COMMUNICATIONS QUARTERLY

THE JOURNAL OF COMMUNICATIONS TECHNOLOGY

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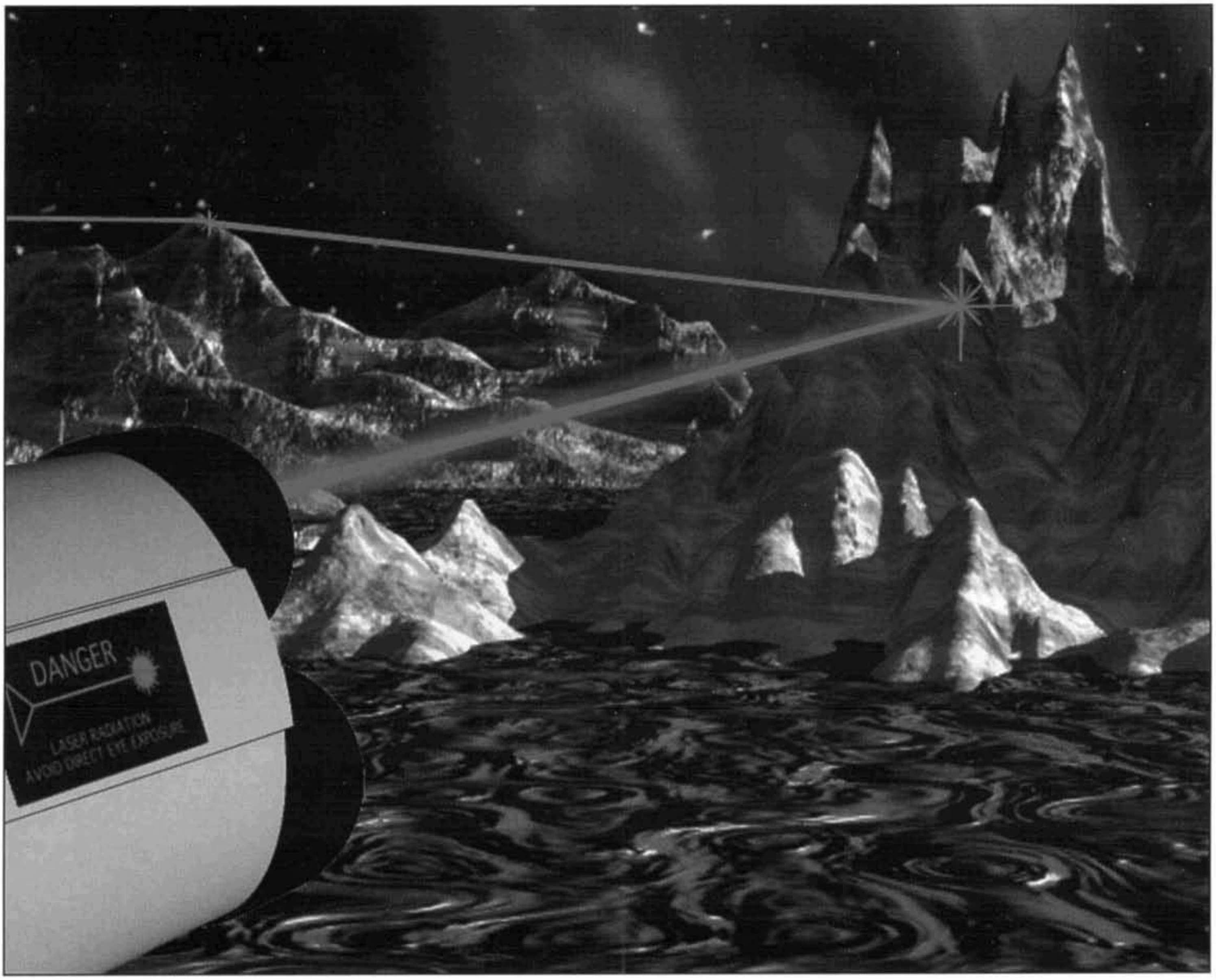

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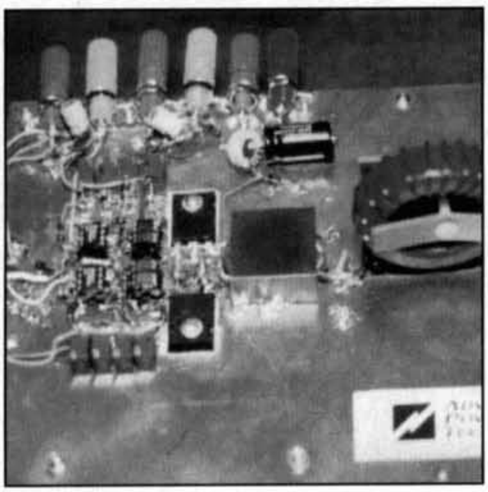

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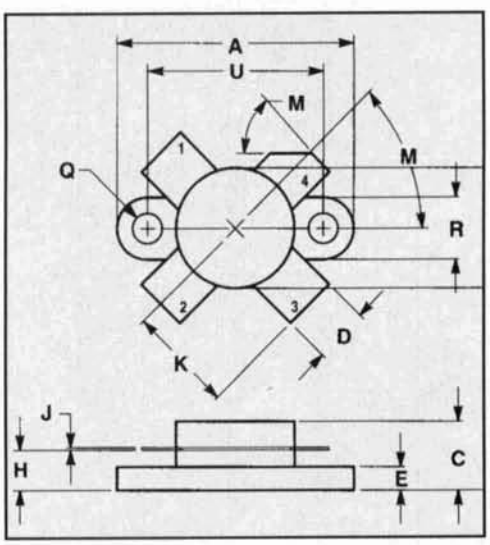

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On the Cover: There has always been a group of radio amateurs interested in pushing communication toward higher frequencies. In this month's feature "Optical Communications" Richard Bitzer, WB2ZKW shows you how. Photo by Bryan Bergeron, NU1N.

EDITORIAL

Surplus Test Equipment—Bonanza or Bust???

Readers of the many amateur radio publications are familiar with photos depicting the best operators and shacks in the country. These guys have Alpha amplifiers, an array of antenna control boxes (presumably attached to stacked multiband monobanders on 65-meter towers), and the best radios that money can buy—all displayed in a 40 by 40 room with carpeting and paneling.

On the other end of the spectrum, there are those of us with workbenches bigger than our operating desks, rigs that are 20 years old (that still work great!), and more test equipment than radios. And, if you check in the "reading room," you'll probably find the current Tucker Test and Measurements Catalog, instead of the latest amateur radio buyers guide! Most of us are surplus equipment junkies. You'll find us scouring the flea market tables at hamfests across the country looking for the right deal.

The recent rash of military base closings has flooded the market with state-of-the-art test equipment—often at prices a fraction of the original cost to the government. The time appeared right to upgrade my workbench arsenal. Unfortunately, for the novice, the surplus market can be both unforgiving and costly. For instance, one item on my shopping list is a good spectrum analyzer. One model that caught my eye has been listed for as high as $2000. It's now being offered for $1500—if you search out the lowest price. Because the unit is both portable and solid-state, offers coverage from 10 MHz to 40 GHz, and is manufactured by one of the top names in the industry, you'd expect it to be a bargain. Unfortunately, the fine print usually fails to point out that this analyzer uses two ceramic triodes that cost about $400 each—if they can be found—and a TWT than can't be replaced. Add the fact that the analyzer can only display 100 MHz on screen (due to a swept IF that operates from 150 to 250 MHz), and you've got problems. The IF range makes the analyzer particularly unsuited for most VHF work—unless you're willing to suffer with false on-screen signals due to mixer blow-by! What was an engineering marvel some 20-odd years ago, is a plague today.

Is this an isolated example? No. Recently, I found two synthesized signal generators that cover from 1 to 520 MHz in 1 kHz steps—again produced by a well-known American manufacturer. These generators currently sell for about $800 to $1995. On the surface, these units appear to be a bargain: an excellent attenuator system with accurate calibration, FM and AM modulation capabilities, and small size. I considered myself both shrewd and lucky to find two of these units in nonworking, but complete, condition for the normal dealer cost of $250 each. Sure, they were broken and fixing them took time, but I enjoyed the challenge. Was there a catch? You bet!! The noise floor of the beasts was, at best, 70 dBc, from 0 to 520! While this is fine for SINAD or sensitivity tests, it's certainly unsuited for receiver IMD or blocking measurements. And, with everything locked to an uncompensated 40-MHz overtone oscillator, warm-up drift exceeds 4 kHz. An expensive lesson learned, but it could have been several times worse! Alas, for a few dollars more, I could have purchased a military version of the HP8640, an industry standard for many years. However, even this venerable and reliable workhorse can contain hidden pitfalls. HP stopped producing many of the hybrid modules used in this model back in 1990; often the only source of replacement parts is from a cannibalized unit.

Another item to be wary of is "options." Most models have a variety of available options; some are most desirable, some are not! For example, the designation "option 001" may indicate that a specific frequency counter includes a more accurate TXCO timebase. On other equipment, it may indicate something as mundane as the presence of a tilt bail for easier viewing. Option 001 could also forewarn that a signal generator or spectrum analyzer is a 75, not 50 ohm, unit! Be sure to question the dealer as to which options are installed in a unit you may be interested in buying!

Luckily, all my purchases weren't as ill-advised as the RF generators. I found a Boonton 92AS2 RF milli-voltmeter for $150. The meter came with the original probes and 50-ohm termination adapter, with serial numbers that matched the unit. It was calibrated, and works as well as those selling for about seven times the amount I paid. A catch? Again, yes. But I knew beforehand to make sure the vendor guaranteed the matching serial numbers, and I knew replacing the mechanical chopper used in these instruments could be expensive if it ever failed. If there's a lesson to be learned here, it's to network!!! Had I invested the same amount of time I spent investigating the spectrum analyzer in buying those RF generators, I could have saved myself some money and embarrassment! I talked to a few *CommQuart* writers who are RF engineers, and learned that they had the same bad experiences with the particular series of generators I own. I could have been forewarned before jumping into a deal I knew little about.

So what about the spectrum analyzer? Rather than buy, I decided to build my own! As senior technical editor, I know spectrum analyzer how-to articles will highlight the summer 1996 issue of *Communications Quarterly*.

The bottom line is that, if you're interested in purchasing surplus test equipment, there are tremendous bargains to be found—along with expensive potential "white elephants." If you're brave, you can save better than 50 percent if you're willing to accept equipment needing repair or calibration. You can save even more if you're willing to waive the usual right-of-return offered by most reputable dealers. On the other hand, expect to pay a small premium for a calibrated and tested unit backed by a an extended guarantee. Be an informed shopper; if it looks to good to be true, it just may be . . .

Peter Bertini, K1ZJH
Senior Technical Editor

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TECHNICAL CONVERSATIONS

Dear Editor:

The following is based on physical interactions plus correlation with historical sunspot data for the years 1701 to 1994. The correlation combines a 22 year cycle modulated by a cycle of 264 years. There is evidence of a longer cycle, but the data is not sufficient to determine its parameters.

The correlation indicates that the current cycle 22 will end in 1966.5, and that cycle 23 will end in the year 2007.5. The estimated error is ±0.5 years. Note that the first spots of cycle 23 have appeared, but that the exact point of cycle end cannot be determined until the cycle has been passed.

The magnitude of cycle 23 is predicted for the period 2000–2001, with a peak sunspot number of 140. The error in this prediction is some ±40 units.

A better indication of the peak sunspot number can be obtained from the change in sunspot number for the first full year of the cycle, from the relation SSpeak=60+1.37*SSchange. Solar flux can be used with the usual conversion to sunspot numbers.

A report on these techniques and detailed results is in preparation.

R.P. Haviland, W4MB
Daytona Beach, Florida

Dear Editor:

Doesn't anyone use Smith charts anymore? I thought of this when I read the article on stub matching in the Fall 1995 issue of Communications Quarterly.

Stub matching using a Smith chart is ridiculously easy. Why go to a computer?

The Smith chart lets you see what is happening to a transmission line all along its length; it gives you a picture, something that a computer cannot do.

I really regret that the wonderful invention of Philip Smith seems to have fallen into disuse.

Harry R. Hyder, W7IV
Tempe, Arizona

Charles Kitchen, N1TEV, received this letter about his article "Regenerative Receivers" in our Fall 1995 *issue.*

Dear OM:

Congratulations on your "Regenerative Receivers" article, which was published in the Fall 1995 issue of *Communications Quarterly*. It took me back almost 70 years and at that time I did not realize what an engineering marvel my regenerative detector and two audio stages was. It was not only a great receiver, but a fair transmitter with across town range. I used to loop modulate (carbon mike) the receiver, and friends across town could receive me. Start of narrow band FM?

Several tubes were tried as the detector. As I recall, the screen voltage of the type 224 (RF pentode) with regeneration control was the smoothest in and out of oscillation. I do not recall if I used AC or DC on the filaments of the 224. With some triodes as detectors when the tube went in and out of oscillation, there was a slight howl. This condition was cured by shunting the primary of the first audio transformer with a resistor.

If I had more ambition, I would build the FET detector together with a type 47/210 transmitter and try it on today's bands.

Your trick of microwave coil forms is new to me. We have had a microwave oven for only five years. Thanks for the tip.

Keep up the good work.

Johnny Johnson, W1JY
Bristol, New Hampshire

Here's another letter N1TEV received on his regen article. Two thumbs up, Charles!

Dear Charles:

Congratulations! I thoroughly enjoyed your article in the Fall issue of *Communications Quarterly.*

In the early 1930s, when I was in my early teens, I spent a lot of time building regenerative sets, using mostly parts from discarded broadcast sets of the 1920s. I changed my set almost monthly, always looking for that "perfect" circuit.

I tried tickler regeneration, cathode tap regeneration, plate voltage control, screen voltage control and capacitor control. They all seemed to work about the same, but I had fun. When I got my ticket for my first couple of years on the air I used a two-tube regenerator before graduating to a five tube home-built superhet.

I last built a regenerative set in 1986. I described it in *QST* September 1986: "A 1935 Ham Receiver."

You brought back to me many fond memories.

73,

Harry R. Hyder, W7IV
Tempe, Arizona

Dear Editor:

I am not renewing my subscription to *Communications Quarterly* for several reasons, three of which are as follows:

1. Spring 1994 issue—17 pages devoted to arc and spark transmitters (pp 27–43).

2. Fall 1994 issue—14 pages devoted to superregeneration (pp 27–40). Your editors should be aware that *superregeneration* is one word, according to *Webster's Third New International Dictionary*, not two words.

3. Fall 1995 issue—20 pages devoted to regenerative receivers (pp 7–26).

You claim to be the successor to *Ham Radio* magazine. I knew the late Jim Fisk personally and wrote many articles for the magazine, some specifically at his request. Jim would never have published material such as that indicated above.

Robert S. Stein, W6NBI
Los Altos, California

Those of you who enjoyed Peter Bertini, K1ZJH's, editorial "Thank You Mr. Morgan!" in the Fall 1995 *may find the following correspondence interesting.*

Dear Mr. Bertini:

I noticed your editorial on Alfred Morgan in the latest *Communications Quarterly.* I also started out with one of his books, but a much older one, in the local public library when I was 12. I might have build a one-tube radio too, but my only information on that subject was in a Cub Scout Handbook which, as I look back at it now, was incomplete and misleading. When Morgan wrote the book I had, vacuum tubes were barely invented. I'm enclosing a page from the Unites States Catalog showing books published from 1912 to 1917, also the chapter on Morgan's manufacturing company Adams-Morgan (Paragon) from my books on radio manufacturers. Morgan, himself, had more success as a writer than as a manufacturer, and Adams-Morgan did not survive past the mid-1920s (only a few did, as radio moved from hamshacks to big business).

