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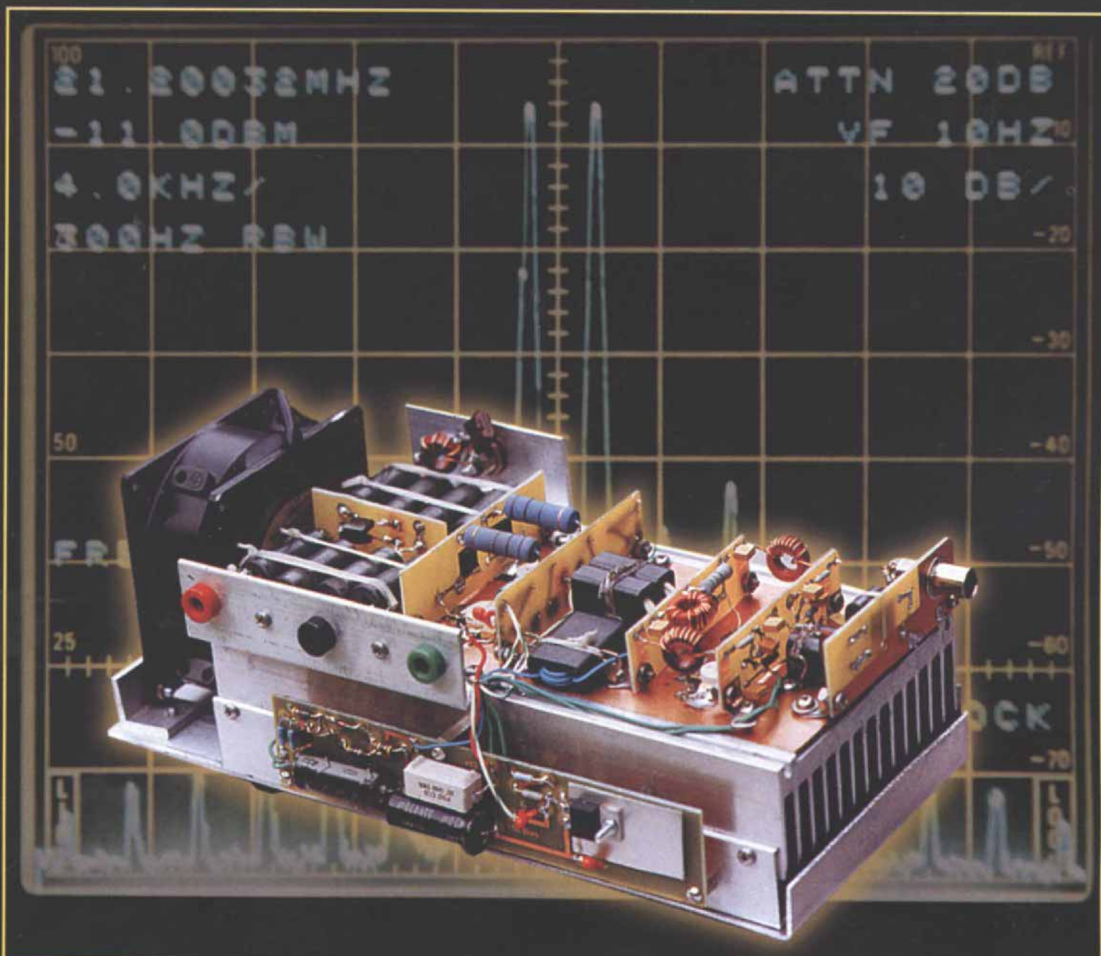
November/December 1999

Forum for Communications Experimenters



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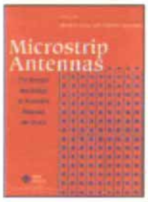

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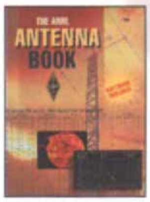

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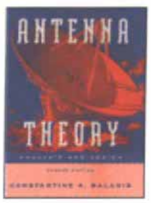

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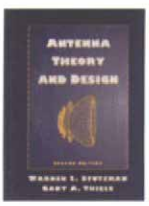

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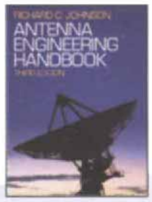

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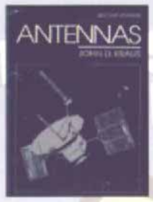

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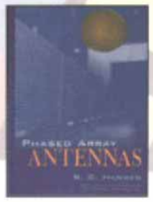

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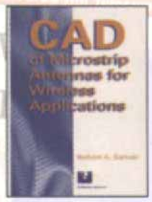

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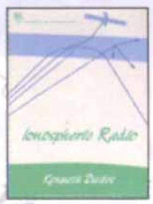

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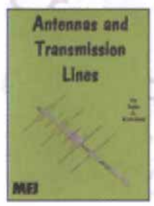

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QEX

QEX (ISSN: 0886-8093) is published bimonthly in January '99, March '99, May '99, July '99, September '99, and November '99 by the American Radio Relay League, 225 Main Street, Newington CT 06111-1494. Subscription rate for 6 issues to ARRL members is $22; nonmembers $34. Other rates are listed below. Periodicals postage paid at Hartford, CT and at additional mailing offices.

POSTMASTER: Form 3579 requested. Send address changes to: QEX, 225 Main St, Newington, CT 06111-1494

Issue No 197

Offices

225 Main St, Newington, CT 06111-1494 USA
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Subscription rate for 6 issues:

In the US: ARRL Member $22, nonmember $34;

US, Canada and Mexico by First Class Mail: ARRL member $35, nonmember $47;

Elsewhere by Surface Mail (4-8 week delivery): ARRL member $27, nonmember $39;

Elsewhere by Airmail: ARRL member $55, nonmember $67.

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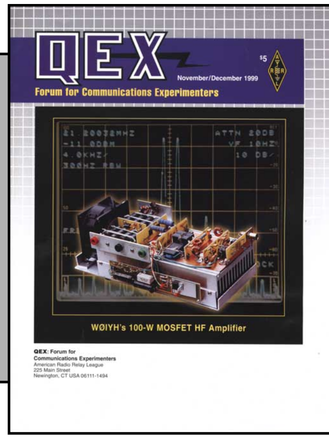

About the Cover

Bill Sabin's 100-W PA.
See the article on p 31.

Features

Columns

Nov/Dec 1999 *QEX* Advertising Index

The purpose of *QEX* is to:

1) provide a medium for the exchange of ideas and information among Amateur Radio experimenters,
2) document advanced technical work in the Amateur Radio field, and
3) support efforts to advance the state of the Amateur Radio art.

All correspondence concerning *QEX* should be addressed to the American Radio Relay League, 225 Main Street, Newington, CT 06111 USA. Envelopes containing manuscripts and letters for publication in *QEX* should be marked Editor, *QEX*.

Both theoretical and practical technical articles are welcomed. Manuscripts should be submitted on IBM or Mac format 3.5-inch diskette in word-processor format, if possible. We can redraw any figures as long as their content is clear. Photos should be glossy, color or black-and-white prints of at least the size they are to appear in *QEX*. Further information for authors can be found on the Web at **www.arrl.org/qex/** or by e-mail to **qex@arrl.org**.

Empirically Speaking

The ARRL Technology Task Force (TTF) and Technology Working Group (TWG) are open for business. As reported here in May, the League's Board of Directors charged these committees with identifying and evaluating promising technologies for Amateur Radio. ARRL President Rod Stafford, W6ROD, appointed leading amateurs to serve on the panels.

The TTF comprises members of the ARRL Board family and HQ staff. Chairing is ARRL First Vice President Steve Mendelsohn, W2ML. Other TTF members are Roanoke Vice Director Dennis Bodson, W4PWF; Hudson Director Frank Fallon, N2FF; New England Director Tom Frenaye, K1KI; Southwestern Vice Director Art Goddard, W6XD; Pacific Director-elect and current Vice Director Jim Maxwell, W6CF; ARRL International Affairs Vice President Larry Price, W4RA; and Rocky Mountain Director Walt Stinson, W0CP.

The TTF will receive evaluations of technical proposals from the TWG, then make recommendations to the Board. TWG members represent a broad range of interests and expertise. *CQ* Editor Rich Moseson, W2VU, is the chairperson. Others include AMSAT-NA President Keith Baker, KB1SF; Peter Coffee, AC6EN; Mike Cook, AF9Y; Gene McGahey, NR0NR; ARRL Technical Relations Manager Paul Rinaldo, W4RI; Dennis Silage, K3DS; and yours truly. ARRL Laboratory Supervisor Ed Hare, W1RFI, serves as HQ staff liaison to both committees.

We're inviting you to submit ideas through the TTF Web page, **www.arrl.org/news/ttf/**; by e-mail to **ttfinput@arrl.org**; or by Pony Express to *ARRL Technology Task Force, c/o Ed Hare, W1RFI, 225 Main St, Newington, CT, 06111*. We're counting on you, the technically savvy readers of *QEX*, to provide input to the TTF that will help shape League policy into the next decade. Please submit by November 30. Elsewhere in this issue, I have some thoughts on technology and the future.

Some possibilities for future systems arose when the FCC relaxed the spread spectrum (SS) rules for Amateur Radio in their Report and Order of WT Docket 97-12. Effective November 1, the new rules allow stations to transmit SS emissions other than frequency hopping and direct sequence. Limitations on spreading codes are removed. SS contacts are allowed between US hams and DX in countries that permit SS. Record-keeping requirements have also been eased.

The compromise made in obtaining our new freedom is that stations using between 1 W and the maximum 100 W must employ automatic transmit-power control. Since the criterion is SNR, it will be interesting to see what means we can devise to comply. Part-15 wireless SS modems and chip sets for the ISM bands are now widely available; these bands nearly coincide with our 33-, 13- and 5-cm bands. At 1 W maximum, ISM modems already satisfy the power rules, but modification is needed to provide identification as required in §97.119. The Report and Order may be viewed at **www.arrl.org/announce/regulatory/wt97-12**.

In This *QEX*

Harke Smits, PA0HRK, presents his design of a noise/gain analyzer. It's nice to measure the noise performance of your actual circuit rather than extrapolating it from data sheets. Also of interest to UHF-and-above enthusiasts is John Stephensen, KD6OZH's crystal oscillator. Improve your dynamic range and get more contacts with this low-noise unit.

Bruce Pontius, N0ADL, brings his talent to bear on the elimination of wasted current in signal generators. Naturally, the work is equally applicable to LO synthesis.

Bill Sabin, W0IYH, contributes the crowning segment of his HF PA design—the MOSFET amplifier itself. This one ought to motivate quite a few to get out the tools and start building.

We continue to follow the non-synthesized approach of Mark Mandelkern, K5AM, in the third part of his series. This time, it's the RF board and all second-conversion functions.

I was compelled to do some thinking about double-balanced mixers, so I offer a bit of analysis. Zack Lau, W1VT, explores modeling of capacitors with *ARRL Radio Designer*. *—73, Doug Smith, KF6DX,* **kf6dx@arrl.org**.

Technology and the Future of Amateur Radio

Observations, suggestions and opinions about technical advancement

By *QEX* Editor Doug Smith, KF6DX

It has been said that change is the only constant in life; certainly Amateur Radio is no exception. What changes should we make? Which should we shun? Specifically, how best can technology enhance the enjoyment and use of our resources? What recommendations should the Technology Task Force (TTF) make? How can we identify technical ideas that should be considered? As we look to the future, our charter principles and "reasons for being" serve as useful guides. I paraphrase them for emphasis here:[1]

[1]Notes appear on page 4.

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(a) Recognition and enhancement of volunteer, public-service communications, especially during emergencies or disasters
(b) Extension of technical development and testing
(c) Emphasis on education and encouragement of learning
(d) Providing a resource of communicators, technicians and engineers
(e) Promotion of international social contact on a personal level

I'd like to concentrate on paragraphs b, c and d. A complex interrelation seems to exist that needs some discussion.

Think Creatively

Many of the best inventions elegantly solve some persistent problem. Yet, some don't solve a problem, but better our lot nonetheless. How many of us knew we needed a VCR before they became popular? So let's ask two questions, first "What are the most pressing challenges facing Amateur Radio?" then dream of "What else could be?"

It's not difficult to name a few challenges: crowding on our bands, our need to coexist with other services and our desire to bring in new talent. I'm sure we can come up with others. Technology can provide some solutions, but others seem to be purely sociological.

What technical solutions can help? Which ones aid Amateur Radio growth and make it more fun? PSK31 is an excellent illustration: It requires little occupied bandwidth (reduces QRM) and has generated fresh excitement in the RTTY world. CW can occupy even less bandwidth; this mode still allows maximum spectral occupancy. Extremely narrow DSP filtering and noise-reduction techniques boost the potential of both these modes still further.

The union of Amateur Radio with other communications media has been quite productive. APRS, HF e-mail and DX-cluster services stand as proof. I'm confident we can discover other fruitful combinations.

QEX contributors have created powerful additions to our legacy of innovation in Amateur Radio. I am certain that this will not wane. Someday I'll look back and say: "By Jove, those folks were on the ball!"

Change the Rules?

The laws governing the Amateur Radio Service provide the framework within which we operate, but the rules can be changed. The ARRL has steadfastly led the way to many sensible revisions over the years. Your input can influence the development of policy that encourages the good operational and technical skills we need to further Amateur Radio.

While not all of Part 97 deals with technical matters, you may identify areas with potential to change for the better. Look at the many recent, significant proposals put forth for examples.[2] Do you see anything hindering technical progress?

Given the accelerating pace of technology, what will Amateur Radio look like in 5, 10 or even 50 years? It is difficult to gaze into a crystal ball that deeply, but now is the time to step back and look at the complete picture.

Education

Rapid and continual technical advancement requires a concomitant and equal increase in education. DSP is an example: It is nearly everywhere in Amateur Radio these days, but how many of us really understand it? Will we be ready to learn something from the future success of digital-audio-broadcasting, auditory-psychophysics or image-reconstruction technologies?[3]

At Amateur Radio shows I'm repeatedly amazed by the many very young people who walk right up to computer-controlled transceivers and begin operating with no prior instruction. An entire generation is growing up immersed in knowledge of computing. Surely, we can draw on this to our benefit.

Information technology has its own acronym these days (IT) and encompasses ever-growing realms of inquiry and transformation, each having widespread impact on our society. The ARRL Web site and various Internet news groups have, in my opinion, enhanced the Amateur Radio experience. What else can IT contribute to Amateur Radio?

Technical education is only part of the picture. Let's find ways to communicate the *fun* of Amateur Radio. Make sure folks know how much good it does, and that it is continually satisfying its "basis and purpose." Those who spread good information about Amateur Radio help threefold: They enrich public relations and general goodwill toward the hobby. They add to the remarkable heritage of hams worldwide as volunteer servants. They fulfill our charter of increasing new operational and technical talent by stimulating interest.

Build Motivation

Many of us have mourned the dwindling technical content of other Amateur Radio magazines. I believe it is, in general, a reflection of our changing interests. Can we reverse the trend? We can't force new interest in technical topics; that interest comes from within. We can work to motivate our community. What motivates most hams? Competition, operational and technical challenges, social interaction and public service are some answers.

Consider that we enhance our pleasure and understanding by finding and fortifying the threads that unite us. An example is the growing interest in high-quality HF SSB audio. This pursuit of more pleasurable listening emphasizes technical fundamentals and encourages in-depth understanding of transceiver-design elements. By prizing low-distortion signals, it fosters interference reduction. The rules about good engineering practice and occupied bandwidth are part of the goal. This is just one example; be quixotic—find new ones.

When many of us started in electronics, things were simpler. It was feasible to study and master most of the basic concepts. Over the last three or four decades, dozens of specialized disciplines have developed; mastery of them all is daunting. The first big split began with the development of microprocessors in the early '70s. Now even digital territory is divided into several distinct specialties.

Has the diversity of our interests has led to a decline of our focus? Perhaps, but we can choose to let our diversity bring us together. Personally, I can't say what Amateur Radio will look like in 50 years, but I know hams will be active because the hobby is fun and it's necessary. They will not lack new hills to climb. Please help inspire new climbers by telling them how beautiful the view is from up here.[4]

Notes

[1]FCC Part 97, §97.1-Basis and purpose. FCC Part 97, the Amateur Radio Service regulations is available on the ARRL Web site at **http://www.arrl.org/field/regulations/news/part97/** or the US government site at **http://www.access.gpo.gov/nara/cfr/waisidx_98/47cfrv5_98.html**.

[2]For information on recently proposed rules changes, visit the ARRL Web site at **www.arrl.org/announce/regulatory/**.

[3]Further reading on these topics can be found in *The Digital Signal Processing Handbook*, ed. V. K. Madisetti and D. B. Williams (Boca Raton: CRC Press LLC, 1998; ISBN 0-8493-8572-5).

[4]TTF Web page: **www.arrl.org/news/ttf/**.

A Noise/Gain Analyzer

This home-built test instrument simultaneously measures system noise and gain. It uses an Analog Devices log-converter chip with 90 dB dynamic range up to 500 MHz.

By Harke Smits, PA0HRK

If you work with low-noise amplifiers (LNAs), you need a noise/gain analyzer. I wanted one for years, but the one I built a long time ago was not good enough and suffered from several drawbacks. Just when I finally decided to buy one—a surplus AIL 7380 for about US $750—I read an article by Luis Cupido.[1] I liked the idea very much, but disapproved of using the vertical amplifier of a spectrum analyzer to get a log-converted signal. At the same time, I learned about the AD8307, a log converter with a frequency range of at least 500 MHz and about 90 dB of dynamic range. That was it! I was able to make my own noise/gain analyzer of high quality. Here, I describe my homebrew noise/gain analyzer, how it works, how to build and operate it. With this project, you can measure the noise figure and gain of your device under test (DUT) simultaneously (and in real time!) at a fraction of a commercial unit's cost.

[1]Notes appear on page 10.

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How It Works

In a lab environment, it's common to measure noise figure by switching a noise source at the input of the device under test (DUT) on and off, then calculate the noise figure from the measured RF output power in both cases. One may use an ordinary bolometer to measure the RF power and a hand calculator to calculate the noise figure—quite slow, cumbersome and error-prone. In the analyzer, we use a log converter IC (the AD8307 from Analog Devices) to convert the RF signal into "power" signals of 100 mV/dB. The two resulting signals (noise source on and off) are stored in two sample-and-hold circuits and fed through some op amps to calculate the noise figure and gain. These values are displayed on two LCD displays. The noise figure—as indicated by the instrument in real time—is the composite system noise figure. You will need an adaptation of the Friis formula (Eq 4) to calculate the device noise figure.

The Circuits

The noise/gain analyzer consists of three blocks: the RF-input section with log converter, the signal processor with displays and the power supply (dc voltages: +12 V, –9 V and +28.0 V, switched). Please see Fig 1.

The RF-input section contains a low-pass filter, a MMIC amplifier, some switched filters followed by the AD8307 log converter and an op amp to normalize the output to 100 mV/dB.

The low-pass filter at the input (PLP 550 from Mini Circuits) prevents any unwanted DUT signals from reaching the amplifier. In some transverters, considerable leak-through of LO signals causes distortion or even saturation in the input amplifier. This amplifier is a standard MMIC (eg, ERA-5 or MAV-11) with a gain of about 13 to 20 dB. Its output compression point must be +17 dBm or more. I found that the ERA-5 amplifier can show unstable behavior in combination with the low-pass input filter, so I use a ferrite bead in series with its input-coupling capacitor. In my ham practice, I normally use IFs of 30, 150 and 432 MHz. I use helical band-pass filters for the latter two frequencies and I have inserted a low-pass filter (PLP 150) to cover both 144 and 30 MHz. All filters are switched with miniature relays (Fujitsu 21006, Conrad or equivalent).

To gain experience with the instrument, I advise that you first construct a simple input section at your favorite IF. Start with an input filter (eg, two-section helical or low-pass), then the MMIC followed by a second filter and so on. Once familiar with this instrument, you can build a new RF-input section with switchable filters to cover more IFs. Beware: Helical filters tend to have spurious resonances at higher frequencies. Consider using a low-pass filter at the input.

The circuit around the AD8307 closely follows the datasheet.[2] It has a high input impedance, so we need a 51-Ω resistor across its input to match the filters. Use an additional 78L05 to supply voltage for the AD8307 properly. The LM158 (or 358) is used here with a single supply (+12 V, only this voltage is fed to the RF input section). I selected it because it can use input signals very near ground or the supply rail. This section is only sketched because I suppose that if you understand the workings of, and the need for, a noise/gain analyzer, you most likely also know how to build the RF-input part.

The output of the RF section is fed to a sample-and-hold circuit (a '4066 with output capacitors C2 and C3) to get steady signals for both the on and off states. (Refer to Fig 2.) The basic task of the subsequent op-amp circuit is to calculate the noise and gain values. The formula is:

$$NF_{\mathrm{dB}} = \log\left(\frac{ENR}{(Y-1)}\right) \quad \text{(Eq 1)}$$

where

$$Y = \frac{P_{\mathrm{hot}}}{P_{\mathrm{cold}}} \quad \text{(Eq 2)}$$

ENR = Excess noise ratio: The noise power of the noise source (on) divided by the noise generated by a 50-Ω resistor at room temperature (off). Normally, the *ENR* values are frequency-dependent and stated on professional noise heads after laborious calibration procedures. We, as amateurs, also need to know—as precisely as possible—the *ENR* value (in decibels) of our head at the frequency of interest.

P_{hot} = measured output power of the DUT with the noise source on.

P_{cold} = measured output power of DUT with the noise source off.

The bad news is that this instrument can only calculate NF_{dB} as:

$$NF_{\mathrm{dB}} = \log\left(\frac{ENR}{Y}\right) = \log(ENR) - \log(Y) \quad \text{(Eq 3)}$$

So, an error is introduced. This result is obtained by mere subtraction of the P_{hot} and P_{cold} voltages, as produced by the RF section. The good news is that the error is small, especially for small values of *NF*, as shown in Table 1.

The noise figure of the device under test (NF_{DUT}) can be calculated with an adaptation of the Friis formula:[3]

$$NF_{DUT} = NF - \frac{F_{NMI} - 1}{G_{DUT}} \quad \text{(Eq 4)}$$

Where G_{DUT} is the gain of the device under test; notice *NF* is no longer in dB.

All signal levels in the op-amp section are normalized to 100 mV/dB, so the *ENR* value is simply generated with a potentiometer and a buffer op amp. To indicate whether the cold signal level is high enough, I made a small indication circuit with an LED. The LED (green) lights if the cold value is higher than 1.3 V. Normally during a measurement, the LED must be on. I use two LCD digital panel meters to display the measured gain and noise values. They have a range of 20 V, and the decimal point is moved one place to the right. I use two toggle switches to switch the input of the meters between gain/hot values and between the noise-figure/ENR set. Further details can be found in *VHF Communications*, and I advise reading the original article before heating up the soldering iron.

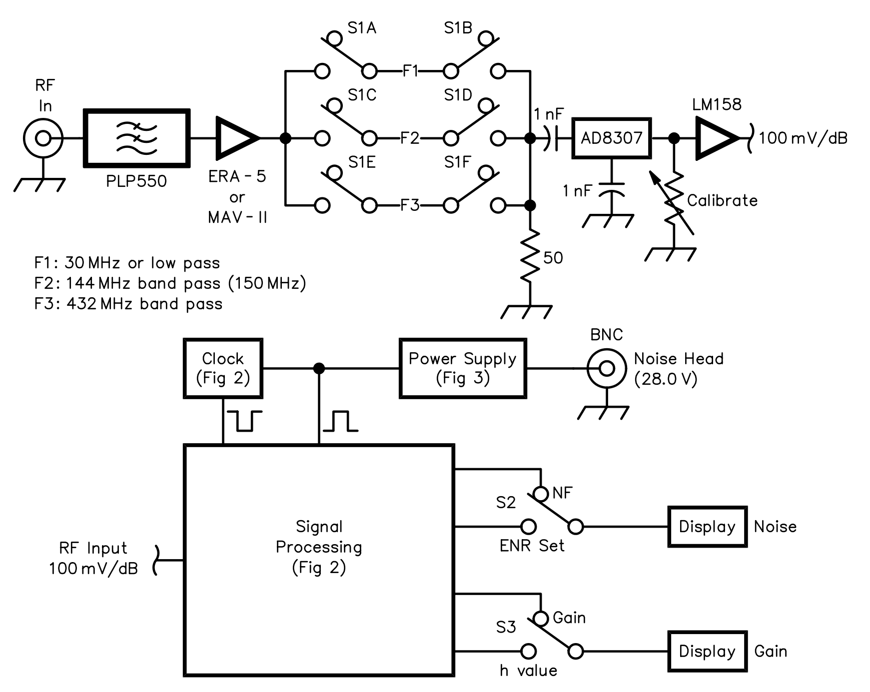

Fig 1—Noise/gain analyzer block diagrams.

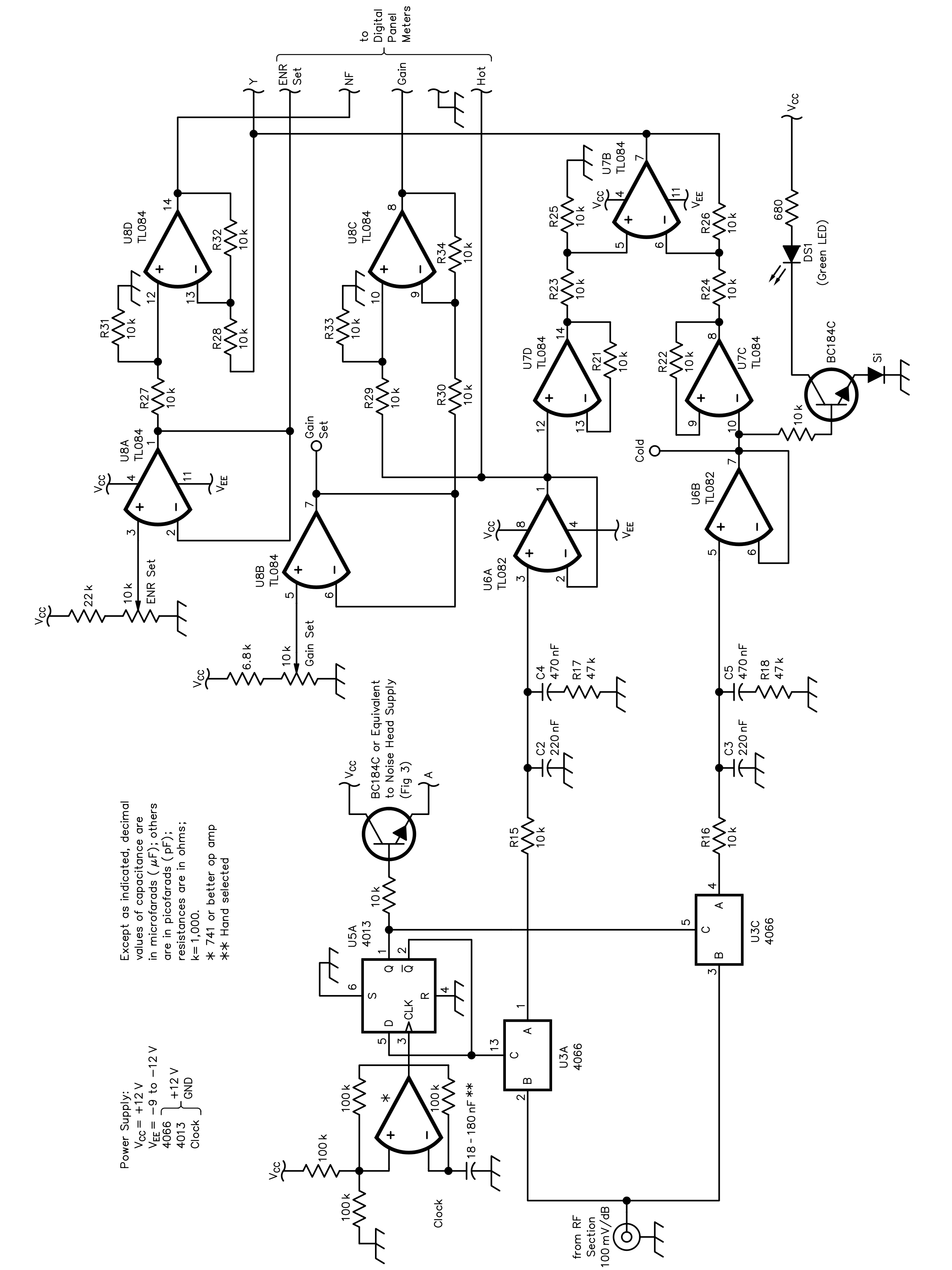

Fig 2—Signal-processing section schematic diagram.

The fixed power supplies are made with three-terminal regulators and need no further clarification. The 28-V supply (refer to Fig 3) is built around a good-old classic IC: the 723. Yes, dear readers, this IC is so old that many of you did not even exist when the it was designed in the late sixties. The transistor (2N3053 or equivalent) at the input simply reduces the voltage to prevent exceeding the 723's limit. The supply circuit itself is a textbook standard. The 723 is switched by a transistor driven by an opto-coupler. This prevents ground loops, which might impair the instrument's performance.

To increase versatility, I inserted a three-position switch: normal operation, off (to change heads for instance) and continuous on. In normal operation, the noise head is driven with a pulsed signal that switches the source on and off at a rate of 10 to 100 Hz. This signal is derived from a clock (op-amp oscillator) followed by a D flip-flop. Both outputs are fed to the sample-and-hold circuit to obtain steady hot and cold signals; one output is also used to switch the 28-V supply to the head.

How to Build It

Well, to start with, there are no printed circuit boards. After all, this is not a beginner's project. To build it, you need at least a good working knowledge of digital circuits, power supplies, op amps, RF techniques and noise.

All power-supply circuits are built on Veroboard; the RF and op-amp circuits are built on double-clad epoxy boards. The RF circuit board is soldered into a tinned box with SMC connectors and soldered feed-through capacitors to the outside world. In my instrument, the RF section comprises all circuitry from the input filter up to and including the LM358.

I used no sockets. IC leads are bent 90° upward, except for those that are ground connections, which have the small end section cut off and are then soldered directly to the ground foil. Mount ICs with the markings on top. Use small 100-nF caps at each power-supply terminal for decoupling and additional support. Use only metal-film resistors; buy a hundred 10-kΩ units, you'll need most of them. Use wire-wound, multiturn pots for the *ENR* and gain adjustments and a good multiturn trimmer at the calibration point near the LM358.

I normally build step-by-step with lots of intermediate function checks—it saves a lot of debugging time. Use a complete metal enclosure for the finished instrument. I built my first prototype, with one display, in about 10 evenings. The final version, with two displays in a neat cabinet, took much more time.

Alignment

There are only two alignment points: the 28-V supply for the noise head and the log converter. The 28-V supply needs to be adjusted to precisely 28.00 V. Adjust this noise-source supply with the head connected—after at least 30 minutes of warm-up time—using a good voltmeter.

Adjustment of the log converter is a bit more cumbersome. I use a precisely specified 40-dB N-type attenuator (eg, Narda model 777). You also need a signal generator with an accurately adjustable output level. I use my HP-8640B. Connect the signal generator to the RF section and set it to the highest input frequency (in my case, 432 MHz). Set the RF level so that the output of the RF section (output of LM358) reads 8.0 V. Insert the attenuator. Adjust the trimmer for 4.0 V. Remove the attenuator. Repeat this procedure several times until the difference exactly meets the specified attenuator value. In my case, 4.0 V for 40 dB (equals 100 mV/dB). Don't trust the signal generator's output attenuator: It's simply not accurate enough. Verify the input section's dynamic range by varying the input RF level; below 1 V output and above about 9 V, it will show nonlinearity. Take your favorite 1- or 10-dB step attenuator, step through the values and see what happens.

The Noise Source

Obviously, the noise source is the essential component. If you have great luck, you will find one looking for a second owner for a reasonable price. In general, this will not be the case, and you must make one yourself. I made a very simple one using the reverse biased base-emitter junction of an RF transistor. It can be constructed on a flanged connector (eg, BNC or N) in half an hour. It allows measurements up to 2.3 GHz when followed by a suitable attenuator (eg, 20 dB) to get an *ENR* value of about 15 dB. (Refer also to the article by DB2GM.[4]) Much better—and much more expensive—noise generators are also described in other ham literature.[5] Specific noise diodes are required for these high-performance noise generators. I use an AIL 7616 noise generator with an *ENR* value of 15.5 dB (specified from 1 to 12.4 GHz) but it operates well down to 100 MHz.

Actually, there is a tendency to use noise generators (eg, HP-346A) with lower *ENR* values (around 5 dB) to measure modern LNAs with noise figures well below 1 dB. It is of paramount

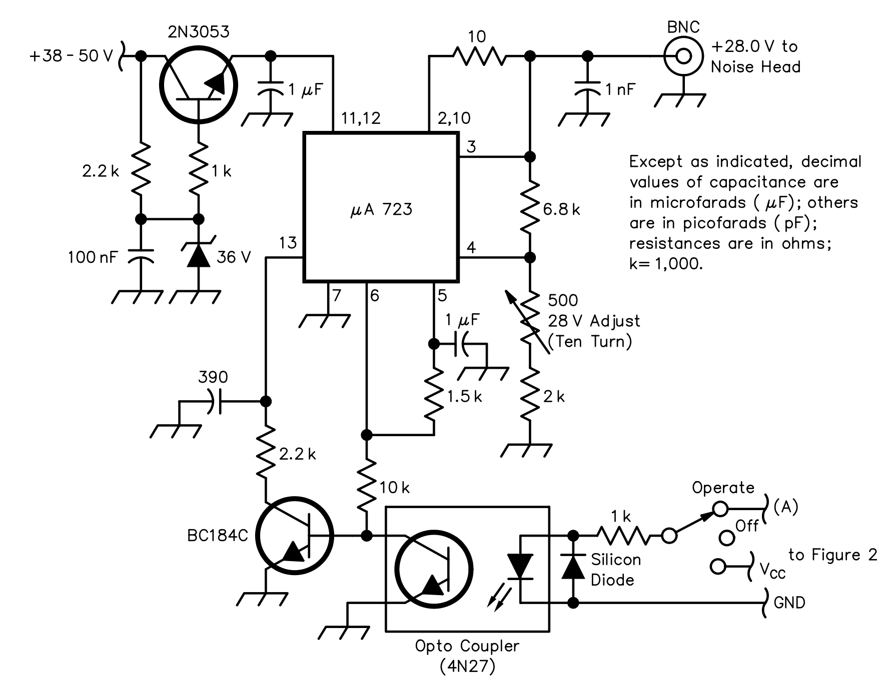

Fig 3—Noise-head power supply.

importance that the output impedance of the head—including attenuator—does not deviate from the desired 50-Ω value in either the hot or the cold state.

I also happen to have some high-*ENR* noise generators, originally designed for a BITE (Built-in Test Equipment) application. They produce about 35 dB *ENR* and need a good attenuator before they can be usefully applied. It's very interesting, however, that they produce a nice signal on 24 GHz without an attenuator! In that case, I use a waveguide isolator between noise head and DUT.

Some noise sources don't switch fast. In a normal measurement setup (noise source, DUT and instrument properly interconnected), check the rise time with an oscilloscope at the RF-section output. The rise time with my AIL head is less then 10 µs. If not this good, the switching speed of the head is too low; decrease the clock frequency. Switch frequency is about 10 Hz for slow heads, 100 Hz for fast heads (AIL). You may go down to 10 Hz in any case. Of course, this does not decrease the rise time in an absolute sense, but it reduces the ratio between rise time and on period.

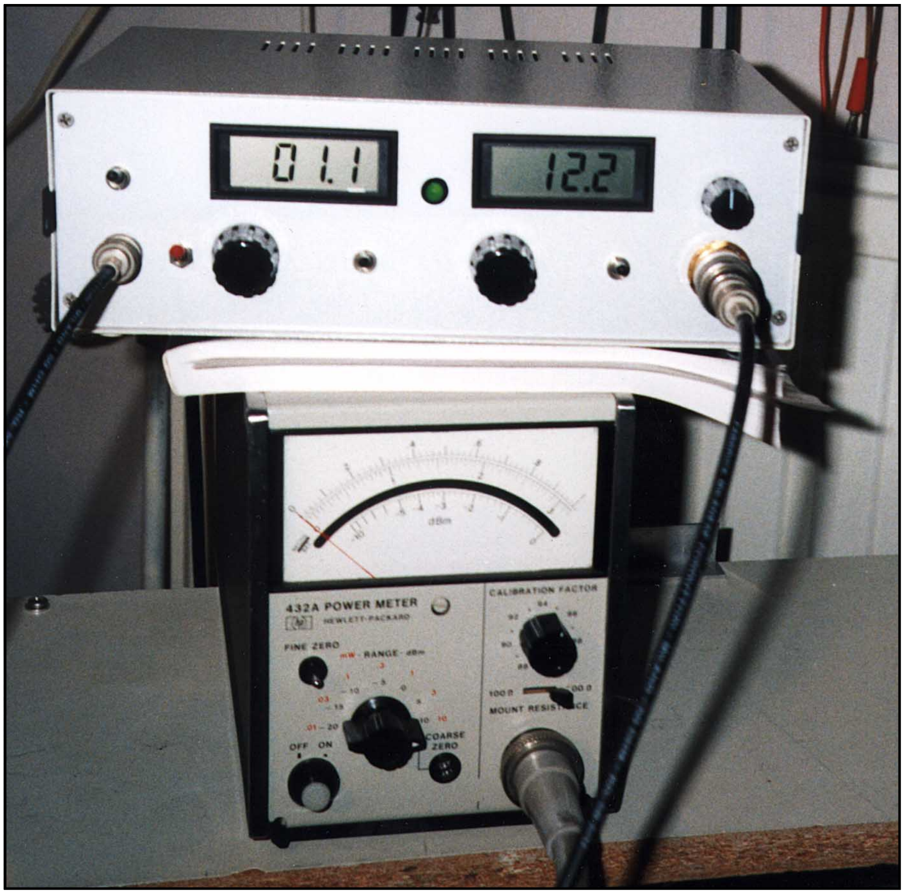

Fig 4—The noise-figure meter in PA0HRK's shack. The left indicator shows the noise figure: 1.1 dB for a DB6NT preamp. The right meter shows an uncalibrated gain value. Below the meters are the ENR pot and NF/ENR SET switch (left), relative-gain pot and GAIN/HOT-VALUE switch (right). At far right are the RF-section input selector switch and RF input connector. At top left the power switch. Lower left is the noise head connector and mode switch. Between the meters, the green LED lights when the signal level exceeds the minimum needed.

Operation

Apart from a few peculiarities, operation is quite simple. If your DUT is an LNA, you probably need a transverter. Connect the noise head to its input and the output of the transverter to the instrument's input. Adjust the *ENR* pot to set the value of the head at the transverter input frequency. Measure the noise figure of the transverter to see if it makes sense. Always be sure to measure well within the working range of the noise/gain analyzer. Adjust the gain pot so that the indicated gain is 0 dB; the actual gain set value is irrelevant. Insert the DUT, then read the noise figure and gain directly. If the DUT is an attenuator, cable, relay or other passive component, read the negative gain value indicating a loss. Note that it is not possible to measure the NF of the instrument directly because of insufficient gain. You always need some gain preceding the RF section. You may add more gain in this section, but take care to ensure linearity and add filters.

