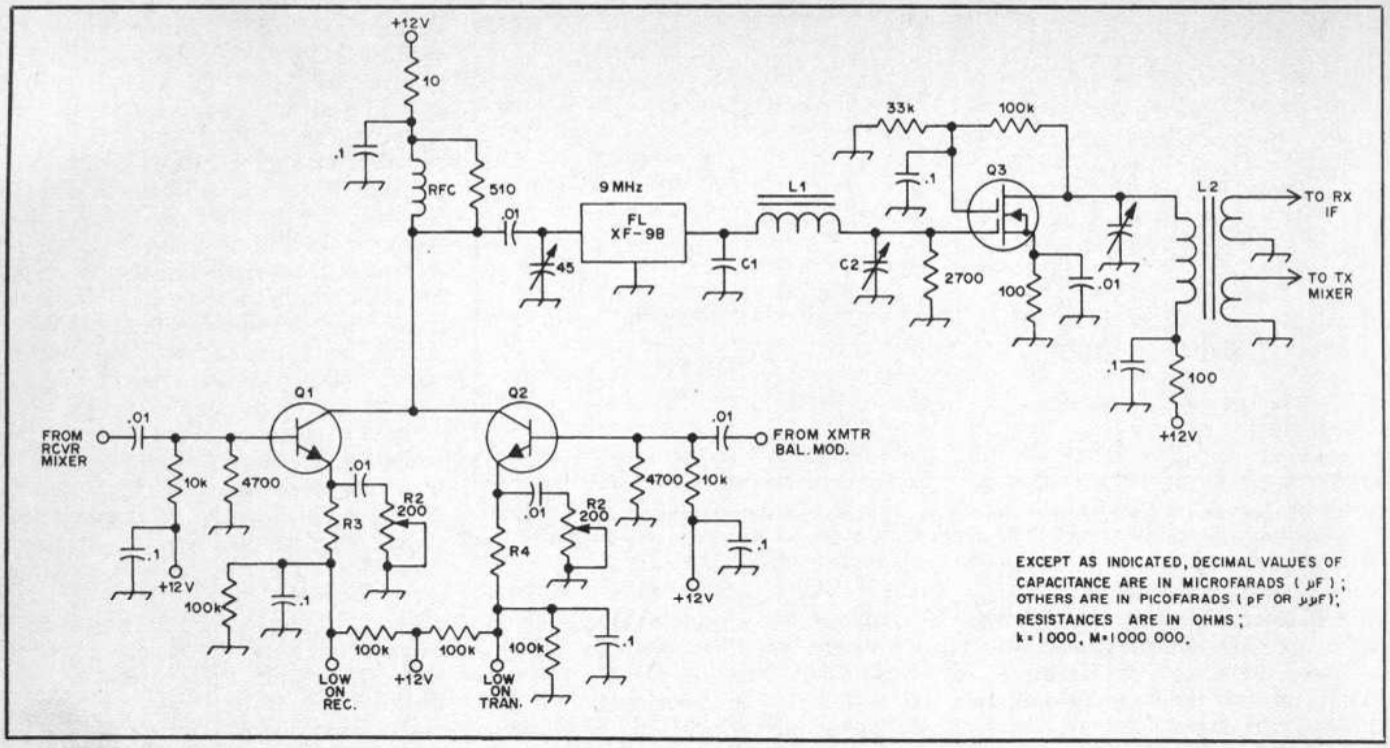


Fig. 25 — Circuit for sharing a crystal filter between receive and transmit functions in a transceiver. Bipolar transistors are used at the input, while a dual-gate MOSFET is employed at the output.

mixers simultaneously. If diode-ring mixers are used, it may be necessary to buffer each mixer input separately to ensure that proper LO injection levels are maintained.

The third major component that is shared between the two functions is the crystal filter. It is usually necessary that switching be provided at at least one end of the filter, if not both. One solution would be to use diodes as the switching elements. A sample circuit is presented in Fig. 24. Only one side is shown, although the other side would be similar. Four diodes are used in this scheme. If input A is selected, CR1 will be conducting a dc current of about 25 mA. CR2 is reverse biased by 6 volts. At the same time, the off channel (input B) is shunted to ground with CR3 which is conducting approximately 6 mA while additional isolation is provided by CR4 which is back biased with 6 volts. The diodes may be 1N914 switching types for casual applications. However, better performance will probably be provided by using PIN diodes or low-speed high-voltage rectifier diodes. The reader is referred to the i-f amplifier discussion in chapter 5 for details.

Another approach to filter switching is shown in Fig. 25. A pair of bipolar transistors is combined with a common collector connection to feed a 500-ohm crystal filter. The collector current in each transistor is determined by picking R3 and R4 appropriately. Small 200-ohm controls are used at R1 and R2 to

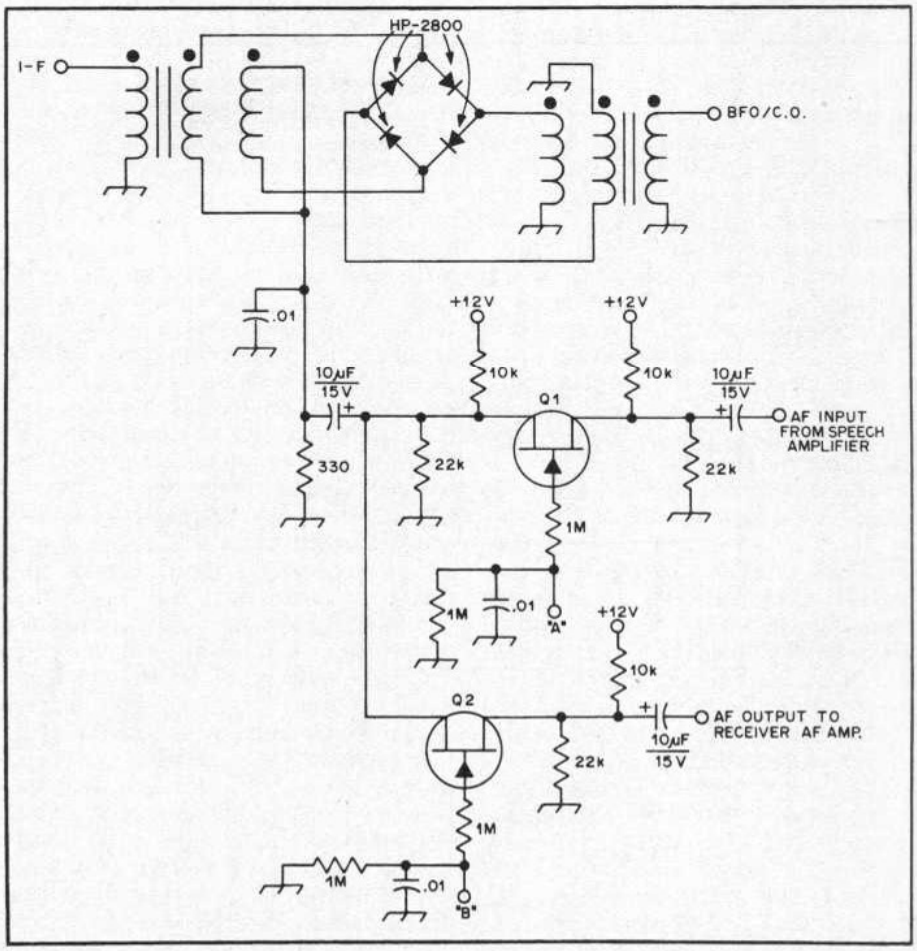


Fig. 26 — Application of a diode ring as a balanced modulator during transmit periods, and a product detector for receiving. FETs are used for switching the audio. Q1 and Q2 may be general-purpose FETs such as the MPF102.

establish the gain of each stage. The output of the filter is applied to a pi network consisting of C1, C2 and L1. This network should be designed for a Q of 10 to 15, with resistances of 500 and 2,700 ohms. The 2,700-ohm resistor at the gate of Q3 ensures that the crystal filter has a proper termination. The output of the MOSFET amplifier is tuned to 9 MHz with a low-Q circuit. Two output links are used. One drives the receiver i-f amplifier while the other is applied to the transmit mixer.

An innovative and unique means for ssb transceiver design is through the use of bidirectional circuits. These are circuits that will function with signals flowing in either direction. One example that has been discussed in detail is the diode-ring mixer. An example is shown in Fig. 26 with a circuit that functions both as a balanced modulator during transmit periods and as a receiving product detector. JFETs are used as switches at the audio port. Point "A" should be high (+12 volts) during transmit periods and point "B" positive for receiver operation.

Shown in Fig. 27 is an amplifier that will provide gain in either direction. The direction is controlled by choosing which power-supply port is activated with +12 volts. Each transistor is biased for a current of approximately 35 mA, which is enough to yield good IMD performance. A 2:1-turns-ratio ferrite transformer is used in the output of each collector in order to obtain some impedance matching. However, this could be eliminated if lower gain is desired. Provision is made for the use of both shunt and series (emitter) feedback. Again, depending upon the gain desired, one or the other may be eliminated. Some shunt feedback would be desirable in order to preserve stability, since the transistors specified have an f_T of over 1 GHz. It is important that the two stages share a common emitter resistance as part of the dc biasing scheme. This will ensure that the "off" transistor has both of its junctions reverse biased. This circuit is an adaptation of one designed by W7UDM.

Bidirectional circuits are ideal for driving the rf and i-f ports of a diode-ring mixer. When used in this way, the only switching required would be that for controlling the direction of the amplifiers.

Shown in Fig. 28 is a partial block diagram of a possible ssb transceiver that could be built with diode-ring mixers and bidirectional amplifiers. If desired, a PIN diode attenuator or two could be inserted in the signal path for control of gain in both modes. Techniques of this kind have been used in a commercially built multiple-conversion transceiver. However, the amplifiers used germanium transistors biased for minimal current. Many of the mixers were designed similarly. The dynamic range of the system was disastrous! On the other hand, using these concepts a good 20-meter ssb transceiver has been designed and built by W7UDM. By using diode-ring mixers with proper LO injection and amplifiers with adequate current, and by using single conversion, a receiver dynamic range of 90 dB has been obtained. The advantage of this scheme is that virtually all of the filtering in the system can be used for both transmit and receive. This is highly desired. In any transceive system, it is advisable to run the received signal through the low-pass filters that will be needed for harmonic filtering of the power amplifiers.

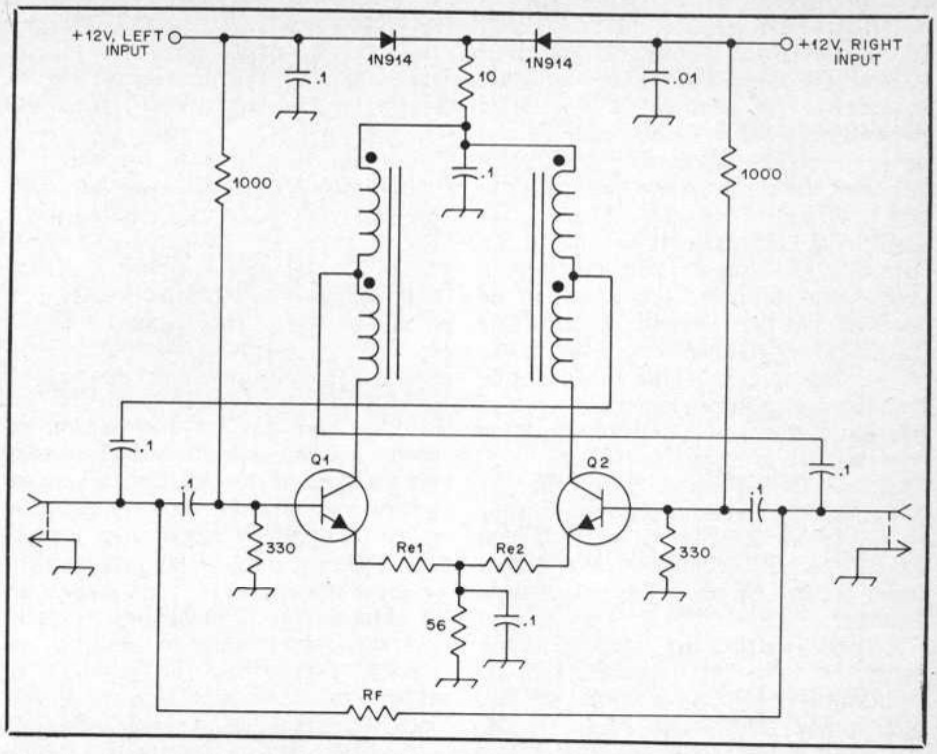


Fig. 27 — Circuit for a bidirectional amplifier using bipolar transistors. Q1 and Q2 are 2N5943 or similar devices with a high f_T.

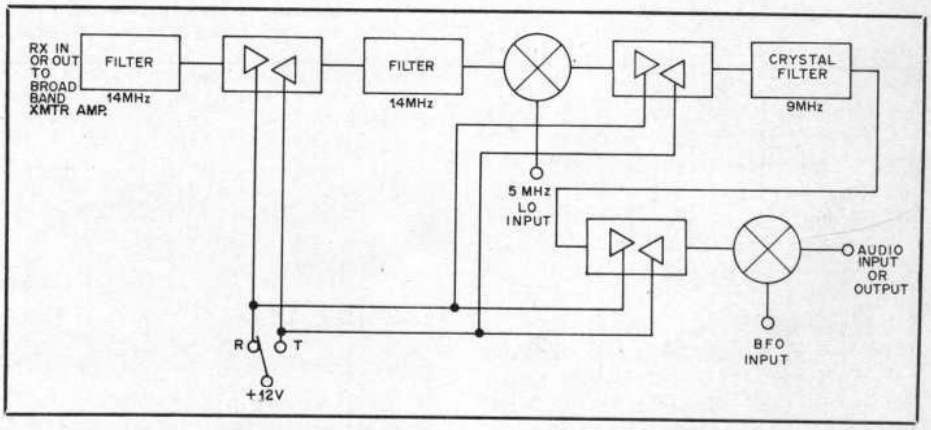


Fig. 28 — Partial block diagram of an ssb transceiver based upon bidirectional circuits.

Double-Sideband Transmitters

In the earlier theoretical discussion we treated suppressed-carrier double side band as a intermediate step toward generation of an ssb signal. While this is normally the case, there are many situations where a double-sideband transmitter is quite useful. An advantage of dsb over ssb is simplicity. The major disadvantage is that extra spectrum is occupied. Sometimes, the tradeoff may favor the use of dsb.

One application for dsb that comes to mind is for the QRP enthusiast. Often he has an interest in working phone, but has little interest in building a complete ssb transmitter. Dsb gives him an alternative. Another point in his favor is that transceivers are built easily

to utilize the VFO which is already present in a direct-conversion receiver. All of the normal advantages of a ssb transceiver (in contrast to a separate transmitter) are available. Specifically, once a station is tuned in with the receiver, the transmitter is automatically on the same frequency. There is an additional advantage: If an unused frequency is found with a direct-conversion receiver, the user can be assured that the segments on either side of his carrier frequency are unoccupied. If he were to call CQ, he would not be causing undue interference as a result of his extra sideband. Additionally, if an ssb station is copyable with a "dc" receiver, the operator knows that he may call that station without causing QRM to an adjacent channel. If that channel were occupied, it would have been heard in the direct-conversion receiver.

There is a liability with the transceiver using a "dc" receiver and dsb transmitter. Two such units are not compatible with each other. A dsb station is not generally copyable with a dc receiver. This is normally not a problem for the QRP operator, for most of his contacts are with higher-power ssb stations. Because of the spectrum used, dsb is not recommended for use on the hf bands except at low powers. A maximum limit might be 10-watts PEP output.

Another application for the dsb transmitter would be for the DX-oriented vhf operator. He often has a desire to converse with local ssb operators with common interests. For such purposes low power is usually sufficient. When band openings occur and the more distant contacts are available, he switches to cw to ensure the contact.

The vhf dsb station again has the liability that half of his transmitted power is in an unwanted sideband. On the portions of the vhf bands where ssb and cw predominate, the extra spectrum space occupied by dsb is rarely a problem. The mountain-topping vhfer might consider it wasteful to throw away 3 dB of extra energy from his battery pack. However, if he were to examine the current that would be required to remove the extra sideband, the difference becomes much less significant. This is especially true for the portable station running less than 1 watt of output.

A final advantage of building a dsb vhf transmitter is that it is expandable. The oscillator (and multiplier chain, if used) can always be adapted for use with a later ssb exciter. The balanced modulator and speech amplifier may also be used later with some modification to another frequency. A linear amplifier chain designed specifically for a dsb transmitter may be used directly with a later ssb replacement.

A Simple DSB Transmitter for Six Meters

Shown in Fig. 29 is a simple QRP transmitter for 6 meters. A third-overtone crystal oscillator is operated directly at the output frequency. This circuit delivers about +10 dBm of drive to the balanced modulator. The balanced modulator is simple, using two hot-carrier diodes driven from a ferrite transformer. This circuit uses a pair of variable capacitors for adjustment of the carrier balance. Over 50 dB of carrier rejection was measured with this circuit on an open bench when driven from a separate signal generator. In the transmitter shown the carrier suppression is less – only 36 dB. This is because the signal from the crystal oscillator is leaking around the balanced modulator to the amplifier chain. Some additional

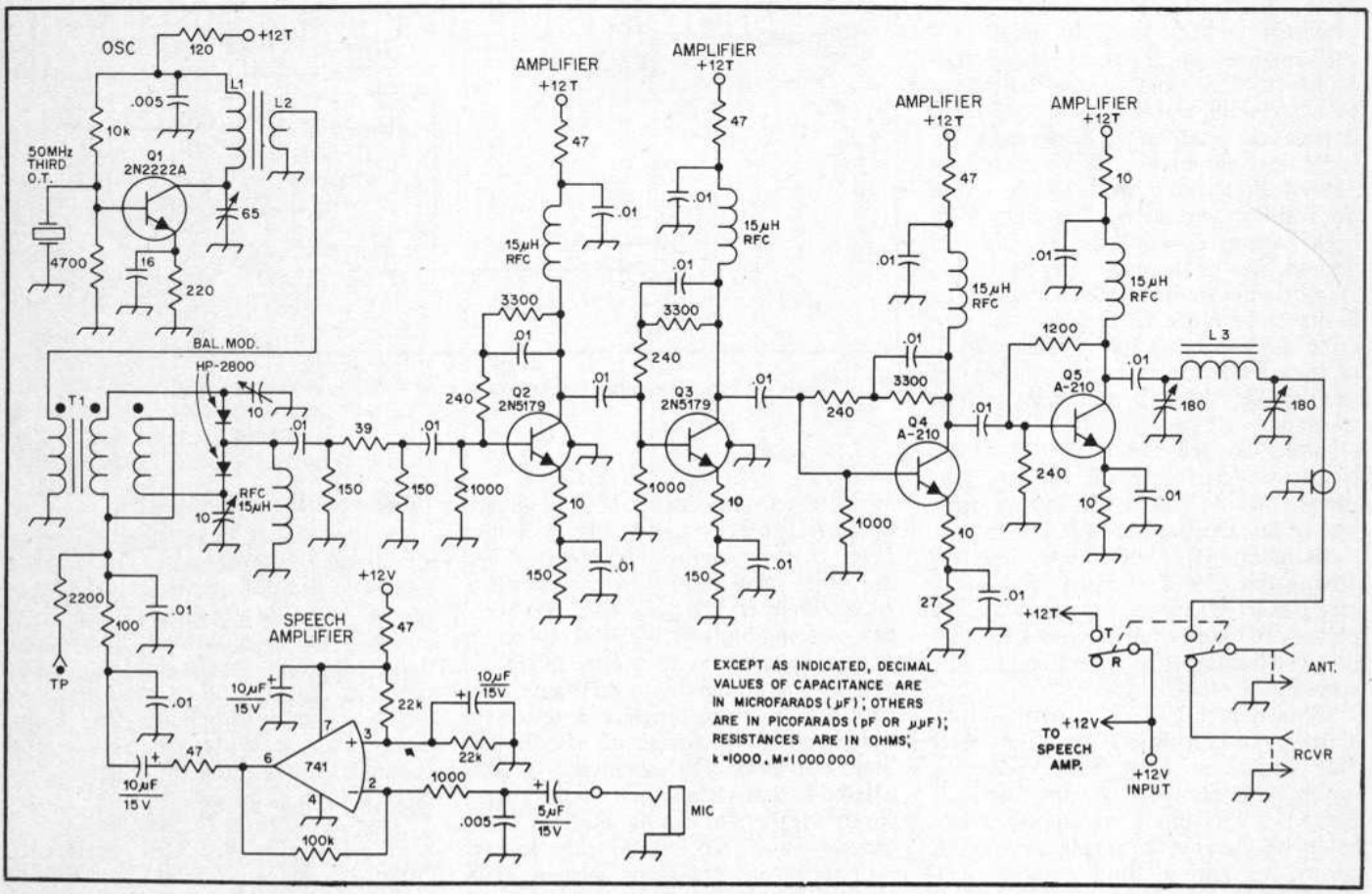
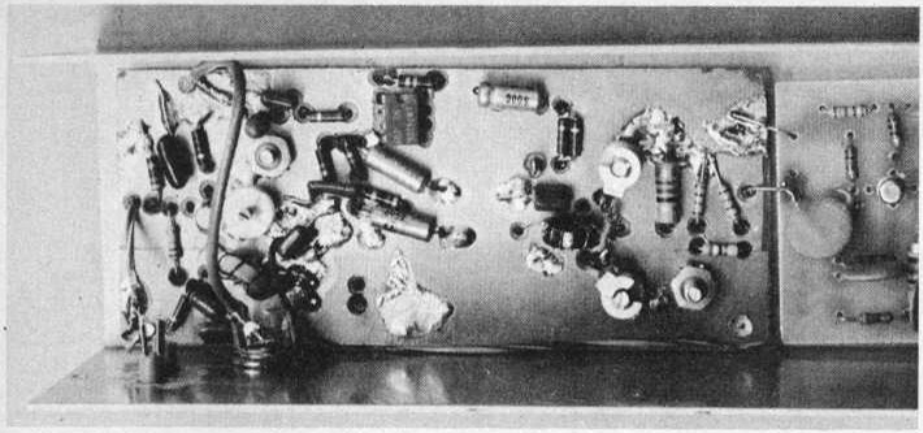


Fig. 29 – Circuit for a 6-Meter dsb QRP transmitter (see text). T-R switching is realized with a double-pole, double-throw slide switch.
L1 – 10 turns No. 24 enameled wire on Amidon T37-6 toroid core.
L2 – 2-turn link over L1.
L3 – 6 turns No. 22 enameled wire on Amidon T50-6 toroid core.
T1 – 10 trifilar turns No. 30 enameled wire on Amidon FT-37-61 toroid core.



Interior view of the 50-MHz dsb transmitter. The crystal oscillator, speech amplifier and balanced modulator are on this circuit board.

isolation would solve this problem.

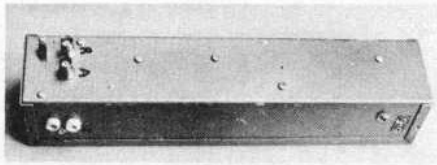
The speech amplifier consists of a single 741 operational amplifier. The feedback resistors were picked to produce a suitable output level while using a microphone from an inexpensive cassette tape recorder.

A test point is provided in the balanced modulator. If +12 volts are applied to this resistor, the circuit is unbalanced, yielding a carrier output for test purposes and alignment. If the transmitter is to be used on cw, this point could be keyed to the +12-volt supply with a pnp switch. In these applications, it would be wise to also key the supply to the linear-amplifier chain.

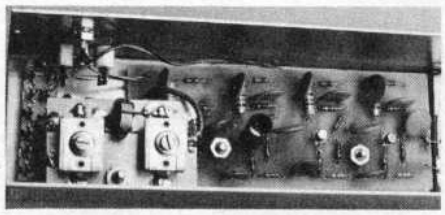
The linear amplifier uses four stages with an output of 400-mW PEP. The first three stages were designed for 10 dB of gain per stage, with heavy negative feedback being employed in each stage. The output has shunt feedback, but there is no emitter degeneration. Because of this the gain is not as flat with frequency as it is in the preceding stages. A 6-dB attenuator is used at the input to the amplifier chain to ensure that a proper termination exists for the balanced modulator.

The first two stages in the amplifier chain use 2N5179 transistors. These devices have an f_T of 1 GHz and a low collector-to-base capacitance. They are recommended for general-purpose vhf use. The driver and output amplifier use Amprex A-210s. This transistor is rugged and has an f_T of 1200 MHz. A suitable substitute would probably be the 2N3866 or the 2N3553. Since the standing current is moderatly high (over 100 mA in Q5), heat sinks are needed for Q4 and Q5.

A small piece of double-sided pc board was used for the crystal oscillator and the balanced modulator. The top side, where the components reside, was



External view of the 50-MHz dsb transmitter. The slide type T-R switch is adjacent to the BNC connectors which are used for the antenna and the line to the receiver. Power receptacles are also close to this switch. The crystal socket and microphone jack are at the opposite end of the chassis.



The driver is the transistor with the small heat sink. A slightly larger heat sink is used on the output amplifier, which is hidden below the small board containing the output network.

used as a ground plane. The amplifier chain was built on single-sided board. The extensive use of feedback makes ground-loop problems less severe. The board was originally etched as a general-purpose instrumentation amplifier (described in chapter 7) which dictated the circuit configuration. If higher-gain circuits were used, employing 2:1-turns-ratio transformers in the outputs of the low-level stages, it would be possible to obtain the needed gain with three stages. The present amplifier has a small-signal gain of 45 dB at 50 MHz.

It should be straightforward to adapt this circuit to any of the lower bands. The bypass capacitor at the emitter of Q5 should be removed in order to drop the gain accordingly. The crystal oscillator and the simple pi network output would be replaced with suitable circuits. If the amplifier is to be used on the 160-meter band, it would be advisable to increase the inductance value of the rf chokes to around 50 μH. The output network is adjusted for maximum output with the test point set to +12 volts. The output should be monitored in a high-frequency oscilloscope for flat topping (if such an instrument is available). Good results have been obtained with this transmitter.

A DSB/CW Exciter for 144 MHz

Experience with the 6-meter QRP dsb transmitter was encouraging. A similar unit was built for the 2-meter band. A number of refinements were included for operational convenience and to test a number of experimental ideas. The circuit for the transmitter is shown in Fig. 30.

While crystal control is adequate for some operations, flexibility in frequency coverage is highly desirable. There are a number of ways to achieve this at vhf. The usual one is to use a heterodyne type of transmitter circuit.

An alternative to a heterodyne exciter is to use a low-frequency VXO and a multiplier chain. While a VFO could have been used, it is quite difficult to obtain suitability for cw and ssb at vhf. A Colpitts crystal oscillator was modified with an inductor and variable capacitor in series with the crystal. With this circuit (Q1), approximately 100 kHz of tuning range in the 2-meter band was obtained. The frequency shift could have been extended farther. (See VXOs in chapter 2.)

The frequency-multiplier chain was unconventional, but highly successful. A frequency of 18 MHz was chosen for the VXO, allowing the 2-meter band to be reached by using frequency doublers. The output of the oscillator is buffered and filtered in order to yield a symmetrical waveform with a power of over +10 dBm. This output was then applied to a balanced doubler which uses a pair of silicon switching diodes. The output of the doubler was filtered in a single tuned circuit, furnishing energy at 36 MHz. This was amplified to a +10-dBm level with a broadband amplifier. The same methods were repeated to arrive at 72 and finally 144 MHz. The 144-MHz output was filtered with a double-tuned circuit, providing power output of +11 dBm.

The output of the multiplier chain was carefully investigated with a spectrum analyzer to evaluate the spurious responses. Only one spur could be found. That was at 72 MHz. It was 55 dB down. All other subharmonic spurs were undetectable. This response is a result of using simple balanced circuitry

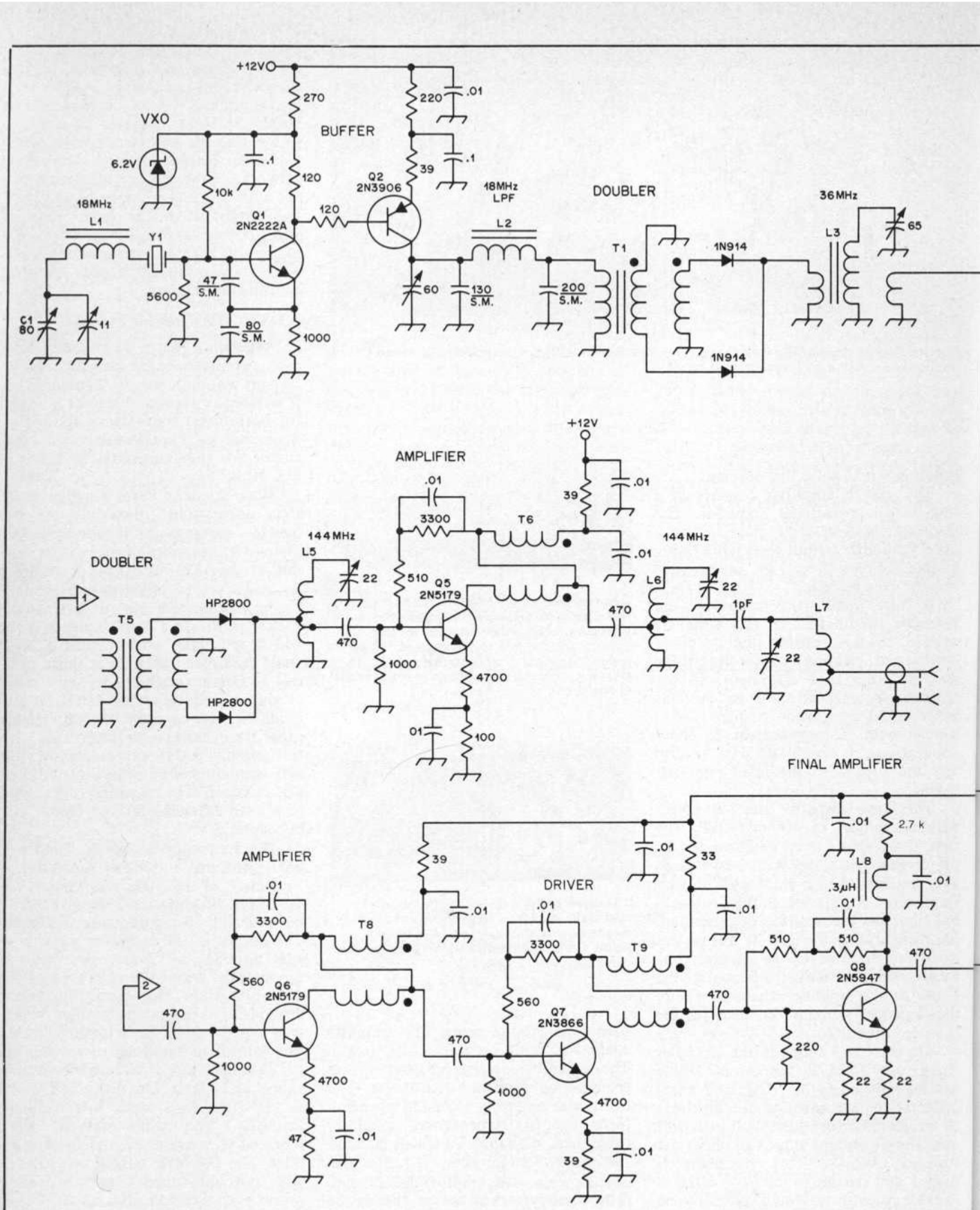


Fig. 30 — Circuit diagram for a 144-MHz cw/dsb transmitter. See text for details. Variable capacitors are air, Teflon, or ceramic-dielectric types. All resistors are 5 percent, 1/4 watt.

C1 — 5-80 pF air variable.

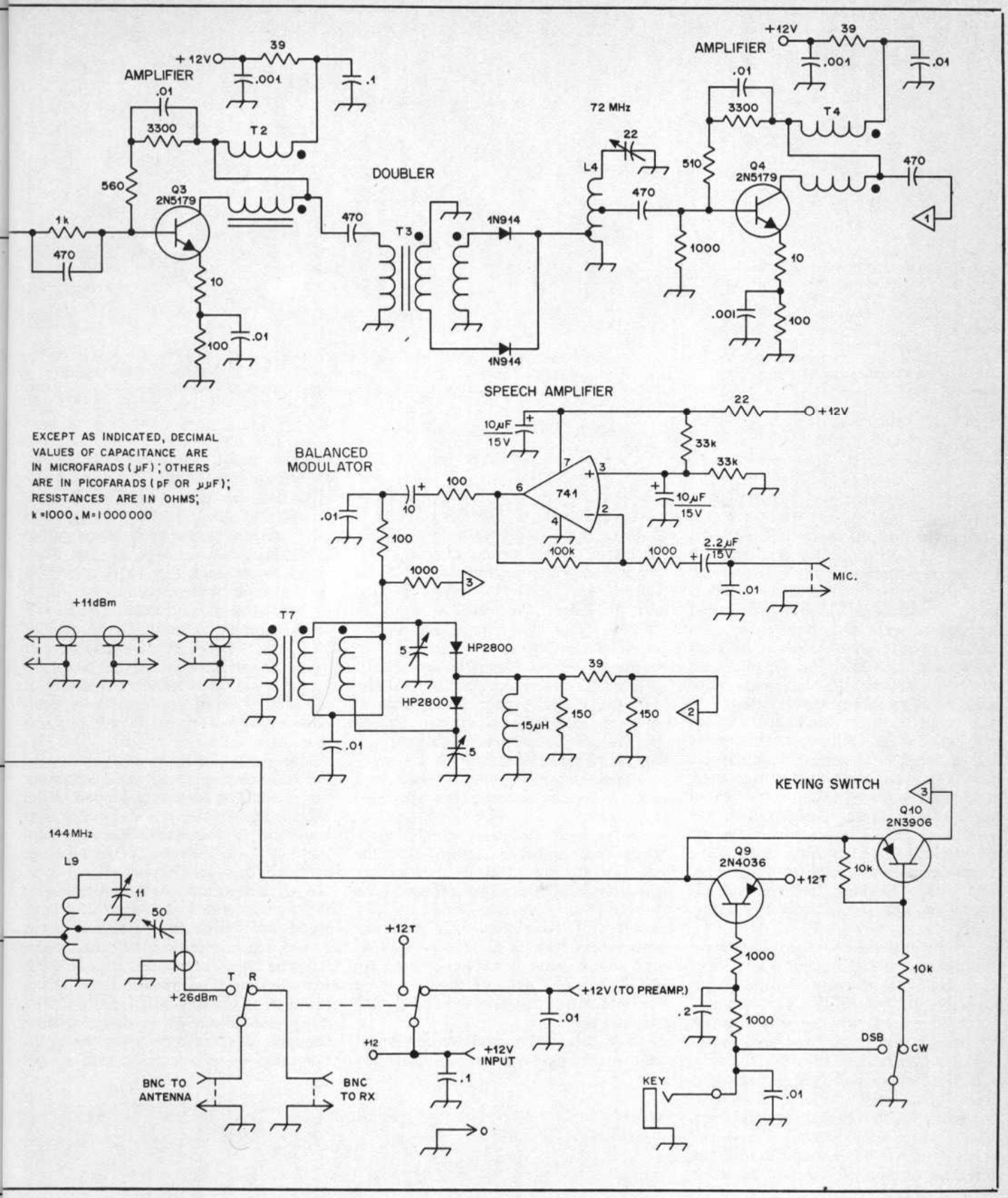
L1 — 24 turns of No. 27 enameled wire on a T37-6 toroid core.

L2 — 14 turns of No. 27 enameled wire on a T37-6 toroid core.

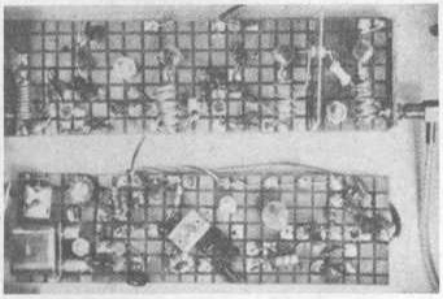
L3 — 12 turns of No. 27 enameled wire on a T37-6 toroid core, 3-turn input link, 2-turn output link.

L4 — Air core, 0.25 ID × 0.65 long (inch). 10 turns of No. 22 enameled wire, taps at 1-1/4 and 1-3/4 turns.

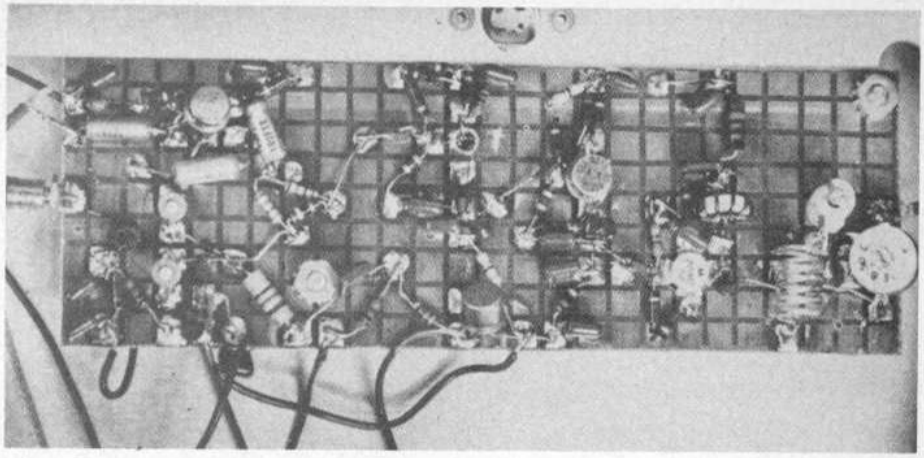
L5 — 5 turns, air core, 1/4 ID × 1/2



long (inch), taps at 1 and 3/4 turns.
L6 — 5 turns, air core, taps at 1 turn and 2-1/2 turns.
L7 — 5 turns, air core, tap at 1 turn.
L8 — 10 turns No. 27 enameled wire on a T37-6 toroid core.
L9 - 7 turns, aircore, taps at 3/4 and 3 turns.
T1, T3, T5, T7 — 7 trifilar turns No. 30 enameled wire on an FT-23-43.
ferrite core.
T2, T4, T6, T8, T9 — 5 bifilar turns No. 30 enameled wire on an FT-23-43 (ferrite) core.



Interior view of the 2-meter transmitter showing the oscillator and multiplier chain. The lower board contains the 18-MHz VXO, a buffer, and the first diode-doubler amplifier combination. The upper board contains two more diode-doubler amplifier combinations and a double-tuned 144-MHz output network. In spite of the open construction, the output of the chain is remarkably free of spurious responses.



This board contains the balanced modulator and rf power-amplifier chain for the 2-meter exciter as well as keying circuits and the speech amplifier. The stud of the 2N5947 output amplifier is attached to a small piece of aluminum which serves as a heat sink.

rather than relying upon shielding or selectivity. It was found that hot carrier diodes gave superior performance in the last frequency doubler. While the output power was sufficient with 1N914s, the 72-MHz component was only 50 dB below the desired output.

Other frequency-multiplier schemes were investigated. While single-ended multipliers were the simplest, double-tuned circuits were required at each frequency in order to keep spurs 50 dB down. Push-push doublers were tried using well-matched transistors. While the suppression of fundamental drive was good, instability problems were encountered in cascading a number of such stages. The diode frequency doublers have been found to be one of the best avenues to follow for frequency multiplication. The broadband amplifiers appear to be unconditionally stable and the tuning is unambiguous. A more exotic filter at the output (L6 and L7) would suppress the spurs by even higher ratios.

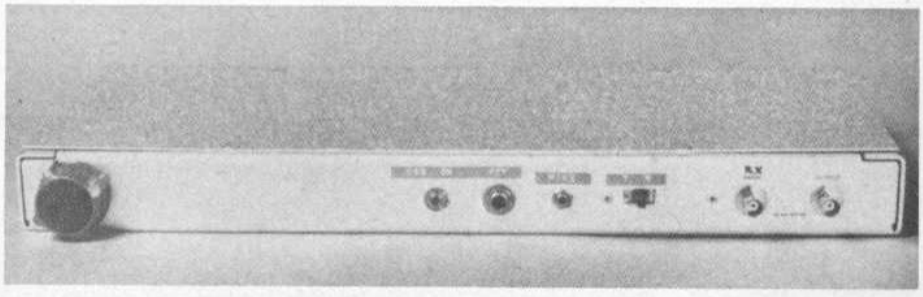
The output of the frequency-multiplier chain is applied to a balanced modulator to generate the dsb signal directly at 144 MHz. The balanced modulator and speech amplifier are virtually identical to those used in the 50-MHz transmitter. The differences are a reduced number of turns on a smaller ferrite core and the use of smaller balancing variable capacitors. The transmitter strip was originally built and adjusted in the home shop. As adjusted, the carrier suppression was 40 dB. When it was adjusted more carefully while using a spectrum analyzer, a suppression of over 50 dB was obtained. Using an outboard signal source (+13 dBm), similar levels were obtained at 14, 28 and 50 MHz. "Retweaking" was required at each band. The carrier suppression was only 35 dB at 220 MHz.

The linear-amplifier chain is similar to that shown for 50 MHz (Fig. 29), although only three stages were used. The input stage, Q6, used a 2N5179 while the driver, Q7, used a 2N3866. Both of these stages were designed for 20 dB of low-frequency gain per stage, and included a ferrite transformer in the collector circuits for matching. The output amplifier, Q8, used a Motorola 2N5947. This stud-mount transistor is specified for Class A linear service. The stage was set for a gain of near 10 dB with a collector current of 120 mA. The collector rf choke is a toroidal inductor. A piece of aluminum with an area of five square inches served as a heat sink for Q8. The weakest link in the transmitter is the output network which used a single tuned circuit. The taps were adjusted for maximum cw output power while using home-lab type equipment. Later measurements revealed that the second harmonic at 288 MHz was only suppressed 20 dB. This presented no problems in operation, since an outboard filter was used. An improved output network is definitely in order and should certainly not be difficult. An *L-C-L* type of T network should provide suitable performance, as would a double pi circuit.

The balanced modulator, 6-dB pad and output network were disconnected and the amplifier was evaluated over a wide frequency range. The gain at 144 MHz was 37 dB, while 48 dB was available at 50 MHz. The gain at 220 MHz was down to 31 dB. Some inductance in the collectors of the three stages would peak this up if operation on that band was contemplated. Alternatively, another low-level 2N5179 amplifier could be used. The gain at 28 MHz was nearly 50 dB. However, at lower frequencies the gain began to drop. This is predominantly because of the 470-pF coupling capacitors used. The output power was 400-mW PEP dsb or cw at 144 MHz.

Cw operation is provided by keying the +12-volt supply to the total amplifier chain. The dc that is applied to the balanced modulator was also keyed. The backwave of this transmitter was measured at –75 dB. An *RC* network is included for shaped keying.

The construction method used for this rig was unorthodox for vhf. A large piece of double-sided pc board was etched to form some breadboard material. The top side was a matrix of copper islands, 1 cm on a side. The back of the board was a continuous ground foil. The same results can be achieved with a hacksaw. The capacitances of the board presented no problems because almost



Exterior view of the 2-meter dsb/cw transmitter. The knob controls the frequency of the VXO at 18 MHz.

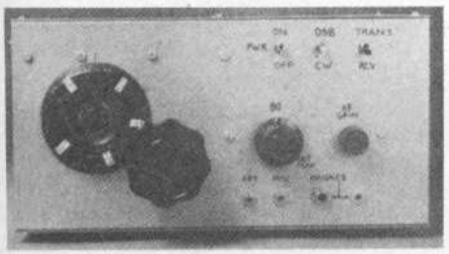
all of the high-frequency circuitry was at a low impedance level. The capacitance of each pad section was less than 0.5 pF. Holes may be drilled through to the ground foil wherever a ground connection is needed. The VXO and first doubler are on one board. A second board contains the other two frequency doublers. A third board contains the balanced modulator, speech amplifier and linear-amplifier chain. Results with this transmitter have been good.