Morgan's electrical books are hard to find nowadays, and I thing that's because they were all worn out by kids! I've never seen another copy of *The Boy Electrician*, and when I went back to the library twenty years ago or more, that book was nowhere to be found.

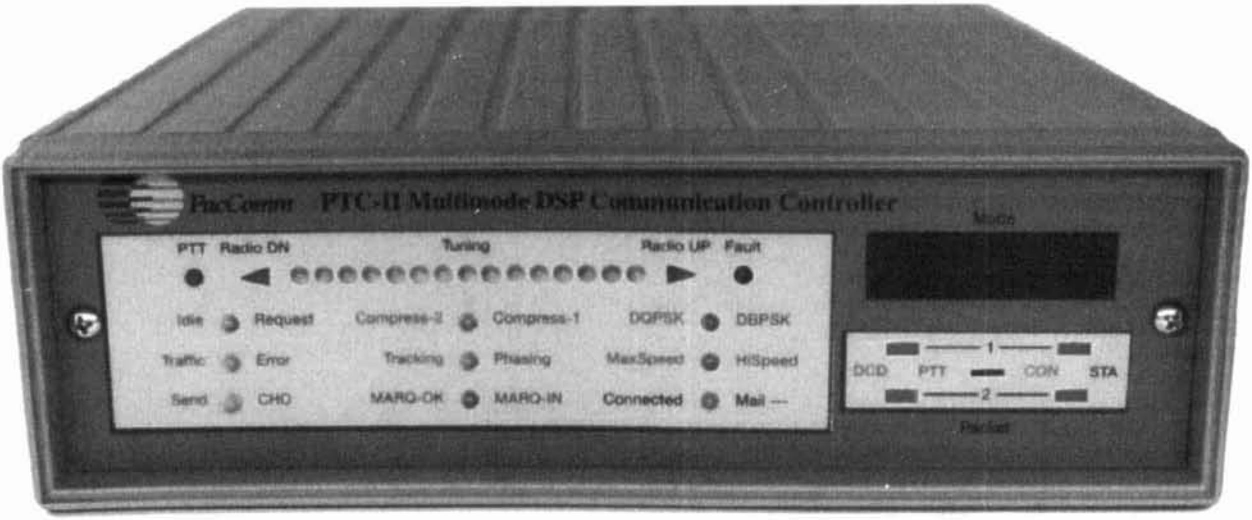
Alan Douglas
Pocasset, Massachusetts

Dear Mr. Douglas:

Thank you for your letter and the related material concerning Alfred Powe Morgan. Although I am an antique radio collector, I had never made the connection between Alfred Morgan and the Adams-Morgan or the Paragon radio manufacturing companies! This information certainly places Mr. Morgan in a different historical perspective in my eyes—dating him with other such radio pioneers as Armstrong and Godley! It is even more amazing how the timeless value of his books have influenced so many generations of young readers.

I suspect many of the Morgan books have ended up in the hands of Paragon radio collectors—as often related ephemera is as sought after as the radios themselves. Perhaps this explains the numerous ads seeking these volumes that have appeared in *Antique Radio Classified* over the past several years. But, I would rather believe, as you do, that these books have all been worn out by kids of all ages.

Peter Bertini, K1ZJH
Senior Technical Editor



You've been hearing about it for several years.... Now it is available. PACTOR-II.

The PTC-II is a new multi-mode controller and "communications platform" which contains very powerful and flexible hardware and firmware.

It is being built in the United States by PacComm under license from S.C.S., the group that developed both the original PACTOR and PACTOR-II.

The PTC-II offers the most robust HF digital protocol available to radio amateurs, but it should not be overlooked that the PTC-II is configurable as a triple-port multimode controller supporting packet data rates of 1200 and 9600 bps and numerous other modes.

What is PACTOR-II?

Like PACTOR-I, PACTOR-II is a communication system. There is a unique protocol, carefully designed and optimized for excellent HF performance, new and powerful modems realized via Digital Signal Processing, and a hardware platform to allow the firmware to realize its full operating potential.

PACTOR-II is the name of the new protocol including modulation specifications. PTC-II is the name of this specific hardware realization of the PACTOR-II architecture.

PACTOR-II offers all the following advantages:

- A step-synchronous ARQ protocol.
- Full support of memory ARQ.
- Independent of sideband; no mark/space convention.
- Center frequency adjustable between 400 and 2600 Hz to exactly match your radio's filters.
- Long-path capability for worldwide connectivity.
- Full compatibility with PACTOR-I (the original PACTOR system), AMTOR, and RTTY.
- All-mode mailbox with up to 32 megabytes of storage.
- Automatic switching between Level-1 (PACTOR-I) and Level-2 (PACTOR-II) at contact initiation.
- Occupies a bandwidth of under 500 Hz - use your 500 Hz CW filters.
- Constant bandwidth irrespective of actual transmission speed.
- Differential Phase Shift Keying with two continuously transmitted carriers. 100 symbols per second.
- DBPSK modulation yields 200 bps (uncompressed). DQPSK modulation yields 400 bps (uncompressed). 8-DPSK modulation yields 600 bps (uncompressed). 16-DPSK modulation yields 800 bps (uncompressed).
- Powerful Forward Error Correction (FEC):
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- All units available in 220 VAC input voltage (except for SL-11A)

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MODEL	Colors Gray	Colors Black	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H × W × D	Shipping Wt. (lbs.)
SL-11A	•	•	7	11	2⅝ × 7⅝ × 9¾	12
SL-11R	•	•	7	11	2⅝ × 7 × 9¾	12
SL-11S	•	•	7	11	2⅝ × 7⅝ × 9¾	12
SL-11R-RA		•	7	11	4¾ × 7 × 9¾	13

RS-L SERIES

• POWER SUPPLIES WITH BUILT IN CIGARETTE LIGHTER RECEPTACLE

MODEL	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H × W × D	Shipping Wt. (lbs.)
RS-4L	3	4	3½ × 6⅛ × 7¼	6
RS-5L	4	5	3½ × 6⅛ × 7¼	7

RM SERIES

MODEL RM-35M

• 19" RACK MOUNT POWER SUPPLIES

MODEL	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H × W × D	Shipping Wt. (lbs.)
RM-12A	9	12	5¼ × 19 × 8¼	16
RM-35A	25	35	5¼ × 19 × 12½	38
RM-50A	37	50	5¼ × 19 × 12½	50
RM-60A	50	55	7 × 19 × 12½	60
• Separate Volt and Amp Meters				
RM-12M	9	12	5¼ × 19 × 8¼	16
RM-35M	25	35	5¼ × 19 × 12½	38
RM-50M	37	50	5¼ × 19 × 12½	50
RM-60M	50	55	7 × 19 × 12½	60

RS-A SERIES

MODEL RS-7A

MODEL	Colors Gray	Colors Black	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H × W × D	Shipping Wt. (lbs.)
RS-3A		•	2.5	3	3 × 4¾ × 5¾	4
RS-4A	•	•	3	4	3¾ × 6½ × 9	5
RS-5A		•	4	5	3½ × 6⅛ × 7¼	7
RS-7A	•	•	5	7	3¾ × 6½ × 9	9
RS-7B	•	•	5	7	4 × 7½ × 10¾	10
RS-10A	•	•	7.5	10	4 × 7½ × 10¾	11
RS-12A	•	•	9	12	4½ × 8 × 9	13
RS-12B		•	9	12	4 × 7½ × 10¾	13
RS-20A	•	•	16	20	5 × 9 × 10½	18
RS-35A	•	•	25	35	5 × 11 × 11	27
RS-50A	•		37	50	6 × 13¾ × 11	46
RS-70A	•		57	70	6 × 13¾ × 12⅛	48

RS-M SERIES

MODEL RS-35M

MODEL	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H × W × D	Shipping Wt. (lbs.)
• Switchable volt and Amp meter				
RS-12M	9	12	4½ × 8 × 9	13
• Separate volt and Amp meters				
RS-20M	16	20	5 × 9 × 10½	18
RS-35M	25	35	5 × 11 × 11	27
RS-50M	37	50	6 × 13¾ × 11	46
RS-70M	57	70	6 × 13¾ × 12⅛	48

VS-M AND VRM-M SERIES

MODEL VS-35M

• Separate Volt and Amp Meters • Output Voltage adjustable from 2-15 volts • Current limit adjustable from 1.5 amps to Full Load

MODEL	Continuous Duty (Amps) @13.8VDC	@10VDC	@5VDC	ICS* (Amps) @13.8V	Size (IN) H × W × D	Shipping Wt. (lbs.)
VS-12M	9	5	2	12	4½ × 8 × 9	13
VS-20M	16	9	4	20	5 × 9 × 10½	20
VS-35M	25	15	7	35	5 × 11 × 11	29
VS-50M	37	22	10	50	6 × 13¾ × 11	46
• Variable rack mount power supplies						
VRM-35M	25	15	7	35	5¼ × 19 × 12½	38
VRM-50M	37	22	10	50	5¼ × 19 × 12½	50

RS-S SERIES

MODEL RS-12S

• Built in speaker

MODEL	Colors Gray	Colors Black	Continuous Duty (Amps)	ICS* Amps	Size (IN) H × W × D	Shipping Wt. (lbs.)
RS-7S	•	•	5	7	4 × 7½ × 10¾	10
RS-10S	•	•	7.5	10	4 × 7½ × 10¾	12
RS-12S	•	•	9	12	4½ × 8 × 9	13
RS-20S	•	•	16	20	5 × 9 × 10½	18
SL-11S	•	•	7	11	2¾ x 7⅝ x 9¾	12

*ICS—Intermittent Communication Service (50% Duty Cycle 5min. on 5 min. off)

Richard Bitzer, *WB2ZKW*
7 Oaktree Road
Monmouth Junction, New Jersey 08852

OPTICAL COMMUNICATIONS

Equipment for the radio amateur

There has always been a group of radio amateurs interested in pushing communication toward higher frequencies. Today there's a respectable amount of activity in the microwave bands, so it's not unusual to expect the next push to be into the near infrared and optical frequencies. Transmitters and receivers for the optical frequencies center around the use of lasers and a variety of photodetectors that are preceded by narrow filters tuned to the laser frequency in use. Very efficient solid-state semiconductor lasers are now coming down in price. However, the main drawback to their use is the collimating optics required because their optical output is fan-shaped rather than focused in a beam. I'll describe a laser transmitter and optical receiver that can be built with equipment purchased at ham flea markets and through parts houses that cater to the radio amateur who wants to experiment in the optical frequency range.

The laser transmitter

The HeNe laser is easy to find at ham flea markets and through surplus dealers. Its main advantage is that it can project a narrow beam over distances of several miles without external optics. The HeNe laser typically consists of a set of concentric gas-filled tubes with internal electrodes to initiate and maintain a gas discharge. The thinner tube, called a capillary tube, has mirrors at each end that act as an optical resonator and concentrate the discharge for higher efficiency. The capillary tube is surrounded by a reservoir tube that contains a mixture of Helium and Neon gases—typically about five times more Helium than Neon, up to

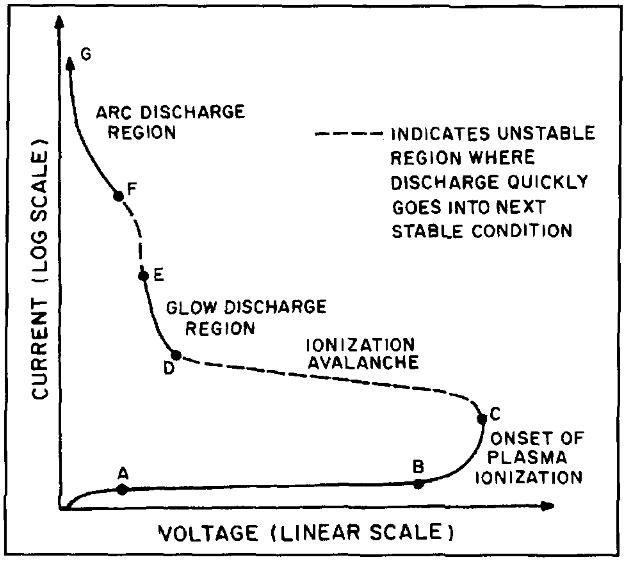

Figure 1. Volt-ampere characteristics of a gas discharge.

several Torr of pressure. This discharge is initiated by a high-voltage pulse in the 10 kV range. Once the discharge has started, the ignition voltage is turned off and a sustaining power supply of 1.2 to 3 kV is required to maintain the discharge.

The HeNe laser follows the typical volt-ampere characteristics of a gas discharge, as shown in **Figure 1**. The current is very low as the voltage is increased from A to B, typically in the nanoamperes range. From B to C, the current begins to increase at the onset of plasma ionization. For a HeNe laser, the voltage at this point is typically around 10 to 12 kV. Once the ionization avalanche is reached, the gas becomes increasingly ionized and the voltage drops to point D. This is the region where the

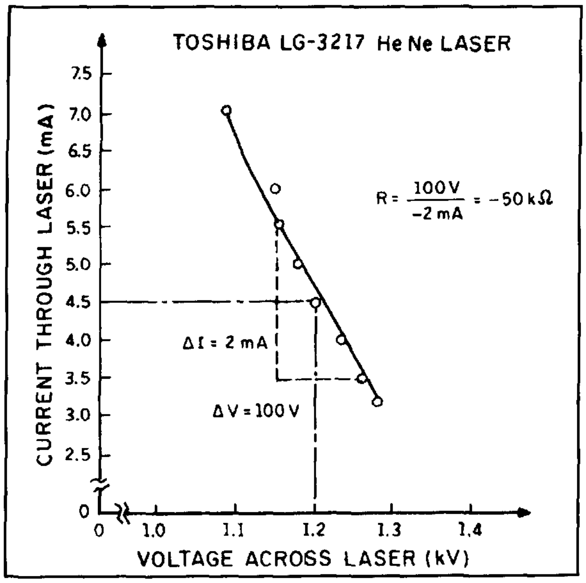

Figure 2. V-I characteristics of Toshiba LG-3217 laser.

laser typically operates; i.e., 1.2 to 3 kV at 4 to 8 mA. As the plasma becomes more fully ionized, the current rises rapidly to the arc discharge region between F and G. The region of normal operation, between D and E, exhibits negative resistance. This is indicated by a decrease in voltage across the discharge as the current increases through the discharge. The HeNe laser requires a ballast resistor to stabilize the discharge and limit the current through the laser tube.

The explanation above shows the necessity of an ignition voltage to initiate ionization (breakdown of gas), and a second supply to maintain discharge in a HeNe gas laser circuit (the former has been neglected in many HeNe laser articles). To start, I'll describe the typical operating parameters of a gas laser upon which the appropriate power supplies can be designed. I obtained a set of Toshiba HeNe lasers at a flea market and will use these as an example.

Figure 2 shows the measured V-I characteristics of a typical Toshiba LG-3217 HeNe laser. This corresponds to region D through E of **Figure 1.** The laser exhibits a negative resistance characteristic within the desired operating region between 3.5 and 5.5 mA. A quiescent operating point of 4.5 mA at 1.2 kV is chosen. The slope of the curve at the Q point is -50 k. The minus sign indicates a negative resistance in this region. A "ballast" resistor is placed in series with the laser anode circuit whose value is equal to or greater than the negative resistance of the discharge. The resistance found in a commercial Toshiba unit is 51 k, with a 2-watt dissipation. This satisfies the need for a series resistor to stabilize the gas discharge. The laser power supply must provide a voltage at least equal to the laser voltage drop plus the voltage drop across the ballast resistor, while maintaining the discharge current through the tube. The sustaining supply would then require it to provide around a 1.45 kV output of at least 5 to 7 mA. The ballast resistor *must* be placed close to the laser anode terminal if stable operation is to be maintained. If the ballast resistor chosen is too large, the discharge can't be maintained, and the laser won't start. If the ballast resistor is too small, the current through the discharge will be high, and there is danger the laser will be destroyed.