Performance and Possible Improvements

How good is its accuracy? Noise measurement is not a trivial job.[6,7] Remember, you are working with signal levels near the device's noise floor. Be aware

Table 1—Correction factors for the noise/gain analyzer.

ENR = 16 dB, NF' = log(ENR)–log(Y), NF = log[ENR/(Y–1)]

NF' (*measured, dB*)	*Correction* (*dB*)	*NF* (*actual, dB*)	*Y*
0.50	0.14	0.64	35.48
1.00	0.16	1.16	31.62
1.50	0.18	1.68	28.18
2.00	0.20	2.20	25.12
2.50	0.22	2.72	22.39
3.00	0.24	3.24	19.95
3.50	0.27	3.77	17.78
4.00	0.30	4.30	15.85
4.50	0.34	4.84	14.13
5.00	0.38	5.38	12.59
6.00	0.48	6.48	10.00
7.00	0.60	7.60	7.94
8.00	0.77	8.77	6.31
9.00	0.99	9.99	5.01
10.00	1.28	11.28	3.98

of all possible pitfalls, especially when you measure noise figures below 1 dB. You need a noise source with a well-known noise power output (the *ENR* level), since it directly impacts accuracy. The head must have a constant 50-Ω output impedance in the frequency range of interest, in both the on and off states. You must work in an RF-quiet environment—hamfests are measurement disasters! The instrument needs to work in a linear mode. Watch the LED (low voltage) and for high input voltage! You need good filtering at the input to remove unwanted signals from transverter DUTs that may distort the log conversion. Take care of the gain distribution in the measurement setup. Other sources of error seem trivial given the requirements stated above. In general, most sources of error will be outside the instrument.

Apart from the linearity of the log converter, the basic instrument accuracy is estimated at 0.1 dB in a temperature-stable environment. You may improve this a little by using offset-adjustable op amps; in my opinion, this is worth neither the cost nor the much-increased complexity. The specification of the log converter itself is given at ±0.3 dB over an 80-dB range at frequencies below 100 MHz. It will turn out to be even better over the small range of Y factors we normally encounter during a measurement. For the HP-8970, the manufacturer claims a system accuracy of ±0.2 dB, so we are in good company. The systematic error resulting from the way the noise figure is calculated is easily corrected by means of Table 1.

Can this nice instrument be improved? Well it can, especially if you are familiar with programming microcontrollers. A microcontroller with an analog-to-digital converter (ADC), such as one in the PIC family, could be programmed to take over all display and calculation tasks. This would eliminate use of the correction table. The controller could also check input signal level and automate the gain measurement by storing the actual value and displaying the actual gain after insertion of the DUT and pushing a button. A nice feature would be the calculation of the real DUT noise figure, rather than the system noise figure. Beware of RF interference from the controller. You can think of many more functions, but then you have another project.

Specifications

Here are some of the instrument's specifications to get you hungry:

- Noise figure measurement range: 0-30 dB (depends on *ENR*)
- Gain measurement range: ≈ 60 dB
- Dynamic range, RF signal: >80 dB
- Resolution: 0.1 dB (with digital panel meter)
- Basic instrument accuracy: estimated 0.2 dB
- Frequency range of input RF signal: < 500 MHz
- Frequency range of noise measurement: depends on available noise head and converters
- Bandwidth: ≈ 500 kHz, determined by filters used
- Supply of noise head: 28.0 V switched
- Total cost (without noise head): ≈ 200 Euro

Acknowledgements

Luis Cupido deserves much credit for the original idea. The designers of the AD8307 at Analog Devices surely deserve the Nobel Prize for their work. They made a very versatile IC with unprecedented specifications that is useful in lots of other interesting applications. Ever dreamed of making your own spectrum analyzer? If you need further help or clarification, please contact me.

Notes

[1]L. Cupido, CT1DMK, *UKW Berichte,* No. 2, 1998, pp 77-84.

[2]AD8307 data sheet, Analog Devices; **http://products.analog.com/products/info.asp?product=AD8307**.

[3]R. Straw, N6BV, Ed., *The 2000 ARRL Handbook* (Newington, Connecticut: ARRL, 1999), pp 26.34-26.35, Eq 19. ARRL publications are available from your local ARRL dealer or directly from the ARRL. Mail orders to Pub Sales Dept, ARRL, 225 Main St, Newington, CT 06111-1494. You can call us toll-free at tel 888-277-5289; fax your order to 860-594-0303; or send e-mail to **pubsales@arrl.org**. Check out the full ARRL publications line at **http://www.arrl.org/catalog**.

[4]M. Ulbricht, DB2GM, *UKW Berichte,* No. 2, 1981, pp 148-153.

[5]R. Bertelsmeier, DJ9BV, "A Broadband Noise Source," *DUBUS,* No. 2, 1996; also see *UKW Berichte*, No. 2, 1984.

[6]HP Application Note 57-2: *Noise Figure Measurement Accuracy*. Beware: Once you have read this application note, you probably will not go into noise figure measurement at all. (An Acrobat file of AN57-2 is available through a link on **http://www.tmo.hp.com/tmo/datasheets/English/HP346A.html**—*Ed.*)

[7]A. Ward, WB5LUA, "Noise Figure Measurements," *Microwave Update* (Newington, Connecticut: ARRL, 1997).

UKW Berichte and *DUBUS* are German publications, each publishes four issues per year. *UKW Berichte* is also known as *VHF Communications*. It is published by UKW Berichte, PO Box 80, D-91083 Baiersdorf, **ukwbericht@aol.com**; **http://www.ukw-berichte.de/**. *DUBUS* is published by: DUBUS Verlag, Grützmühlenweg 23, D-22769 Hamburg, BRD, Editor: Rainer Bertelsmeier, DJ9BV, **dubus@t-online.de**. English subscription information is available at **http://www.marsport.demon.co.uk/dubus.htm**. DUBUS is a bilingual magazine in German and English. Try to get a copy of these German magazines, they might very well interest you.

Harke Smits (born in 1953) was first licensed in 1973 (Dutch class C: 144 MHz and up), when he had already been interested in electronics and RF techniques for about 10 years. A year later he mastered the art of telegraphy well enough to obtain an A-class license. He quickly began building receivers, transmitters, radios and test / measurement equipment for frequencies from 80 meters up to 47 GHz. Actually, he has never stopped doing just that. He has never owned commercial ham equipment—with only one exception: an IC-402 (inherited from his friend PA0MGA) is his IF on 24 and 47 GHz. He strongly prefers SSB, and uses CW only if necessary. Harke likes to make rover expeditions with his microwave stations.

He owns and maintains a collection of pre-digital HP test and measurement equipment. He has written articles for Electron, DUBUS *and* Hyper.

In 1982, Harke received a masters degree in electronics from the Technical University of Twente. He then held several positions in industry before joining TNO, the largest Dutch organization for applied scientific research. As this is written, he is a business consultant working on a program to help SME's develop a new business strategy.

A Stable, Low-Noise Crystal Oscillator for Microwave and Millimeter-Wave Transverters

Would you like to try narrow-band modes in the gigahertz bands? If so, you'll need a very stable and ultra-clean local oscillator. This project fills that need.

By John Stephensen, KD6OZH

Amateurs are using narrow-band modulation—including CW, SSB and NBFM—on ever-higher frequencies. In the US, SSB is commonplace on all microwave bands through 10 GHz and is spreading to the 24- and 47-GHz millimeter-wave bands. In Europe, narrow-band operation has taken place as high as 411 GHz.[1]

The local oscillator (LO) used at these higher, millimeter-wave frequencies must be much more stable than at lower microwave frequencies. On SSB, the indicated frequency should be within 500 Hz at both the transmitter and receiver, or you may not hear the station calling you. During a microwave contest, you don't want to adjust both the antenna *and* the frequency while trying to make a contact. I wasted many hours during the last 10-GHz-and-Up contest because the LO in my transverter was 85 kHz off frequency at 24.192 GHz.

A lesser-known problem is phase noise. When extended to millimeter-wave bands, most published LO designs for amateur microwave transverters have excessive phase noise that limits the dynamic range of the transverter. Many contesters on mountaintops have experienced this desensitization from commercial equipment on nearby frequencies or amateur beacons operating at the same site.

I decided to replace the LO in my 24-GHz transverter and solve both the phase-noise and stability problems. This article describes the crystal oscillator and multiplier designs that resulted. They work nicely with existing transverters using the KK7B LO design[2] with a few modifications and can be adapted to others. The KK7B LO was originally designed for a 2304-MHz transverter and was extended by N1BWT (now W1GHZ) with an additional ×5 multiplier for his 10-GHz transverter.[3] Replacing the original LO with the circuit described here and modifying the first multiplier board make this technology much

[1]Notes appear on page 17.

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more usable on the millimeter-wave bands at 24 GHz and above.

The Phase Noise Problem Explained

The LO for a microwave or millimeter-wave transverter is usually a crystal oscillator operating around 100 MHz, which is then multiplied up to the amateur band in use. It is critical that this oscillator have a very low noise floor because each stage of multiplication adds noise to the signal at a rate of 6 dB each time the frequency doubles. In addition, the frequency multiplication process is lossy; signal levels can decrease rapidly with high-order multiplication, and the noise inherent in low-level amplifiers can degrade the noise floor even faster. Multiplying the crystal frequency by 100 to reach 10 GHz increases the noise floor by at least 40 dB. At 250 GHz, the added noise would be 68 dB, or more.

Many VHF crystal oscillator circuits have a noise floor no better than –155 dBc/Hz. This means that noise 155 decibels below the carrier power will appear in each hertz of bandwidth at the oscillator output. Multiplying to 245 GHz increases this to at least –87 dBc/Hz. This is the best case, and the degradation will always be several decibels worse due to the noise figure of components within the multiplier chain.

The problem with a high LO noise floor is that signals outside the receiver passband mix with this noise and appear as increased noise within the passband, reducing sensitivity. If NBFM is being used, the noise that appears in the 16-kHz bandwidth will be –87 + 42, only 45 dB below the level of the interfering signal: The receiver dynamic range is 45 dB. Therefore, any signal appearing within the passband of the RF circuitry (several gigahertz wide) that is 45 dB (about 7½ S-units) stronger than the desired signal will mask it completely. When transmitting, broadband noise will be radiated that is only 45 dB below your signal. The problem is smaller at lower frequencies but still results in a receiver that is easily overloaded. The same example at 10 GHz results in a 71-dB dynamic range, which is more acceptable, but still 15 dB worse than most VHF/UHF transceivers and more than 30 dB worse than many HF transceivers.

Improving Stability

The first issue is frequency stability. Quartz crystals have inherent temperature sensitivity that can vary the resonant frequency by ±10 ppm from 0 to +70°C, which amounts to ±100 kHz at 10 GHz. Temperature-compensated crystal oscillators and those in ovens do better at ±0.3 ppm or ±3kHz at 10 GHz. This is totally unacceptable at millimeter-wave frequencies, where the drift is multiplied to ±7 kHz at 24 GHz or ±75 kHz at 250 GHz.

To provide a rock-stable LO, the only solution is to phase lock the crystal oscillator to something more stable. Small rubidium frequency standards are now available at moderate cost on the surplus market. Typically, they are removed from obsolete radio-navigation equipment.[4] The long-term accuracy over temperature is 1 part in 10^9 or 0.001 ppm. This results in an accuracy of ±250 Hz after multiplication to the highest amateur band at 241-250 GHz.

Improving Phase Noise

The phase-noise problem is solved by building a crystal oscillator with a lower noise floor. The ratio of the noise at the input of the transistor to the signal arriving from the crystal ultimately determines this noise-floor level.

The traditional common-base Butler oscillator described in *The ARRL UHF/Microwave Experimenter's Manual* (see Fig 1) shows the problem. The input resistance of the 2N5179 emitter is very low ($26/I_e$ Ω). With the transistor biased for 5 mA emitter current, as shown, the input impedance is approximately 5 Ω. At resonance, the fifth-overtone crystals used in this circuit have a series resistance of about 60 Ω. In this type of oscillator, the peak current through the crystal is approximately equal to the standing current; the RMS current through the crystal is 3.5 mA, and the power dissipated in the crystal is I^2R or 0.735 mW.

The trouble is that the amount of power appearing at the 2N5179 emitter is only $(0.0035)^2(5)$ = 0.061 mW or –12 dBm, and the noise figure of the 2N5179 is probably about 10 dB in this configuration. The noise floor can be calculated by taking the noise power caused by circuit resistance, –177 dBm, adding the noise figure and subtracting input power level. This yields –177 + 10 – (–12) = –155 dBc. This is only a first-order estimate, and the actual oscillator could be several decibels worse.

The power level in the crystal cannot be raised, as it would cause excessive aging and instability. We must provide more input to the transistor by increasing its input impedance to provide a better match to the crystal. With a bipolar transistor amplifier, this means reducing the emitter current; but this would also reduce the power available, exacerbating the problem.

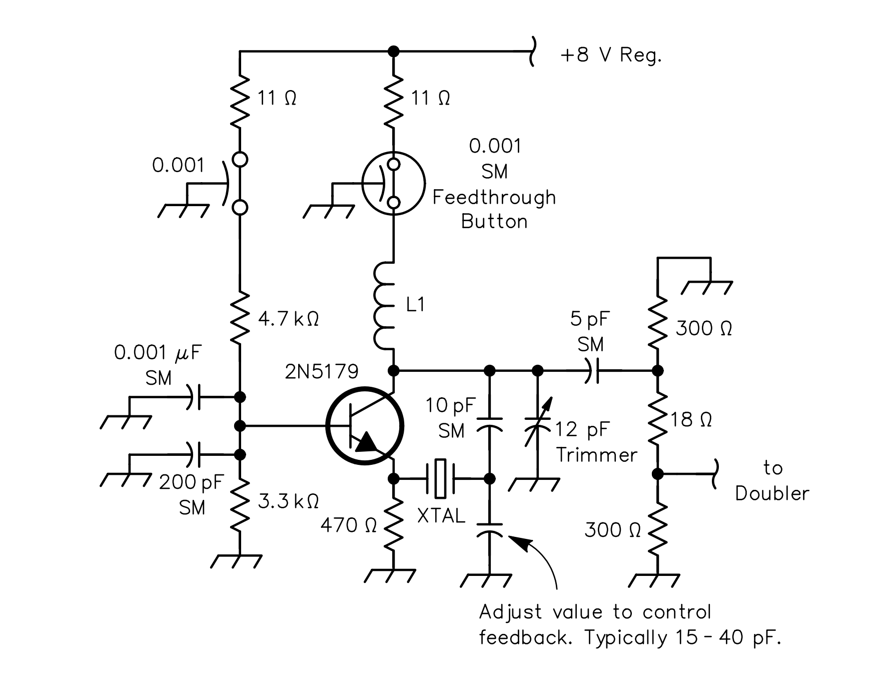

Fig 1—Conventional Butler Oscillator

The input impedance of a common-gate FET is the inverse of the transconductance and is independent of the standing current. The transconductance for a J310 FET is about 8,000-18,000 µS; its input impedance in a common-gate configuration is, therefore between 56 and 125 Ω. It also has a noise figure of less than 2 dB at 100 MHz. Replacing the 2N5179 with a J310 and keeping the same crystal dissipation results in an input power of $(0.0035)^2(56) = 0.686$ mW or –2 dBm in the worst case (lowest input impedance). We can also assume that the J310 noise figure may be degraded somewhat, to 3 dB, by noise in later stages. The noise floor is lowered to –177 + 3 – (–2) = –172 dBc—an improvement of 17 dB.

The increased input impedance does have one drawback: It is in series with the crystal, so the loaded Q of the crystal is lower than with a low-input-impedance circuit. In this application, it is not a problem. The motional capacitance of a 90 to 100-MHz crystal is about 2.4 fF (femtofarads), which results in a reactance of about 1.6 MΩ. The unloaded Q is about 27,000 (1,600,000/60) and the loaded Q is about 8700 in the worst case (1,600,000/(60+125)). This results in a bandwidth of about 11 kHz, which is fine for SSB and CW operation. To ensure that there is no additional degradation in Q, the crystal must be driven from a low-impedance source.

Once we have a low-phase-noise oscillator, we need to make sure that it is not degraded by succeeding stages in the LO chain. The amplifier(s) immediately following the VCXO must contribute very little noise, and the initial frequency multiplication must be done in small increments to minimize reduction of the LO level at any intermediate point. As the multiplication process proceeds, the LO noise floor rises. As it does, we can be less stringent in our requirements by using higher-order multipliers and having less constraint on noise figure.

The VCXO Circuit

The basis of the low-noise, phase-locked crystal oscillator (LNPLXO) is the voltage-controlled crystal oscillator (VCXO) circuit shown in Fig 2.

A J310 JFET (Q1) is a common-gate amplifier providing a high-impedance input for the signal from the crystal (Y1). A 2N5179 bipolar transistor (Q2) is an emitter follower providing low-impedance drive to the crystal. Y1 is a fifth-overtone, AT-cut crystal ground for operation in the series-resonant mode with a load capacitance of 30 pF. The feedback path is completed through a resonant circuit consisting of C1, C2 and L1 that selects the desired overtone. R14 loads the drain of Q1 to ensure linear operation. It also sets the loop gain of the oscillator. The dual varactor diode, D1, in series with the crystal, provides for pulling of the crystal frequency by ±500 Hz. L2 cancels out the parallel capacitance of the crystal to enable a wider pulling range.

The circuit composed of D2, R2, R3 and C9 controls the amplitude of the oscillator. The voltage across R3 is about 1.6 V. When the RF voltage on Q1's drain reaches –2 V, D2 conducts, and the signal is limited. This is done without affecting the operating point of Q1, so it remains in a linear, low-noise mode. R2 can be adjusted to change the output level of the oscillator, while making sure that the power

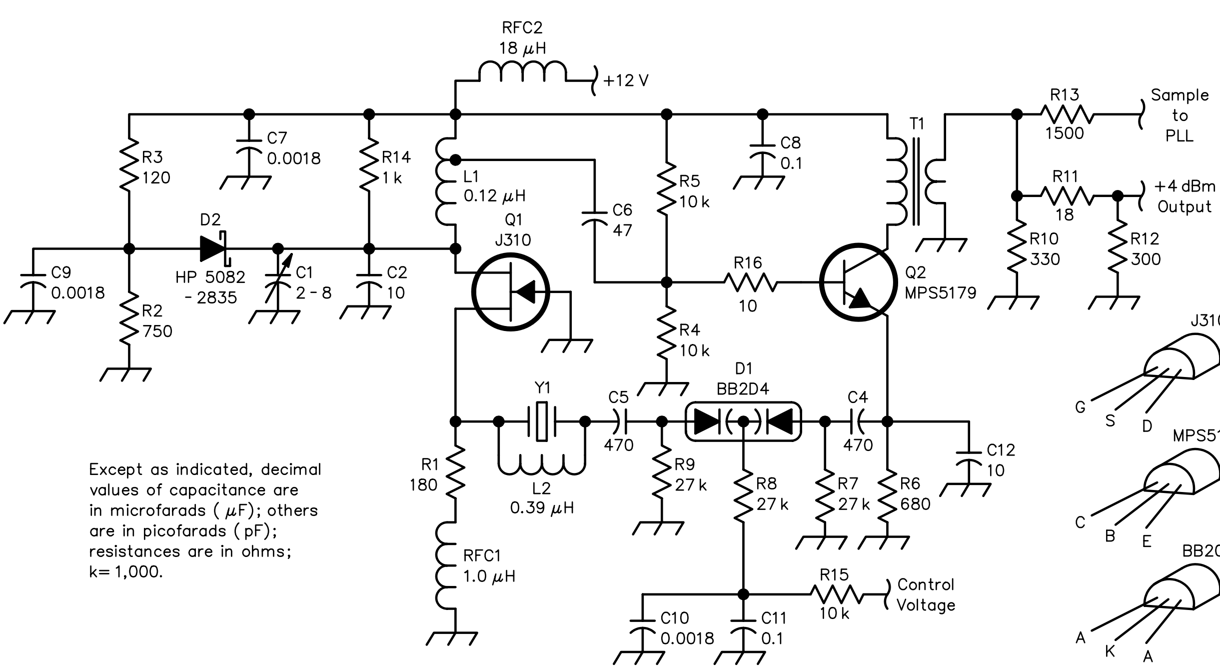

Fig 2—VCXO schematic diagram. Unless otherwise specified, use 1/4 W, 5%-tolerance carbon composition or film resistors.

C1—2-8 pF NP0 ceramic trimmer
C2—10 pF NP0 ceramic disc
L1—0.12 µH, 8 turns #22 AWG enameled wire on T30-12 with a tap 2 turns from cold end
L2—0.39 µH, 15 turns #26 AWG enameled wire on a T37-12 powdered-iron core.
T1—Primary 6 turns #30 AWG enameled wire, secondary 2 turns #28 AWG enameled wire on a BN-61-2402 ferrite core.
Y1—Fifth-overtone crystal, series resonant, 30 pF load capacitance, series resistance less than 60 Ω.

dissipated in Y1 never exceeds 1 mW.

To ensure minimum effect on loop gain by external load variations, the output of the oscillator is taken from the collector of Q2. T1, a 9:1 broadband transformer, provides impedance matching; 6 to 7 dBm is available at its secondary. A 3-dB attenuator provides more isolation and reduces the signal level to prevent overdrive of the next stage in the LO chain. The 1500-Ω resistor is used to couple some of the output to the PLL circuitry.

The PLL Circuit

Fig 3 shows the PLL components required to lock the VCXO frequency to the 10-MHz output of a rubidium frequency standard. A sample of the VCXO output is applied to a dual-modulus prescaler chip, U2. This can be a Motorola MC12019 for division by 20/21, or a Motorola MC12015 for division by 32/33. The prescaler divides the VCXO frequency so that it will not exceed the 20-MHz maximum clock frequency of the PLL chip, U1.

The Motorola MC145158 PLL chip at U1 is the heart of the circuit. It accepts the 10-MHz reference frequency at pin 1 and divides it to a user-settable internal reference frequency using the R-counter. The prescaled VCXO output is applied to pin 8, the input of the A and N counters. The A counter determines when the prescaler will be switched from the N to N+1 mode. The N counter divides the prescaler output before application to the phase detector. The PLL will lock at the frequency determined by all of these counters as shown in Table 1.

The phase-detector output at pin 4 of U1 is filtered by R19 and C13, then amplified by U3. U3 is required to increase the phase detector output, from 5 to 13.5 V. Note that R19 and C13 are *not* the values predicted by the equations Motorola supplies for the PLL. The crystal is a high-Q device and there is a time delay when changing its frequency. The low-pass-filter time constant had to be determined experimentally. The loop has been verified to be adequately damped with three crystals of different manufacture, so the time constant should not need to be changed. To ensure stability, the PLL reference frequency should not be set below 50 kHz. It should also not go above 200 kHz to ensure that the phase-detector output is accurate. This frequency range should be adequate for all amateur microwave and millimeter-wave LO requirements. Choose your division ratios carefully.

The PLL (U1) is configured by data entered serially on the clock, data and enable pins. A PIC16F83 microcon-

Table 1—Frequency vs Divider Values R, A, and N.

MC12019 prescaler: Frequency = (10 MHz / R)×(A + N×20)
MC12015 prescaler: Frequency = (10 MHz / R)×(A + N×32)
Note that *N* cannot be less than the prescaler division ratio.

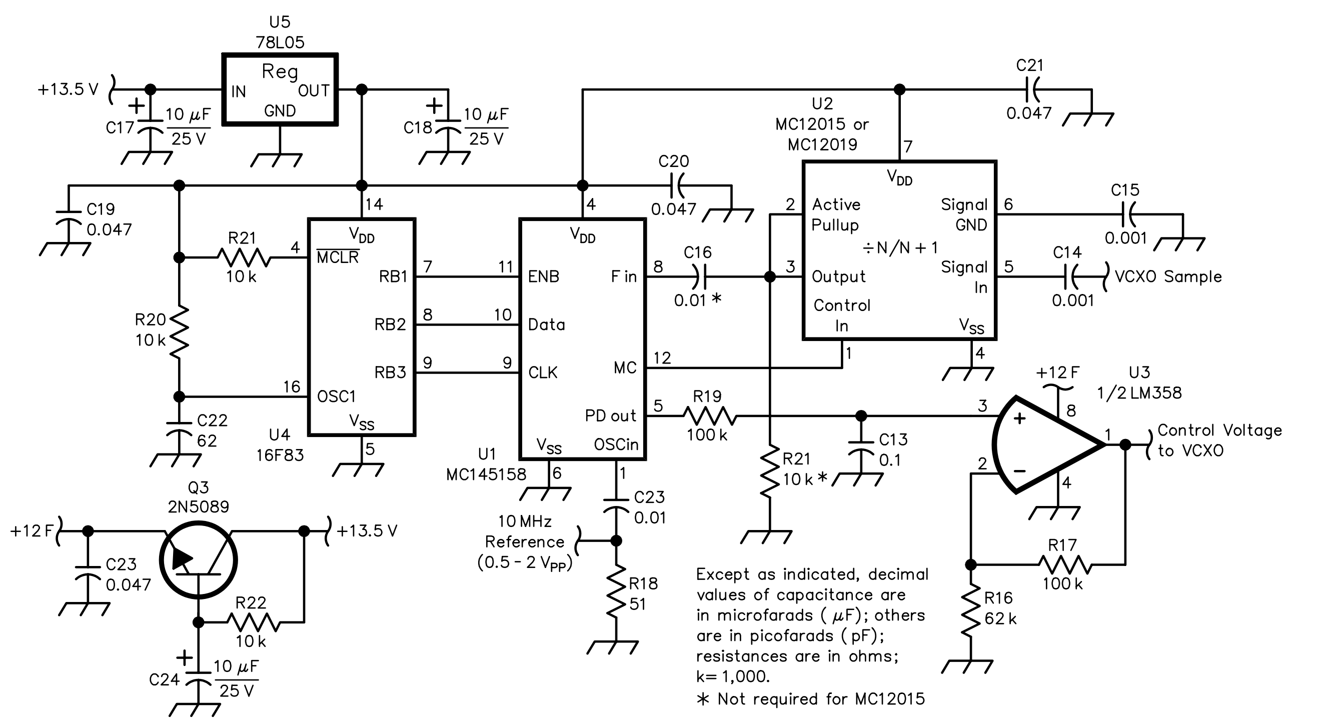

Fig 3—PLL schematic diagram. Unless otherwise specified, use ¼ W, 5%-tolerance carbon composition or film resistors. C16 and R21 are not required if U2 is a MC12015.

C22—62 pF ceramic, ±10% tolerance
C14-C16, C23—ceramic, ±20% tolerance
C13—Monolithic ceramic, ±20%
C19, C20, C21, C23—Monolithic ceramic, +80%/–20%
C17, C18, C24—Tantalum Electrolytic
U1—MC145158 PLL
U2—MC12019 (∏20/21) or MC12015 (∏32/33) prescaler
U3—LM358
U4—PIC16F83 microcontroller
U5—78L05 5-V regulator

troller unit (U4, MCU) provides these data. Bits 1, 2 and 3 of U4 port B are used to drive these lines. Since the MCU does not require an accurate clock, the RC oscillator configuration is used. The use of an MCU may seem like overkill, but a PIC16F83-04 with 1 KB of program memory and 68 bytes of RAM is only $6.25 in single-unit quantities. The single 18-pin DIP package is also smaller than any other solution.

LO Chain

My 24.192-GHz transverter uses a harmonic mixer with an anti-parallel diode pair that requires injection at half the LO frequency. With a 1296-MHz IF, this works out to 11.448 GHz. Thus, the output of the VCXO at 95.4 MHz needs to be multiplied by 120. I planned to use the same frequency-multiplication chain that N1BWT used in his 10-GHz transverter (see Note 3). This multiplied the crystal frequency by six in the first stage, followed by two stages of multiplication by four and five. Each was a single PC board. However, measurements showed that it degraded the phase noise significantly. The first multiplication in this LO chain reduced the LO signal level below –23 dBm. Coupled with a noise figure of 6 dB, this could not support an LO-to-noise ratio of more than –148 dB—a degradation of 7 dB on the theoretical minimum of –155 dBc.

The LO chain was redesigned to add an additional stage, resulting in successive multiplication by 2, 3, 4 and 5. The KK7B LO board was modified to act as a tripler and a new doubler stage was designed to precede it.

Frequency Doubler

The doubler stage in Fig 4 is essentially a full-wave rectifier. This circuit suppresses the fundamental and odd harmonics,[8] which results in much cleaner output from the following tripler. U1, an MSA-1104, provides the low noise figure and high output level that are critical to maintaining a low noise floor. D1 and D2 are driven with about +16 dBm and produce about +3 dBm of output, which is then filtered by L1 and C2. Minimal filtering is required here, given the six poles of filtering in the tripler that follows.

Frequency Tripler

Modifying a KK7B ×6 multiplier board to perform the tripler function turned out to be fairly easy. Remove U1, Q2 and R7 from the board (see Fig 5) to disable the original crystal oscillator. Connect the doubler output to the free end of C10. Replace U2 with an MSA-0485—less gain is needed, as the input level is larger. It amplifies the +3 dBm from the doubler to +11 dBm, which is enough to drive the following diode tripler. Also, replace L3 with a 0.22 µH RFC.

The original diode sextupler is converted to a tripler by removing L4 and C12 and changing C11 to 8.2 pF. This creates a low-Q series resonant circuit at about 190 MHz to isolate the tripler output from U2. D1 is also removed and replaced with two PIN diodes in anti-parallel. PIN diodes produce about 3 dB more output in this application than Schottky diodes[7]. The anti-parallel circuit also results in 6 dB more output than a single diode and good suppression of even harmonics, resulting in a cleaner LO. I used two ECG-553 diodes because they were easily available, but an HP HSMP-3821 or two 5082-3188s should work just as well. I thought about designing a more complex impedance-matching circuit for the PIN diodes, but tripler output was better than predicted by the HP application note, so I discarded the idea in favor of simplicity.

U3 is replaced with an MSA-0685 to reduce the noise figure. Since the signal levels are much higher, U4 is removed and replaced with a copper strap to the filter. Also remove C15 and R9. Replace C15 with a copper strap and change R10 to a 360-Ω, 1/2-W resistor. U5 can be an MSA-0485 for +12 dBm output or an MSA-1104 for +16 dBm output. The MSA-0485 is adequate to drive the KK7B ×4 multiplier board.

Construction

I built the VCXO and PLL on copper-clad Vectorboard. Make sure to shield the VCXO from the PLL, otherwise leakage from the PLL digital circuitry will show up as spurs at –50 to –60 dBc after multiplication to 11 GHz. One doubler was constructed in a minibox for testing and the other on the KK7B board by chopping up the traces in the old crystal-oscillator area. It could probably be added to the VCXO board.

Circuit construction on copper-clad perfboard was described in amateur literature in the late 70s, but it is less well known today. It requires the use of old-fashioned leaded components, but avoids the need to etch PC boards. I've found that this method can be used up to 150 MHz if care is taken to provide large interconnected areas of copper as a ground plane.

The board I use is made by Vector Electronics.[5] It is an epoxy-glass material with 0.042-inch-diameter holes punched on a 0.1-inch grid and is coated with copper on one side. Components are mounted on the epoxy-glass side and the grounded ends of components are just soldered to the copper. Where ungrounded connections are required, a pad-cutting tool is used to remove the copper around the holes. With good layout, most interconnections in analog circuits can be accommodated by mounting the components in adjacent holes and soldering the leads together on the non-component side of the board. Occasionally, you will need to run a wire between holes to make connections.

MCU Programming

The PIC16F83 MCU must be programmed to load the appropriate values into the MC145158 PLL chip. A listing of the program used may be

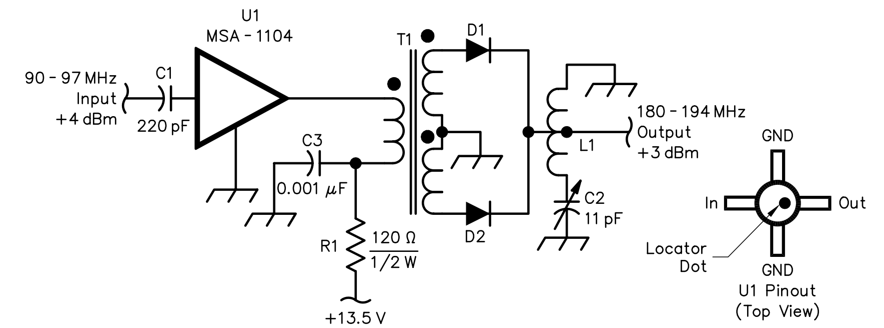

Fig 4—Frequency doubler schematic.

C2—11 pF trimmer
D1, D2—HP5082-2835
L1—6 turns #24 AWG enameled wire 1/4-inch long on a 3/16-inch-diameter form. Tap at 1 1/2 turns from grounded end.

T1—5 trifilar turns #26 AWG enameled wire on a FT37-61 ferrite-toroid core.

downloaded from the ARRL Web page.[10]

The program is very simple. On power up, it initializes the port used to communicate with the PLL, executes a delay routine to ensure that the PLL chip has powered up, programs the PLL registers then shuts off. The PLL counter division values are sent one bit at a time, with the most-significant bit first. In this example, the PLL phase detector runs at 100 kHz, so the R counter is set for division by 100. The N and A counters are set for division by 49 and 15, respectively, to achieve division by 995 when used with a MC12015 prescaler.

The program uses common subroutines to send ones and zeros to the PLL chip and enable latching of the values sent. The program may be easily altered by changing the calls to the subroutines ONE and ZERO to reflect the binary values to be loaded into the counters. For information on programming PIC MCUs, see the October 1998 *QST*.[6]

Adjustment

Adjustment of the LNPLXO is straightforward. Apply power and adjust C1 for maximum output on a power meter. This can be as simple as a 50-Ω resistor, a Schottky diode rectifier and a voltmeter. Then connect the 10-MHz reference oscillator and adjust C1 for approximately 5 V on D2. This detuning pulls the crystal frequency slightly to center it in the PLL lock-in range. Check the power output again to be sure that it has not dropped by more than 0.5 dB. If it has, add a capacitor in series with the crystal to raise the frequency, or an inductor to lower the frequency and repeat the adjustment. Crystals ordered with a 10-ppm tolerance and 30-pF load should not require any circuit modifications. If you have crystals on hand that are calibrated for series resonance without a reactive load (so-called 0-pF load), insert a 100-nH inductor in series with the crystal.

Results

The output at 552 MHz was examined on a spectrum analyzer and all spurious outputs—except the second harmonic—were below –55 dBc, as shown in Table 2.

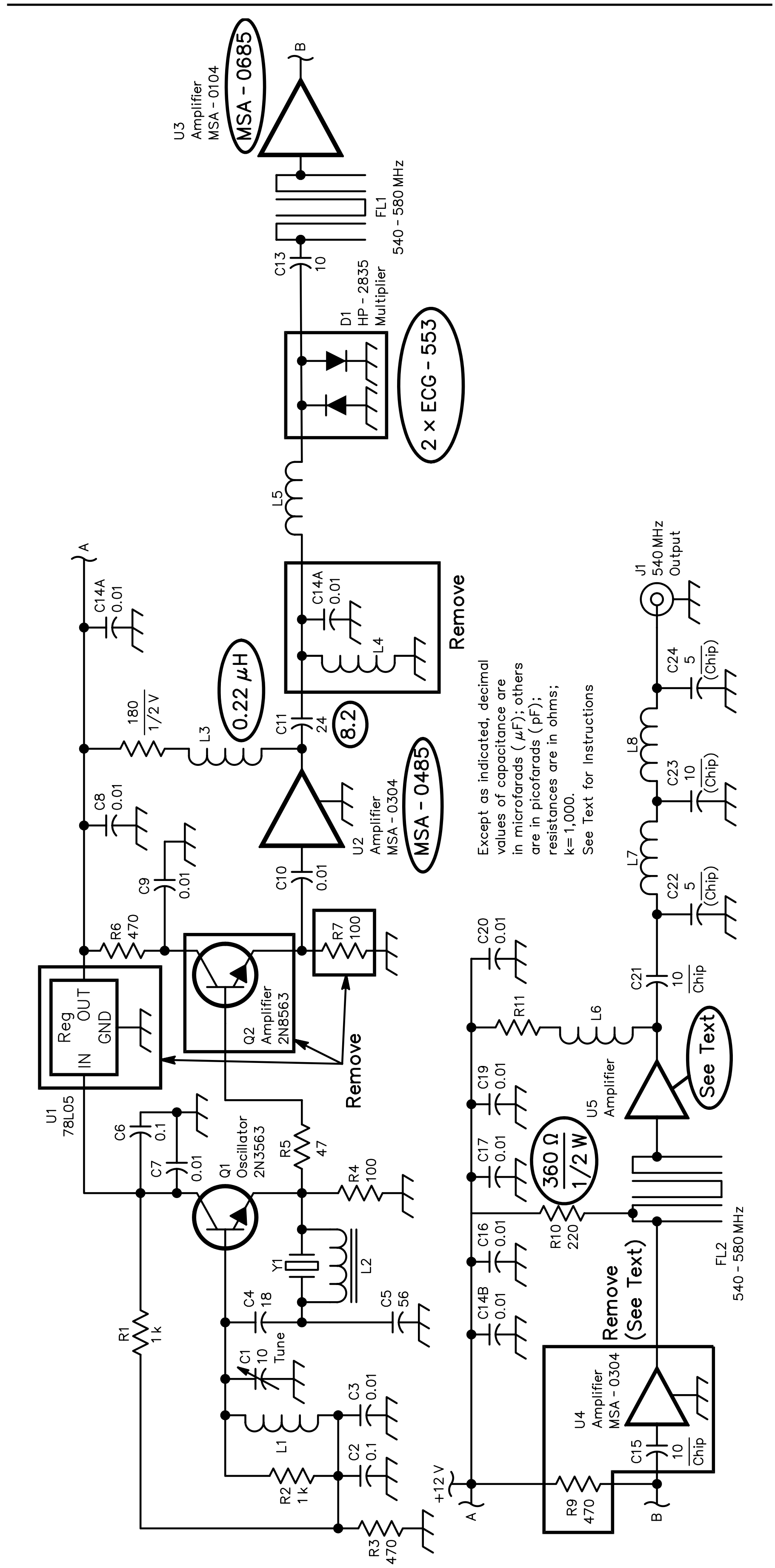

Fig 5—(right) Modifications to convert KK7B's ×6 multiplier into a frequency tripler. Detailed instructions are given in the text.

Table 2—Output Spectrum

Harmonic	*Frequency (MHz)*	*Level (dBc)*
1	92	< –75
2	184	–62
3	276	< –75
4	368	–56
5	460	–70
6	552	0
7	644	< –75
8	736	–67
9	828	< –75
10	920	–72
11	1012	< –75
12	1104	–43

Phase-noise measurement was done by building two identical 92-MHz LNPLXOs and multiplier chains then connecting them to a common 10-MHz crystal oscillator. The outputs (approximately +17 dBm) were then applied to a double balanced mixer (DBM) through 10- and 20-dB attenuators, resulting in a +7 dBm LO and –3 dBm at the RF port. The output of the DBM was ac-coupled to a low-noise amplifier. This arrangement converts the carrier to dc and the phase noise sidebands on each side of the carrier to frequencies between 2 kHz and 500 kHz, where they were measured on a HP 8553 spectrum analyzer. See Note 9 for more information on this technique.