A 75-Meter Transceiver – Direct-Conversion Receive and DSB Transmit

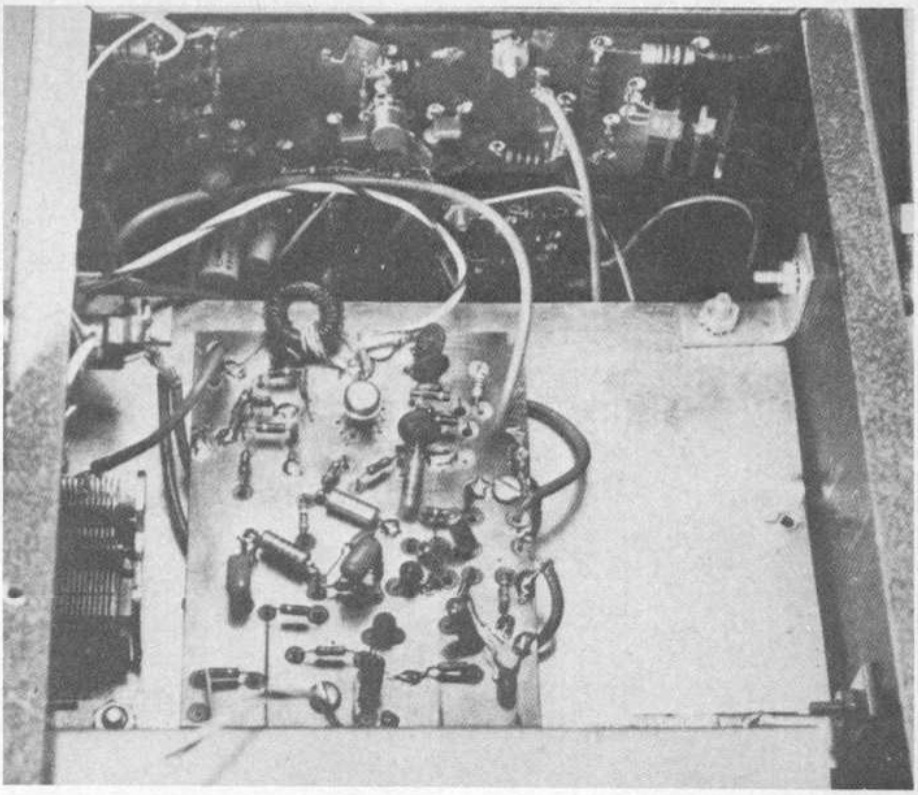
The transceiver described in this section covers the 80-meter cw and 75-meter phone bands. It provides full transceive and has an output of over 1 watt. This rig was built by Jeff Damm, WA7MLH, and is used for home station and portable operation.

The VFO section of the transceiver is shown in Fig. 31. This circuit is similar to many that have been used in other projects. The Hartley configuration is used. Reasonable stability is obtained by using capacitors of both the NP0 ceramic type and air variables. An MPF102 JFET is used and is Zener-diode regulated. The VFO is tuned with a capacitor from a surplus BC454 receiver. This capacitor has a maximum range of nearly 200 pF. The VFO requires that the variable capacitor (in parallel with the inductor) cover a range of 33 to 68 pF in order to tune the range from 3.5 to 4 MHz. In the WA7MLH transceiver a combination of fixed-value ceramic NP0 and air-variable capacitors was used in series with the main tuning capacitor to obtain the proper range.

The VFO is built in a separate box that is contained within the main cabinet. Since the oscillator operates at the same frequency as the transmitter output, it is important that good isolation be maintained. The oscillator is buffered with a feedback amplifier consisting of Q2 and Q3. The output power available is +10 dBm into a 50-ohm termination. The emitter current in Q3 was chosen large enough to maintain a sine-wave output under a 50-ohm load.



External view of the 75-meter dsb/cw transceiver built by WA7MLH. The VFO control is at the left.



Interior of the 75-meter transceiver. The VFO compartment is at the bottom of the photograph, and the receiver board is seen at the center. The transmitter output amplifier, balanced modulator and speech amplifier are mounted on the end panel at the top of the picture.

In transmit the VFO output is applied to the balanced modulator shown in Fig. 32, using a two-diode circuit. Carrier balance is adjustable with a 100-ohm control between the diodes. The carrier suppression was 36 dB. The 1N914 diodes were matched for forward resistance by means of an ohmmeter.

The output of the balanced modulator is applied to a 6-dB pad to assure proper termination, and is then routed to a broadband amplifier, Q4. This stage provides nearly 20 dB of gain. The balanced modulator and the first linear amplifier (Q4) are contained on a single circuit board.

Another circuit board contains the speech amplifier and a pnp transistor, Q5, for cw keying. The speech amplifier uses a pair of 741 op amps. Keying is realized through addition of Q5, a 2N3906.

The output amplifier is shown in Fig. 33. A 2N5189 transistor is biased for a current of 50 mA and serves as the

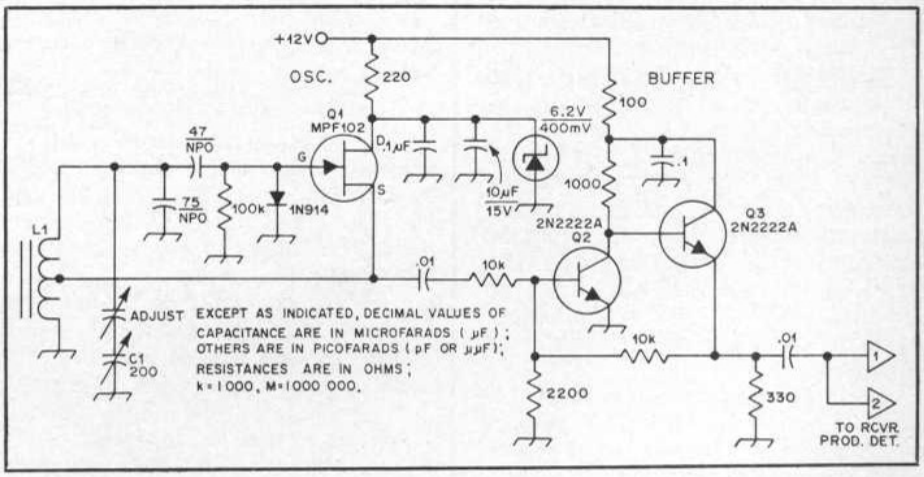


Fig. 31 – 3.5- to 4-MHz VFO for the WA7MLH 75-meter dsb transceiver.
C1 – 200 pF. Air variable capacitor.
L1 – 51 turns No. 26 enameled wire on Amidon T68-2 toroid core, tapped 12 turns from ground.

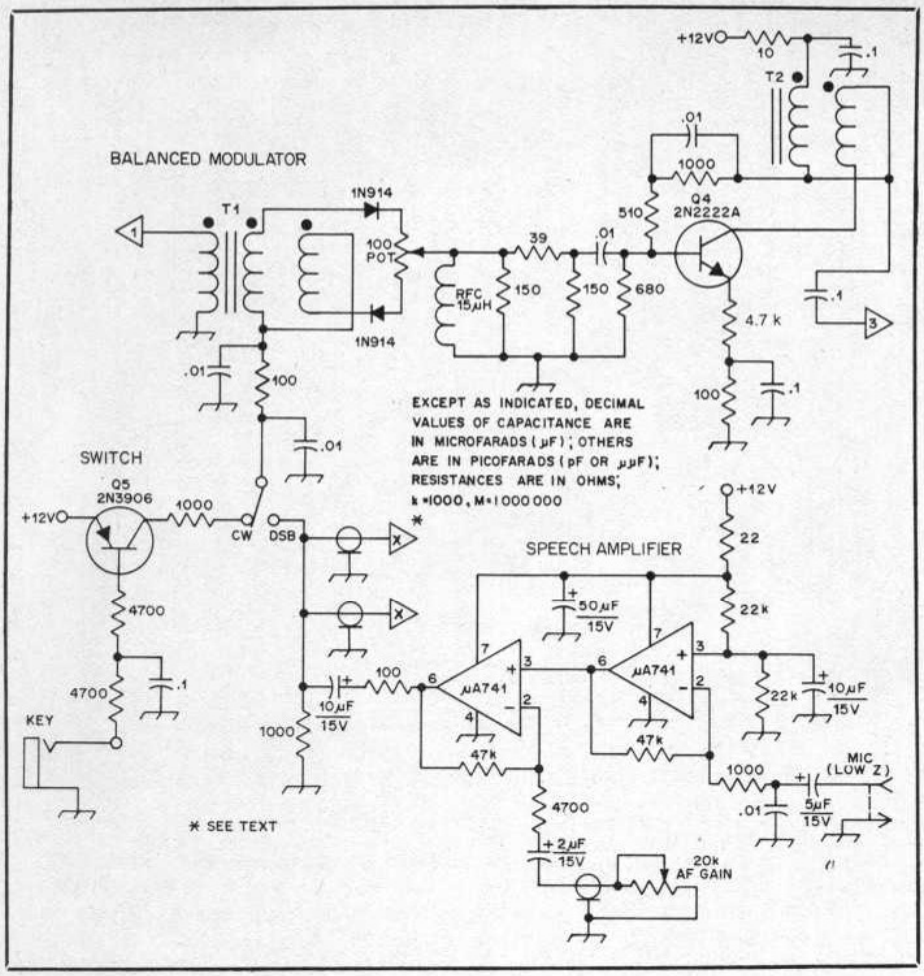


Fig. 32 — Balanced modulator, speech amplifier and keying switch for the WA7MLH transceiver.
T1 — 15 trifilar turns No. 30 enameled wire on an Amidon FT-37-61 toroid core.
T2 — 12 bifilar turns No. 30 enameled wire on an FT-37-61 toroid core.

with a dual-gate MOSFET product detector. While sensitivity was more than sufficient, severe problems were encountered with square-law detection of a-m stations. The MC1496G detector eliminated these problems with no penalty in sensitivity.

Transmit-receive control is provided by S1, a double-pole, double-throw toggle switch. One set of contacts switches the 12-volt supply between the transmitter and the receiver. A 12-volt relay is controlled by this line to change the antenna from the receiver to the transmitter input. The other contacts on S1 disconnect the output of the speech amplifier from the balanced modulator during receive periods. Without this measure, the operator's voice could be heard in the receiver during that mode. The +12-volt supply should be applied to the speech amplifier continuously.

A useful addition to this transceiver would be a meter (0-1 A) to monitor the total power-supply current. The operator could then adjust his voice level and microphone gain such that the current remained constant during transmit periods. An increase in current would indicate that the final amplifier was being overdriven. This would increase the distortion products significantly. Excessive "flat topping" was observed on an oscilloscope when the linear amplifier was overdriven.

A Universal Exciter for SSB and CW

The transmitter described in this section was designed to provide good driver. A small heat sink is used on this stage. The output amplifier, Q7, is an inexpensive plastic power device, a GE type D44C6. It is biased for Class A operation with a standing dc collector current of 250 mA. The saturated cw output power of this stage is 1.5 watt. About 1-watt PEP of dsb output power is obtained. A half-wave filter serves as the output network.

Measurements have been performed on a similar breadboarded version of this amplifier. The overall gain of the linear chain (Q4, Q6 and Q7) is over 40 dB. The same gain is available in the 40-meter band. The gain drops significantly at 14 MHz. This results from the limited f_T of the output transistor. If a similar transmitter were to be used on the higher amateur bands, an output transistor with a higher f_T would be desirable.

The receiver used in the WA7MLH transceiver uses an MC1496G product detector which is followed by a pair of audio amplifiers containing 2N3565s. This receiver was described in detail in chapter 5. The original version of the WA7MLH transceiver used a receiver

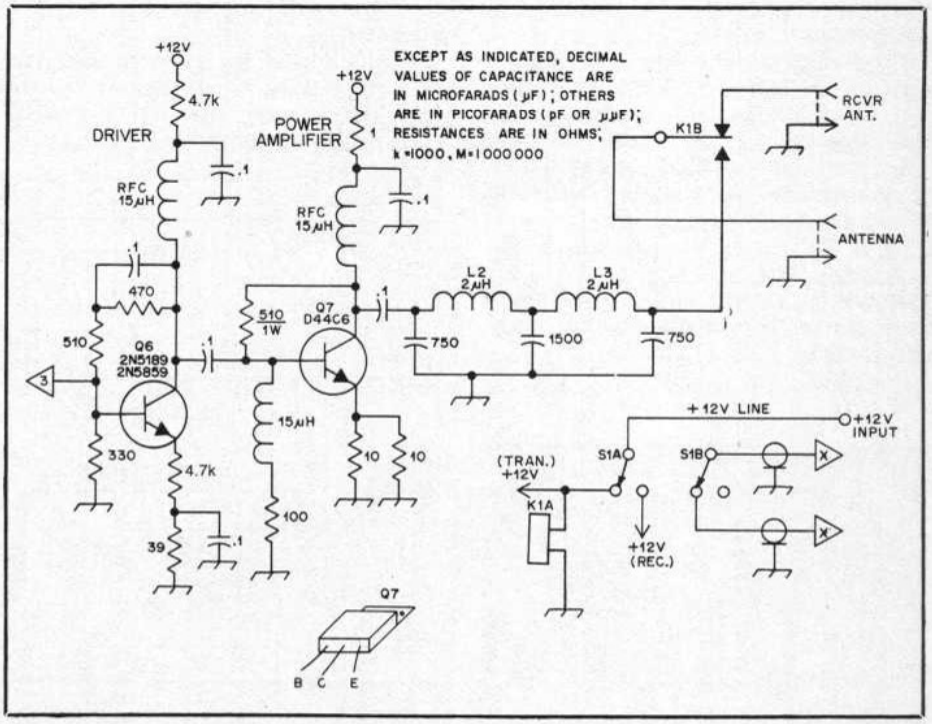


Fig. 33 — Rf-output amplifier and details of T-R switching for the WA7MLH dsb transceiver. System is shown in the transmit position.

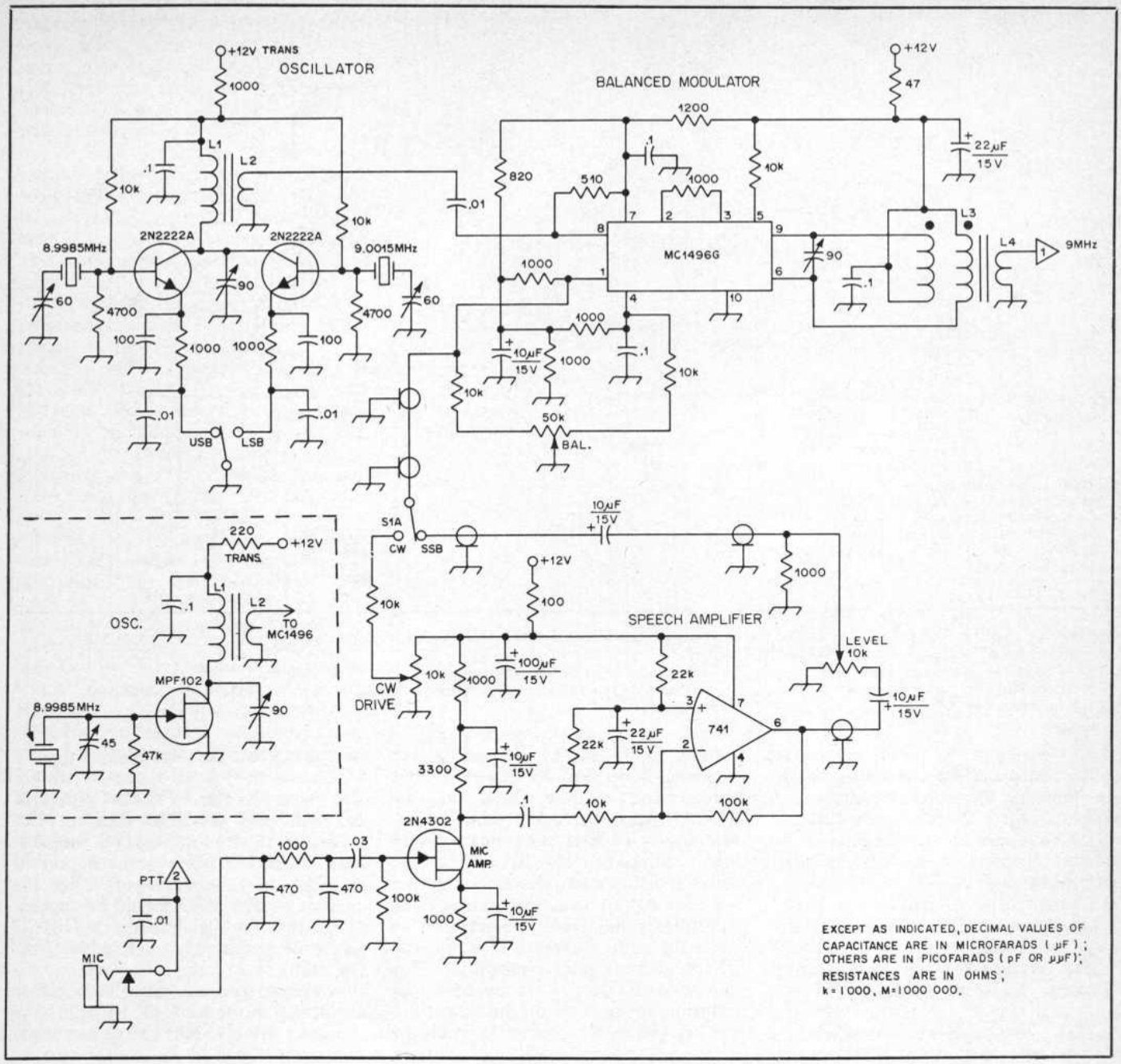


Fig. 34 — Carrier oscillator, speech amplifier and balanced modulator for the universal ssb transmitter. Insert shows the FET oscillator used in the KL7IAK version. Coils are identical for either circuit. A double-pole, double-throw switch (S1) serves as the mode switch. Any type is suitable since no rf is switched. The other half of the switch is shown in Fig. 41. All variable capacitors are mica compression or ceramic trimmer types.

L1 — 45 turns No. 28 enameled wire on an Amidon T50-2 toroid core.

L2 — 3 turn link over L1.

L3 — 20 bifilar turns No. 28 enameled wire on an Amidon T50-6 toroid core.

L4 — 6-turn link over L3.

performance on ssb and cw. It was intended primarily for QRP work. The output power is enough that higher power linear amplifiers may be driven directly. Data are given for operation on any amateur band from 1.8 to 50 MHz. The original unit was built by Terry White, KL7IAK.

The transmitter was a single-band unit for 20 meters. The filter approach to side-band selection was used and a narrow-band design was adopted for the rf power chain.

Shown in Fig. 34 is the carrier generator, balanced modulator and speech amplifier. The carrier oscillator uses a pair of bipolar transistors. A common tuned circuit is shared by the collectors of the two oscillators. However, only one transistor is biased "on" at a time. This allows the operator to choose the desired sideband. The JFET oscillator used for usb generation in the original KL7IAK unit is also shown in the insert in Fig. 34.

An MC1496G is used as a balanced modulator. Means are provided for adjusting the carrier balance. Measured carrier suppression was over 50 dB. Code operation is realized by inserting dc into the balanced modulator. This allows sufficient carrier energy to ride through for cw operation.

The speech amplifier uses a JFET input amplifier, making the circuit compatible with high- or low-impedance microphones. The FET is followed by a 741 op amp which provides a voltage gain of 10. If additional gain is needed, a second op amp could be cascaded with the first.

Shown in Fig. 35 is the i-f and output mixer system for the transmitter. A pair of dual-gate MOSFETs is used as 9-MHz amplifiers. They provide

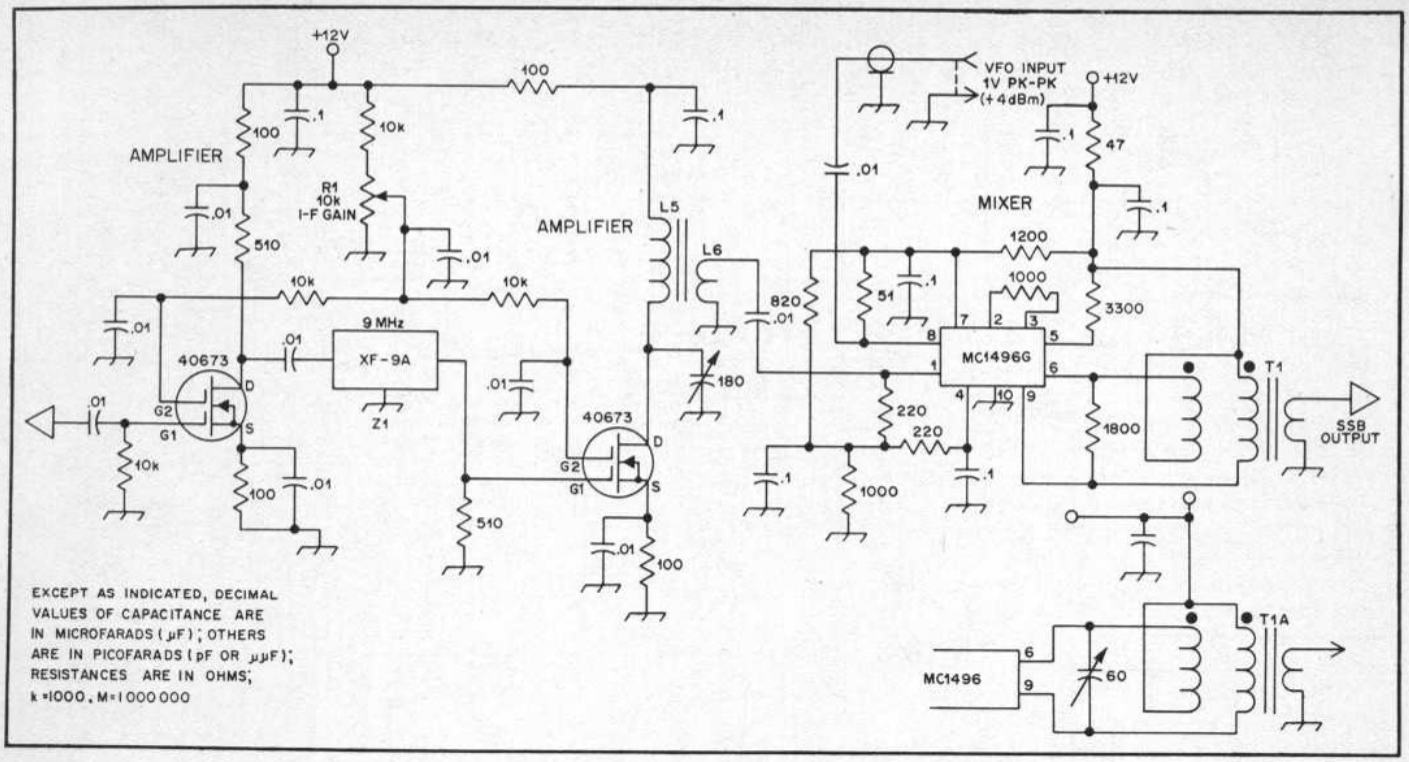


Fig. 35 — I-f amplifier and transmit mixer for the universal ssb transmitter. The insert shows the mixer output circuit used in the KL7IAK version of this exciter. R1 is a pc-board-mounted control.

L5 — 28 turns No. enameled wire on an Amidon T50-6 toroid core.

L6 — 3-turn link over L5.

T1 — 15 bifilar turns of No. 30 enameled wire on an FT-37-43 (primary). 5-turn secondary.

T1A — see text.

Z1 — 9-MHz crystal filter, KVG type XF-9A.

some signal gain, terminate the crystal filter, and provide a convenient means for adjusting the gain. The application of gain control to gate 2 of a dual-gate MOSFET amplifier was discussed in the receiver chapters. While this can cause IMD to be generated, the signal levels in this i-f amplifier are low enough that it is not a problem. If desired, an alc signal could be applied to the two gates. The reader is referred to the receiver chapters for the discussion of agc systems.

The output mixer transfers the 9-MHz ssb signal to the output frequency of interest. An MC1496G is used as the mixer. The IC is biased for larger currents than are normally used with this device. This enhances the linearity (the output intercept is increased). Broadband and narrow-band output networks are shown. The broadband transformer will provide a 50-ohm output over a wide range of frequencies, making it suitable for driving multi-section filters at the desired output frequencies. See Fig. 38. The alternative narrow-band output (used in the KL7IAK version) uses a tuned transformer. For 14 MHz, the primary has 20 bifilar turns of No. 30 wire on an Amidon T50-6 core. The secondary has a 3-turn output link. The narrow-band transformer has enough bandwidth to cover the entire 20-meter band, but still offers some image rejection. The narrow-band output is suitable for adaption to most of the hf bands. For use on 160 or 75 meters, it would be desirable to use the wide-band design.

Shown in Fig. 36 is the circuit for the 5- to 5.5-MHz VFO that is used in the KL7IAK 14-MHz version. The reader is referred to the 80- and 20-meter superhet receiver in chapter 5, and to the discussion of VFOs in chapter 3. The VFO should be capable of delivering a signal to the MC1496G mixer of 1 volt pk-pk across 50 ohms (+4 dBm).

The narrow-band linear rf-amplifier chain used in the KL7IAK transmitter is shown in Fig. 37. This circuit uses three



The 9-MHz i-f amplifier used in the universal ssb/cw transmitter.

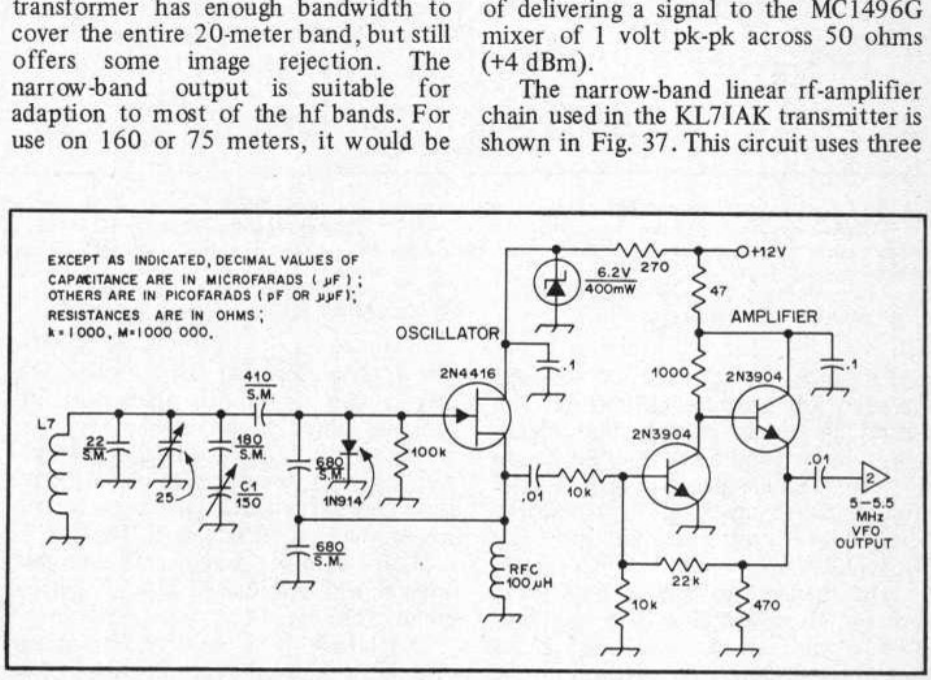
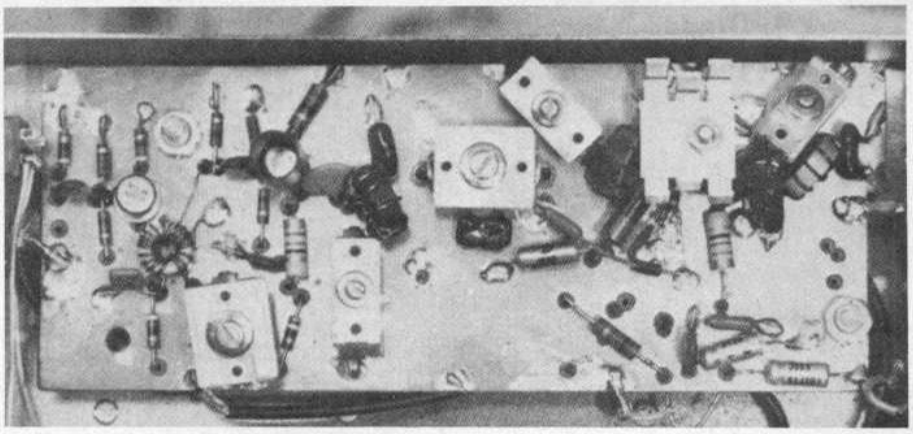


Fig. 36 — A 5.0 to 5.5-MHz VFO for the universal ssb transmitter. This circuit may be used as shown for 3.5 to 4, or 14 to 14.5-MHz operation. For other bands it is heterodyned to the appropriate injection frequency. This is presented in Fig. 40. L7 is a 3.4-μH inductor on a 3/8-inch diameter ceramic form (no tuning slug used). C1 is a 150-pF air variable.

stages and delivers an output of 2.5-watts PEP, or cw with a total small-signal gain of 57 dB. The input stage is a 2N5859 biased for a current of 25 mA. This is followed by another 2N5859 which runs at a collector current of 60 mA. A tuned transformer is used in the output of the input stage. A pi network matches the driver to 50 ohms. The first two stages are capable of delivering 100 mW of output with excellent linearity. This stage was matched to 50 ohms rather than directly to the base of the final amplifier to allow the low-level output to be extracted for driving vhf transverters.

The output amplifier contains a Motorola 2N6366. The networks were designed from the impedance data supplied by the manufacturer. A *C-C-L* type of T network is used for base matching and a pi network was employed for the output. Originally, the circuit was built following the sample presented in the manufacturer's literature. The transistor was bolted to a heat sink and the reference diode was soldered to a lug that was fastened to the stud of the transistor. The performance appeared to be exactly that specified by the manufacturer when R1 was set for an idling current of 15 mA. However, the amplifier could be run only for very short periods in a cw or two-tone ssb test. If the operating period exceeded half a minute, the transistor would go into thermal runaway. If rf drive was applied for 1 minute, then removed, the collector current was near 200 mA. At this level the heating was enough without rf drive that current would slowly increase.

In order to ensure thermal stability, emitter degeneration was inserted in the circuit. Four 2.7-ohm, 1/4-watt resistors were paralleled to provide a resistance of 0.68 ohm. The bias in the diode was then readjusted for 15-mA collector current with no rf drive. The stage gain was decreased by means of the emitter degeneration. A slight instability was cured by placing a 220-ohm resistor across the collector rf choke.



The 20-meter power-amplifier chain used in the universal ssb/cw transmitter. The input stage is seen at the left. To the right is the Class AB output amplifier. Power output is approximately 2-1/2 W cw or PEP. Third-order IMD products are 30 dB below the 2-W PEP two-tone output with this amplifier.

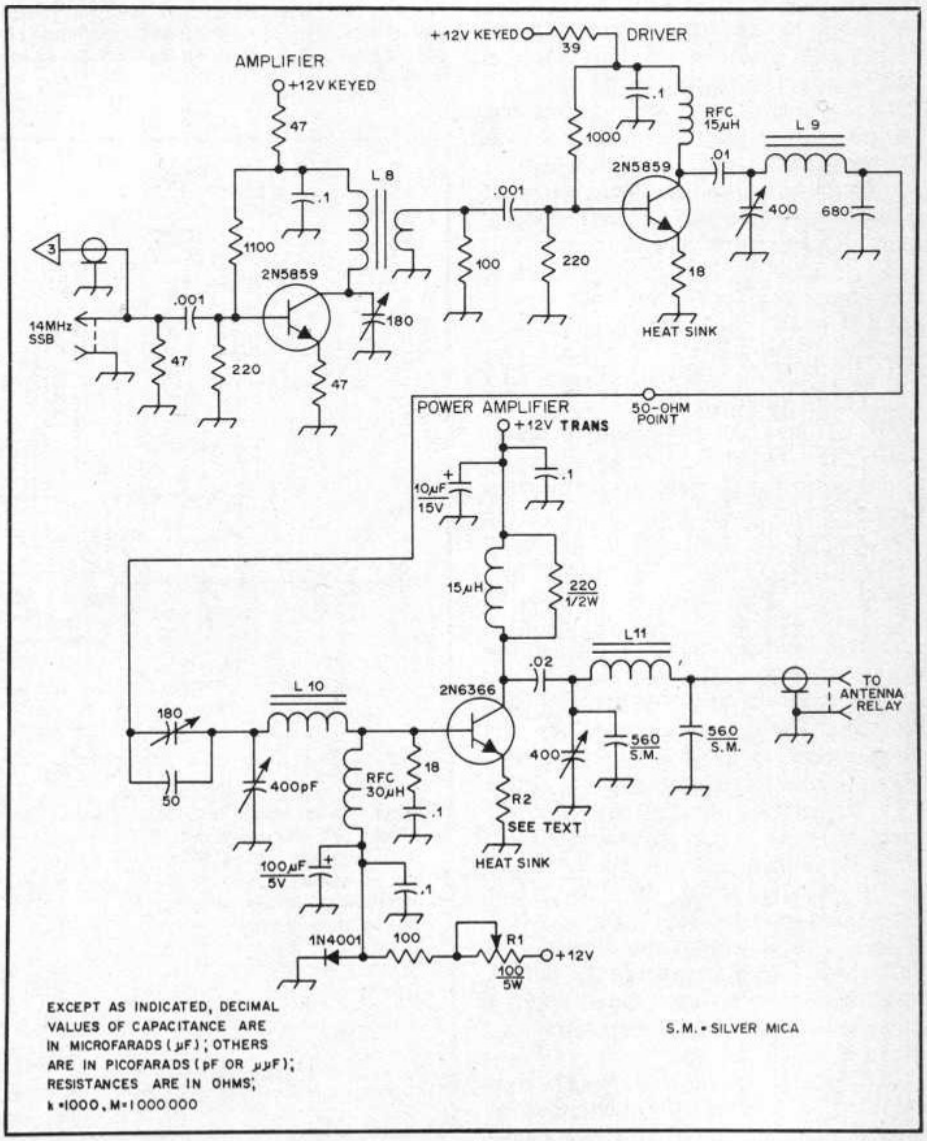
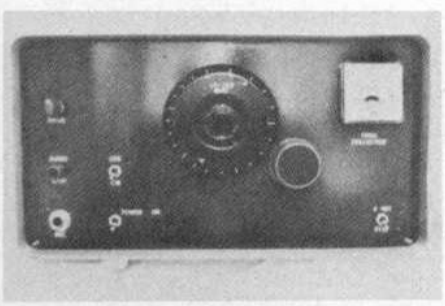
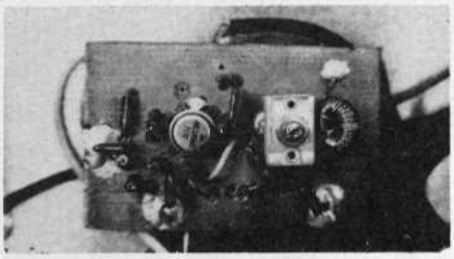


Fig. 37 — A 14-MHz narrow-band rf-power amplifier chain used in the KL7IAK version of the universal ssb transmitter. All variable capacitors are mica-compression types. R2 is 0.68 ohm (four 2.7 ohm, 1/4-watt, 5-percent resistors in parallel).

L8 — 18 turns No. 22 enameled wire on a T50-6 toroid core.
L9 — 11 turns No. 22 enameled wire on a T50-2 core.
L10 — 9 turns No. 22 enameled wire on a T50-6 core.
L11 — 8 turns No. 22 enameled wire on a T50-6 core.



Front view of the 20-meter version of the universal ssb/cw transmitter (built by KL7IAK).



The transmit mixer used by KL7IAK.

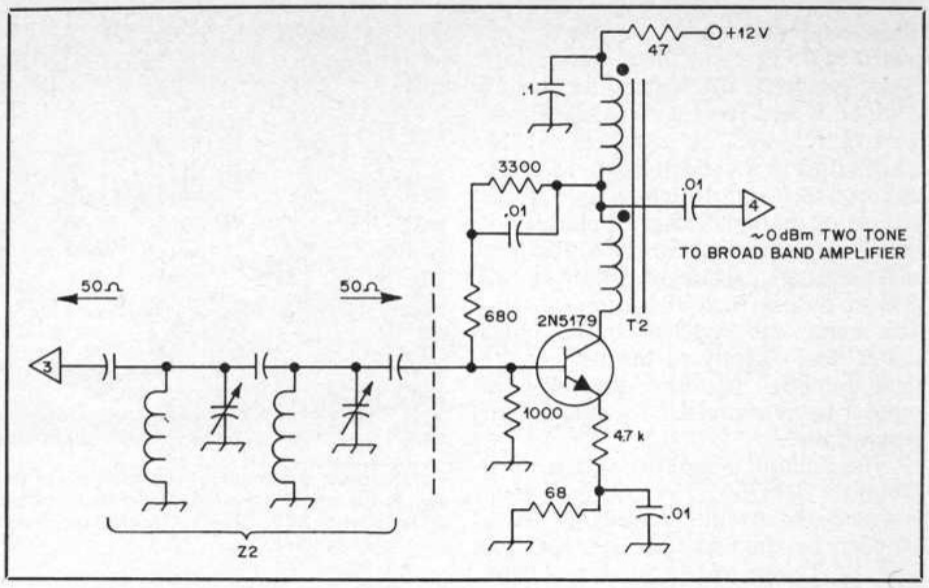


Fig. 38 — Bandpass filter and broadband preamplifier for the rf-output chain of the universal ssb transmitter. This circuit follows the transmit mixer of Fig. 35. T2 consists of 10 bifilar turns of No. 28 enam. wire on an Amidon FT-23-43 toroid. Z2 is a bandpass filter from the tables in the appendix.

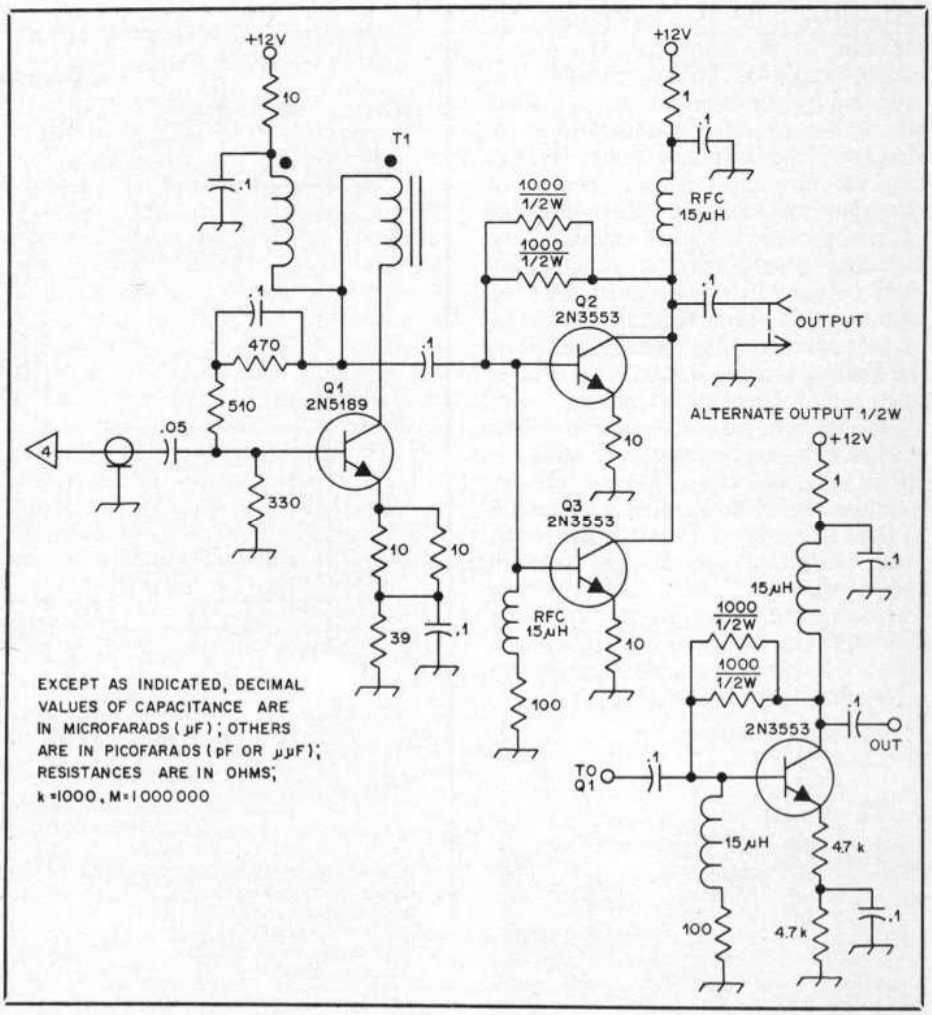


Fig. 39 — Broadband Class A power amplifier for the universal ssb transmitter. The output power is 1-watt cw or PEP linear. This circuit is suitable to follow the filter and amplifier of Fig. 38 for any band from 1.8 to 30 MHz. Heat sinks should be used on all three transistors in the amplifier. The insert shows a modification which is suitable for 0.5 watt of output. This filter should be followed by a low-pass filter for the band of application. Suitable filters were described in chapter 4. T1 is 12 bifilar turns No. 30 enameled wire on an Amidon FT-37-61 ferrite toroid.

As modified with the emitter degeneration, the amplifier appeared to be thermally stable. The amplifier chain was run at full output for a five-minute period. When the drive power was removed, the collector current in the output stage was 50 mA, and quickly decreased to the previously established 15 mA. A two-tone test on the total power chain produced IMD products over 30 dB below each output tone. The output power during the test was 2.5-watts PEP.

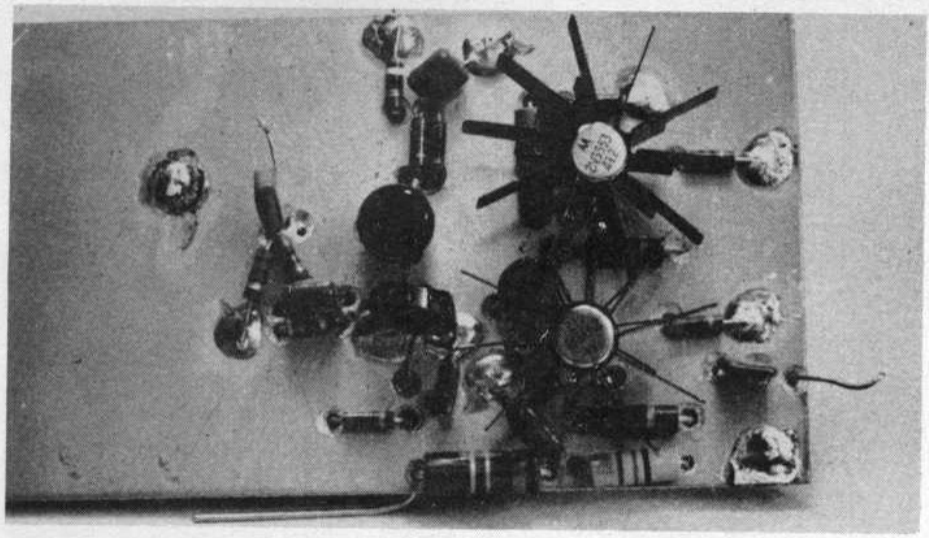
The output-amplifier chain was built in Oregon where instrumentation was available for careful evaluation. The experience with thermal runaway was very impressive for the writers, suggesting that emitter ballasting is a necessity for any Class AB ssb amplifier. This includes units for QRP operation as well as the higher power versions.

If this transmitter is built for other bands, the circuits must be changed. The narrow-band power chain shown in Fig. 37 could be adapted to any of the amateur bands from 1.8 to 30 MHz. However, a more modern approach would be to utilize broadband designs.