Laser power supplies

The HeNe laser requires two voltages: a voltage in the range of 10 kV that starts the laser and then turns off once the discharge begins, and a lower voltage power supply to sustain the discharge. The circuit diagrams of **Figures 3** and **4** illustrate the two different ways in which a momentary 10-kV ignition voltage can be connected to the sustaining supply. **Figure 3** illustrates one method of generating the laser ignition voltage. A fraction of the main supply voltage is used to charge up a capacitor. When triggered, it's discharged through a high-voltage ignition transformer in series with the main power supply output (a television flyback transformer will do the job nicely). When the supply is first turned on, a delay circuit allows the main supply to stabilize at full voltage and charge up the capacitor. A nonlatching relay operates to discharge the capacitor on a one-shot basis into the ignition coil after the timing delay. The HV diode in series with the ignition coil and the capacitor across the supply rectifies the damped oscillatory waveform out of the ignition coil, producing a single positive pulse across the laser tube ionizing the gas in the laser. Once the laser is ignited by the HV pulse, the main power supply maintains the discharge.

Figure 4 illustrates a second way in which a momentary 10 kV voltage pulse can be generated to initiate the plasma discharge across the laser tube. A diode voltage multiplier circuit is connected in series with the main supply and obtains its input power across one of the voltage multiplier diodes in the main supply. The voltage across this diode is typically 1.8 kV p-p. With a ten-section multiplier, the output voltage is approximately 9 kV in series with a 1.8 kV sustaining supply. The capacitance value of the capacitors in the multiplier chain are much smaller than those of the main supply. When the power supply is first turned on, the capacitors in the voltage multiplier charge up to the ignition voltage. As the plasma forms, the laser tube draws more current and the multiplier capacitors can't maintain their charge. As a result, the volt-

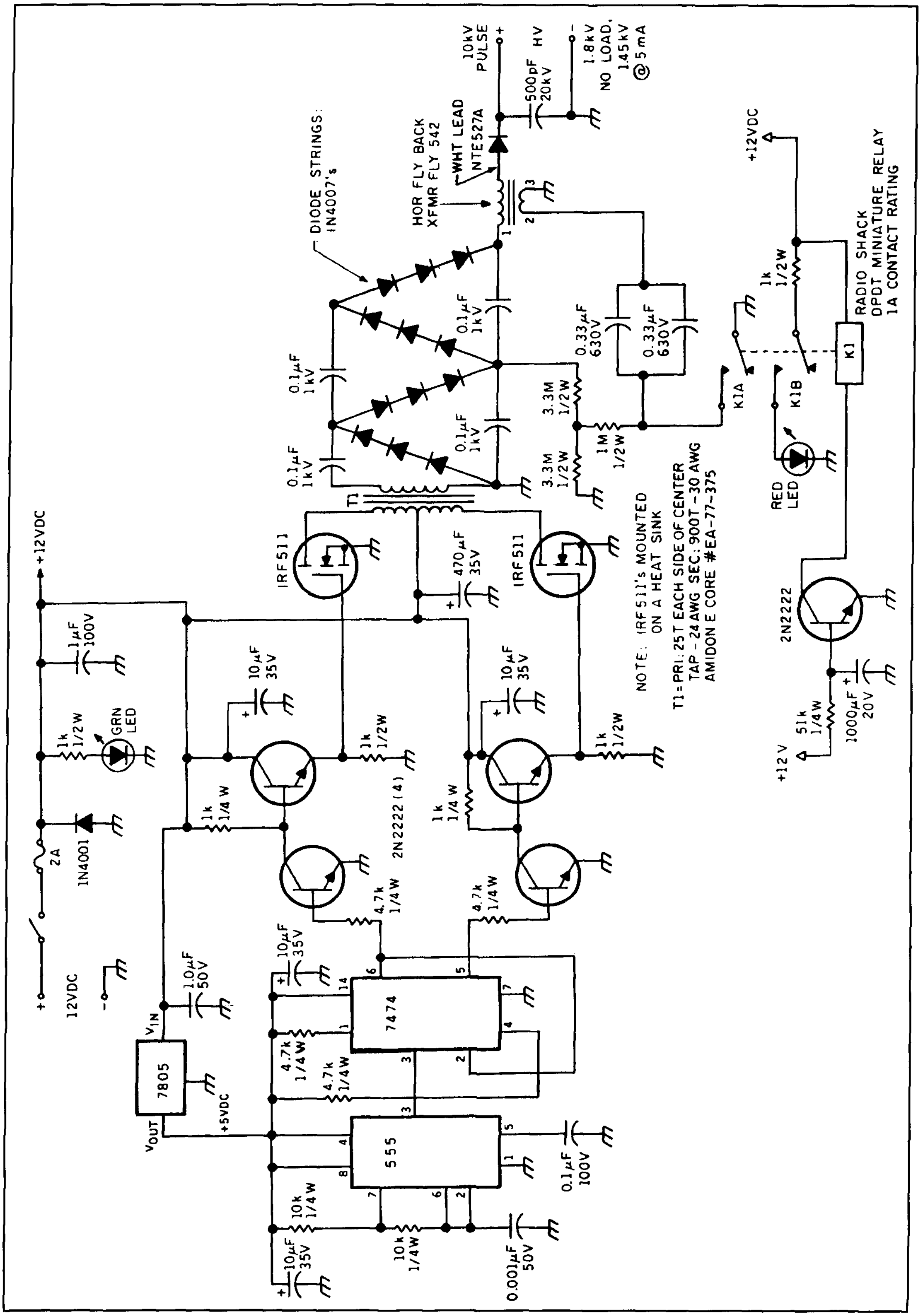

Figure 3. HeNe laser DC converter with HV pulse ignition.

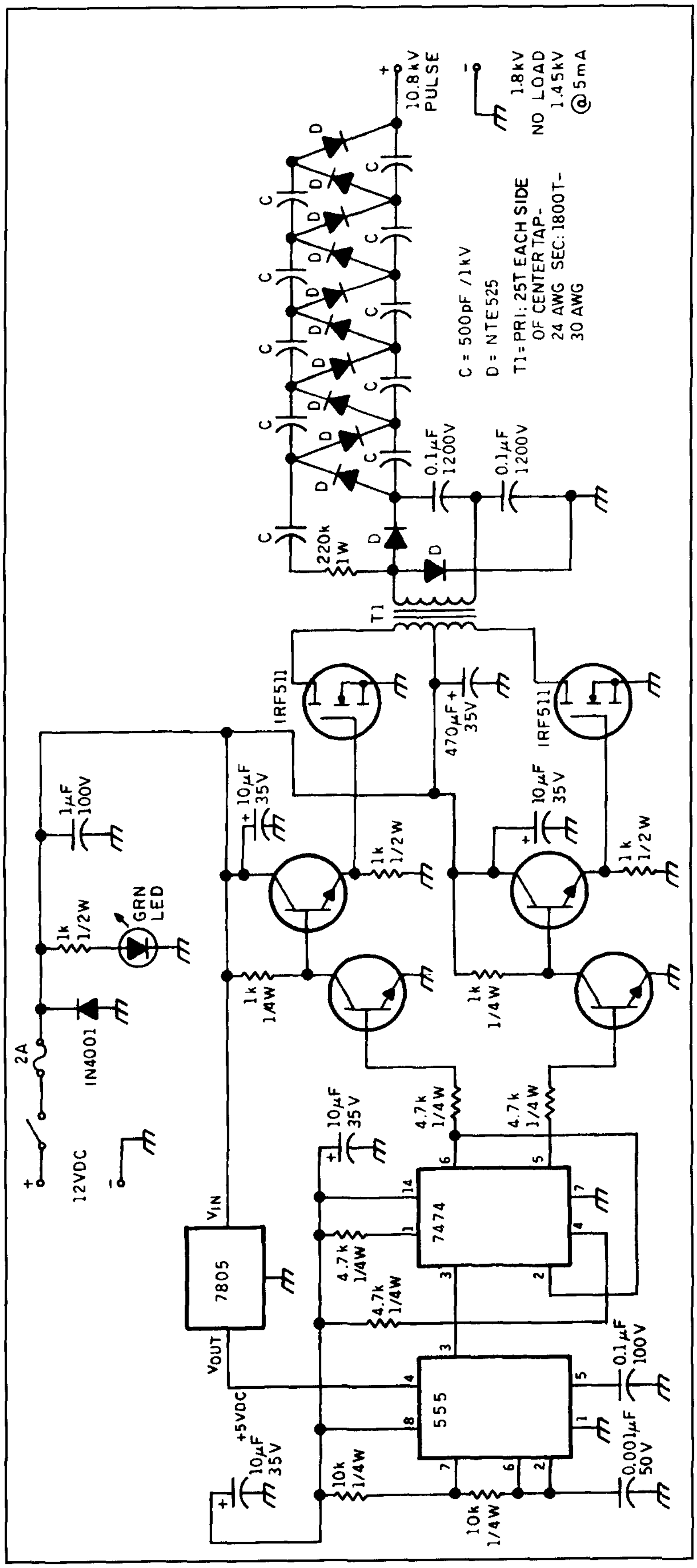

Figure 4. HeNe laser DC converter with HV multiplier ignition.

age immediately drops to that of the main sustaining supply, with all the diodes in the multiplier chain forward biased. In this manner, an HV pulse is generated to ionize the gases.

All rectifier diodes used in these circuits must have low-junction capacitance. Because the voltage across the transformer secondary is a square wave, the efficiency of the supply is inversely proportional to the junction capacitance, as the switching time of the rectifier is kept small compared to the square wave period. If high-speed rectifier diodes aren't available, up to four general-purpose high-voltage silicon diodes can be connected in series to reduce the total junction capacitance.

The circuits of the main sustaining high-voltage supplies shown in **Figures 3** and **4** are identical up to the transformer. The supply is a switching power converter topology operated from a 12 volts DC power source like a car battery. The 555 IC generates a nonsymmetrical 40-kHz square wave. The 7474 flip-flop will provide two symmetrical square waves opposite in phase at half the converter frequency (20 kHz) to drive the switching FETs. These 5-volt p-p outputs must be stepped up to 10 volts p-p by low-impedance drivers that drive the FETs between cutoff and saturation. A simple, direct-coupled transistor driver pair is used here. The first transistor acts as an amplifier, while the second acts as an emitter-follower providing a low-impedance source to drive the large input capacitance of the FETs. The power FETs are toggled through the transformer producing a high-voltage square wave across the transformer secondary. Low-capacitance rectifier diodes with fast switching times are used on a full-wave voltage doubler configuration to produce a 1.8 kV output.

Both sustaining supplies for the laser in the example above use an oscillator providing push-pull signals to a pair of FETs. This is by no means the only way to drive the FET switches. There are many laser high-voltage supply circuit topologies available. This circuit topology was chosen because it allows a non-saturatable transformer to be used, which simplifies transformer design.

There are two important cautions to observe when working with lasers and high-voltage power supplies:

(1) All precautions must be observed, as these are high-voltage supplies.
(2) Never look into a laser output or at a fully reflected laser output beam.

Laser modulation

Many lasers can be modulated by two methods: (1) varying the current through the laser (referred to as pump modulation) and (2) modulating the laser light beam external to the laser. The HeNe laser can only be modulated by varying current through the discharge up to a maximum of 15 percent. The HeNe plasma column in the laser acts as a constant current load and tends to resist changes in current. If the current modulation is pushed beyond the 15-percent range, the plasma extinguishes, and the laser stops functioning. Therefore, for short-haul communication links, this shallow modulation percentage is satisfactory because it provides a simple means of modulating the light output of the laser. When the HeNe laser is to be used for DX work, the output beam must be modulated directly by either a mechanical chopping wheel and shutter for CW or an electro-optic modulator. The latter is very expensive. I'll describe laser current (pump) modulation because it's the most straightforward to implement.

To determine the modulation current range for the HeNe laser one first measures its V-I curves. You'll remember that the Toshiba LG-3217 typically operates with approximately 1.2 kV across the tube drawing 5 mA, providing an optical output power of 1 mW. The V-I characteristics were measured as shown in **Figure 2**. During these measurements, the optical output was also measured as a function of current (see **Figure 5**). The V-I curves indicate a device with a negative resistance characteristic, specifically 50 k. Looking at the optical output of the laser as a function of laser current, the optical output is linear between 3.4 and 4.8 mA. For linear modulation, a change in 1.4 mA centered around 4.1 mA will cause a linear change in optical output. A meter connected at the output of an optical detector shows the light causing a current change from 151 µA to 183 µA in a linear fashion. Therefore, the percentage modulation is:

%mod=100x(183-151)/(183+151)=9.6%

This is the best modulation that can be accomplished with this laser, as extending the current range downward causes nonlinear modulation and eventually extinguishes the plasma. Pushing the current upward causes optical saturation and increased laser dissipation. For this HeNe laser, the quiescent operating point is set at 4.1 mA with the current varying ±0.7 mA around this value, which can be verified by a current probe when placed in the ground lead of the laser.

Pump modulation (changing the laser current) can be accomplished by placing a transformer in series with the ground lead to the laser, as shown in **Figure 6**. This is by far the simplest means of intensity-modulating the

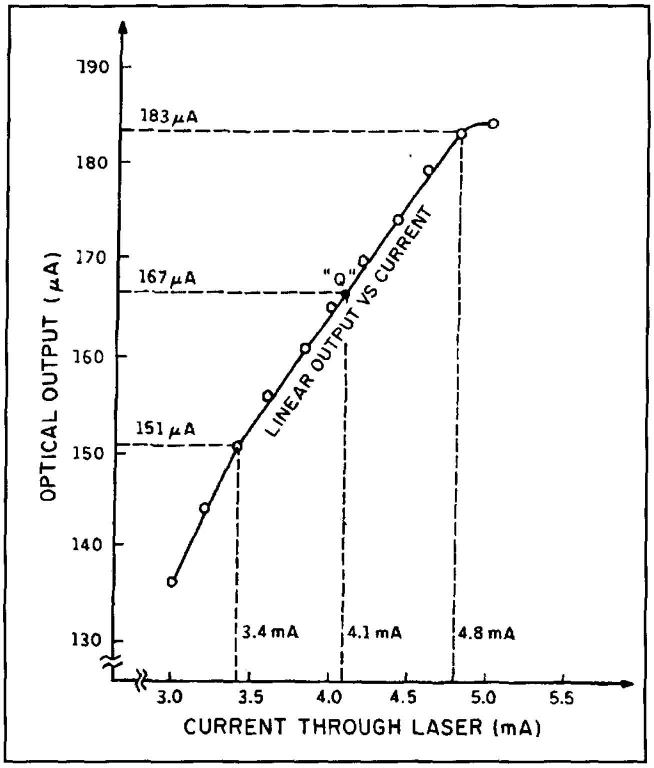

Figure 5. Optical output versus laser current.