To measure the phase noise in the absence of the multiplier chain, the LNPLXOs were connected to low-noise isolation amplifiers and a level-23 DBM. This DBM did not support an IF below 100 kHz, so measurements were limited to offsets between 100 and 500 kHz.

The noise floor of the basic LNPLXO was measured to be –171 dBc. After multiplication by six to 552 MHz, the noise floor at a 500-kHz offset was measured at –154 dBc. This is very close to the theoretical minimum, considering that my measurements are only accurate within ±2 dB. The phase noise versus frequency for these two cases is shown in Table 3. The SSB phase noise rises slowly from –150 dBc at 50 kHz to –143 dBc at 5 kHz, and then jumps to –133 dBc at 2 kHz. The slow rise is probably caused by flicker noise in the J310 FET and AM-PM conversion in the multipliers. The more rapid rise below 5 kHz occurs within the passband of the crystal itself, and is as expected.

Table 3—SSB Phase Noise at 552 MHz

Offset (kHz)	*Noise (dBc)*
2	–133
5	–143
10	–146
25	–148
50	–150
500	–154

This performance should be adequate for my 24-GHz transverter, which has a noise figure of 1.9 dB and a calculated third-order input intercept of –28 dBm. This would result in a two-tone dynamic range of 76 dB with a perfect local oscillator. I estimate that the LO phase-noise floor will degrade by 34 dB (32 dB for multiplication by 40, plus an extra 2 dB for circuit losses) to –120 dBc after multiplication to 22.896 GHz. This would limit dynamic range to 78 dB in a 16-kHz bandwidth, so LO noise is not the limiting factor for NBFM, SSB and CW operation on the 12-mm band.

Low LO phase noise will be even more important for the 6-mm band, where LO phase noise will degrade by 7 dB, reducing the dynamic range to 71 dB. This results in a dynamic range at 47 GHz that is similar to that of 10-GHz transverters using the traditional Butler oscillator circuit. In addition, the frequency will stay within 50 Hz—even if it is 115° in the shade.

Notes

[1]H. Neidel, DL1IN, "First QSO on 411 GHz," *DUBUS*, 2/98, p 44.

[2]R. Campbell, KK7B, "A Clean, Low-Cost Microwave Local Oscillator," *QST*, July 1989, pp 15-21.

[3]P. Wade, N1BWT, "Building Blocks for a 10-GHz Transverter," *ARRL UHF/Microwave Projects Manual*, Vol 2, pp 3-33 to 3-40.

[4]I use a surplus Efratom FRK-L rubidium oscillator obtained from Lehman Scientific, 85 Surrey Dr, Wrightsville, PA 17368. They will also perform calibration of oscillators, traceable to NIST. Don't try to calibrate these using WWV or WWVH. Received WWV and WWVH signals are only accurate to 0.1 ppm because of propagation variations caused by short-term instability of the ionosphere. These used oscillators sell for about $500.

[5]Vector Electronics products are available from most distributors and some local electronic parts stores in the US. Vector part number 169P84WEC1 is a 17× 8.5-inch board. Vector part number P138 is a pad-cutting tool.

[6]J. Hansen, W2FS, "Using PIC Microcontrollers in Amateur Radio Projects," *QST*, Oct 1998, pp 34-40.

[7]*Low Cost Frequency Multipliers Using Surface Mount PIN Diodes*, Application Note 1054, Hewlett Packard. (A PDF file of this note is available at **http://ftp.hp.com/pub/accesshp/HP-COMP/rf/4_downld/lit/diodelit/an1054.pdf**.—*Ed.*)

[8]W. Hayward, W7ZOI and D. DeMaw, W1FB, *Solid-State Design for the Radio Amateur*, p 42.

[9]V. Manassewitsch, *Frequency Synthesizers—Theory and Design* (New York: John Wiley and Sons, 1987).

10 You can download this package from the ARRL Web **http://www.arrl.org/files/qex/**. Look for STEP1199.ZIP.

John Stephensen, KD6OZH, has been interested in radio communications since building a crystal radio kit at age 11. He went on to study Electronic Engineering at the University of California and has worked in the computer industry for 26 years. He was a cofounder of Polymorphic Systems, a PC manufacturer, in 1975 and a cofounder of Retix, a communication software and hardware manufacturer, in 1986. Most recently, he was Vice President of Technology at ISOCOR, which develops messaging and directory software for commercial users and ISPs.

John received his Amateur Radio license in 1993. He is active on amateur bands from 28 MHz through 24 GHz. His interests include designing and building Amateur Radio gear, operation through digital and analog amateur satellites, VHF and microwave contesting and 10-meter DX. His home station is almost entirely homebrew and supports operation on SSB, PSK31, RTTY and analog and digital satellites in the 28, 50, 144, 420, 1240, 2300, 5650 and 10,000 MHz bands from grid square DM04 in Los Angeles. The mobile station includes 10-meter SSB, 144/440 MHz FM and 24 GHz SSB.

Signal Sources

Do you need signal sources for RF evaluation and test work? Perhaps you would like to replace some big, hot boxes with smaller, cooler ones.

By Bruce Pontius, N0ADL

I am fortunate to have some professional-level RF signal generators and other test equipment in my home workshop. However, most of these pieces are very large and heavy; they consume lots of power, generate lots of heat and have noisy fans. The heat, humming and size of the large instruments make my workroom into a hot, noisy, cluttered place. I decided to replace some of this older (but good) equipment with new, compact, cool-running, quiet instruments where possible. These new pieces can be used for most of the less-critical or less-demanding jobs, even if their performance is not quite as good as the larger, more-elaborate equipment.

I will describe a few signal sources intended for applications in RF-communication work, including component evaluation, circuit design and testing and some equipment-level testing. These instruments are not intended for precision local-oscillator or carrier-generator use in narrow-band or digital radio systems, but they could be used for some tests in such cases. I will describe performance factors and limitations, and compare features and performance with typical professional-grade equipment. The sources described will serve as models for other frequency ranges.

The equipment described can be built by a motivated experimenter at lower costs than buying large, heavy surplus equipment. Features and ease of use will be compromised somewhat, however.

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Sources for 0-50 and 2100-2500 MHz

Before I got started on RF signal generators, I wanted to clear off my bench to make more room to work. For some of my overgrown test setups, I have simultaneously used 10 or more dc power supplies, from 3 to 28 V, so I had already built several small dc sources using LM317 regulators. These units are all fed from a fixed-output, 15.5-V supply. In another case, I built a parallel low-current supply into one of the small RadioShack dc-supply boxes by adding another small power transformer and some regulator ICs. This supply provides 18 V for tuning diodes and various linear instrumentation amplifiers. These additions significantly reduced the bench space formerly taken by several separate, large power supplies, which can be retrieved from storage when needed.

The next candidates for replacement were my vacuum-tube HF (100 kHz-30 MHz) signal generator and an HP-8614A 800-2400 MHz signal generator, which also has some hot tubes in it. While I was at it, I planned to build two or more of each generator for all of those times when multiple sources are needed. If the units were small, cool and stackable, I could have a whole pile of them without unduly overloading my workspaces. Actually, one HP-8640B or HP-8614A generator takes up more space than dozens of the new little ones.

A Replacement HF Generator

When I say replacement, I mean with no additional features, unless they would be very easy to realize or merely incidental to the process. In the case of the HF generator, all I wanted was a small box that could be easily set to a desired frequency at a known power level.

Frequency and power settings should be virtually drift-free. Tuning should be nearly continuous or with less than 10-Hz steps. Energy isolation without elaborate shielding was desired to make mechanical packaging easy and compact. Reliable-magnitude, low-level signals (less than 0.1 µV) should be readily achievable.

The box does not need a built-in digital readout of frequency or level because my test setups are usually instrumented with variable attenuators, power meters, high-resolution counters and spectrum analyzers.

Computer control or interface is not desired. Modulation, if required, is usually handled outside of the generator box if other than AM or audio square-wave modulation. Low phase noise or high short-term stability (within a few dozen hertz) is not required of these general-coverage sources. Fast frequency changing is not necessary.

There are several ways to build an HF generator. A very brief review of some possibilities includes:

1. *On-Frequency, Wide-Range Capacitor/Switched-Inductor, Mechanically-Tuned Oscillator:* A Motorola MC12148 (MC1648) would probably do a good job.[1] This type of generator could be electronically tuned or swept over parts of the range and could be frequency-stabilized if necessary. However, I did not want to struggle with shielding RF circuits or mounting coils and a large tuning capacitor.

2. *PLL and DDS Methods:* The wide frequency range, 30 MHz maximum output frequency and small-step tuning made these approaches too complex for my taste. There have been several examples of DDS and PLL synthesizers in *QEX* and *QST* in the last few years.[2, 3, 4] A person could use the local-oscillator synthesizer from a spare HF receiver, but extraction and repackaging sounds like a lot of hassle.

3. *A VHF Oscillator Heterodyned Down to Cover HF.* This approach requires frequency stabilization. Shielding at the desired output frequency might be readily achieved. This approach could possibly provide the opportunity for electronic tuning over a wide frequency range or electronic sweep from near zero to 50 MHz. The design can be readily built with simple building blocks. Frequency stabilization techniques and circuits are presented in several references.[5, 6, 7, 8] The CycleMaster Plug-'n-Play Frequency Control by Lee Richey[9, 10] is particularly easy to build and apply.

For several reasons, I chose the third option for building the first unit. This approach was perceived as the simplest and easiest path to achieving the design goals of the project. By using the CycleMaster frequency controller, a versatile multi-application system can be built with modular building blocks. After procuring a CycleMaster kit, I had all of the components on hand to build several generators at various frequencies, up through 2500 MHz.[11]

Fig 1 is a block diagram of a simple approach to building an HF generator. It has other uses as well, such as a controlled VHF source for heterodyning to higher frequencies.

The CycleMaster frequency controller microcomputer/counter can count to just over 50 MHz, and it has an integrated LED digital frequency readout. The controller is designed to hold the frequency presented at the frequency counter input within about ±10 Hz. Since the VHF oscillator frequency is well above 50 MHz, the oscillator output is mixed down to a range of 0-50 MHz. The controller then holds the oscillator frequency to within ±10 Hz. This signal is also the HF output of the source, as shown in Fig 1.

It is a small disadvantage of this arrangement that the desired output frequency (0-50 MHz) is present at a moderate level in the controller /counter input signal-conditioning circuits. Some signal leakage can make low-level receiver-sensitivity measurements difficult, thereby requiring shielding of the counter circuits.

The output from near zero to over 50 MHz is a clean signal of predictable level, flat to within ±1 dB over most of the range. Spurious signals are virtually eliminated by a good low-pass output filter (with a cutoff just above 50 MHz) and a low-pass harmonic filter on the VHF oscillator. There is one harmonically related spur 50 dB or more below the desired signal (–50 dBc). In some small frequency ranges, it rises to around –30 dBc. The spurious signal travels at twice the rate of the desired signal, so it is easy to move that spur slightly away from the desired signal if necessary.

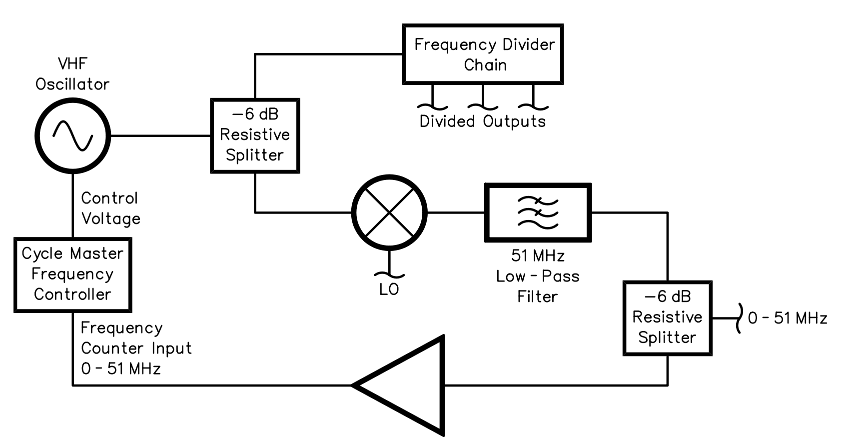

Fig 1—Block diagram of a frequency-stabilized signal source from 100 kHz to 50 MHz, with some VHF coverage as well.

[1]Notes appear on page 29.

The spur is close to the desired signal only in a narrow frequency range near 36 MHz.

Use of the frequency-divider circuits can provide cleaner and much more stable signals if the addition of simple range-switching (by switching dc power to dividers) and low-pass filters is acceptable. In fact, the arrangement shown in Fig 2 is very simple and easy to build. The mixer and local oscillator can be eliminated for simplicity. Of course you could use only the mixer and LO and omit the frequency dividers. We will consider some of the trade-offs later, in the performance-review sections.

The CycleMaster boards are easy to assemble, as are the MAR-6 amplifier and the three divider ICs. The small-package VHF oscillator is simply mounted as an IC. A Mini-Circuits POS-150 provides a simple, quick and adequate oscillator for most purposes. The cost-to-performance ratio is hard to beat (about the same cost as a large pizza). The JTOS-150 has almost the same specifications in a surface-mount package. It costs about \$2 more than the POS package. I had a couple of the JTOS parts on hand, so I used those. The JTOS runs from 72 to over 160 MHz.

The circuits around the oscillator are given in Fig 3. When not using the frequency controller, the control-voltage tip plug is removed and the 100 kΩ resistor is left open at that end. The circuit then reverts to manual control only.

With this arrangement, the VHF VCO can be manually tuned across its range, even with the CycleMaster (CM) control voltage plugged in. The CM can be set in the counter mode to

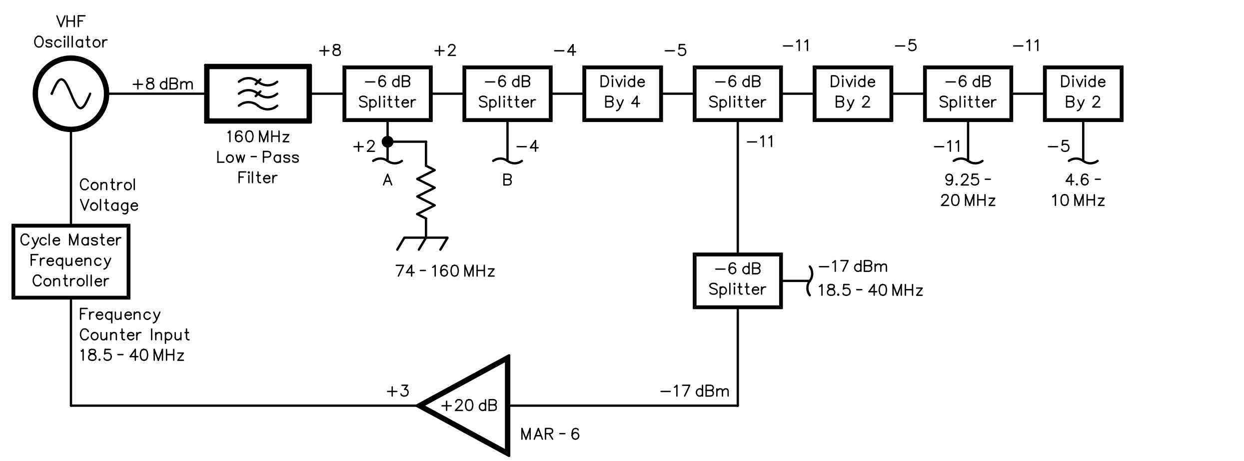

Fig 2—An HF signal source using a few easy-to-build blocks to cover 4.6-40 MHz. The VHF oscillator is a MCL POS-150. CycleMaster is a kit from Radio Adventures Corporation (see Notes 9 and 10). Amplifier is an MCL MAR-6. Dividers are Motorola MC100EL33 and 32, HEL-33 (divide by four) and HEL 32 (divide by two). Splitters are 6-dB resistive types built with 1/4-W resistors. The 160-MHz LP filter is described later. (It is not needed if only the output of the frequency dividers is to be used.)

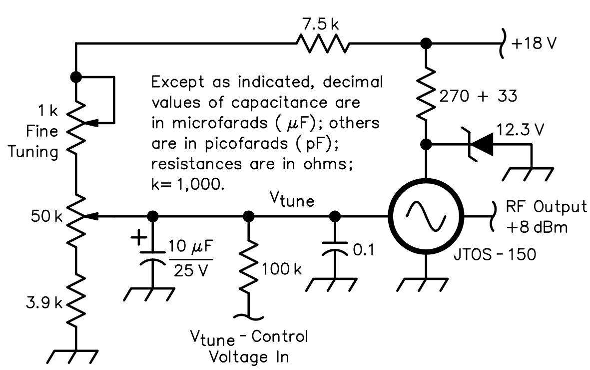

Fig 3—JTOS-150 or POS-150 circuit. The 18-V supply is required to obtain the full tuning range at the high end. The controller outputs 0 to 8.5 V at V_{tune}.

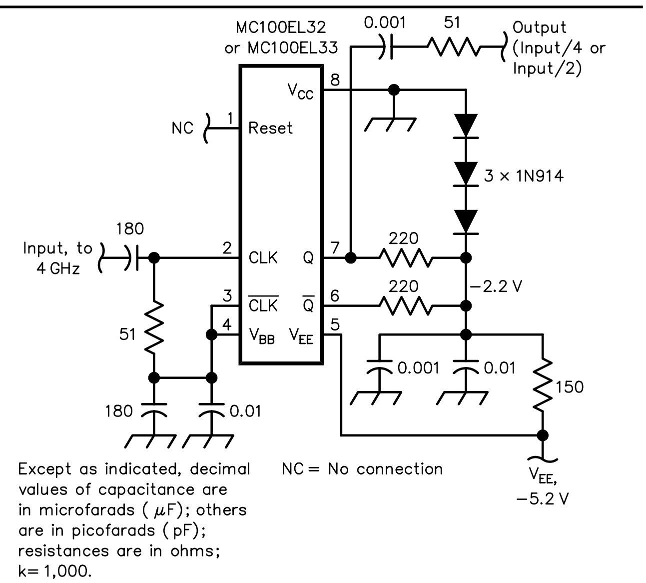

Fig 4—Frequency dividers. All components are surface-mount chips except the diodes and 150-Ω resistor.

read frequency directly on the digital readout. When fine tuning or frequency stability is desired, the CM is put in control and tuning is done with the CM dial. Easy, no-backlash tuning is provided, and the frequency is locked to the CM setting when tuning stops. Two VFO settings, A and B, and 16 programmable—or preset—frequencies are available. See the referent of Note 9 for further description of the operation of the CycleMaster.

The speed of control is not an issue for this signal generator. A control-voltage-integrating resistor (1 MΩ) in the CM is used in most cases. To ensure stability while under electronic frequency control from the CM, the electronic tuning range is held to less than 1 MHz/V of CM tuning voltage in all VCO applications. A typical frequency-change time from VFO A to VFO B is a few seconds. The electronic tuning operates in parallel with manual knob tuning.

The wide-range VCOs used in this project are very sensitive to tuning voltage (from 2.6 to 44 MHz/V). Various VCOs from 25 to 2500 MHz were satisfactorily controlled with the CM as long as tuning rates at the control-voltage input point from the CM were not too fast (less than 1 MHz/V).

In the circuit of Fig 2, the oscillator frequency is divided by four and frequency control is applied at that frequency; the oscillator is controlled within about ±40 Hz, but the signal output after the division by four is about ±10 Hz again. The 4.6-10 MHz

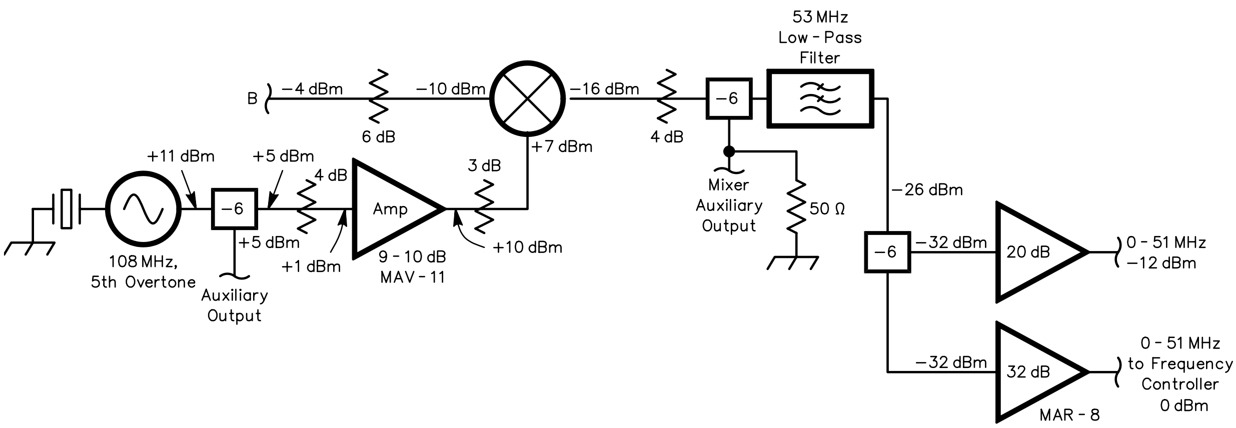

Fig 5—A mixer and local oscillator can be added at point B of Fig 2 to provide 0-50 MHz for frequency control and for continuous HF output. The prototypes used two MAR-6 ICs in series for 40 dB gain, with appropriate attenuation. The mixer is an MCL SBL-1. The MAR-8 amplifiers replace the MAR-6 in Fig 2. The fifth-overtone oscillator is shown in Fig 7.

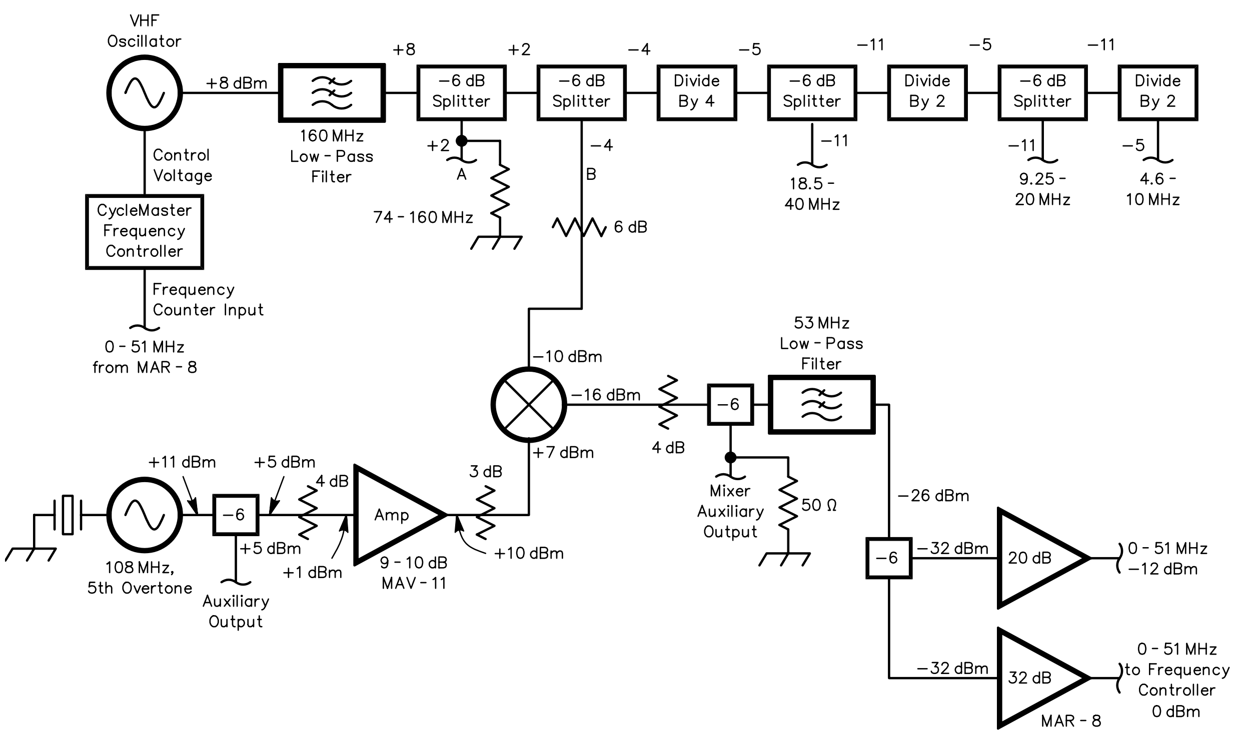

Fig 6—Composite block diagram for both heterodyning and frequency dividing. These blocks would replace Figs 2 and 5. The MAR-8 amplifiers replace the MAR-6 in Fig 2.

output is held within about ±2.5 Hz (40/16 Hz).

The frequency dividers are Motorola MC100ELs, labeled HEL 33 (÷4) and HEL 32 (÷2) on the device packages. I chose these because they were on hand from a 4-GHz application (dividing down the frequency of a YIG oscillator). They are also used to divide the 2400-MHz source in the next application. These ICs cost about $4.95 each. Only one divide-by-four is required (at the 160-MHz input point), then lower-frequency CMOS dividers can be used with appropriate signal-level changes. Some 74AC74s that I tested toggled at over 160 MHz. Motorola says 160 MHz is typical for the ACs, but their MC74VHC74 is typically 170 MHz. However, the HEL units are convenient since one divider will drive a following divider through a 50-Ω, 6-dB resistive splitter, allowing for outputs without need for buffers. Fig 4 shows a schematic. A mixer and LO can be added at point B in Fig 2, as shown in Fig 5.

If the source is to be used for very low-level outputs, the 20-dB amplifier should not be fixed in place inside the box. It will be easier to achieve low signal levels without the additional internal gain. If desired, a few more decibels of RF drive to the mixer can be achieved by adjusting the resistors in the attenuator at the RF port. The MAR-8 in Fig 5 replaces the MAR-6 amplifier in Fig 2. For convenience and clarity, a composite diagram is included in Fig 6.

The frequency controller's input requires 0.1 to 1 V for proper operation. Use 0 to +4 dBm into 50 Ω. I had 40-dB gain amplifiers already built from another project, so I used those with an attenuator to set the level. The single MAR-8 shown in Fig 5 should do the job, however.

The exact frequency of the LO is not important. In one version, it is a 100-MHz crystal oscillator in a can for computer applications. I also had a 108-MHz crystal that was convenient. Of course, if a different VHF oscillator is used, the VHF oscillator's low-end tuning frequency must be near the LO frequency to obtain near-zero output frequencies from the mixer. To use the full counting-range capability of the CM, I wanted a 50-MHz tuning range for the VHF VFO/VCO while keeping the LO frequency high enough to easily filter the LO signal out of the 0-50 MHz output. For signal quality, I wanted to keep the VHF VFO/VCO as low in frequency as possible, commensurate with the foregoing considerations.

The MCL JTOS-150 does a pretty

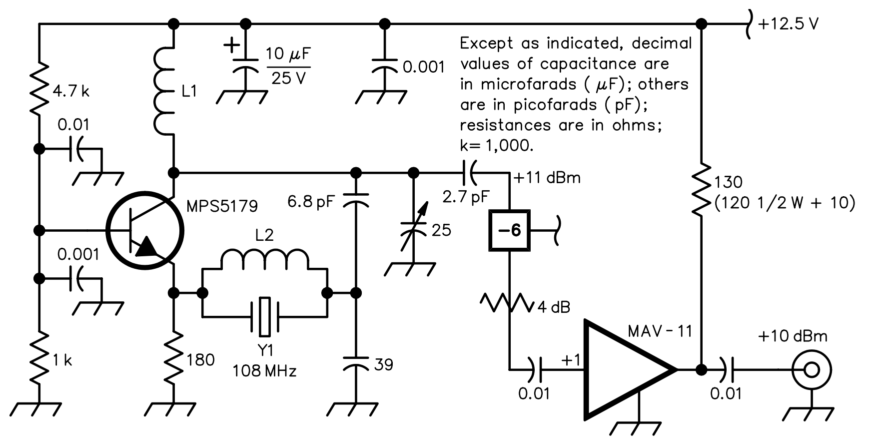

Fig 7—The fifth-overtone oscillator is taken from *The ARRL Handbook*. Cs (from transistor collector to the crystal) is 10 pF at 90-100 MHz and 6.8 pF at 100-108 MHz. L1 is 9½ turns (7½ for the higher range), 0.10-inch ID, tightly wound. L2 is 15 turns (13 for the higher range) of #28 enameled wire on a T25-6 core.

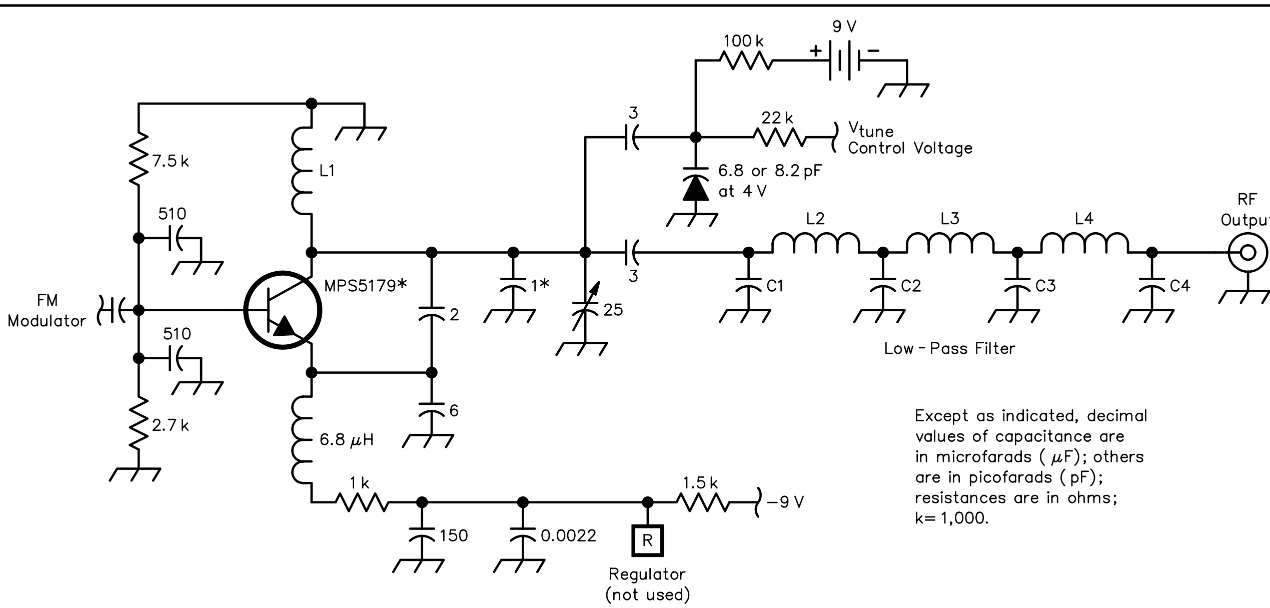

Fig 8—A mechanically tuned VHF VFO/VCO with small-range electronic tuning for higher circuit Q. Data were taken with a 2N5130 in place of the MPS5179. L1 is 3½ turns #18 wire, ¼-inch ID, 0.34 inch long. The tuning capacitor (labeled 25) in this version was an E. F. Johnson 193-6, 1.7 to 21 pF. *1 pF selected to adjust tuning range.

good job. If you want greater short-term stability (short-term drift within less than the CycleMaster control range for several seconds at a time) and better noise performance, a narrower-tuning-range higher-circuit-Q oscillator can replace the JTOS. A VHF oscillator was available, built for another project way back before there was an MCL company. It uses an E. F. Johnson air-variable capacitor and an air-core inductor, with a small-range electronic tuning diode to facilitate phase or frequency locking. This is not the best VHF oscillator, but it was easy to use for this project. It provides less phase noise and slightly better short-term stability (slightly less warble) than the JTOS oscillator. It is a little better for use in up-conversion for a UHF or microwave source, so I tried a version of Figs 1, 2, 3, 4, 5 and 7 using this oscillator, as shown in Fig 8. In addition, this oscillator draws less than 2 mA, so it can be battery operated to test for reduced power-supply noise, ac hum and pickup. Simply using battery power does not prevent noise pickup. The circuit and batteries must be shielded from all forms of electromagnetic radiation to exhibit a low-noise signal. In the present mechanical configuration, it is microphonic and picks up physical noise from humming/buzzing equipment and from florescent lights, even when shielded in a die-cast box. However, it is somewhat more stable and less noisy than the wide-tuning range JTOS-150, so it was used in addition to the convenient and flexible JTOS.

I once worked on a 2-megabits-per-second digital radio modem breadboard lying open and unshielded on the bench-top. The modem's bit error rate was dependent on the florescent lights and other equipment running in the vicinity. It would run for hours at a BER of near 10^{-10} late at night or on weekends with the florescent lights off. During the busy day or with the nearby lights on, the BER would drop to 10^{-7}.

The highest frequency resolution is obtained by mixing the VHF signal down and feeding it into the frequency counter/comparator of the CycleMaster. When using a wide-range oscillator like the JTOS or POS-150, a lower LO frequency in addition to the 108 MHz is useful. Then the two LOs can be combined in the mixer and switched on or off. This provides ±10-Hz control over the whole range from 74 to 160 MHz. The CM counter works to about 51 or 52 MHz. This approach reduces warbling in the output because short-term drift is held to a tighter range. Fig 9 is an approach for combining two LOs. The dc power to the oscillators is switched to select one oscillator or the other, but the class-A amplifiers are left on to provide an absorptive, linear termination for the splitter. Some noise is added to the LO signal, but it is insignificant. The 108-MHz LO is used to generate the 0-50 MHz signal, while the 74-MHz LO is used to provide frequency stabilization when using the divider circuits below 108 MHz on the VHF VFO. If you take precautions regarding the LO signal at 74 MHz, the VHF VFO (whatever one is in use) can be used in the lower portion of its tuning range. This yields a better mixed-down signal between 34 MHz and 50 MHz (108 – 74 = 34). This is especially true of an LC oscillator where the high-C end has higher Q.

If additional LOs are not used, an amplifier and additional splitter can be added after the divide-by-four circuit. Then control will always be available below 108 MHz when a wide-range VHF VFO is in use (down to the 74 MHz of the JTOS or POS). The amplified leg of the addition goes to the CM frequency-control counter input; the other leg is an output at 18.5 to 40 MHz, as shown in Fig 2. Some means of switching between CM frequency-counter inputs would be needed in that case, probably by simply changing a cable.

Low-Pass Filters

Low-pass filter tables are given in *The ARRL Handbook*.[12] Mini-Circuits BLP-50 and BLP-150 filters were used in these tests because they were available and reduced assembly time. Both of the MCL BLP filters are a little low in cutoff frequency, since they are stock values. We need somewhat higher values in both cases. However, the BLP filters worked well enough to give the results shown over the ranges discussed here. These filters should be satisfactory:

160-MHz low-pass (Fig 8): L2, L3 and L4—4 turns #26 enameled wire, 0.125-inch ID, close wound, (approximately 0.08 (µH); C1, C4—22 pF; C2, C3—39 pF;

51-MHz low-pass: L2, L4—10 turns #26 enameled wire, 0.125-inch ID, close wound, (approximately 0.22 µH); L3—11 turns #26 enameled wire, 0.125-inch ID, close wound, (approximately 0.26 µH); C1, C4—62 pF; C2, C3—120 pF.

Suitable filters to follow the frequency dividers can be found in Chapter 30 of the *Handbook*. Two filters for each octave will suffice.

The CM frequency-counter amplifier used in the tests is shown in

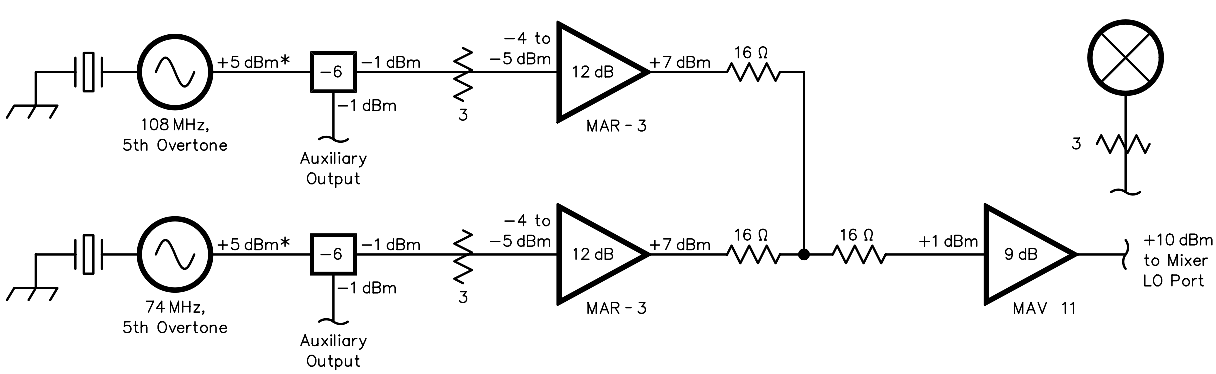

Fig 9—Two LOs combined using class-A amplifiers and dc switching to select one oscillator or the other. The amplifiers are always on. Some insignificant noise will be added to the LO signal. *Adjust coupling from the oscillators and the attenuators to set the power level to +10 dBm output.

Fig 10. This configuration allowed varying power levels during the experiments using high gain and a variable attenuator preceding the amplifier. This circuit is used in a phase-noise-measurement test setup, so it has 10-µF coupling capacitors to provide gain down to 500 Hz. The 10-µF capacitors are not needed in this application. The transmission line between the ICs provides physical separation to help protect against instabilities.

Performance

Phase Noise

Minimal phase noise was not a particular criterion for this general-coverage signal source. If a stable, quiet source of +17 dBm or more is required for a local-oscillator application, then a special-purpose source using a lower-frequency VFO might be a better choice. However, the phase noise of several different signal sources was measured or taken from data sheets and compared. See Fig 11. The measurements were made as described in my earlier article (see Note 15).

Noise on the divided-down frequencies was not measured, but it is quite predictable at close to –6 dB per octave of division. When divided by 16, to a 9-MHz output (144-MHz oscillator frequency), the wide-range VCO would be about –116 dBc/Hz at a separation of 20 kHz. The mechanically tuned VFO, divided down to 9 MHz, would be close to –146 dBc/Hz at a separation of 20 kHz. The short-term frequency drift or warbles within the control range are gone also. These results are within several decibels of the best that can be expected from a tunable local oscillator at 9 MHz. As we will see later, these numbers are better than a HP-8920A RF communication test set. Of course, the 8920 exhibits no warbling.

I plugged a MCL POS-50 (24-50 MHz) into the circuit of Fig 3 and ran it with the CM directly controlling the VCO. The signal was divided by four for a 6 to 12-MHz output. The signal quality is adequate for a local oscillator in a narrow-band radio. Divided phase noise is higher than a good fundamental oscillator, but it is a reasonable signal—stable with T9 tone quality. If divided by eight for a 3 to 6-MHz output, it is quite good. It was certainly easy to find parts and build it.