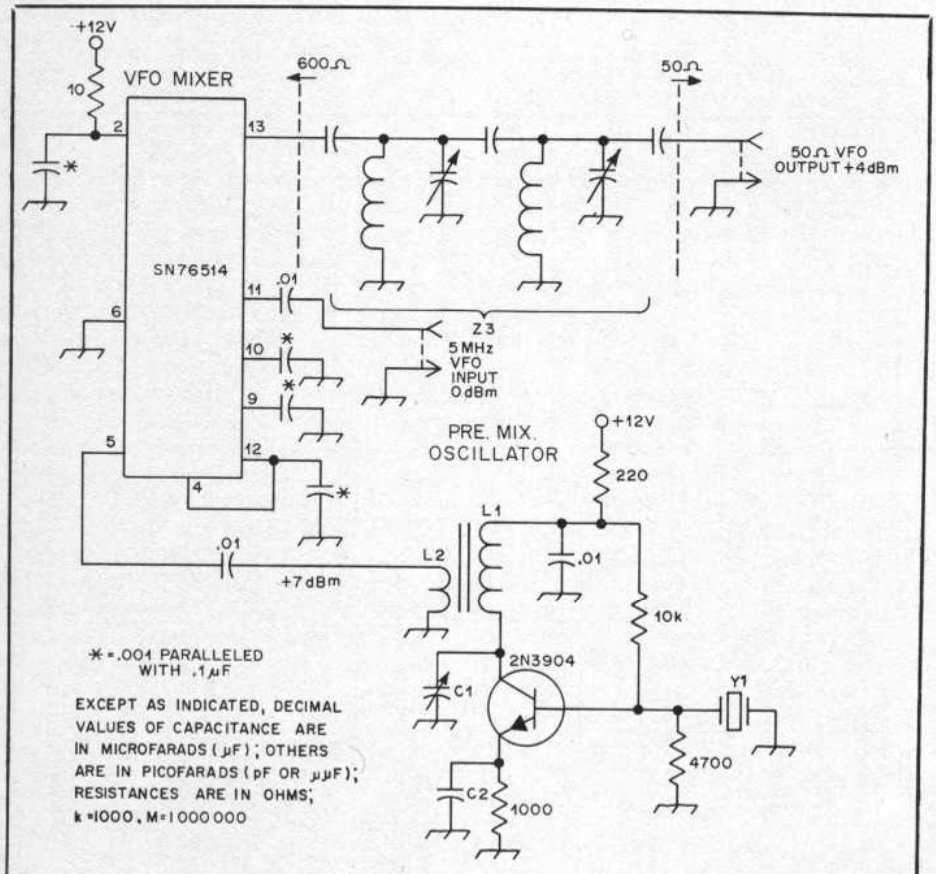
Shown in Fig. 38 is the circuit of an amplifier suitable for following the mixer of Fig. 35. Network Z2 is a double tuned circuit for the band of interest. The component values for these filters are listed in the computer-generated tables in the appendix. The 2N5179 amplifier is flat into the vhf spectrum with a gain of almost 20 dB.

A broadband Class A power amplifier is presented in Fig. 39. This circuit has two stages, provides gains up to 36 dB, and will deliver an output of 1-watt PEP or cw. The transistors in the output stage should be fastened carefully to a suitable heat sink, since the current in the final is 250 mA.

The driver in the rf-power chain is identical to the amplifier described in Fig. 38, except that the collector current is higher. The output amplifier uses a pair of 2N3553s in parallel. Emitter degeneration is used in this amplifier for bandwidth extension. The emitter resistors further ensure that the dc current in the transistors is divided equally. Also



Breadboard version of a 1-watt output Class A broadband power amplifier. The circuit provides over 30 dB of gain over most of the hf region. Heat sinks are used on the parallel 2N3553 output amplifiers.



BAND	Y1 MHz	F_{out} (Z_3)	L1	L2	C1 (pF)	C2 (pF)
40M	11	16-16.3	24 ts	2 ts	100	75
15M	17.5	12-12.5	22 ts	2 ts	50	47
10M	14	19-19.5	21 ts	2 ts	75	47
6M	36	41-41.5	13 ts	2 ts	30	15
160M	5.8	10.8-11	45 ts	4 ts	100	100

≠ = Nominal C1 required.
Z_3 = Filter from tables.
F = F_{out}

Fig. 40—Circuit for a heterodyne conversion system for the VFO. A 5- to 5.5-MHz VFO such as that shown in Fig. 36 is heterodyned to the needed output frequency for operation on any amateur band from 1.8 to 50 MHz. Note that the values given in the table for C1 are nominal values. A slightly larger mica compression trimmer should be used. All coils for the crystal oscillator are wound on Amidon T50-6 toroid forms. Z3 is a 2- or 3-pole bandpass type from the appendix (see text). The SN-76514 mixer IC has been reidentified as TL-442-CN by Texas Instruments. It may be procured under either part number.

shown in Fig. 39 is an adaptation of the circuit using a single 2N3553. This circuit should provide identical gain and bandwidth, but will have an output power of only 1/2 watt.

The broadband amplifier was evaluated for IMD while using a pair of signal generators at 14 MHz, and a spectrum analyzer. The output intercept was +43.5 dBm. When the amplifier was run at 1-watt PEP output (+24 dBm per tone) the IMD was 39 dB down. The maximum gain was 36 dB in the 3.5- and 7-MHz bands. The gain was down to 34.5 dB at 14 MHz and was 29 dB at 29 MHz. If the transmitter is built for the 6-meter band, it is suggested that the power amplifier used in the previously described 144-MHz dsb transmitter be used. The output network must be altered.

The broadband amplifier (Fig. 39) should be followed by a low-pass filter. Half-wave filters are suitable (see chapter 4).

When the transmitter is used on bands other than 20 or 80 meters, a different VFO system is needed. A solution is to use a heterodyne VFO. Shown in Fig. 40 is a schematic for a proposed system that could be built for any of the bands from 1.8 to 50 MHz. A 5- to 5.5-MHz VFO is used. Its output is heterodyned to the suitable injection frequency. An SN76514 double-balanced mixer IC is used. A crystal-controlled oscillator is employed as the other input to the VFO mixer. Values are given for the oscillator components for all bands.

The output of the premixer (Fig. 40) must be filtered well in order to suppress spurious responses. A double- or triple-tuned circuit is used. The circuit should be terminated in 50 ohms at the output. The input termination should be 600 ohms to match the output of the SN76514. Filters values are given in the appendix. They are designed for a 50-ohm termination at each end. The methods to adapt them to other terminations are also presented. Either 2- or 3-pole filters may be used. For most cases the double-tuned circuit will be sufficient. The 3-pole filters are preferable for the 10- and 15-meter bands.

The transmit mixer requires an injection power of +4 dBm. If this level is not available at the output of the filter, it may be increased by means of a broadband amplifier.

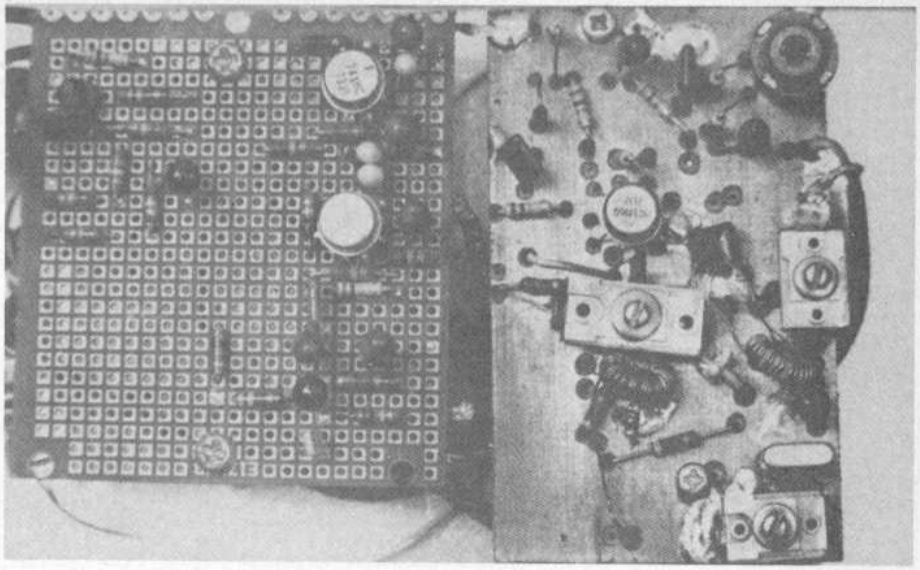
While the circuit shown in Fig. 40 has not been built, we feel that it should present no problems. Two other projects in the book use a similar circuit in a virtually identical application. No problems were encountered with those designs as long as the proper filter terminations were used.

A control system for the ssb exciter is shown in Fig. 41 (see chapter 7). All

switching functions are done with transistors except for the antenna section which utilizes a relay. A delay is built into the system to ensure that the antenna relay is closed prior to generation of rf from the transmitter. Shaped keying is also included. Two pnp transistors are used for switching. These supply the +12T and +12K (keyed) lines in the transmitter. These transistors should be capable of switching up to 1 ampere.

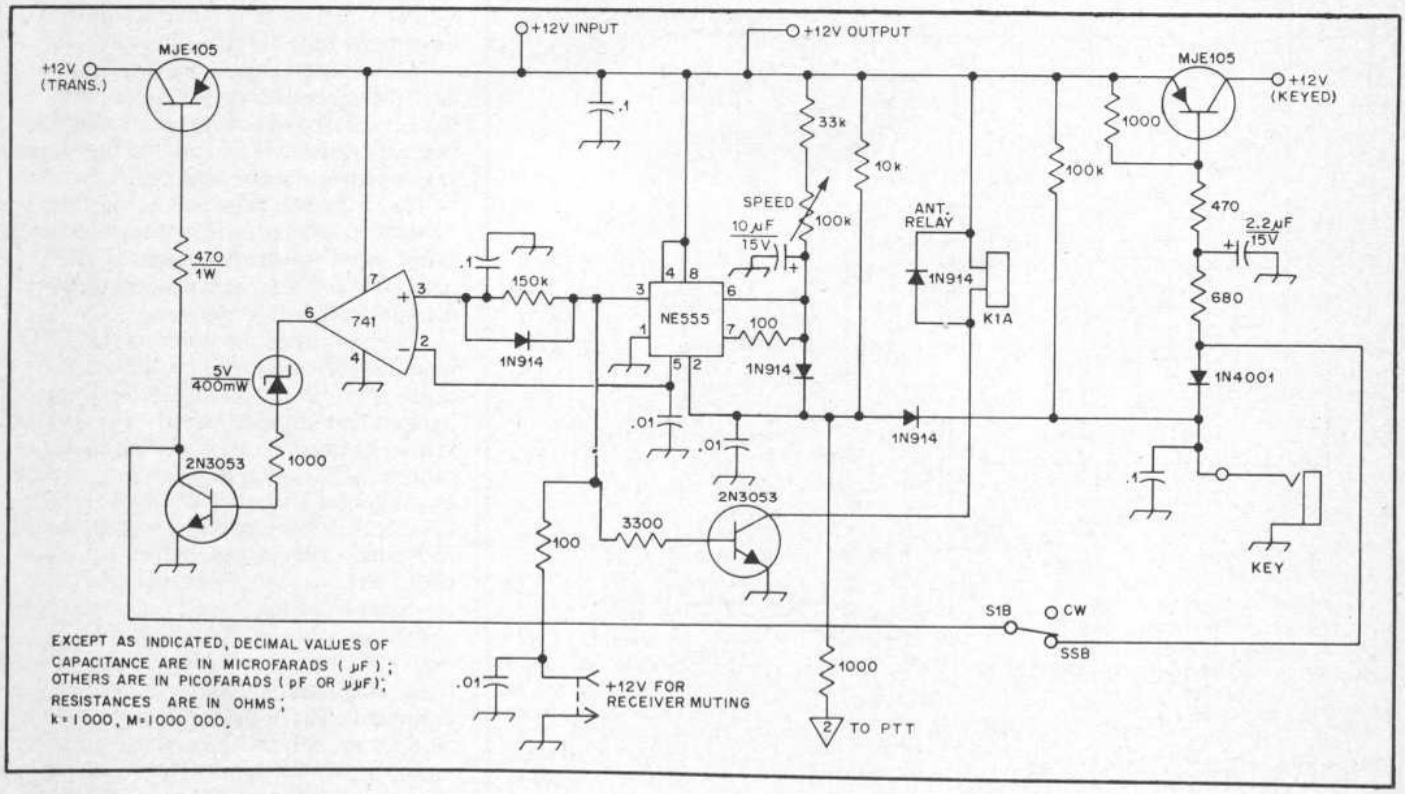
At this writing only one of these transmitters has been built – the original KL71AK unit. It has been highly successful on both ssb and cw.

It should be emphasized that the universal ssb system described in the preceding pages is an advanced project. Although well within the capabilities of the amateur with construction experience, it should not be attempted by the beginner. There is no printed-circuit layout information available on any of the projects described in this chapter.



The right-hand board contains the carrier oscillator and balanced modulator of the KL71AK transmitter. The circuit at the left is the audio section. Two stages of audio were used, but later found to be unnecessary.

Fig. 41 – Control system for the universal ssb system. This circuit provides automatic T-R switching on cw and push-to-talk operation on ssb. The design details of these control systems were presented in Chapter 7.



Chapter 9

Field Operation, Portable Gear and Integrated Stations

Most of the equipment described in this book is suitable for field use, be the application one of weekend camping, mountain climbing, hiking, boating, or long-term vacationing abroad or in the USA. The exact nature of the material taken afield will depend to a large extent upon the environment in which the gear shall be used. In more definitive language, the equipment must be designed for extreme compactness in some instances, and must be capable of operating from batteries. The backpacker and hiker are especially mindful of the foregoing requirements, and would add to their list of accessories a lightweight antenna system, headphones, key, and/or microphone.

The lakeshore or river-side camper might elect to carry larger, more powerful radio equipment with him. He could utilize the automobile battery or a gasoline-powered generator to obtain the needed source of energy. His antenna system could be more rugged and elaborate than that used by the backpacker.

Those who operate from motels or hotels, stateside or in some distant land, would be more apt to employ an ac-operated power supply which was compatible with the line voltage and frequency in the area where operation was planned. However, a rechargeable battery might also be included in the travel kit for use at times when local power failures occur – and they do in many foreign countries!

There is a mystique connected with portable operation, for in many instances the amateur is using home-made equipment which was tailored to the application. Furthermore, low power is employed much of the time, and conditions are seldom ideal with respect to operating conveniences. Being heard, and having other station operators copy your signal solidly, not only is a measure of your station effectiveness, it's a self-satisfying feather in the cap of the designer/operator. "Doing it the hard way" does not necessarily denote a twinge of masochism. Rather, it proves that QRP gear is worth its weight when applied properly.



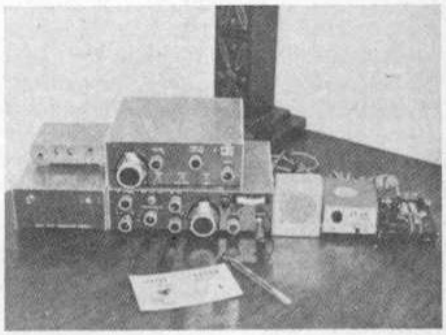
Low -power station equipment can be used in place of commercial gear when the QRP challenge inspires the operator. On the left side of the operating position is the W1CER 40- and 20-meter 10-W station. The power supply and Transmatch for the homemade setup are on the shelf behind the QRP gear.

Equipment Characteristics

The environment at the site of portable operations is of major importance to the designer. For example, the mountain climber will encounter extremes of cold, which can affect the performance of his equipment if certain design steps aren't taken. His transceiver and related apparatus must be small and light of weight – and rugged – if it is to suit his particular needs correctly (more on this subject later).

The camper needs equipment that can function properly in damp weather. It should be reasonably immune to dirt and temperature extremes, and requires a quality of ruggedness which most home-station gear need not have.

The foreign traveler will often choose compact equipment, owing to the inconvenience of lugging a large, heavy commercial transceiver. Lightweight, compact gear can be carried aboard an airplane without the penalty of being "overweight." The latter can become rather expensive! Also, the station equipment is less likely to be damaged if kept out of the hands of baggage men during air travel. Being able to take the package of radio equipment to one's seat on the plane will also prevent misrouting of the parcel to some destination other than the intended one! The writers recall an unhappy event that found the entire DXpedition radio package missent to Trinidad, when the operators and their personal effects were destined to land on Barbados (W1KLK, W1CKK and W1CER). Not only did the radio gear become lost temporarily, the suitcases



Solid-state QRP station used by W1CER at ZF1ST. A backup keyer and the station power supply are at the left. The top-center unit is the 40- and 20-meter 10-W cw transmitter. A 160- through 15-meter superheterodyne receiver is below the transmitter. A small speaker flanks the receiver to the right, then comes a homemade keyer with a commercial paddle.



Portable operation can take place from a makeshift table. Here, the trunk of a VW fastback is used as an operating position. A 12-V battery is used to power this 1-W 80- and 40-meter cw transceiver. A homemade keyer is visible at the right. This station was used during an ARRL CD Party by W1CER during a New Hampshire camping trip.

containing clothing and cosmetics vanished at the same time. The errant luggage turned up a few days later at the seaside resort on Barbados, but the radio equipment had been damaged severely. The lesson learned was that hand-carried QRP equipment was the better choice for traveling by air!

Tent Camping

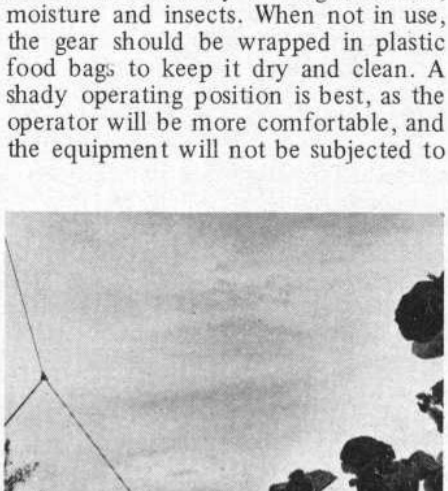
Most "purist" campers who dwell in tents will not be situated where ac power is available. Chances are that they will not be close to an automobile, which will rule out "snitching" equipment power from a car battery. Not many ardent campers will justify polluting the serenity of the wilderness by using a noisy, gas-gulping power plant. Therefore, various types of battery power supplies become the order of the day. Some camper/amateurs use series-connected 6-volt lantern batteries to obtain 12 volts for the QRP gear. Others employ Gel-Cell or NiCad batteries. Still others obtain good results with flashlight cells connected in series to provide the required operating voltage. The choice is based usually on what's available at the time, and on the power consumption of the field equipment. Another excellent power source is a 12-volt motorcycle battery, or two 6-volt ones hooked in series. If the automobile is close enough to the campsite to permit occasional recharging of the batteries, NiCads, Gel-Cells and motorcycle batteries are the best bet. If dry batteries are used exclusively, it's wise to carry enough spares to bracket the arrival and departure dates adequately.

Assuming that battery power is used, the equipment should not consume more than a few hundred milliamperes with everything running. The cw operating mode will probably be the most efficient one. Effective communications should be possible from 160 through 10 meters while using power levels from 0.5 to 3 watts, assuming that a reasonably good antenna is employed, and that band conditions are suitable. A good day- and night-time frequency combination is 20 and 40 meters. Propagation on those bands will permit round-the-clock operating, most of the time.

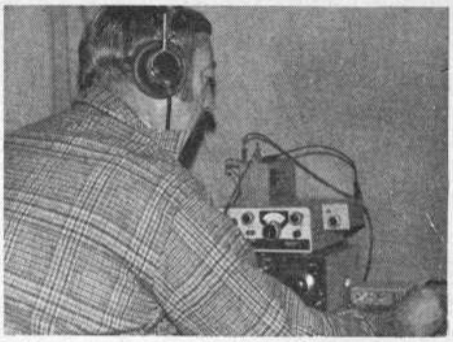
Dipole antennas are among the easiest to transport and erect when camping. They can be supported by tall trees or cliffs – erected as inverted Vs, sloping dipoles, or in a traditional format – horizontally. A bow and arrow is useful when erecting antennas, for it permits a pilot line to be fired and snaked through a treetop, preparatory to pulling the antenna aloft with a heavier line. Those skilled with a spinning rod can shoot a quarter-ounce practice lure or sinker over a treetop, then pull the antenna line up by hooking the monofilament fishing line to the main one.

An inexpensive but good antenna-support line is the Nylon type which can be bought in many hardware and discount stores in the USA. A 500-foot roll will last a long time and will cost less than $3. The writers prefer the small-diameter kind which has a tensile strength of 100 pounds or greater. When the campout is finished, the cord can be placed back on the spool for use another time.

It may be necessary to use the radio equipment on the ground, as some campers do not carry tables and chairs afield. Therefore, the equipment should be sealed reasonably well against sand, moisture and insects. When not in use, the gear should be wrapped in plastic food bags to keep it dry and clean. A shady operating position is best, as the operator will be more comfortable, and the equipment will not be subjected to extremes of heat from the sun. The latter can cause expansion of critical tuning mechanisms (trimmers and coil slugs), leading to degraded performance and a need to readjust the circuits.



A sloping dipole strung near the seashore makes an effective antenna for QRP DX-peditions. Shown here is the ZF1ST/W1CER 40-meter dipole used on Grand Cayman Island (Spanish Bay Reef). Power output from the transmitter was 7 watts, and RST 589 reports were received from JA stations during the operation.



One advantage of QRP gear is that it doesn't occupy a great deal of room. In this photograph the station (HW-7 and a homemade cw transceiver) occupy one corner of a tent during a camping trip. The equipment is powered from a 12-V battery.

Since accessory equipment for camping and out-of-country operations is similar, that subject will be covered singly, later in this section. Generally speaking, the same kinds of antennas are adequate for both applications.

QRP DXpeditioning

There is probably no greater thrill in amateur radio than that of *being* DX with QRP equipment. W1CER has made several trips to islands in the West Indies for the purpose. Much of the work was done as 8P6EU from Barbados, with XYL Jean, W1CKK/8P6FJ, as a second operator. Other operations took place from Grand Cayman Island as ZF1ST. Propagation from that part of the world is superb to the USA and Europe, making it practical to employ low-power transmitting gear. The antennas have always been half-wave dipoles (coax fed) which were erected as "slopers" at whatever height was possible. Because salt water constitutes a superb ground medium, the antennas were slug over the seashore to assure best performance. The maximum transmitter output power used was 7 watts. Much of the work was done, however, with 1-1/2 to 2.5-watts output. The primary bands of operation have been 40, 20 and 15 meters. Cw was the operating mode.

Solid QSOs were had with many amateurs from Europe, South America and the USA. From Grand Cayman during October of 1974, a number of Japanese stations were worked on 40 meters at sunrise, local time (1000 GMT). Power output was 7 watts, and the antenna was a sloping dipole, the center of which was 15 feet above ground! Signal reports both ways were RST 589. ZLs and VKs have been

worked with 2 watts and a sloping dipole from Barbados. The period was early sunrise, and the band was 20 meters. Contacts like that are the exception rather than the rule, but they can be made with QRP equipment. Some signal enhancement from 8P6EU to Oceania probably resulted from having the 20-meter sloping dipole facing west on the western side of the island. Furthermore, a 30-foot coral cliff was behind the antenna (east), helping to effect some directivity.

There are many fine Caribbean islands from which to operate. Prior familiarity with government regulations is recommended, lest an amateur arrive and not be granted operating privileges. On Barbados a license can be acquired only in person. One must present his U.S. license to the Government Electrical Inspector, Old Hospital Bldg., Bridgetown. The fee for 12 months is nominal, and the license can be renewed yearly by mail. A Caymanian reciprocal permit can be obtained by mail if the fee is sent along with a photocopy of the U.S. license. The call will be your U.S. one, slant ZF1. Applications must be addressed to Her Majesty's Postmistress, Licensing Division, Post Office, Georgetown, GCI, BWI. The fee for one year on Grand Cayman is fairly stiff, and the rate changes from time to time.

It is wise to write to the local radio club on the island one plans to visit. Data can thus be obtained on Customs regulations and licensing. Some countries will not grant a license, and others make it practically impossible to bring equipment in. On some of the islands one must post a bond which represents 80 percent of the face value of the radio gear! Some amateurs have reported great difficulty getting all of the bond money returned at the end of their vacations! On some islands it takes a year or more to get a license, owing to government red tape. Be sure to check first; then make vacation plans.

Accessory Equipment

Campers and DXpedition types should anticipate equipment failures and prepare accordingly. If a backup station is not carried afield, spare parts and tools are a must. It is wise to conclude that most of these things will not be available once the operator reaches his destination. Radio stores just don't exist in the back woods or on many West Indies islands, so take what you need with you.

The following list of tools is suggested when space permits taking them along: Diagonal cutters, jack knife, electrical tape, screwdrivers, pliers (needle-nose and regular), small VOM, solder, soldering iron (battery operated), clip leads (6), cube taps, extension cord and hookup wire. These items will require very little additional space in the travel case, and may prove useful when setting up the station. Schematic diagrams of the equipment should also be taken afield, should troubleshooting be required.



A typical collection of equipment, antennas, spare parts and tools for a QRP DXpedition. The materials in the picture were packed into the portable typewriter case seen at the rear of this illustration, then transported in hand to Barbados during operation as 8P6EU/8P6FJ in 1973.

Spare parts are important when operating portable, and a few components thrown into the tool kit could be helpful. Critical components, such as the PA and driver transistors of the transmitter, should be taken along as spares. Fuses, spare batteries, and a collection of capacitors are often handy when a failure occurs. The W1CER parts kit contains .001, .01, 0.1, 2, 10 and 50-μF capacitors. Included also are rectifier diodes, high-speed switching diodes, general-purpose FET and bipolar small-signal transistors. Depending on the kind of circuit used, certain ICs are also included in the kit.

Salt water, and the air near salt water, has a notable effect upon some kinds of amateur equipment. The keyer paddle will develop poor electrical contacts after a period of time near the sea. A machinist's point burnishing file is handy for restoring the contacts of a key. Antennas mounted near the seashore for long periods should be coated with silicone grease to prevent corrosion. This is especially true if aluminum tubing is used in the antenna system. All joints in wire antennas should be soldered rather than twisted together. That will prevent salt air and spray from causing poor connections.

An SWR indicator is useful when afield: Some antenna pruning is usually required to provide a low SWR. A Transmatch can be taken afield for use with antennas that must accommodate more than one band of operation – end-fed wires or a 40-meter dipole that will be used also on 15 meters.

It is wise to check before traveling to a foreign land, to learn what the local power service is. Some parts of the world still use 25-Hz lines, while others use 50- or 60-Hz lines at some unusual voltage amount. It may be necessary to carry a power converter when ac operation is contemplated. Furthermore, the wall outlets in some countries are pretty strange to U.S. amateurs – an adapter may be necessary.

One final word of advice: When abroad it is important to exhibit proper radio conduct. Be especially courteous to the local amateurs you meet and talk with on the air. If you're operating from a hotel, use headphones rather than causing disturbances by operating with a loudspeaker. Be on the watch for interference to TV sets and radios. If the fault can't be corrected, cease operating. Also, a secondary frequency standard should be included with your radio gear. Straying out of an authorized amateur band could be embarrassing and expensive. Some foreign governments require that you have a crystal-controlled secondary standard before they will allow you to operate. A 100-kHz calibrator is usually adequate.

Wilderness Operation

The preceding section dealt with the problems encountered during operation at camping and DX locations. While such activities are certainly glamorous, especially for the DX operator, other portable ventures can produce similar rewards.

For over a decade a dominant activity at W7ZOI has been operation in connection with mountaineering and backpacking trips. The equipment requirements are different than they would be for other portable stints. All of the equipment must be carried on the back of the operator. This presents no problem if the walk is short and the purpose of the trip is specifically for "hamming" for a short duration. However, when the operation is secondary to a physically more ambitious goal, such as reaching the summit of a major peak, the criteria change.

There is a philosophy practiced by backpackers when assembling equipment for an extended outing. Simply stated, it is "Worry about the ounces – the pounds will take care of themselves." This approach must be extended to the design of any radio equipment that will be taken aloft.

The equipment should be designed, built and tested in the winter months. The gear is then ready when spring arrives. Hasty construction just before a trip invites equipment malfunction.

The primary consideration is weight rather than volume. Excessively dense packaging may be entertaining for the builder. However, if it makes the equipment less reliable and versatile due to



Author Hayward is seen here during a mountain trip on which he took the 40-meter Ultra Portable Transceiver described in this chapter. The battery pack is in his jacket pocket.

component crowding, it should be avoided. A reasonable size for a complete rucksack station is 2 X 5 X 7 inches. This allows ample room for circuitry while keeping construction straightforward. Batteries should be external.

The heaviest items to be considered are the batteries and antenna. The size of the batteries required will depend upon the power level of the transmitter and on the expected period of operation. This brings us to a major constraint – keep the power as low as possible. It is difficult to say how low it should be. At W7ZOI, the portable power levels used, mainly at 7 MHz (cw), have ranged from 8 W down to 250 mW of output. Our impression is that an output of 0.5 to 1 watt is near optimum for use in the contiguous states and in the less remote parts of Canada. This allows the use of Penlight cells or NiCads for short operating periods. Higher powers are useful in the more remote areas or for contest work.

Temperature extremes can have a dramatic effect on equipment performance. Oscillator instability is one common problem. In one cold experience (Mt. Adams in Washington State), a germanium transistor oscillator would not start. During another trip with vhf gear, low temperatures caused severe detuning of a frequency-multiplier chain.

There are a number of factors to consider when designing for temperature extremes. Semiconductors with wide operating-temperature ranges are suggested. Crystal control is recommended. While this is not mandatory for a receiver, it is highly desired for a transmitter. Frequency accuracy is of importance if schedules are made with other stations. The crystals should be built into the equipment and selected with a switch. Loose crystals are easily lost. Critical tuned circuits should be avoided. Finally, the equipment should be tested under severe conditions. While the home freezer can be used, it's usually more fun to take the equipment into the field when the first winter storm arrives.

There are other criteria that might be applied when designing equipment for mountaineering application. They might seem extreme, but have been found useful. First, it is desirable that the equipment be operable with gloves being worn by the operator. Sharp edges should be removed. This lessens the possibility of tearing holes in a tent or sleeping bag. It's worthwhile to build the equipment so that it may be operated in the dark. This is particularly useful during winter trips when the rig must reside *inside* the sleeping bag with the operator. Both must be warm to function well. Most batteries will decrease in output voltage and energy capability when cold. Provision should also be made to keep the battery pack warm.

As an aside, a winter trip on skis or snowshoes is an especially enjoyable time for taking the radio gear along. Owing to the long nights, it is often necessary to spend from 12 to 14 hours at a stretch in a mountain tent or snow cave. The ham gear helps to pass the time. Also, the possibility of being stranded by a change in the weather is greater. Reliable communications capability could be very valuable.

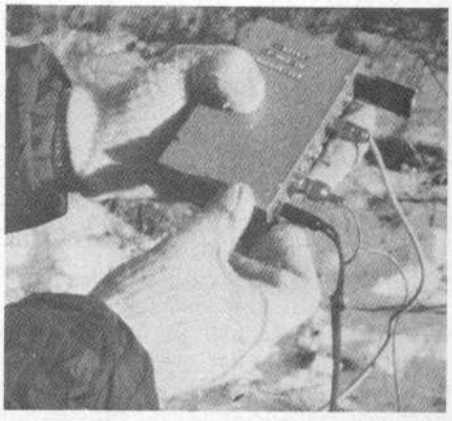
Questions asked by the prospective portable operator are, "What band and mode to use?" First, cw is preferred over ssb. The equipment tends to be more reliable owing to the simplicity. The narrower information bandwidths help. However, the operator should be proficient with conversational cw – that is, he should be able to copy the code without having to put anything on paper other than logging information and a few notes. Physical strain and the effects of a harsh environment make normal operation difficult. The less proficient cw operator should consider ssb or dsb equipment.

The choice of frequency is difficult, and partially subjective. For summer operation, 40 meters is ideal. The band remains open for short hops during the daylight hours and well after sundown. Eighty meters is better for winter use. Noise levels are too high for 80-meter effectiveness in the late summer. The 20-meter band is excellent for the operator with an interest in evaluating unusual locations for DX effectiveness. In Field Day contests where the writers have participated with QRP, 20 and 40 meters have produced the largest number of contacts.

The vhf bands are, in many ways, the optimum bands to explore. It is at these frequencies where the real benefits of a high, mountain-top location are obtained. Equipment is more complex, but not unreasonable. Sideband or cw are still the recommended modes. Cw has the edge for long-haul work. However, more vhf operators will be comfortable with phone.

Antennas can be a problem in the mountains. The standard carried in the W7ZOI rucksack is a 7-MHz dipole with a 40-foot transmission line of RG-174. While a larger size cable would have less loss, the added weight would be intolerable. The 40-foot cable mentioned has 0.7 dB loss at 7 MHz. The center insulator for the dipole is made from a scrap of pc board (unclad). The antenna is always operated in the inverted-V configuration. This has the advantage that only one support is needed. Trees are ideal when available. Above timberline, a small telescoping whip antenna is carried. The unit used collapses to 14 inches and is 12 feet long when extended. It has always been more effective to use the whip as a center support for the dipole than to load it against a group of radials. This whip was a surplus item and the source is unknown. A good substitute would be a long fishing rod. Fiberglass rods up to 20 feet long are available through the *Sears Fishing and Boating Catalog.*

The usual methods for putting a line into a tree are effective. A lightweight fishing line is preferred over a heavier nylon cord. Also, it is wise to carry a 1-ounce weight with the other gear. On one occasion, one of the writers found himself without an adequate weight to tie to the end of a nylon cord. Not wanting to miss out on Field Day, a



A close-up view of W7ZOI's hands holding the 40-meter transceiver. Despite the gloves he was able to operate the built-in key lever during cw transmissions.

Transmatch was attached to the rope and hurled aloft.

A small beam is recommended at vhf. This may be lashed to an ice ax for above-timberline operation. The antenna should be capable of easy assembly with a minimum of loose parts needed.

No matter what equipment is used, or what the goals of the operator are, respect should be maintained for other mountain travelers. Rambling into country which is devoid of roads or even trails offers an escape from the daily pressures and routine that have become a dominant part of our society. Backpacking and mountaineering are increasing in popularity and, unfortunately, the "wilderness" is often an area with a number of visitors. The last thing a fellow hiker wishes to hear is the blare of cw rushing from an overdriven speaker. Headphones should always be used!

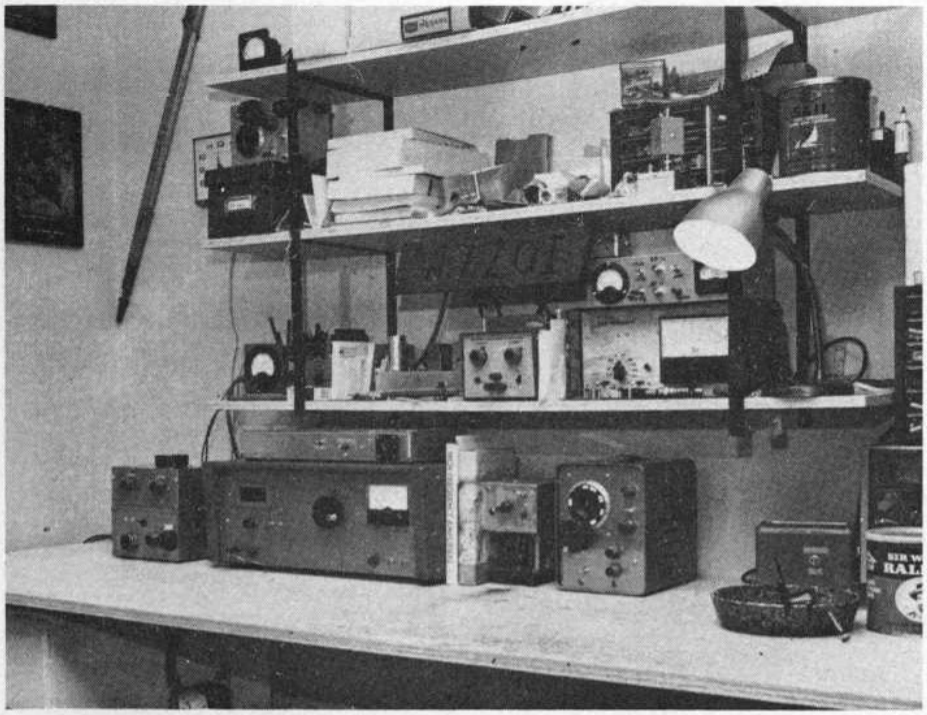
Finally, the radio amateur who carries his hobby into the back country should be prepared for an occasional emotional dilemma. Should he compromise his hiking or climbing goal in order to get on the air or should he pursue the primary goal? In this age dominated by high technology, the answer is obvious. Climb the mountain!

QRP Operation

Although portable operation has been the motivation for the work of the writers respective to QRP equipment, this is not typical. The more common QRP operation occurs from the home station. The operator's motivation is to add excitement and adventure to contacts that would otherwise offer minor challenge. Much of the present QRP popularity results from ready availability of commercial equipment at reasonable prices. Fortunately, the excitement has spurred many amateurs to build their own gear, allowing them to gain doubly from their operating activities.

The criteria for success with QRP gear are not all that different than they are for high power. The key is in the antenna system and in a wise choice of operating frequencies for a given time of the day or year. These decisions are more critical with low power.

There are a few operating techniques that can aid the QRP operator. Generally, he will be more successful if he calls other stations rather than calling CQ. Often, it is better to call a station as he is finishing a contact rather than answering a CQ from a loud station. Another trick is to add some additional information to a call, letting the fellow on the other end know that there is a reason for the signal being weak. This can be successful even when calling CQ. However, it is usually not enough to tack a "QRP" on to a CQ. To some operators, this merely implies that the station signing "QRP" is running less that 100 watts, only 10 dB down from the legal limit. A much more effective format is CQ CQ de QRP 1 watt, W7ZOI W7ZOI and so forth. The writers feel that these methods should not be applied except for output powers of less than 1 or 2 watts.



Photograph of the W7ZOI home station. All of the amateur equipment and test gear is homemade. The operation position serves double duty by also being a workbench.

An excellent time for the QRP operator to make a large number of contacts is during contests. Here there are a larger number of stations available to be worked. Of greater significance, they are anxious to work anyone they can, and will not be upset with a less than ear-shattering signal.

One of the best contests for QRP work is Field Day, for a large number of similar stations are active during the same period. This has been aided by the individual listing of low-power stations in the *QST* results. A club Field Day using QRP is an interesting and unusual experience. TERAC (Tektronix Employees' RAC, K7AUO) has participated in the QRP category for several years. While many of the more competitive, contest-oriented members have avoided the activity, others with general interests have participated and have enjoyed low-power work. A QRP Field Day tends to be a more relaxed affair. This is aided by the conspicuous absence of the roar of a generator.

Although QRP operation may seem casual, there are some who have become accomplished in this area. Many operators have achieved WAS and WAC with quite low powers, and a few have qualified for DXCC with less than 5 watts of rf output. Generally, the more successful QRPers are cw enthusiasts.

An interesting experiment is to attempt contacts with as little power as possible. Minimum-power experiments were performed during a number of contacts between W7ZOI and WA6YVT in 1969 and 1970 on 40 meters. To attach some legitimacy to the contacts, a strict format was established. Contact was established initially with an output of 1 to 3 watts from W7ZOI. If the reports from WA6YVT (in the Los Angeles area) were favorable, the output power would be decreased. A step attenuator was used in a matched 50-ohm antenna system to ensure that the output power at W7ZOI was well defined. At each power level, an arbitrarily chosen 4- or 5-letter word would be sent. The word was repeated several times. WA6YVT would then repeat the word to confirm that information had actually been exchanged. It was not possible for a vivid imagination to serve as a substitute for actual copy.

While experiments were conducted to evaluate the power levels that would be suitable for portable equipment, they turned out to be generally interesting. In nearly all cases where the attenuator was put into the transmission line, information was exchanged at 100-mW output. Often 50 mW was successful. The lowest power producing a *real* exchange of information was 2.5-mW output. Immediately after that contact, the output power was confirmed with a high-frequency oscilloscope. One con-

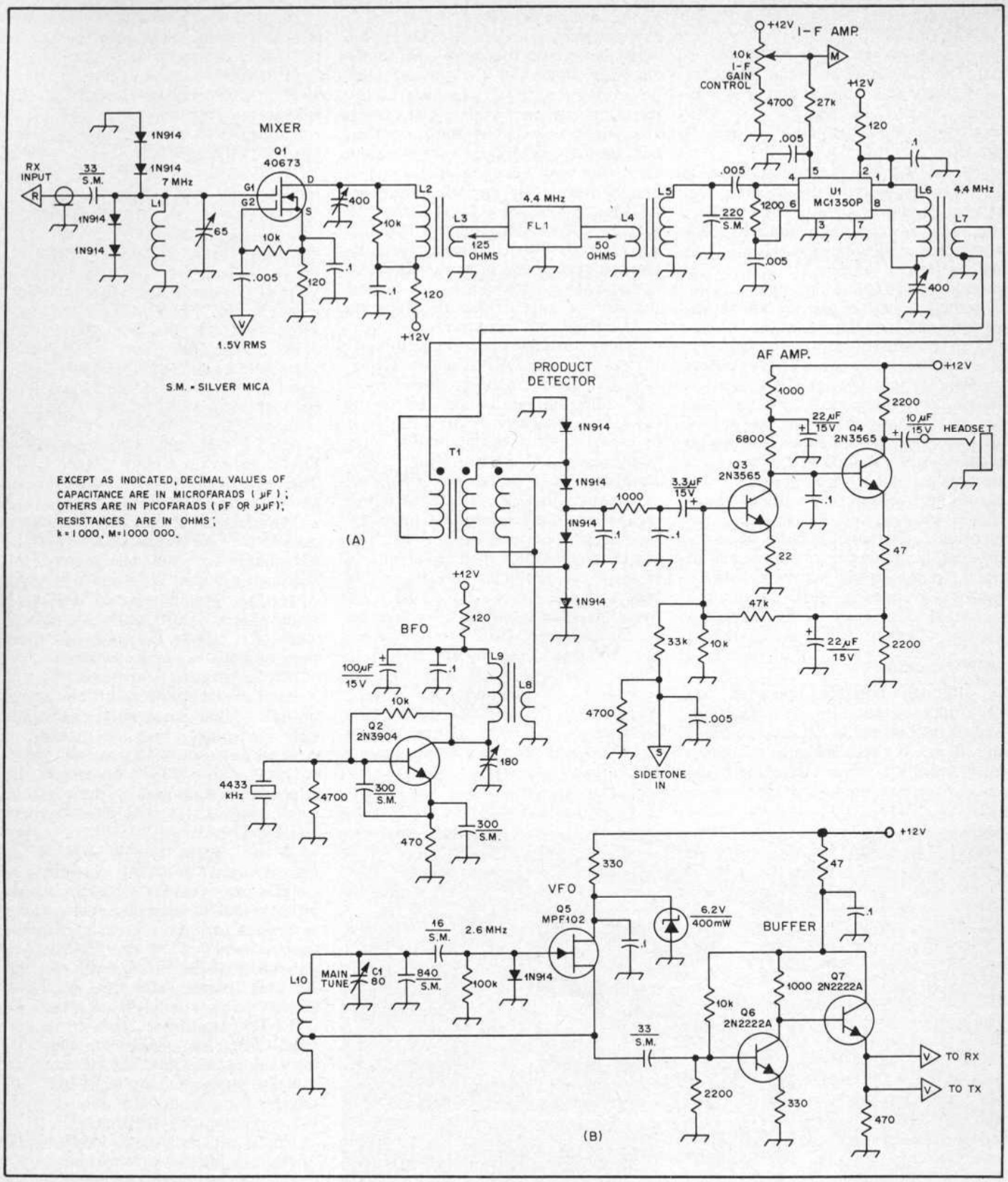
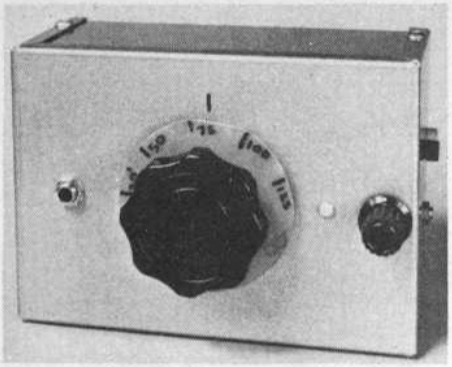


Fig. 1 — Schematic diagram of the VFO and receiver portions of the 7-MHz transceiver. Fixed-value capacitors are disk ceramic unless otherwise noted. Fixed-value resistors are 1/4- or 1/2-W composition. Variable capacitors without part numbers can be mica compression trimmers (surplus Teflon or ceramic trimmers were used in the authors' unit).