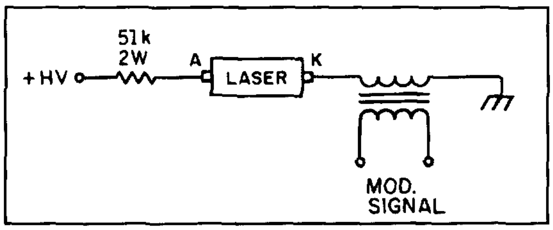

Figure 6. Laser current modulation through transformer.

beam. Because atmospheric turbulence can become bothersome, any low-frequency baseband intensity modulation will become corrupted over long distances. To overcome the this limitation, the information must be frequency modulated onto a subcarrier that intensity modulates the optical carrier. HeNe lasers can be subcarrier modulated up to 200 kHz in this manner. To obtain a more desirable current modulator for subcarrier modulation, use a transistor as a variable current source to vary the current through the laser (see **Figure 7**). Here the quiescent operating current through the laser is set at 4.1 mA by a potentiometer in the base bias circuit. Assuming a 0.7 volts DC drop across the transistor base-emitter junction and a 1000-ohm resistor in the emitter lead, the quiescent bias voltage from base to ground is 4.8 volts DC. The input voltage must vary 1.4 volts p-p to cause a ±0.7 mA current variation through the laser. To effect this, the modulation

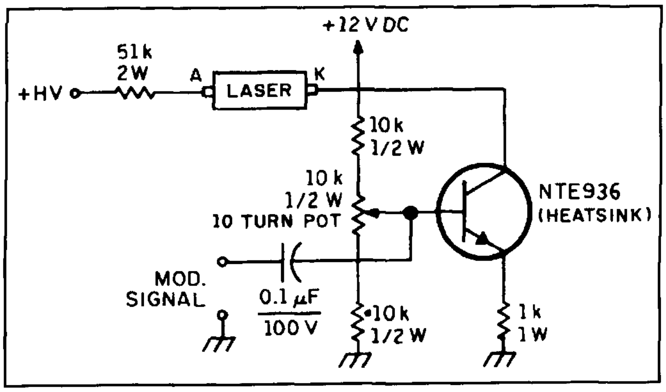

Figure 7. Laser current modulation through current source.

signal source must provide up to a 1.4 volts p-p signal into the laser current modulator circuit.

The HeNe laser transmitter

Figure 8 illustrates how the entire laser transmitter comes together. The current modulator shown in **Figure 7** is connected to the cathode of the laser. The quiescent current is set at 4.1 mA, and the modulating signal is set to vary the current only by ±0.7 mA. A well-protected and bypassed 0 to 10 mA meter is used in the ground leg to set and monitor the laser's quiescent operating current. A 51-k, 2-watt ballast resistor close to the anode lead of the laser stabilizes the discharge. The voltage across the sustaining supply should be 1.45 kV DC at full load. I'll leave the design of the FM subcarrier modulator to the individual experimenter. Just remember that the subcarrier frequency must be no greater than 200 kHz, and that the peak-to-peak signal voltage into the modulator must not exceed 1.4 volts p-p. (There are any number of ICs that operate as voltage controlled oscillators to produce the FM modulated subcarrier.)

The optical receiver

In this section, I'll describe a direct-detection optical receiver. It consists of a lens system acting as an antenna focusing the optical signal through an optical bandpass filter onto a photodetector. The photodetector converts the information on the carrier down to the baseband for further signal processing. A typical optical receiver block diagram is shown in **Figure 9**. There are a number of techniques

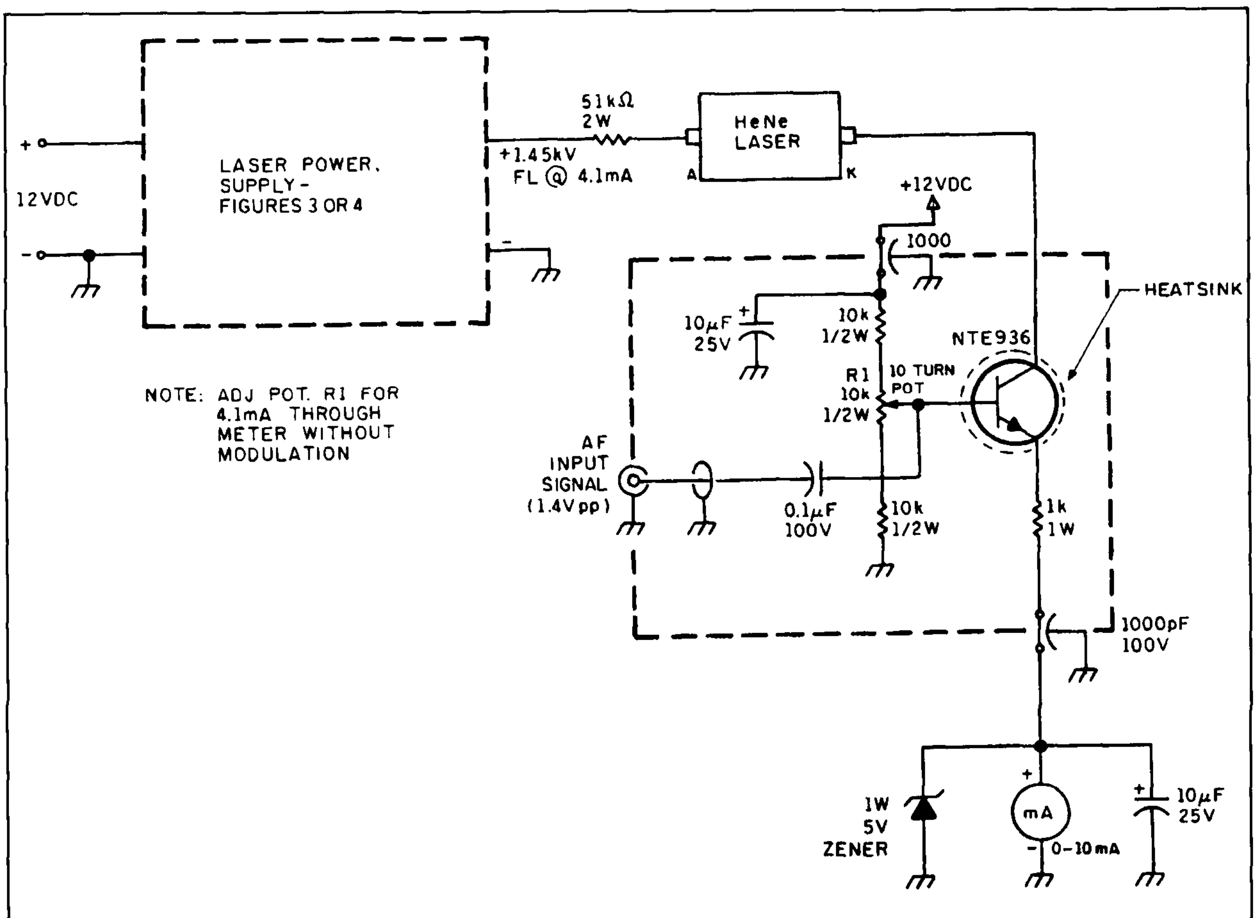

Figure 8. Laser transmitter.

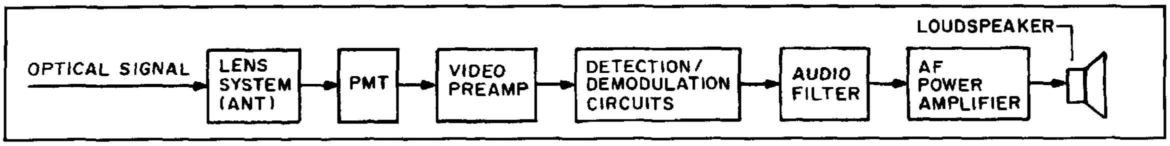

Figure 9. Optical receiver block diagram.

that must be implemented to ensure success of the receiver. Because optical detectors operate on the principle of area detection, there is no discrimination between the optical signal from the desired source and ambient light from many other sources. To aid in the rejection of undesired optical inputs, the following techniques should be used: (1) baffling (such as a tube) around the lens antenna to limit the field of view, (2) allowing only the light coming through the lens to enter the photodetector, and (3) using a narrow interference filter tuned to the laser output. This allows the receiving system to be used both day and night.

Lens antenna and input filter

There are a great variety of lens systems available from simple lenses to surplus camera lens combinations. The latter can be purchased through parts houses or at flea markets. I prefer the camera lens systems because they are easier to mount mechanically. Many also have an adjustable iris and adjustable focus. They also can be made to fit inside black paint covered PVC tubing forming an optical baffle that limits the antenna's field of view. The output side of the lens system can be mounted to an enclosure surrounding the optical filter and photodetector, making a light-proof housing as shown in **Figure 10**. Make sure the lens system focuses the source image in such a way that it fills the entire photodetector surface as illustrated in **Figure 10**. Construction detail of the optical receiver warrants careful attention, as it is part of the critical alignment for the receiver and laser transmitter.

Photodetector

There are a wide variety of photodetectors available. Many of the solid-state type peak their spectral response in the near infrared wavelengths. A number of the vacuum phototubes and photomultiplier tubes (PMT) peak their spectral response in the green/blue portion of the spectrum. The photomultiplier has good high-frequency response to a modulated signal, an acceptable spectral response to 632.8 nanometer laser radiation, and very high gain—typically approaching 1 million for a 931 type PMT. On the negative side, PMT usage requires a regulated 1000 volts DC power supply that provides 10 mA to the PMT and its resistor divider network. However, the advantages outweigh the disadvantage involved with the regulated HV supply.

Figure 11 shows the circuit diagram of a regulated 1000 volts DC power supply driven from a 12 volts DC source. A low-voltage secondary

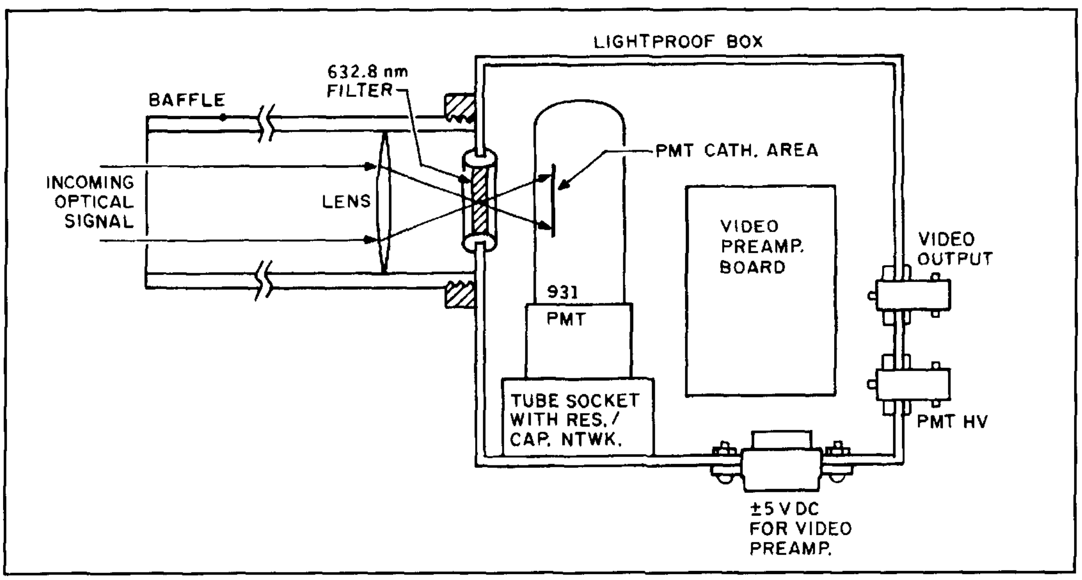

Figure 10. Optical receiver front end.

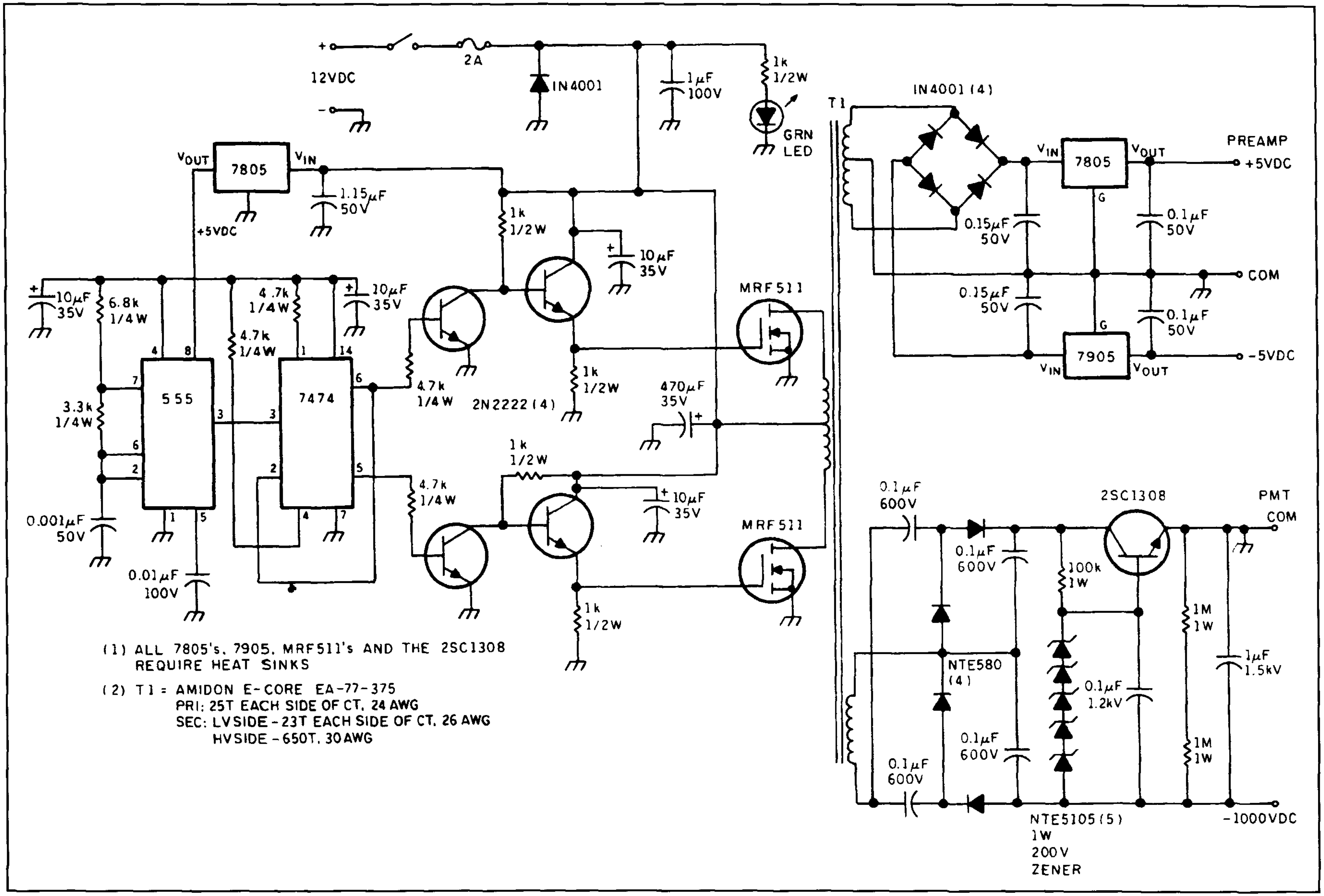

Figure 11. Regulated PMT/preamp power converter.