Almost none of the special noise-reducing precautions recommended by Mini-Circuits and others were applied in these prototypes. The noise results for the wide-range, varactor-tuned VCOs are probably near the worst case. A new assembly on a PC board will incorporate their recommendations, though, which follow.[13]

The phase noise generated by a VCO is determined by:

1. Q factor of the resonator
2. Q of the varactor diode
3. Noise figure of the active device used as the oscillating transistor
4. Power supply noise
5. External tuning-voltage supply noise

The noise contribution made by items 4 and 5 can be minimized by careful choice of the power supplies. The phase noise of the VCO is therefore determined primarily by the Q of the circuit overall. To design a circuit with high Q, the tuning bandwidth must be made small. Therefore, a VCO designed for good phase-noise performance will have a smaller tuning range.

Ways to Minimize Noise

The following steps are recommended for obtaining the best performance overall from Mini-Circuits VCOs:

1. Power Supply (V_{cc}) and tuning voltage (V_{tune}) returns must be connected to the PC board ground plane. The VCO ground plane must be the same as that of the PC board;

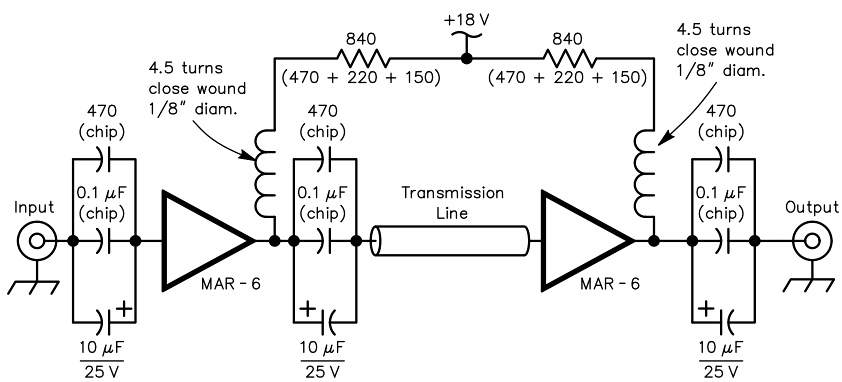

Fig 10—Preamplifier, 40-dB gain from 500 Hz to 500 MHz, with useful gain to 1300 MHz.

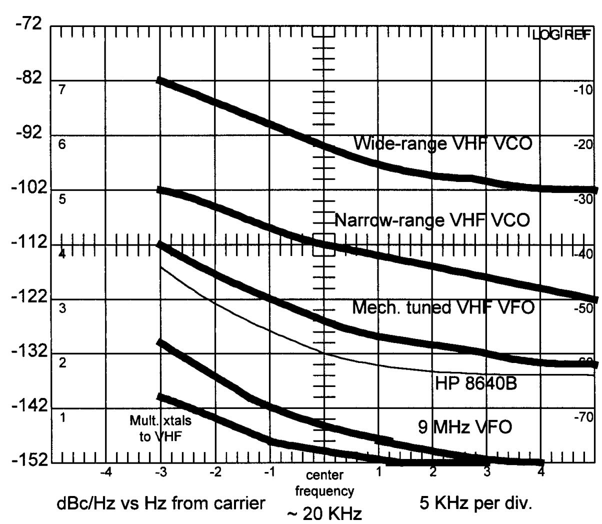

Fig 11—Phase noise of several different types of equipment. The VHF sources are for the low to middle parts of their frequency ranges (120-125 MHz). The HP-8640B data (the thin line) are for a fundamental of about 380 MHz. The 9-MHz VFO shown is a fundamental LC oscillator tuning over ±270 kHz.

therefore, all VCO ground pins must be soldered directly to the ground plane.

2. Adequate RF grounding is required. Several chip-decoupling capacitors must be provided between the V_{cc} supply and ground.

3. Good, low-noise power supplies must be used. Ideally, batteries for both supply (V_{cc}) and tuning (V_{tune}) voltages will provide the best performance.

4. Output must be correctly terminated. It is also a good practice to use a resistive pad between the VCO and the external load.

5. Connections to the tuning port must be as short as possible and with PC board partitions, shields and decoupling to prevent the VCO from being modulated by external noise sources. A low-noise power supply must be used for the tuning-voltage (V_{tune}) supply.

Comparison with the HP8920A RF Communications Test Set

A spectral plot of the HP-8920A with output on 9 MHz, taken with a HP-8561E spectrum analyzer, is shown in Fig 12. This shows about –91 dBc/Hz at a separation of 2.5 kHz. The mechanically tuned oscillator, built as shown in Figs 1, 2, 5, 7 and 8 and mixed down to 9 MHz, is almost identical, except that within 500 Hz of the carrier, the oscillator is less noisy than the 8920A. The JTOS version is a little better than the 8920A near the carrier, but several decibels more noisy further out. If the measures previously listed were taken, it would improve by several decibels. As stated though, minimum phase noise is not that important. Of course, as previously discussed, the divided-down signals are far less noisy. The noise performance at the oscillators' direct frequencies is the same as for the mixed down signal. The mix down does not add measurable additional noise at these levels.

A Heath HF vacuum-tube generator is less noisy than any of those so far mentioned, and is quite stable after a few hours of warm-up. However, ac hum makes it sound like a buzz saw above 5 or 6 MHz, and it is very difficult to set on frequency on higher bands. It is as microphonic as the mechanically tuned VFO. The unit is large, gets hot and does not lend itself well to stacking. In spite of all these negatives, I might have reworked it to reduce the hum and noise and to add electronic and fine-tuning were it not for the tubes and heat.

How do the signals sound in a communications receiver? A revealing spectral measurement can be made by listening to a signal source that is trying to earn the name "signal generator" in a narrow-band CW receiver. Some pieces of expensive professional-grade gear bearing that name from the not-too-distant past would not qualify for my list now. A subjective signal-quality evaluation system was derived from Amateur Radio's RST system. The *modified* rating chart is given in Table 1 with emphasis on the applicable ***TONE*** section.

The tone report refers only to the purity of the signal and has no connection with its long-term stability or freedom from clicks or chirps. Short-

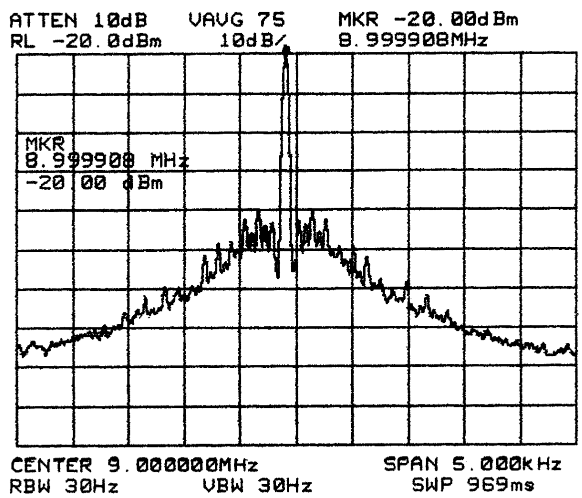

Fig 12—A spectral plot of the output from a HP-8920A RF Communications Test Set at 9 MHz using an HP-8561E spectrum analyzer.

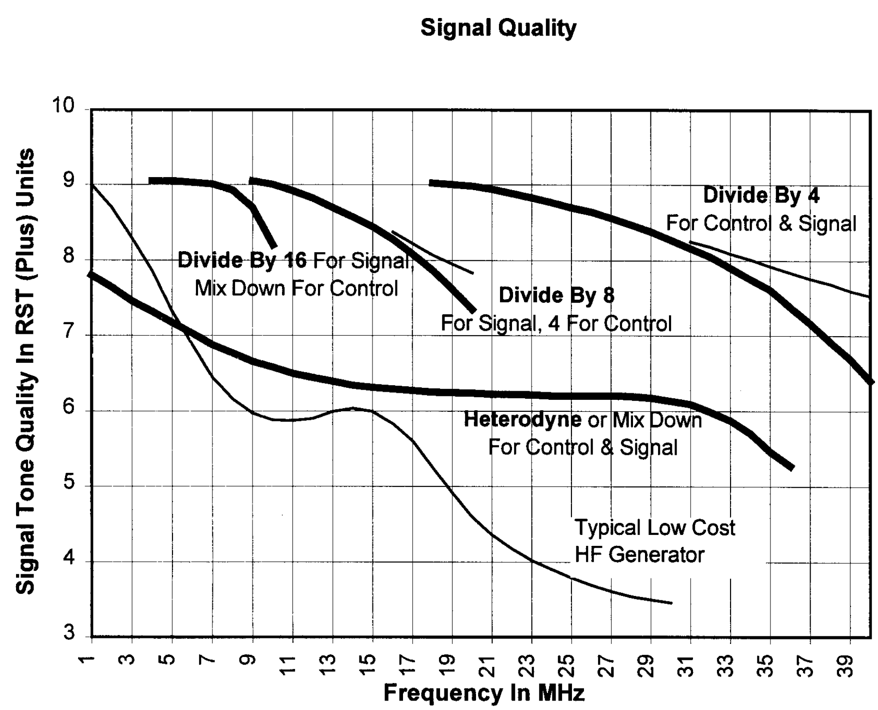

Fig 13—Signal tone quality in modified RST units (see Table 1) for several sources. For perspective, just about any synthesized generator or transmitter will yield a T9 tone, regardless of the phase noise. The fundamental 9-MHz LC VFO previously mentioned gives an excellent signal at T9.

term drift, or warble, in the range of from one tone shift per second to a few seconds between tone shifts was added to the tone report for this evaluation. If the signal has the characteristic steadiness of crystal control, add X to the report (for example, RST 469X).

Fig 13 gives the results of subjective tests on several sources. The heavy lines at the T8 to T9 level are for a JTOS-150 or the mechanically tuned VHF VFO using the divider circuits. The lighter lines (just above those) indicate results for heterodyned (mixed-down) frequency control and division of the output signal. The signal quality holds up better at the higher-frequency ends with heterodyne frequency control.

The heavy line in the middle, varying slowly from T8 down to T6 is for the heterodyned, or mixed-down, 0-50 MHz continuous signal. This is not as good, mainly because of translated hum, noise pickup and short-term instability within the control range. The lower thin line is for a Heath HF generator and the low ratings are caused primarily by hum and noise modulation of the otherwise good signal.

Recommendations

The circuits in Figs 1, 2, 3, 4, 5, 6, 7 and 9 provide both the heterodyned-down signal and the divided-down signals to use as the particular application might require. Using the POS or JTOS-150 and a single fifth-overtone LO will provide the continuous 0-50 MHz signal and will also allow heterodyne frequency control for the higher ends of the divider frequency ranges. Adding a second LO frequency will allow heterodyne frequency control at the oscillator's output frequency over the whole range of the POS oscillator.

Additional LOs will allow the VHF oscillator to be operated only in the lower end of its tuning range, yielding a better signal. This is particularly applicable to the mechanically tuned version. Low-pass filters to follow the dividers can be built as needed. The source can be frequency controlled by the CycleMaster, which is in a separate box. The CM can then be easily used for other applications simply by plugging in the control-voltage plug.

Up-Converting the Frequency-Stabilized VHF Signal to Higher Bands

If the source will be used for up-conversion to UHF or microwave frequencies, a narrow-tuning-range mechanically tuned VFO/VCO covering only, say, 108-128 MHz with a few megahertz of electronic tuning range can be used in place of the JTOS. This will provide the best possible signal on the higher frequencies, where a narrower range of coverage is acceptable. Up-converter LO crystals or a synthesizer LO can be changed to provide greater coverage.

One easy way to get the VFO signal on the 23-cm, 13-cm, 2400-MHz ISM and higher bands is to use the standardized Down East Microwave LO modules and transverter building blocks.[14] These, like the CycleMaster, are truly "plug-and-play." They require no-tuning and work well.

The 1152-MHz output for the LO

Table 1—The Modified RST System

Readability
1—Unreadable
2—Barely readable, occasional words distinguishable
3—Readable with considerable difficulty
4—Readable with practically no difficulty
5—Perfectly readable

Signal Strength
1—Faint signals barely perceptible
2—Very weak signals
3—Weak signals
4—Fair signals
5—Fairly good signals
6—Good signals
7—Moderately strong signals
8—Strong signals
9—Extremely strong signals

TONE
1—Sixty-cycle ac or less, very rough and broad
2—Very rough ac, very harsh and broad
3—Rough ac tone, rectified but not filtered. Considerable hum and noise
4—Rough note, some trace of filtering. Moderate hum and noise
5—Filtered rectified ac but strongly ripple-modulated. Some hum and noise
6—Filtered tone, definite trace of ripple modulation. Slight hum and noise, and warble
7—Near pure tone, trace of ripple modulation. Low hum and noise, only small warble
8—Near perfect tone, slight trace of modulation. Hum and noise not noticeable, only slight warble
9—Perfect tone, no trace of ripple or modulation of any kind. Clean, pure, stable tone, with rich, full, low frequency tone near zero beat
X—Characteristic steadiness of crystal control. Clean, pure, stable tone, with rich, full, low frequency tone near zero beat.

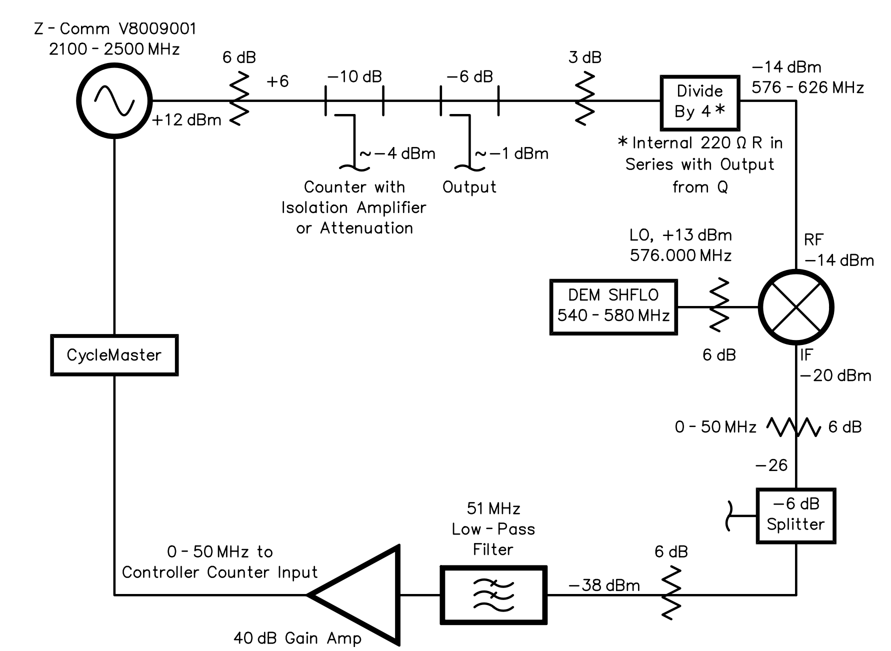

Fig 14—A 2400-MHz source with frequency stabilization at one-quarter of the output frequency.

injection in a DEM transverter module was used to up-convert the VHF-oscillator signals to a range of 1260 to 1310 MHz. A 96-MHz crystal was used. The same stability and signal quality from VHF was then available at 1260-1310 MHz. It is sure nice to have smooth tuning and precise control available in that frequency range.

A DEM 540-580 MHz SHF LO board driving a DEM ×4 multiplier board (to 2160-2320 MHz) was used to convert the VHF signal to the range of 2268-2478 MHz. An appropriate crystal in the SHF LO board in the range 90-96.667 MHz will hit any frequency within 2268-2478 MHz, in bands of 50 MHz or more, when mixed with the VHF output. This covers the 13-cm amateur band and the ISM band at 2400-2485 MHz, for testing components. Of course, the outputs contain the usual unwanted mixer outputs to be filtered out. Many times, filters are not even needed, if the components or equipment under test have sufficient frequency selectivity or if a spectrum analyzer is used to monitor and measure signals.

Replacement of the HP-8614A 800 to 2400-MHz Signal Generator

In addition to heterodyning the frequency-controlled VHF source—or any other suitable VHF or UHF source—to higher frequencies, an on-frequency oscillator can be frequency controlled with the circuits previously discussed. I do not have a manual for HP-8614A, but the FM input is apparently not dc coupled. If it were, it could be frequency-stabilized using that input (it needs it). The instrument can probably be modified to add electronic tuning, but it is still large, hot and noisy. The following discussion covers points regarding the replacement of the HP-8614A.

Controlling a VCO in the 2100-2500 MHz Range

I have several Z-Communications VCOs covering 900 to 2700 MHz. Three models are required to cover this range: 900-1900 MHz, 1600-2200 MHz and 2200-2700 MHz. I had a few of the V8009001, 2200-2700 MHz models, so these were used in tests for the 2400-MHz range. This unit was priced at around $55. It has been replaced by the V805 ME03 at $55 each ($39 each for five). The units I have tune from 2100-2600 MHz. These were built into a circuit as in Fig 3, except the resistors and pot in the tune-voltage string from 18 V should be 2.7 kΩ; a 10-kΩ, 10-turn pot and a 500-Ω resistor to ground. The 12 V to the oscillator was supplied by an LM317 fed from the 18-V supply. The design recommendations for low noise listed previously should be followed for a new assembly. The VCO was connected into the circuit shown in Fig 14.

The divide-by-four circuit is the same as in Fig 4 except the 51-Ω output resistor is 220 Ω. Other crystals can be used in the LO to provide 540-580 MHz injection that allows VCO control anywhere in the range of 2160-2520 MHz. This circuit allows fairly easy tuning below 2480 MHz and can be set to a predetermined frequency within about 5 kHz. With some effort, it can be set to within 1 kHz. There is some skittishness, a little

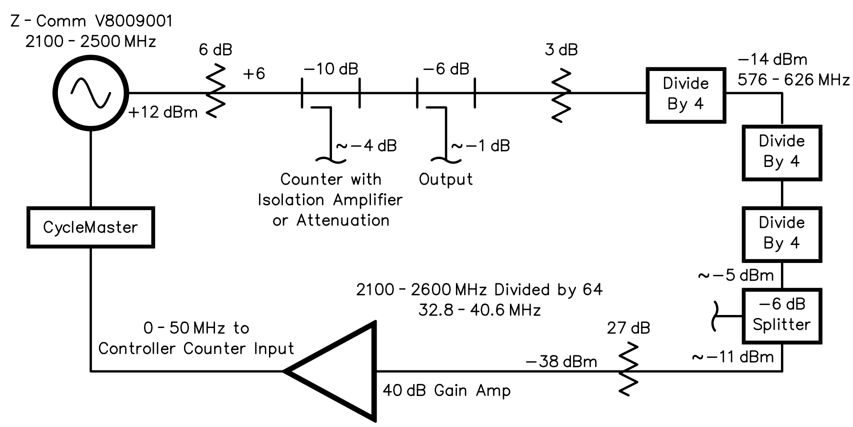

Fig 15—Using a divide-by-64 circuit to divide the control signal. The full range of the oscillator can be used. Splitters could be inserted between dividers to obtain additional frequencies. Two divide-by-two stages could be used in place of the first divide-by-four stage to give 1050-1300 and 525-650 MHz.

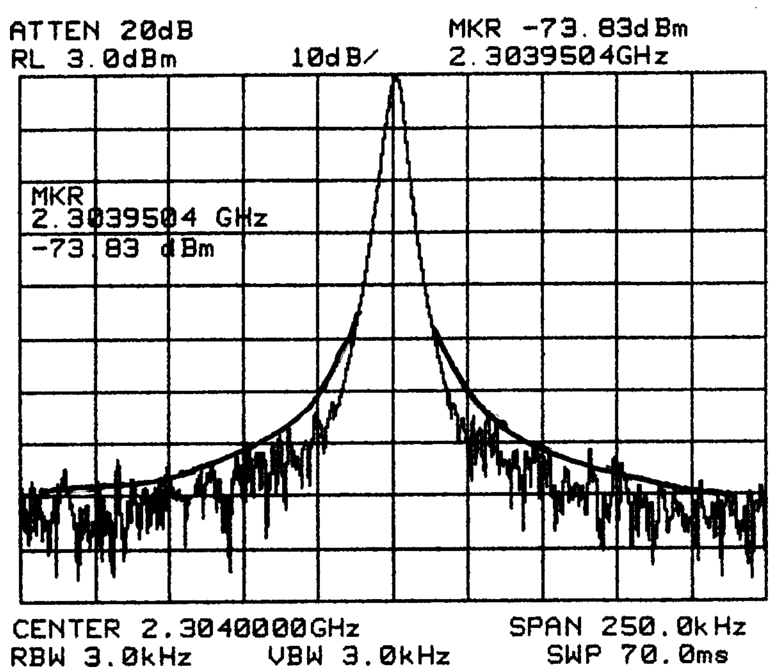

Fig 16—Spectrum plots of an HP-8648C Economy RF Signal Generator and two VCOs in the 900-2500 MHz range.

backlash or "rubberiness" because of the 35-MHz/V sensitivity of the VCO and caused by its inherent instability. The oscillator's output can jump considerably from one perturbation or another, and the CM must work hard to get control again. Once settled on a count in the CM, the locked frequency stays put, but it just may be a kilohertz or two from the desired frequency. The CM control-voltage input is divided down considerably by the 100 kΩ in series with the mechanical tuning circuit. The tuning rate is a little less than 1 MHz/V of CM tuning voltage. This slows the operation of the VCO considerably. It won't do for fast-frequency-changing electronic-countermeasures equipment, but is good enough for a cheap signal generator.

When set on a count, the frequency stayed within ±1 kHz most of the time, with a few excursions out to ±2 kHz. Frequency pulling caused by load pulling or SWR changes was corrected automatically—a relief from the constant knob-twiddling usually required on a manual oscillator at these frequencies. It can take many seconds to correct, depending on how far off it got pulled.

Additional divide-by-four circuits were added to achieve division by 64 and the down-conversion circuits were removed as shown in Fig 15. This simple arrangement actually provides easier "setability" than the previous arrangement. A desired frequency can be set within 100 Hz on the CM dial, and the output at 2400 MHz follows. Monotonic tuning with less skittishness and backlash is evident. The Startek counter used for these tests has a gate time of about 2.7 seconds for 100-Hz resolution at 2500 MHz.

Recorded frequency readings at 2470 MHz over ten minutes showed about half the readings within ±640 Hz or one control step (that is ±10 Hz ×64). A little over 80% of the readings were within ±1280 Hz, and 98% were within ±2 kHz of the average, 2470 MHz.

The same techniques can be used at frequencies between 0 and 2500 MHz. Any VFO/VCO in that range can be frequency-controlled or stabilized, and the outputs can be heterodyned or frequency divided for a variety of applications. I will need to experiment with higher-frequency VCOs, which tend to get flaky and require phase-lock techniques to get good results. Microwave-cavity-stabilized or dielectric-resonator oscillators, which are further stabilized by this frequency-control technique, are a possibility for flexible signal-generator applications.

Recommendations for the 2100-2500 MHz Range and Comparisons with Commercial Equipment

Test setups for component measurements (filters, couplers, attenuators, amplifiers, mixers) using wider-bandwidth monitoring and measurement instruments—such as power meters or spectrum analyzers—can make use of the simple and relatively inexpensive VCO and stabilization circuits described in this paper. Narrow-band or high-stability applications such as local oscillators or carrier generators should employ crystal-controlled frequency multipliers or PLL synthesizers for this frequency range, except for rough functional testing.[14, 16]

As the frequency range is lowered from here, performance becomes more like the VHF frequency-stabilized oscillators. A 24-50 MHz POS-50, for example, divided by four to the 6-12.5-MHz range (or divide by eight to the 3-6 MHz range) acts more like a crystal oscillator than a wandering VCO.

To replace the major function of the HP-8614A and extend coverage to the ISM band at 2400-2485 MHz, three VCOs are required, but they can all be mounted in the same small box and make use of common support circuitry.

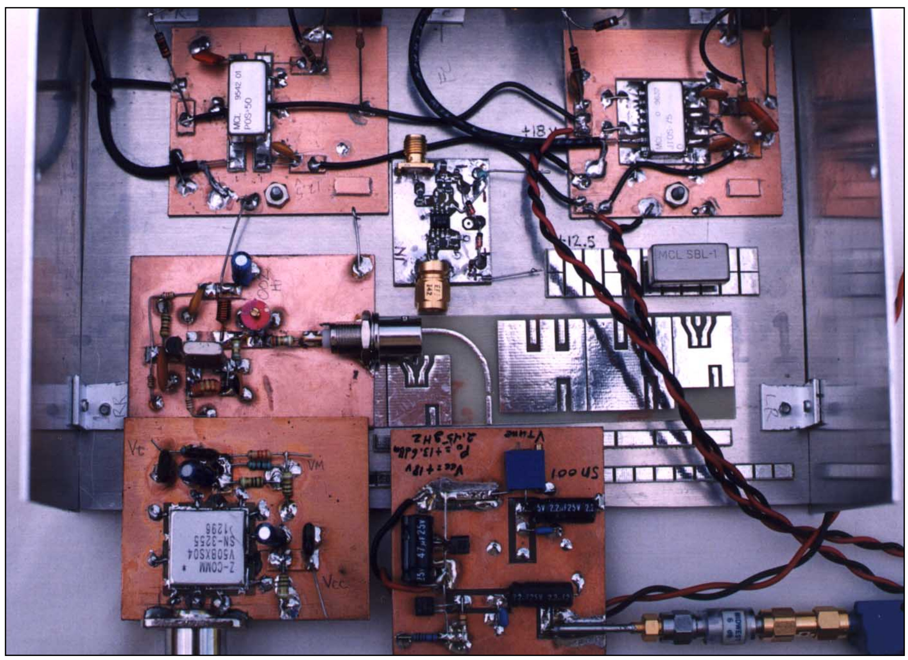

Fig 17—Some of the circuits used in the tests and evaluations of signal sources. The JTOS-150, in the upper right corner of the photo, is mounted on a small circuit board made for the JTOS package on $^1/_{32}$-inch-thick single-sided FR-4. One of the small boards can be seen in the lower middle of the photo. A POS-50, in the upper left of the photo, is mounted on some standardized, utility-type, etched board pieces. These strips are sheared from a larger PC board with many of the strips on it. A couple of the strips can be seen in the lower right center of the photo. One has pads spaced on 0.400-inch centers, and the other strip has smaller pads spaced on 0.200-inch centers. They are made of $^1/_{16}$-inch-thick FR-4 material. Many circuits can be assembled using pieces of these strips glued to a ground plane as standoffs. One of the fifth-overtone oscillators is in the middle lower portion of the photo. It is built with the standoffs on a ground plane.

A HEL-33 divide-by-four IC is shown in the middle of the photo. It is built on a small PC board suitable for 4-GHz operation using $^1/_{16}$-inch-thick FR-4 single-sided copper material.

A VCO board is shown in the lower left of the photo. This 900-1900 MHz unit is assembled on a ground plane. A couple of the standoffs are used for V_{cc}, V_{tune} and a monitoring point. The ground plane is relieved (cut away) with a large drill bit under the signal pads on the VCO package, then the VCO package is soldered to the ground plane with several short wires. A 2100-2600 MHz VCO is in the lower middle. The VCO package is the same size as the one just described, but has pins on it instead of surface-mount connections. The package is on the board underside with the pins stuck through and soldered to the topside copper. This board has copper foil on both sides, tied together at a few points by the VCO grounding pins.

The circuits for an HF/VHF signal source can be integrated into a box of the size shown: approximately $5^1/_2 \times 7^1/_2$ inches.

The CycleMaster

The CycleMaster (CM) is a highly flexible frequency stabilizer and controller. It measures the frequency of a VCO, reads a shaft encoder, controls frequency, provides RIT capability, stores frequencies and setup parameters in any of 32 memories and displays measured frequency with many available offsets and multipliers. It does not require a controller preprogrammed with specific frequency limits or steps. The CM is a general-purpose controller that provides 10-Hz tuning resolution over a range from near dc to 50 MHz and works at higher frequencies with prescaler or mixing techniques.

Compared to PLLs and DDSs, the CM cannot reduce microphonics or noise near the oscillator frequency, nor can it switch frequency very quickly (in microseconds). The CM can tune in small (10-Hz) steps without adding noise or spurs in the process. It also provides crystal-like long-term stability.—*Bob Schetgen, KU7G, QEX Managing Editor*

Fig A—Front Panel of the CycleMaster. Notice that the seven-segment displays are inverted, so that the decimal point can serve as an A/B VFO and RIT indicator.

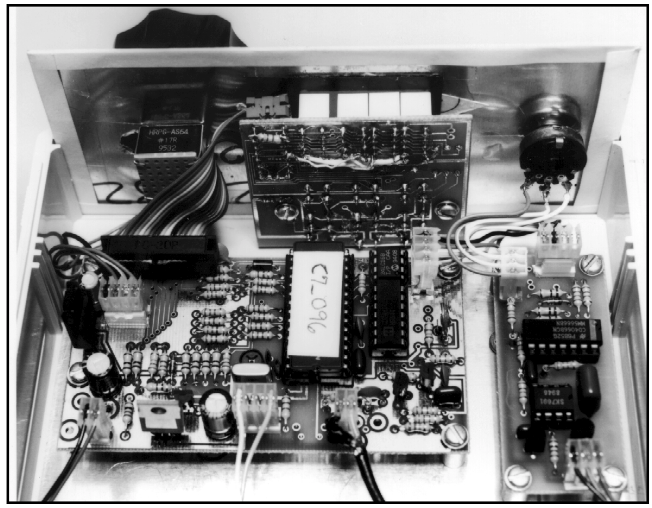

Fig B—An inside view of the CycleMaster. At top is the front panel. From left to right we see the shaft encoder, display/switch board and RIT pot. On the chassis from left to right: main logic board, integrator board.

By using two VCOs along with divide-by-two circuits, most of the range from 800-2500 MHz can be covered. The 8614A is not easy to set to a specific frequency, and it won't stay there for long. It does have a calibrated attenuator from 0 down to –130 dB. For calibrated low-level signals, a heterodyne approach will handle the job. In that way, calibrated attenuation is accomplished at the IF, rather than the output frequency, where variable attenuators are more difficult to obtain.

A spectrum plot of an HP-8648C Economy RF Signal Generator (about $8900 in the 1997 HP catalog) made with an HP-8561E spectrum analyzer is shown in Fig 16. The center frequency is 2304 MHz. A plot of the Z-Comm V8009001 is drawn as the solid line just above the printer data for the HP-8648C. A plot of a Z-Comm V50BXSO4 at 1600 MHz falls right on top of the 8648 data. The V50BXSO4 covers 900-1900 MHz. Noise performance of the VCOs might be better than an HP-8648C if the design recommendations for low noise listed previously in this report are followed. That will be done in a new circuit-board layout later. Right now, performance is good enough for the intended applications.

Ceramic-resonator oscillators for land mobile, cellular, PCS and other commercial applications in the 800-1100 MHz and 2400-2485 MHz ranges can provide fairly stable, easily controlled sources in the circuits of Figs 1, 2, 3, 4, 5, 7 and 8. Tuning ranges are much reduced because of the high Q of the ceramic resonator and the nature of the original intended applications. Some of those devices are inexpensive.

Conclusions

Compact, cool-running signal sources from 0 to 2500 MHz can be built in the home workshop with little effort and reasonable expense. Calibrated outputs with smooth, easy frequency setting can be achieved. Using a simple frequency-stabilization system, long-term stability can be nearly as good as a crystal source. Even at 2500 MHz, the frequency can be held to within less than 0.8 ppm plus the reference crystal drift. A number of programmable frequency settings can be stored in memory and two VFOs are available for each stored program. All components and materials required are readily available. The HF source with two LOs and frequency dividers can be built on a 5×5-inch circuit board (not including the various low-pass filters for the divided-down signals).

Notes

[1] A. Tilley, WM6T, "A Comprehensive Antenna Analyzer," *QEX,* Aug 1994, p 3.
[2] D. Kirk, WD8DSB, "The Ultimate VFO," *QEX*, Apr 1996, p 13.
[3] W. Cross, KA0JAD, "The Flexible Frequency Generator," *QEX*, May/June 1998, p 50.
[4] C. Preuss, WB2V, "Building a Direct Digital Synthesis VFO," *QEX*, July 1997, p 3.
[5] U. Hadorn, "Better Frequency Stability for the Drake TR7," *Ham Radio*, Aug 1987.
[6] K. Spaargaren, PA0KSB, "Frequency Stabilization of LC Oscillators," *QEX*, Feb 1996, p 19.
[7] J. Makhinson, "A Drift-Free VFO," *QST*, Dec 1996, p 32.
[8] P. Lawton, G7IXH, "The 'Fast' Digital Oscillator Stabilizer," *QEX*, Nov/Dec 1998, p 17.

[9]L. Richey, "The CycleMaster," *QST*, Sep 1997, p 37.

[10]Radio Adventures Corporation, RD 4, Box 240, Summit Dr, Franklin, PA 16323; fax 814-437-5432, **lee@radioadv.com**, **www.radioadv.com**.

[11]A circuit board for building an HF/VHF signal source is available from the author. The board is about 4.5×5 inches and uses mostly surface-mount parts (the larger 1206 size for easy assembly) and a JTOS-150 oscillator as discussed (other oscillators can be used as well). The PC board contains the circuits as shown in Figures 6 and 9 (dual LOs) and adds circuits for the low-cost 74VHC74 divider ICs as well as the faster HEL devices. Through use of the various circuits and outputs available on the circuit board—heterodyning down and up, frequency division, and direct signal output—stable signals of known amplitudes are available from a few kilohertz to 268 MHz in overlapping ranges. The PC-board source can be used with or without the CycleMaster. Write or e-mail the author for information at 15802 N 50th St, Scottsdale, AZ 85254; **bepontius@aol.com**. No phone calls please.

[12]R. Straw, N6BV, Editor, *The ARRL Handbook for Radio Amateurs*, 2000 edition, ARRL Order No. 1832. ARRL publications are available from your local ARRL dealer or directly from the ARRL. Mail orders to Pub Sales Dept, ARRL, 225 Main St, Newington, CT 06111-1494. You can call us toll-free at tel 888-277-5289; fax your order to 860-594-0303; or send e-mail to **pubsales@arrl.org**. Check out the full ARRL publications line on the Web at **http://www.arrl.org/catalog/**.

[13]From the Mini-Circuits Labs literature [also see Leeson's modified equation in *QEX*, May/Jun 1998, p 30, and Jul/Aug 1998, p 5.—Ed.]

[14]Down East Microwave, **www.downeastmicrowave.com**, frequency multipliers and LO modules.

[15]B. Pontius, N0ADL, "Measurement of Signal-Source Phase Noise with Low-Cost Equipment," *QEX*, May/June 1998, p 38.

[16]S. Rumley, KI6QP, "The KI6QP Dual Synthesizer Module," *QEX*, Jan 1996, p 3.

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He served as Vice-President of Engineering at E. F. Johnson Company for 15 years and worked with other companies in similar roles after that on development of communication equipment and systems. He has recently retired from Harris Digital Radio.

Mr. Pontius is now President of TRM Associates and works with various companies, assisting in technical and management consulting.

Mr. Pontius has a number of papers and patents in the communication field.

Bruce has been building and testing radio circuits since high school, and he was first licensed in 1978. His primary interests have been in test and measurement and in VHF and higher-frequency SSB radio circuits. □□

A 100-W MOSFET HF Amplifier

We had the power supply in Mar/Apr[1] and the diplexer filters in Jul/Aug,[2] here's the main event: a reliable FET power amplifier that needs only 10 dBm of drive to produce a pristine 100 W output.

By William E. Sabin, W0IYH

The two-stage amplifier described in this article and shown in Fig 1 is intended for SSB/CW/Data operation on all nine HF amateur bands. An input ($R_{in} \approx 50\ \Omega$) of about 10 mW (+10 dBm) is amplified to 100 W (PEP or average), continuous duty, with a gain of 40 ± 1.0 dB from 1.8 to 29.7 MHz. Third-order, two-tone intermodulation distortion (IMD) products are 35 to 40 dB below 100 W, and higher-order products are also within the high-quality range for amateur SSB equipment, as shown in Fig 2. The main goal for this amplifier is to operate as a driver, with low adjacent-channel interference, for a legal-limit 1500-W linear amplifier. At this power level,

[1]Notes appear on page 40.

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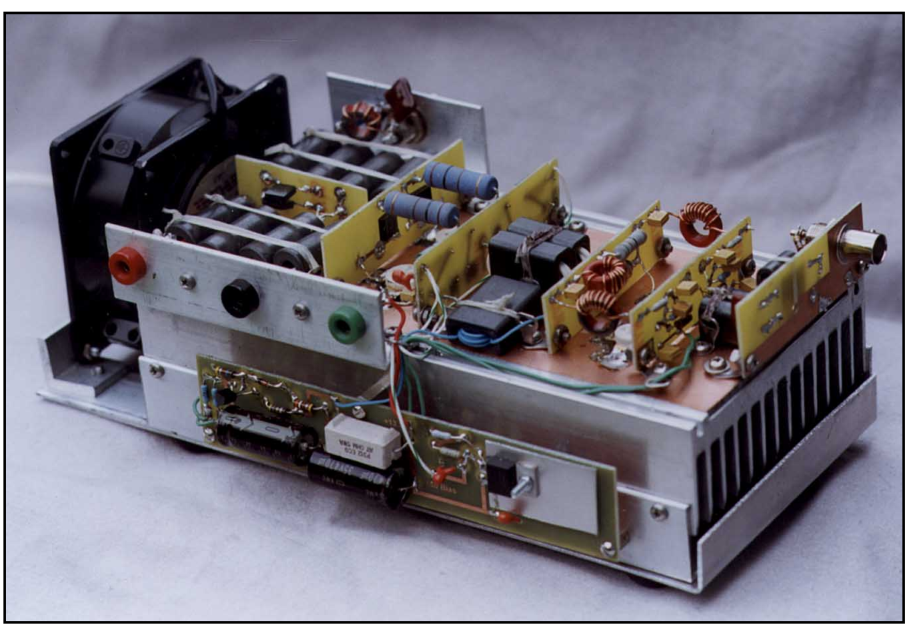


Fig 1—The 100-W broadband amplifier.