C1 — 80-pF air variable.
FL1 — Crystal filter, ladder type (see text).
L1 — 30 turns No. 26 enam. wire on T50-2 toroid core.
L2 — 36 turns No. 26 enam. wire on T50-2 toroid core.
L3 — 3-turn link over L2 winding, No. 26 enam. wire.
L4 — 8-turn link over L5 winding, No. 26 enam. wire.
L5 — 34 turns No. 26 enam. wire on T50-2 toroid core.
L6 — 36 turns No. 26 enam. wire on T50-2 toroid core.
L7 — 5-turn link over L6, No. 26 enam. wire.
L8 — 8-turn link over L9, No. 26 enam. wire.
L9 — 43 turns No. 26 enam. wire on T50-2 toroid core.
L10 — 29 turns No. 24 enam. wire on T50-2 toroid core, tapped 7 turns from ground end. Coat with *Q* dope.
T1 — 12 trifilar turns No. 30 enam wire on Amidon FT37-61 ferrite toroid core.
U1 — Motorola IC.
VR1 — 6.2-V, 400-mW Zener diode.



Front of the 7-MHz superheterodyne QRP transceiver. The jack at the left accommodates an electronic keyer. Audio output is taken from the side panel (right). The case measures 3 X 5 X 7 inches.

clusion was that the method was an accurate means for evaluating the overall condition of the path within an accuracy of 3 dB, far more accurate than an S-meter reading. However, the lowest powers were successful only when the propagation conditions were favorable and while noise levels were low. Similar methods would be useful for the study of vhf propagation. Additionally, the weak-signal experience would be valuable to the operator with an interest in modes such as moonbounce.

A Superheterodyne CW Transceiver for 7 MHz

For the beginning experimenter with an interest in QRP and portable operation, a direct-conversion transceiver is ideal. Construction is straightforward, owing to the simplicity of design. When a higher level of performance is desired, especially in the receiver, it is better to build a superheterodyne system. Transceive operation is still desirable for some applications. Contests such as the ARRL Field Day are an example.

The transceiver described in this section is based upon the preceding design criteria. The unit tunes a 100-kHz segment of the 40-meter cw band. A full transceive type of transmitter with an output of 1.5 watts is employed. The receiver selectivity is provided by a homemade 3-pole crystal filter of the lower side-band ladder type. The bandwidth is 250 Hz and the rejection of the undesired sideband is approximately 60 dB. A completely electronic T-R system is included, providing smooth, transient-free control. Owing to the subtleties of the design, especially in the construction and alignment of the crystal filter, this project is not recommended for the inexperienced experimenter. No pc information is available.

Shown in Fig. 1A is the receiver section of this transceiver. The front end employs a dual-gate MOSFET as the mixer. A single tuned circuit (L1) serves as the preselector. Four 1N914 diodes are used to protect the input. This would not be needed if the electronic T-R system were not employed. The mixer was not damaged when the diodes were omitted. However, they were included as a precaution against an improper termination at the antenna terminal. This could lead to high rf voltages at gate 1 of Q1. The output network of the mixer (L2-L3) is designed to present a termination of 125 ohms to the crystal filter. The design of the 4.4-MHz filter will be presented later.

At the output of the crystal filter is a network (L4-L5) that presents a 50-ohm termination to the filter. This is followed by an i-f amplifier which uses an MC1350P IC. This circuit provides a gain of approximately 40 dB, and allows for a gain variation of 60 dB. No agc system is included in this transceiver. Receiver muting is realized by application of +12 volts to the arm of the manual gain control potentiometer.

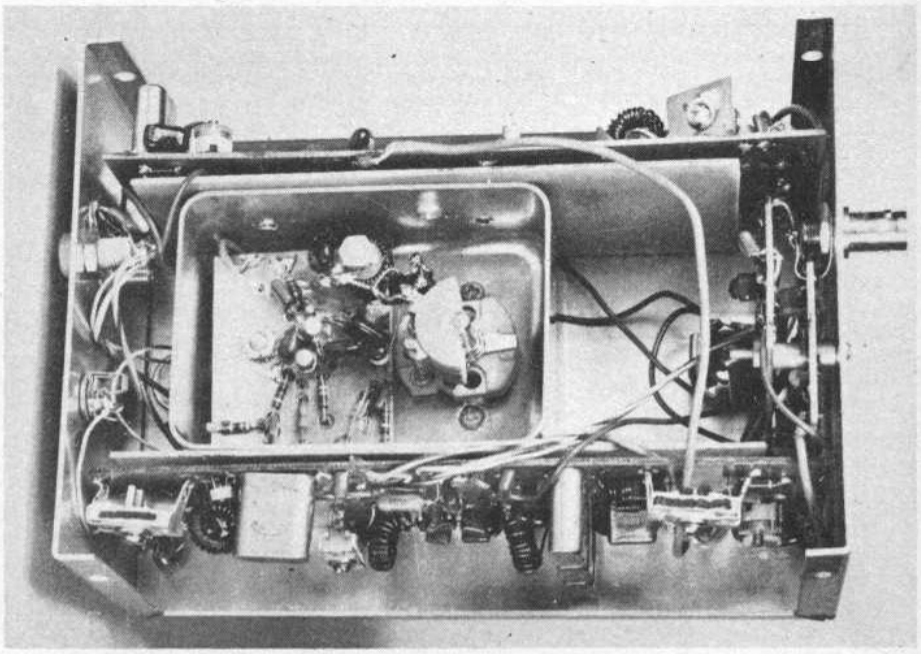
The i-f output is matched to 50 ohms (L6-L7) and then routed to a product detector utilizing four diodes. Originally, only two diodes were used. However, it was found that the improved balance obtained with four diodes provided less noise modulation of the BFO signal that found its way into the i-f amplifier. Note that the primary of T1 is balanced, being grounded only at the output of the i-f (L7). This also improved the balance. Such precautions would not be necessary if a less dense packaging format were used. Ground loops in a single continuous ground foil can cause these problems. They are avoided in systems employing a number of smaller circuit boards.

BFO injection is provided by Q2. This oscillator is standard except that some means must be provided for adjusting the crystal to the proper frequency. All of the crystals used in the transceiver, including those in the filter, were cut for the same frequency. Experimentation may be required on the part of the builder to establish the proper capacitance across the crystal.

A two-stage audio amplifier is used. Emitter degeneration is employed in both stages (Q3 and Q4) to ensure that linearity is preserved under large-signal conditions. The sidetone signal is injected into the base of Q3 during transmit periods.

The VFO for the transceiver is shown in Fig. 1B. A JFET is employed as a Hartley oscillator. Because the frequency is low (2.6 MHz), stability is excellent. The oscillator is tuned by means of an 80-pF air variable capacitor, which is driven by a Jackson Brothers vernier-drive mechanism. The fixed-value capacitance across the oscillator coil (L10) was chosen to provide the desired tuning range. A trimmer might be a useful addition to ease alignment. The oscillator is buffered with a two-stage amplifier, Q6 and Q7.

The VFO is built in a small aluminum box. This box is fastened securely to the front panel by means of standoff posts. Because all of the oscillator components are mounted securely to the smaller housing, mechanical stability is good. It was found that the transceiver



Interior of the 7-MHz superheterodyne transceiver. At the center is the 2.6-MHz VFO compartment. The bottom pc board contains the receiver. At the top is the transmitter module. The small assemblies (2) at the right are the TR and sidetone boards.

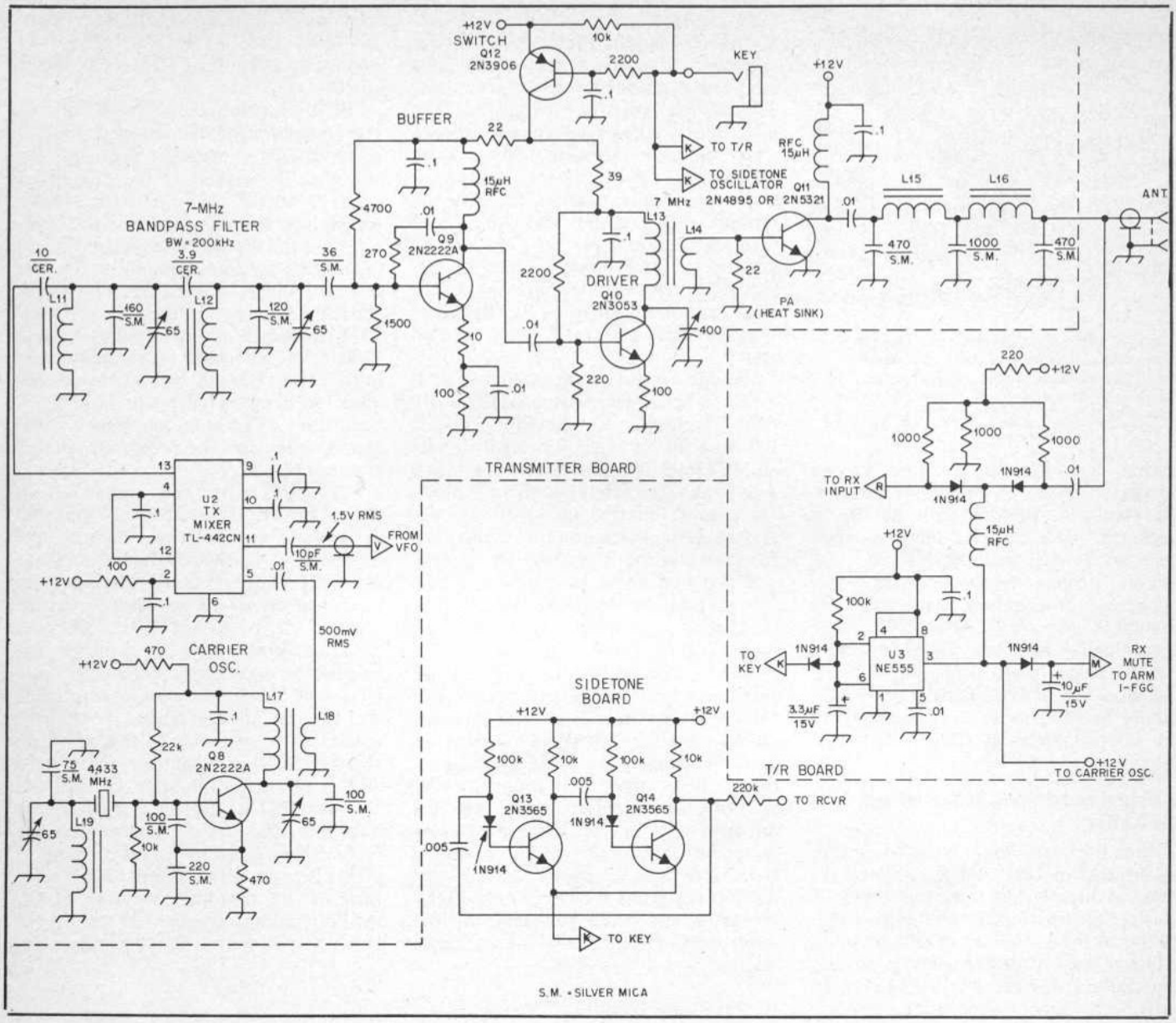


Fig. 2—Schematic diagram of the transmitter, T-R switching and sidetone circuits of the 7-MHz transceiver, Fixed-value capacitors are disk ceramic unless otherwise noted. Fixed-value resistors are 1/4- or 1/2-W composition. Variable capacitors can be mica compression trimmers. The SN-76514 mixer IC has been reidentified as TL-442-CN by Texas Instruments. It may be procured under either part number.

L11, L12—23 turns No. 24 enam. wire on T50-2 toroid core.
L13—20 turns No. 22 enam. wire on T50-2 toroid core.
L14—2-turn link over L13, No. 22 enam. wire.
L15, L16—14 turns No. 22 enam. wire on T50-2 toroid core.
L17—42 turns No. 26 enam. wire on T50-2 toroid core.
L18—4-turn link over I17, No. 26 enam. wire.
L19—40 turns No. 26 enam. wire on T50-2 toroid core.

could be dropped 2 or 3 inches onto the operating table with no detectable frequency shift.

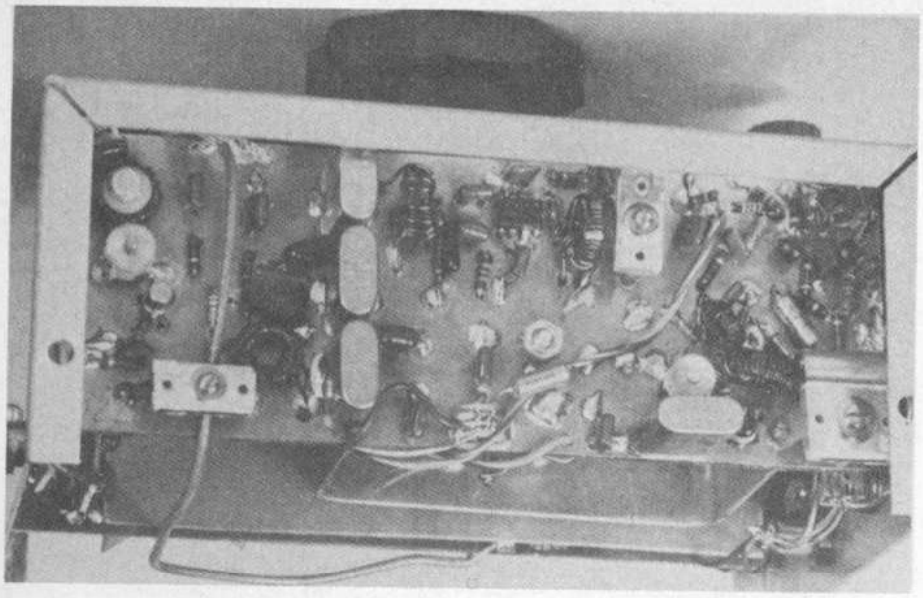
Shown in Fig. 2 is the transmitter portion of the transceiver. Three circuit boards are employed. One board is used for the sidetone oscillator, with a second containing the electronic T-R switch. The third board contains the rest of the transmitter.

The carrier oscillator (Q8) is a bipolar transistor operating in the Colpitts configuration. To adjust the frequency to the center of the i-f passband it was necessary to add inductance (L19) and capacitance to the circuit. The crystal oscillator delivers 0.5 volt rms to the transmit mixer, U2. An SN76514 was selected for the mixer, owing to the internally contained biasing resistors. An MC1496G could be used in this application.

The output of the mixer is applied to a two-pole bandpass filter. The coupling capacitors between the resonators and into and out of the filter are critical and should not be substituted casually. The output of the filter is terminated in the 50-ohm input impedance of a feedback amplifier, Q9. The use of feedback is very useful where a well defined input immittance is desired. The buffer is followed by a driver, Q10. Both Q9 and Q10 are keyed by a pnp switch, Q12.

The final amplifier, Q11, uses a Fairchild 2N4895 TO-5 type of transistor. A 2N5321 would function as well in this circuit. The output network is a half-wave filter. The output stage should have a small heat sink.

The electronic T-R switch uses a pair of silicon switching diodes. The antenna is permanently connected to the transmitter. The receiver is also connected when the switching diodes are biased to an *on* condition. When the key is depressed, the 555 timer IC (U3) is triggered on. The output at pin 3 is then in a high state and supplies power to the transmitter carrier oscillator, Q8. The receiver is also muted, and the T-R diodes are reverse biased slightly. When the key is opened, U3 begins to time



Close-up view of the receiver pc board. The front end is at the left, followed (right) by the three-pole ladder filter. The i-f amplifier is at the center of the board, with the product detector and audio amplifier at the upper right. In the lower right corner is the BFO.

out. There is a short period before the receiver again becomes operational.

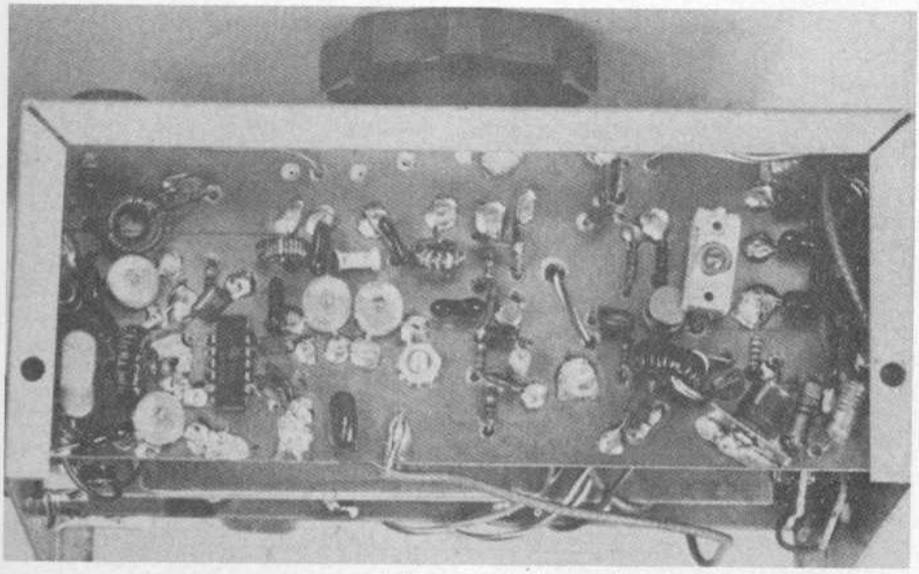
Crystal-Filter Construction

The crystal filter is shown in Fig. 3A. All three crystals were at the same frequency with a maximum deviation of 10 Hz. The crystals were measured using the methods outlined in chapter 5. This information was then used to design the filter using the methods outlined by Zverev (see bibliography). The design predicted the values of the coupling capacitors and the resistances needed to properly terminate each end of the filter. The measured filter response was in excellent agreement with the design goals and no empirical changes were required for any of the values. The crystals used were surplus European TV color-burst types with a frequency of 4433 kHz.

Shown in Fig. 3B is the circuit of a two-pole filter of similar characteristics. This filter should be much easier to build on an empirical basis. Four crystals should be ordered at one time. An oscillator is built next. A frequency counter can be used to select the two crystals that are closest in frequency (Y4 and Y5). The other two are set aside for use in the BFO and carrier oscillator. Y4 and Y5 are soldered into place between the mixer and the i-f amplifier (Fig. 1A). Various values of C2 are tried until the desired results are obtained. For a cw filter, a good starting point for C2 would be 470 pF. It may be necessary to change the terminating impedances. This can be done by experimenting with the number of turns on L3 and L4. Although this procedure may sound a bit terrifying to the beginner, it is not difficult to obtain suitable results. Experimentation *will* be required though!

A receiver was described in chapter 5 which uses a filter with a single crystal. This circuit could be used in this transceiver. However, the performance difference between a single-pole response and that realized with two or three crystals is profound.

The performance of this transceiver has been excellent. It has been used for portable and home-station QRP operation. Especially enjoyable has been the crisp response of the receiver and the smoothness of the control circuitry.



Close-up view of the 40-meter transmitter module. The carrier oscillator is at the left. Moving to the right, one can see the IC transmit mixer, double-tuned output filter, two keyed Class A amplifiers, and finally the PA stage.

Transceivers and Integrated Stations – Construction and Operation

In this section we are presenting construction projects for complete stations. Most are suitable for home-station use and operation from a portable location. Various degrees of sophistication are considered. The simplest station represents perhaps, the most elementary "stripped-down" station that is suitable for communications. Included also is a station which approaches the ultimate that the amateur can construct with limited tooling and test facilities.

Transceivers and Trans-Receivers

A *transceiver* is a unit which shares some of the circuits during the transmit and receive modes. Although an outboard VFO can be used with some commercial transceivers to provide separate frequency control for the transmit and receive functions, the composite transceiver contains a single local oscillator which serves both modes.

Conversely, a trans-receiver contains in its cabinet an independent transmitter and receiver, each of which has its own tunable local oscillator. More often than not the power supply is shared by the two circuits, as are the changeover relay (or solid-state TR circuit) and cabinet.

Frequency Offset

An important part of a transceiver is the frequency-offset circuit. When the equipment is designed to accommodate both sideband modes (upper and lower sideband), the tunable local oscillator must be shifted in frequency when going from upper to lower sideband,

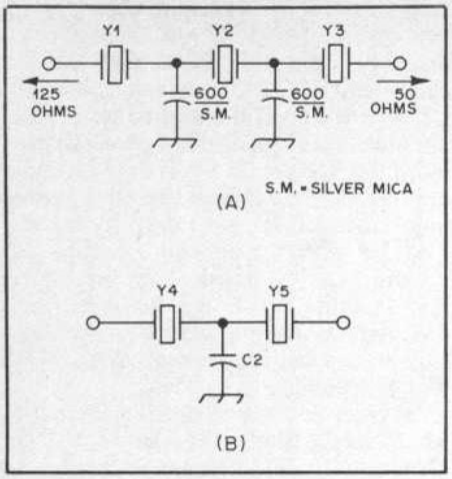


Fig. 3 — Circuit of FL1, the crystal filter shown in Fig. 1. At A is the filter used in the 7-MHz transceiver. A simplified version of the filter is shown at B (see text). The crystals should be on the same frequency — within 20 percent of the filter bandwidth. Bandwidth of the circuit at A is 250 Hz. The crystals are 4.433-MHz units.

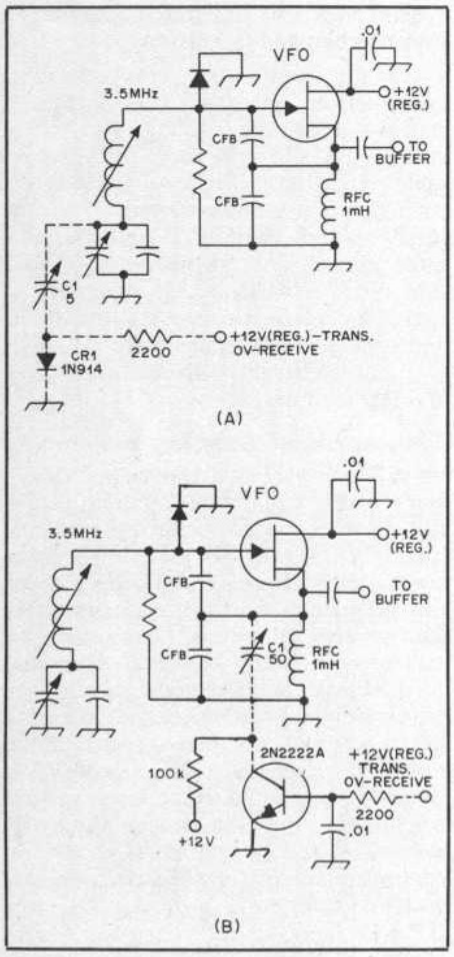


Fig. 4 — Examples of VFO offset circuits. In the circuit at A, a low value of capacitance (C1) will cause a large change in frequency. A larger amount of capacitance (C1) is required in the circuit at B to effect a similar frequency change. A diode or a transistor can be used as an electronic switch in either circuit.

and vice versa. In a heterodyne type of transceiver the operating frequency of the BFO must be shifted also, placing the injection frequency on the proper side of the i-f filter-response curve (usually 1.5 kHz above or below the center frequency of the i-f filter). The shift in LO frequency is necessary in order to maintain accurate dial calibration for the main-tuning control.

Direct-conversion cw or dsb transceivers need to have a frequency-offset circuit if they are to be compatible with other transceivers employed during QSOs. With no offset circuit in a direct-conversion transceiver, the transmitted signal from the latter would probably appear at or near zero beat on the other station's receiver (undesirable). As a consequence — if the other station happened to discover someone calling at zero beat (no audio beat note in his phones), he would compensate by moving his tuning dial. This would necessitate readjusting the main-tuning dial of the direct-conversion transceiver. The process would be repeated during each transmission, and the two stations would be "walking" across the band until they signed off!

Most cw operators prefer an audio beat note which occurs between 500 and 1000 Hz. If, for example, the operator liked to listen to a 700-Hz note during cw operation, the local oscillator offset would be 700 Hz when changing from transmit to receive. The transmit frequency in such a case would be 700 Hz *lower* than the receive frequency to assure compatibility with most commercial transceivers in use.

Fig. 4 shows circuits in which a diode or a transistor can be used to actuate an offset circuit in the tunable local oscillator. Ordinarily, the offset is turned on and off by means of the transceiver TR circuit (relay or solid-state logic).

Using RIT

A useful feature in a transceiver is RIT (receiver incremental tuning). The addition of RIT permits the operator to tune his receiver a few kHz above and below the receive frequency without disturbing the transmit frequency. The RIT circuit enables an operator to select the desired audio pitch during cw reception, or to tune in an ssb signal so that the voice quality suits his listening tastes. An RIT circuit is beneficial in DX pileups, when the DX station is listening a kHz or two away from his transmit frequency. For all practical purposes in this discussion we can call RIT an *ultra* fine-tuning control. During the transmit mode the transceiver changeover circuitry disables the RIT so that the transmit frequency remains the same as indicated on the frequency-readout dial.

Electrically, the RIT circuit is similar to that for the frequency-offset system discussed earlier. The principal difference is that with RIT one can control the amount of offset from the front panel of the transceiver. Fig. 5 shows an RIT circuit which can be connected to a tunable local oscillator.

The offset circuits of Fig. 4 are identical in principle, but the circuit at A requires a fairly small capacitance value at C1 to keep the offset amount within practical limits. If the same circuit were connected across the lower feedback capacitor, C_{fb}, C1 would have

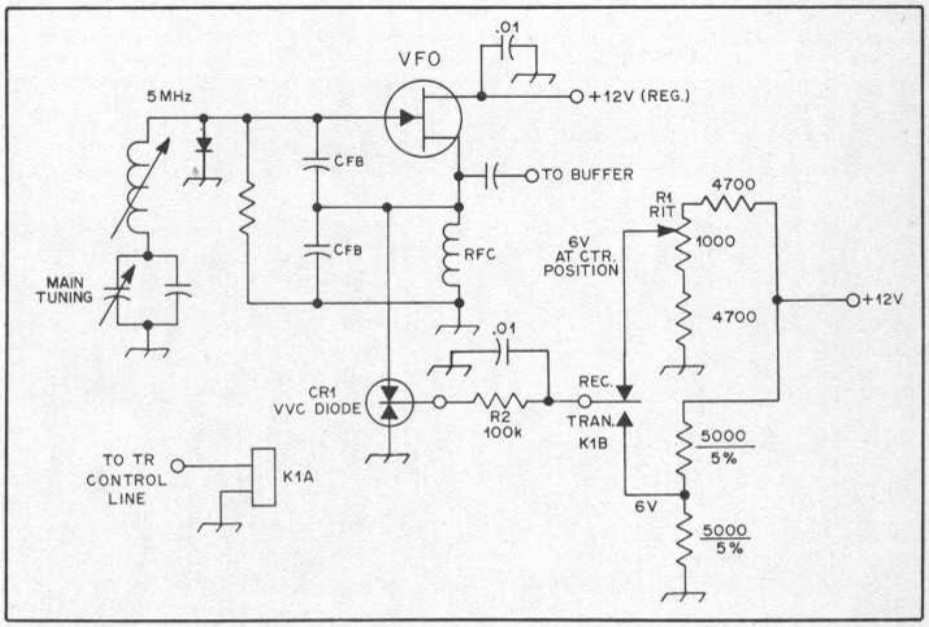


Fig. 5 — Circuit example of an RIT circuit for use in the VFO of a transceiver (see text). CR1 is a voltage-variable capacitor diode, which by means of R1 can move the VFO frequency a few kHz above or below the transmit frequency during the receive mode.

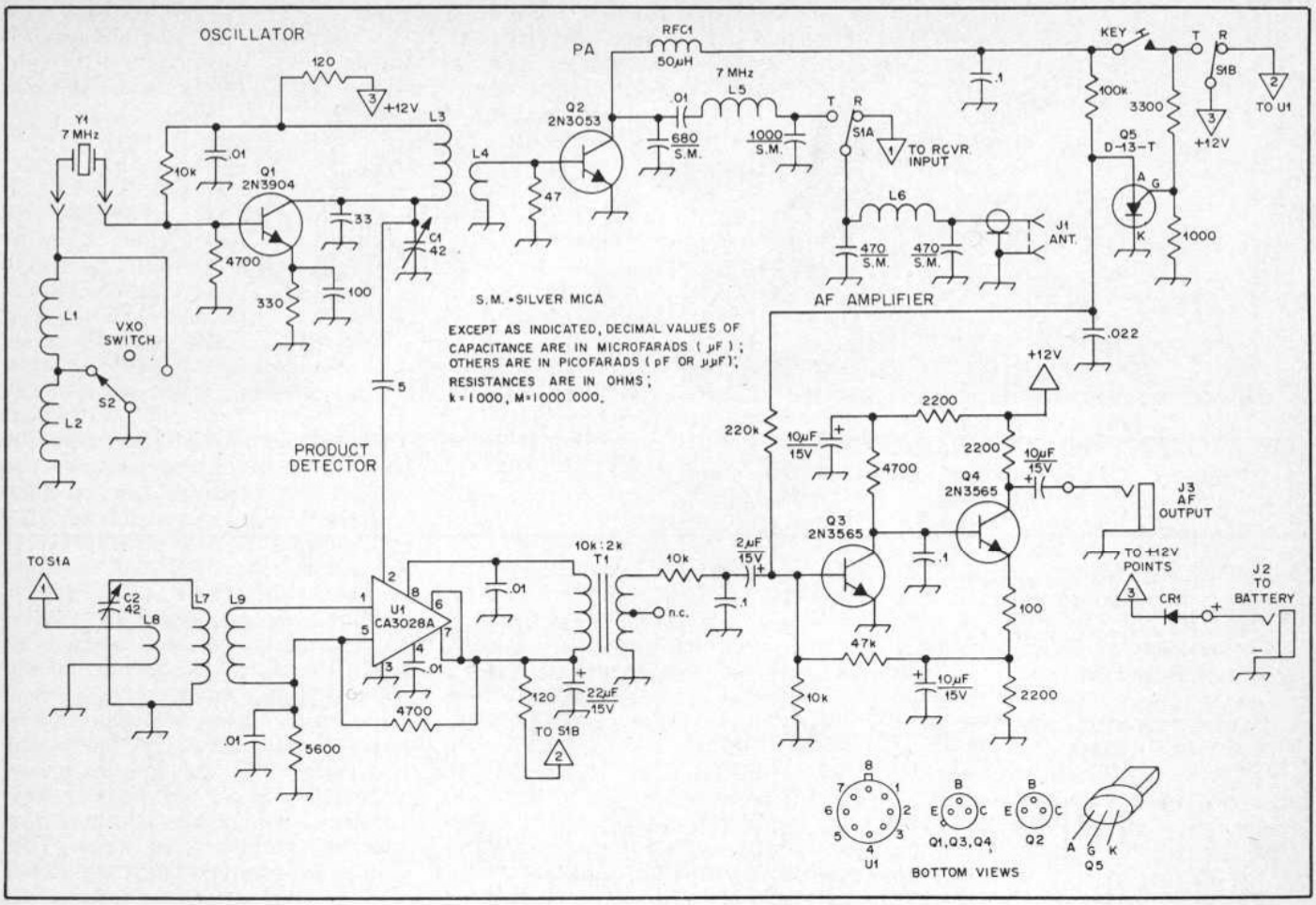


Fig. 6 — Schematic diagram of the ultra-portable transceiver. Fixed-value capacitors are discussed in the text. Resistors are 1/4-watt composition.
C1, C2 — Subminiature ceramic trimmer, 42pF maximum.
CR1 — Silicon rectifier diode, 50 PRV, 500 mA.
J1 — BNC chassis-mount coax connector.
J2, J3 — Phone jack.
L1, L2 — 20 turns No. 22 enam. on Amidon T-50-2 toroid core (Amidon Assoc., 12033 Otsego St., N. Hollywood, CA 91607).
L3 — 44 turns No. 28 enam. on T50-2 core.
L4 — 4 turns No. 22 enam. over L3 winding.
L5, L6 — 14 turns No. 22 enam. on T50-2 core.
L7 — 60 turns No. 28 enam. on T50-2 core.
L8 — 5 turns No. 22 enam. over L7 winding.
L9 — 10 turns No. 22 enam. over L7 winding.
T1 — Miniature 10,000-ohm to 2000-ohm transformer. Center tap not used.
U1 — RCA integrated circuit.
Y1 — 7-MHz crystal.

to be somewhat greater in capacitance to effect the required offset amount, as shown at B. A diode or a transistor can be used as a switch at either point in the VFO circuit. Regulated voltage should be supplied to the switching device to assure frequency stability. C1 should be an air-dielectric trimmer or glass piston trimmer — a further aid to stability (mechanical and electrical).

The RIT example at Fig. 5 is a simplified one. In a practical transceiver some additional switching provisions would be included to remove the RIT from the circuit altogether when normal transceiving was desired. This would require placing a fixed-value resistive divider in the circuit to replace the potentiometer, R1, during receive. The Varactor diode, CR1, should have identical voltages applied to it during the transmit and receive modes when RIT is not needed. When the RIT is actuated, the center position of R1 should provide the same dc voltage to CR1 that is present in the transmit mode. Then, the receive frequency can be varied above and below the transmit frequency (± 3 kHz, typically) by means of R1. The VFO readout dial should be calibrated with K1B in the transmit position.

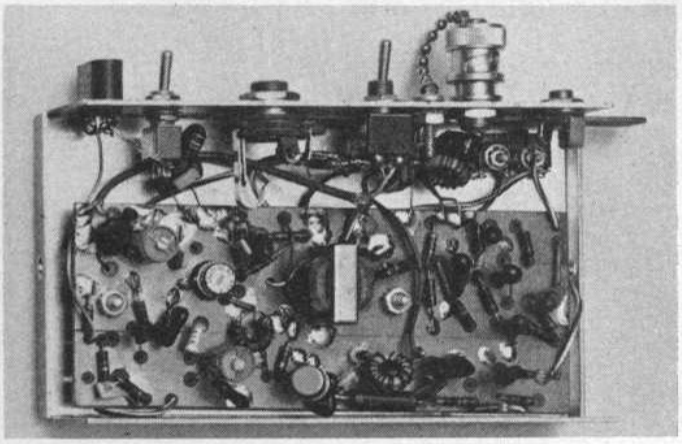
It is worthy of mention that addition of the offset or RIT circuits to a VFO can increase the drift of an oscillator. This can result from the heating of the Varicap-diode junction, or from the junction-capacitance changes in the switching transistor or diode in the offset circuits we have illustrated.

An Ultra-Portable CW Transceiver for 7 MHz

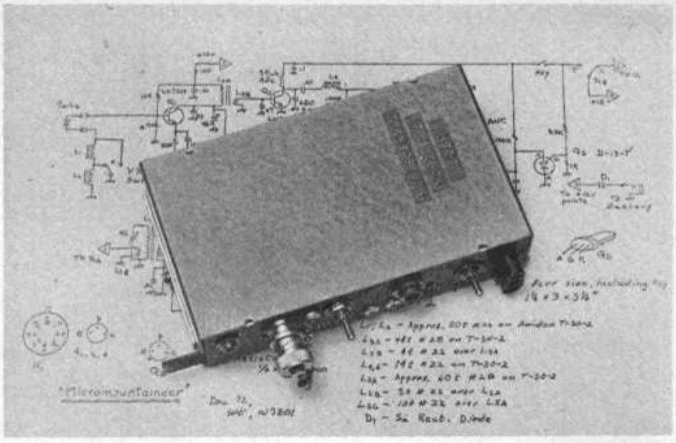
The design of any equipment is dictated to a large extent by the intended application. Home-station equipment may be large physically, and may contain as much sophistication as the builder desires. For portable operation, however, it is desirable that the equipment be physically small. A major criterion for miniaturization is simplicity. This forms the basis of the transceiver described in this section.

Shown in Fig. 6 is the circuit. Q1 functions as a crystal-controlled oscillator operating at 7 MHz. This stage serves a dual role. It drives, Q2, the power-output amplifier of the transmitter. Second, it provides BFO injection for the direct-conversion receiver.

Initially, it may seem limiting to utilize crystal control for both the transmitter and the receiver. But, if the transmitter is to be crystal controlled, it is generally unnecessary for the receiver to have the ability to receive on different frequencies. On the hf bands contacts occur rarely on a split-frequency basis. It is mandatory though that the crystal oscillator have capability for slight adjustment. If this were not present, it would be possible for another station to be exactly zero beat with the transceiver without the operator realizing its presence, as mentioned earlier in this chapter. This tuning is achieved by moving the crystal frequency slightly by switching in series inductors, L1 or the



Interior of the ultra-portable transceiver. Double-sided board is used in this equipment.



This 40-meter transceiver measures only 1-1/2 X 3 X 5-1/4 inches. The key paddle is visible at the lower left.

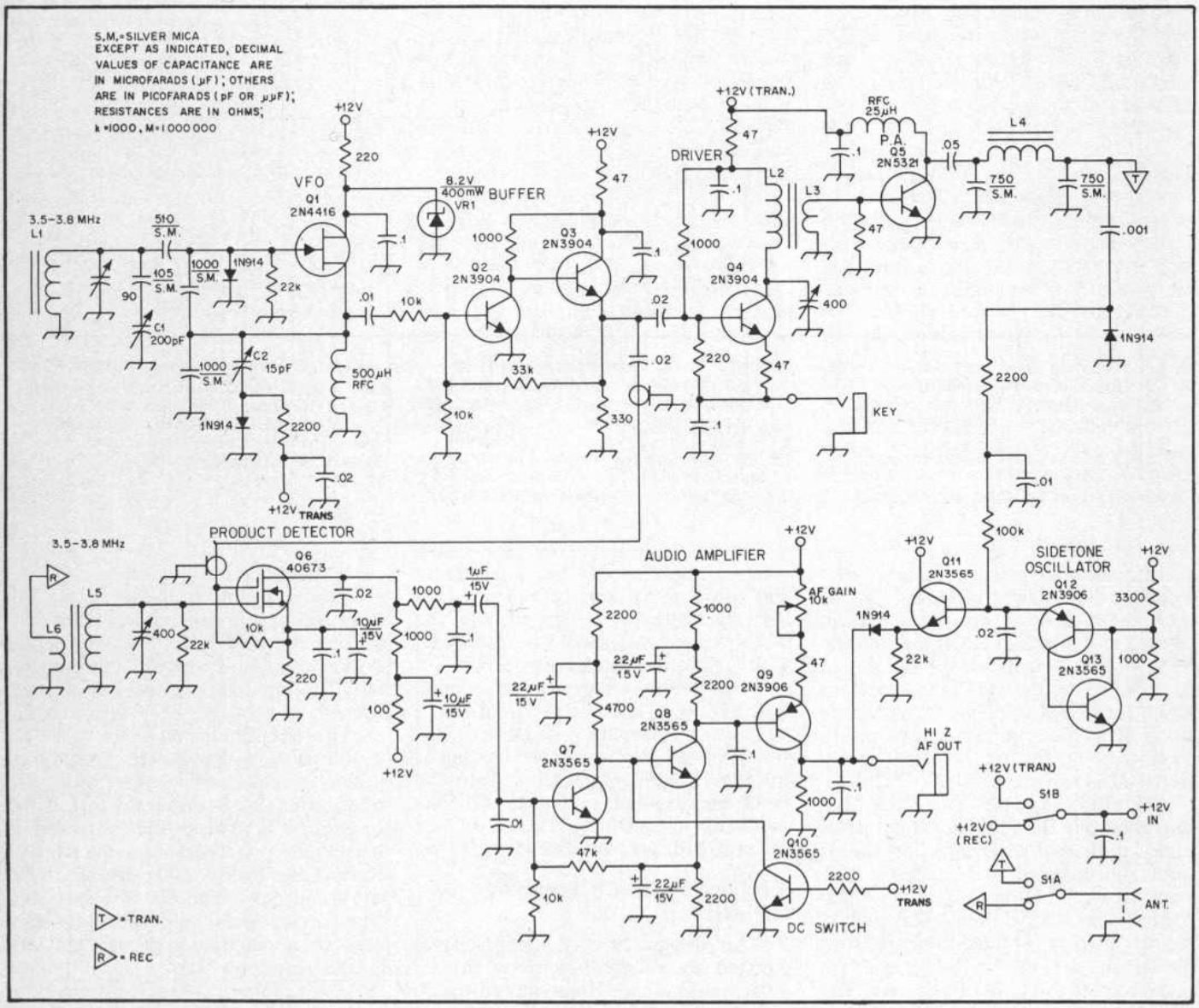


Fig. 7 — Schematic diagram of the KL7IAK 80-meter transceiver. Fixed-value capacitors are disk ceramic unless otherwise noted. Polarized capacitors are electrolytic. Unlabeled variable capacitors are mica compression trimmers. Fixed-value resistors are 1/4- or 1/2-W composition.

C1 — 200-pF air variable with vernier drive.
C2 — Pc-board-mount 15-pF air variable.
L1 — 30-turns No. 24 enam. wire on T68-2 toroid core.
L2 — 43 turns No. 28 enam. wire on T50-2 toroid core.
L3 — 5-turn link over L2, No. 28 enam. wire.
L4 — 21 turns No. 24 enam. wire on T50-2 toroid core.
L5 — 44 turns No. 28 enam. wire on T50-2 toroid core.
L6 — 6-turn link over L5, No. 28 enam. wire.
S1 — Dpdt slide switch (shown in receive mode).

series combination of L1 and L2. With the component values shown in Fig. 6, the shift is −0.5 or −1 kHz. The shift will vary with different crystals: Experimentation may be required.

The receiver is similar to others described in previous chapters. An RCA CA3028A serves as a product detector. The output is transformer coupled to a two-stage audio amplifier which utilizes a pair of bipolar transistors. In the interest of simplicity, no audio-gain control was included. The only selectivity in the receiver is that which is provided by the low-pass characteristic of the audio amplifier and the limited bandwidth of T1.

The transmitter portion of the circuit consists of the crystal oscillator (Q1) and the keyed power amplifier (Q2). Keying is by means of a microswitch in series with the supply to the collector. The microswitch is activated by a strip of pc board which serves as a paddle. The details may be seen in the photographs. Keying is clean, although with this method the backwave is only suppressed by approximately 30 dB. Owing to the low power output of the transmitter (0.5 watt), the backwave presents no problem.

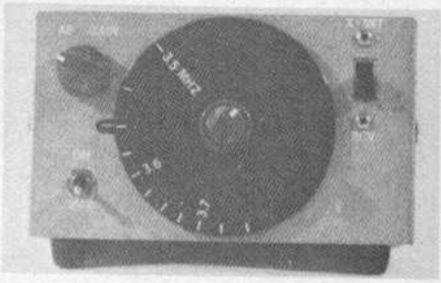
A General Electric D13-T type of programmable unijunction transistor (PUT) serves as a sidetone oscillator. The output is injected into the input of the two-stage audio amplifier.