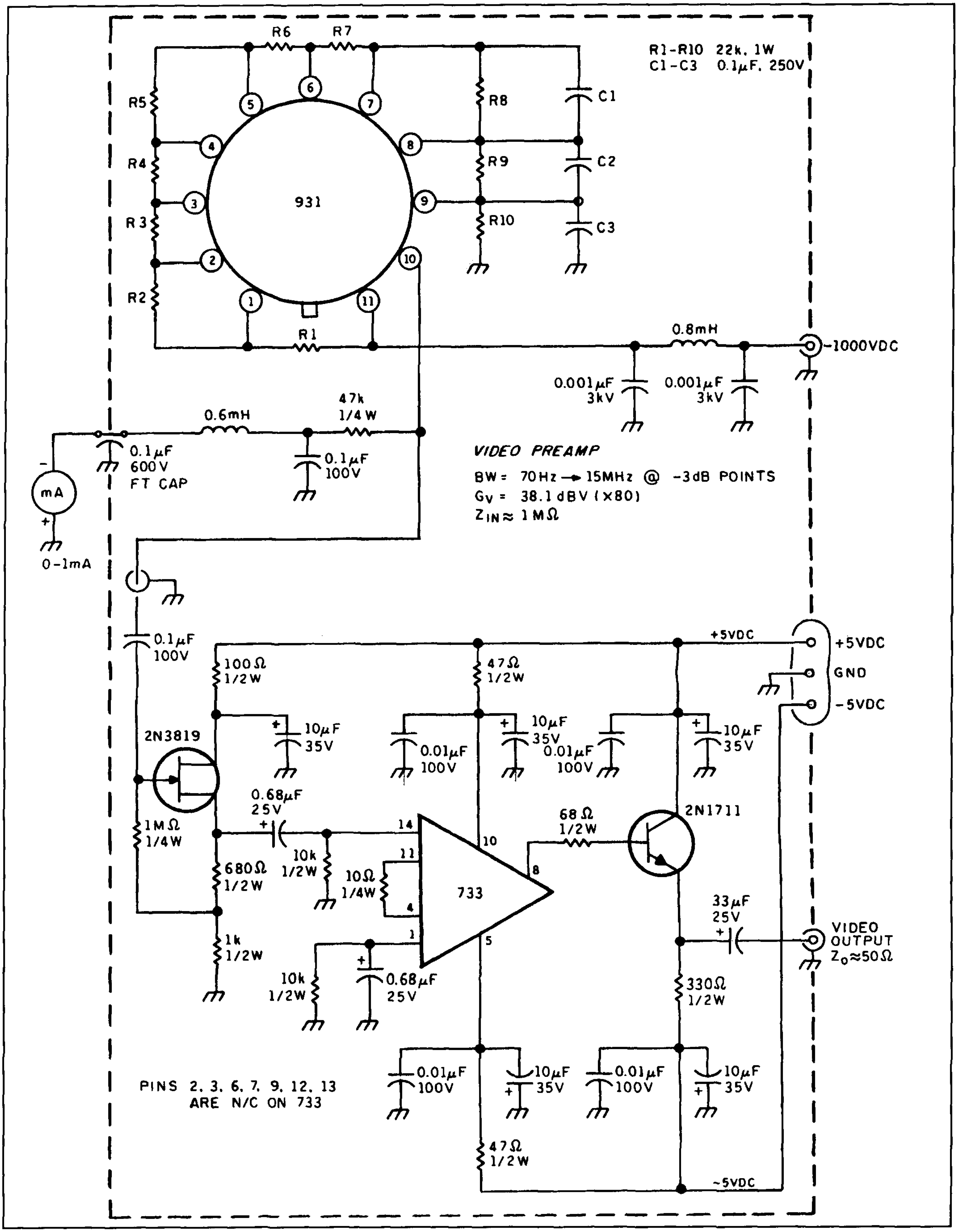

Figure 12. PMT and video preamp circuits.

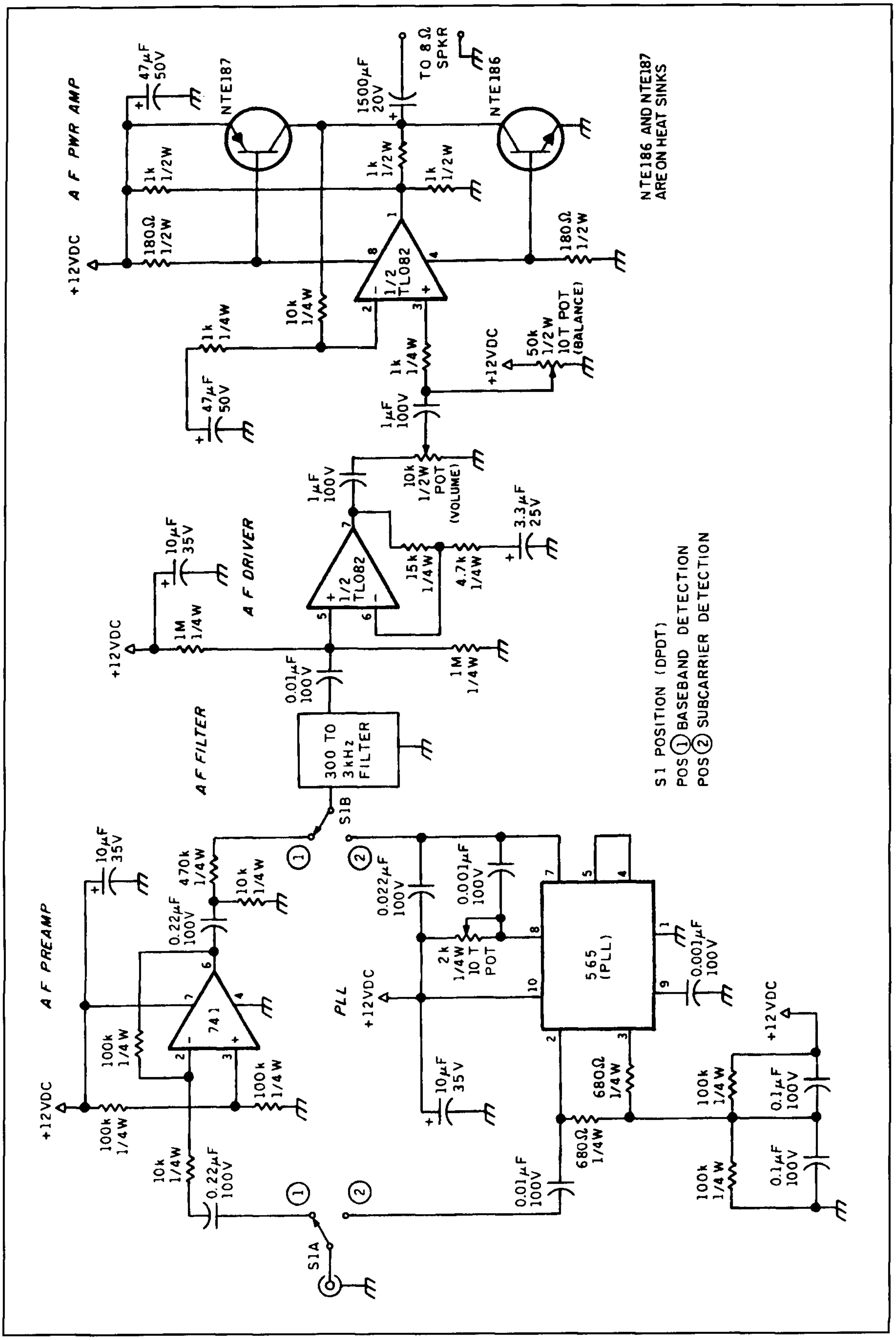

Figure 13. Receiver detection and audio amplifier circuits.

regulated power source supplying power to the PMT video preamplifier is included. Again, follow the caution regarding PMT power supplies.

All precautions must be observed when working with the pmt high-voltage supply.

Figure 12 is a circuit diagram for the inexpensive 931 side-looking PMT and its resistor divider network. This circuit maintains the proper voltages for the dynodes. The capacitors across the last three network resistors improve the frequency response of the PMT to modulated signals. A milliammeter in the PMT anode circuit registers the average PMT current under medium to high illumination encountered during diagnostic tests. A PMT preamplifier uses a 733 (or equivalent) wide-band amplifier. Because the demodulated signal can be either a direct baseband signal or a baseband signal modulated onto a subcarrier, the preamplifier must be capable of amplifying all the demodulated signal frequencies. The preamplifier is designed for a 70 Hz to 15 MHz bandwidth at the –3dB points and for a gain of 80. Because the signal detection circuits will probably be placed in a separate housing, the video preamplifier is designed to drive a 50-ohm cable. With the high current multiplication inherent in the PMT, the PMT noise will predominate over the video preamplifier IC noise.

Signal detection/audio power amplifier circuits

The video preamplifier output can be switched between an audio amplifier that establishes the correct level of the demodulated baseband audio when used for short distance test links or a phase-locked loop (PLL) that demodulates an FM modulated subcarrier when used for longer distances. The circuits are shown in **Figure 13**. The PLL VCO is set to a 100 kHz subcarrier frequency. The error signal generated as the VCO tracks the FM signal produces the demodulated audio. Either of these outputs are sent to a 300 to 3000 Hz bandpass filter and on to an audio power amplifier capable of driving low impedance headphones or speaker. The DC power supply for these circuits is provided by a filtered 12-volts DC input. The heart of the receiver can be put together by combining the power supplies for the PMT and preamplifier of **Figure 11**, the PMT and preamplifier circuits of **Figure 12**, and the signal detection and audio power amplifier circuits of **Figure 13**. I leave the 12 volts DC power supply and the antenna and optical filter to you. **Figure 10** will give an idea of the mechanical layout.

Conclusion

This has been a brief glimpse into what is possible when constructing equipment to set up an optical link. The receiver and transmitter described can be used separately or in a transceiver operation; I leave that up to you. The HeNe laser and photomultiplier tube are just one possible optical link combination. Visible red semiconductor diode lasers are dropping in price and offer attractive alternatives to HeNe lasers, because they are more efficient and don't require high-voltage supplies. They can also be intensity modulated at frequencies in the MHz range to nearly 100 percent. Avalanche PIN diodes may be used as optical detectors, although their current multiplication isn't as high as the PMT and their spectral response peaks into the near infrared. Aside from these changes, the modulator and demodulation/signal conditioning circuits presented here are common. ■

PRODUCT INFORMATION

BASIC Stamp II Programmable Module

Parallax, Inc. has released The Stamp II. The Stamp II is a complete BASIC-programmable computer in a 24-pin DIP package. It has 16 I/O lines, 2K of non-volatile memory, and a clock speed of 20 MHz.

The Stamp II's I/O lines are used to connect the Stamp to the outside world. Most I/O functions are digital, and include serial communications, pulse measurement, button input, transition counting, etc. A few functions are pseudo-analog, such as resistance measurement and PWM. The Stamp II even has functions for transmitting X-10 powerline control signals, as well as for generating accurate audio frequencies (including DTMF tones for telephone dialing).

An 8-pin EEPROM provides 2K of non-volatile memory. This memory is used for both program and data storage. Each BASIC instruction takes 3–4 bytes of space, so the memory can store about 600 program instructions.

Stamp II modules are available for $49 in single quantities. For more information, please call Parallax at 916-624-8333; fax 916-624-8003; or e-mail info@parallaxinc.com.

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Madison, Wisconsin 53711-4738

THE SOLAR SPECTRUM

Ulysses verifies the shape of the interplanetary magnetic field

After a journey of more than 3 million, million kilometers, the Ulysses spacecraft has now completed the first observational flight over the Sun's polar regions. This extraordinarily successful mission has provided astronomers and other researchers with a vast store of information about the forces and phenomena at work in previously unexplored high-latitude regions. Over the course of its five-year journey, the NASA/European Space Agency-directed Ulysses project has confirmed a number of long-held theories, and uncovered a few surprises. For example:

- Ulysses verified global differences in solar wind velocity (finding a polar wind speed that is nearly double that at low latitudes), composition and temperature.
- The spacecraft observed outward-propagating, high-speed, long-period Alfven waves to be continuously present in the solar wind emanating from the Sun's polar regions. Such waves move along magnetic field lines and accelerate charged particles.
- While it has long been postulated that the intensity of the solar magnetic field above the poles would increase substantially, Ulysses found a uniform magnetic field with an intensity that did not change from equator to pole.

One of the more intriguing findings, however, came from a clever radio-based experiment designed to obtain information about the shape of the solar interplanetary magnetic field.

Energetic solar phenomena such as solar flares inject huge numbers of highly energetic, high-temperature, electrons into the interplanetary medium. As they propagate outward, these particles follow the Sun's extended magnetic field structure, interacting with the slower solar wind plasma and generating radio emissions along the way. Since these emissions are in part based on solar wind density—which decreases as distance from the Sun increases—the signals vary in frequency from several hundred MHz near the Sun to about 50 kHz near the orbit of the Earth. The radio manifestation of such interactions can be observed as a moving Type III radio burst, which we briefly described in the Fall 1995 issue of *Communications Quarterly*.

The instrumentation carried aboard Ulysses includes a sensitive 76-channel radio receiver coupled to both dipole and monopole antennas; a configuration which allows frequencies between 1 and 940 kHz to be observed. Furthermore, the experiment was designed to determine the direction of arrival of the radiation from modulation measurements. Thus, observations permit tracking of the bursts through interplanetary space. The speed of the exciter electrons—typically between 0.1 and 0.3 the speed of light—causes the radio emission to occur at progressively later times. It takes about 20 minutes for the particles to travel from Sun to Earth.

This portion of Ulysses instrumentation was developed in a cooperative effort with experts at the Paris Observatory in France, the University of Minnesota, and the NASA Goddard Space Flight Center, in Maryland. Even though solar activity is rapidly moving towards an expected minimum sometime in 1996, numerous Type III bursts have been detected by this device during the course of the Ulysses mission.

The shape of the solar interplanetary magnetic field has long been thought to be in the form of an "Archimedean Spiral" (**Figure 1**), a distinctive pattern of field lines that results from the rotation of the Sun. For an observer on the Earth, the solar coronal plasma appears to expand outward radially, while the field lines form spirals with footprints anchored to the Sun itself. (Deviations from the precise spiral path

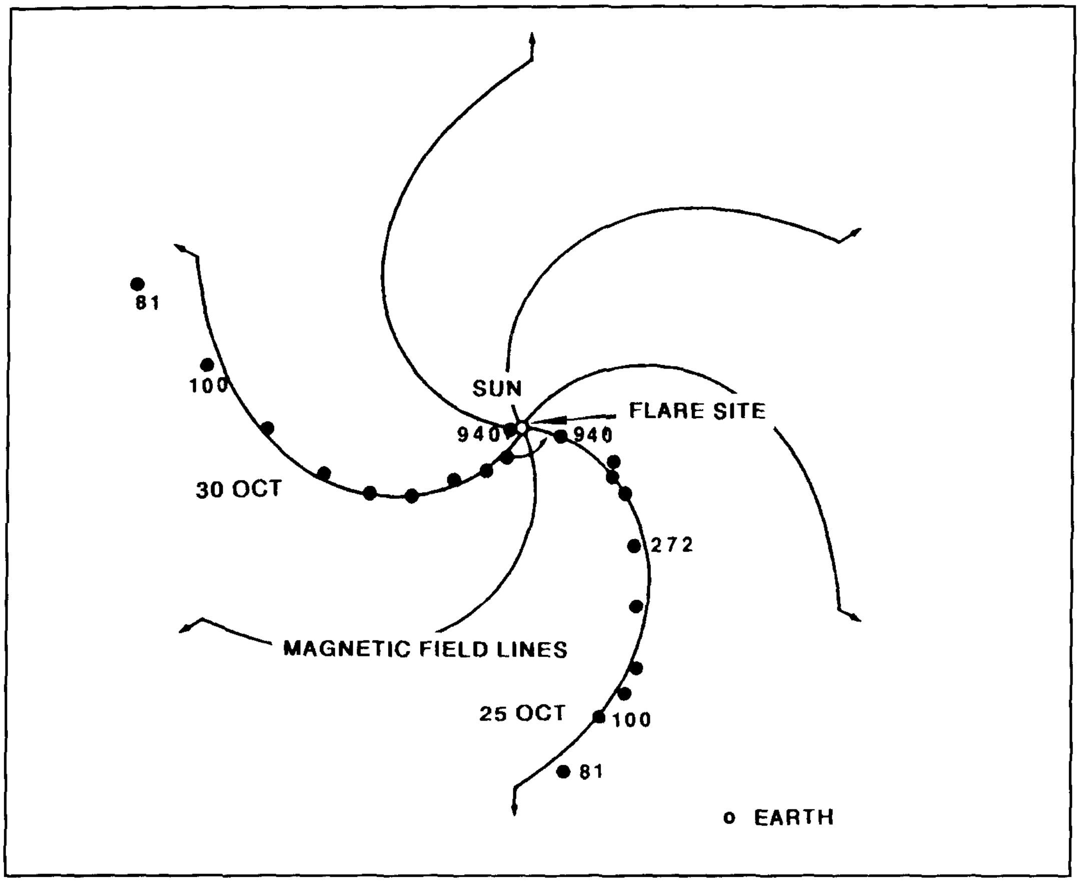

Figure 1. *Looking at the ecliptic plane.* This representation shows the measured trajectories of two moving Type III radio bursts observed by the Ulysses spacecraft on October 25 and 30 1994. Both trace out Archimedean Spiral-like paths (magnetic field lines) through interplanetary space, from near the Sun to near the orbit of the Earth. Numbers refer to frequency in kilohertz. The October 25 flare site is also shown. (Diagram prepared from a NASA illustration.)

result primarily from kinks spawned by variations in solar wind speed.)