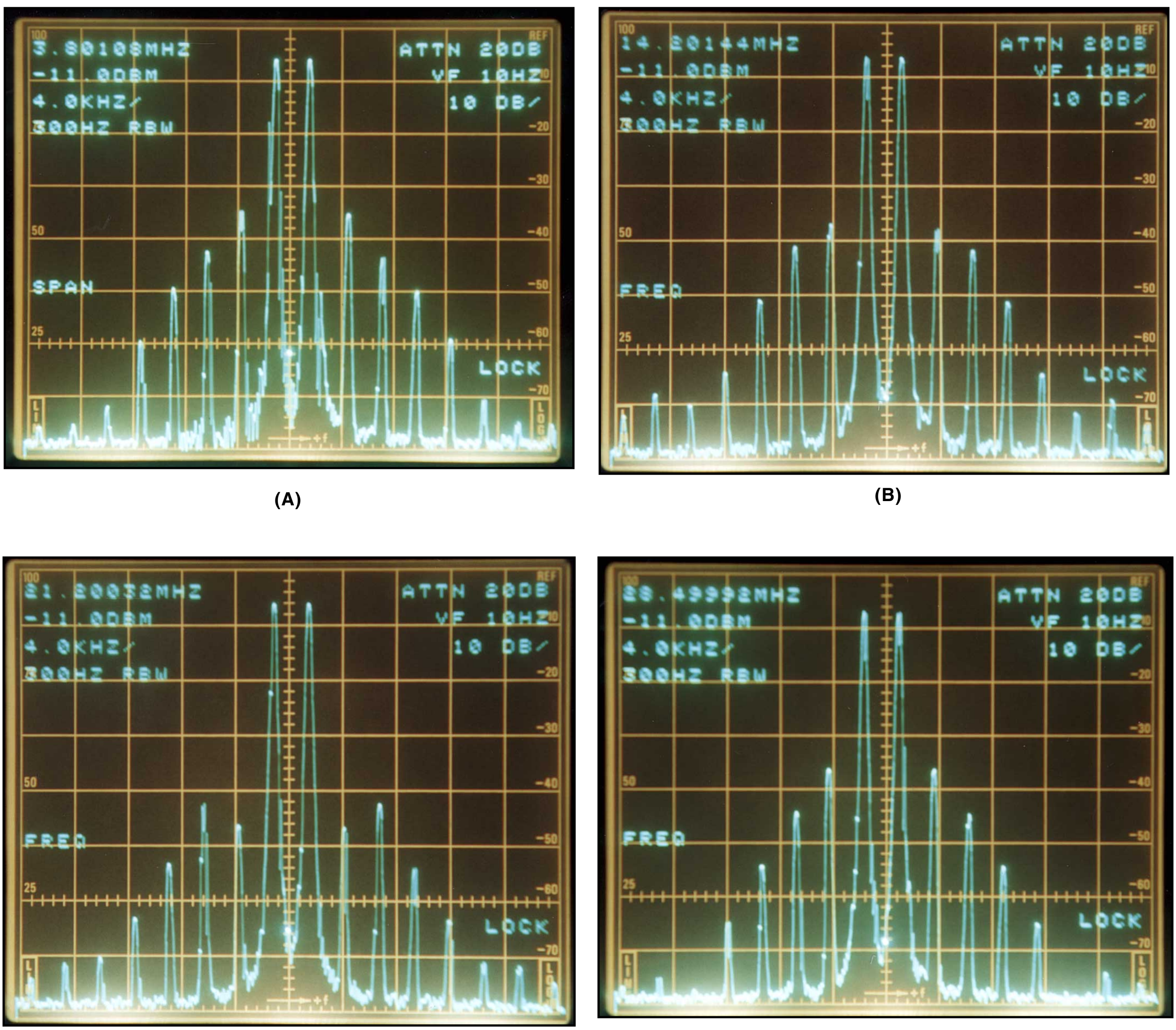
adjacent-channel reduction is especially important. And, of course, it is used in the "barefoot" mode as well.

The power supply (40 V at 8 A) for the push-pull, class-AB MRF150MP (matched-pair) MOSFET output stage was described previously (see Note 1). Six diplexer filters (see Note 2)—also known as invulnerable filters—provide more than adequate harmonic attenuation for all nine HF bands. They present a broadband load impedance to the MOSFETs that helps to assure freedom from regeneration and oscillation, and good IMD performance. A resistive load impedance between 45 Ω and 55 Ω is recommended for best performance.

The MRF150 was chosen because it is designed for linear, class-AB SSB operation and because it has high gain (g_m) at the 30-MHz end of the HF spectrum. A desirable feature of the MOSFET power transistor is its ability to achieve low values of the higher-order IMD products.[3, 4] As mentioned previously, these products contribute to adjacent channel SSB interference. The first stage uses high-gain, class-A push-pull MRF426 (matched-pair) BJTs that require 13.5 V at about 1.0 A from a separate supply. This supply is also the main supply for other system components. Matched pairs of both transistors are available for a small extra fee from at least two sources.[5, 6] They are both listed in the current Motorola manual.[7]

One main idea for this amplifier is to operate it in a very low-stress manner that helps assure a low probability of failure for a very long time, which offsets the initial cost of the high-quality transistors. The MRF150 is a 50-V transistor operated at 40 V; the MRF426 is a 28-V transistor operated at 13.5 V. The required input level is low enough that most of the amplification at the signal frequency occurs in one "gain block." Because of the good layout, circuit design and decoupling, the 40-dB gain value does not result in any stability problems.

The balanced amplifier greatly



(A) (B) (C) (D)

Fig 2—Two-tone IMD products: (A) 3.8 MHz, (B) 14.2 MHz, (C) 21.2 MHz, (D) 28.5 MHz.

reduces even-order harmonics—especially the second—prior to any output filtering, as shown in Fig 3 for a 7.0-MHz signal. It should be 40 dB or more below a 100-W CW signal for each amateur band. This reduction has been found reliable, once achieved. The low-level signal source that drives this amplifier must have at least 50 dB of second-harmonic attenuation, since this amplifier will not suppress that harmonic. This is easy to accomplish, but must be considered during the equipment system design (see Fig 9) and while bench-testing the amplifier as shown in Fig 4.

Circuit General Discussion

Fig 5 is the schematic of the two-stage amplifier. It utilizes 1:1 choke baluns and 1:4 (impedance) step-up and step-down transmission line (Guanella) transformers. The choke baluns significantly improve the balance of the input and output stages. Notice also that T2, T3 and T4 have "floating" center taps rather than bypasses to ground. This is recommended to improve the even-harmonic balance[8] (verified). T4 and T5 run slightly warm as compared to conventional transformers that get quite hot. The 5-W feedback resistors get warm, but their large surface area limits their temperature rise. They also receive cooling air from the fan. All power resistors in the RF circuits are the excellent metal-oxide types (see Note 5) that are quite stable with age and temperature, and have very low reactance at 30 MHz (measured). Metal-film 1% resistors are used in several critical locations.

The first stage has resistance and inductance loading from collector to collector. A powerful free-running oscillation in the first stage at about 32.5 MHz was being triggered occasionally at the moment of dc supply turn-on while a fairly large 28.0 to 29.7 MHz input was present. The loading reduces stage gain and suppresses the parasitic collector circuit resonance that produced the oscillation. It also helps to flatten the amplifier frequency response.

The second stage has a very low value of RF resistance, 15 Ω, from each gate to ground that helps to assure stability. Because these resistors are in parallel with the large input capacitance of the MOSFETs, they also help to flatten the frequency response. Both stages have negative feedback networks that further assure stability and flatness of frequency response.

Two biasing networks are employed. The LM317 provides a highly regulated gate voltage for the second stage. The

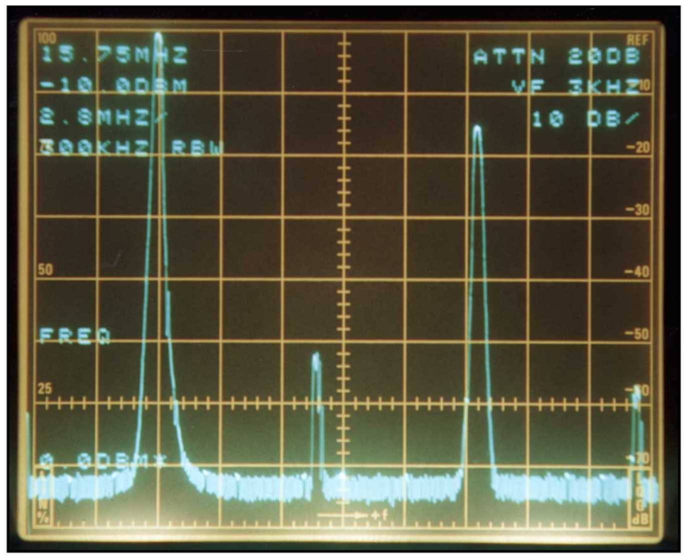


Fig 3—Wideband spectrum for a 7-MHz signal.

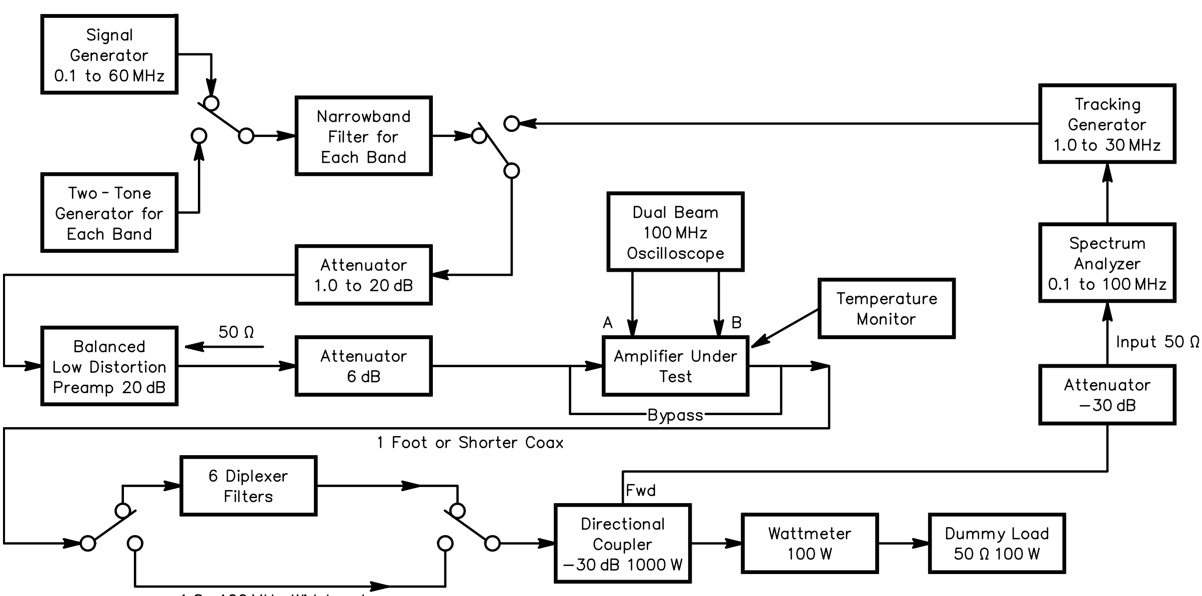


Fig 4—Block diagram for equipment setup, design and testing of the MOSFET amplifier.

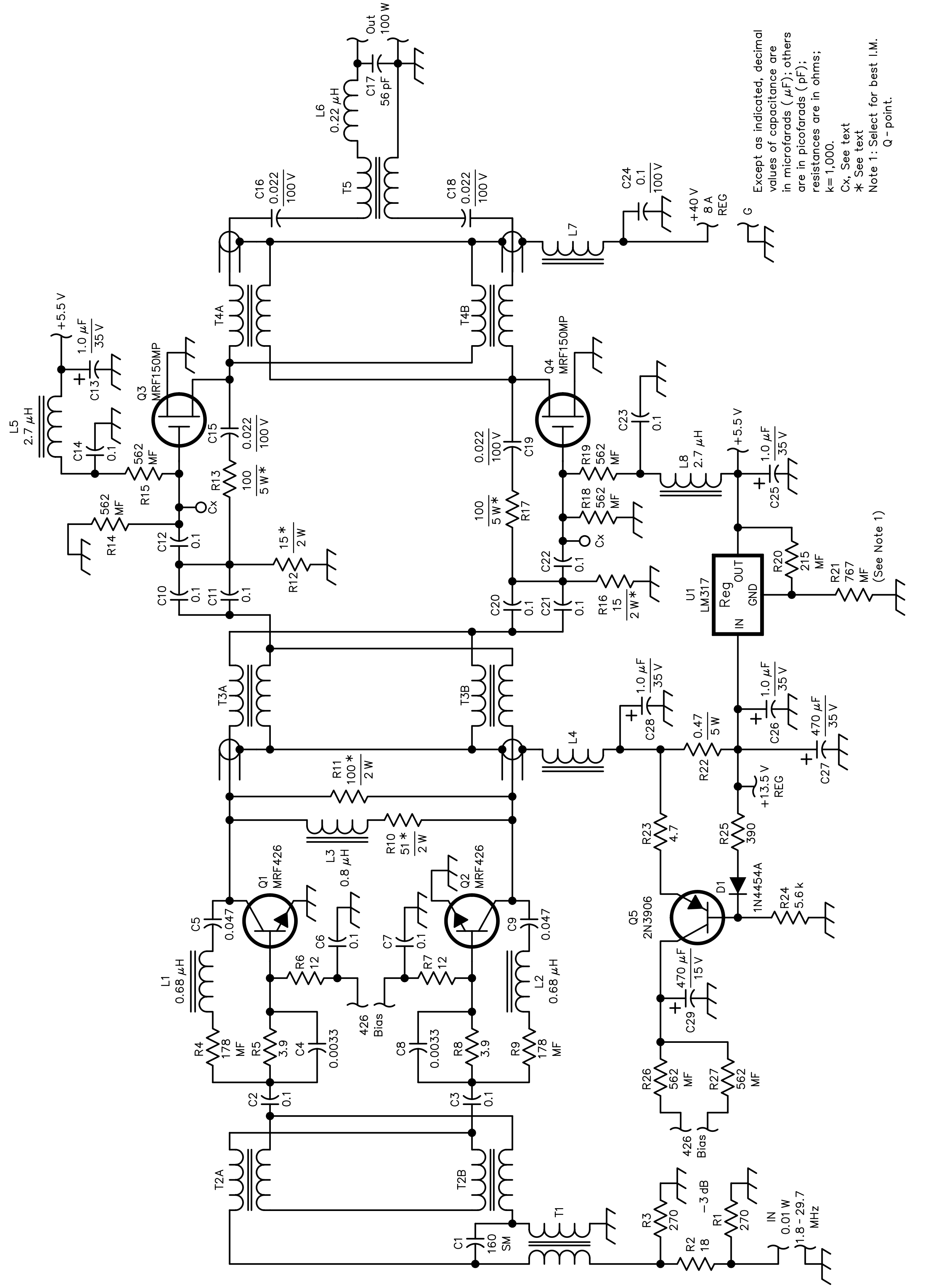

Out
100 W
IN
0.01 W
1.8 - 29.7 MHz
Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms; k=1,000.
Cx, See text
* See text
Note 1: Select for best I.M. Q-point.

Fig 5

FET dissipation and linearity are very sensitive to this voltage; the bias voltage for SSB is fine-tuned for best signal purity using two-tone tests and a spectrum analyzer. Note the four 562-Ω resistors. If the LM317 fails short circuit, the voltage on the FET gates does not exceed 6.8 V, which will not damage the gates. The large drain current that results from this failure also does no damage because the 40-V power supply has current limiting and voltage fold-back that prevent harmful FET dissipation. The FETs are thus kept well within the safe-operating-area (SOAR) as defined in data sheets[9] (verified). It is important that each 562-Ω resistor from gate to ground be permanently attached directly between the gate and source tabs of the FETs themselves so that the gates are never floating. Wrap the resistor leads around the tabs so that they cannot come loose. This avoids accidental static charges that might ruin them. On the other hand, it is better to "tack" and not wrap the base-collector and gate-drain signal leads so that they can be easily disconnected. We want to be able to test individual segments of the circuitry easily. In my experience, the MRF150 has proven to be a rugged transistor, much more so when operated conservatively and with the power-supply safeguards that I mentioned.

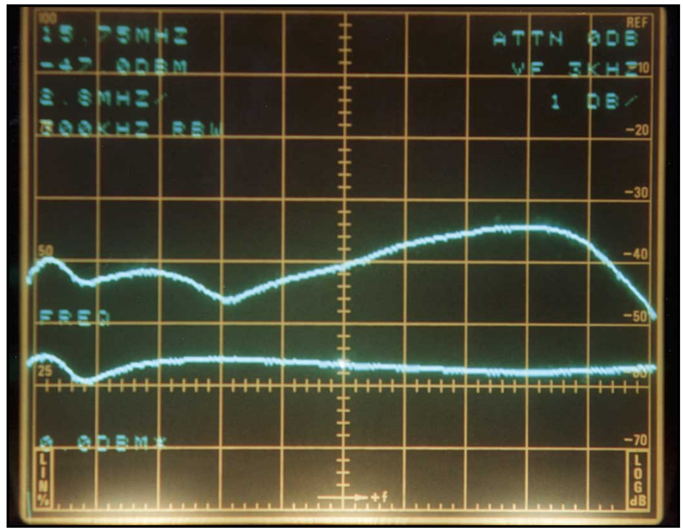


Fig 6—Gain variation, 1.8 MHz to 29.7 MHz, 1.0 dB per division vertical scale—The lower trace is a reference sweep that bypasses the 100-W amplifier—The response bump at low frequency is an artifact of the tracking generator.

The LM317 provides about 5.8 V for the particular pair of FETs that I used. Individual pairs of FETs will probably require an adjustment of this voltage for idling current and for IMD products that resemble Fig 2. The adjustment procedure is described later. This bias value is set to emphasize the reduction of the higher-order products rather than third- and fifth-order products. These lower-order products do not contribute as much to adjacent channel interference; in fact, an SSB speech processor will make them worse anyway. The higher-order products need to be reduced and the MRF150 has this capability.

The 2N3906 PNP transistor is a bias-current source. The value of this current is determined by the 4.7-Ω resistor and the base-to-ground voltage of the 2N3906. The two 562-Ω resistors force equal base currents into the MRF426s. This helps to assure equal performance. The 0.47-Ω resistor is part of a negative-feedback bias loop. If collector current increases, the voltage across this resistor increases and this reduces the base current. Thermal runaway is avoided by this strategy because the reduced base bias restricts the current increase to a small value. Because of the low dissipation of the MRF426s and the good heat sink, this method is very effective. These transistors are also well inside their SOAR requirements. This stage is very linear and contributes almost nothing to the overall IMD products. To verify this, it is necessary to turn off the drain voltage to the MRF150s and connect a spectrum analyzer across one of the 15-Ω resistors.

The diplexer filter method was decided upon—after a lot of experimentation with other low-pass filter methods—as a solution that is free of problems caused by complex interactions between the filters and the output transistors. A special peculiar problem is discussed later. The MRF150 has high gain into the VHF region; the diplexers eliminated all problems associated with this fact. This approach is recommended as a simple way to assure correct operation for HF amateur-band operation of these high-frequency MOSFETs. I was able to get "good-enough" operation with the more conventional low-pass filters, but this approach was by far the most satisfactory for an amateur-band amplifier, as confirmed by swept-frequency tests at all power levels into a 50-Ω load. More about complex loads later.

Frequency Compensation

The bipolar transistors have gain values that decrease as frequency increases. The MOSFETS have large capacitances that also affect frequency-response roll-off. It was a major exercise to design networks that flatten the response from 1.8 to 29.7 MHz. Fig 5 shows the approach. The idea is to compensate smoothly from input to output in such a way that neither of the two stages is over-driven at any frequency. The criterion for this is to check IMD products and harmonics during the design process, which involves approximate analysis (see the "MOSFET Stage Simulation" sidebar) and negative feedback.[10] Beyond about 32 MHz, the gain falls off fairly rapidly (but not too rapidly). This is also desirable.

When testing the frequency response using the test setup of Fig 4, it is necessary to measure the frequency response of the signal path from tracking generator to spectrum analyzer while bypassing the 100-W amplifier. This reference response is then compared with the response with the amplifier in the path, as shown in Fig 6. Note the vertical scale: 1.0 dB per division. For the most credible results and ease of measurement, it is very desirable that the impedance looking back from the input be 50 Ω.

Fig 5—(left) Schematic diagram of a two-stage, 100-W amplifier. Unless otherwise specified, use 1/4 W, 5%-tolerance carbon composition or film resistors. Resistors marked with an asterisk are metal-oxide units. MF indicates metal-film resistors. Both the MRF426 and MRF150 devices are matched pairs. All capacitors are 50 V, unless otherwise indicated.

Signal Level Testing

We want to verify that the first stage is operating normally by measuring its RF voltages at 4.0 MHz. The class-A first-stage input impedance is close to 50 Ω, and the input level is +10 dBm. Temporarily disconnect the signal leads to the gates of the FETs. The voltage at each output of T3 is about 1.35 V. Reconnect the gate signal leads. The final output over the entire range into an unfiltered, wide-band 50-Ω load is then observed using the setup in Fig 4. A spectrum analyzer with tracking generator is very valuable for this. The accuracy of the wattmeter should be verified or calibrated by some means at both the 100-W and the 25-W levels (the power of each tone of a two-tone, 100-W PEP signal) in each amateur band.

If the second stage is working correctly, the output power is 100 W, or 71 V RMS across 50 Ω. These procedures assure that both stages are working properly and amplifying as intended. Because of the variations in the fabrication of the transistors, the total gain can vary a decibel up or down despite the use of negative feedback in each stage. I suggest the 3-dB attenuator at the input not be modified for simplicity reasons.

The drain-to-drain load impedance of the class-AB output stage is 12.5 Ω and the CW output power is 100 W, so the drain-to-drain ac voltage is 35.4 V. Fig 7 shows the dual-trace scope waveforms (chop mode) on each drain, superimposed on the 40-V supply, and also the drain-to-drain waveform that is confined to the linear region. The third-harmonic content is visible. It is interesting to note that although these waveforms show considerable non-linearity, the fundamental component is quite linear with respect to the gate signal level. I also found that a 50-V dc supply created more heat, but at 100 W, did not improve linearity enough to make it worthwhile. The 35.4 V ac also appears across the two series-connected 100-Ω feedback resistors; they dissipate about $(35.4^2)/200 = 6.3$ W, or 3.2 W per resistor (64% of the 5 W rating).

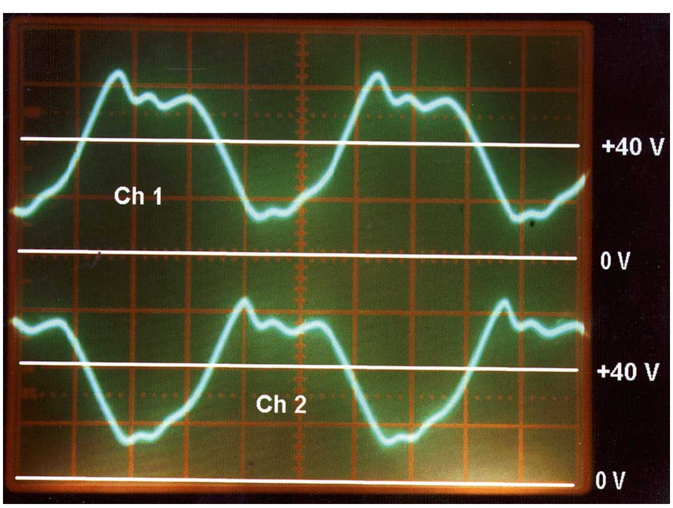
A broadband, untuned power amplifier with flat frequency response and low distortion is not, by necessity, especially energy-efficient (ratio of RF output power to dc power). The output stage is less than 40% efficient for this reason. For SSB use, where the average power is not more that 25 W, even with speech processing, this is no problem at all. For continuous key-down at 100 W, the cooling fan is more than adequate.

Here is an important caution about using oscilloscope probes at the FET drains and the output connector: A 10:1 probe could be damaged (it happened to me) if used directly at this RF voltage level, especially at the upper end of the HF range. Use instead the homemade probe described previously (see Note 2) with a 50-Ω terminating resistor.

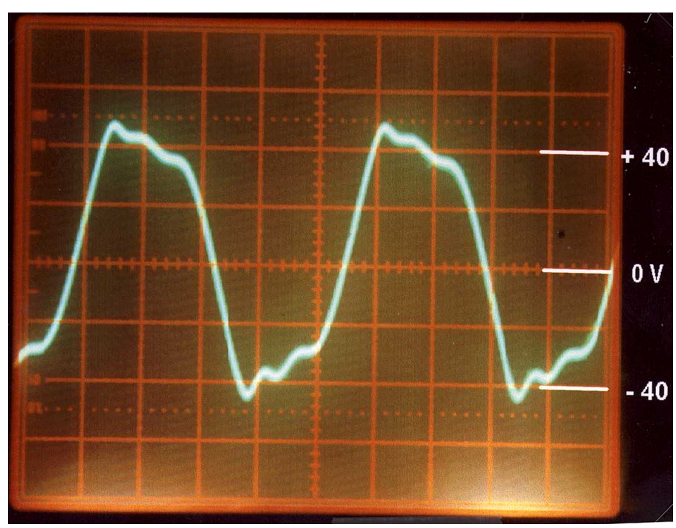
The quality of balance is checked by looking at the second harmonic on a spectrum analyzer using the setup in Fig 4 with the diplexer filters out of the signal path. At 100 W output, check the harmonics in each of the nine amateur bands. Capacitor C_x in Fig 5 is used (if needed) to improve the second-harmonic phase balance on the higher amateur bands. If the value is not well below –40 dBc, use C_x at one gate or the other to achieve –45 dBc. Extra pads are provided on the PC board for this capacitor. A value of between 22 pF and 56 pF should be adequate. Overkill is neither necessary nor desirable: The output filters will do the rest. For each decibel of power output below 100 W, the second harmonic "normally" drops about two decibels.

Construction Notes

Fig 1 shows how my version of the amplifier is constructed. A 0.125-inch (or 0.062-inch) double-clad PC board, 3.25×8.0 inches, is firmly attached to the heat sink. I suggest using this compact size for best reproducibility. The heat sink shown may not be presently available from RF Parts (see Note 5), but I also purchased a Model 99 sink from CCI (see Note 6) that is 6.5×12 inches. This can be easily tailored to the appropriate size with a band saw that has a metal-cutting blade. Pieces of angle and sheet aluminum can then be creatively fashioned to accommodate the fan (RadioShack #273-242) and the PC board that contains the bias circuitry. The transistors are bolted directly to the heat sink (through cutouts in the PC board) using heat-sink compound and tapped (and carefully deburred) #4-40 holes. Careful mounting of the transistors is essential. The main PC board has small sections of copper removed underneath the base and collector tabs of the MRF426s and underneath the gate and drain tabs of the MRF150s, so that accidental grounding is avoided. Use a hobby



(A)



(B)

Fig 7—(A) drain and (B) drain-to-drain oscilloscope waveforms.

knife to define the areas and a hot soldering iron to peel off the copper.

A bottom cover is advised, as shown, so that air is funneled through the heat-sink fins efficiently. For a 100-W, continuous, single-tone output, the exhaust air temperature reaches 50°C. I used a RadioShack #22-174 multimeter with its temperature probe mounted inside the fins to monitor temperature during the design and testing phase.

The components are mounted on a set of seven small PC boards, six of which are mounted vertically (as shown) and bolted to the drilled-and-tapped heat sink through the PC board, using #4-40 screws and small angle brackets. I used stiff, right angle #6 solder lugs that worked out very nicely. The seven PC boards are cut from a single 4×6-inch two-sided PC board, shown in Fig 8. The bias circuit board has a ground plane; the others do not. This set of boards is available from FAR Circuits.[11]

I chose this method because it is easy, makes the amplifier more compact, reduces stray L and C that can degrade the wide-band frequency response and reduces stray couplings that can impair stability and harmonic balance. It also allows the ground plane to be one continuous surface, which is a plus factor. This approach worked out very well and I recommend it as a simple approach.

Temperature Rise

The cooling fan is important. This amplifier, as designed, should have the fan running, and I have found that a *simple* and reliable way to keep everything safe. It has been tested at 100 W continuously, for several hours. The MOSFETs—well separated on a good heat sink and with efficient airflow—do not have a temperature-controlled gate-bias arrangement because I did not find it necessary at this power level. Because of the "spread" in threshold voltage of individual matched pairs, a one-time adjustment of gate bias is needed. The following simple procedure is used:

- Replace R21 with a resistor decade box set at 750 Ω.
- Place a 0-10 A meter in the +40-V line.
- With *no* signal input, switch on the amplifier and let the current reach its final value, which should reach about 1.5 A as the FETs warm up.
- Adjust the resistor value until I_{DD} is 1.5 A. After each adjustment, allow time for the FETs to reach a steady current value. This final value is not critical, but should be within 10%.
- Check the two-tone IMD at 100 W PEP (50 W average) and make further small changes, if needed, to resemble the IMD patterns in Fig 2.

In this amplifier, we are able to use the self-limiting feature of the MOSFET. This approach would not be appropriate in higher-power, higher-temperature amplifiers because of thermal runaway possibilities, where the decrease with temperature rise of the gate threshold voltage exceeds the decrease of the dc transconductance.[12, 13] If fan reliability is a concern, add a thermal cutout switch to the +40-V line.

Load Impedance

Swept- and fixed-frequency tests at many signal levels and voltage values—using various values of parallel conductance and capacitive/inductive susceptance—did not reveal instability problems at SWR values less than 2:1. A tendency to oscillate in a peculiar manner occurred in the 20.0 to 29.7 MHz range, but only with input signal applied (a "driven" oscillation)

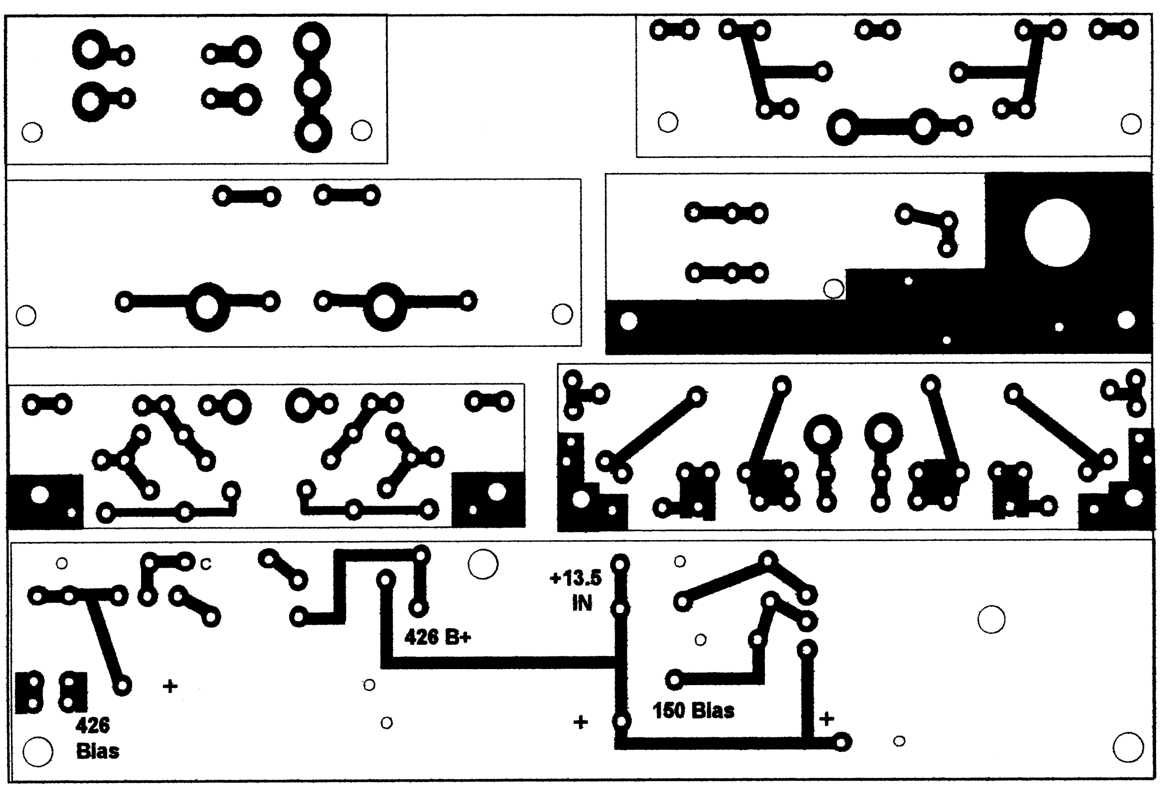


Fig 8—Etching pattern for a 4×6-inch (finished size) PC board that provides the seven individual boards.

and certain values of drive level, supply voltages less than 30 V, SWR of 3:1 or more, and a certain range of shunt capacitance at the output of the amplifier itself or at the output of the 12/10 meter diplexer. The instability shows up as a "Christmas tree" pattern in which discrete, uniformly-spaced 2- to 4-MHz sidebands appear on the main carrier. It is due to a parametric effect[14] involving the voltage-variable capacitances of the MOSFET transistors that is somehow enhanced by the shunt test capacitance. Shunt inductance did not produce the problem. The amplitude modulation of the carrier by the low-frequency oscillation is clearly visible on an oscilloscope. The low-frequency oscillation itself can be seen on a spectrum analyzer. At SWR values less than 2:1 and drain voltages greater than 35 V (40 V is recommended), the safety margin is large. The 0.22-µH, 56-pF network at the output greatly assisted with this problem. A coax cable from the PA to the diplexer assembly that is a foot or less duplicates my test setup. Under these conditions, there is no stability problem on any band caused by load-impedance values that my testing could identify.

Actually, a resistive load between 45 Ω and 55 Ω is recommended for best SSB linearity, as mentioned previously. While driving a high-power linear amplifier to its rated two-tone SSB output, adjust its input impedance to 50 Ω resistive as closely as possible on each band, using a 50-Ω directional wattmeter. A broadband (untuned) solid-state driver amplifier such as this one requires such attention to load value for best results as compared to pi-network vacuum-tube PAs that can transform a fairly wide range of complex load im-

MOSFET Stage Simulation

A simplified analysis of the second MOSFET stage is presented (Fig A) to illustrate the effect of the resistive negative feedback of the 100-Ω resistors and the 40-pF gate-to-drain capacitance. The simulation diagram shows the voltage-controlled current sources with a G_m of 6.0 S, which is assumed constant over frequency. The gate-to-source capacitance is 360 pF and the drain-to-source capacitance is 200 pF. These numbers are from the MRF150 data sheets.

The frequency plot (Fig B) obtained from the *ARRL Radio Designer* program shows the gain, MS_{21} (dB), with (lower trace) and without (middle trace) the 100-Ω feedback resistors. The gain variation is reduced from about 6 dB to about 1.5 dB. In this simplified model (good enough for this illustration), the gain drop is caused by the capacitors, especially the 40-pF capacitor whose influence is greatly magnified by the Miller effect. If this 40-pF capacitance is eliminated from the simulation, the frequency response variation—even without the feedback resistors—is less than 1.0 dB, as seen on the upper trace.

If the power output at 15 MHz is to be held constant, the turns ratio of the output transformer and the value of the feedback resistors can be manipulated to achieve that end. The simulation shows that a turns ratio of 1: $\sqrt{2}$ and a resistor value of 120 Ω does this. The FET load impedance is now 25 Ω instead of 12.5 Ω. The increased voltage gain makes the Miller effect greater and increases the gain variation (lower trace) to 2.3 dB. The drain-to-drain voltage is now 50 V instead of 35 V, and this results in higher drain efficiency for the MOSFETs. However, the push-pull 1: $\sqrt{2}$ -turns-ratio transmission-line transformer is not as simple as the 1:2 that uses 25-Ω coax, and for me, that was the determining factor.

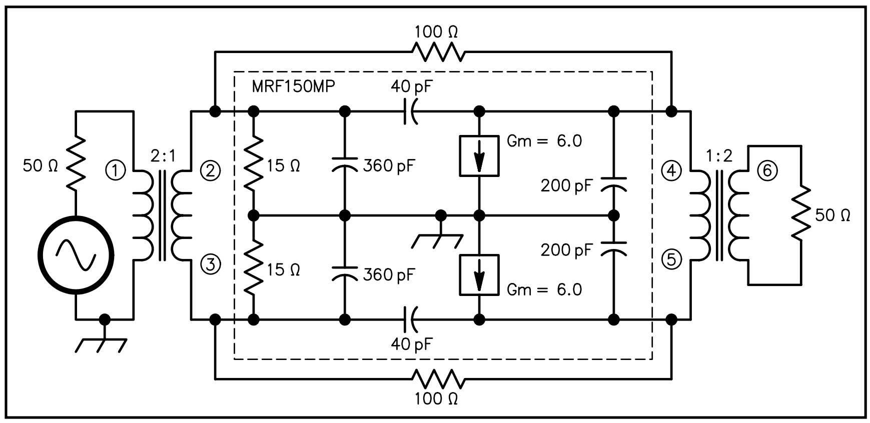


Fig A—A circuit used to analyze the PA output stage with ARD.

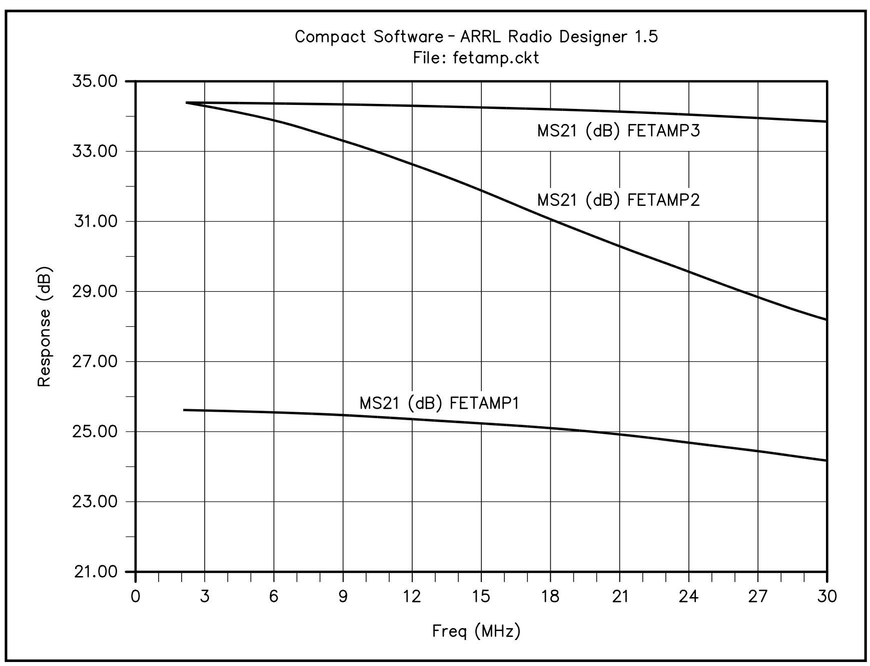


Fig B—The analysis results.

pedance values to the correct tube plate load resistance at resonance.

It is difficult to achieve stability in a broadband, transistor power amplifier. Having done so, it is still wise to operate into the correct 50-Ω load as closely as possible. The monitoring of forward and reflected power in Fig 9 is helpful for this and assures clean, stable and reliable operation. This circuitry will cut back the drive level to a value that protects the output stage from excessive drain current or gate drive. Incidentally, the first stage goes into saturation far below the level that would damage the MOSFETs.