Transmit-receive switching is realized with a double-pole, double-throw toggle switch, S1. One section switches the antenna while the other controls the power-supply output. Receiver muting is done by removing the operating voltage from the detector during transmit periods.

A low-pass filter section (L6) is included at the antenna jack of the transceiver. This provides harmonic suppression at the transmitter output. Additionally, it adds preselection to the receiver front end. This was found to be helpful when the transceiver was operated in close proximity to TV broadcast stations.

The station is built on a 2 X 5-inch double-sided pc board. The side containing the components is the ground foil, with the interconnecting runs on the back of the board. The box size is 1-1/2 X 3 X 5 inches. Locations of the components may be seen in the photographs. Placing all of the controls on one side of the chassis permits convenient operation. The transceiver is normally held in the left hand, with the right hand activating the controls and key. The battery pack is composed of AA-size NiCads, and usually resides in a parka pocket.

This transceiver has been used for several years, predominantly on backpacking and mountain-climbing trips. While never needed, it has been available for emergency communications on trips of a more committed nature. A deficiency of the design, as presented here, is the need for plug-in crystals. Not only are loose crystals lost easily, but the pins are subject to corrosion. The next version of this transceiver will contain switched crystals. VFO operation has not been considered because of environmental extremes that are encountered during use. No pc information is available for this project.



Front panel view of the 80-meter direct-conversion cw transceiver.

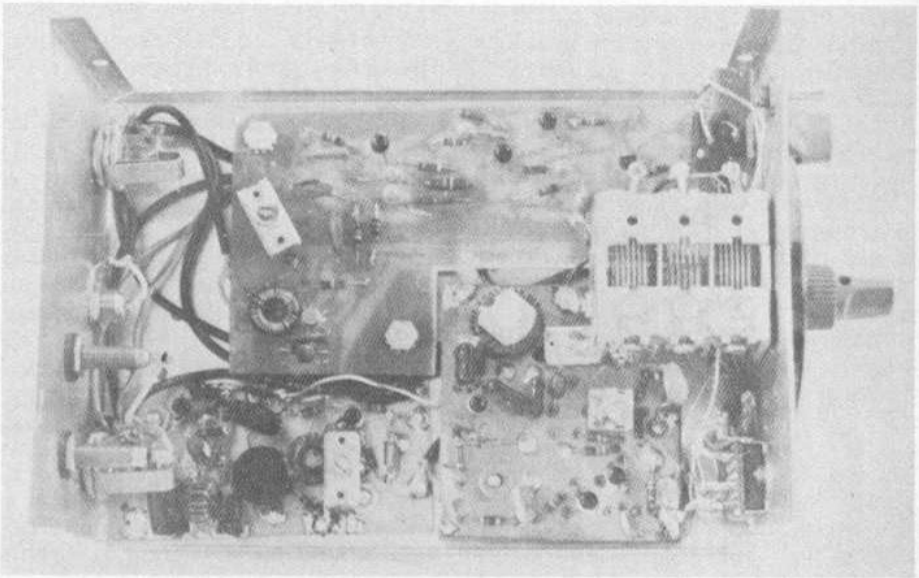
Direct-Conversion VFO Transceivers for 40 and 80 Meters

For general purpose portable operation, or for "sport" QRP work from the home station, a direct-conversion transceiver is ideal. Construction is simplified if a single-band design is used. This section describes two VFO-controlled "dc" transceivers. The 80-meter unit was built by KL7IAK. The 40-meter transceiver was constructed by one of the writers. Both transceivers have a transmitter output of approximately 1.5-watt, and they are physically compact.

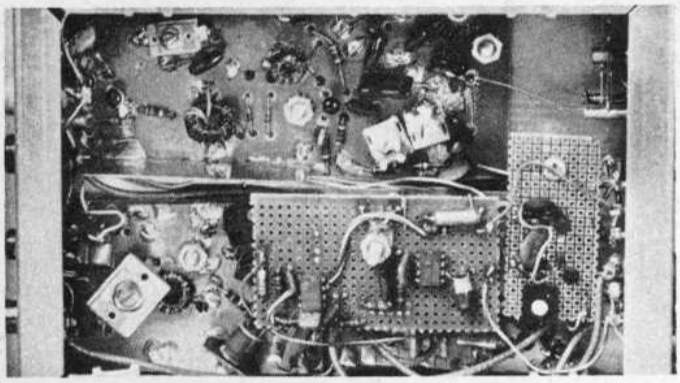
Shown in Fig. 7 is the 80-meter transceiver. A VFO (Q1) operates directly in the 80-meter band and is buffered with a two-stage feedback pair of bipolar transistors. The output of the buffer is applied to the transmitter and receiver simultaneously. The VFO is tuned with two sections of a capacitor that was scavenged from an old broadcast receiver. The capacitor had a built-in vernier-drive mechanism, simplifying the physical construction. The total capacitance required was approximately 200 pF. The VFO is built on a double-sided circuit board.

The transmitter board consists of Q4, a keyed driver, and Q5, the output amplifier. This circuit is virtually identical to the universal QRP transmitter described in an earlier chapter. One of the boards from that layout could be adapted for this transmitter if desired. The output amplifier uses a 2N5321 with a small heat sink. A large number of transistors could be substituted for this part if desired. The GE D44C6 used in a number of earlier transmitters could provide an output power of several watts. Different network constants at L4 would be required. (Early chapters should be consulted.)

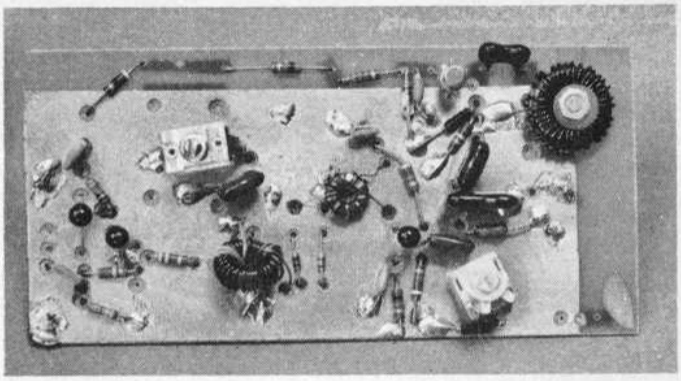
The receiver was adapted from the "TERAC Mountaineer." This was a transmitter-receiver combination that was built as a club project by the Tektronix Employees' Radio Amateur Club and was originally described in *QST* for August, 1972. The original version was for 40 meters, but was adapted for 80 by KL7IAK. These boards are no longer available although pc information may still be obtained in accordance with the reference in the original paper.



Interior of the 80-meter transceiver. At the lower right is the VFO. The receiver can be seen on the L-shaped board. At the lower left is the transmitter output circuit.



Interior of the 40-meter unit. The upper board contains the VFO, frequency doubler and 7-MHz buffer amplifier. The transmitter pc board is at the lower left. At the lower right can be seen a small board on which the sidetone oscillator is built. Seen at the lower center is the *RC* active filter assembly. The main receiver board is buried beneath the audio filter.



Close-up view of the 40-meter transceiver VFO. The 80-meter oscillator is at the right. The offset circuit was incomplete when this photograph was made. At the center is the diode frequency multiplier, with the output buffer at the left end of the board.

The basis of the receiver is a product detector using a dual-gate MOSFET, Q6. This is followed by a three-stage audio amplifier. More than ample audio is available to drive 2000-ohm headphones.

A sidetone oscillator is included. This circuit (Q12 and Q13) uses a pair of transistors to synthesize the action of a programmable unijunction transistor. A GE type D13-T PUT could be substituted directly. Q11 provides a low-impedance drive for the headphones from the sidetone oscillator. Detected rf is used to activate the sidetone, offering a built-in indication of rf output. The pitch of the audio note will depend upon the output level. Transistor Q10 is included to mute the receiver during transmit periods.

Transmit-receive control is achieved with S1, a double-pole, double-throw slide switch. The transceiver is built in a 3 X 5 X 7-inch box. Parts placement can be seen in the photographs. Shielding is not necessary.

The VFO circuit includes an offset capacitor, C2. This is switched in during transmit periods to place the outgoing signal at approximate zero beat with the station being contacted. It is necessary that the receiver be tuned on the high-frequency side of the other station.

The results obtained with this unit have been excellent. While most of the contacts have been with other Alaskans, the "lower 48" have also been worked from KL7IAK.

Shown in Fig. 8 is the circuit for the 40-meter transceiver. This unit is similar in design to the one for 80 meters. The VFO (Q1) is virtually identical. It operates at 3.5 MHz. The main-tuning capacitor, C1, has a range of approximately 10 pF. The series capacitor that is used with it provides a tuning range of 50 kHz on the 7-MHz band. While a larger range would be desirable, the restricted one has the advantage that no vernier drive is required. Frequency calibration is not included in the transceiver. The frequency-offset method used in the KL7IAK transceiver is used in the VFO. However, in this model the diode is activated during receive periods. A toggle switch, S2, is included on the front panel to interrupt the diode bias current. When S2 is open, the transceiver may be tuned to zero beat with an arriving signal. S2 is then closed. This causes the VFO to decrease in frequency by the proper amount to produce an output tone of 800 Hz.

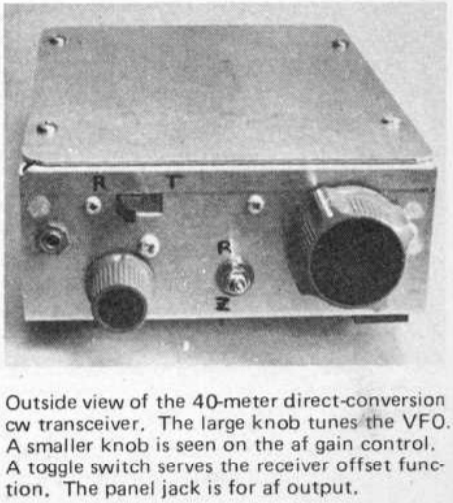
The output of the FET VFO is applied to a single-stage bipolar buffer amplifier. The buffer output drives a frequency doubler which uses a pair of silicon switching diodes. The resultant 7-MHz output is filtered with a single tuned circuit (L3) and then routed to a two-stage feedback amplifier (Q3 and Q4). This signal is applied to both the receiver detector and the transmitter board.

The transmitter is nearly identical in design to the 80-meter circuit used by KL7IAK. A 2N3904 keyed driver is followed by a 2N5321 power amplifier. The output power is slightly over 1.5 watts.

The receiver is conventional in design. It uses CA3028A product detector. Balance at the rf port of the detector is enhanced through the use of a bifilar link to drive the IC. The detector output is amplified with a two-stage audio amplifier, Q9 and Q10. The resultant signal is filtered with a four-pole *RC* active low-pass filter. U2 serves as an impedance-transforming element to ensure proper drive for the following stages. A dc level shift is also provided to properly establish the bias on the filter ICs. A low-pass filter with a 1-kHz cutoff frequency was chosen over a bandpass circuit. This allows a received signal to be tuned to zero beat with greater ease.

The sidetone oscillator in the 7-MHz transceiver is a free-running multivibrator. A three-pole, double-throw slide switch, S1, serves as the TR control. One set of contacts transfers the headphone jack between the receiver output and the sidetone oscillator. A board-mounted potentiometer is included on the sidetone-oscillator board for level adjustment.

At first glance the amount of circuitry used in this transceiver may seem excessive. Certainly, some simplification is possible, just as further refinement might be desired. The audio filter may be eliminated. However, the filter is so simple, and adds so much to the performance, that this is not suggested. The use of diodes as the multiplier might also be questioned. The total parts count is somewhat higher than might be realized with other circuits. However, no special equipment is required for adjustment. An oscilloscope is not needed to obtain balance to ensure rejection of the 3.5-MHz fundamental. Also, diodes do not oscillate! The output of this transmitter was studied with a spectrum analyzer. At 1.5-watts output, the 3.5-MHz fundamental compon-



Outside view of the 40-meter direct-conversion cw transceiver. The large knob tunes the VFO. A smaller knob is seen on the af gain control. A toggle switch serves the receiver offset function. The panel jack is for af output.

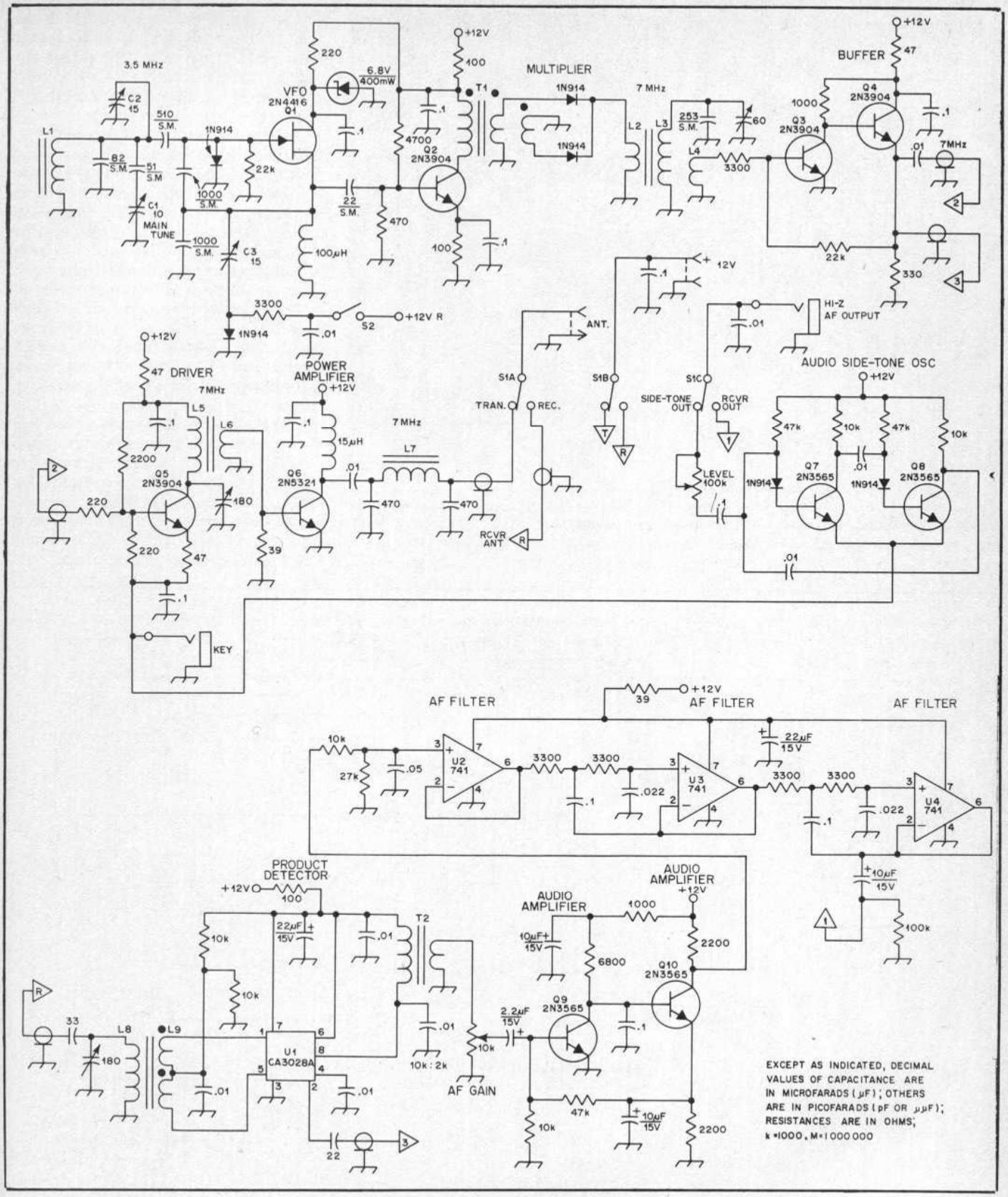
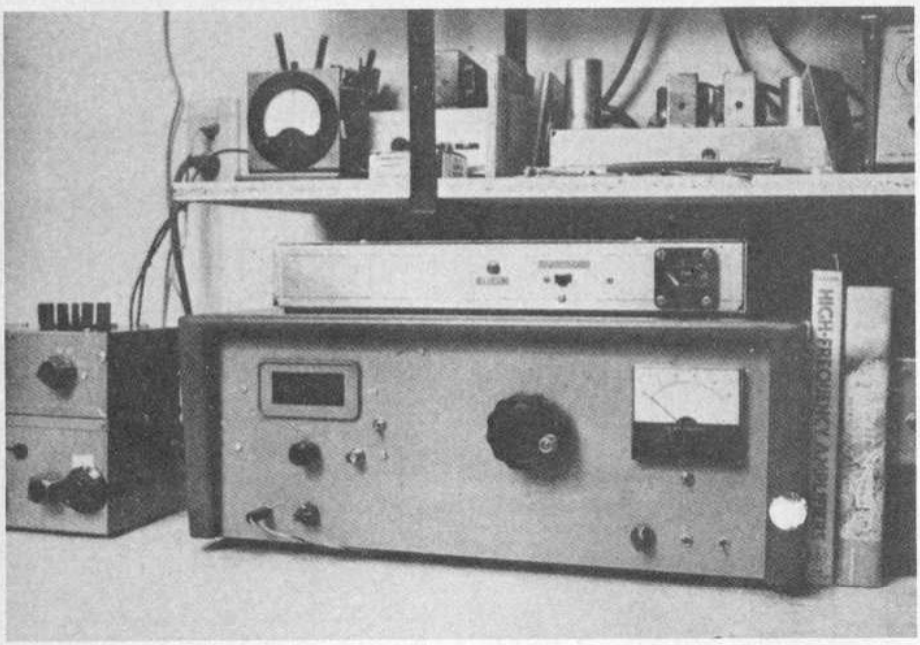


Fig. 8 — Schematic diagram of the 7-MHz direct-conversion cw transceiver. Fixed-value capacitors are disk ceramic unless otherwise indicated. Polarized capacitors are electrolytic. Variable capacitors without numbers are mica compression trimmers. Fixed-value resistors are 1/4- or 1/2-W composition.

C1 — 10-pF air variable, panel mounted.
C2, C3 — 15-pF pc-board-mount air variable.
L1 — 31 turns No. 22 enam. wire on T68-2 toroid core.
L2 — 5-turn link over L3, No. 24 enam. wire.
L3 — 18 turns No. 24 enam. wire on T50-2 toroid core.
L4 — 3-turn link over L3, No. 24 enam. wire.
L5 — 35 turns No. 26 enam. wire on T50-2 toroid core.
L6 — 4-turn link over L5, No. 26 enam. wire.
L7 — 14 turns No. 22 enam. wire on T50-2 toroid core.
L8 — 30 turns No. 28 enam. wire on T50-2 toroid core.
L9 — 5-turn bifilar winding of No. 28 enam. wire over L8.
S1 — 3-pole, double-throw slide switch.
S2 — Spst toggle.
T1 — 10 trifilar turns No. 32 enam. wire on Amidon FT37-61 ferrite toroid core.
T2 — 10,000-ohm pri., 2000-ohm sec., miniature audio trans.
VR1 — 6.8-V, 400-mW Zener diode.



Exterior of the contest-grade cw station. The large unit is the receiver. Atop the receiver is the exciter, and at the left is the PA module. The latter is affixed to a Transmatch (lower box). Station power supplies are on the shelf above the main equipment.

ent was 52 dB down, and the backwave was 76 dB below the key-down output. This performance would be difficult to achieve with unbalanced circuitry unless much more selective filters were used at 7 MHz.

The transceiver is built in a 2 X 5 X 7-inch chassis, which serves as the cabinet. Shielding is not necessary between sections. The general placement of components may be seen in the photographs. The main part of the receiver (U1, Q9, and Q10) is on a 2 X 2-inch board that is buried below the active filter. Although not shown in the schematic, two key jacks are provided. One is for a hand key. The other is a stereo-type of headphone jack. The extra output has +12 volts applied to it. This provides power for a portable electronic keyer.

This transceiver has been used for a two-year period during portable applications. While most of the service has been casual (family picnics and Field Day), the transceiver has also been carried to some of the high elevations in the

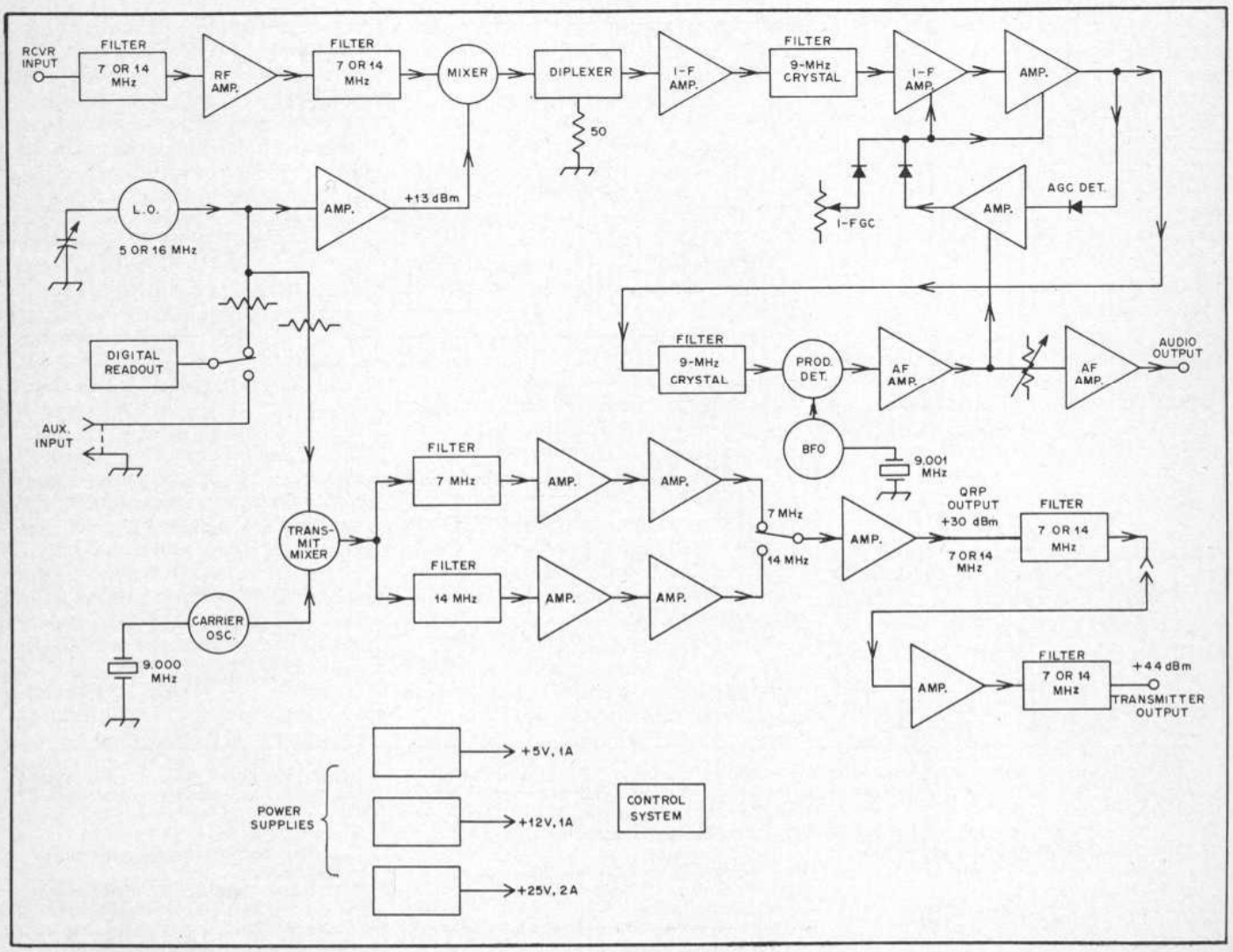


Fig. 9 — Block diagram of the integrated cw station for 7 and 14 MHz.

Pacific Northwest. The compact format makes this realized easily.

An Integrated Contest-Grade CW Station

Most of the equipment described in this book has been comparatively simple. One- or two-band designs have been more prevalent than multiband systems. Equipment has, more often than not, been designed with ease of duplication as a major objective. There is good reason for this: Our motivation is to encourage the amateur to construct his own equipment. This is more easily realized if extremes of complexity are avoided. The writers have followed these guidelines for their own equipment in many cases.

During all of the experimentation and design work required for the simpler projects, there has always been the question, "What would happen if all of the constraints were lifted? What level of equipment performance can the amateur experimenter expect to achieve without the aid of sophisticated instrumentation?" The station described in this section is aimed at providing one answer to those queries.

The station is an outgrowth of a receiver that was described initially in *QST* for March and April, 1974. Since that time several refinements have been incorporated to provide improved performance. A transmitter has been built to operate in a full transceive mode with the receiver. The power output is 1 watt for QRP work or 25 watts for DXing and contesting. The performance of the system is excellent, and appears to equal or exceed that of commercially available equipment with which we are familiar.

Some semblance of simplicity is retained in this station by confining the operation to cw and to only two bands, 7 and 14 MHz. No other constraints are imposed other than that of low power, which is a matter of personal choice. Owing to the relative complexity of this station, it is not recommended as a construction project except for the amateur with *considerable* experience. No pc information is available. However, every effort has been made to include all pertinent circuit information.

A project such as this serves a multiple purpose. First, it provides high-quality equipment for communications. Of greater significance, the gear functions as an experimental vehicle – a means of trying new ideas as they occur. As such, this station is in a constant state of change. To enhance this flexibility, no attempt has been made toward miniaturization.

System Details – The Receiver

A block diagram of the total station is shown in Fig. 9. The receiver is a single-conversion design with a 9-MHz i-f. The local oscillator, which is at either 5 or 16 MHz for 20- and 40-meter operation, respectively, is the only circuit that is shared with the transmitter except for the power supplies. A digital display is employed to read the LO frequency. Because the i-f is exactly at 9.000 MHz, no special programming is needed for the counter, allowing its use for general-purpose test applications.

Shown in Fig. 10 is the receiver preselector function. Four poles of filtering are used on each band with an rf amplifier embedded within the filter. A JFET is used for 40 meters while a dual-gate MOSFET is employed at 14 MHz. There was no special justification for this choice since both are capable of low noise figure and high output intercept. The dual-gate MOSFET is probably the better choice since it tends to be more stable in the common-source configuration. Relays are used for band switching. This has the advantage of placing the switches where they are

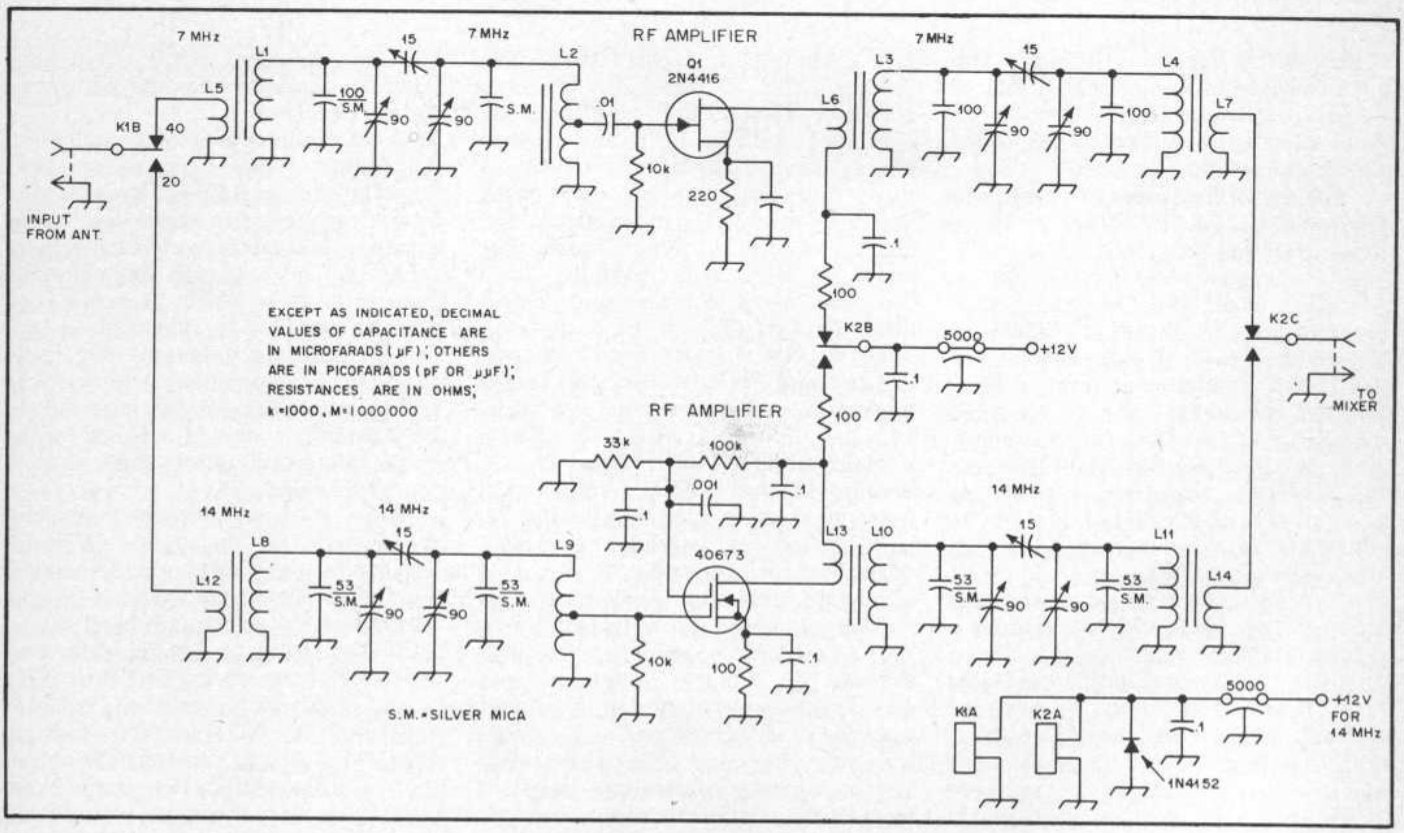


Fig. 10 – Schematic diagram of 7- and 14-MHz receiver preselectors. Fixed-value capacitors are disk ceramic unless noted differently. Variable capacitors are mica compression trimmers. Fixed-value resistors are 1/4- or 1/2-W composition.

K1, K2 – Double-pole, double-throw 12-V dc relay with 800-ohm field coil.

L1-L4, incl. – 30 turns No. 27 enam. wire on T37-6 toroid core. L2 tapped at 15 turns.

L5 – 4-turn link No. 27 enam. over L1.

L6 – 6-turn link No. 27 enam. over L3.

L7 – 4-turn link No. 27 enam. wire over L4.

L8-L11, incl. – 18 turns No. 24 enam. wire on T37-6 toroid core. L9 tapped 8 turns from ground.

L12 – 3-turn link No. 24 enam. over L8.

L13 – 4-turn link No. 24 enam. over L10.

L14 – 3-turn link No. 24 enam. over L11.

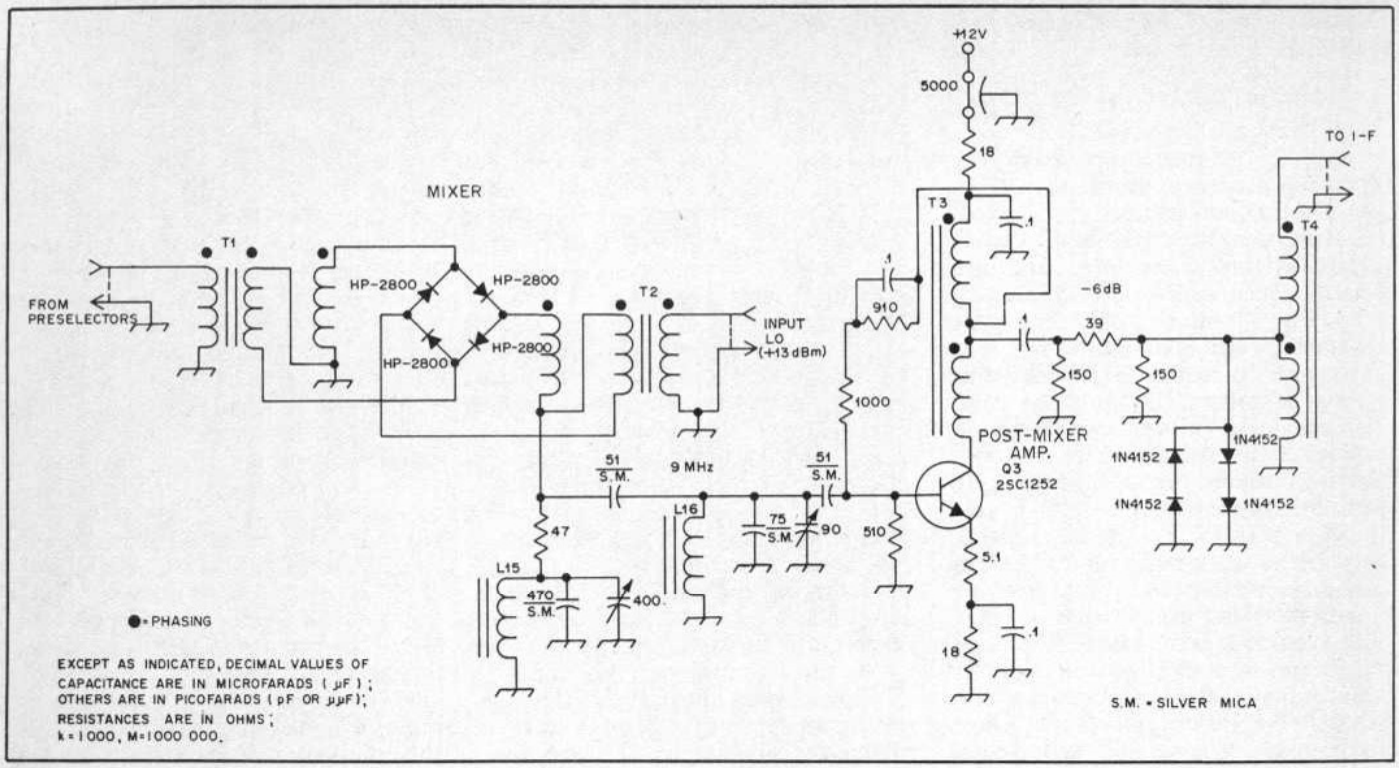


Fig. 11 — Schematic diagram of the front-end mixer and post-mixer i-f amplifier. Fixed-value capacitors are disk ceramic unless noted differently. Variable capacitors are mica compression trimmers. Resistors are 1/4- or 1/2-W composition.

CR1-CR4, incl. — Hewlett-Packard hot-carrier diode or equiv.

L15 — 12 turns No. 24 enam. wire on T37-6 toroid core.

L16 — 17 turns No. 22 enam. wire on T68-6 toroid core.

Q3 — Nippon Electric 2SC1252 (California Eastern Labs., Inc., One Edwards Ct., Burlingame, CA 94010).

T1, T2 — 10 trifilar turns No. 30 enam. wire on FT37-61 ferrite toroid core.

T3, T4 — 10 bifilar turns No. 30 enam. wire on F37-61 ferrite toroid core.

needed within the circuitry, while adding no mechanical complexity. Shielding integrity may be easily maintained. Band selection is realized by means of a panel-mounted toggle switch.

The use of four poles of preselection is quite worthwhile. The measured image rejection on both bands is 95 dB. Similar numbers were obtained for the i-f feedthrough. The preselectors were adjusted for a bandwidth of 100 kHz on 40 meters, with a slightly wider one at 20 meters. Careful adjustment is necessary to ensure that the double-tuned circuits are not over-coupled. The amplifiers are biased for high gain. However, by purposeful impedance mismatching the net gain of this section is set at 10 dB. This is mandatory to maintain reasonable gain distribution.

Although detailed measurements have not been made with this module, it is possible that some intermodulation distortion is occurring within the toroid cores used in the filters. It might be desirable to replace the T37-6 cores with the larger T68-6 units. Suitable circuits are presented in the appendix tables. An alternative approach would be to eliminate the rf amplifiers completely, using only passive preselector filters. Such networks were used in a family of crystal-controlled converters described in chapter 6. The preselectors here, with their rf amplifiers, are housed in a separate box, shielded from other circuits. This is mandatory if filter stopband rejection is to be maintained.

Shown in Fig. 11 is the receiver mixer and the associated circuitry. A ring of hot-carrier diodes is used, owing to its relatively low noise figure and high intercept of this type of mixer. The mixer is terminated carefully on a broadband basis at the i-f port. This is done through the use of a diplexer. These circuits were described in chapter 6. L16 and the related capacitors form a single-pole bandpass circuit at 9 MHz. L15 and the capacitors associated with it form another 9-MHz tuned circuit. Because of the parallel resonance, at frequencies other than 9 MHz the 47-ohm resistor is attached to ground through a low reactance, serving as a termination for out-of-passband energy.

The diplexer is followed by a "strong" 9-MHz i-f amplifier. Through the use of feedback (both shunt and series) the input impedance of this amplifier is very close to 50 ohms over a wide frequency range. The 6-dB attenuator at the output helps to ensure that impedance variations resulting from the following crystal filter do not reflect back through the amplifier to alter the input immittance. Even with the 6-dB attenuator, the gain of this amplifier is 17 dB. The amplifier noise figure is 6 dB and the output intercept is +35 dBm. The high intercept results from good transistor characteristics and a high bias current of 65 mA.

A set of silicon switching diodes is at the output of the i-f post-mixer amplifier. They protect the following crystal filter from excessive signal levels. The amplifier has an output capability of about 250 mW, enough to potentially damage the filter. While the receiver will never (hopefully) be subjected to such signals from the antenna, they could result during experiments with break-in keying or from an antenna-relay failure. T4 provides a source impedance for the crystal filter (200 ohms) which is close to that specified.

Fig. 12 shows the local-oscillator system used for the station. A three-terminal regulator (U1) provides a stable 5 volts for the oscillators. Two separate LOs are used, one for each band, with a relay for band switching. Motorola MC1648Ps serve as the oscillators. The output power of these circuits is low — only about 1 mW. A broadband amplifier (Q4) is used to boost the LO output to +13 dBm. Attenuated outputs are provided to drive the digital readout and the transmit mixer.

The stability of these oscillators is more than sufficient. Temperature compensation was required in the 16-MHz oscillator used for 40 meters. This was accomplished experimentally by repeat-

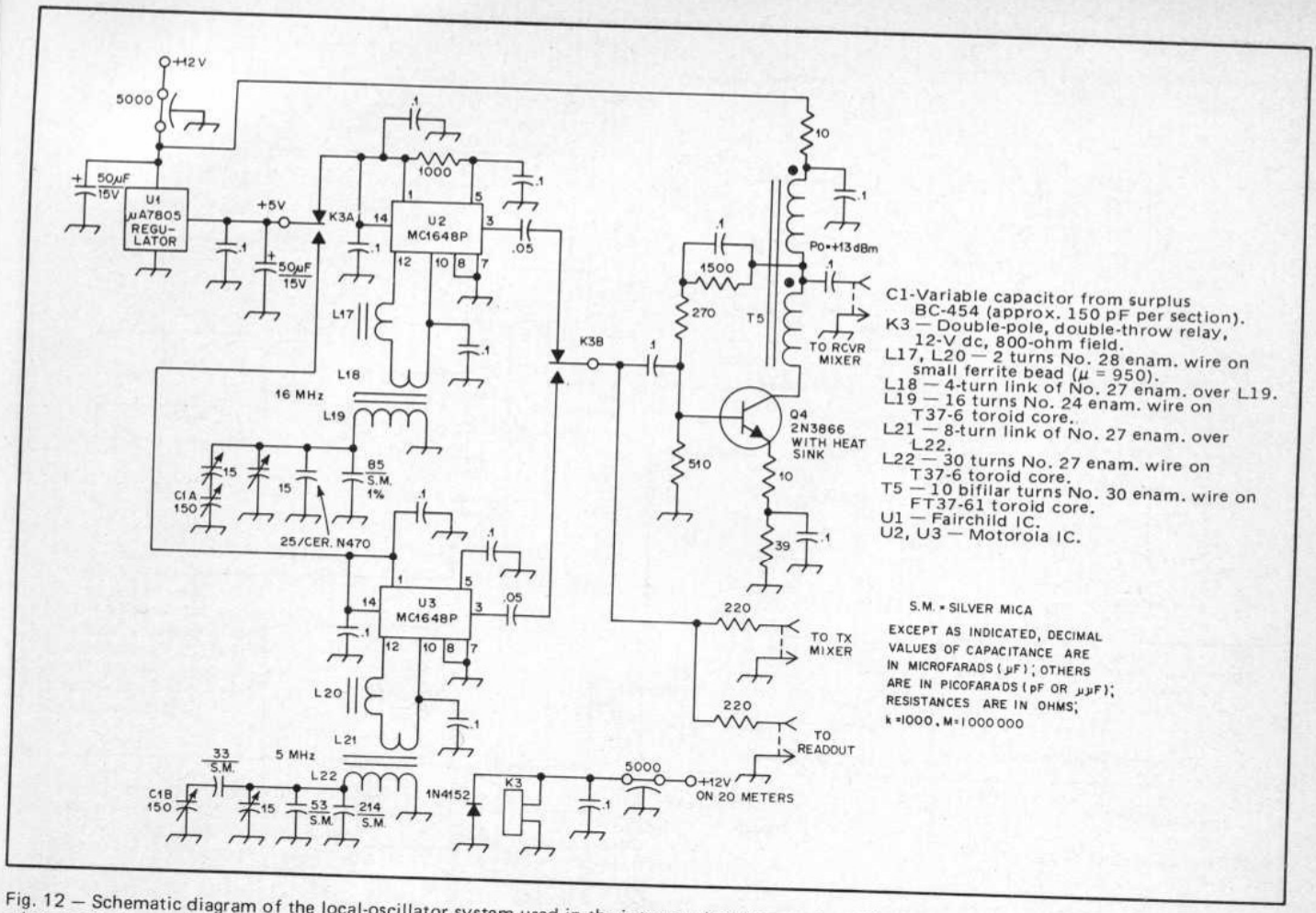


Fig. 12 — Schematic diagram of the local-oscillator system used in the integrated station receiver. Fixed-value capacitors are disk ceramic unless otherwise indicated. Variable capacitors without parts numbers are air-dielectric pc-board-mount trimmers. Fixed-value resistors are 1/4- or 1/2-W composition.

edly opening the window to the shop, and through application of heat from a desk lamp. One might question the advisability of using a free-running oscillator at a frequency as high as 16 MHz. However, the low-noise characteristics, and the lack of other output components that might lead to spurious responses, is ample justification for the minor job of temperature compensation.

The oscillator is tuned with C1, a three-section capacitor from a surplus receiver. The built-in gear-reduction was found to be superior to commercially available drive mechanisms. The large reduction ratio employed leads to a tuning rate of less than 10 kHz per revolution of the knob. The high selectivity justifies this.

The circuitry used in the original version of this receiver did not include the amplifier in the LO chain. Also, the mixer was poorly terminated. The result was good sensitivity but a dynamic range of only 85 dB. The increase in LO drive power and improved mixer termination, along with the addition of a better post-mixer amplifier, increased the dynamic range to 95.5 dB. The noise figure of the receiver was virtually unchanged at 7 dB. The MDS was −141 dBm. The measurements were performed with high-quality laboratory instrumentation. Noise-figure measurements correlated well with direct MDS measurements. Further study is required to determine the factors that are presently limiting the dynamic range. Effects like nonlinearity in transformers and coils, and IMD due to crystal filters, could be of significance.

The i-f amplifier and agc system are shown in Fig. 13. Most of the selectivity of the receiver is provided by FL1. The filter used in this receiver has ten crystals and a 3-dB bandwidth of 500 Hz. The shape of the filter was Gaussian near the peak of the response, a characteristic that provides improved transient response. This filter (KVG-XL-10M) is no longer available. However, KVG has recently introduced a similar unit, the XF-9NB. This filter should be an excellent substitute. The termination resistance for the XF-9NB is 500 ohms instead of the lower values used for the XL-10M. Circuit changes will be required. A pi-network matching scheme would be ideal at the input, while output termination can be realized by replacing the present 300-ohm resistance. KVG filters are available from Spectrum International, Box 1084, Concord, MA 01742.