As Ulysses passed over the Sun's southern polar regions on 25 October 1994, a medium-intensity, but long-duration solar flare accompanied by material ejection and radio emission erupted in NOAA/USAF Region 7792—a rather unspectacular sunspot group in the Southern Hemisphere.

Because the entire radio emission trajectory that resulted from this activity was visible to the high-flying Ulysses spacecraft, the entire path of radio bright spots could be measured in less than one hour. A second series of radio burst observations, this time without an obvious source, was undertaken five days later. Thus, each of these unique observational runs provided near instantaneous views of the spiral form of the interplanetary field.

Dr. Michael Reiner, chief scientist at Hughes STX in Lanham, Maryland, explains, "The aerial view of the interplanetary magnetic field became possible with the flight of Ulysses over the South Pole of the Sun in 1994. Now we could look down on the solar system, and these radio observations gave us the first direct observation of the spiral structure in space between the Sun and Earth. The radio emissions, caused by fast electrons moving through with the slower solar wind, allow us to trace out the magnetic lines of force much like you might deduce the course of a road at night from an airplane by tracking the headlights of moving cars."

With its northern pass completed, Ulysses has begun the long voyage back out to the orbit of Jupiter, reaching the giant planet's distance of about 800 million kilometers during April 1998. Once there, Ulysses will again loop back and return to the Sun, arriving in its vicinity in the year 2000; a time when solar activity is expected to be very high and the solar magnetic field will have reversed magnetic polarity. According to Dr. Edward J. Smith, Ulysses project scientist at the California Jet Propulsion Laboratory where the U.S. portion of the mission is managed, "We expect the profile we obtain five years from now will be dramatically different and give us many new insights into the dynamics of this star at the center of our solar system.

WIND spacecraft encounters mass ejection from the sun

In a separate space-related development nearly one year after Ulysses observations verified the suspected shape of the solar magnetic field, a second spacecraft has detected a huge interplanetary disturbance that struck the Earth's protective magnetic field on October 18, 1995. The collision produced a magnetic storm and auroral display that persisted for two days and was visible in the United States as far south as Denver, Colorado.

The information was relayed by NASA's Goddard Space Flight Center in Greenbelt, Maryland, to the U.S. Air Force and to the National Oceanic and Atmospheric Administration's Space Environment Laboratory, in Boulder, Colorado, where evaluators issued alerts to commercial satellite operators, electrical utilities, and other organizations throughout the world that were likely to be affected.

By the time it was detected by NASA's WIND spacecraft, the disturbance—a giant cloud of material ejected from the outer atmosphere of the Sun (corona)—had grown to a diameter of some 104 million kilometers across and was rushing towards the Earth at 3.4 million kilometers per hour.

WIND is an unmanned spacecraft pointed towards the Sun that patrols interplanetary space over one million kilometers from the Earth. Invisible to normal telescopes and to the human eye, the cloud was composed of magnetic fields and electrified subatomic particles ejected from the solar corona.

About thirty minutes after the front edge of the cloud passed over the WIND probe, it swept over Japan's GEOTAIL satellite, which was located on the sunward side of the Earth in its 193,000 x 64,000 kilometer elliptical orbit. GEOTAIL also gathered important scientific data on this phenomenon. Minutes later, the cloud of ejected material struck the outer limits of the Earth's magnetic field, which acts as a protective buffer. The impact compressed the magnetic field on the sunward side of the Earth and stretched it out and away from the Sun on the night-side, triggering the magnetic storm and aurora.

WIND was launched on November 1, 1994, and is the first of two NASA spacecraft in the Global Geospace Science initiative. Together with GEOTAIL, Polar, SOHO, and Cluster, WIND is part of the *International Solar Terrestrial Physics* (ISTP) Project. The main purpose of the WIND spacecraft is to measure the incoming solar wind, magnetic fields, and particles; however it also observes Earth's foreshock region. Another of the principle science objectives of the WIND mission is to provide baseline ecliptic plane observations to be used in heliospheric latitudes from Ulysses.

A complete analysis of the October WIND data—and data from other spacecraft and instruments—may take months or years to complete, but the information is expected to tell scientists much about how interplanetary disturbances propagate through space and affect the Earth's environment. The rate at which such disturbances occur is expected to rise sharply as the new sunspot cycle, expected to commence in 1996, peaks sometime around the year 2000.

Information for this article was obtained from a paper published in the journal *Science* (M.J. Reiner, J. Fainberg, and R.G. Stone, **Volume 270**, 20 October 1995), and from material furnished by the *National Aeronautics and Space Administration*. Michael Boschat, Nova Scotia, Canada, provided the NASA diagram from which **Figure 1** was prepared. ■

PRODUCT INFORMATION

New Digital Data Buffer From Harris Corp.

Harris RF Communications now has a new digital data buffer which uses multiple error-correction techniques to provide data transmission via VHF/UHF radio. The RF-3490C is a waterproof unit designed to operate with AN/PRC-117A, AN/VRC-94A(V) and other VHF radios directly, or through digital encryption devices using MIL-STD-188(C) interface ports. The unit interfaces to terminals via synchronous or asynchronous formats.

The data buffer provides front-panel selectable data rates and modes, including 1200 or 2400 bps synchronous or asynchronous, 50–19,200 bps synchronous or asynchronous, and data throughput rates from 820 to 16,000 bps. The unit also provides automatic switch over to voice mode, and automatic or manual keying operation.

For further information, contact Harris Corporation, RF Communications Group, 1680 University Avenue, Rochester, NY 14610; phone 716-244-5830.

Wade G. Holcomb, *W1GHU*
185 Linden Street
New Haven, Connecticut 06511-2407

TRY NMR WITH YOUR OLD CW RIG

Using amateur radio equipment to perform nuclear magnetic resonance experiments

Want to try something new and different with your old CW rig? Consider building your own experimental nuclear magnetic resonance (NMR) instrument. With it, you can experience the thrill of sending and receiving radio signals to the protons of hydrogen atoms. As a matter of fact, it's entirely possible to duplicate discoveries made shortly after World War II with that old CW rig of yours, plus a surplus magnet similar to those that formed part of a radar magnetron. Of course, some readjustment will be necessary to get your old rig tuned to the correct frequency. You'll also need an oscilloscope and an automatic keying circuit.

For those who enjoy construction and troubleshooting, this experiment could be the basis of a science fair project using dated ham rig components. Special interests in RF circuits or computer software are very useful in building

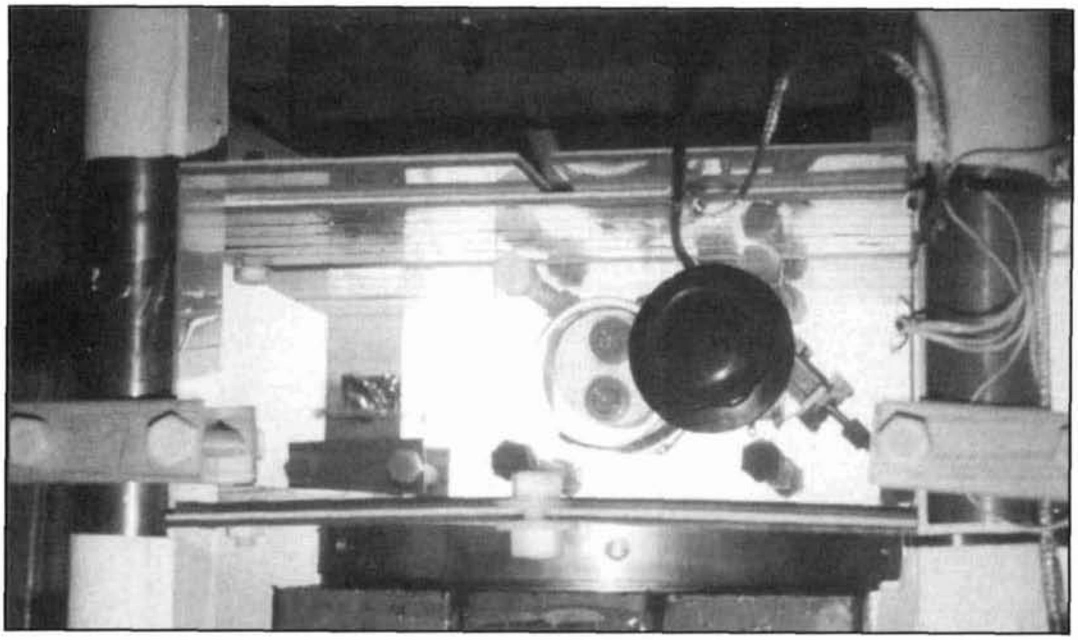

Photo A. Magnet with RF tank coil with two tubes of salad oil. Four steel support columns also serve as the return magnetic field circuit. The field is about 731 Gauss.

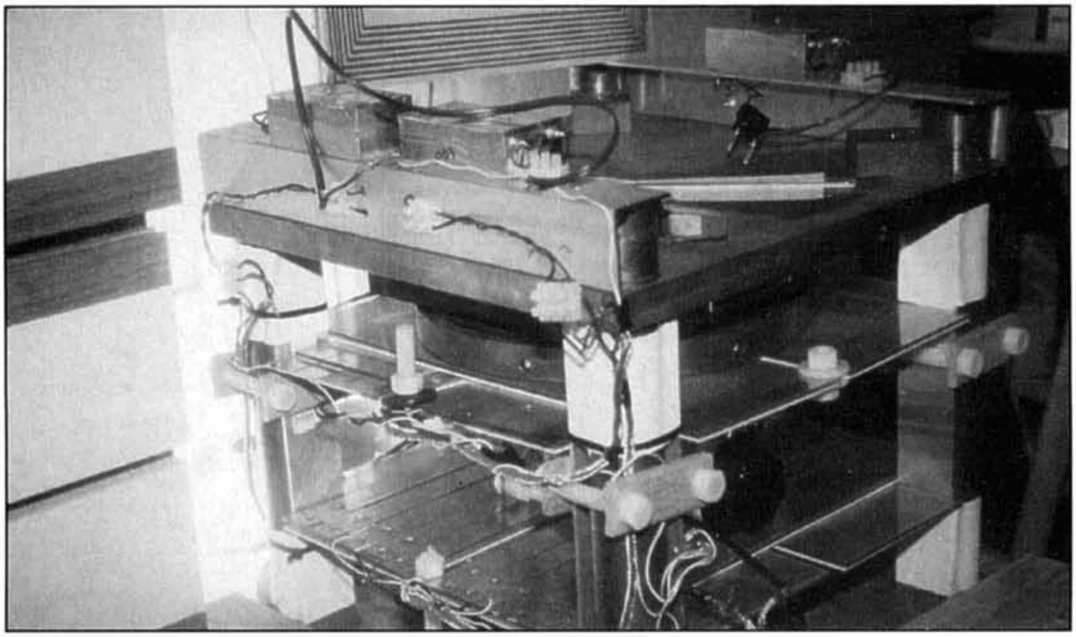

Photo B. The four-poster magnet is 18 inches on each side. A bottle of salad oil is inserted inside a 3.11 tank circuit. Credit cards can be erased if one is not careful.

your own amateur NMR system. **Figure 1** shows a functional block diagram of the major components required to perform amateur NMR.

What is nuclear magnetic resonance?

The hydrogen atom contains one proton at its center. Nuclear magnetic resonance (NMR) and magnetic resonance imaging (MRI) techniques make use of two magnetic fields—a fixed field and a variable radio frequency (RF) field—in a manner that lets an observer make physical measurements based on the proton's reaction to these fields. This method allows one to study the properties of many common substances using components familiar to radio amateurs.

While information on NMR is mostly accessible to those with training in one of the physical sciences, **Reference 1** offers detailed explanations of the fundamentals of NMR using a descriptive, mostly nonmathematical approach. The rapid development of medical MRI systems required that a trained support force be available. This book is often used by institutions to teach support personnel, and is one of several books written to fill this need.

Many atomic nuclei have "spin" and charge. Spin is the atomic equivalent of angular momentum in everyday life. According to quantum theory, a nucleus with spin can only take certain energy levels in a magnetic field. We can visualize the nucleus spinning like a bar magnetic on its axis, producing an associated magnetic field. It is the interaction of this field with external fields that separates nuclear energy levels and allows NMR to occur. The magnetic moment (current times enclosed area) is sometimes called a nuclear magneton. The hydrogen atom has a 2.79 nuclear magneton value.

A small bottle of salad oil contains a large number of possible radio signal sources (about 6 x 10E+22 per cubic milliliter). **Photo A** shows two tubes of salad oil inside a tank circuit between the poles of my magnet. In my magnetic field, only about one atom per million atoms is a potential contributor, on a chance basis, to a detectable RF signal following an RF pulse. A huge number of such atoms results in a detectable signal. The strength of the detected signal can be as much as 5 µV.

The duration of the RF keying pulse and its power level must be determined by experimentation to find the correct amount of energy to "flip" protons. Best results are obtained when the flip is 90 degrees from the static field. For instance, it's possible to have too great a pulse duration or power level, which might result in flipping the protons 450 degrees, a complete circle plus 90 degrees. The detectable signal would be similar to the correct amount!