System Design

Fig 9 suggests a system implementation of the amplifier. For the flattest frequency response of the complete system from 1.8 to 29.7 MHz, the circuitry that drives this amplifier should also have a flat response and a 50-Ω output resistance—both easy to achieve. If it is not perfectly flat or not exactly 50 Ω, it may be necessary to slightly adjust the drive level on each band, which is not difficult. As mentioned before, the second-harmonic production of the circuitry preceding this amplifier must be of sufficiently low level (–50 dBc) that balanced, push-pull operation is necessary. Narrow-band resonator filters are also very commonly used for amateur-band harmonic reduction in low-level exciter circuitry.[15, 16]

The directional coupler detects excessive forward and reflected power. Both of these are displayed on a panel meter. The directional coupler's forward port is also a source of ALC

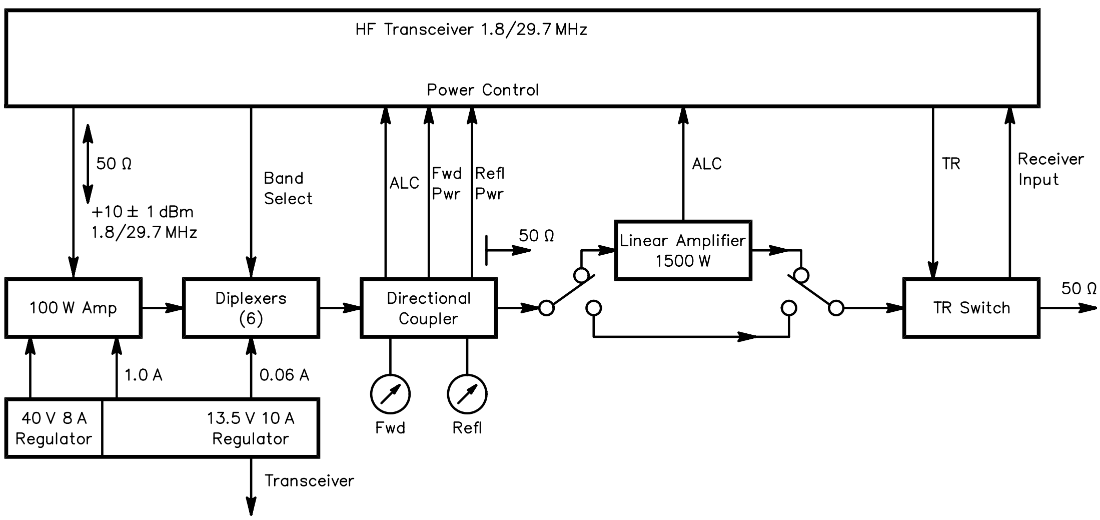


Fig 9—System implementation of the 100-W amplifier.

Parts List

C1—160 pF SM
C2, 3, 6, 7, 10-12, 14, 20-23—0.1 µF, 50-V CK05
C4, 8—0.0033 µF, 50-V CK05
C5, 9—0.047 µF, 50-V CK05
C13, 25, 26, 28—1.0 µF, 35-V tantalum
C15, 16, 18, 19—0.022 µF, 100-V CK05
C17—56 pF SM
C24—0.1 µF, 100-V CK05
C27—470 µF, 35-V aluminum
C29—470 µF, 15-V aluminum
D1—1N4454A or equivalent
L1, 2—0.68 µH, T50-2 core, 10 turns #26 AWG
L3—0.80 µH, T50-2 core, 12 turns #26 AWG
L4—BN-43-3312, 4 1/2 turns #22 hookup wire
L5, 8—2.7 µH molded
L6—0.22 µH, T50-2 core, 5 turns #26
L7—2 FB-43-5621 cores, 1 1/2 turns #12 stranded
Q1,2—MRF426 matched pair
Q3,4—MRF150 matched pair
Q5—2N3906 PNP
R1, 3—270 Ω, 1/4 W, 5% tolerance
R2—18 Ω, 1/4 W, 5% tolerance
R4, 9—178 Ω, 1/8 W metal film, 1% tolerance
R5, 8—3.9 Ω, 1/4 W, 5% tolerance
R6, 7—12 Ω, 1/4 W, 5% tolerance
R10—51Ω, 2 W metal oxide
R11—100 Ω, 2 W metal oxide
R12, 16—51 Ω, 2 W metal oxide
R13, 17—100 Ω, 5 W metal oxide
R14, 15, 18, 19, 26, 27—562 Ω, 1/8 W metal film, 1% tolerance
R20—215 Ω, 1/8 W metal film 1% tolerance
R21—767 Ω, 1/8 W metal film 1% tolerance, test selected
R22—0.47 Ω, 5W wire-wound
R23—4.7 Ω, 1/4 W 5% tolerance
R24—5.6 kΩ, 1/4 W, 5%
R25—390 Ω, 1/4 W, 5%
T1—BN-43-202, 2 1/2 turns #32 AWG, bifilar
T2A, B—BN-43-202, 2 1/2 turns #32 AWG, bifilar
T3A, B—BN-43-3312, 2 1/2 turns 25-Ω miniature coax*
T4A, B—(2) FB-43-5621, 1 1/2 turns 25-Ω miniature coax*
T5—(2) FB-43-5621, 2 1/2 turns 50-Ω miniature coax
U1—LM317 adjustable voltage regulator

Notes
*Microdot D260-4118-0000 available from Communication Concepts, Inc.
All cores are available from Amidon.
Closely matched transistor pairs are from RF Parts (see Note 5).

voltage. These control voltages are fed back to the appropriate gain-controlled stages in the exciter. Gain control of the transistors in the 100-W amplifier is not recommended because changes in base or gate bias will degrade amplifier linearity, IMD products and possibly the flat frequency response. There are better ways to accomplish the gain-control task, such as preferably in a low-level IF amplifier. With respect to IMD, we are trying to control the odd-order curvature of the MOSFET transfer characteristic—especially for the higher-order products—by setting the gate-bias point. The bias value is found by looking at IMD on each amateur band and selecting the best compromise value. This was discussed previously in this article.

The diplexer filters are linked to the other band-switch circuitry so that the correct filter is always switched in. The TR relay is at the output of the diplexers. Do not "hot-switch" the diplexer filters because this might damage the inexpensive relays.

Conclusion

In conjunction with the power supply (see Note 1) and the diplexer filters (see Note 2), the amplifier described here is a basic module for homebrew equipment that should satisfy the requirements for a 100-W power level, a 1.8 to 29.7 MHz bandwidth and a high-quality signal. It should run trouble-free for a very long time. The initial cost of the four transistors is about $180, but they will last indefinitely if cared for properly. The ones that I use were abused considerably during the experimentation and continue to work perfectly. The approach that I suggest is best implemented with diplexer filters and a preamplifier that has very low second-harmonic output and low IMD—both easy to get at +10 dBm. The payoff for me is excellent performance and reliability, once the design was completed.

This project is suggested for the homebrew enthusiast who has at least some part-time access to lab-quality test equipment. Others who do not care to build may find the article, together with numerous other sources, interesting background information regarding MOSFET power-amplifier design and test methods.

Any attempt at building and testing this amplifier should also use the 40-V power supply in the referent of Note 1 or something similar. The automatic current limiting at 8 A, the automatic reduction of drain voltage, the short-circuit limiting to 4 A and the manual control down to 24 V help to protect the MOSFETs from mishaps that are bound to occur.

Notes

[1]W. Sabin, W0IYH, "Power Supply for a MOSFET Power Amplifier," *QEX*, Mar/Apr 1999, pp 50-54.

[2]W. Sabin, W0IYH, "Diplexer Filters for an HF MOSFET Power Amplifier," *QEX*, Jul/Aug 1999, pp 20-26.

[3]N. Dye and H. Granberg, *Radio Frequency Transistors* (New York: Butterworth-Heinemann, 1993), pp 42-43.

[4]W. Sabin and E. Schoenike, *HF Radio Systems and Circuits* (Tucker, Georgia: Noble Publishing Co, 1998), Chapter 12, "Solid State Power Amplifiers," by R. Blocksome. This chapter is highly recommended regarding solid-state SSB power amplifiers. This book is also ARRL Order No. 7253. ARRL publications are available from your local ARRL dealer or directly from the ARRL. Mail orders to Pub Sales Dept, ARRL, 225 Main St, Newington, CT 06111-1494. You can call us toll-free at tel 888-277-5289; fax your order to 860-594-0303; or send e-mail to **pubsales@arrl.org**. Check out the full ARRL publications line on the World Wide Web at **http://www.arrl.org/catalog**.

[5]RF Parts Co, 435 S Pacific St, San Marcos, CA 92069; tel 760-744-0700, 800-737-2787 (orders only), fax 760-744-1943; **rfp@rfparts.com**, **http://www.rfparts.com/**.

[6]Communication Concepts, Inc. (CCI), 508 Millstone Dr, Beavercreek, OH 45434; tel 937-426-8600, fax 937-429-3811; **cci.dayton@pobox.com**, **http://www.communication-concepts.com/**; ask for a catalog.

[7]*Motorola Wireless Semiconductor Solutions, Device Data*, Rev 9, Vol. II; Motorola #DL110/D. The data book is available from RF Parts (see Note 5). (A Motorola Selector Guide is available at **http://www.mot-sps.com/books/sguides/pdf/sg46rev18.pdf**. There is a MRF150 page at **http://www.mot-sps.com/products/rf_and_if/rf_transistors/high_power/power_mosfets/mrf150.html**. A MRF426 datasheet (MRF426/D)is available at **http://www.zettweb.com/CDROMs/cdrom013/pdf/mrf426rev0.pdf**, which is *not* a Motorola site.—*Ed.*)

[8]Dye and Granberg, pp 113-115.

[9]*Motorola Wireless Semiconductor Solutions*; see pp 4.2-133 for the MRF151, similar to the MRF150. (See Note 7 for an MRF150 URL—*Ed.*)

[10]Dye and Granberg, Chapter 12.

[11]The set of boards costs $12 plus shipping and handling. FAR Circuits, 18N640 Field Ct, Dundee, IL 60118-9269; tel 847-836-9148 (Voice mail), fax 847-836-9148 (same as voice mail); **farcir@ais.net**, **http://www.cl.ais.net/farcir/**.

[12]Dye and Granberg, Chapter 4.

[13]H. Granberg, "Wideband RF Power Amplifier," *RF design*, Feb 1988; also *Motorola RF Application Reports*, AR313, p 424.

[14]Dye and Granberg, similar to p 124, Fig 7.7.

[15]*The ARRL Handbook* (Newington, CT: 1995-2000 editions), ARRL Order No. 1832; pp 17.60-17.66 and Fig 17.69. See Note 4 for purchasing information.

[16]W. Sabin, W0IYH, "Designing Narrow Band-Pass Filters with a BASIC Program," *QST*, May 1983, pp 23-29.

A High-Performance Homebrew Transceiver: Part 3

Mixing, premixing, dual receiving, IF shift and CW offset—all these topics are covered in this description of the 40-MHz RF board.

By Mark Mandelkern, K5AM

To many builders, the RF board in any radio is the most interesting. It includes the receive mixer, a prime component in determining the dynamic range of the receiver. The RF board in this radio also contains the transmit mixer, BFOs, premixers for LO injection, LO amplifier and the tunable noise channel. The basic radio covers 40-39 MHz; three front-end sections covering the ham bands from 160 to 2 meters are on separate panels. This RF board, part of the main panel, establishes the 40 MHz to 9 MHz transitions.

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Part 1 gave a general description of the K5AM homebrew transceiver, built for serious DX work and contest operating.[1] Part 2 described the IF board.[2] The RF board described in this article is shown in Fig 1.

Features

The main features of the RF board are:

- Balanced JFET receive mixer
- Balanced MOSFET transmit mixer
- Premixing for the IF-shift circuit
- Adjustable-waveform keying circuit
- Offset oscillator for panel-adjustable CW offset
- Tunable noise channel for the noncrunching blanker

[1]Notes appear on page 50.

Circuit Description

A general description of the RF board has been given in Part 1. The block diagram in Fig 2 shows the arrangement of the various RF-board stages. Figures 3, 4, 5, 6, 7, 8, 9, and 10 show schematic diagrams of the different RF-board sections.

BFO

See Reference 1 for a discussion of the mixing scheme. To provide IF shift (IFS) operation, the BFOs are premixed along with the PTOs to obtain the LO injection frequencies for the receive and transmit mixers. BFO frequencies are also routed to the AF board to provide LO injection for the product detector and balanced modulator.

The BFO circuit is shown in Fig 3. Although the BFO is tunable, a wide-range VXO circuit (with its inherent drift) was rejected in favor of a simple, adjustable crystal oscillator. A narrower tuning range results, but it is more than adequate for normal operating. For simplicity, two separate oscillators are used: one for USB/CW, the other for LSB. The components are selected to obtain the required IFS range. For normal SSB use, ±500 Hz is adequate.

The CW filters on the IF board have a center frequency of 8815.7 kHz. Thus, to provide full CW tone range for receiving, the USB oscillator must tune downward at least 800 Hz.

The circuit is somewhat unusual, as the variable tuning element is in the feedback loop. One would expect this to cause unwanted oscillator output-level changes as the BFO is tuned, but level changes also occur when pulling the crystal: The output drops with increased capacity across the crystal. With the varactor diode in the feedback loop, the feedback increases with increasing diode capacity; this counteracts the aforementioned effect. The result is much less output variation than that with the varactor across the crystal. The output is essentially constant over the range normally used in operation.

The transmit-frequency trimpot adjustments are critical. The BFOs must be positioned at the proper points on the SSB filter passband skirt to obtain the best transmitted-audio-frequency response. The initial settings have held within 10 Hz during the last seven years of operating; no doubt, the choice of quality crystals was a factor in this happy situation.

The simple IFS circuit allows BFO tuning while receiving and automatic return to the proper frequency when transmitting. When receiving, the µIFS control line is nominally –15 V. The IF-shift control on the front panel may then vary the voltage at terminal IFS1 from 0 to –15; this tunes the varactor diode VC1 in the oscillator tank circuit. At the same time, the transistor in the IFS circuit is cut off, and the **USB XMIT SET** trimpot has no effect. When transmitting, the µIFS line shifts to zero. Now the panel IFS control has no effect, but the transistor is turned on, allowing the **USB XMIT SET** trimpot to set the BFO to the proper frequency. When the CW offset spotting button is pressed, the µIFS control line also shifts to the transmit state to ensure proper offset adjustment. The Zener diode in the IF-shift circuit is needed because the control line shifts to –15 V only approximately. In practice, the op amps on the logic board that drive the control lines provide about –14 V. According to the op-amp data sheet, only –13 V can be assumed. The Zener holds the transistor's emitter voltage below –12 V, so the control line easily keeps the transistor turned off. The 10-kΩ resistor at the IFS1 terminal provides a load to ensure conduction in the diodes. Without this resistor, one may observe floating and drifting of the bias voltage applied to the varactor diode.

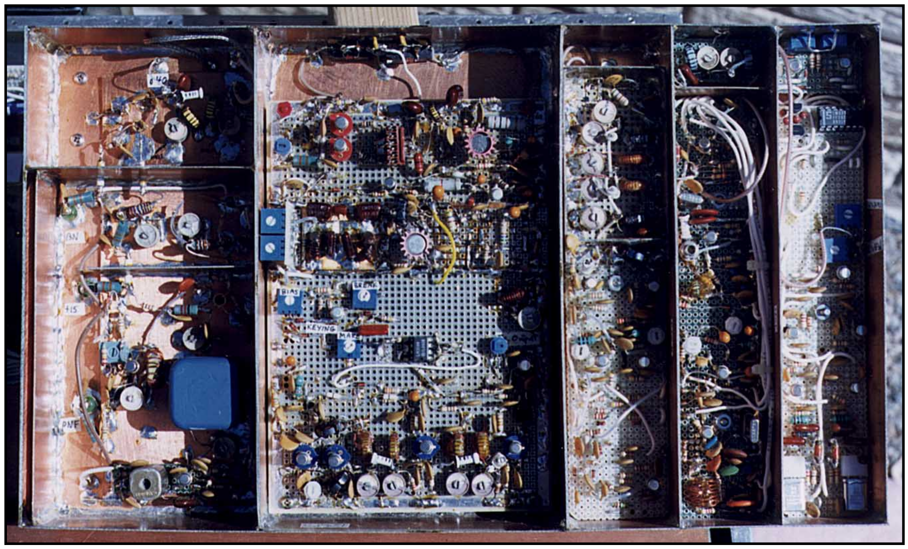


Fig 1—Top view of the RF board in the K5AM homebrew transceiver. From right to left, the sections are: the BFO, BFO mixer, PTO mixer, LO amplifier and main mixers and tunable noise channel. Several section-shielding covers have been removed for the photo.

BFO Mixer and Offset Oscillator

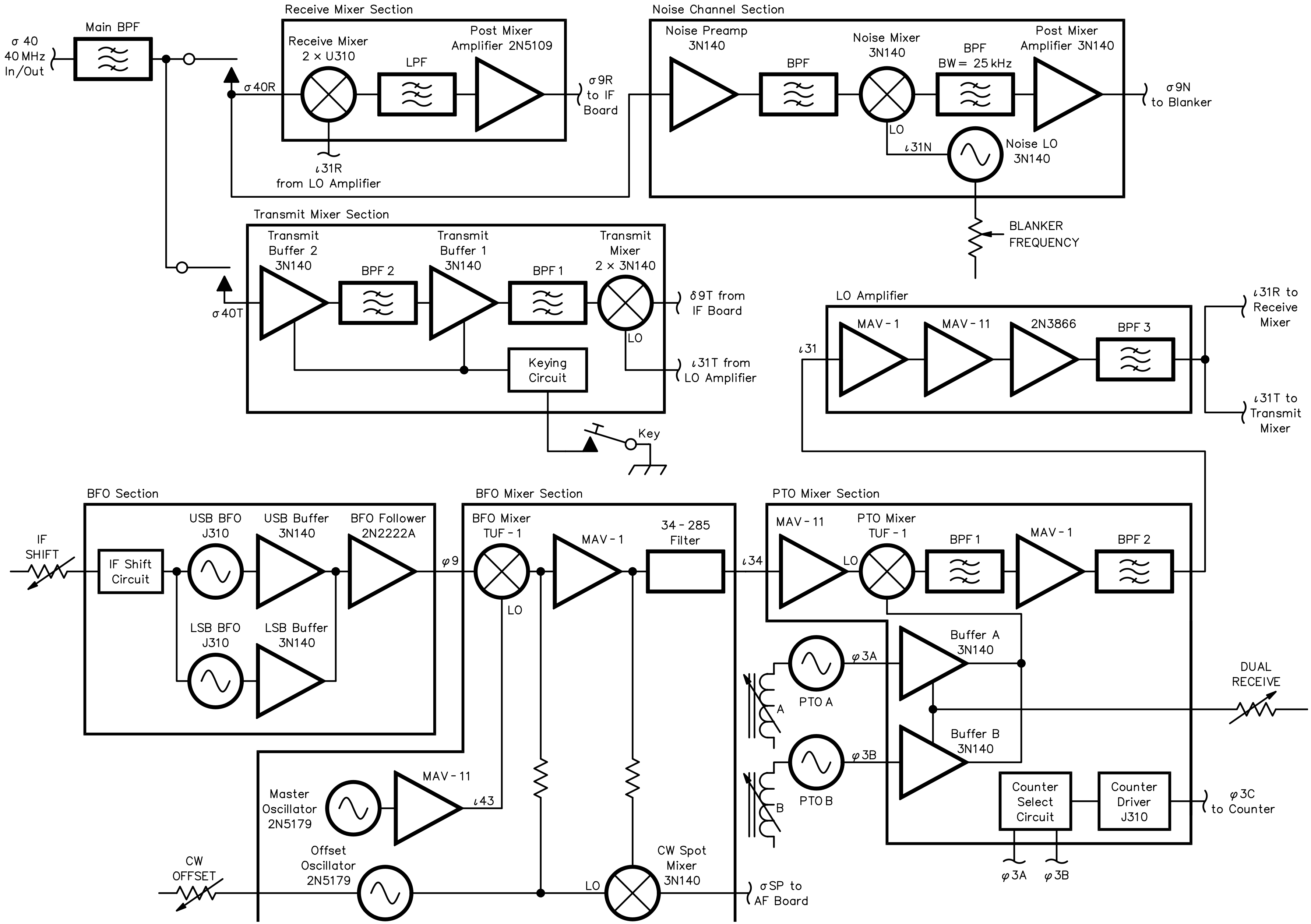
In this mixer, the BFO frequencies are mixed with the fixed 43.1-MHz master oscillator. The circuit is shown in Fig 4. The output of this mixer is nominally 34.285 MHz, shifting slightly with sideband selection and IF-shift operation. For CW-offset operation, this mixer is switched off (ignore the BFO frequencies here), and the panel-adjustable offset oscillator is used instead. Considerable temperature compensation is used to cancel the drift of the varactor diode in the offset oscillator.[3] The offset spotting mixer is enabled by the **CW SPOT** push-button on the front panel, which also enables both the normal BFO mixer and the offset oscillator. The resulting audio tone represents the actual offset; it is fed to the headphones by the AF board. A panel control sets the headphone level.

In the Signal/One CX7, the 43.1-MHz oscillator was tuned by a varactor diode, adjustable from the front panel. This was used in conjunction with an HF (100-kHz marker) calibration oscillator—itself adjusted using WWV—to calibrate the radio on each band. This resulted in some drift, destroyed the CW-offset settings and made band changing very inconvenient. Here the oscillator is fixed; the highest-quality crystals are used in the front-end sections for each band. This results in maximum convenience and read-out accuracy within 100 Hz. An important feature is the use of 10 separate oscillators on the HF panel in lieu of a crystal switch, which can cause frequency errors, instability and even total failure.

PTO Mixer and Dual-Receive Circuit

This premixer produces the final LO-injection frequencies for the main receive and transmit mixers. The cir-

Fig 2—RF board block diagram. Terminal σ40 is the 40-MHz input/output terminal from the rear panel. When receiving, the output is at terminal σ9R. For transmitting, the input is at terminal σ9T. Potentiometers labeled in all-capital letters are front-panel controls. An explanation of the terminal designations is given in Part 2, Table 1. The control lines are provided by the logic board.


Receive Mixer Section
Noise Channel Section
Main BPF
σ 40
40 MHz
In/Out
Receive Mixer
2 × U310
LPF
Post Mixer
Amplifier 2N5109
σ 40R
σ 9R
to IF
Board
ι31R
from LO Amplifier
Noise Preamp
3N140
BPF
Noise Mixer
3N140
BPF
BW = 25 kHz
Post Mixer
Amplifier 3N140
σ 9N
to Blanker
LO
ι31N
Noise LO
3N140
BLANKER
FREQUENCY
Transmit Mixer Section
Transmit
Buffer 2
3N140
BPF 2
Transmit
Buffer 1
3N140
BPF 1
Transmit
Mixer
2 × 3N140
σ 40T
δ9T from
IF Board
LO
ι31T from
LO Amplifier
Keying
Circuit
Key
LO Amplifier
MAV - 1
MAV - 11
2N3866
BPF 3
ι31
ι31R to
Receive
Mixer
ι31T to
Transmit
Mixer
BFO Section
BFO Mixer Section
PTO Mixer Section
IF
SHIFT
IF Shift
Circuit
USB BFO
J310
USB Buffer
3N140
BFO Follower
2N2222A
LSB BFO
J310
LSB Buffer
3N140
φ9
BFO Mixer
TUF - 1
LO
MAV - 1
34 - 285
Filter
ι34
MAV - 11
LO
PTO Mixer
TUF - 1
BPF 1
MAV - 1
BPF 2
Buffer A
3N140
φ3A
PTO A
A
DUAL
RECEIVE
Buffer B
3N140
φ3B
PTO B
B
Counter
Select
Circuit
Counter
Driver
J310
φ3C
to Counter
φ3A
φ3B
Master
Oscillator
2N5179
MAV - 11
ι43
CW
OFFSET
Offset
Oscillator
2N5179
LO
CW Spot
Mixer
3N140
σSP to
AF Board

cuit is shown in Fig 5. To minimize spurious responses, three band-pass filters are used for the LO injection: two here and a third in the LO-amplifier section.

The dual-receive feature is very simple; it is not equal to the much more elaborate subreceivers found in some contemporary commercial radios. The two PTOs each produce an LO injection frequency, so the receiver responds to two different frequencies at the antenna. Although simple, it can be very effective in certain DX split-frequency situations. With appropriate switching and combining of the external front-end sections, it can also be used to monitor two VHF DX calling frequencies simultaneously, or an HF and a VHF frequency. Control line µD energizes the **DUAL RECEIVE** control on the front panel. This applies gain-control voltage to the PTO buffers, resulting in adjustable balance control.

This section also contains a diode switch that routes the appropriate PTO frequencies to the counter on the front panel. Signals are selected by the readout-control line βR from the logic board. The control lines µIA and µIB control the PTO buffers and thus the received signal. This allows some flexibility. For example, PTO B can be read out and tuned while listening to a signal on PTO A.

LO Amplifier

The grounded-gate balanced JFET mixer used for the receiver has LO injection applied to the JFET source ter-

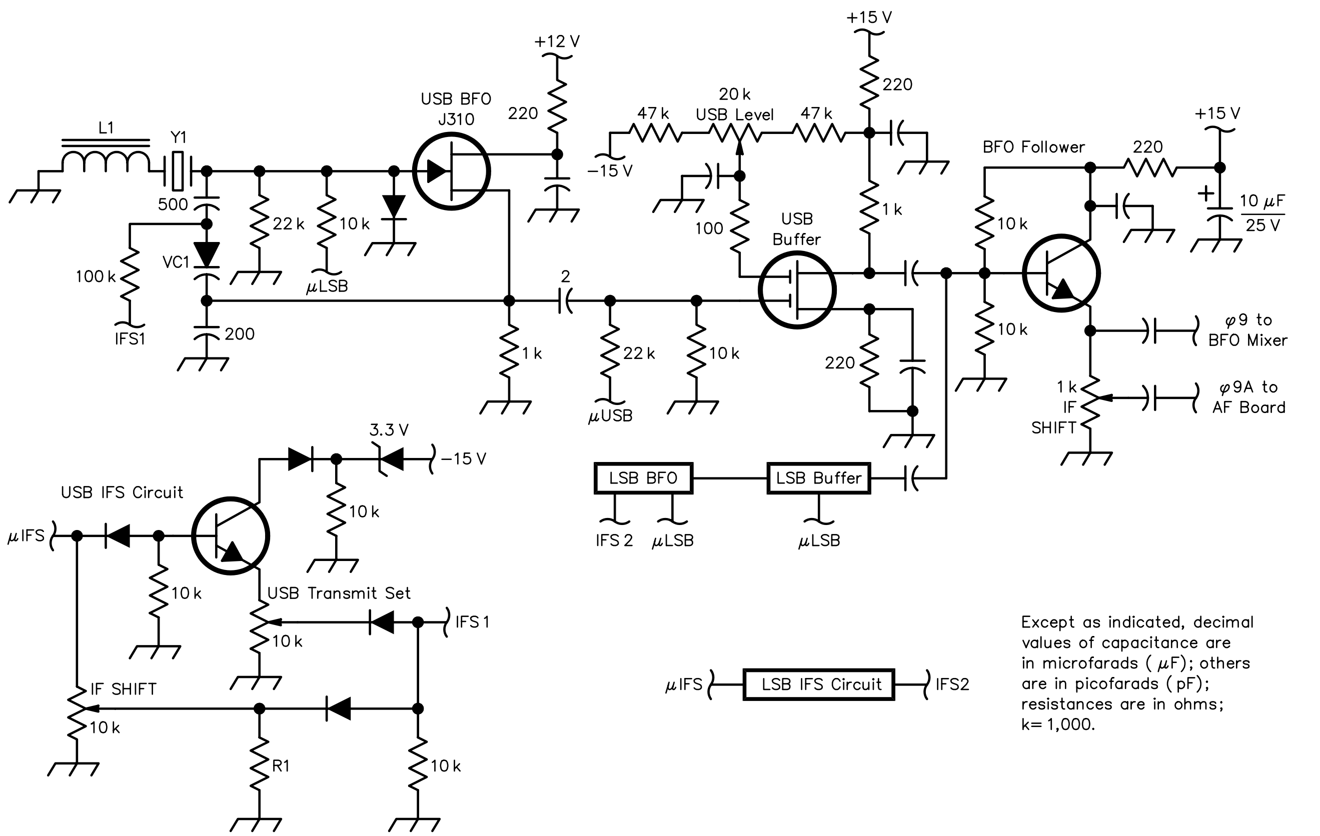


Fig 3—BFO schematic diagram. Except as noted, each resistor is a 1/4-W, carbon-film type. All trimpots are one-turn miniature types, such as Bourns type 3386; Digi-Key #3386F-nnn. (See Reference 8.) The unmarked coupling and bypass capacitors are all disc ceramic types; 1 nF in circuits above 30 MHz, 10 nF below. Also, each control and power terminal has a bypass capacitor that is not shown. Except as noted, the trimmer capacitors are Erie series 538, a sturdy 9-mm-diameter type that will withstand extensive testing and adjustment. These trimmers are available on the surplus market. (See Reference 9.) Xicon 7-mm ceramic trimmers are possible substitutes. (See Reference 10.) Except as noted, other capacitors are silver-mica types. Electrolytic capacitors are tantalum. Values of RF chokes (RFC) are given in microhenries. Potentiometers labeled in all-capital letters are front-panel controls; others are circuit-board trimpots for internal adjustment. All coils are wound with #26 enameled wire. The control signals are provided by the logic board. Some part designators differ from *QEX* style so they conform to the author's diagrams.

MOSFETs are small-signal VHF dual-gate types. Type 3N140 is used here, but any similar type may be substituted. Type NTE 221 is available from Hosfelt (Reference 11). Except where otherwise indicated, the diodes are all small-signal silicon types, such as 1N4148, and the bipolar transistors are type 2N2222A (NPN) or 2N2907A (PNP).

For clarity and to save space, LSB circuits (which are identical to the USB circuits) are indicated only as blocks. The only variation concerns the IFS panel control, which is a dual control (single shaft). The LSB section is wired so that the clockwise indicator arrow points towards ground. This is done so that the control functions the same on either sideband with respect to received audio passband.

L1—Small, molded RFC. Select, if needed, to adjust oscillator range; 1.6 µH used here.
R1—Select as needed to obtain center BFO frequency at center position of panel control. Used here: USB, none; LSB, 2.2 kΩ.
VC1—Varactor diode, nominal 33 pF. Motorola type MV2109. NTE type 614 (see Reference 11).
Y1—Fundamental crystal, USB/CW, 8816.5 kHz, type CS-1, ICM #433375-8.8165. Socket type FM-2, ICM #035007 (Reference 12).
Y2—Same as Y1, except LSB, 8813.5 kHz.

minals. It requires more LO power than other configurations. The transmit mixer also requires considerable LO power. The circuit for the amplifier that provides it is shown in Fig 6; the amplifier is operated in the linear region. The LO power for the two mixers is provided through adjustable trimpots.

Receive Mixer

The second mixer in a dual-conversion receiver is the most important single stage in the radio (see the discussion in Part 1). The circuit is shown in Fig 7. The singly balanced JFET mixer results in excellent dynamic range. The circuit is based on ideas found in a Siliconix manual.[4] There is no balance control; best mixer performance is achieved with a matched pair. A pair was selected from a batch of 10 devices. Best IMD performance and stability are obtained with the common-gate (grounded-gate) configuration.[5] In this configuration, the manufacturer suggests from +12 to +17 dBm of LO power for best performance. The dc source bias is set at 2 V, as measured at test point TP1. This dc bias is half the –4 V gate-cutoff bias of the selected JFETs; this allows full LO voltage swing without cut-off or gate conduction. The LO injection level is set for a dc reading of 1.1 V at TP2 (on a 10M Ω meter). This represents about 3.1 V P-P, or +14 dBm.

Ahead of the receive mixer and following the transmit stages are the main 40-MHz band-pass filter and the TR relays. These are shown in Fig 10. Mixer gain is only about 3 dB. A strong

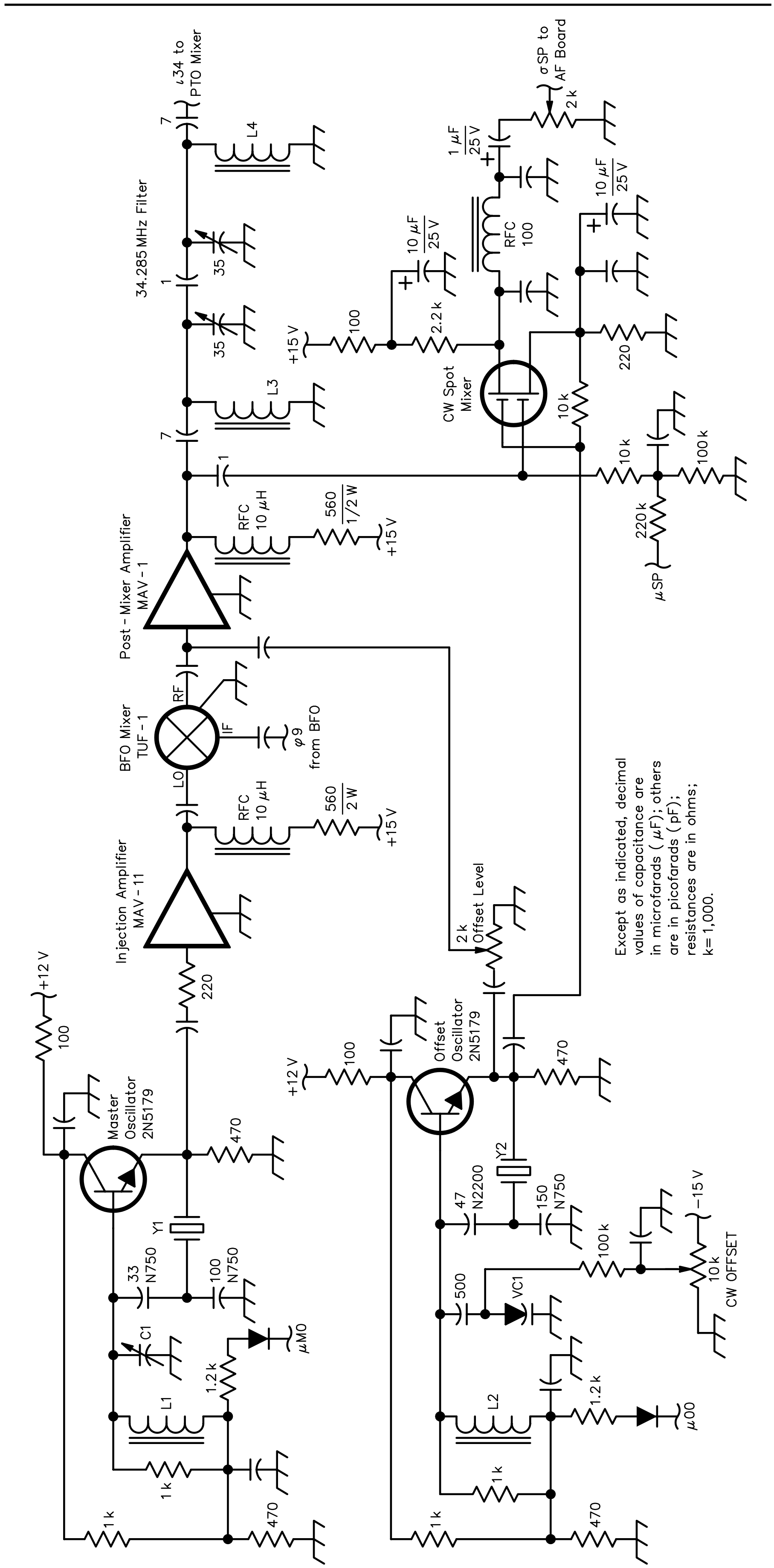


Fig 4—BFO mixer schematic diagram. For general notes on the schematics, refer to the caption for Fig 3. DBMs and MMICs may be obtained in small quantities directly from the manufacturer (Reference 13).

C1—Glass piston trimmer, 1-5 pF

L1—Master oscillator coil, 450 nH, T-37-17 powdered-iron toroidal core, 17 turns. Adjust turns to pull the crystal to the proper frequency. Compressing the winding or expanding to fill the core greatly affects the inductance. (See Reference 14.)

L2—Offset oscillator coil, 500 nH, same as L1 except 18 turns.

L3, L4—570 nH, T-37-6 powdered-iron toroidal core, 14 turns.

VC1—Varactor diode, nominal 6.8 pF. Motorola type MV2101. NTE type 610 (see Reference 11).

Y1—Master-oscillator crystal, third overtone, 43.1 MHz, type CS-1, ICM #471360-43.1 (see Reference 12).

Y2—Offset-oscillator crystal, same as Y1 except 34.2835 MHz; ICM #471360-34.2835 (see Reference 12).

bipolar amplifier follows the receive mixer. It overcomes filter losses at the input to the IF board and provides enough signal at the first IF amplifier to ensure high IF sensitivity.

Transmit Mixer

The balanced MOSFET transmit mixer circuit is shown in Fig 8. The mixer is followed by several buffers and careful filtering. There is no impedance matching at the mixer input, since high mixer gain is not required here. The voltage level on the σ9T input line is sufficient for the mixer gates. Keeping the level from the IF board high on this line minimizes the effect of any possible carrier leakage into the circuit. The total residual hum, noise and carrier on the transmitted SSB signal is more than 65 dB down.

CW keying is accomplished in the buffer stages. The keying waveform is adjustable using two trimpots in the simple timing circuit shown in Fig 9. The make and break trimpots act independently. The bias trimpot is needed because of the variation in individual MOSFET cutoff voltages. A fixed bias level high enough to accommodate any MOSFET would result in an unwanted lag between the key closure and the start of the transmitted element. The bias is set just high enough to achieve full cutoff. The timing trimpots are set for 2- to 3-ms make and break times. To ensure the absence of key clicks, the adjustment is made while monitoring with a receiver.

Tunable Noise Channel

The noncrunching noise blanker has been described previously in *QEX*.[6] It consists of several parts. The tunable noise channel is situated here on the RF board. The noise amplifier, pulse detector, signal channel and blanker gate are located on the IF board, described in Part 2. The circuit for the tunable noise channel is shown in Fig 10. The main 40-MHz filter and the 40-MHz TR reed relays are also shown on this diagram.