The gain in the i-f strip is provided by a pair of MC1590G ICs. While sufficient gain could be realized with one stage, a larger agc range is available with two. In the circuit shown, the gain is approximately 65 dB. Over 120 dB of gain variation are achieved, however. The output of U5 is applied to a FET, Q5. This unit buffers the output from the agc takeoff point.

The agc system is one that was described in detail in chapter 5. A full hang action is employed. The main memory capacitor, C2, should be a low-leakage type such as a disc ceramic. The agc is defeated by S1 while S2 changes the decay-time constant. Both switches are toggle types. R1 at the inverting input of U6 should be adjusted for +5 volts at pin 6 of U6 with the agc off. The FETs used in the agc are not commonly available. A modified circuit that is compatible with more common FET types was shown in chapter 5.

The product detector and BFO are shown in Fig. 14. The input to the product detector is filtered with a four-pole crystal filter, FL2. This restricts

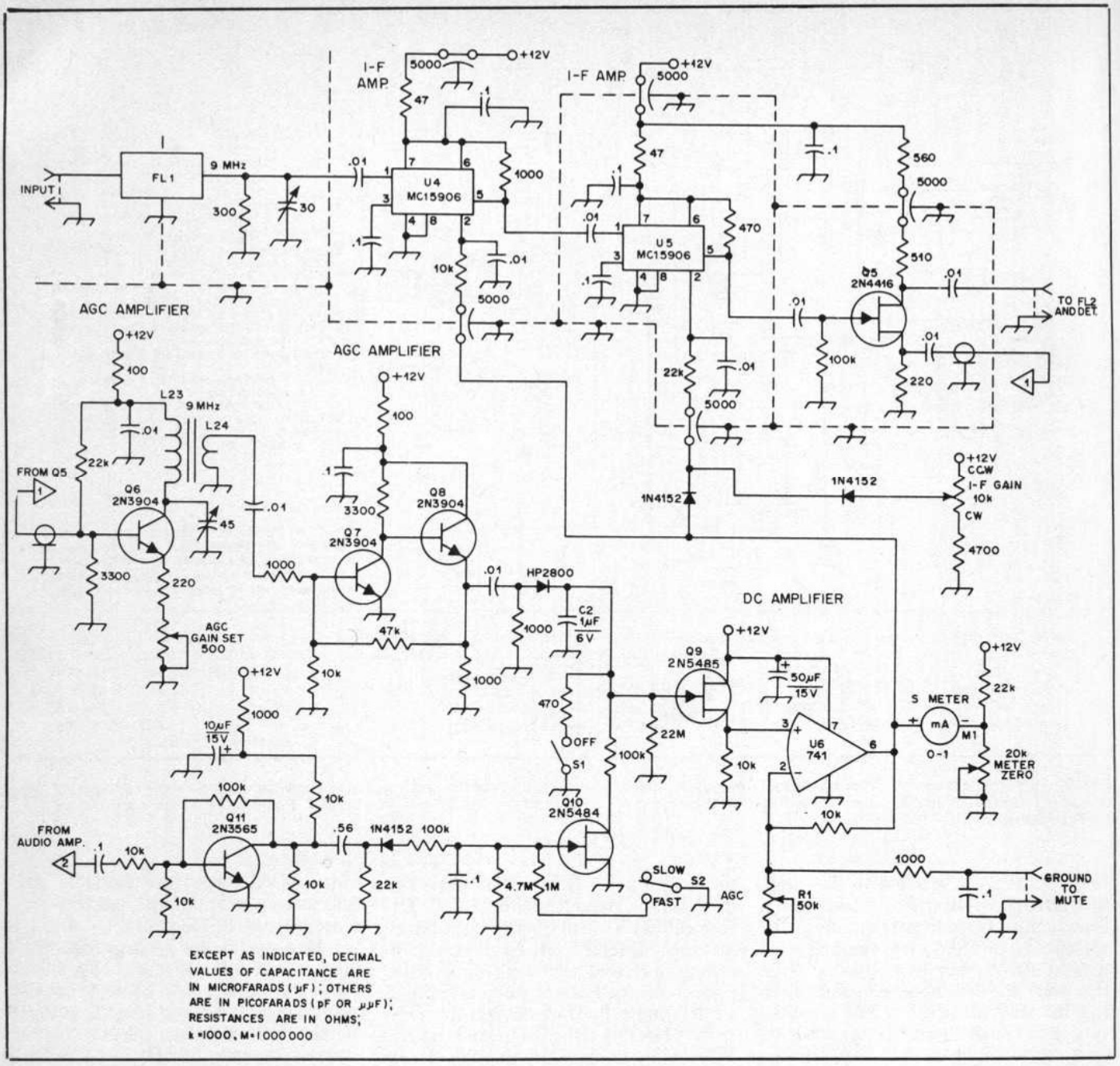


Fig. 13 — Schematic diagram of the receiver i-f amplifier and agc system. Dashed lines indicate shielding, which is extensive throughout the receiver. Fixed-value capacitors are disk ceramic except those with polarity marked, which are electrolytic. Fixed-value resistors are 1/4- or 1/2-W composition. Variable resistors other than R1 are pc-board-mount controls. Variable capacitors are mica compression trimmers.

FL1 — 9-MHz 10-pole crystal filter, Spectrum International type KVG (see text).
L23 — 50 turns No. 34 enam. wire on T37-6 toroid core.
L24 — 10 turns No. 34 enam. over L23.
M1 — 1-mA meter of builder's choice.
R1 — 50,000-ohm linear-taper control, panel mounted.
S1, S2 — Spst toggle.
U4, U5 — Motorola IC.

the noise bandwidth of the i-f energy reaching the detector. FL2 should be matched to FL1 in frequency. A pi network is used to match the 500-ohm output impedance of FL2 to the 50-ohm input of the detector. A diode ring serves as the product detector. Originally an MC1496 was used. However, this led to IMD and excessive noise. The performance of the diode ring is much better.

The BFO uses a JFET, Q12. The output is filtered with a single-section low-pass filter. This ensures that the waveform for the detector is symmetrical, a requirement for best balance. The power available to the detector is +12 dBM, which is enough to provide good IMD performance from the diode ring.

The audio system for the receiver is presented in Fig. 15. An LM301A is used as an audio preamplifier. Owing to the high closed-loop gain of this circuit, a noisier device (741) should not be substituted here. A 50-kΩ linear potentiometer serves as the audio gain control. An audio-taper unit was not available, but was simulated by loading the arm of the control with a 4700-ohm resistor.

The output amplifier operates in Class A. While this has the liability of consuming considerable current, the fidelity is excellent. The maximum output power is under 100 mW, but is enough to drive a small monitor speaker or headphones. A sidetone oscillator (Q14) is included. This circuit is activated with a +12-volt source that is derived from the station keyer.

The performance of this receiver is

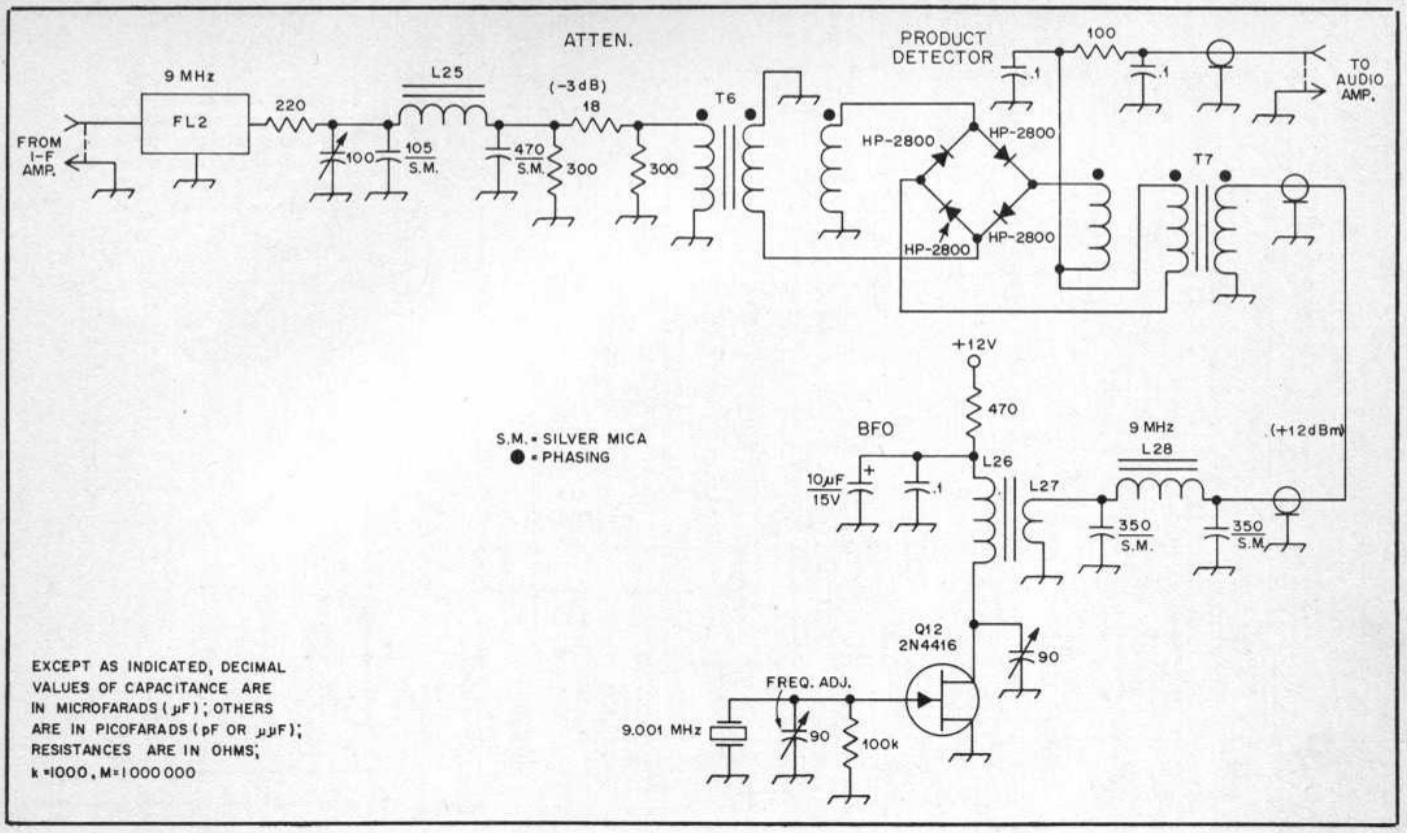


Fig. 14 — Schematic diagram of the receiver product detector and BFO. Fixed-value capacitors are disk ceramic unless noted differently. Variable capacitors are mica compression trimmers. Fixed-value resistors are 1/4- or 1/2-W composition.

FL2 — Four-pole, 9-MHz crystal filter (Spectrum International type KVG). See text.

L25 — 26 turns No. 26 enam. wire on T37-6 toroid core.

L26 — 40 turns No. 26 enam. wire on T50-2 toroid core.

L27 — 6-turn link of No. 26 enam. over L26.

L28 — 17 turns No. 26 enam. wire on T37-6 toroid core.

T6, T7 — 10 trifilar turns of No. 30 enam. wire on FT37-61 toroid core.

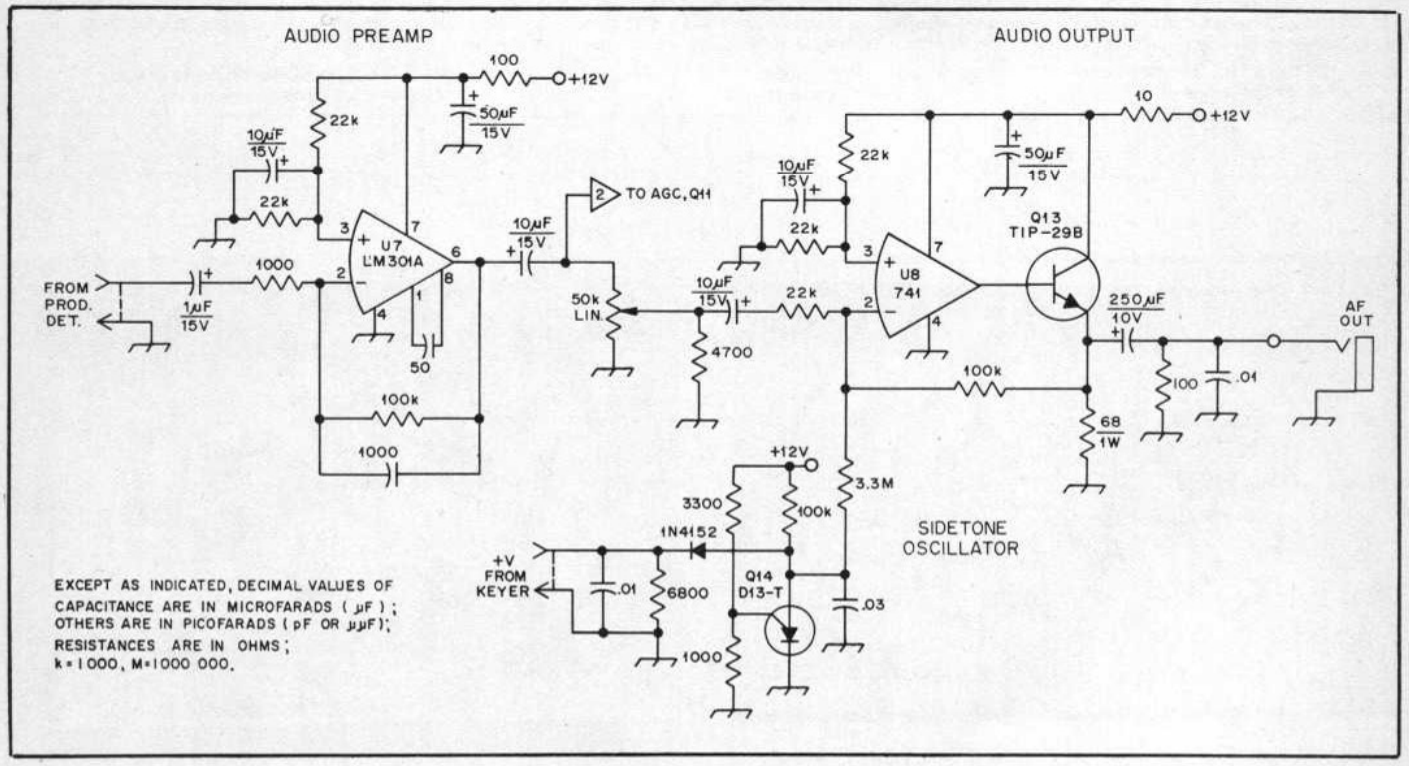


Fig. 15 — Schematic diagram of the audio system and side-tone oscillator for the integrated station. Capacitors are disk ceramic except those with polarity marked, which are electrolytic. Fixed-value resistors are 1/4- or 1/2-W composition unless otherwise noted. R2 is a linear-taper composition control, panel mounted.

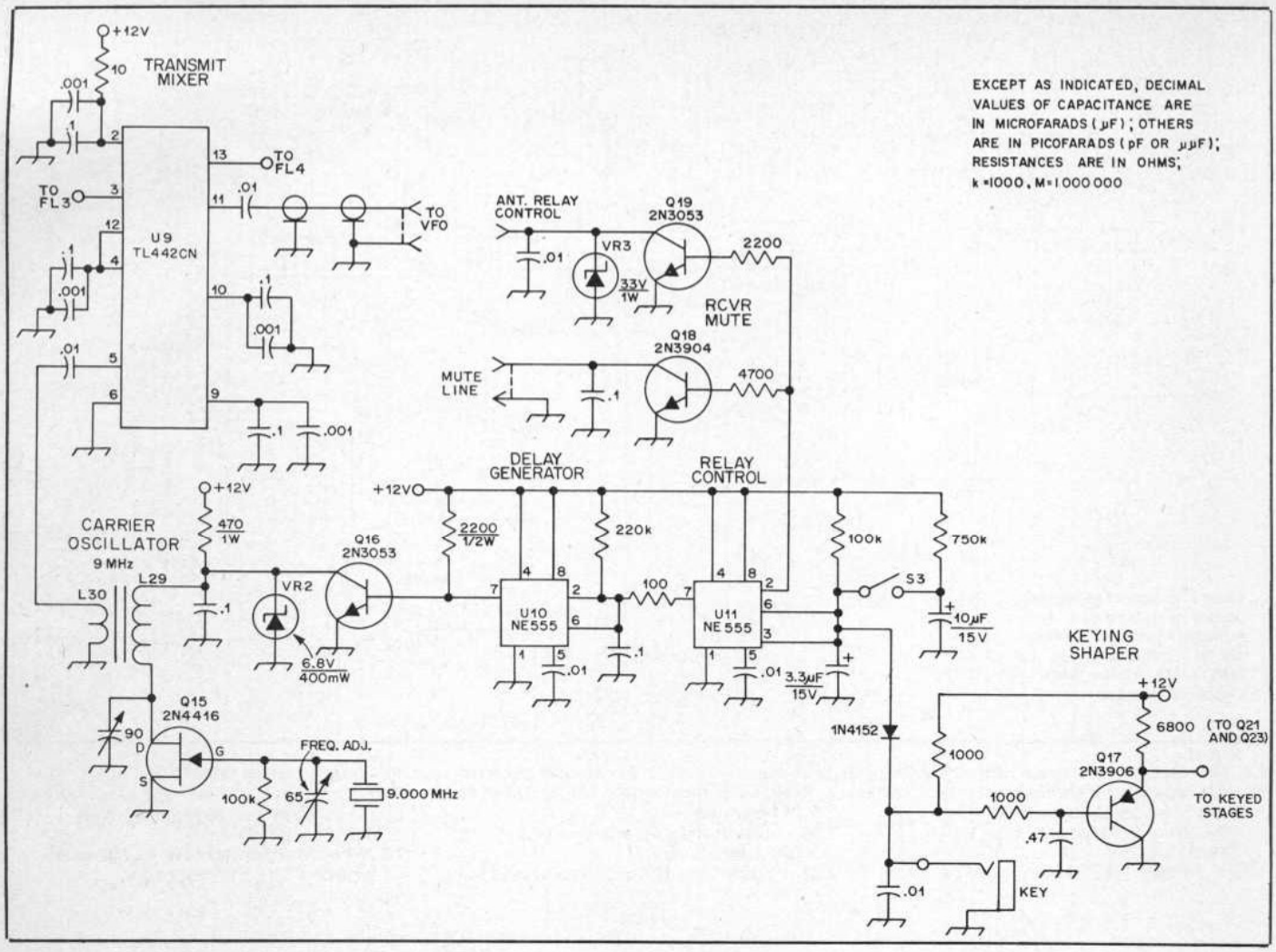
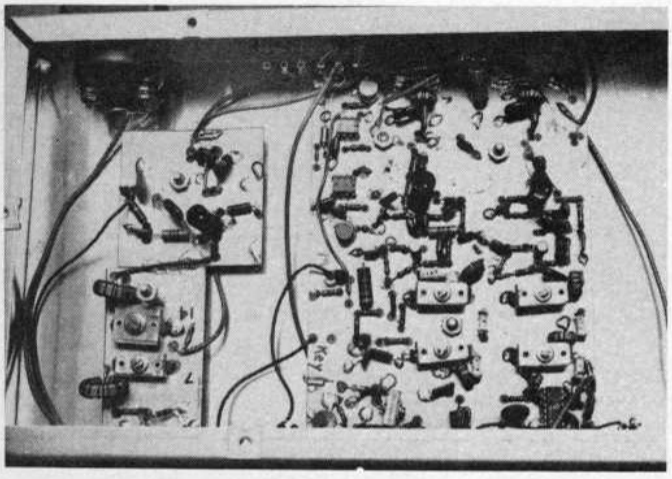
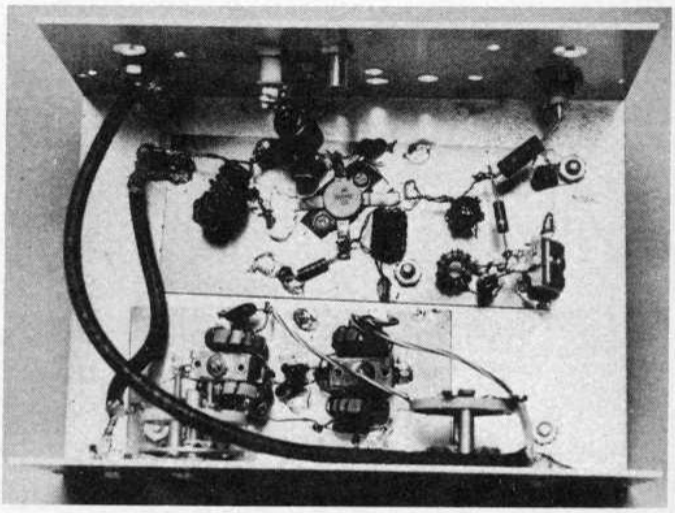


Fig. 16—Schematic diagram of the carrier oscillator, transmit mixer and control circuits. Fixed-value capacitors are disk ceramic, Mylar, or monolithic chip types. Variable capacitors are mica compression trimmers. Fixed-value resistors are 1/4-or 1/2-W composition. Polarized capacitors are electrolytic. The SN-76514 mixer IC has been reidentified as TL-442-CN by Texas Instruments. It may be procured under either part number.

L29 — 38 turns No. 28 enam. wire on T37-6 toroid core.
L30 — 2-turn link No. 28 enam. over L29.
S3 — Spst toggle.
U9 — Texas Instrument IC.
U10, U11 — NE555 timer ICs.
VR2 — 6.8-V, 400-mW Zener diode.
VR3 — 33-V, 1-W Zener diode.



Interior look at the exciter for the integrated cw station. The main board in the center contains the transmit mixer, 7- and 14-MHz bandpass filters, and the individual amplifier chains. Control and key-shaping circuits are on the same board. At the upper left is the 1-W PA. The PA output network is seen at the lower left.



Inside look at the 25-W PA. The switches control the networks in the amplifier output. There are separate networks for 7 and 14 MHz.

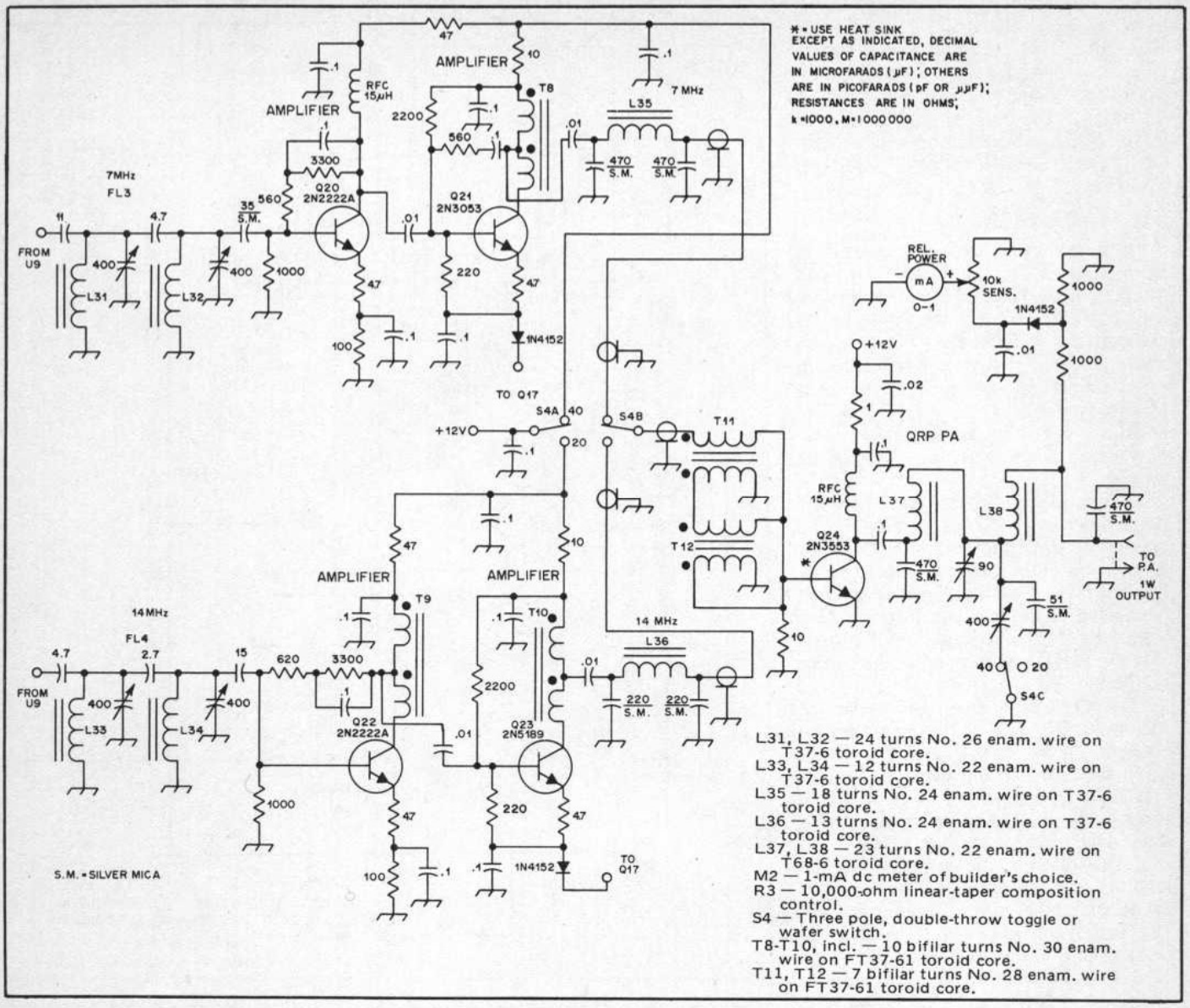


Fig. 17 — Schematic diagram of the 7- and 14-MHz exciter circuit. Included also is the 1-W output PA stage. Fixed-value capacitors are disk ceramic unless noted differently. Variable capacitors are mica compression trimmers. Resistors of fixed value are 1/4- or 1/2-W composition.

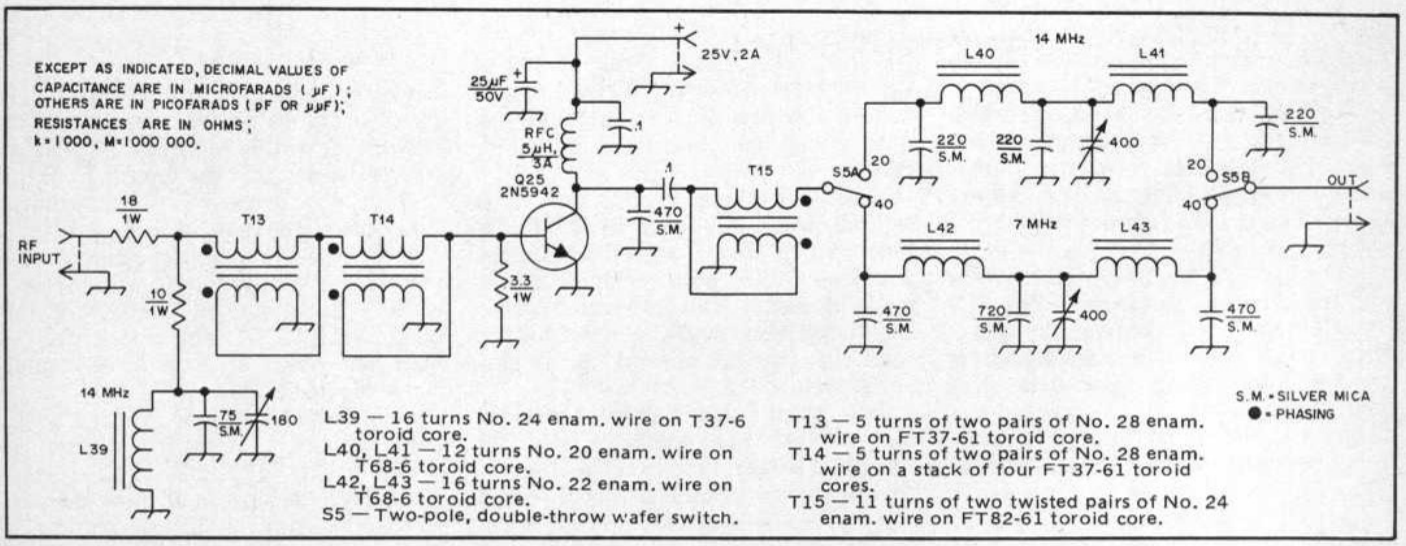


Fig. 18 — Schematic diagram of the 25-W rf amplifier. Fixed-value capacitors are disk ceramic unless otherwise noted. Capacitors with polarity marked are electrolytic. Variable capacitors are mica compression trimmers.

exceptional. It has been in use for over three years. While sophisticated instrumentation has been used for evaluation, none was available during construction or testing. The builder should have at least one signal generator and a high-frequency oscilloscope available, though.

The Transmitter Circuit

A heterodyne approach is used in the design of the transmitter. This was done to allow full transceive operation, a desirable feature for the lower power contest station. Separate VFOs could be added to the existing equipment in order to make it more effective during DXing. Room is available in the exciter enclosure for this. An alternative solution would be incremental tuning of the receiver.

Shown in Fig. 16 are circuits for the transmit mixer, carrier oscillator and system-control functions. An SN76514 is used as the transmit mixer, with filters for each of the bands of interest connected to the two output ports. This simplifies the band switching considerably. A JFET is employed as the carrier oscillator.

The control system uses a pair of 555 timer ICs. U11 is the main control element. When the key is depressed, U11 switches on and remains in that condition while the key is closed. U10 provides a short delay before activating the carrier oscillator. This ensures that there has been sufficient time for the antenna relay to operate. When the key is opened, the state of the control system remains constant until U11 has "timed out." At that instant U11 and U10 revert to their initial condition. With the delay switch, S3, open, the hold-in time is about 0.5 second. Closing S3 extends this to more than 1 second, a more desirable period for casual operation. Shaped keying is provided with Q17. It operates as an emitter follower. Chapter 7 should be consulted for more information on control systems.

The antenna relay often used is a 24-volt-dc coaxial type requiring 100 mA of coil current. However, other relays are sometimes used. A 33-volt Zener diode is used to protect Q19 from voltage transients.

The heart of the exciter is shown in Fig. 17. The two bandpass filters, FL3 and FL4, provide selection of the appropriate mixer output. Ceramic capacitors are used as coupling elements in these filters. Their values should not be changed casually. The appendix on filters provides additional information on filter design.

Separate two-stage amplifiers are provided for each band. Feedback is used to establish the desired gains and to provide proper termination impedances for the bandpass filters. A single-section low-pass filter is used at the output of the individual amplifier chains.

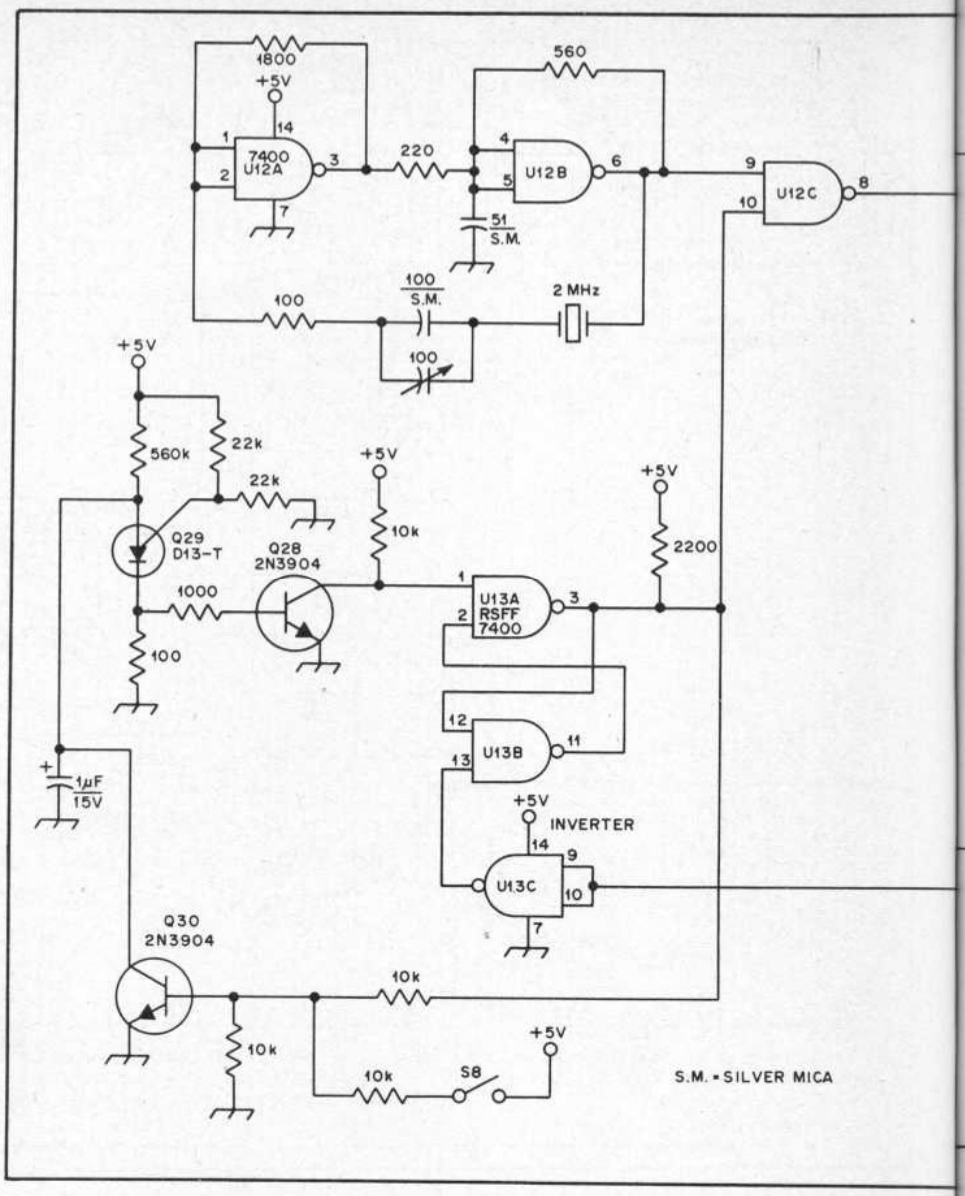
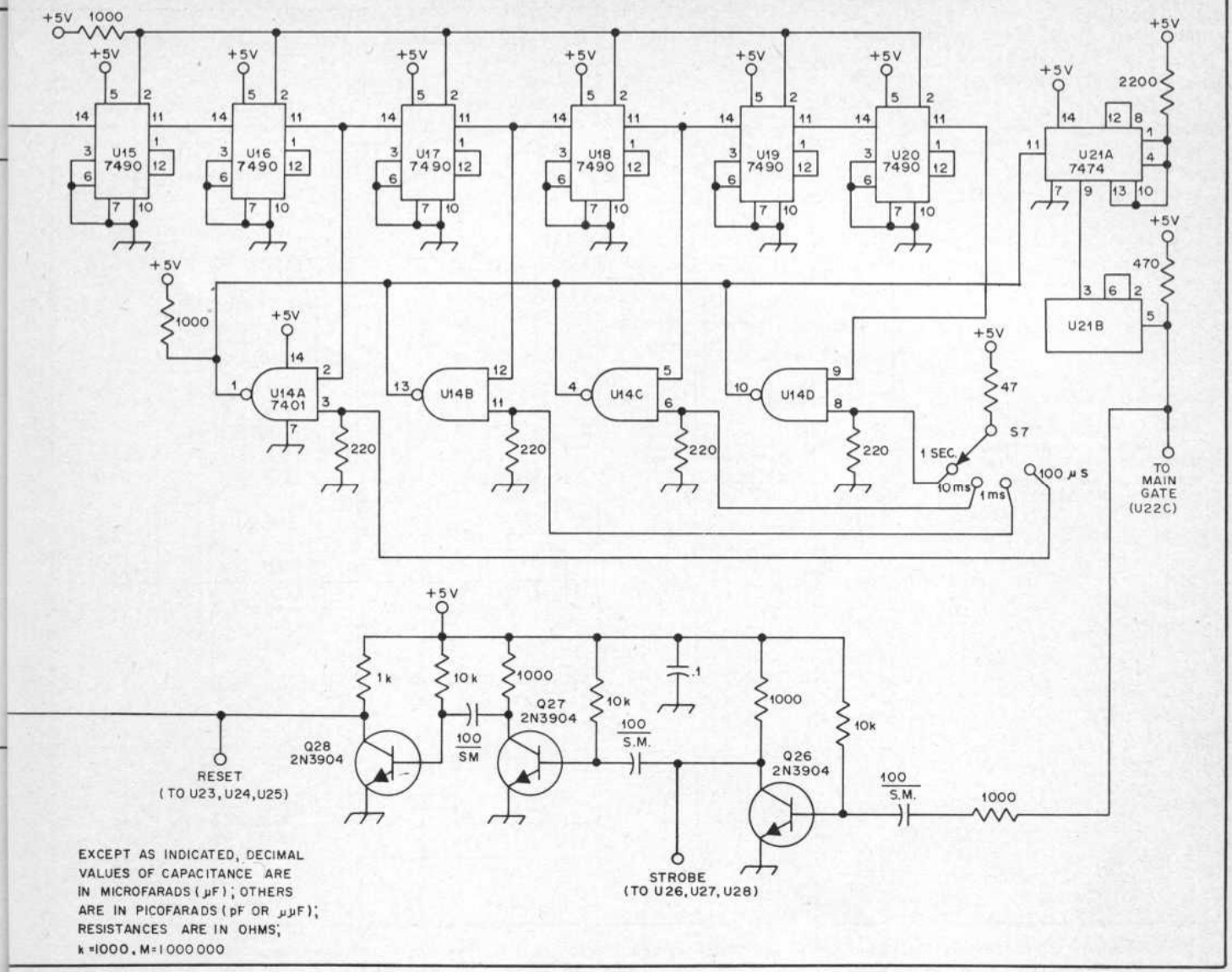


Fig. 19 — Schematic diagram of the time base and control system for the frequency counter. Fixed-value capacitors are disk ceramic unless otherwise indicated. Polarized capacitors are electrolytic. The variable capacitor is a mica compression trimmer. Resistors are 1/4- or 1/2-W composition.

Band switching in the exciter is realized with a three-pole, double-throw slide switch. The circuits that require switching are the positive supply to the amplifier chains and the output of the amplifiers. Also, switching is needed for band changing the network at the output of the exciter.

The output stage delivers 1 watt on each of the two bands. A 2N3553 transistor is used because of its ruggedness. Input matching is provided with a composite 9:1 impedance-ratio transformer consisting of T11 and T12. The output is matched by means of a modified T network. It was found that one network would function for two bands by changing the center capacitance. The network Q will be higher at 14 MHz. It was also noted that the typical T configuration was prone to vhf instability. This was eliminated by using a 470-pF capacitor at the collector. Another was required at the output to preserve network symmetry. One section of the band switch (S1C) adds capacitance for 40-meter operation. Front-panel tuning is not provided.

The output amplifier, which is packaged separately from the exciter, is shown in Fig. 18. A 2N5942 is used. This device is capable of more than 80 watts of output. In this application, it runs conservatively. Input matching is performed with a composite 16:1 impedance-ratio transformer formed



U12, U13 – 7400 quad, dual-input, NAND gate.
U14 – 7401 quad, dual-input, open-collector NAND gate.
U15 – 7490 decade counter.
U21 – 7474 dual D-type flip-flop.

from T13 and T14. Some attenuation is present at the input. The amount of attenuation is higher at 7 MHz than at 14. This results from the network associated with L39, which is tuned to 14 MHz. The action of this network is similar to that used in a diplexer. The transformers (T13 and T14) were units on hand. They seem to do the job adequately. The reader is referenced to the earlier discussion of impedance-matching methods for power amplifiers.

The output of the amplifier is matched with T15. This broadband 4:1 impedance-ratio transformer presents a 12.5-ohm termination to the collector of Q25. Output filtering is performed with a double pi network for each band. The filters are selected by means of two wafer switches. A double wafer switch would function as well. The output of this amplifier is 25 watts on each band.

Digital Readout

Early in this project, it was decided that digital methods would be used for frequency readout. The advantages of this, along with some general discussion of methods, were presented in chapter 6. A further motivation was a need for a general-purpose counter for experimentation.

The time-base and counter-control section is shown in Fig. 19. TTL logic is used exclusively. The clock for the circuit operates at 2 MHz, using an oscillator composed of a pair of NAND gates (U12A and B). This frequency was chosen since it has no harmonic output at the 9-MHz i-f of the receiver. The output of the clock oscillator is applied to a gate (U12C), which then drives the count-down chain. This divider is composed of six 7490 decade dividers and a divide-by-two divider using half of a dual-D flip-flop (U15 through U21A). Four different outputs may be selected from the divider chain. This is done with S7, a multiposition wafer switch. S7 controls the appropriate inputs of a 7401 quad NAND gate. This type differs from the usual 7400 in that the outputs are open collectors. Also, the pin-out is different! The outputs of the 7401 (U14) are "wire ored" to drive U21A. Depending upon the position of S7, the time base will have a period of 100 μs, 1 ms, 10 ms, or 1 second. During normal receiver operation, the 10-ms time base is used, allowing read-

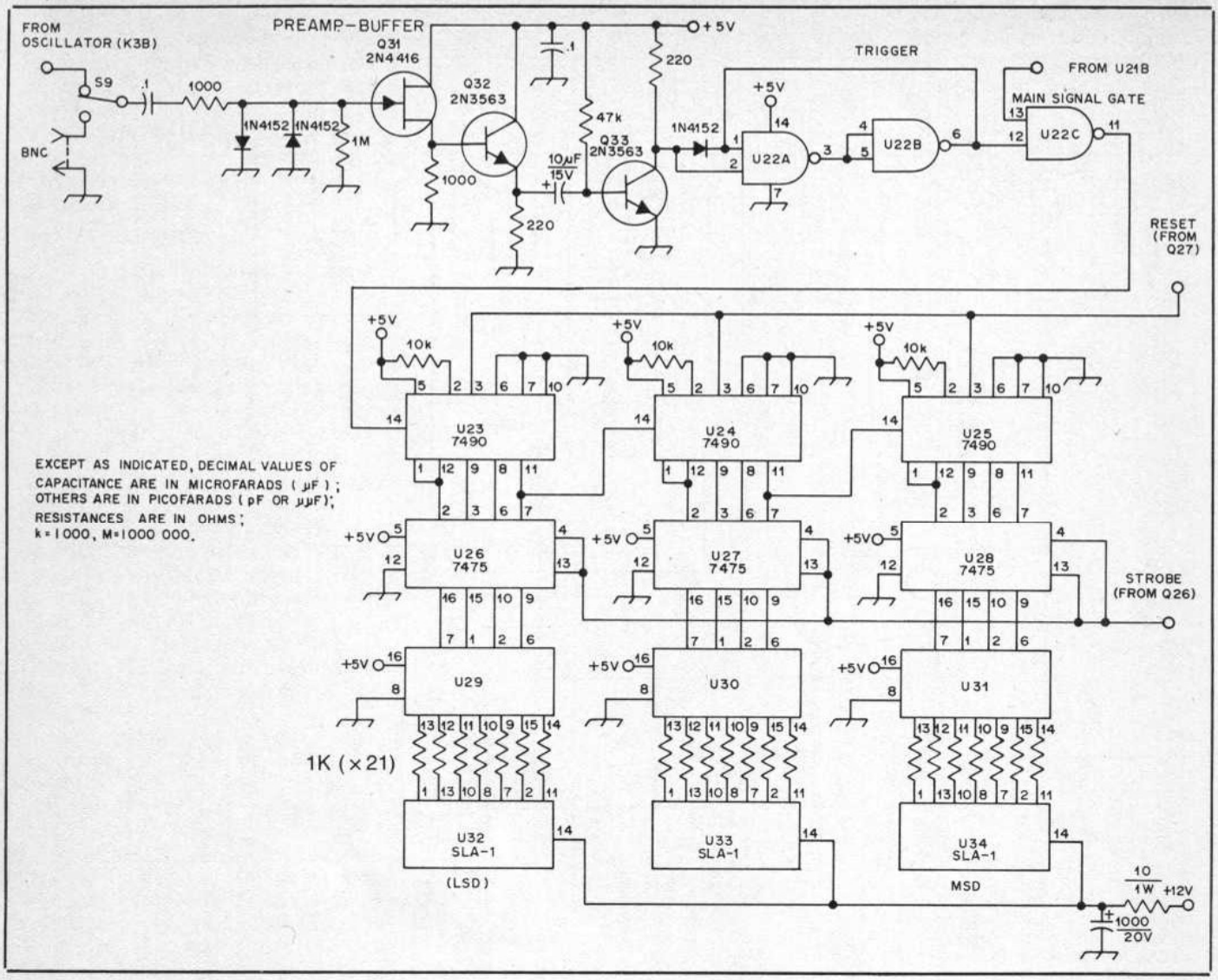


Fig. 20 — Schematic diagram of the signal-conditioning and frequency counters for the digital readout. Fixed-value capacitors are disk ceramic except those with polarity marked, which are electrolytic. Fixed-value resistors are 1/4- or 1/2-W composition.
U22 — 7400 quad, dual-input NAND gate.
U23-U25, incl. — 7490 decade counter.
U26-U28, incl. — 7475 quad latch.
U29-U31, incl. — 7447A BCD to 7-segment decoder/driver.
U32-U34, incl. — Opcoa SLA-1 7-segment LED display.

out to 0.1-kHz resolution.