Finding a magnet

Magnets are still available from surplus catalogs. When choosing a magnet, remember that the RF signal frequency's purity is a function of the field's uniformity. The magnet's uniformity is equal in importance to its field strength in procuring good results. Obtaining a uniform field is a never-ending goal for NMR and MRI

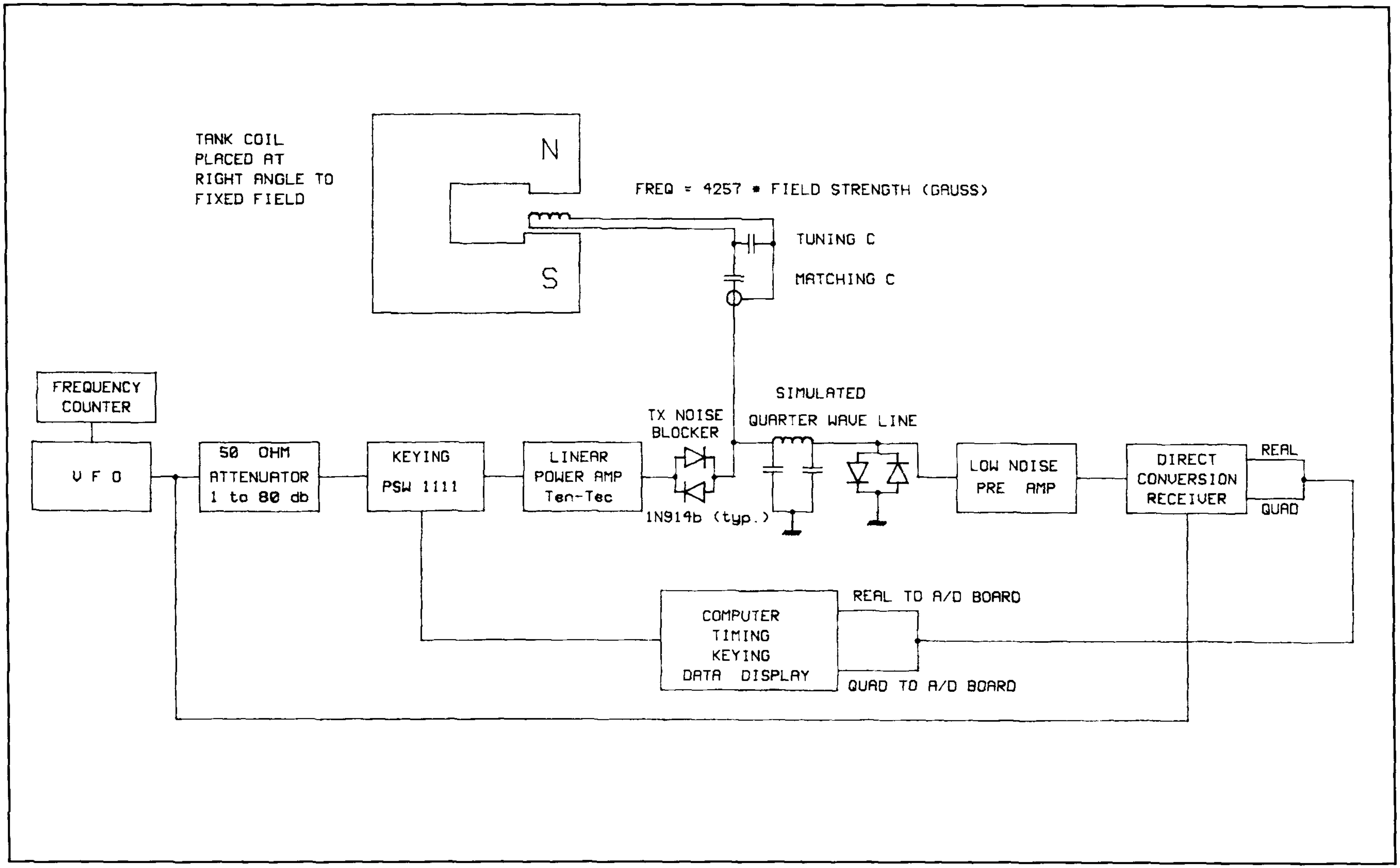

Figure 1. Functional block amateur NMR system.

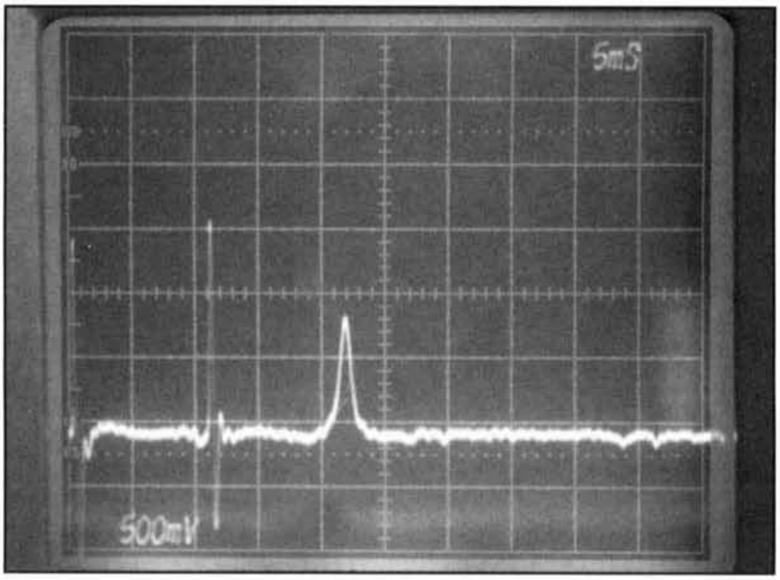

Photo C. A dot/dash RF pulse to the tank coil holding an oil sample in a magnetic field sends back an RF Hahn echo. This is one of the first subjects a new NMR student finds out about (see references).

workers. A tolerance of 5 to 10 parts per million over a volume the size of a golf ball would make a very useful amateur magnet. A change of 1 gauss will mean a change of 4257 Hz in the observed frequency. Moving a metal chair near the magnet can distort the magnetic field and detune your system.

It's even possible to make tests using the Earth's magnetic field at a frequency about 2000 Hz, using audio in place of RF equipment. Perform these tests in your backyard, away from cars or other large metal objects. Several papers appeared during the 1950s showing excellent results in measuring small variations in the Earth's magnetic field.[2]

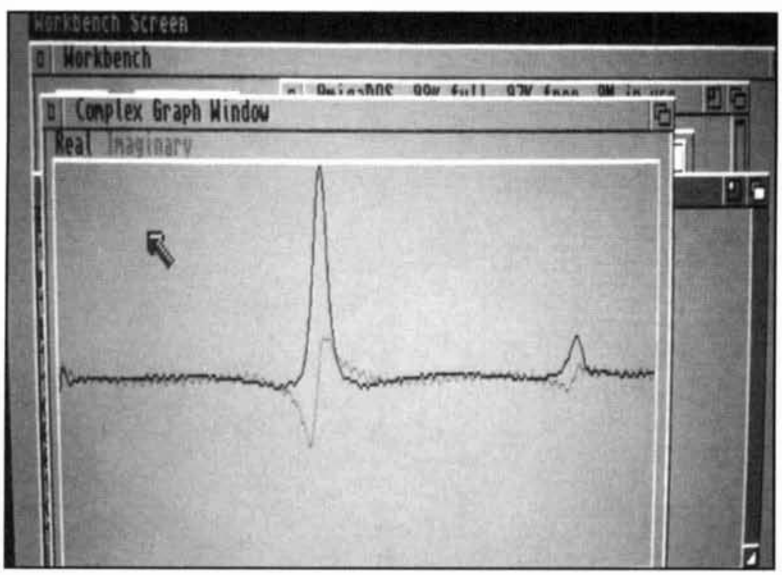

Photo D. Amiga screen shows the real and quadrature of the Hahn echo held RAM memory, this display is the average of 16 echoes. A dual A/D converter board suitable for stereo music will do this nicely.

Simple NMR experiments

The vertical field strength of my 500 pound magnet (see **Photo B**) is about 731 gauss, approximately 1400 times the Earth's magnetic field at my QTH. This magnet is quite temperature sensitive, almost 1 gauss/degree C. I usually have to readjust my master oscillator to find the hydrogen proton frequency if the room temperature changes. My magnet's field strength increases in cold weather.

Once I find the resonant proton frequency, I measure it within one cycle using a frequency counter. This frequency allows a very accurate method of determining the magnetic field strength. The relationship of frequency to magnetic field strength is given by Larmor's constant:

f—magnetic field in gauss x 4257

In my magnet, the NMR frequency is 3.11 MHz, for a field strength of 0.0731T, (The ST unit of Tesla, T, equals 10,000 gauss.) This is near the amateur 80-meter CW band.

My RF tank circuit looks like an 80-meter final tank coil (see **Photo B**). It's driven by short duration RF pulses at 3.11 MHz. When the RF field is applied, the protons spinning in the plane of the static field rotate out of the plane of the field. When the RF field is turned off, the protons return to the plane of the static field, with two degrees of rotational freedom.

The protons' spins, after the RF pulse is turned off, go through a spiral trajectory—like an orange being peeled from one end to the other—emitting a weak RF signal into the resonant tuned tank circuit. The detected RF signal takes the form of a damped sine wave. This damped wave is called a free induction decay (FID), which can last several seconds in a very uniform field, or perhaps only a few milliseconds in a non-uniform field. I sometimes judge the best spot in my magnet by positioning my sample for the longest FID.

This recovery is described by two time constants, T1 and T2, which can be measured later if the data is stored in computer memory. These two time constants, longitudinal (T1) and transverse (T2), describe these return spins to the static field, and can indicate the effect of nearby atomic neighbors on the observed hydrogen protons. For instance in pure water, the two time constants are equal to each other, but this isn't so in salad oil or other complex compounds.

System requirements

The amateur radio requirements needed to

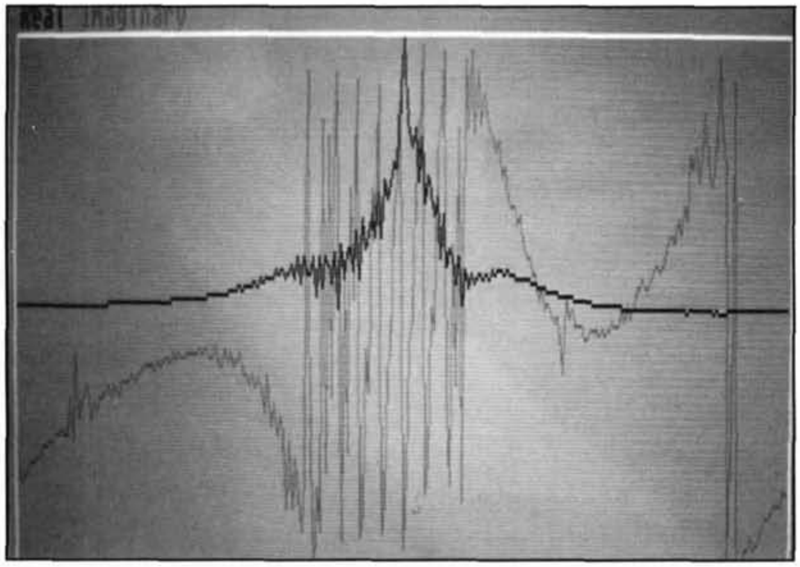

Photo E. Frequency spectrum of Hahn echo shown in *Photo D*, found by using computer software. Baseline is 10 kHz wide. Width at the 50 percent amplitude point is about 200 Hz and may be used to judge magnetic field uniformity. Phase spectrum is shown in background.

bounce an RF signal off the earth-moon-earth (EME) are equivalent to those required for listening to the proton's spin (see **Figure 1**). As you know, these are a transmitter, receiver, antenna, keyer, a low-noise receiver front end, a T/R system, and a display. The keyer in my system is a computer interface board and software. I use a direct conversion receiver.

I use a computer with a timer board to generate a dot and dash pattern to key the transmitter with the two required pulses—a 90-degree dot followed by a 180-degree dash. The dot lasts 100 μS and the dash 200 μS in a typical pattern, with a 25 mS spacing. This is repeated after a 500-mS delay. Several different timing patterns are required to determine the proton spin time constants (T1 and T2). You could try it with a hand key, but you wouldn't get the accuracy you need.

The Hahn echo,[3] in **Photos C** and **D**, appearing at 25 mS from my "dot" 90-degree pulse, is captured with a computer analog-to-digital board and stored in computer memory, much as one digitizes a note of music. Later, I use computer software to determine the frequency spectrum (**Photo E**) of the stored echo by Fast Fourier Transform (FFT). The spectrum line width helps me determine the magnetic field uniformity at the position of my sample.

History

I.I. Rabi was known to have been a radio amateur, and was photographed at the controls of this "wireless telegraph" station as a teenager, around 1912.[4] He's given credit for the general concepts of using two magnetic fields to overcome the field created by the atom's rotating electron, which shields the atomic nucleus. He was awarded an unshared Nobel prize in 1944 for this work, while doing radar development for the war effort. More Nobel awards were presented to others for carrying out advances on this method in the months following the end of World War II using circuits developed by the wartime radar laboratories.[5–7] No complete study has been published covering the scientific history of the development of NMR and MRI.

Work in progress

At present, I'm measuring time constants and doing spectrum analysis of Hahn echoes to measure field purity. This should be easy for amateurs to repeat using almost any computer. I did my first Fast Fourier Transform on an Apple II+ based on an article in *BYTE* for viewing music spectrum. This required writing a 6502 machine language FFT routine. This allowed the Apple to become my first audio spectrum display about 10 years ago. I hope to obtain my first 2-dimensional MRI picture, perhaps an image of a sectional slice through an orange, soon.

I'll have to develop computer software and gradient amplifiers to drive the gradient coils shown in **Photo A** before this is possible. Complex patterns of gradients and RF pulses are needed to acquire a 2-D image plane, which must then be "decoded" using 2-dimensional spectrum analysis. With the help of Dave Reddy, N1RBJ, I've developed computer software that will perform a double-precision 128 x 128 2-D FFT on a generic 486DX 66-MHz PC clone in about 4 seconds—much faster than the expensive array processors used for these kinds of reconstructions in the recent past.

We've tested this software by reconstructing raw data of a water-bottle phantom originally acquired on a Yale University experimental NMR system. **Photo F** shows the raw data, which looks like ripples spreading in water, and **Photo G** depicts the finished magnitude and phase images. Note that the finished images are inverted, and the air bubble at the top of the bottle with its meniscus is shown at the bottom.

Summary

If you're interested in transmitters, receivers, or computer software, you'll find the effort required to capture the radio signals emitted by the proton's spin a challenge. Everything I've done can be recreated using common amateur

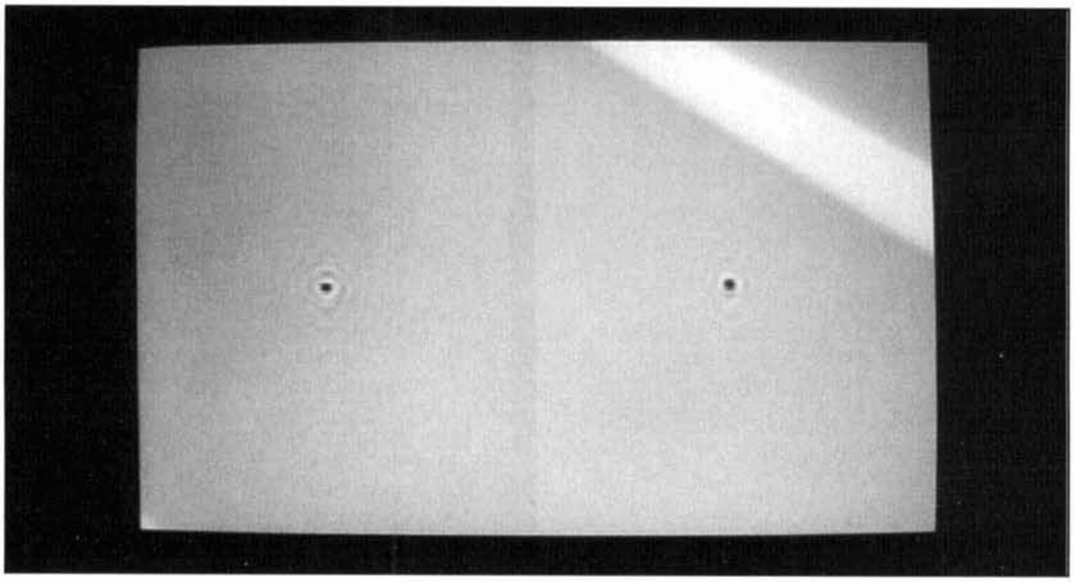

Photo F. Raster display of two 64K arrays showing RF data received from an oil sample. MRI images look like holograms before the 2-D FFT data reduction. This represents a 128 x 128 x 12 bit array.