The MOSFET noise preamplifier, with its high input impedance, taps

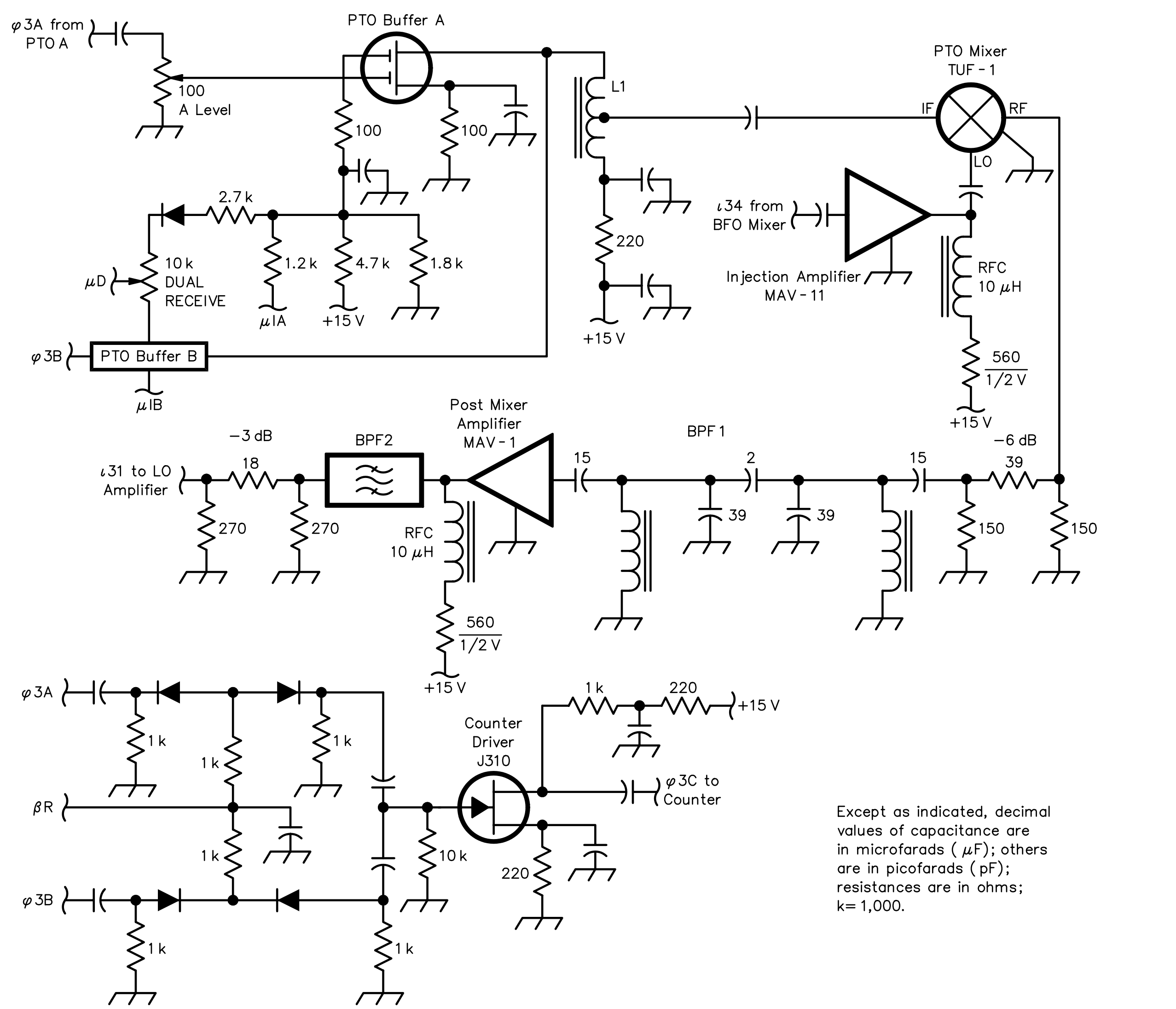


Fig 5—PTO mixer schematic diagram. For general notes on schematics, refer to the caption for Fig 3.

BPF1, BPF2—Band-pass filter, bandwidth 2 MHz, center 30.685 MHz. The coils are each 520 nH, T-37-10 powdered-iron toroidal core, 13 turns. Formulas predict 14.4 turns, but coils on these cores often have more inductance than indicated by the manufacturer's data. Adjust turns for proper inductance before assembly and again in-circuit with a sweep generator.

L1—FT-37-43 ferrite toroidal core, 12 turns, tap 3 turns from low end.

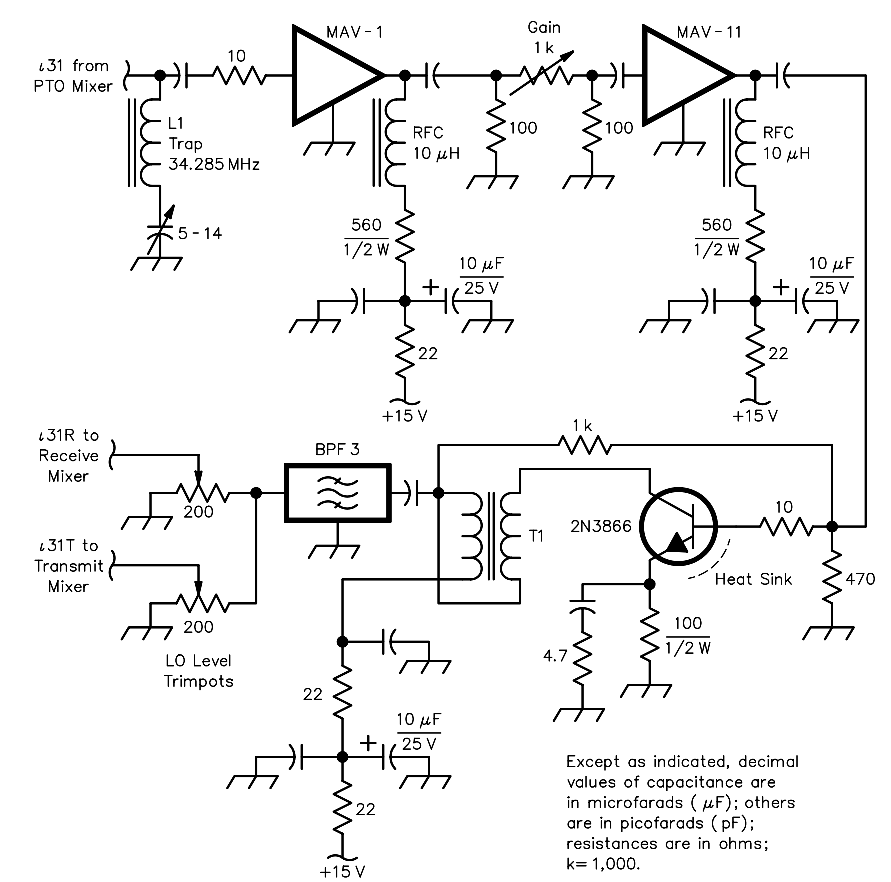
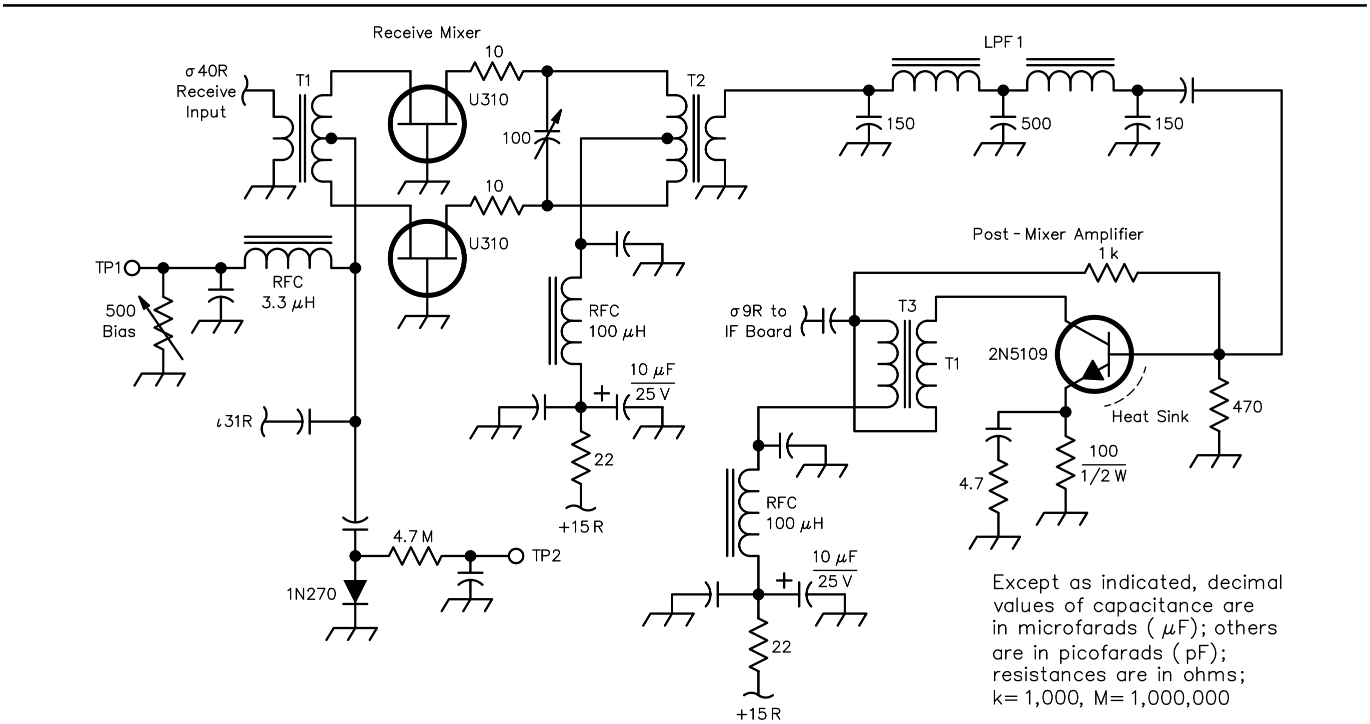


Fig 6—LO-amplifier schematic diagram. For general notes on the schematics, refer to the caption for Fig 3.

BPF3—Same as BPF1 (see Fig 5).
L1—2 μH, T-37-10 powdered-iron toroidal core, 28 turns.

T1—FT-37-61 ferrite toroidal core, 8 bifilar turns.

Fig 7 (below)—Receive-mixer schematic diagram. For general notes on the schematics, refer to the caption for Fig 3. The JFETs comprise a matched pair. Resonance in the mixer drain circuit is with about 80 pF. The tuning capacitor may be made up of convenient components; for example, a 62-pF silver-mica capacitor in parallel with a 9 to 35-pF ceramic trimmer. The 64:1 transformer in the drain circuit presents a load of 1600 Ω to each drain, which is optimum for best IMD performance (see Reference 4).
LPF1—Low-pass filter, cutoff frequency 13 MHz. Inductors, 1 μH, T-37-6 powdered-iron toroidal core, 18 turns.
T1—Mixer input transformer, FT-37-61 ferrite toroidal core, 8 trifilar turns.
T2—Mixer output transformer, T-50-6 powdered-iron toroidal core. Primary, 4 μH, 32 turns, wound over full length of core, center-tapped. Secondary, 4 turns, close-wound at the center of the primary winding.
T3—FT-37-43 ferrite toroidal core, 6 bifilar turns.



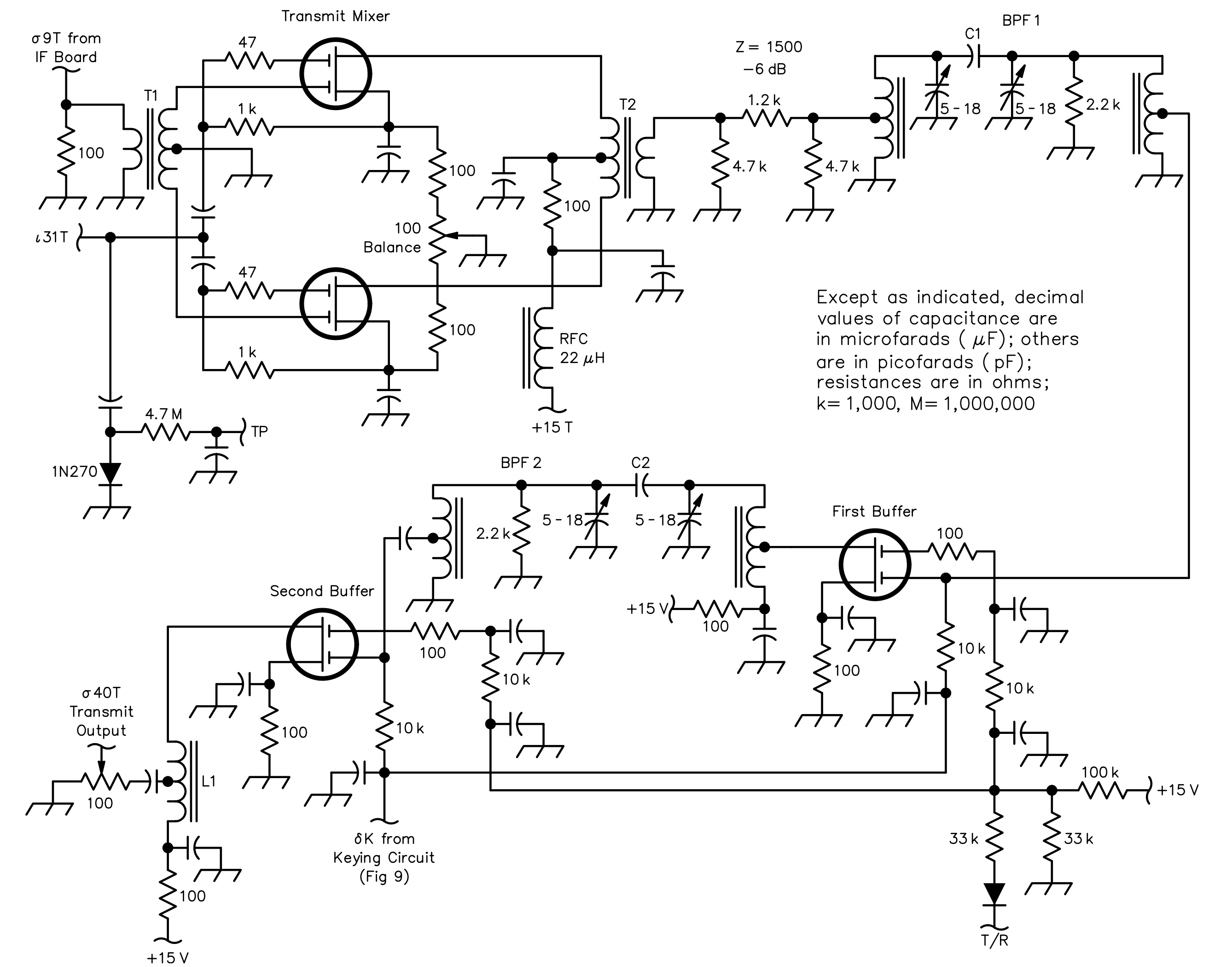


Fig 8—Transmit-mixer schematic diagram. For general notes on the schematics, refer to the caption for Fig 3.

BPF1-BPF2—Band-pass filter, bandwidth 2 MHz, center 39.5 MHz. Inductors, 1 µH, T-37-6 powdered-iron toroidal core, 18 turns, center-tapped.
C1, C2—Nominally 0.5 pF; adjust for desired passband with sweep generator. Homebrew fractional-pF piston trimmer capacitor; cover a 1-inch piece of #16 bare wire with teflon-sleeving, then wrap #24 bare wire completely over a 1/2-inch range of the teflon. Adjust by sliding the teflon, with wrapping, partially off the larger wire.
L1—FT-37-61 ferrite toroidal core, 11 turns, tap 2 turns from ground end.
T1—FT-37-43 ferrite toroidal core, 6 trifilar turns.
T2—FT-37-61 ferrite toroidal core, 6 trifilar turns.

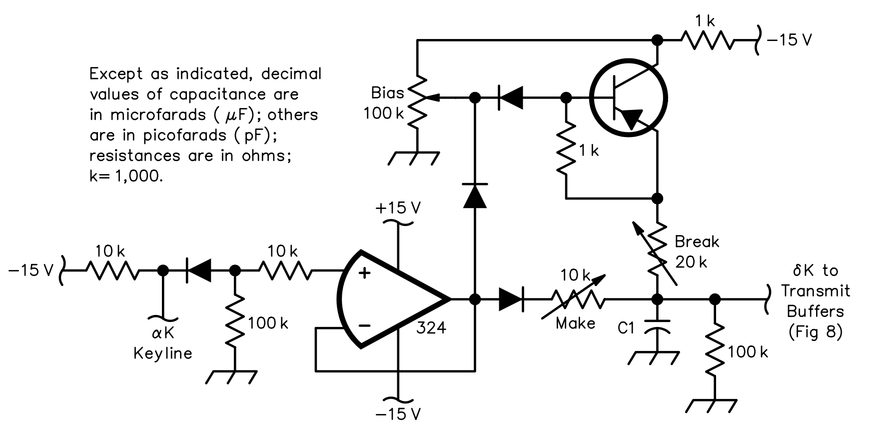


Fig 9—Keying-circuit schematic diagram. For general notes on the schematics, refer to the caption for Fig 3. The bias trimpot sets the cutoff bias. When the key is closed, the transistor is turned off and the Break trimpot has no effect; the Make trimpot determines the rise time. When the keyline is open, the Make trimpot has no effect and the transistor is turned on; the Break trimpot determines the decay time. The voltage-follower op amp isolates the timing circuit from external influences. C1—Low-loss timing capacitor, 0.47 µF. Panasonic P-series polypropylene, #ECQ-P1H474GZ. Digi-Key #P3474 (see Reference 8).

noise off the main 40-MHz line with no significant loading of the main receiver circuits. With no input matching, the preamplifier has only moderate gain; it serves mainly as a buffer. The noise mixer is a simple MOSFET, which is adequate for this task. The LO injection power is provided by a free-running oscillator tuned by a varactor diode with a front-panel control. The total range is about 2 MHz, providing considerable over-range. The 1 MHz

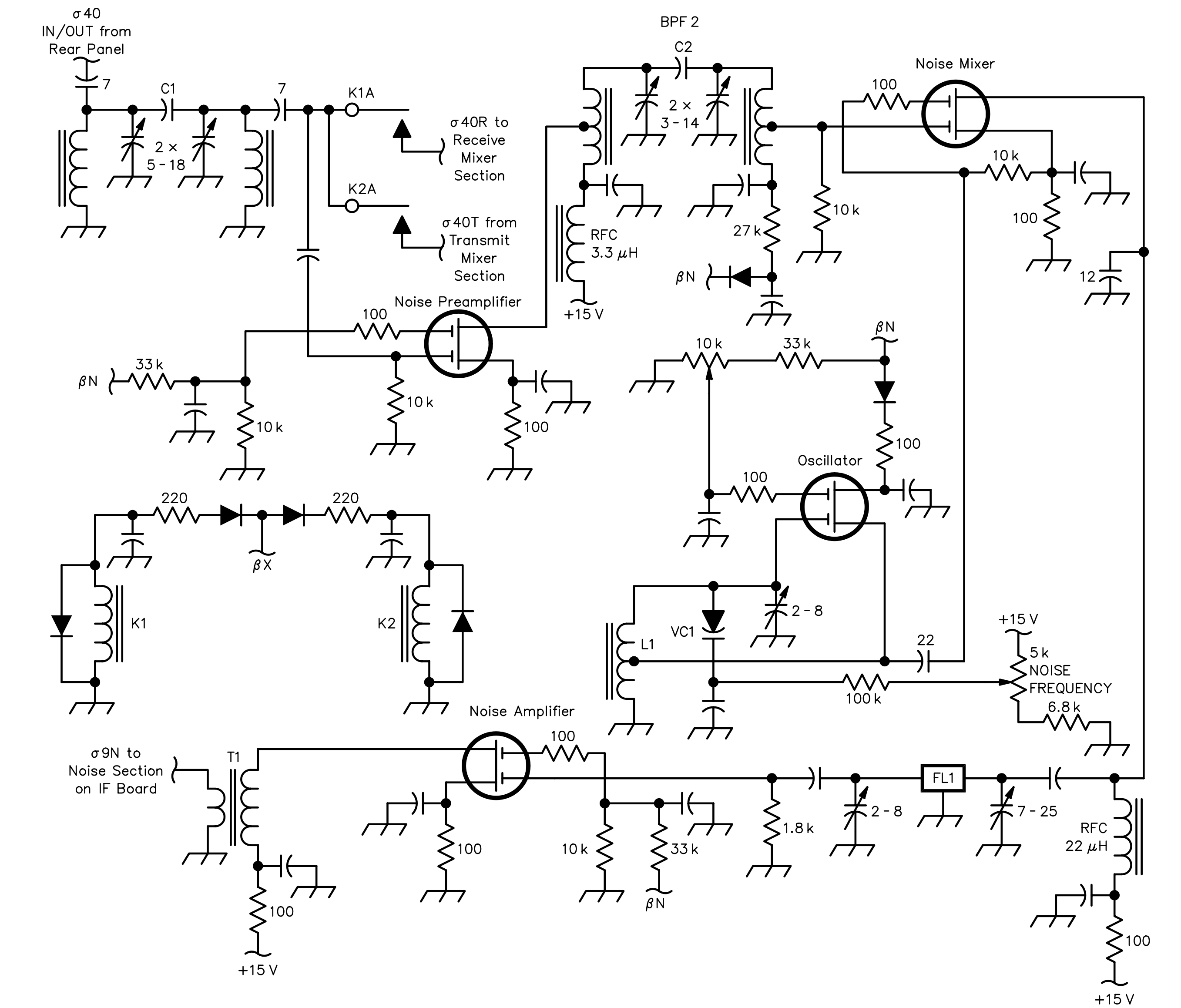


Fig 10—Tunable noise-channel schematic diagram. For general notes on the schematics, refer to the caption for Fig 3. This diagram includes the main 40-MHz band-pass filter and the TR relays. Reed relays are used to allow high-speed, break-in CW (QSK) operation.

The βN control line is provided by the Logic board. It switches from –15 to +15 V when the blanker is turned on. The control line is used here in three different ways. Most conveniently, for the noise-channel preamp and the noise-channel post-mixer amp, it controls the MOSFETs by means of gate 2. For the noise mixer, a diode is added; this simulates a μ-type control line and controls the MOSFET using gate 1. The βN line is also used to directly power the noise channel's local oscillator. Later designs avoid μ-type control lines and use only β-type control lines. This simplifies the logic board circuit; only a diode is needed for a β-type control line to simulate an μ-type line. The βX line switches from –15 V in receive mode to +15 V in transmit. Thus it powers one or the other of the TR reed relays.

BPF1—Main band-pass filter, center 39.5 MHz, bandwidth 1.4 MHz. Inductors, 800 nH, T-37-6 powdered-iron toroidal core, 15 turns.

BPF2—Noise band-pass filter, center 39.5 MHz, bandwidth 2 MHz. Inductors, 1.2 μH, T-37-6 powdered-iron toroidal core, 19 turns, center-tapped.

C1, C2—Same as C1 and C2 in Fig 8.

FL1—Crystal filter, 25-kHz bandwidth, four-pole. This filter was salvaged from an irreparable CX7; it has an impedance of 3200 Ω.

K1, K2—Miniature reed relay, SPST, NO, 12 V dc coil, 1450 Ω, 8 mA; Gordos #0490-1478DZ; Hosfelt #45-191 (see Reference 11); Jones #3806-RL (see Reference 15).

L1—Noise local-oscillator coil, 1.8 μH, T-50-6 powdered-iron toroidal core, 21 turns, tapped 5 turns from the ground end.

T1—FT-37-43 ferrite toroidal core. Primary, 11 turns; secondary, 2 turns wound on low end of primary.

VC1—Varactor diode, nominal 6.8 pF. Motorola type MV2101; NTE type 610 (see Reference 11).

desired range is conveniently obtained over the upper 180° swing of the panel knob. Drift problems were anticipated with this free-running circuit, but in seven years of operating, this has not been a problem. The diode typically covers a range of 4 to 14 pF when reverse biased from 15 to 0.1 V. Less range is actually required here, and the panel control is wired accordingly. This prevents conduction in the varactor diode, which might otherwise result from the RF voltage in the tank circuit. The mixer output at 9 MHz is filtered by a two-pole crystal filter and amplified by a MOSFET, with the output going to the noise amplifier on the IF board.

Construction

The RF board is shown in Fig 1. The general method of construction was described in Part 1, where the need for careful shielding and lead filtering was discussed. Part 2 gave further construction details, most of which also apply to the RF board. The power and control leads are filtered as described in Part 2, except for the π-section filters in the power and control lines feeding the circuits operating above 30 MHz. Those instead consist of two 1-nF bypass capacitors and a 100-µH RFC. The board's underside is shown in Fig 11.

Some of the circuits are built on perfboards: some with no solder pads, some with pads and some with a ground plane on the underside. These were all poor choices for RF circuits. Plain copper board and true dead-bug construction—as used in the noise channel and later in the IF board (Part 2)—is much better. In addition, the LO amplifier should be in a separate compartment. All these circuits are scheduled for rebuilding.

Test and Alignment

The BFO trimpots in the IFS circuit are adjusted in the transmit mode for 8816.5 kHz in USB and 8813.5 kHz in LSB. This assumes selection of first-rate, prime-condition SSB crystal filters with the correct passband of 8814-8816 kHz. The USB and LSB BFO-buffer-level trimpots are adjusted to obtain –16 dBm at the IF port of the BFO mixer. The trimpot at the BFO section output (terminal φ9A leading to the AF board) is adjusted for 100 mV P-P output. The BFO voltage on this line is kept low to minimize leakage into the IF strip and elsewhere. A BFO amplifier on the AF board provides the proper LO injection levels for the product detector and balanced modulator.

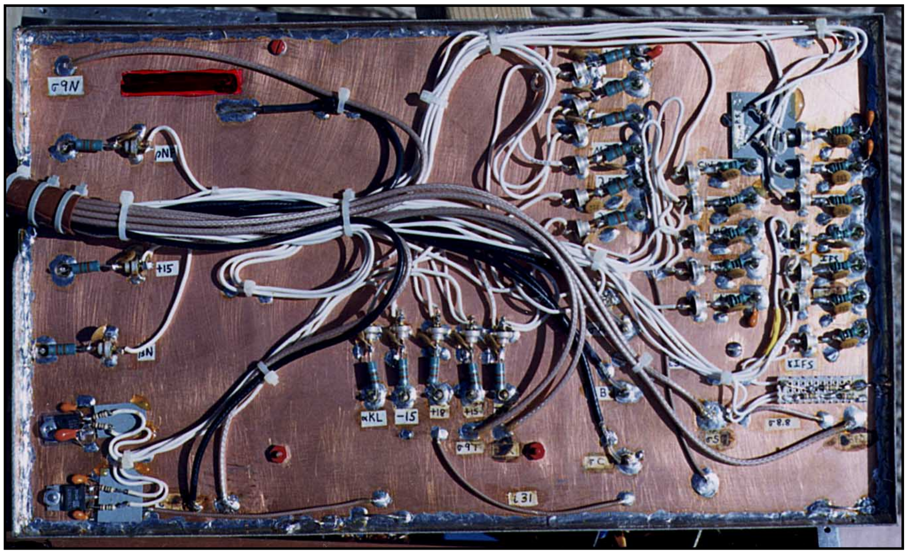


Fig 11—Bottom view of the RF board. Effective filters are installed at each terminal and coax cables are soldered directly to the double-sided circuit board; see the discussion in Part 2. To minimize connector troubles, the board is hard-wired to the radio; a 12-inch-long bundle of wires and cables allows the board to be easily lifted and serviced.

The master oscillator is adjusted for 43.1 MHz. The offset oscillator is adjusted for a range of 34.2835 MHz (for 0 Hz offset) to 34.2825 MHz (for 1000-Hz offset). The PTO trimpots are adjusted to provide –16 dBm at the IF port of the PTO mixer. Injection to both the BFO and PTO mixers at the LO ports is about +4 dBm. Although the doubly balanced mixers (DBMs) are +7 dBm devices, the lower LO power results in only about 1 dB more conversion loss. This lower LO power is in accordance with the manufacturer's suggestions for situations where dynamic range is not a factor.[7] Lessened LO power reduces harmonic mixing and decreases stray LO power in the system.

For convenience, RF probes that measure LO injection level at the main mixers are permanently built into the circuits. This enhances the repeatability of measurements, difficult to achieve with external RF probes and their questionable grounding leads. The proper dc voltages at the receive- and transmit-mixer LO test points are 1.1 and 1.4 V, respectively (on a 10 MΩ meter). The two trimpots at the LO-amplifier output compensate for the different impedances of the two mixers and for circuit reactance; one trimpot is kept at maximum. The **GAIN ADJUST** trimpot in the LO amplifier is set as required to obtain the specified injection levels.

The RF-board receiving circuits operate at a gain of 15 dB, overall, between terminal σ40 (the rear-panel 40-MHz jack) and terminal σ9R, which leads to the IF board. The mixer drain-circuit tuning capacitor is peaked on a weak signal. The input to the transmit mixer at terminal σ9T is adjusted for 200 mV P-P using trimpots on the IF board. The transmit mixer's balance trimpot is adjusted to minimize LO energy at the output. The trimpot at the output of the transmit-mixer section is set to obtain –7 dBm at the 40-MHz jack on the rear panel of the radio. Other alignment specifications are included above in the discussions of the individual circuits.

Summary

This article gives a complete description of the RF board in a high-performance homebrew transceiver. The board establishes the 40 MHz to 9 MHz transitions.

References

[1]M. Mandelkern, K5AM, "A High-Performance Homebrew Transceiver: Part 1," *QEX*, Mar/Apr 1999, pp 16-24.

[2]M. Mandelkern, K5AM, "A High-Performance Homebrew Transceiver: Part 2," *QEX*, Sep/Oct 1999, pp 3-8.

[3]A long-term temperature-compensation program was carried out over a six-month period while work continued on the next board in the radio. A garage workshop, quite cold on winter mornings, provided daily temperature cycling.

[4]E. Oxner, KB6QJ, "Designing FET Balanced Mixers for High Dynamic Range," Application Note LPD-14, pages 9-114 to 9-145 in the Low Power Discretes Data Book, 1989, Siliconix Incorporated, 2201 Laurelwood Rd, Santa Clara, CA 95054-1516; tel 408-988-8000, fax 408-727-5414; **www.siliconix.com.**

[5]J. B. Compton, "High-Frequency JFET Characterization and Applications," *Design Electronics*, Mar, 1970.

[6]M. Mandelkern, K5AM, "Evasive Noise Blanking," *QEX*, Aug 1993, pp 3-6.

[7]*RF/IF Designer's Handbook*, Mini-Circuits Labs, Brooklyn, NY, 1992, page 1-14 (also see Reference 13).

[8]Digi-Key Corporation, 701 Brooks Ave S, PO Box 677, Thief River Falls, MN 56701-0677; tel 800-344-4539 (800-DIGI-KEY), fax 218-681-3380; **http://www.digikey.com/**

[9]Surplus Sales of Nebraska, 1502 Jones St, Omaha, NE 68102; tel 402-346-4750, fax 402-346-2939; **www.surplussales.com**. See p 82 in catalog 8 for Erie ceramic trimmers.

[10]Mouser Electronics, 2401 Hwy 287 N, Mansfield, TX 76063; tel 800-346-6873, fax 817-483-0931; **sales@mouse.com**; **http://www.mouser.com/**

[11]Hosfelt Electronics, 2700 Sunset Blvd, Steubenville, OH 43952; tel 800-524-6464, fax 800-524-5414; **hosfelt@clover.net**, **http://www.hosfelt.com/**.

[12]International Crystal Mfg Co, 10 North Lee, PO Box 26330, Oklahoma City, OK 73126-0330; tel 800-725-1426, 405-236-3741, fax 800-322-9426; e-mail **customer-service@icmfg.com**; **http://www.icmfg.com/**.

[13]Mini Circuits Labs, PO Box 350166, Brooklyn, NY 11235-0003; tel 800-654-7949, 718-934-4500, fax 718-332-4661; **http://www.mini-circuits.com/**

[14]Z. Lau, W1VT, "Designing Band-Pass Filters," *QEX*, Jul/Aug 1998, pp 53-56.

[15]Marlin P. Jones & Associates, Inc, PO Box 12685, Lake Park, FL 33403-0685; tel 800-OK2-ORDER (800-652-6733), 561-848-8236 (Technical), fax 800-4FAX-YES (800-432-9937); **mpja@gate.net**; **www.mpja.com**.

Notes on "Ideal" Commutating Mixers

Here's a little light for a topic that confuses many hams.

By Doug Smith, KF6DX

In the Nov/Dec 98 *QEX* "Letters to the Editor" column, Byron Blanchard, N1EKV, raised a question regarding the following statement from my article in the Mar/Apr 98 issue: "It's no wonder a mixer's conversion loss is about 6 dB—precisely what physics predicts." I had given the proof that two equal-amplitude cosine waves, when multiplied, produced sum and difference products at half the input amplitude, yielding a 6-dB conversion loss. Of course, real reverse-switching mixers don't contain little analog computers that do the math, but simply commutate the input signal based on LO drive polarity. I'll show how this can result in a conversion loss of less than 6 dB, as Byron rightly claimed.

A misconception persists, however, about the "ideal" sine-wave mixer. The fallacy indicates a 3-dB conversion loss is possible for a "loss-less," passive sine-wave mixer. I'll attempt, again, to show this isn't possible, at least by any mechanism of which I'm aware. I'm writing this to share what I learned, but also with hopes of mitigating any confusion I caused in the earlier exchange. I also hope you'll be able to follow my pedantic methods—and help me out!

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Multipliers as Mixers and Vice Versa

Any nonlinear circuit or process acts as a mixer. It doesn't necessarily follow that every mixer employs a nonlinear process. In an AM transmitter, for example, the process of modulation produces sidebands at the sum and difference frequencies of the carrier and baseband signals. Nothing need be nonlinear to do it. This was the first "brain-lock" I had to overcome, believe it or not.

Below is the "textbook" derivation for the output of an analog multiplier, as I gave before:

$$(\cos\omega_0 t)(\cos\omega t) = \left[\frac{(e^{j\omega_0 t} + e^{-j\omega_0 t})}{2}\right]\left[\frac{(e^{j\omega t} + e^{-j\omega t})}{2}\right] \qquad \text{(Eq 1)}$$

[In order to reduce clutter, parentheses around the arguments of trigonometric functions have been eliminated when those arguments are single terms or simple products. For example, sin(ωt) is written simply sin $\omega_0 t$.—*Ed.*]

$$= \frac{\left[e^{j(\omega_0+\omega)t} + e^{-j(\omega_0+\omega)t}\right] + \left[e^{j(\omega_0-\omega)t} + e^{-j(\omega_0-\omega)t}\right]}{4} \qquad \text{(Eq 2)}$$

$$= \frac{1}{2}\left[\cos\left(\omega_0 t + \omega t\right) + \cos\left(\omega_0 t - \omega t\right)\right] \qquad \text{(Eq 3)}$$

Note that this represents a linear process, since we can alter the amplitude of either input term, and the output will change by the same factor; that is:

$$(A\cos\omega_0 t)(B\cos\omega t) = \frac{AB\left[\cos(\omega_0+\omega)t+\cos(\omega_0-\omega)t\right]}{2} \quad \text{(Eq 4)}$$

The goal of LO drive to a double-balanced mixer (DBM) is to alternately switch diode pairs in the quad on and off at the LO frequency. In the limiting case, the LO drive is a square wave. At no time is any of the diodes in a "half-on" state. This is equivalent to multiplying the input signal by either 1 or –1. The product of such a multiplication, as in every other case, is dependent on the spectral content of the inputs. Here, we have a sine wave at one input and a square wave at the other. See Fig 1.

The square-wave LO drive obviously contains harmonics. If we modify Eq 4 above to describe the harmonic content of the LO, we can use it to find the output level at the sum or difference of the fundamental input frequencies. We define the conversion loss of the mixer as the output level of the desired sum or difference product relative to the non-LO input level. For square-wave LO drive sufficient to approach perfect switching, its amplitude is ±1. We have to find the amplitude of the square wave's fundamental component.

Fourier analysis does a great job of finding spectral content of waveforms, but the integrals are intractable for exact mathematical results. We end up calculating Fourier series instead. I offer here—with some qualifications—another technique that I call the "mean-squares" method. When the phases of harmonic components are known, it produces exact harmonic amplitudes. We will first use it to find the amplitude of a square wave's fundamental component.

The "Mean-Squares" Method of Harmonic Analysis

If some repetitive function of time $f(t)$ is known to be a component of another function of time $g(t)$, and its relative phase is known, the amplitude of function $f(t)$ can be found using the following procedure. In this case, we have $g(t) = \mathrm{sign}(\sin x)$, a square wave of unity amplitude, and $f(t) = A \sin x$, its fundamental component. See Fig 2.

I say the difference of the mean-squares of $g(t)$ and $f(t)$ is equal to the mean-square of the difference $g(t)– f(t)$. Put another way, were $g(t)$ a waveform delivering power to a load, the power delivered is proportional to the sum of the RMS value of $f(t)$ squared plus the RMS value of $[g(t)–f(t)]$ squared.

"What's that?" you say. Here's how it works: Find the mean-square (RMS^2) value of $g(t)$, the square wave. Then find the mean-square value of $g(t) – f(t)$. Subtract from the mean-square of $g(t)$, and solve for A. The mean-square value of our unity-amplitude square wave is 1. The peak amplitude of sine wave $f(t)$ is A, and its mean-square value is $A^2 / 2$.

To find the mean-square value of a function, find the average value of the square of the function over some interval. The function $g(t) – f(t)$ represents the shaded area between the curves in Fig 2. The mean-square value of this function is written as:

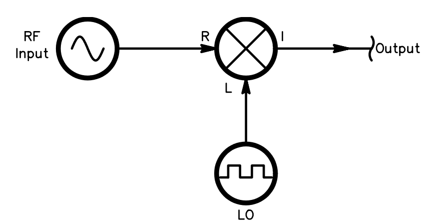

Fig 1—DBM with square-wave LO drive.

$$MS_{g(t)-f(t)} = \frac{1}{\pi}\int_0^{\pi}(1-A\sin t)^2\,dt$$

$$= \frac{1}{\pi}\int_0^{\pi}(1-2A\sin t + A^2\sin^2 t)\,dt$$

$$= \frac{1}{\pi}\left[t+2A\cos t + A^2\left(\frac{t-\sin t\,\cos t}{2}\right)\right]_0^{\pi} \quad \text{(Eq 5)}$$

$$= \frac{A^2}{2} - \frac{4A}{\pi} + 1$$

We can solve for A by writing:

$$MS_{g(t)-f(t)} + MS_{f(t)} = MS_{g(t)}$$

$$\left(\frac{A^2}{2} - \frac{4A}{\pi} + 1\right) + \frac{A^2}{2} = 1$$

$$A^2 - \frac{4A}{\pi} = 0$$

$$A\left(A - \frac{4}{\pi}\right) = 0 \quad \text{(Eq 6)}$$

$$A = \frac{4}{\pi}$$

So the fundamental's peak amplitude is $4/\pi$, or greater than unity! It turns out the amplitude of each harmonic in a square wave is inversely proportional to its harmonic number, and the phase of each one is opposite from its neighbor. To generate a square wave of unity amplitude, we can use:

$$g(t) = \frac{4\cos\omega_0 t}{\pi} + \frac{4\cos 3\omega_0 t}{3\pi} + \frac{4\cos 5\omega_0 t}{5\pi} \quad \text{(Eq 7)}$$

To find out what happens when we use this series as an LO, we substitute it into Eq 4.