The time-base output (pin 9 of U21A) is applied to a second flip-flop (U21B). This circuit produces pulses which are positive for the length of the time-base period. It controls the main counting gate (U22C in Fig. 12). At the end of a counting period, U21B will reset to a logical zero. The negative-going transition will be differentiated by the *RC* network driving Q26. The result is a pulse that is utilized to strobe the latches in the main counter (Fig. 19). The strobe pulse is also differentiated, producing a reset pulse. This resets the signal counters. It is also inverted in U13C and then used to set an RS flip-flop composed of cross-coupled gates (U13A and B).

When the RS flip-flop is set, a number of things happen. First, the gate in the time-base chain (U12C) is inhibited. This prevents the 2-MHz pulses arriving from the clock from triggering the divider chain. Hence, there is no digital circuitry operating except for the clock. This can benefit in reducing the level of digital noise injected into the receiver. The flip-flop also causes Q30 to be cut off, allowing a relaxation oscillator (Q29) to begin. This circuit has a time period of about 0.5 second. At the end of that period, a positive pulse is produced at the cathode of the PUT (Q29). This resets the RS flip-flop, allowing the clock to again be divided. Counter operation commences again and the cycle is repeated. The PUT oscillator establishes the update rate on the display. Switch S8 was not found necessary in this unit. If closed, it will completely inhibit the counter, leaving the last frequency that was counted in the displays.

The main signal-counting portion of the circuit is shown in Fig. 20. Switch S9, a front-panel-mounted toggle type, selects either the internal LO or an input from a panel-mounted BNC connector. With this, the counter may be used for general purpose applications. The preamp/buffer amplifier allows signals as low as 50 mV pk-pk to be counted. A pair of gates are connected in a trigger circuit (U22A and B). The main counting gate is U22C. This determines the number of input pulses that reach the counting chain.

Three decades of counting are provided with 7490 decade counters. The BCD outputs are applied to three 7475 quad latches. The resulting signals feed 7447A 7-segment decoder-drivers. The displays are Opcoa SLA-1 types, which are 1/3 inch high, common-anode types. At the time the counter was built, the

cost of each display was $5. At this writing, similar or better versions are available for as little as $2 each. Because of the decreased prices, it is highly recommended that four or even five decades of counting be used. This circuit is capable of operation up to 25 MHz, even though this is slightly higher than the specification of the TTL devices.

Concluding Thoughts

The station just described is not the sort of project that is undertaken casually. Including the power supply components, a total of 37 transistors and 36 integrated circuits are employed. A much simpler station with an equivalent power output would probably yield an equal number of contacts.

On the other hand, there are some circuit features that cannot be found in commercially manufactured equipment. Fourteen poles of crystal filtering leads to selectivity that is better than the writers have experienced in any other equipment they have used, homebuilt or commercial. The agc system is totally uncompromising. The dynamic range is better than any commercially manufactured amateur receivers that we have reviewed.

Above all of the features listed, the station is personal. Not only does operation of such equipment offer more satisfaction than might be realized with an "appliance," but the operator has gained the experience of learning and understanding. That's "where it's at!"

Appendix 1

The Phasing Method of SSB

Single sideband a-m phone may be generated with two balanced modulators, each driven with identical amplitude carrier and audio voltages. The two voltages applied to one modulator are out of phase by 90 degrees from those applied to the other. See Fig. 1 where the voltages are analytically defined. Note that the phase-shifted signals are denoted with a prime sign throughout this discussion. The "c" and "a" subscripts denote carrier and audio signals.

Each balanced modulator is assumed to be a perfect multiplier. Thus the output voltages are given by

$$E_o = KE_cE_a$$
$$E_o' = KE_c'E_a' \qquad \text{(Eq. 1)}$$

We will assume $K = 1$ for simplicity. If we insert the voltages from Fig. 1, we obtain for the two modulator outputs

$$E_o = V_cV_a \sin\omega_c t \sin\omega_a t$$
$$E_o' = V_c'V_a' \sin(\omega_c t + \pi/2) \sin(\omega_a t + \pi/2) \qquad \text{(Eq. 2)}$$

Three standard trigonometric identities which we will use are

$$\sin A \sin B = 1/2\left[\cos(A - B) - \cos(A + B)\right] \qquad \text{(Eq. 3)}$$

$$\sin(A + \pi/2) = \cos A \qquad \text{(Eq. 4)}$$

$$\cos A \cos B = 1/2\left[\cos(A + B) + \cos(A - B)\right] \qquad \text{(Eq. 5)}$$

Applying the identity of Eq. 3 to E_o as given in Eq. 2, we obtain

$$E_o = 1/2V_cV_a\left[\cos(\omega_c - \omega_a)t - \cos(\omega_c + \omega_a)t\right] \qquad \text{(Eq. 6)}$$

The two terms represent the lower and upper sidebands respectively. Applying the other identities to the E_o' expression of Eq. 2, the output of the other modulator is given by

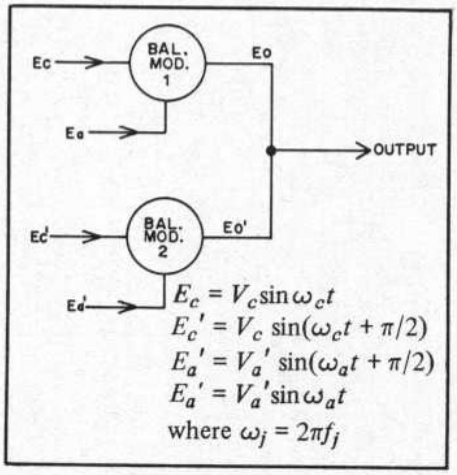


Fig. 1 – Phasing method of ssb generation. The equations define the applied voltages.

$$E_o' = V_c'V_a' \cos\omega_c t \cos\omega_a t = 1/2V_c'V_a'\left[\cos(\omega_c - \omega_a)t + \cos(\omega_c + \omega_a)t\right] \qquad \text{(Eq. 7)}$$

Again, both lower and upper sidebands are represented.

If the 2 outputs, E_o and E_o', are added as shown in Fig. 1, the resultant output voltage E_{net} is given by

$$E_{net} = E_o + E_o' = 1/2\left(V_cV_a + V_c'V_a'\right)\left[\cos(\omega_c - \omega_a)t\right] + 1/2\left(V_c'V_a' - V_cV_a\right)\left[\cos(\omega_c - \omega_a)t\right] \qquad \text{(Eq. 8)}$$

If $V_c = V_c'$ and $V_a = V_a'$, the second term vanishes leaving only the first term which is the lower sideband. The upper sideband may be obtained by subtracting the two modulator outputs. Alternatively, one balanced modulator may be driven by E_c and E_a' with the other operating on E_c' and E_a. The outputs are then added.

Clearly from the equations, the carrier voltages and the audio voltages must be equal in amplitude to obtain complete cancellation of the unwanted sideband. Consider the effect of a phase difference, θ, other than 90°. This is shown in the phasor diagram of Fig. 2, where θ is slightly under 90°. However, Eq. 8 gives the output in terms of the amplitude of the two phase quadrature (90° difference) signals. It may be shown that E_a' of Fig. 2 may be resolved into the sum of a voltage in phase with E_a and another 90° out of phase. These are also shown in Fig. 2. We see that a phase difference less than 90° tends to increase the V_a terms and decrease the V_a' terms in Eq. 8.

As an example, assume that the magnitude of all voltages is 1, but the audio phase difference, θ, is 88° instead of 90°. In this case the effective value of V_a is $1 + \cos 88°$ while V_a' is $\sin 88°$. The values, respectively, are then $V_a = 1.0349$, and $V_a' = 0.9994$. The amplitude of the two sidebands is then

$$E_{lsb} = 1/2\,(1.0349 + 0.9994)$$
$$E_{usb} = 1/2\,(0.9994 - 1.0349) \qquad \text{(Eq. 9)}$$

Taking 20 times the log of the ratio of the two, we find that the suppression of the undesired sideband is 35.2 dB. Slight phasing errors are of consequence.

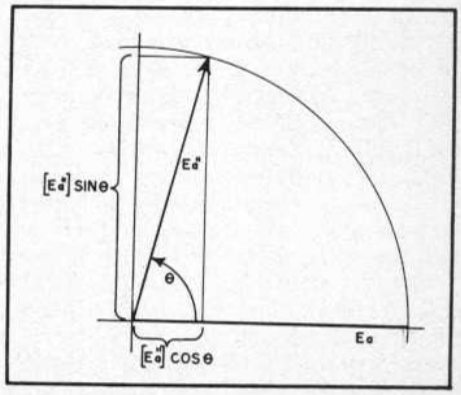


Fig. 2 – Phase relationships which pertain to Eq. 8.

Appendix 2

Band-pass Filters

A number of 2- and 3-pole band-pass filters have been designed. The filter synthesis was performed with computer programs using the predistorted Butterworth tables of Zverev (see the bibliography). Linear interpolation between the data given in the Zverev tables was used in the program.

Several of the filters have been built for evaluation. Others have been studied using computer techniques. These included both programs for nodal analysis and for microwave-network analysis and optimization. In all cases, excellent correspondence has been obtained with the data presented in the included tables.

Predistortion implies that the unloaded Q of the resonators must be known. The *filter* loaded Q is defined as $Q_L = F_o/\mathrm{BW}_{3\mathrm{dB}}$ where F_o is the center frequency and BW is the 3-dB bandwidth. A parameter called the normalized Q is defined as $Q_o = Q_u/Q_L$ where Q_u is the unloaded Q of the resonators used in the filters. It is assumed that all of the resonators have equal Q_u. While not mandatory for accurate synthesis, we have used identical inductance values in a given filter. Once the normalized Q is known, the insertion loss of the filter is well defined. Shown in Fig. 1 are curves of insertion loss vs. Q_o for Butterworth filters with from 1 to 4 poles.

Six inductors were wound on Amidon toroid cores. They were evaluated over a wide range of frequencies with a Boonton 160 Q meter. This data was used in calculating the filter components presented in the tables. The winding data for the six inductors are presented in Table 1. It is important that these inductors be duplicated exactly when building filters from the tables.

The circuit of a doubly terminated 2-pole filter is shown in Fig. 2. The form shown in A is one where the filter is terminated in a high impedance, characteristic of the filter. The form shown in Fig. 2B uses capacitors for transforming an external load, R_L, to present a proper termination to the filter. Methods for link coupling will be shown later.

Table 2 presents the data for 31 two-pole band-pass filters. Data presented include the band-edge frequencies (3-dB points), the inductor from Table 1 to be used, the normalized Q, Q_o, the nodal capacitance, C_o, the coupling capacitor between resonators, C_{12}, the capacitance at the ends of the filter required to match 50 ohms and finally, the resistance that must be placed across the ends of the filter. The nodal capacitance, C_o, is that required to resonate the inductor at the center frequency. The resistor values are given for calculations in cases where capacitive or inductive connection to a well defined external impedance is used. It is

Table 1

INDUCTOR NUMBER	*WINDING DATA*	*CORE TYPE*
L1	10 turns No. 24	T50-6
L2	12 turns No. 22	T68-6
L3	20 turns No. 22	T68-6
L4	30 turns No. 22	T68-2
L5	38 turns No. 24	T68-2
L6	33 turns No. 20	T106-2

Winding data for the toroidal coils used in the band-pass filters. All cores are available from Amidon Associates, G. R. Whitehouse and Palomar Eng. (See *QST* ads.) Enameled wire is used for all windings. The wire is distributed evenly over the core and is wound tightly. A layer of polystyrene *Q*-dope is applied after winding.

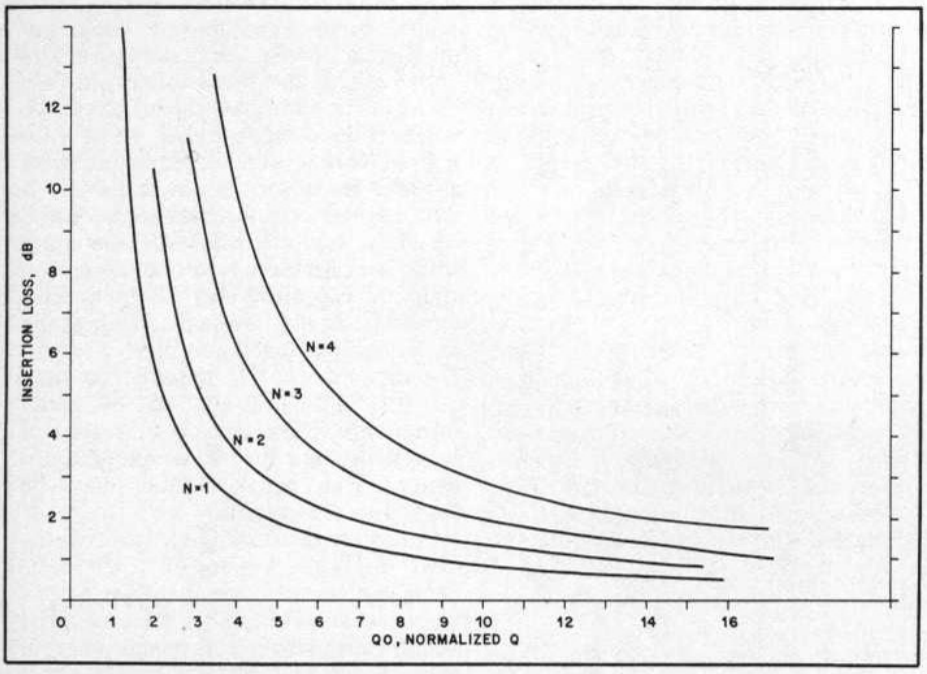


Fig. 1 — Insertion loss versus Q_o for Butterworth filters with one to four poles.

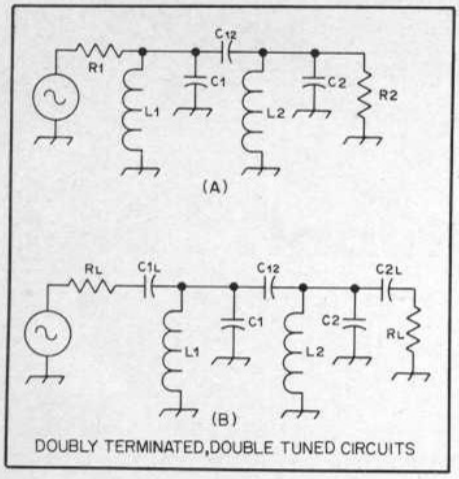


Fig. 2 — Examples of doubly terminated double-tuned circuits. See text for C1 and C2 value.

usually not necessary that the resistors be included when building the circuits.

Since only the nodal capacitances are given in Table 2, it will be necessary for the builder to calculate the actual capacitance, C1 and C2 (see Fig. 2), that will be used in a practical circuit. The equations are also given in Table 2. All of the filters in the tables are doubly terminated. That is, each end of the filter must be terminated with a resistive load of the proper magnitude. All filters are not necessarily doubly terminated, however. An example of a familiar singly terminated filter is the pi network used in transmitter-output stages. Singly terminated 2-pole filters will be discussed later.

Shown in Fig. 3 is the circuit for a 3-pole filter. This filter is also doubly terminated. The circuit at Fig. 3A is the complete 3-pole filter. The subscripting has been chosen to signify the position of that component in the filter. That is, C2 is the capacitor needed to tune the second resonator while C_{23} is the coupling capacitor between resonators 2 and 3. All inductors being labelled L signifies that all are identical.

Table 3 gives data for 30, three-pole band-pass filters. The normalized Q is given, allowing the insertion loss to be evaluated from Fig. 1. C1, C2 and C3, the capacitors to tune the three resonators, are given as are the respective coupling capacitors C_{12} and C_{23} Also presented are the proper end capacitors, C_{01} and C_{30} needed to match to 50 ohms.

Sometimes it is desirable to match the ends of the filter with links (perhaps to impedances other than 50 ohms) or to terminate the filter in a high value of resistance. The appropriate methods are shown in Fig. 3B and C. Note that it has been necessary to replace the resonator

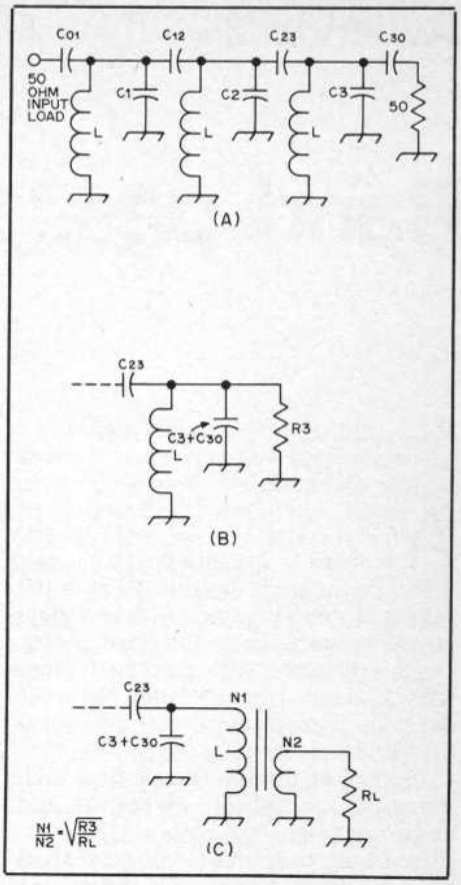


Fig. 3 — Doubly terminated 3-pole filter.

Table 2

3-dB FREQ.'S MHz	L	Q_o	C_o, pF	C_{12}, pF	C_{end} TO 50Ω	R1 & R2, kΩ
1.8 - 1.85	L5	5.4	870	16.8	148	7.02
1.8 - 1.9	L5	10.6	847	32.4	221	3.06
1.8 - 1.85	L4	5.2	1485	28.8	192	4.12
1.8 - 1.9	L4	10.3	1446	55.3	291	1.8
3.5 - 3.7	L5	12.5	224	8.8	83.4	5.67
3.5 - 3.6	L5	6.3	230	4.6	56.6	12.6
3.8 - 4.0	L5	11.5	191	6.9	70.7	6.7
3.8 - 4.0	L4	12.7	325	11.8	93.2	3.89
3.5 - 3.7	L4	13.7	382	15.0	110	3.28
3.5 - 3.6	L4	7.0	393	7.8	75.1	7.19
5.0 - 5.2	L5	8.3	111	3.1	40.1	12.1
7.0 - 7.1	L3	3.8	248	2.5	26.6	14.4
7.0 - 7.2	L3	7.5	245	4.9	42.4	5.6
7.0 - 7.3	L3	11.2	242	7.2	53.2	3.55
7.0 - 7.2	L5	4.9	57.5	1.1	19.1	27.4
10.7 - 11.1	L3	9.9	102.7	2.7	25.9	6.4
10.8 - 11.0	L3	4.6	102.7	1.3	16.7	15.3
14.0 - 14.2	L3	3.6	61.6	0.6	9.2	30.0
14.0 - 14.4	L3	7.2	60.4	1.2	14.7	11.6
14.0 - 14.4	L2	3.3	158.7	1.6	14.4	12.3
16.0 - 16.5	L2	7.6	118	2.6	20.4	4.7
19.0 - 21.0	L1	19.5	129	9.2	37.8	0.94
19.0 - 20.0	L1	10.0	136	4.9	26.6	1.93
19.0 - 19.5	L1	5.1	139	2.6	17.6	4.49
21.0 - 21.5	L2	5.7	68.4	1.1	11.4	8.7
21.0 - 21.3	L1	2.8	115	1.2	9.2	13.3
21.0 - 21.5	L1	4.6	114	1.9	14.1	5.69
28.0 - 29.0	L1	6.3	64	1.6	11.8	4.5
28.0 - 28.5	L1	3.2	65.3	0.8	7.2	12.4
41 - 42	L1	3.1	29.9	.05	4.6	13.7
41 - 43	L1	6.2	29.3	0.1	7.6	4.98

Data for doubly terminated double-tuned circuit. See Table 1 for coil data and Fig. 2 for schematic of filter. Note that the resonator capacitances are not given. $C_j = C_o - C_{12} - C_{end}$. See text for data and explanation.

capacitor, C3, with a larger one to account for the capacitive reactance of C_{30} presented by the end loading shown in Table 3. While these methods are shown only for the "output" end of the filter, they may be applied equally to the input section.

If capacitive coupling to a resistance other than 50 ohms is desired at one or both ends of the filter, this is possible. Considering the input, the nodal capacitance of the resonator is $C_0 = C_{01} + C1 + C_{12}$. This value will be the same for all three resonators. Given in Table 3 is the required resistive load at the input, R1. If it is desired to load the input with a resistance lower than R1, a coupling capacitor may be used. This capacitor should have a reactance at the center frequency of $X_c = \sqrt{R_1 R_L - R_L^2}$. The capacitance, C1, required to tune the first resonator will be the nodal capacitance less the inter-resonator capacitance and the end coupling capacitance. The same method may be applied to the output.

Consider an example. One of the filters in Table 3 is for 28 to 29 MHz. From the table we see that C1 = 46 pF. C_{01} = 16.6 pF and C_{12} = 1.6 pF. The nodal capacitance is then the sum of these, C_o = 64.2 pF. Assume that it is desired to couple a 600-ohm load into

Table 3

F_{3dB} MHz	*L No.	Q_o	C1	C2	C3	C_{01}	C_{30}	C_{12}	C_{23}	R1 KΩ	R3 KΩ
1.80 1.85	4	5.2	1186	1430	1289	273	286	26.9	28.6	2.0	5.4
1.8 - 1.9	5	10.6	505	783	605	308	212	33.9	30.7	1.6	3.4
1.8 - 1.85	5	5.4	646	838	724	209	130	15.9	16.7	3.5	9.0
1.8 - 2.0	5	21	314	684	438	426	311	65.2	55.9	0.83	1.5
1.8 - 1.85	6	9.0	360	522	416	172	117	10.8	10.0	5.1	11.1
1.8 - 1.825	6	4.6	431	541	477	115	68.7	4.8	5.4	11.7	32.7
3.5 - 3.6	5	6.3	146	221	174	79.3	51.2	4.5	4.5	6.4	15.4
3.8 - 4.0	5	11.5	85.9	177	116	97.5	67.8	7.3	6.5	3.6	7.3
3.8 - 4.0	4	12.7	184	302	224	129	90	12.5	11.2	2.0	4.2
3.5 - 3.7	4	13.8	214	352	261	152	107	16.0	14.1	1.8	3.5
3.5 - 3.6	4	7.0	280	377	316	105	68.8	7.8	7.6	3.7	8.6
3.9 - 4.0	4	6.3	228	306	258	84	54	5.5	5.6	4.7	11.2
3.5 - 4.0	5	33.1	88.5	286	154	229	168	36.3	31.1	0.74	1.3
5.0 - 5.5	5	24	52.9	155	84.6	114	84	13.2	11.3	1.5	2.7
7.0 - 7.1	3	3.8	208	244	220	38.4	25.8	2.1	2.6	7.0	15.0
7.0 - 7.2	3	7.6	181	236	201	59.1	39.1	4.9	4.7	2.9	6.6
7.0 - 7.3	3	11.2	161	227	184	73.7	68.0	7.5	7.0	1.87	3.86
10.7 - 11.1	3	10.1	63.9	97.4	75.4	36.0	24.8	2.7	2.5	3.3	7.0
12.0 - 12.5	3	10.8	46.9	76.7	57.0	32.0	22.1	2.5	2.2	3.4	7.0
14.0 - 14.2	3	3.6	47.4	60.1	51.7	13.2	8.9	0.5	0.6	14.5	32.1
14.0 - 14.4	3	7.18	38.6	58	45.6	20.5	13.6	1.2	1.1	6.0	13.7
16.0 - 16.5	2	7.6	87.3	113.2	97	28.4	18.8	2.6	2.5	2.4	5.5
19 - 20	1	10.0	93	126	106	37.2	25.3	5.1	4.7	1.0	2.1
19.0 - 19.5	1	5.1	112	134	121	25	15.3	2.4	2.6	2.2	5.9
21.0 - 21.5	2	5.7	51.3	66.2	57.2	16.0	10.1	1.0	1.1	4.4	11.1
28.0 - 28.5	1	3.2	54	63.9	59	10.5	5.4	0.6	0.9	2.2	5.8
28 - 29	1	6.3	46	61.1	52	16.6	10.7	1.6	1.6	2.3	5.5
41 - 42	1	3.1	22.8	29	26.0	6.7	3.5	.4	.5	2.5	6.4
41 - 43	1	6.2	17.6	27.3	21.4	10.7	6.9	1.0	1.0	2.5	6.1
50 - 52	1	4.7	12.3	18.8	15.1	7.0	4.2	0.50	0.55	4.0	11.0

*Refer to inductors in Table 1.

this filter. From the table, we see that R1 = 2300 ohms. The reactance of the capacitor required will be $X_c = (2300 \times 600 - 600^2)^{1/2} = 1010$ ohms. The center frequency is 28.5 MHz (actually, the geometric mean should be used). The capacitance at this frequency with 1.01-kΩ reactance is 5.5 pF. The resulting input part of the filter is shown in Fig. 4. Note that C1 has changed slightly from the value used for a 50-ohm termination.

While computer analysis is handy when designing a large number of filters, it is not necessary. Shown in Fig. 5 is a set of nine equations which may be followed to design a 2-pole filter. The designations follow those used in the schematic of Fig. 2. To design a filter, all that is required to be known are the 3-dB frequencies (in Hz), the inductor (in henrys) and the unloaded Q of that inductor at the center frequency. With reference to Fig. 5, Eq. A gives the center angular frequency. Eq. B is the nodal capacitance in farads while Eq. C gives the loaded filter Q. Eq. D shows the coupling capacitance between resonators. Eq. E gives the net Q that each end section must be loaded to, while Eq. F gives the external Q. In a 2-pole, doubly terminated filter, these values are the same for each end. Eq. G gives the end loading resistance required to establish the previously defined external Q. Eq. H gives the capacitor needed to couple to a given R_L. R_L must be less than the corresponding R_e. Eq. I completes the calculations with the values of the capacitors to tune each resonator.

Fig. 6 shows an application of these calculations. The filter covers the 14.0- to 14.4-MHz range. The inductor is L3 from Table 1 with $L = 2.08$ μH with $Q_u = 255$ at 14 MHz. Note that this filter is included in the catalog, Table 2.

While not complicated, the exact design of filters with a larger number of poles is more involved. This results not only from additional component values that must be calculated, but from the so-called normalized coupling coefficients and end section Qs. These values are dependent upon the normalized Q of the filter. They are independent of Q_o in a two-pole filter, however, and are contained within the equations of Fig. 5.

Sometimes it is desirable to couple a filter with mutual inductors instead of capacitors. This is done easily using data

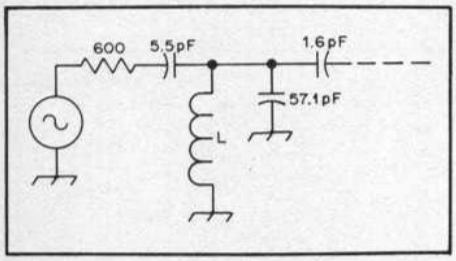


Fig. 4 — Input portion of a filter from Table 3.

$$\omega_o = 2\pi\sqrt{f_1 f_2} \quad \text{(Eq. A)}$$

$$C_o = (L\omega_o^2)^{-1} \quad \text{(Eq. B)}$$

$$Q_L = \omega_o / [2\pi (f_2 - f_1)] \quad \text{(Eq. C)}$$

$$C_{12} = C_o / (Q_L \sqrt{2}) \quad \text{(Eq. D)}$$

$$Q_j = \sqrt{2}\ Q_L \quad \text{(Eq. E)}$$
for $j = 1, 2$

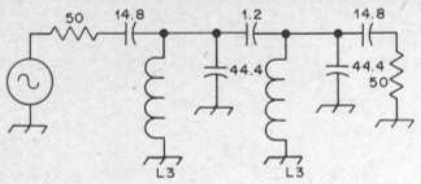
$$Q_{ej} = \left(\frac{1}{Q_j} - \frac{1}{Q_u}\right)^{-1}$$
for $j = 1, 2$ (Eq. F)

$$R_{ej} = Q_{ej}\ \omega_o L$$
for $j = 1, 2$ (Eq. G)

$$C_{jL} = \frac{1}{\omega_o \sqrt{R_{ej}R_L - R_L^2}}$$
for $j = 1, 2$ (Eq. H)

$$C_j = C_o - C_{jL} - C_{12}$$
for $j = 1, 2$ (Eq. I)

Fig. 5 — Algorithm for the design of a doubly terminated double tuned circuit.



$$\omega_o = 2\pi\sqrt{14 \times 10^6 \times 14.4 \times 10^6}$$
$$= 8.92 \times 10^7 \text{ radians per sec.} \quad \text{(Eq. A)}$$

$$C_o = (\omega_o{}^2 \times 2.08 \times 10^{-6})^{-1}$$
$$= 6.04 \times 10^{-11} \text{ Farad} = 60.4 \text{ pF.} \quad \text{(Eq. B)}$$

$$Q_L = \omega_o / (2\pi \times 0.4 \times 10^6) = 35.5 \quad \text{(Eq. C)}$$

$$C_{12} = 60.4 \times 10^{-11} / (35.5 \times 1.4.4)$$
$$= 1.2 \times 10^{-12} \; F = 1.2 \text{ pF.} \quad \text{(Eq. D)}$$

$$Q_1 = Q_2 = 1.414 \times 35.5 = 50.2 \quad \text{(Eq. E)}$$

$$Q_{e1} = Q_{e2}$$
$$= \left(\frac{1}{50.2} - \frac{1}{225}\right)^{-1} = 62.5 \quad \text{(Eq. F)}$$

$$R_{e1} = R_{e2}$$
$$= 62.5 \times \omega_o \times 2.08 \times 10^{-6}$$
$$= 11.6 \text{ k}\Omega \quad \text{(Eq. G)}$$

$$C_{1L} = C_{2L}$$
$$= \frac{1}{\omega_o\sqrt{11.6 \times 10^3 \times 50} - 2500}$$
$$= 14.8 \times 10^{-6} = 11.6 \text{ k}\Omega \quad \text{(Eq. H)}$$

$$C_1 = C_2 = 60.4 - 14.8 - 1.2$$
$$= 44.4 \text{ pF.} \quad \text{(Eq. I)}$$

Fig. 6 — Design of a filter using the method of Fig. 5. The filter is terminated in 50 ohms at each end and has a bandpass of 14 to 14.4 MHz.

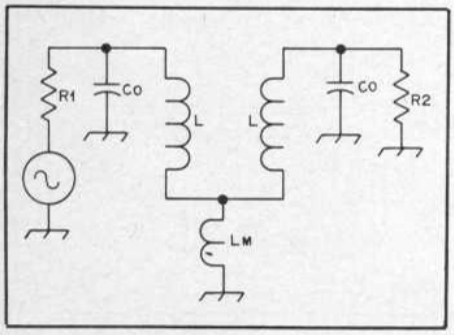


Fig. 7 — Mutual-inductor coupling method.

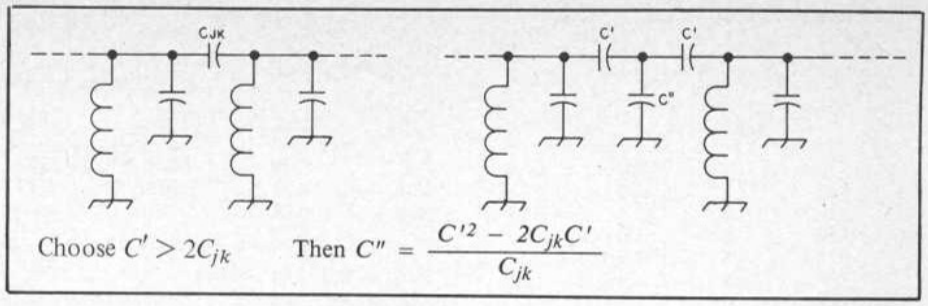


Fig. 8 — Method for dealing with small values of coupling capacitors.

implicit in the tables. The tables give the nodal capacitance, C_o, and the coupling capacitance between two resonators, C_{jk}. (In Table 2, C_o is given directly. In Table 3 it must be calculated.) The coupling coefficient between resonators is $K_{jk} = C_{jk}/C_o$. If a mutual inductor is to be used (see Fig. 7), its value is $L_m = LK_{jk}$. L_m is the value of the nodal inductor.

Ideally, for the calculations described it is best to measure the value of the inductance and Q_u at the frequency of application. However, we have found the data in the Amidon catalog to be accurate and suitable for these network calculations.

One of the practical problems encountered when building multipole filters is that of component selection. The capacitors used to tune each resonator are not difficult to realize. Usually, a combination of a fixed-value unit and a mica compression trimmer will serve adequately. The most severe problem is with the coupling capacitors. Tables 2 and 3 reveal a number of small, nonstandard values. One way to circumvent this problem is shown in Fig. 8. A desired small capacitor, C_{jk}, may be replaced by a network of three capacitors, two of value C' and a third of value C''. The equations for selection are given in the figure. Similar methods may be applied at the end sections. Link coupling can also be used at the ends, saving in component count and space. Alternatively, a combination of link coupling and a series capacitor can be used.

All of the filters presented have been doubly terminated. However, in the case of the two-pole filter, it is not always necessary that a filter be doubly terminated to function properly. In some cases, it is desirable that a filter not be terminated at both ends. One example might be the input to a receiver where a double-tuned circuit is used to drive the input of an FET mixer or amplifier. The lack of a termination leads to higher voltage transformation ratios, increasing gain of the FET circuit.

Shown in Fig. 9 is a singly terminated filter with two poles. The easiest way to realize such a filter in practice is to use the tables or Fig. 5 as a guideline and then adjust the filter empirically for the desired response. If a filter is designed for a given bandwidth when doubly terminated, but is then constructed according to Fig. 9, the result will usually be a double-humped response. A flat response could be achieved with a terminating resistor at the output of the filter, R2 of Table 2. Alternatively, the coupling capacitor to the load, C_{1L}, could be increased. This increases the loading on the first resonator, decreasing the value of Q_e for that circuit. Generally, a flat response is obtained if Q_e is decreased by a factor of $\sqrt{2}$ (see Fig. 5). When this is done, the ultimate bandwidth will be less than the original, again by a factor of about $\sqrt{2}$. This is often acceptable. It is not recommended that unterminated filters be built which utilize more than two poles.

Q Measurement and Filter Alignment

In the design of predistorted filters, it is necessary that the unloaded resonator Q be known prior to synthesis. While this value can often be measured with a Q meter, an equally viable method is shown in Fig. 10. A 50-ohm signal generator is used in conjunction with a 50-ohm detector. First, the generator (perhaps with an attenuator in its output) is connected to the detector and the response is noted. Then an unknown resonator is inserted, as shown in the figure. C_{in} and C_{out} should be equal and small in value. The capacitors are small enough when the insertion loss through the resulting one-pole filter is 30 to 40 dB. The generator is then tuned through the resonant frequency of the resonator, noting the frequencies where the detector response is down by 3 dB. The difference in the two is the *unloaded bandwidth*. The unloaded Q is

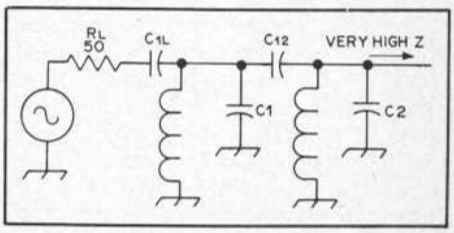


Fig. 9 — Singly terminated 2-pole filter.

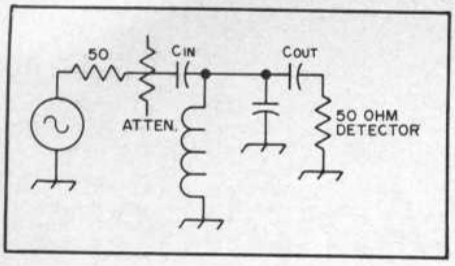


Fig. 10 — Method for determining resonator unloaded Q.

then the center frequency divided by the bandwidth. The advantage of this method over that of using a Q meter is that it is applicable at vhf and uhf, well above the range of Q-measuring instrumentation.

The design of multipole filters is covered by Zverev. The data for three-pole Butterworth filters have been applied here. However, tables are available for a number of response shapes with up to eight poles. Although not immediately obvious, the essence of such a design is to establish the singly loaded Q of the end sections of the filter and the coupling coefficients between resonators. As mentioned previously, the coupling coefficients in the two- and three-pole filters may be inferred from the data we have presented. The values of these parameters will depend upon the normalized Q of the filter.

Once a filter is designed, it still must be built and aligned. This is sometimes more difficult than the original synthesis. There are subtle, but easily performed procedures that can be applied.

One of the advantages of the Butterworth filter is that it is more easily aligned than many others. The signal generator should be set at the geometric mean of the 3-dB frequencies (the square root of the product of the frequencies). The filter is then adjusted for a maximum response at the other end. If the couplings and end loadings are proper, a flat response will usually result. Generally, those filters with a low normalized Q are the easiest to align. Unfortunately, they are also lossier.

Shown in Fig. 11 is a more advanced method of filter alignment. The filter is modified in two ways. First, a low-impedance detector is coupled very loosely to the first resonator. The probing capacitor, C_p, should be much smaller than any of the coupling or end loading capacitors. Second, each resonator has a switch across it. In practice, each may be a small piece of wire that is soldered temporarily to each resonator.

The first step in alignment is to close all switches except S1. The generator is set to the center frequency and C1 is adjusted for a peak in the detector. The generator is then swept around the center frequency, noting the 3-dB frequencies. This determines the loaded bandwidth and thus the loaded Q of the end section. C_{01} may be adjusted, if necessary, to produce the proper end Q. A similar procedure is performed at the output end of the filter to establish that Q, usually different than that of the input.

The next step is to reconnect the generator to the input. With all switches except S1 closed, C1 is peaked at the filter center. Leaving the generator set, S2 is now opened. C2 is adjusted for a *dip* in detector response. Note that the detector is still attached to the first resonator. If desired, at this point the coupling may be checked between resonators 1 and 2. If the generator is swept on either side of the center frequency, peaks will be measured. The coupling coefficient is approximately equal to the separation in frequency divided by the center frequency. (If the methods of Fig. 8 were being used, C'' could then be adjusted properly.) This general method may be used to evaluate the coupling between all of the resonators.

Assuming that the methods outlined have been applied, or the builder has otherwise assured that the coupling and loadings are proper, final alignment may be done. The setup of Fig. 11 is used, as shown. First, all switches except S1 are closed. C1 is tuned for a *peak* at the center frequency. Then, S2 is opened and C2 is tuned for a *dip* in detector response. Following this, S3 is opened and C3 is tuned for a *peak*. The procedure is continued until the filter is completely aligned. The output section should be terminated during alignment. The advantage of this method is that all alignment takes place at one frequency. This technique is attributed to Dishal (see bibliography). Equipment suitable for this method was described in chapter 7. After alignment, the detector is removed and the end of C_p is soldered to ground.

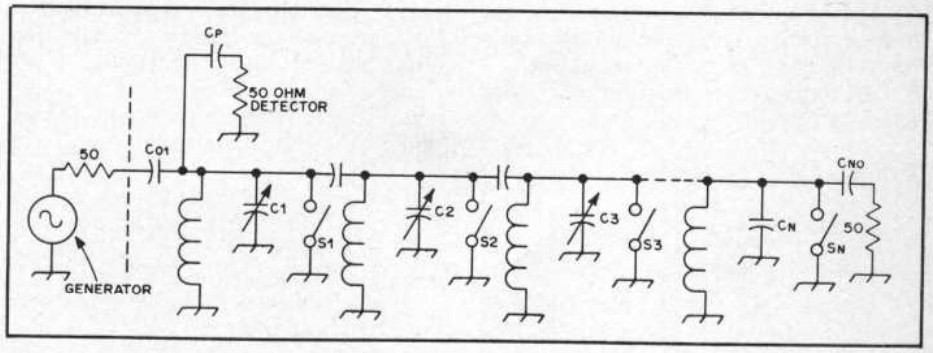


Fig. 11 — Advanced method for aligning a filter (see text).

Appendix 3

Distortion Properties of Amplifiers and Receivers

An amplifier, for the purposes of this discussion, is any nonreciprocal two-port network that might be used for signal processing. The usual function is to provide power gain. However, circuits such as diode mixers are analyzed using the same concepts.

The output of an amplifier can be expressed as a voltage, V_o, across some terminating resistance. This output is the result of an applied input voltage, V_i. The usual relationship of interest is of the form $V_o = G_v V_i$, where G_v is the voltage gain. However, the complete transfer function may be much more complicated. In the most general sense, all that may be assumed is that the output voltage is a continuous monotonic function of the input. As such, it can be expressed in the form of Taylor series expansion.

$$V_o = K_o + K_1 V_i + K_2 V_i^2 + K_3 V_i^3 + \ldots = \sum_0^\infty K_n V_i^n \quad \text{(Eq. 1)}$$

The term K_o merely represents a dc offset resulting from device biasing. The linear term, $K_1 V_i$ is the typically desired output. Harmonic and intermodulation distortion effects arrive from the high-order terms.

Consider an input signal of the form

$$V_i = E \sin \omega t \quad \text{(Eq. 2)}$$

where ω is the usual angular frequency, $2\pi f$, and E is the peak amplitude. If only the linear (first-order) term is considered, $V_o = K_1 E \sin \omega t$. The output signal contains only that frequency presented to the input. If two simultaneously applied signals (two tones) are considered,

$$V_i = E_1 \sin \omega_1 t + E_2 \sin \omega_2 t$$
$$V_o = K_1 E_1 \sin \omega_1 t + K_1 E_2 \sin \omega_2 t \quad \text{(Eq. 3)}$$

Consider now the application of a single tone and the effect of the second-order term. The resulting output will be, for $V_i = E \sin \omega t$

$$V_o \big|_{n=2} = K_2 (E \sin \omega t)^2 \quad \text{(Eq. 4)}$$

Using trigonometric identities, this reduces to

$$V_o \big|_{n=2} = \frac{K_2 E^2}{2} (1 - \cos 2\omega t) \quad \text{(Eq. 5)}$$

Two terms arise. The first is at dc, $1/2\ K_2 E^2$. If the dc output of the amplifier is monitored, a shift will be noted. This change is proportional to the square of the input amplitude. But input power is also proportional to E^2. Hence, for a 3-dB increase in input power, the output dc term will double. This is the basis of a square-law detector.