Mixing with a Square-Wave LO

Taking the first two terms of Eq 7 as our LO (the fundamental and third harmonic only) we have mixer output:

$$y = \left[\frac{4\cos\omega_0 t}{\pi} - \frac{4\cos 3\omega_0 t}{3\pi}\right]\left[A\cos\omega t\right]$$

$$= \frac{2A}{\pi}\left[\cos(\omega_0+\omega)t + \cos(\omega_0-\omega)t\right] + \quad \text{(Eq 8)}$$

$$\frac{2A}{3\pi}\left[\cos(3\omega_0+\omega)t + \cos(3\omega_0-\omega)t\right]$$

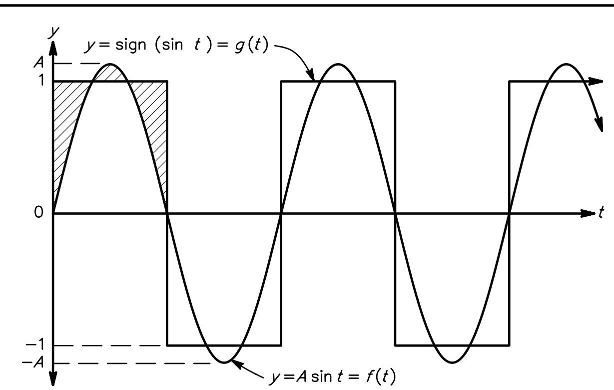

Fig 2—A unity-amplitude square wave and its fundamental component.

and sure enough, for $A = 1$, the amplitude of the first pair of terms in the result is $2 / \pi$. And:

$$20\log\left(\frac{2}{\pi}\right) \approx -3.9\ dB \qquad \text{(Eq 9)}$$

The conversion loss of an ideal square-wave LO mixer is, in fact, 3.9 dB.

As a check, we turn to the example in the *Handbook,*[1] where we set the LO frequency equal to the other input frequency, or:

$$\omega_0 = \omega \qquad \text{(Eq 10)}$$

Note that we maintain square-wave LO drive, and the other input is still a sine wave. It's a doubler, and by inspection of the switching, we find the output is a full-wave rectified sine curve. This obviously has a dc term and a term at the second harmonic of the input frequency. These are the sum and difference frequencies in this case. To find the second harmonic's amplitude, we'll again use the mean-squares method.

Now $g(t)$ is a rectified sine wave, and $f(t)$ is the negative of a cosine wave at twice the frequency. See Fig 3. In finding the mean-square value of $g(t) – f(t)$, we'll avoid integrating over a full cycle of $f(t)$ to avoid a non sequitur result. Instead, we'll integrate over half of $f(t)$'s period to find the mean-square value of the shaded area in Fig 3 and double the result.

We have:

$$MS_{g(t)\text{-}f(t)} = \frac{2}{\pi}\int_0^{\frac{\pi}{2}} (\sin t + A\cos 2t)^2\, dt$$

$$= \frac{2}{\pi}\int_0^{\frac{\pi}{2}} (\sin^2 t + 2A\sin t\cos 2t + A^2\cos^2 2t)\,dt \qquad \text{(Eq 11)}$$

The integral of $\sin^2 t$ can be found in reference tables, but for the other two terms, we utilize some trigonometric identities, namely:

$$2A\sin t\cos 2t = A(\sin 3t - \sin t) \qquad \text{(Eq 12)}$$

and

$$A^2\cos^2 2t = \frac{A^2(\cos 4t + 1)}{2} \qquad \text{(Eq 13)}$$

Substituting:

$$MS_{g(t)\text{-}f(t)} = \frac{2}{\pi}\int_0^{\frac{\pi}{2}}\left[\sin^2 t + A(\sin 3t - \sin t) + \frac{A^2(\cos 4t + 1)}{2}\right]dt$$

$$= \frac{2}{\pi}\left[\frac{t - \sin t\cos t}{2} + A\left(-\frac{\cos 3t}{3} + \cos t\right) + \frac{A^2\left(\frac{\sin 4t}{4} + t\right)}{2}\right]_0^{\frac{\pi}{2}} \qquad \text{(Eq 14)}$$

$$= \frac{A^2}{2} - \frac{4A}{3\pi} + \frac{1}{2}$$

Taking the mean-square value of $g(t)$ as 1/2, and solving for A:

$$MS_{g(t)\text{-}f(t)} + MS_{f(t)} = MS_{g(t)}$$

$$\frac{A^2}{2} - \frac{4A}{3\pi} + \frac{1}{2} + \frac{A^2}{2} = \frac{1}{2}$$

$$A = \frac{4}{3\pi} \qquad \text{(Eq 15)}$$

[1]P. Danzer, N1II, Ed., *The 1997 ARRL Handbook for Radio Amateurs* (Newington, Connecticut: ARRL, 1996), p 15.15 ff.

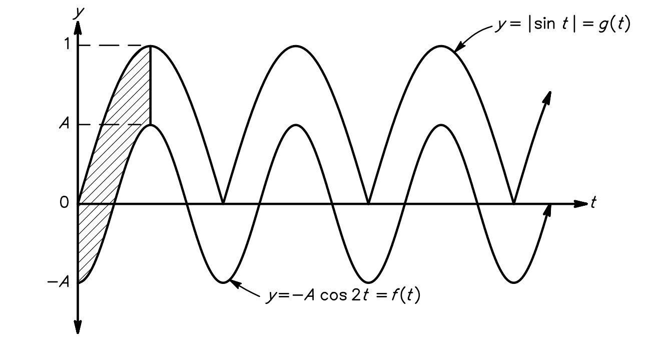

Fig 3—A full-wave rectified sine curve and its fundamental component.

Hey, what gives?! We were expecting to get $2 / \pi$, which we know from power-supply theory is the average value of $g(t)$, so that the sum and difference amplitudes would be equal.

Looking back at Eq 8, we get the explanation. In this case, we had set the two frequencies equal, and therefore:

$$\omega_o + \omega = 3\omega_0 - \omega \qquad \text{(Eq 16)}$$

So the matching terms in Eq 8 appear at the same frequency—the second harmonic—and their amplitudes must be added. The final amplitude ought to be:

$$A_{2\omega 0} = \frac{2}{\pi} - \frac{2}{3\pi}$$

$$= \frac{4}{3\pi} \qquad \text{(Eq 17)}$$

which agrees with the mean-squares analysis.

Initially, this result involving unequal sum and difference amplitudes threw me for another "brain-lock." To use inputs with equal frequencies to demonstrate mixer conversion loss is therefore a somewhat awkward choice.

Conclusion

From this discussion, is it apparent to me that "ideal" mixers must be thought of either as analog multipliers or as commutators. The conversion loss of a DBM using square-wave LO drive is less than that of an ideal analog multiplier because the square wave's fundamental content is greater than unity. The statement "If the LO driving signal contained no harmonics, half of the mixer's output power would show up in the sum product and half in the difference product, resulting in a 3-dB conversion loss...." (see Note 1) doesn't make sense. If the LO had no harmonics, it would not be switching the input signal in square-wave fashion. It is difficult for me to imagine how to devise an LO drive signal that would cause the diodes in a DBM to conduct in such a way as to embody Eq 4 exactly.

Also, the idea that "an ideal reversing-switch mixer is loss less," seems to have led to the misconception that such a device would produce a conversion loss of 3 dB. This statement is almost true if you consider all the harmonics at input and output, but that's not how we defined conversion loss. Besides that, every switch requires some work to throw it. It's even harder to imagine what passive device could perform the linear multiplication of Eq 4, regardless of conversion loss. Anyone care to give us a treatment of Varicap doublers?

RF

By Zack Lau, W1VT

Accurately Modeling Capacitor Q in *ARRL Radio Designer*

I received an e-mail from Lucky, W7CNK, who pointed out that capacitor Q extrapolation models do a poor job of matching real capacitors. I suspect that whoever coded the models borrowed the code for inductor models and never bothered to compare the calculations against measured results. This isn't that surprising, as the effect of capacitor Q is small compared to other losses in typical circuits. Typically, in lumped-component circuits, inductor losses dominate. Of course, if you measure your Q values at the frequency of interest, the errors are small and tough to see.

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The problem is that capacitor Q decreases with increasing frequency. This results from capacitive reactance being inversely proportional to frequency.

$$Q = \frac{\text{reactance}}{\text{resistance}} = \frac{\frac{1}{2\pi f C}}{\text{resistance}} \quad \text{(Eq 1)}$$

In this context, it may be confusing to associate Q with the quality of a capacitor. The degradation of Q has more to do with the way Q is defined, than how the capacitor is built. Similarly, inductor Q increases with frequency—assuming constant resistance—due to the increase in reactance with frequency.

Fortunately, there is an easy way to define Q in terms of frequency in *ARRL Radio Designer* (ARD). ARD allows parameters such as Q to be defined using equations.[1] To help the program, the equation must be put in a pair of parenthesis. An example is:

```
qval:100 ; Q at 50.329 MHz
qvar:(qval*50.329MHz/f)
```

ARD recognizes *f* as a special variable, the frequency of analysis. Thus, as the frequency increases, the value of *qvar* decreases, as one would expect in a capacitor with fixed resistive losses.

Conductor loss, however, increases with frequency. The skin effect drives current to the surface and makes conductors lossier at higher frequencies. This loss varies as the square root of increasing frequency. American Technical Ceramics, Inc, (ATC) suggests an

[1]Notes appear on page 56.

Table 1

Compact Software - ARRL Radio Designer 1.5 02-JUN-99 15:03:35
File: capqex.ckt

Freq MHz	*MS21 (dB) QAT1 R16*	*MS21 (dB) QAT3*	*MS21 (dB) QAT10*	*MS21 (dB) QAT3 1R62*	*MS21 (dB) QAT100*	*MS21 (dB) ATC*	*MS21 (dB) VERQCL*	*MS21 (dB) VERQC*	*MS21 (dB) VERQCH*	*MS21 (dB) VERQ 1CL*
5.033	–2.37	–5.69	–9.57	–10.69	–10.54	–10.33	–10.36	–10.54	–9.57	–10.36
50.329	–1.36	–0.76	–0.27	–0.09	–0.03	–0.00	–0.03	–0.27	–1.36	–0.09
503.290	–10.34	–10.34	–10.33	–10.33	–10.33	–10.34	–10.33	–10.34	–10.39	–10.34

MS21 (dB) VERQ1C	*MS21 (dB) VERQ 1CH*	*MS21 (dB) VERQ 2CL*	*MS21 (dB) VERQ2C*	*MS21 (dB) VERQ 2CH*	*MS21 (dB) VERQ*	*MS21 (dB) VERQ1*	*MS21 (dB) VERQ2*	*MS21 (dB) VERQL*	*MS21 (dB) VERQ1L*	*MS21 (dB) VERQ2L*
–10.41	–10.54	–10.36	–10.36	–10.36	–9.57	–10.69	–10.54	–10.54	–10.54	–10.54
–0.27	–0.76	–0.27	–0.27	–0.27	–0.27	–0.27	–0.27	–0.03	–0.09	–0.27
–10.35	–10.39	–10.39	–10.39	–10.39	–10.33	–10.34	–10.34	–10.33	–10.33	–10.34

MS21 (dB) VERQH	*MS21 (dB) VERQ1H*	*MS21 (dB) VERQ2H*	*MS21 (dB) NET*	*MS21 (dB) NET1*	*MS21 (dB) NET10*	*MS21 (dB) NET100*
–2.37	–9.57	–10.54	–10.62	–2.35	–9.56	–10.54
–1.36	–0.76	–0.27	–2.37	–1.36	–0.27	–0.03
–10.34	–10.34	–10.34	–10.62	–10.34	–10.33	–10.33

exponent of 1.6 above 150 MHz for its 100A chip capacitors.[2]

ATC recommends Eq 2 for extrapolating Q above 150 MHz:

$$Q_f = Q_{150}\left(\frac{150}{f}\right)^{1.6} \qquad \text{(Eq 2)}$$

The metalization layer may be too thin to allow an accurate extrapolation to lower frequencies. Thus, attempting to extrapolate lower in frequency with this equation may result in unreasonably high values of Q. Thus, a 1-pF ATC 100A is shown in the ATC graph as having a Q of 23,000 at 30 MHz, instead of the 49,000 one might expect from extrapolation. The following models the capacitor in ARD.

```
Q150:3700
QATC:(Q150*150MHz/f)^1.6 ; For use
above 150 MHz
CAP 1 2 C=1PF Q=QATC F=150 MHz ;
1 pF ATC chip capacitor between nodes
1 and 2.
```

Those with other simulators may wonder exactly what those programs do. If you don't understand the documentation, I wouldn't worry. Many low-cost products have poor written documentation. Even expensive products can be confusing; although these often have technical-assistance numbers for registered users. Fortunately, the ease of using simulators also makes them easy to test. Just devise a good test circuit and see if the numbers make sense.

The ARD test I ran[3] is a simple, series-tuned LC circuit, something one might use as a tuned-bypass or coupling circuit. The Q of the capacitor can be determined indirectly by the insertion loss of the circuit at resonance, in this case 50.329 MHz. This is necessary, because I've not found a way to display the computed values of parameters in a report. I didn't type in all the circuit blocks at once. Instead, I added a few at a time to see what happens with each change to the circuit model. Initially, I looked at both return loss and insertion loss, but the return loss proved to be redundant, so I removed it from the report in Table 1. Return loss is often a more sensitive indicator of change, however, so it may be more useful with other circuits.

As explained in Chapter 6 of the *2000 ARRL Handbook*, the precise definition of Q becomes important when evaluating low-Q circuits, as the results begin to differ for Qs less than 10. ARD defines Q in terms of conductance, with a conductor in parallel with the capacitor. A bit of investigation shows that Eq 3 matches the Q model in ARD.

$$Q = \frac{R_p}{X_c} \text{ or } Q = \frac{\frac{1}{G_d}}{X_c} \qquad \text{(Eq 3)}$$

Notes

[1]D. Newkirk, WJ1Z, Ed., "Parameter Assignment by Means of Equations," *ARRL Radio Designer Manual,* (Newington: ARRL, 1994, Order No. 6796) p 16-14. ARRL publications are available from your local ARRL dealer or directly from the ARRL. Mail orders to Pub Sales Dept, ARRL, 225 Main St, Newington, CT 06111-1494. You can call us toll-free at tel 888-277-5289; fax your order to 860-594-0303; or send e-mail to **pubsales@arrl.org**. Check out the full ARRL publications line at **http://www.arrl.org/catalog**.

[2]American Technical Ceramics, Inc, Specification #102-584 Rev C 6/81.

[3]You can download this circuit file from the ARRL Web **http://www.arrl.org/files/qex/**. Look for CAPQEX.ZIP.

Test Circuit

```
** Evaluate the Q models in ARRL Radio Designer Manual
** Pages 15-14 and 15-15 of the ARD Manual
** See Note 3
**
** * Loss comments at 50.329 MHz
**
** Zack Lau W1VT 5/28/99

qval:10

qvar:(10*50.329MHz/f); Capacitor with fixed resistive loss.

*******************************************
*Correlate circuit performance with Q

blk
cap 1 2 c=100pf q=1 f=50.329MHz
ind 2 3 l=100nh
qat1:2por 1 3
end
* 1.36 dB loss

blk
cap 1 2 c=100pf q=3.16 f=50.329MHz
ind 2 3 l=100nh
qat3r16:2por 1 3
end
* 0.76 dB loss

blk
cap 1 2 c=100pf q=10 f=50.329MHz
ind 2 3 l=100nh
qat10:2por 1 3
end
* 0.27 dB loss

blk
cap 1 2 c=100pf q=31.62 f=50.329MHz
ind 2 3 l=100nh
qat31r62:2por 1 3
end
* 0.09 dB loss

blk
cap 1 2 c=100pf q=100 f=50.329MHz
ind 2 3 l=100nh
qat100:2por 1 3
end
* 0.03 dB loss

*******************************************

q150:350

qatc:(q150*(150MHz/f)^1.6) ;ATC's suggested Q model

blk
cap 1 2 c=100pf q2=qatc f=50.329MHz
ind 2 3 l=100nh
atc:2por 1 3
end

*******************************************

blk
cap 1 2 c=100pf q=qvar f=5.0329MHz
ind 2 3 l=100nh
verqcl:2por 1 3
end
*0.03 dB loss, Q=100

blk
cap 1 2 c=100pf q=qvar f=50.329MHz
ind 2 3 l=100nh
verqc:2por 1 3
end
*0.27 dB loss, Q=10

blk
cap 1 2 c=100pf q=qvar f=503.29MHz
ind 2 3 l=100nh
verqch:2por 1 3
end
*1.36 dB loss Q=1

*Capacitors show more loss at lower frequency;
*503 MHz Q=10 capacitor is worse than a 5 MHz
*capacitor Q=10 at 50 MHz.
* Not recommended
* program confused by q(f) and qvar(f)
*******************************************

blk
cap 1 2 c=100pf q1=qvar f=5.0329MHz
ind 2 3 l=100nh
verq1cl:2por 1 3
end
*0.09 dB loss Q=31.62

blk
cap 1 2 c=100pf q1=qvar f=50.329MHz
ind 2 3 l=100nh
verq1c:2por 1 3
end
* 0.27 dB loss Q=10

blk
cap 1 2 c=100pf q1=qvar f=503.29MHz
ind 2 3 l=100nh
verq1ch:2por 1 3
end
*0.76 dB loss Q=3.16

*Capacitors show more loss at lower frequency;
*503 MHz Q=10 capacitor is worse than a 5 MHz
*capacitor Q=10 at 50 MHz.
* Not recommended
* program confused by q1(f) and qvar(f)
*******************************************

blk
cap 1 2 c=100pf q2=qvar f=5.0329MHz
ind 2 3 l=100nh
verq2cl:2por 1 3
end
*0.27 dB loss

blk
cap 1 2 c=100pf q2=qvar f=50.329MHz
ind 2 3 l=100nh
verq2c:2por 1 3
end
*0.27 dB loss

blk
cap 1 2 c=100pf q2=qvar f=503.29MHz
ind 2 3 l=100nh
verq2ch:2por 1 3
end
*0.27
dB loss

*Model Q2 works with qvar(f).
*f in the statement line does not affect Q

*******************************************

blk
cap 1 2 c=100pf q=qval f=50.329MHz
ind 2 3 l=100nh
verq:2por 1 3
end
*0.27 dB loss Q=10
```

```
blk
cap 1 2 c=100pf q1=qval f=50.329MHz
ind 2 3 l=100nh
verq1:2por 1 3
end
* 0.27 dB loss Q=10

blk
cap 1 2 c=100pf q2=qval f=50.329MHz
ind 2 3 l=100nh
verq2:2por 1 3
*0.27 dB loss Q=10

end
blk
cap 1 2 c=100pf q=qval f=5.0329MHz
ind 2 3 l=100nh
verql:2por 1 3
*0.03 dB loss Q=100

end
blk
cap 1 2 c=100pf q1=qval f=5.0329MHz
ind 2 3 l=100nh
verq1l:2por 1 3
*0.09 dB loss Q=31.6

end
blk
cap 1 2 c=100pf q2=qval f=5.0329MHz
ind 2 3 l=100nh
verq2l:2por 1 3
end
*0.27 dB loss Q=10

blk
cap 1 2 c=100pf q=qval f=503.29MHz
ind 2 3 l=100nh
verqh:2por 1 3
end
*1.36 dB loss Q=1

blk
cap 1 2 c=100pf q1=10 f=503.29MHz
ind 2 3 l=100nh
verq1h:2por 1 3
end
*0.76 dB loss Q=3.16

blk
cap 1 2 c=100pf q2=10 f=503.29MHz
ind 2 3 l=100nh
verq2h:2por 1 3
end
*0.27 dB loss Q=10

blk
cap 1 2 c=100pf
res 2 3 r=31.4
* Xc/Rs=1 at 50.329 MHz
ind 3 4 l=100nh
net:2por 1 4
end
*2.37 dB loss

blk
cap 1 2 c=100pf
res 1 2 r=31.4
*Rp/Xc=1 at 50.329 MHz
ind 2 3 l=100nh
net1:2por 1 3
end
*1.36 dB loss

blk
cap 1 2 c=100pf
res 1 2 r=314
*Rp/Xc=10 at 50.329 MHz
ind 2 3 l=100nh
net10:2por 1 3
end
*0.27 dB loss

blk
cap 1 2 c=100pf
res 1 2 r=3140
* Rp/Xc=100 at 50.329 MHz
ind 2 3 l=100nh
net100:2por 1 3
end
*0.03 dB loss

freq
5.0329MHz
50.329MHz
503.29MHz
end
```

Letters to the Editor

PSK31: A New Radio-Teletype Mode (Jul/Aug '99)

Doug,

The excellent article by Mr. Martinez, G3PLX, in the July/August *QEX* created some problems for me. The details of the method of demodulating the received signal were difficult to grasp from the somewhat confusing language and the absence of specific timing information in Fig 2. For these reasons, it is possible to draw conclusions that are contrary to the author's intentions. Based on e-mail contacts with the author and my own perceptions that have evolved in that process, I would like to give my views on the demodulation subject. I will use Figs 2 and 3 from the article and some diagrams in this letter, Figs A, B, C and D.

The Waveform

Fig A is a detailed plot of the waveform that shows the phase, amplitude, and time segments that typify a PSK31 received signal. The wave envelope as shown has the half-sine-wave shape, which is not quite correct, but is used here for simplicity. This particular shape has significant side lobes because of the sharp corners at the baseline. The corners are smoothed at the transmitter and receiver by special filtering methods.

Fig A shows that the bit timing starts at point W at the top of the 32-ms pulse, that the logic 0 bit is 32-ms long and reverses phase at the 16-ms midpoint. The second half, which terminates at point X is, in this example, part of the wider 64-ms pulse. However, it could well terminate at the top of an adjacent 32-ms pulse.

Fig A shows that the region between points X and Y has a constant envelope and does not reverse phase. This region is logic 1. We see also that the 1 region has more area under its envelope and by analysis 3 dB more energy than the 0 bit. This would seem to be significant for PSK31 (the E_b/N_0 ratios are different), where a string of zeros is used for "synch" and character-gap purposes, as well as for the data 0.

Another 0 region begins at point Y and ends at point Z. So we see that there are two pulse shapes and widths involved: one 32 ms long and one 64 ms long. A string of ones is just a steady carrier at full amplitude. We see also that phase reversals always occur at the zero-envelope level, which minimizes the spectral widening that is created by phase reversals.

Demodulation

Fig 3 of the article shows that the "present" portion of the signal is multiplied (mixed) with a "previous" portion of the signal. To determine 1 or 0, we could compare the first half of a 0 or a 1 with the second half using a 16-ms delay, instead of the 32-ms delay in Fig 3. Because of the phase reversal between the two halves of a 0, the output of the multiplier is negative-going (+ times –, or – times +). The multiplier output is filtered to remove the twice-frequency component that is an artifact of the mixing process. The result can then be applied to an integrate-and-dump algorithm or some other desirable software that determines 1 or 0 at the end of the bit period. For this approach, the delay time in Fig 3 must be 16 ms.

Fig B shows how the first (delayed) half (A) of a 0 is multiplied with the undelayed second half (B) to produce the output of the multiplier (C), whose envelope is shown by the dotted line. This envelope is then integrated (D) and the bit decision is made at the end. Fig C shows a similar operation for the 1. Here, the multiplier output is positive-going and the integration leads to a larger final value, as mentioned before. These waveforms were all generated by exact equations using *Mathcad*.

Alternate Method

Fig D explains a different approach. The delay in Fig 3 is now 32-ms, not 16 ms. This method also demodulates correctly. In Fig D, there are three columns. Column one is the delayed ("previous") bit and column two is the undelayed ("present") bit. The phase reversals, if any, at the midpoint (16 ms) are illustrated. Column three is the demodulated output (product) of columns one and two. In row one, for a 00 pair, the product (+ times –) is a negative voltage, and if the two inputs are both flipped in

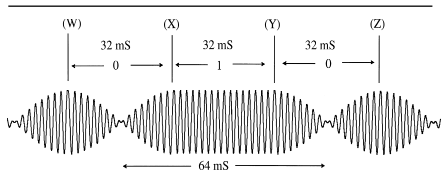

Fig A—A detailed plot of the waveform that shows the phase, amplitude and time segments of a typical PSK31 received signal.

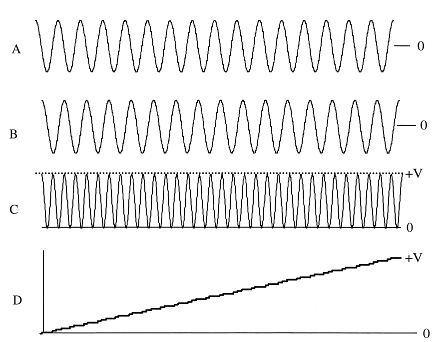

Fig C—The process of demodulating a 1 via a 16-ms delay.

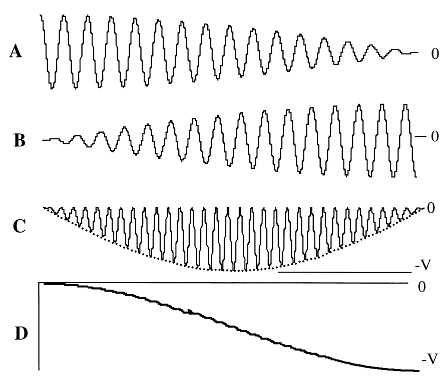

Fig B—The process of demodulating a 0 via a 16-ms delay.

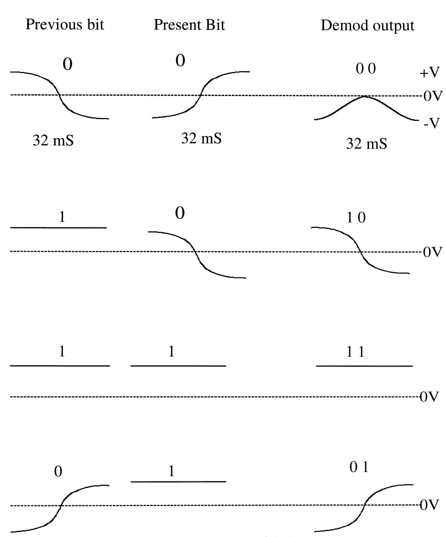

Fig D—W0IYH's alternative demodulation scheme that uses a 32-ms delay.

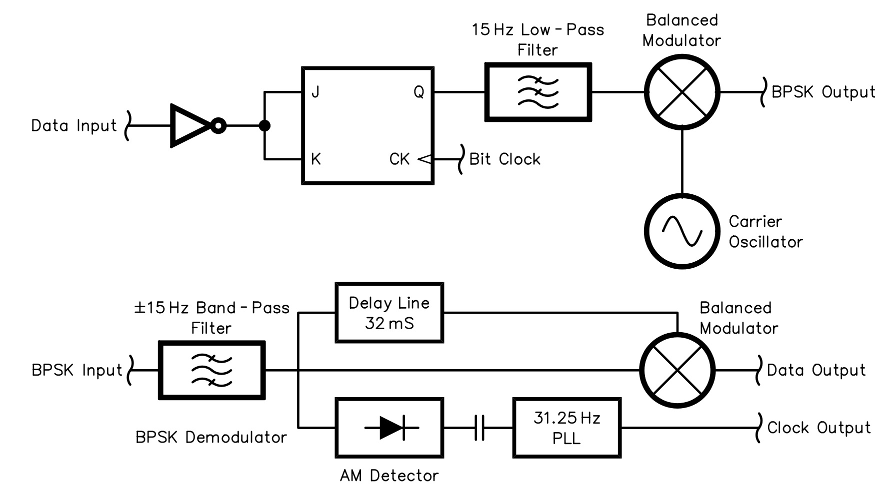

Fig 3—(Revised from P. Martinez, *QEX*, Jul/Aug 1999, p 7) Block diagram of a PSK31 modem.

phase, the product (– times +) remains negative. Adjacent 0s always look like columns one and two or their inverses. Rows two, three and four (or their inverses) show the other possibilities for adjacent pairs of bits.

Consider the sequence "00110." Row one decodes the first two 0s. We then go to row four (01) and the output changes as shown. We now have the first 1. Proceed to row three (11) and get the second 1. Then go to row two (10) and get the last 0.

So every 32 ms, we determine a new bit (the present bit). Observe in Fig D, however, that to decide the new bit, we must make *independent* determination of the first and second 16-ms segments in columns 1 and 2. At the output of the demodulator, these two values are compared independently to determine the new bit. For this reason, choosing between the two alternate methods does not seem to reveal any advantage of one over the other. Either method is correct.

The Ones Problem

The clock shown in Fig 3 of the article is synchronized with the string of 0s that is the PSK31 preamble. The software aligns the clock so that bit timing begins as shown in Fig A. During the time that a string of 1s, as shown in the Varicode list [Table 1 of the article—*Ed.*] is being sent, the clock must not drift out of synch. Between each character and word, some 0s are sent and these help to keep alignment. The software must provide the "inertia" that keeps the clock "flywheeling."—*William E. Sabin, W0IYH, 1400 Harold Dr SE, Cedar Rapids, IA, 52403;* **sabinw@mwci.net**

The author responds in a question-and-answer format (questions posed by the editor):

Q: In the upper half of Fig 3, you show a simple BPSK modulator. A continuous string of 0s obviously does not generate a continuous series of phase reversals, as stated in the text. How should this diagram be modified?

A: The "data input" to the top of Fig 3 should indeed be a string of reversals when idling, not a string of zeros. A flip-flop circuit or a single-stage shift register and an exclusive-OR gate could be used to do this. It's the same process used in AX25. [See the modified Fig 3 above.—*Ed.*]

Q: It is evident that the differential phase-shift-keying (DPSK) demodulator in the lower half of Fig 3 does its job by comparing the phase during the current bit time with that of the previous bit time, hence the 32-ms delay. Is there any advantage in changing the delay to 16 ms so that the first half of each bit is compared with the last half? What SNR performance improvement might be expected, if any?

A: I believe this question may have arisen because some readers have misinterpreted the time scale of Fig 2. The complete diagram is six bits wide, each bit being 32 ms in duration. The waveform is always at a maximum every 32 ms, so a delay line of this length results in the phase comparison being made when both direct and delayed signals are at their peak values. I know of no advantage, SNR or otherwise, of using any other delay.

Q: Varicode may contain sequences of up to nine 1s in a row (exclamation point). Continuous strings of exclamation points would seem to cause the clock regeneration circuit in Fig 3's demodulator to lose proper lock. What techniques are used in clock extraction to prevent this?

A: The clock extraction process is the same as that used in AMTOR, PACTOR or AX25, except that it's driven by the AM component of the signal, rather than the decoded data. Can I suggest thinking of this problem from the other end, in the following way?

Suppose, for a moment, that the transmitter clock and the receiver clock are on exactly the same frequency. They will never drift out of alignment regardless of how many 1s there are in a row. Nonetheless, we need some way to get the receiver clock into the right phase at the start. So, we use the phase of the 31.25-Hz amplitude modulation to advance or retard the phase of the receiver clock. This brings it into the correct alignment (the decoder is clocked when the received waveform is expected to be at a peak). We do this with a slow time constant so that it isn't sensitive to noise spikes. A time constant of one second is about right to make sure we don't miss the first character after the one-second idle preamble. The most that the receiver clock phase will need to advance or retard is half a cycle in this first second.

Another way of saying this is that the receiver clock frequency must be capable of being offset by half a cycle per second, 0.5 Hz, as we say nowadays. This, in turn, means that if the transmitter and receiver clocks are *not* in fact on the same frequency, the system would still just stay locked with 0.5 Hz of clock-frequency difference. The receiver clock phase being advanced or retarded at its maximum rate. The frequency shift of 0.5 Hz is 1.6% of 31.25 Hz, but it occurs with the transmitter idling and the maximum amount of amplitude modulation. When the transmitter is sending continuous exclamation marks, the shift reduces to 0.3% since we only get 18% (2/11) of the amplitude modulation in this worst case. In practice, we can set the clocks much better than 0.3%, so there isn't a problem. Remember that we needed 30 ppm (that's 0.003%) for AMTOR, and that didn't bother us too much!

I have a few additional comments relating to W0IYH's letter. Bill writes, about his Fig A, that its shape is shown wrongly, as a half-sine, and that this shape has bad side lobes. In fact, if we think of this as the waveform at the transmitter, it *is* shown correctly, but there are no side lobes. Continuous half-sine reversals like this are exactly the same as a two-tone emission, the spectrum of which is two pure tones either side of the channel center. The zero crossing at the phase-reversal doesn't just minimize the spectral widening, it eliminates it completely, so long as the transmitter response is linear.

The point about the inequality of energy between the 1s and the 0s is interesting, but I don't know how important it really is. It is possible to remove this inequality by more complex filtering in the transmitter (root raised cosine), but this then has the undesirable side effect that the peak power is 3 dB higher for 0s than for 1s. This, in turn, means that if your transmitter is peak-power limited (they all are), you must back off the drive by 3 dB, transmitting less mean power than before.—*Peter Martinez, G3PLX, High Blakeland Farm, Underbarrow, Kendal, Cumbria, LA8 8HP, England;* **Peter.Martinez@btinternet.com**

There is a correction to the original article. The caption to Peter's Fig 4 should read "... the large ragged hump is random text being transmitted." That's our error, not his.—Ed.

Next Issue in QEX

The Jan/Feb 2000 issue of *QEX* will have come back from the printer before the lights go out on New Year's Day. We don't anticipate any delays in getting that issue out. If you find it's a few days late in arriving, though, we ask you to consider that the delivery system may be having a little trouble.

In the next *QEX*, Steve Hageman presents his 250-MHz synthesized signal source. The project combines "coherent direct synthesis" and EIA-232 serial control. A complete description of Steve's design approach is included, as well as performance data and detailed construction information for those of you who want to replicate this really useful piece of test gear.

Many of the design goals for a generator such as this also apply to local-oscillator circuits for homebrew transceivers. As Steve rightly points out, it can be a lot easier to get the frequencies you want than to remove those you don't! Check it out.

Guy Fletcher, VK2KU, brings us a new approach to the correction of Yagi element lengths based on "plumber's delight" construction. The article focuses on the effects of boom and element diameter on resonance. Guy provides unique insight into these subtleties of Yagi design, especially at frequencies where tube diameters become large fractions of a wavelength. His formulas and theory are supported by many painstaking measurements.

ARRL Technical Specialist Stu Downs, WA6PDP, has given intense study to a problem that has plagued many of us from time to time: automotive RFI. As Stu points out, its solution can be especially troublesome with regard to the HF bands. He gives a good physical explanation of the deleterious effects, followed by suggestions for isolating noise sources and some effective solutions. Should we send complimentary copies of this issue to auto manufacturers?

United States Postal Service

Statement of Ownership, Management, and Circulation

1. Publication Title	2. Publication Number	3. Filing Date
QEX	0886-8093	September 29, 1999
4. Issue Frequency	**5. Number of Issues Published Annually**	**6. Annual Subscription Price**
Bimonthly in: Jan, Mar, May, Jul, Sep, Nov,	6	$22.00

7. Complete Mailing Address of Known Office of Publication *(Not printer) (Street, city, county, state, and ZIP+4)*
225 Main St., Newington, Hartford County, CT, 06111-1494

Contact Person: Debra Jahnke
Telephone: (860) 594-0297

8. Complete Mailing Address of Headquarters or General Business Office of Publisher *(Not printer)*
225 Main St., Newington, Hartford County, CT, 06111-1494

9. Full Names and Complete Mailing Addresses of Publisher, Editor, and Managing Editor *(Do not leave blank)*

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David Sumner, 225 Main Street, Newington, CT, 06111-1494

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American Radio Relay League, Inc.	225 Main St., Newington, CT 06111-1494

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13. Publication Title: QEX		14. Issue Date for Circulation Data Below	
15. Extent and Nature of Circulation		**Average No. Copies Each Issue During Preceding 12 Months** (Sep 98 - Jul 99)	**No. Copies of Single Issue Published Nearest to Filing Date** (Sep 99)
a.	Total Number of Copies *(Net press run)*	6,371	6,595
b. Paid and/or Requested Circulation	(1) Paid/Requested Outside-County Mail Subscriptions Stated on Form 3541. *(Include advertiser's proof and exchange copies)*	4,388	4,750
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	(3) Sales Through Dealers and Carriers, Street Vendors, Counter Sales, and Other Non-USPS Paid Distribution	185	227
	(4) Other Classes Mailed Through the USPS	1,035	1,124
c.	Total Paid and/or Requested Circulation *[Sum of 15b. (1), (2),(3),and (4)]*	5,608	6,101
d. Free Distribution by Mail *(Samples, compliment ary, and other free)*	(1) Outside-County as Stated on Form 3541	69	65
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e.	Free Distribution Outside the Mail *(Carriers or other means)*	0	0
f.	Total Free Distribution *(Sum of 15d. and 15e.)*	88	76
g.	Total Distribution *(Sum of 15c. and 15f)*	5,696	6,177
h.	Copies not Distributed	675	418
i.	Total *(Sum of 15g. and h.)*	6,371	6,595
j.	Percent Paid and/or Requested Circulation *(15c. divided by 15g. times 100)*	98.45%	98.77%

16. Publication of Statement of Ownership
[XX] Publication required. Will be printed in the Nov/Dec 1999 issue of this publication. [ ] Publication not required.

17. Signature and Title of Editor, Publisher, Business Manager, or Owner
Business Manager
Date: September 29, 1999

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(All dates are 1999 unless otherwise indicated.)

Upcoming Conferences

The Southeastern VHF Society will host its fourth annual conference on April 14-15, 2000, at the Atlanta Marriott Northwest (I-75 and Windy Hill Rd, Marietta, Georgia; the same location as the first three conferences.) This is the *first call* for presentations to be made at the 2000 conference and papers to be published in the conference *Proceedings* (submissions due by February 18, 2000). Contact program chairman Bob Lear, K4SZ, at PO Box 1269, Dahlonega, GA 30533; tel 706-864-6229; **k4sz@arrl.net**. In addition to the technical program, there will be preamp noise-figure testing, antenna-gain measurements, a flea market, banquet and door prizes. Further details will be announced here and at **http://www.svhfs.org/svhfs/**.

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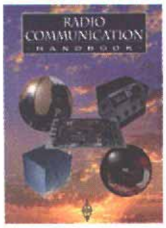

Covers semiconductors, HF and VHF/UHF receivers, transmitters and transceivers, microwaves, propagation, HF and VHF/UHF antennas, amateur satellites and space communications, image techniques, power supplies, and much more. PCB layouts included. 760 pages.
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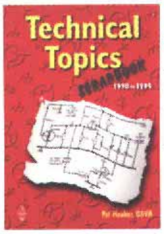

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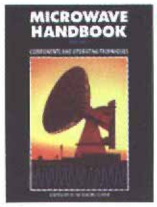

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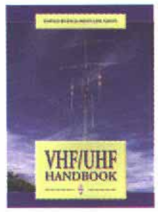

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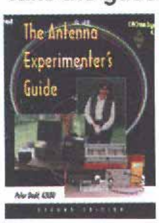

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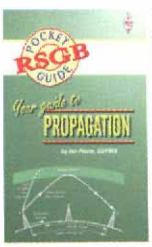

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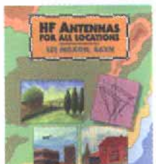

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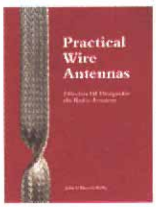

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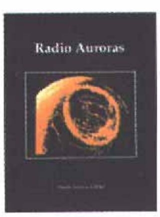

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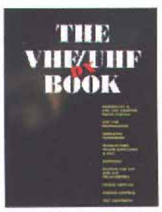

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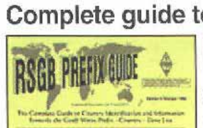

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