The second term to appear is $-1/2\ K_2 E^2 \cos 2\ \omega t$. This is an output component at twice the input frequency. It describes frequency-doubler action.

Consider now a two-tone input with the second-order term.

$$V_i = E_1 \sin \omega_1 t + E_2 \sin \omega_2 t$$
$$V_o = K_2 (E_1 \sin \omega_1 t + E_2 \sin \omega_2 t)^2$$
$$= K_2 [E_1^2 \sin^2 \omega_1 t + E_2^2 \sin^2 \omega_2 t + 2E_1 E_2 \sin \omega_1 t \sin \omega_2 t] \quad \text{(Eq. 6)}$$

The first two output terms are the same as obtained for the single-tone case, leading to square-law detection and frequency multiplication. The third term, $2K_2 E_1 E_2 \sin \omega_1 t \sin \omega_2 t$, is a result of having two input tones present. Again using trig identities, this term becomes V_o'

$$V_o' = K_2 E_1 E_2 [\cos(\omega_1 - \omega_2)t - \cos(\omega_1 + \omega_2)t] \quad \text{(Eq. 7)}$$

The resultant frequencies are sums and differences of the input frequencies. This accounts for the mixer behavior of devices with square-law responses. Note that the output amplitude is proportional to the product of the input amplitudes, E_1 *and* E_2. Hence, the popular terms "multiplier" and "product detector."

A similar procedure is used to evaluate the effect of the third-order term. For an input (single tone) $V_i = E \sin \omega t$,

$$V_o \big|_{n=3} = K_3 E_3^3 \sin^3 \omega t \quad \text{(Eq. 8)}$$

With trig identities, this reduces to

$$V_o \big|_{n=3} = \frac{K_3 E^3}{4} [3 \sin \omega t - \sin 3\omega t] \quad \text{(Eq. 9)}$$

The output signal contains the fundamental drive frequency and its third harmonic. Note, however, that the fundamental is three times as strong as the harmonic output, accounting for the stringent selectivity requirements at the output of frequency triplers.

If a two-tone input is considered, the third-order term yields

$$V_o = K_3 [E_1 \sin \omega_1 t + E_2 \sin \omega_2 t]^3$$
$$= K_3 [E_1^3 \sin^3 \omega_1 t + E_2^3 \sin^3 \omega_2 t + 3E_1^2 E_2 \sin^2 \omega_1 t \sin \omega_2 t + 3E_2^2 E_1 \sin^2 \omega_2 t \sin \omega_1 t] \quad \text{(Eq. 10)}$$

The first two terms are the superimposed single-tone outputs, which will reduce to Eq. 9. The third and fourth terms lead to more complicated properties. Consider the third term

$$3K_3 E_1^2 E_2 \sin^2 \omega_1 t \sin \omega_2 t$$
$$= 3E_1^2 E_2 K_3 \sin \omega_2 t (1 - \cos^2 \omega_1 t)$$
$$= 3E_1^2 E_2 K_3 \sin \omega_2 t - \frac{3K_3 E_1^2 E_2}{2} (\sin \omega_2 t \cos 2\omega_1 t + \sin \omega_2 t)$$
$$= \frac{3E_1^2 E_2 K_3}{2} \{\sin \omega_2 t - 1/2 \times [\sin(2\omega_1 + \omega_2)t - \sin(2\omega_1 - \omega_2)t]\} \quad \text{(Eq. 11)}$$

Similarly, the fourth term in Eq. 10 reduces to the expression

$$\frac{3K_3E_2{}^2E_1}{2}\left\{\sin\omega_1 t - 1/2\right.$$
$$\left.[\sin(2\omega_2+\omega_1)t - \sin(2\omega_2-\omega_1)t]\right\} \quad \text{(Eq. 12)}$$

Examination of the various resultant terms is enlightening. For each of the input frequencies, ω_1 and ω_2, we see terms where the overall oscillation at one frequency is dependent upon the amplitude at the other frequency. This is the phenomenon of cross-modulation.

The other terms of interest are the intermodulation ones at frequencies $2f_1 \pm f_2$ and $2f_2 \pm f_1$. The sum terms are normally not of great significance in amplifier design, for they are far removed from the desired output frequency. However, the differences are of major significance, for they lie very close to the desired outputs of f_1 and f_2. These are the common third-order IMD products. It is these components which, if excessive, cause an ssb signal to appear broad. Furthermore, it is this phenomenon which we have used to define receiver dynamic range, as well as the more fundamental input or output intercepts.

The higher-order terms in the basic transfer function are not analyzed here. However, the results are similar qualitatively. The fourth-order term will lead to outputs at $2f$ and $4f$. The fifth-order term will cause a number of components to exist including those at $3f_1 - 2f_2$, and $3f_2 - 2f_1$. These are the commonly referred to as *fifth-order IMD products*.

On the basis of this analysis, one would be quite fearful of building an amplifier and expecting anything approaching linearity. For example, consider an amplifier with inputs at 20 and 21 MHz. If we consider components in the transfer function only up to the third order, outputs would be expected at dc and at 1, 19, 20, 21, 22, 40, 42, 60, 61, 62 and 63 MHz, not to mention cross-modulation effects. The redeeming feature is that usually the K_1 constant in the series expansion is dominant, with high-order coefficients being progressively smaller (mathematically, the Taylor series expansion is rapidly convergent).

Consider the first- and third-order responses together with two equal input tones E at ω_1 and ω_2. We will assume that $K_1 >> K_3$. Hence, the output voltage amplitudes at ω_1 and ω_2 are each K_1E. The output voltages at the third-order *IMD* frequencies will each be $0.75K_3E^3$

The ratio of the voltages will be

$$R_{IMD} = A\frac{E_{in}}{E_{in}{}^3} = AE_{in}{}^{-2} \quad \text{(Eq. 13)}$$

where $A = 4K_1/3K_3$. (A may be determined from a spectrum analyzer measurement, as an amplifier example.)

Assume now that the input voltage is doubled. The desired output voltage will double, resulting in a four times (6 dB) increase in output power. However, the IM voltages will increase by the factor 2^3, or eightfold. The power increase will be 64 times (18 dB). The ratio of output voltages is 8/2 = 4, while the power ratio is R^2 or 64/4 = 16(12 dB). The input intercept will be the power corresponding to the input voltage causing $R = 1$ (see Eq. 13).

Consider a receiver where the input intercept, P_i, is known. If two tones X dB below the intercept are applied to the input, the *IM* responses will correspond to inputs 3 X dB below the intercept, and the *IMD* ratio will be 2 X. The two-tone dynamic range of a receiver is defined as the ratio of one of two equal tones causing an *IMD* response equal to the *MDS* (minimum discernible signal) to the level of the *MDS*. These relationships can be expressed analytically.

For a given input intercept, P_i, and input two-tone power, P_{ant} per tone, the *IMD* power, P_{IM}, is

$$P_{IMD} = P_i - 3(P_i - P_{ant}) \quad \text{(Eq. 14)}$$

where powers are in dBm. However, the dynamic range, DR, is defined as $P_{ant} - MDS$ for the condition that $P_{IM} = MDS$. Inserting this into Eq. 14, we have

$$MDS = P_i - 3P_i + 3P_{ant}$$
$$= -2P_i + 3P_{ant}$$
$$P_{ant} = \frac{MDS + 2P_i}{3} \quad \text{(Eq. 15)}$$

Hence,

$$DR = \frac{MDS + 2P_i}{3} - MDS$$
$$= \frac{MDS + 2P_i - 3MDS}{3}$$
$$= 2/3\,(P_i - MDS) \quad \text{(Eq. 16)}$$

This relationship is extremely useful for receiver system design and evaluation.

Appendix 4

Transistor Models and Amplifier Analysis

The small-signal model used for many of the designs in the book is repeated in Fig. 1. The transistor is assumed to consist of an input resistance of $26\beta/I_e(\text{mA})$ with a current source in the collector. The beta of this source (ratio of collector current to base current) is approximated by f_T/f at high frequencies.

As a practical application of this model, consider an amplifier with both emitter degeneration and shunt feedback. The circuit is terminated in a 50-ohm load and is driven from a 50-ohm generator with 2 volts, open circuit. The maximum power available, P_a, from this generator occurs when it is terminated in a 50-ohm load. In our example, P_a is 1 volt across 50 ohms, or 20 mW. The circuit for the feedback amplifier is shown in Fig. 2. Let $R_e = 10$ ohms and $R_f = 250$ ohms. The transistor is assumed to have an f_T 10 times higher than the operating frequency, resulting in $\beta = 10$. Assume that the emitter current is 10 mA, leading to an input resistance for the transistor of 26 ohms.

The various currents in the amplifier are defined according to the direction of the arrows. They are purely arbitrary. Final analysis will reveal the actual direction of current flow.

The circuit of Fig. 2 is analyzed by writing nodal equations. At each node in the circuit, the net current *entering* will be zero. In this amplifier, the nodes occur at the base, collector and emitter of the transistor. The equations are

(At the base node)

$$I_i + I_f = I_b$$

$$\frac{V_s - V_b}{R_s} + \frac{V_c - V_b}{R_f} = \frac{V_b - V_e}{R_b} \qquad \text{(Eq. 1)}$$

(At the collector node)

$$I_L = I_f + I_c$$

$$\frac{-V_c}{R_L} = \frac{V_c - V_b}{R_f} + \frac{\beta(V_b - V_e)}{R_b} \qquad \text{(Eq. 2)}$$

(At the emitter node)

$$I_b + I_c = I_e$$

$$\frac{V_b - V_e}{R_b}(1 + \beta) = \frac{V_e}{R_e} \qquad \text{(Eq. 3)}$$

There are now three equations in the three unknowns, V_b, V_e and V_c. This set of simultaneous equations may be solved using standard algebraic methods. There are two approaches that may be taken. One is a general solution, resulting in a set of equations for the three voltages. Direct substitution of the resistor and β values into the answer set will yield the needed output information. The advantage of this approach is that once the equations are solved, a wide variety of feedback elements and current gains may be evaluated.

The alternative solution is to immediately substitute the resistance and beta information into Eqs. 1 through 3. The results will then be quite specific. If the constants given earlier are placed into the three simultaneous equations, we obtain the results $V_b = 0.9315$ volt, $V_c = -2.6986$ volts, and $V_e = 0.7534$ volt. The negative sign on the collector voltage indicates that the amplifier is inverting.

The output power is $V_L^2/R_L = 146$ mW. The available generator power, P_a, was 20 mW. The transducer gain, G_T, is defined as the power output divided by the available generator power. This is the gain that would be observed if a 50-ohm matched line was broken and the amplifier was inserted. In this amplifier, $G_T = 7.28$, or 8.6 dB.

There are many other gains that may be defined. The one termed "power gain" is the output power divided by the actual power delivered to the input. Another would be the "maximum available power gain." This is the gain that would result if both the input and output were conjugately matched.

We calculated V_b as 0.9315 volt. From this, we can calculate the input current. $I_{in} = (V_s - V_b)/R_s = (2 - 0.9315)/50 = 21.4$ mA. But, the input resistance is V_{in}/I_{in}, or 43.8 ohms. This amplifier would present a good match to the 50-ohm source.

Calculation of the output resistance of the circuit is more complicated. The input generator is replaced with a 50-ohm resistor, while the output load is replaced with a generator of 50 ohms characteristic resistance. The nodal equations are written for this circuit and solved. The results will yield the output resistance and the reverse transducer gain. The circuit for this calculation is shown in Fig. 3. The equations and their solution are left as an exercise.

Experimentally, we find that the results predicted above correspond well

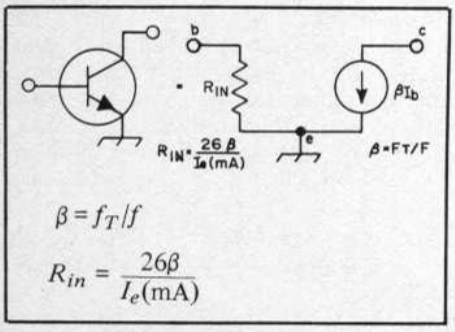


Fig. 1 — Simplified model for a bipolar transistor.

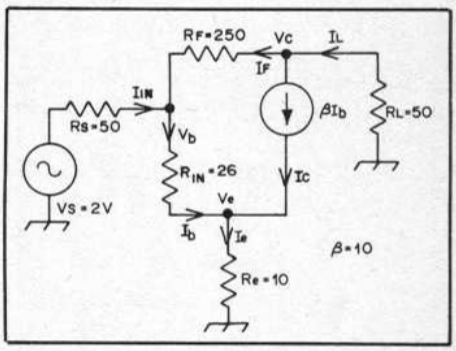


Fig. 2 — Small-signal amplifier using shunt and series feedback. The model of Fig. 1 is used for analysis. Arrows show the assumed direction of current flow.

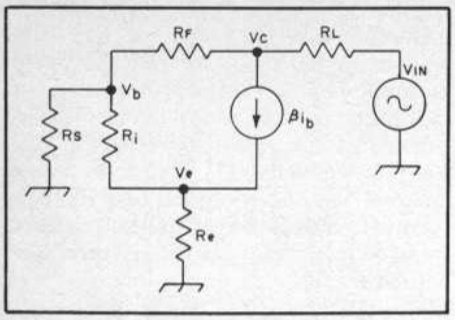


Fig. 3 — Circuit used for evaluation of the output impedance of the feedback amplifier.

with measured data. This is predominantly because of feedback. As we emphasized in the text, one of the major virtues of feedback is predictable circuit behavior, independent of active device characteristics.

The simple model of Fig. 1 is limited. It always predicts an output which is 180 degrees out of phase with the input signal. This is because we have neglected any reactive elements. A more complete model, known as the *hybrid pi*, is shown in Fig. 4. In this model, R_b' is a base resistance that is independent of current in the transistor. R_e is a built-in emitter resistance with a magnitude of $26/I_e$(mA). If this model is analyzed, we find that the emitter resistance, R_e, transforms to a base input resistance of $(\beta + 1)R_e$, similar to that used in the simplified model of Fig. 1. The most unique feature of the hybrid-pi model is the complex (algebraically) nature of beta. At very low frequencies, beta has the value β_o. However, as frequency increases, the magnitude of the effective beta decreases and becomes more reactive. A significant frequency is f_β which is defined as $f_\beta = f_T/\beta_o$. At this frequency, $\beta = (\beta_o/2)(1 - j1)$. The magnitude of beta is reduced by the factor $\sqrt{2}$ and the phase angle is –45 degrees (the collector current is 45 degrees out of phase with the base drive current). As frequency is increased further, beta becomes predominantly imaginary with the phase angle approaching 90 degrees. Many of the transistors used routinely in amateur applications are operated above f_β. For example, for the 2N3904 with $f_T = 300$ MHz and $\beta_o = 100$, f_β is only 3 MHz.

The final feature of the hybrid-pi model is a collector-base capacitance. This built-in feedback element leads to a further decrease in gain as f_T is approached over that implicit in the decrease in beta. It also leads to reverse gain and, sometimes, instability.

The implications of a complex beta can be profound. Consider the slightly simplified hybrid-pi model in the circuit of Fig. 5. In this analysis, we have neglected the collector-to-base capacitance. The results will be qualitatively the same if it is included.

The impedance that is in the emitter lead of the circuit is a paralleled 100-ohm resistor and a 100-pF capacitor. Assume that the operating frequency is 30 MHz. At this frequency, the emitter impedance is $Z_e = 21.9 - j41.4$ ohms. (This is arrived at by writing the admittance $Y = 1/R + j\omega C$. The impedance is then the reciprocal of the admittance, $Z = 1/Y = Y^*/YY^*$ where the asterisk signifies the complex conjugate.)

If the model is analyzed, we find that the input impedance is given by

$$Z_{in} = R_b' + (Z_e + R_e)(\beta + 1) \qquad \text{(Eq. 4)}$$

Assume that the transistor has $f_T = 300$ MHz and $\beta_o = 50$. Using the formula of Fig. 4, $\beta = 1.92 - j9.62$ (predominantly imaginary). Substituting the complex β and Z_e into Eq. 4 and assuming that $R_b' = 20$ ohms and $R_e = 2.6$ ohms ($I_e = 10$ mA), we find that $Z_{in} = -306 - j356$ ohms. It is significant that the real part of this impedance, the input resistance, is negative. This implies that if the input of this amplifier is terminated in a low value of resistance, perhaps with some inductive reactance to tune out the input capacitance, the stage will oscillate! If the original goal were to design an amplifier rather than an oscillator, stability could be regained with a larger emitter bypass capacitance. Alternatively, a series (positive) base resistor will improve stability. A shunt resistance, however, would not.

This analysis demonstrates some of the features of stability analysis. We will have more to say about stability later. Also, the example of Fig. 5 shows why emitter-follower amplifiers sometimes oscillate, especially when terminated in a capacitive load.

Another application of emitter reactance is shown in Fig. 6. Here, Z_e is a small inductor. The value of Z_e is $0 + j\omega L$, a positive imaginary element. If this is inserted into Eq. 4, the input impedance may be calculated. With no emitter inductance, the input impedance (assuming the same transistor parameters as were used above) is $27.6 - j25$. With a 100-nanohenry emitter inductor, $Z_{in} = 189 + j30$. The input is now predominantly real and much higher than before.

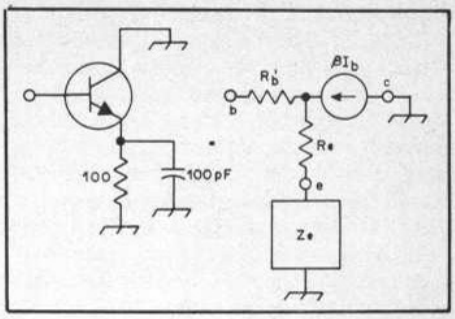


Fig. 5 — Circuit analysis showing the effect of a reactive emitter bypass.

This effect can be of profound importance in the design of very low noise amplifiers. Noise modeling generally attributes much of the excess noise output of a transistor to noise from R_b' and R_e of the hybrid-pi. To achieve a low noise figure, the input must be terminated so that much of the noise power of these elements is shunted to ground. However, a reactive element in a circuit contributes virtually no noise. The increased input impedance of the amplifier with emitter inductance results predominantly from two reactive effects — the reactance of the inductance and the complex, capacitive-like effect of an almost all reactive beta. The clever circuit designer may utilize this phenomenon to achieve a very low noise figure simultaneously with a 50-ohm, resistive-input impedance. This has the virtue of allowing the use of a multiple resonator filter ahead of the amplifier to protect it from strong out-of-passband signals. Typical multipole preselector filters must be doubly terminated.

Amateurs are presently in the process of rediscovering this phenomenon and are applying it to the design of receiver preamplifiers for moonbounce at 432 MHz. However, an exhaustive

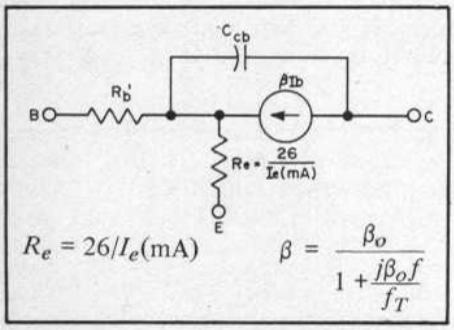


Fig. 4 — Refined model of a bipolar transistor. Note that beta is now a complex number.

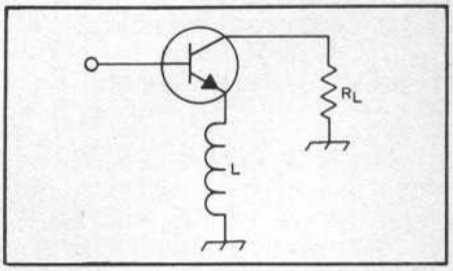


Fig. 6 — Amplifier with an inductive-emitter termination. This method is often used with microwave amplifiers to achieve a proper input impedance match while preserving amplifier-noise figure and stability.

search of the literature reveals that reactive feedback was used in low-noise amplifier design over 30 years ago (MIT Rad. Lab. series, see bibliography). These applications were with tubes. However, the same concepts apply. This early work in no way discredits the more recent efforts of enterprising amateurs. It does exemplify the value of reading the classic literature, even if it does not deal specifically with the latest semiconductor techniques.

Analysis of an FET Amplifier

Shown in Fig. 7 is a model for a JFET or MOSFET operating at low signal levels. In this model, the input is assumed to be an open circuit. This assumption is very good with MOSFETs in the hf region, and is usually good with JFETs as well. At vhf, the input impedance becomes lower. This is because of feedback from the drain to the gate through the capacitances.

Shown in Fig. 8 is a circuit of an amplifier using an FET. Each end of the amplifier is "matched" with a transformer. The transformers could be tuned, although this is not mandatory. Because the input to the FET is virtually an open circuit, it presents no loading to the transformer. The gate voltage is then N_1 times the *open-circuit* voltage of the generator.

The drain of the FET is modeled with a current generator. Unlike the one used for the bipolar transistor, this generator is voltage- rather than current-controlled. The drain current is $g_m V_{gs}$. V_{gs} is the small signal gate-to-source voltage and g_m is a parameter called the transconductance. This is short for transfer conductance. Note that conductance, which is the reciprocal of resistance, has the dimensions of amperes per volt. In this case, g_m specifies the current flowing in one leg of the circuit per volt in another part.

The equations relating the output voltage and current (in the load resistor) to the input signal are given as

$$V_g = E_s N_1 \qquad \text{(Eq. 5)}$$

$$I_d = g_m V_{gs} = g_m E_s N_1 \qquad \text{(Eq. 6)}$$

$$I_L = N_2 I_d = g_m E_s N_1 N_2 \qquad \text{(Eq. 7)}$$

$$V_L = I_L R_L = g_m E_s N_1 N_2 R_L \qquad \text{(Eq. 8)}$$

The power delivered to the load is

$$P_{out} = \frac{V_L^2}{R_L} = g_m^2 E_s^2 N_1^2 N_2^2 R_L \qquad \text{(Eq. 9)}$$

But, the power available from the source, P_a, is $E^2/4R_s$ where R_s is the source impedance. Using this, the transducer gain may be calculated

$$G_T = 4 g_m^2 N_1^2 N_2^2 R_L R_s \qquad \text{(Eq. 10)}$$

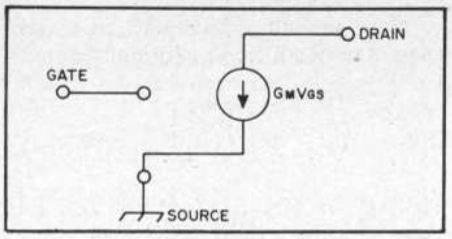


Fig. 7 – Simplified model of a field-effect transistor.

The term $N_1^2 R_s$ is just the source resistance seen by the gate. Similarly, $N_2^2 R_L$ is the load presented to the drain. Attaching primes to these parameters, the transducer gain of Eq. 10 is given by

$$G_T = 4 g_m^2 R_s' R_L' \qquad \text{(Eq. 11)}$$

Typical values for R_s and R_L would be 50 ohms. Turns ratios of 5 would be representative, leading to R_s' and R_L' of 1250 ohms. A common value for g_m of a MOSFET would be 10,000 micromho $= 10^{-2}$ mho. Inserting these parameters into Eq. 11, we arrive at $G_T = 625$, or 28 dB.

If the FET is operated as a mixer with optimum LO injection, the conversion transconductance is 0.25 that displayed by the same device operated as an amplifier. Using the same transformers, the conversion transducer gain would be 16 dB, 12 dB lower than that of the amplifier. These calculations are remarkably consistent with measurements with a 3N140 in the hf region in spite of the simplicity of the model.

Because the input to the FET is virtually an open circuit, no power may be delivered to the input. There is, nonetheless, a finite power output. The power gain is infinite. The theoretical maximum available gain (MAG) with this simple model is also unbounded. In practical applications, the MAG will be limited by stability considerations and by losses in the input transformer.

Another implication of the infinite input resistance of the FET is that the input VSWR is quite high. Again, in a practical amplifier it will be determined by loss elements in the input. If a good input VSWR is desired, a resistor is usually required from the gate of the FET to ground. This should equal R_s to provide a good input impedance match. This will reduce the voltage on the FET gate by a factor of 2 causing a 6-dB decrease in transducer gain. However, this will now allow multipole preselector filters to be used. A degraded noise figure is typical in such amplifiers.

Feedback may be applied to FET amplifiers to provide a controlled, real input impedance. While further work is required, it appears that the same concepts that were used with reactive feedback in bipolar transistor amplifiers may be employed to achieve a simultaneous input match and a low noise figure. Following the work reported in the Vally & Wallman (MIT Rad. Lab. series) volume, it appears that a combination of source inductance and resistive drain-to-gate feedback would produce the desired results.

Linear Two-Port Network Concepts

The nodal analysis presented in the preceding sections may be continued. The models will become more sophisticated, using perhaps over two dozen elements to describe just one bipolar transistor instead of the three or four we have considered. This technique is common practice in industry, usually through the realm of computer-aided design. This is especially useful for the evaluation of large-signal phenomenon. For most rf applications, however, small-signal analysis is adequate. There are more refined approaches to small-signal analysis. They are not only more convenient for calculation, but can be applied with measured device data without resorting to models (which may be limited from oversimplification). These methods are, however, still applicable with models. The vehicle is linear two-port network theory.

A complete treatment of two-port networks is beyond the scope of this text. Not only is the subject exhaustive, but it depends heavily on matrix algebra. This subject is not difficult, but is probably not in the background of many radio amateurs. Furthermore, the calculations, which are in principle straightforward, tend to become complicated in practice. In this section we will present basics of a few of the concepts available to the designer. This is intended to aid the amateur in understanding some of the terminology used in modern electronics. The analitically inclined reader may be inspired to investigate the subject in more detail. A recent text on the subject by R. Carson (*High-Frequency Amplifiers*, see bibliography) is highly recommended.

Shown in Fig. 9 is the circuit of an amplifier. A generator of known impedance, R_s, is applied at the input. The output is terminated in a load resis-

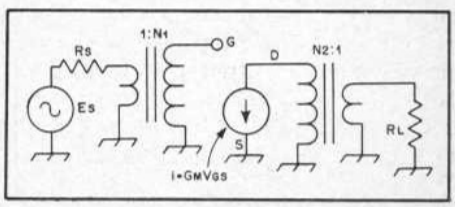


Fig. 8 – Common-source FET amplifier with transformer coupling at the input and output.

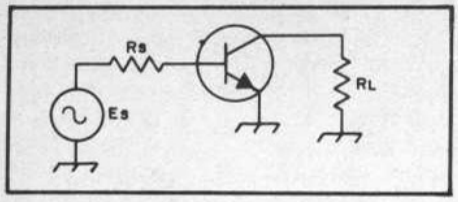


Fig. 9 – Bipolar transistor amplifier.

tance, R_L. While the open-circuit voltage of the generator may be known, the input voltage or current of the amplifier is not specified. It will depend upon the impedance presented to the generator. Similarly, the output voltage presented to the load is not known from inspection. All of these parameters will depend upon the characteristics of the amplifier.

In Fig. 10 a more generalized representation of the amplifier is shown. Here, the transistor symbol has been replaced with a box. This could represent a transistor, an FET, a tube, or even a passive circuit such as a filter or attenuator. There are three terminals to most active devices of interest. However, one is common to both the input and the output. This leads to an input terminal pair and a similar one at the output. Each pair of terminals is called a port.

Shown at each port in the amplifier (or whatever) is a current and a voltage. The current is shown entering each port, although this is merely a convention. Also, voltage polarity conventions are shown, associated with the direction of current flow. There are a total of four variables, V_{in}, I_{in}, V_{out} and I_{out}. More frequently, the ports are numbered, leading to subscripted variables, V_1, I_1, V_2 and I_2.

Any two of the variables associated with the two-port network may be chosen as independent. The remaining pair of variables is then expressed as functions of the independent variables.

Assume that the two voltages, V_1 and V_2, are independent variables. The two currents are then written in terms of these voltages.

$$I_1 = Y_{11}V_1 + Y_{12}V_2 \qquad \text{(Eq. 12)}$$

$$I_2 = Y_{21}V_1 + Y_{22}V_2 \qquad \text{(Eq. 13)}$$

That is, the input current is one constant, Y_{11}, multiplied by the input voltage plus a second constant, Y_{12}, times V_2. The output current is similarly expressed. This set of parameters, Y_{jk}, is called the admittance parameters. The fact that the independent variables (V_1 and V_2) are linear implies that only the small-signal behavior of the amplifier is of interest. High-order terms are not considered.

The previous two equations may be rewritten in a different format.

$$\begin{pmatrix} I_1 \\ I_2 \end{pmatrix} = \begin{pmatrix} Y_{11} & Y_{12} \\ Y_{21} & Y_{22} \end{pmatrix} \begin{pmatrix} V_1 \\ V_2 \end{pmatrix} \qquad \text{(Eq. 14)}$$

This is a matrix relationship. The voltages are a set of ordered numbers (a vector) as are the currents. The two sets are related with the Y matrix which is an ordered array of numbers. In circuit analysis, the elements of the matrix are usually complex, containing both real and imaginary components. The matrix relationship (Eq. 14) is nothing more than a restatement of Eqs. 12 and 13.

Consider the Y parameter equations with regard to their physical meaning. Assume first that V_2, the output voltage, is set to zero. That is, the output of the amplifier is short circuited. In this condition, $I_1 = Y_{11}V_1$ and $I_2 = Y_{21}V_{11}$. Equations of this kind are familiar to us. Y_{11} is just an admittance, relating input current to input voltage. Y_{21} is a transconductance (actually, here a transadmittance) relating the output current to the input voltage. A measurement of the three variables left with V_2 set to zero will allow experimental determination of Y_{11} and Y_{21}. Y_{12} and Y_{22} are similarly evaluated by setting V_1 to zero. For this reason, the Y parameters are often referred to as *short-circuit admittance parameters*. As an exercise, the reader may wish to consider some of the models used for transistors and FETs earlier and to calculate the resulting Y parameters.

In much of the engineering literature, the admittance parameters are expressed in a slightly different form. Numbered subscripts are replaced with ones with letters which have a physical significance. This is presented in the following example.

$$\begin{pmatrix} I_{in} \\ I_{out} \end{pmatrix} = \begin{pmatrix} Y_{ie} & Y_{re} \\ Y_{fe} & Y_{oe} \end{pmatrix} \begin{pmatrix} V_{in} \\ V_{out} \end{pmatrix} \qquad \text{(Eq. 15)}$$

In this representation, the "e" shown in the letter subscripts implies that the Y parameters are for a common emitter amplifier. Y_{ie} is an input admittance. It is the input admittance of the amplifier for the special case where the output is short-circuited. Y_{oe} is similarly defined for the output. Y_{fe} is the forward (common-emitter) transadmittance. This is the dominant parameter that determines the gain of the circuit. Y_{re} is a reverse transadmittance. It tells us the current flowing in the input when a signal is applied to the output port. In most of the simple models that we have analyzed, this term has been assumed to be zero. The one exception is the hybrid-pi model of a bipolar transistor. The presence of a collector-to-base

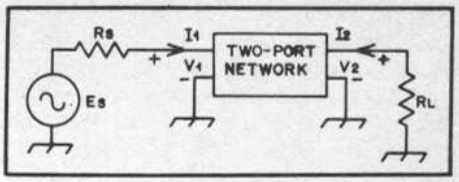


Fig. 10 – Generalized presentation of an amplifier.

capacitance would give rise to a Y_{re} parameter. If the reverse parameter is zero, the amplifier is said to be *unilateral*.

There are several other parameter sets that may be used besides the *Ys*. If the two currents are chosen as independent variables, the result is the Z matrix.

$$\begin{pmatrix} V_1 \\ V_2 \end{pmatrix} = \begin{pmatrix} Z_{11} & Z_{12} \\ Z_{21} & Z_{22} \end{pmatrix} \begin{pmatrix} I_1 \\ I_2 \end{pmatrix} \qquad \text{(Eq. 16)}$$

The Z parameters are evaluated experimentally by allowing each of the independent variables (now currents) to be zero. This is realized with an open circuit. For this reason, the Z parameters are called the *open-circuit impedance parameters.*

A third popular set of two-port parameters are the so-called hybrid, or "h" parameters. They are obtained by allowing I_1 and V_2 to be the independent variables. This is shown in matrix form as

$$\begin{pmatrix} V_1 \\ I_2 \end{pmatrix} = \begin{pmatrix} h_{11} & h_{12} \\ h_{21} & h_{22} \end{pmatrix} \begin{pmatrix} I_1 \\ V_2 \end{pmatrix} \qquad \text{(Eq. 17)}$$

Note that h_{21} is equivalent to beta for a common-emitter transistor amplifier.

Another set of parameters is called the *ABCD* matrix. For this set of parameters the direction of the output current is defined with an opposite sense from that used for the preceding sets. The equation and the defining network are shown in Fig. 11. The virtue of the ABCD matrix is that a number of stages with known parameters may be cascaded easily. The ABCD matrix of the cascaded equivalent network is obtained by matrix multiplication of the individual ABCD matrices. This form is especially useful for evaluation of the transfer function of ladder networks and the like.

If the details of matrix algebra are invoked, there are a number of other sets of parameters that may be generated. For example, with reference to Fig. 10 where the input and output voltages and currents are defined, one might define a new set of variables.

$$\begin{aligned} M_1 &= V_1 + I_1 \\ M_2 &= V_2 + I_2 \\ N_1 &= V_1 - I_1 \\ N_2 &= V_2 - I_2 \end{aligned} \qquad \text{(Eq. 18)}$$

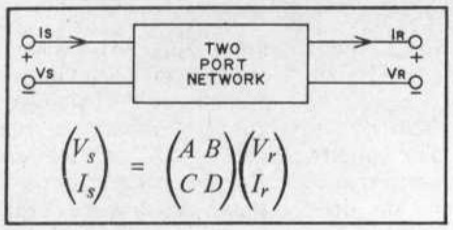


Fig. 11 – Alternative presentation of a two-port network. Note the direction of the output current.

Here, M_1 is a linear combination of I_1 and V_1. M_2, N_1 and N_2 are similarly defined. The M terms are assumed to be the independent variables while the N ones are dependent. This set of variables has no special physical significance. It does show how other families may be assembled, though. The requirement for a set of variables is that they be linearly independent. In a simplistic sense this means that one variable cannot be equal to one of the others multiplied by a constant.

Shown in Fig. 12 are the defining equations and a network diagram for S or scattering parameters. Unlike the N and M parameters, defined merely as an illustration, scattering parameters have physical significance and are extremely useful. The independent variables, a_1 and a_2, may be interpreted as voltage waves incident on ports 1 and 2, respectively. The dependent variables, b_1 and b_2, are voltage waves emanating, or scattered, from the two ports. Z_o is the characteristic impedance of the system used to define the parameters (50 ohms is typical). The b and a variables are related to each other through the S or scattering matrix.

Scattering parameters are more easily measured than Y or Z parameters. Recall that the Y and Z parameters required either short or open circuits at the two ports for measurements. At high frequencies it is often difficult to obtain these conditions. Furthermore, transistors may oscillate under these mismatched conditions, making the measurements difficult if not completely impossible. On the other hand, scattering parameters are measured in a Z_o (usually 50-ohm) system

The basic variables have physical significance. For example, $|a_1|^2$ is the power available from the generator with impedance Z_o and $|b_2|^2$ is the power that would be delivered to a Z_o load. The brackets indicate that the absolute value of the variables is squared to determine the values.

The S parameters themselves have significance. S_{11} is the complex reflection coefficient that would be measured at the input port. S_{22} is similarly defined. The magnitudes of these parameters may be measured with the return-loss bridge and simple detectors described in chapter 7. $|S_{21}|^2$ is the transducer power gain of the amplifier in a 50-ohm (Z_o) system. Also, $|S_{12}|^2$ is the reverse transducer power gain.

Owing to the relative ease of measurement and the physical significance, scattering parameters are employed extensively in the design of microwave-transistor amplifiers. The instrumentation required for the measurements is very expensive, but indispensable for modern high-frequency engineering. S-parameter results are plotted directly on a Smith chart, adding to their utility.

Virtues of Small-Signal Two-Port Analysis

The discussion of two-port analysis presented may appear a bit formal. However, there are a number of calculations that may be performed with the parameters that will enable very complete analysis and design to be done. Some of these will be outlined below, with Y parameters generally used in the examples. Good references are the book by Carson mentioned earlier, *Hewlett-Packard Applications Note 95* and *Motorola Applications Note AN-215A*.

While, in principle, one set of parameters is as complete as any other, some operations are more easily performed with certain parameter sets. As mentioned previously, the cascading of networks is easily analyzed with ABCD matrices. If a shunt feedback element is to be added to an existing network, Y parameters are the most convenient. In this case, the Y matrix for the transistor is added directly to the Y matrix of the feedback element. When series feedback (emitter resistance or reactance) is applied to a stage, it is most easily done with Z matrices. Most measurements are best done with respect to S parameters.

Through the use of the appropriate matrix transformations, any set of parameters describing a two-port can be converted to any other form. This allows complex circuits to be analyzed with relative ease.

If common emitter parameters are available for a given transistor, they can be converted through appropriate transformations to provide data for other circuit configurations. For example, the common-base and common-collector Y parameters are available from common-emitter Y parameters.

Two-port analysis is highly generalized. Although we have considered its application to transistor amplifiers, it may also be used to describe tube or FET amplifiers. The concepts may be extended to more than two ports. A common three-port device is a mixer – it may be analyzed using the concepts. A single-port device of interest might be a reflection amplifier using a tunnel diode or a Gunn-effect diode. Other N port circuits that are often considered

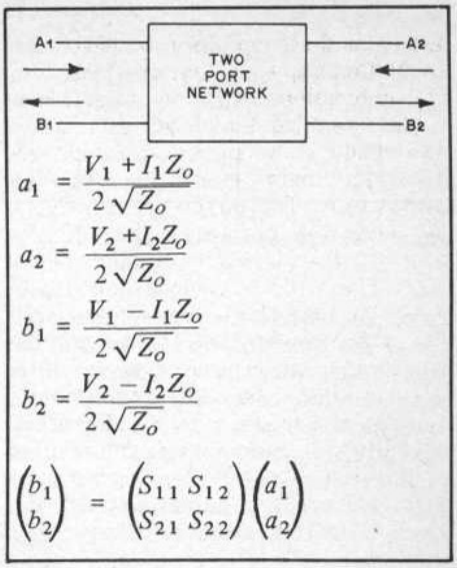


Fig. 12 – Two-port network representation used for the definition of scattering or "S" parameters.

are diplexers and circulators.

Once the two-port parameters of a circuit are known, all of the various gains are defined. For example, the transducer gain is given by

$$G_T = \frac{4Re(Y_s)Re(Y_L)|Y_{21}|^2}{|(Y_{11} + Y_s)(Y_{22} + Y_L) - Y_{12}Y_{21}|^2} \quad \text{(Eq. 19)}$$

where Re means "real part of" and brackets imply the magnitude of the included expression. Note that the transducer gain is a function of both the source and load admittances. Other gains that are available include the MAG and power gain. The source and load admittances required to achieve the maximum available gain are calculated readily.

For a given load admittance, the input admittance of an amplifier may be calculated. Note that this differs from Y_{11}, which was the input admittance when the output was short-circuited. The output admittance may be similarly calculated. These are shown as

$$Y_{in} = Y_{11} - \frac{Y_{12}Y_{21}}{Y_{22} + Y_L} \quad \text{(Eq. 20)}$$

$$Y_{out} = Y_{22} - \frac{Y_{12}Y_{21}}{Y_{11} + Y_s} \quad \text{(Eq. 21)}$$

In an earlier discussion of the hybrid-pi model of a bipolar transistor, we demonstrated that certain types of reactive feedback would lead to an input impedance with a negative real part. If the Y parameters for this circuit were calculated, Eqs. 20 and 21 could be applied and the input admittance, $Y_{in} = G_{in} + jB_{in}$, could be evaluated. G_{in} would be negative. Under other conditions $Re(Y_{out})$ may be negative.

Either of these conditions could lead to oscillation for some terminations. General expressions exist for evaluation of C, the so-called Linvill stability factor. Essentially, a two-port device is allowed to be terminated in any and all passive impedances. The number that results is an indication of the stability. If C is between 0 and 1, the amplifier is unconditionally stable. This means that Y_{in} or Y_{out} will never have negative real parts for any terminations of positive resistances. Oscillation is not possible without additional feedback. If the stability factor, C, is outside of the unconditionally stable region, the circuit may or may not oscillate. It will depend upon the value of the terminations. Other stability factors (such as the Stern factor) take finite terminations into account.

Stability analysis can be of profound value, as any experimenter who has fought with an oscillating amplifier will attest. Care should be used in applying the analysis though. For example, if a 50-MHz amplifier were being analyzed, only the 50-MHz Y parameters would be needed for gain and matching calculations. However, stability should be evaluated over the total frequency range where the transistor remains active. This allows evaluation of the stability at frequencies outside of the operating band.

Two-port parameters are specified by the manufacturers for most high-frequency transistors and FETs. In the vhf region, Y parameters are usually given. For uhf (and higher) applications, S parameters are becoming universal. Usually, detailed specifications are not given for frequencies below 50 or 100 MHz. The reason is that the available low-frequency gain of a transistor with a 1 or 2 GHz f_T is so high that external "strays" will dominate circuit behavior. In these applications, feedback is usually employed to obtain well-defined performance. If detailed calculations are to be done, simple models like the hybrid pi are usually adequate. Y parameters for the hybrid-pi model are calculated easily using the methods outlined by Carson.

Probably the greatest single asset of two-port network theory is that it leads to sophistication and economy in describing circuit behavior. The generality of the methods make them applicable to virtually any active device. The problems of making a transition from "tube thinking" to "transistor thinking" that have discouraged many amateurs from experimenting with solid-state circuits disappear.

Appendix 5

Inductance of Toroidal Coils

The inductance of a toroid of N turns is given by $L = KN^2$ where K is a proportionality constant characteristic of the core. The value of K will depend on both the nature of the core material and the physical size. Values of K for a number of popular powdered-iron cores are given in Table 1.

As an example, the T50-2 core has $K = 5\ nHt^{-2}$ (nanoHenrys per turn squared). The inductance of a 25-turn winding on this core would be $L = 5 \times (25)^2$ nanoHenrys = 3125 nH = 3.125 μH. All of the data in Table 1 was abstracted from the catalog of Amidon Associates.

Table 1

CORE TYPE	*K, nHt^{-2}*	*USEFUL FREQUENCY RANGE*	*CORE TYPE*	*K,nHt^{-2}*	*USEFUL FREQUENCY RANGE*
T30-2	4.3	0.5-30 MHz	T25-6	2.7	3-250 MHz
T50-2	5.0	"	T37-6	3.0	"
T68-2	5.7	"	T50-6	4.0	"
T80-2	5.5	"	T68-6	4.7	"
T94-2	8.4	"	T80-6	4.5	3-100 MHz
T106-2	13.5	"	T94-6	7.0	"
T130-2	11.0	"	T106-6	11.6	"
T184-2	24.0	"	T184-6	19.5	"
T200-2	12.0	"	T200-6	10.5	"

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Solid State Design for the Radio Amateur is among the select few technical books that have sold more than 50,000 copies. Why has it achieved this enviable sales milestone? For one thing, it's chock full of good, basic information—circuit designs and their applications, and descriptions of receivers, transmitters, power supplies and test equipment. Much of the data, such as transistor modeling, cannot be found in other publications. **Solid State Design for the Radio Amateur** should be considered an essential publication for every technical library.