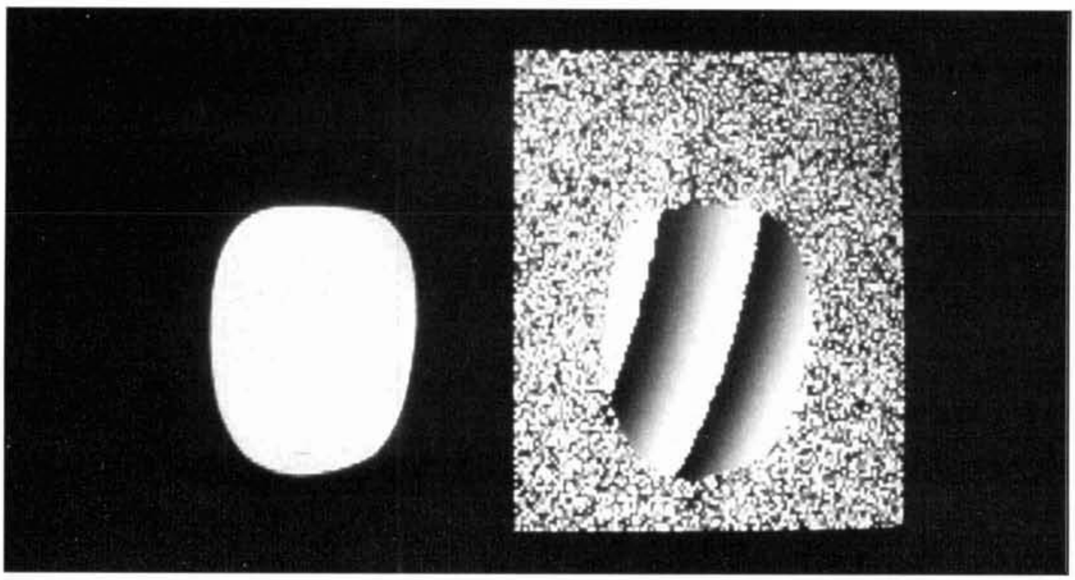

Photo G. After a 2-D FFT computer analysis (*Photo F*) shows a cross-sectional slice through the oil sample bottle. These two images now occupy the same memory space as the images in *Photo F*. Process requires 4 seconds on a 486DX 66-MHz computer.

parts, a magnet, and some patience. Amateurs with RF circuit and computer experience are well-equipped to learn about NMR. I had to learn many new terms—like Larmor's Constant, FFT, FID, T1, T2, and many others—before I was comfortable with this new field that uses RF and computer equipment to perform tasks which would have been material for science fiction stories not too many years ago. ■

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OSCILLATORS WITH LOW PHASE NOISE AND POWER CONSUMPTION

Frequency and time domain analysis and optimization. From a paper presented at RF Expo, East, 1994.

Modern hand-held radios need low phase noise oscillators, which at the same time exhibit low power consumption. Traditional choices are either bipolar transistors or FETs. Typically, FETs have been considered more favorable for frequencies up to 500 MHz. This is because the low phase noise in frequencies above the higher f_{max} of bipolar transistors compared to the N junction transistors make their usage more attractive. In this article, we will evaluate the following topics:

AGC action via diode clamping

- Performance of FET oscillators in Class A Operation, self-limiting.[1]
- Performance of Class A oscillators using a clamping diode between gate and source.[2]
- Self-limiting Class A oscillator with source resistor.[3]

Differential oscillators using FETs and different bias points

```
FET 1 2 0
G=?6.60462E-3?
CGS=?2.94488PF?
F=?4.5638GHZ?
T=?3.2199PS?
CDG=?.657116PF?
CGE=?.046388PF?
CGDE=?.117311PF?
CDE=?.186107PF?
CDS?.573044FF?
RI=?4.619?
RG=?6.39222?
RD=?2.16345?
RS?2.09905?
GDG=?11.2376E-6?
LG=?.022313NH?
LD=?3.57586NH?
LS=?2.1183NH?
```

Table 1. The equivalent linear parameters for the equivalent circuit for the FET shown in *Figure 3*.

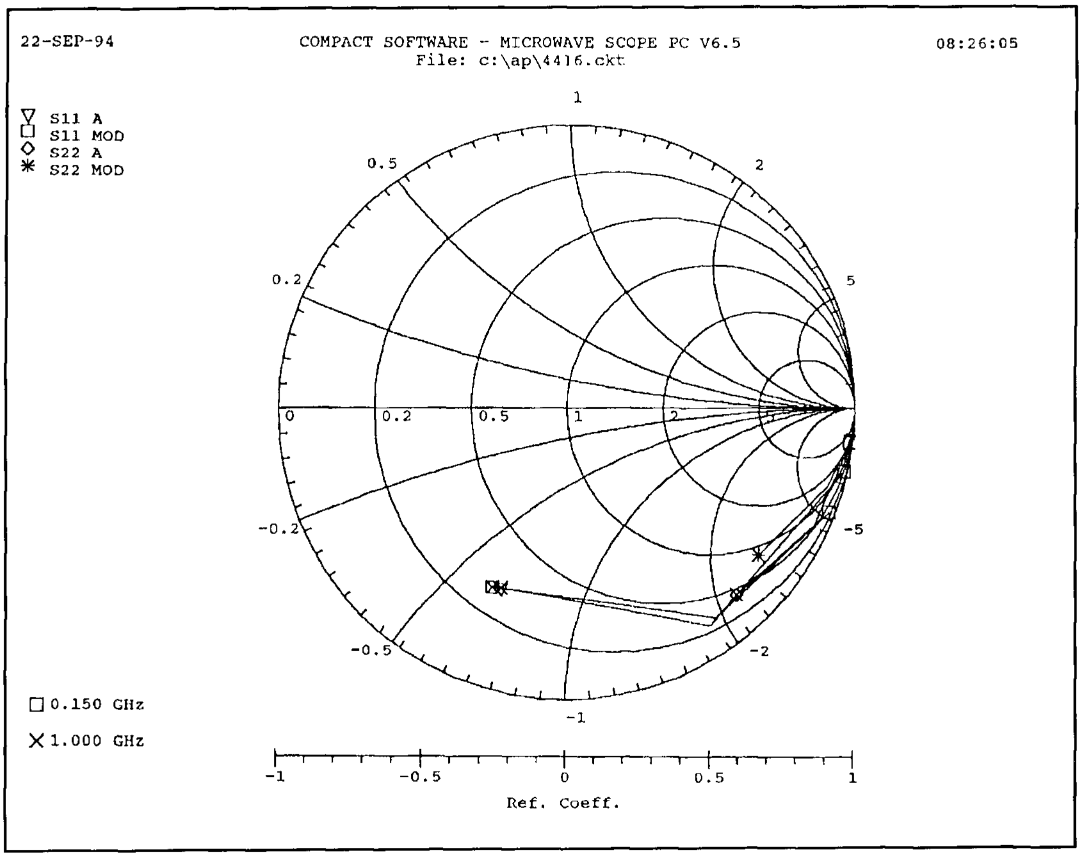

Figure 1. Comparison of measured and predicted S-parameters for the 2N4416, input and output.

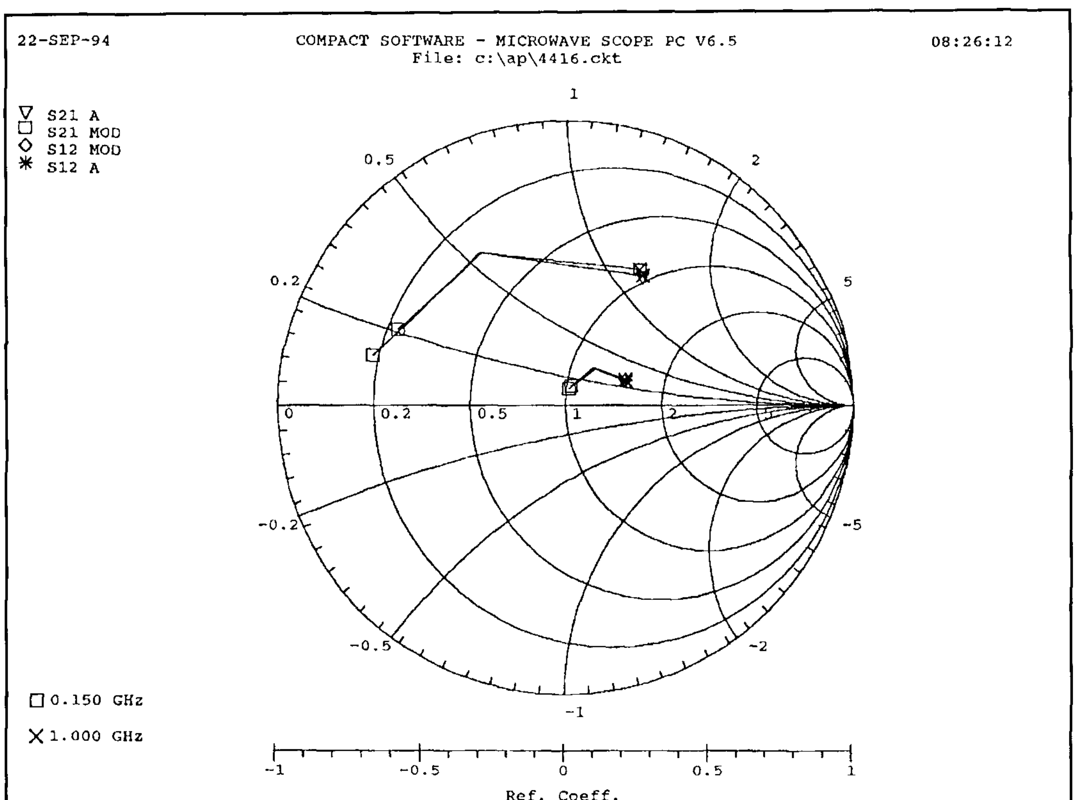

Figure 2. Comparison of measured and predicted S-parameters for the 2N4416, forward and reverse reflection coefficient.

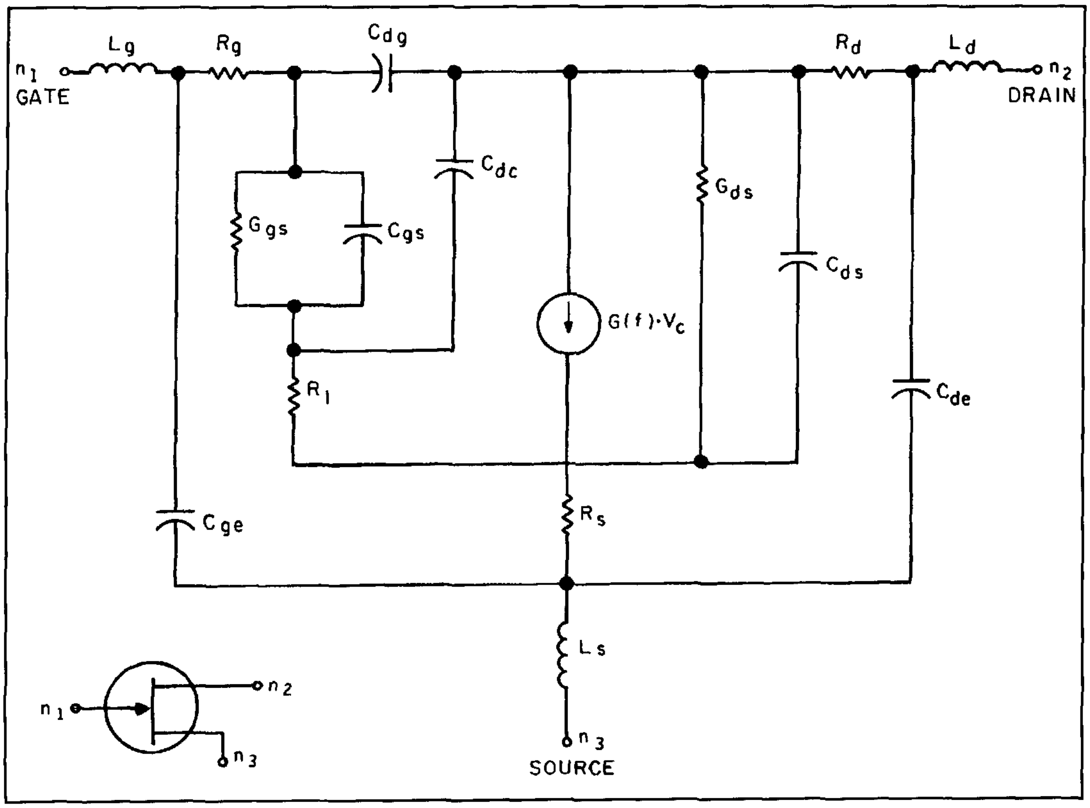

Figure 3. Equivalent circuit for the linear FET. The non-FET model follows the *SPICE* approach by Statz and Curtice.

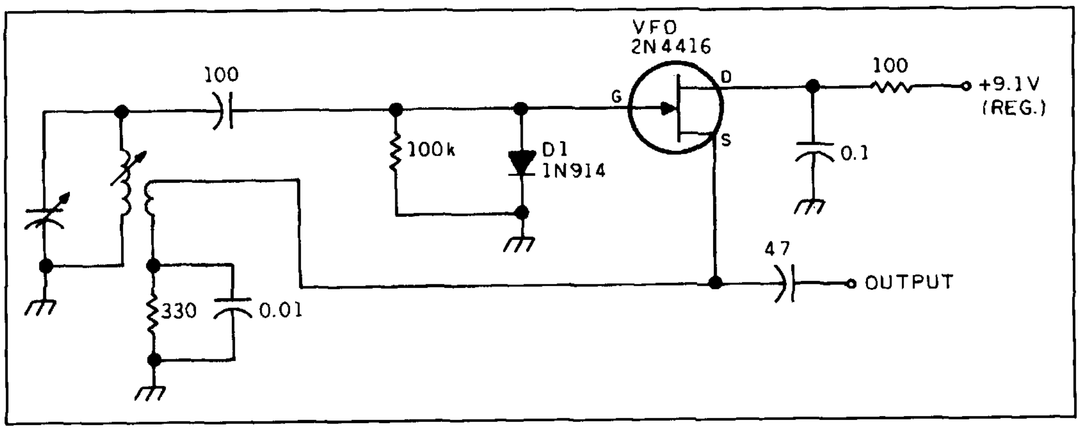

Figure 4. Oscillator test circuit.

The following are bipolar transistor oscillators:

- 800-MHz Motorola VCO for hand-held radios.[4]
- Analysis of the Motorola MC1648M voltage-controlled oscillator circuit.[5]
- BIP differential limiter VCOs from **Reference 6**.

Considerable mystery has surrounded oscillator design. Evaluating the above-mentioned circuits should shed some light in this area.

Another fascinating topic has been the whole modeling issue. To compile data, Compact Software, Inc. has been performing its own parameter extraction and has always maintained that the parameters available from *SPICE* libraries aren't sufficiently accurate to be of use in high frequency applications. In particular, the use of these transistors in oscillators requires reasonable assumptions for flicker noise and other related parameters. One of the tests is to obtain agreement between the measured and simulated values for the same transis-

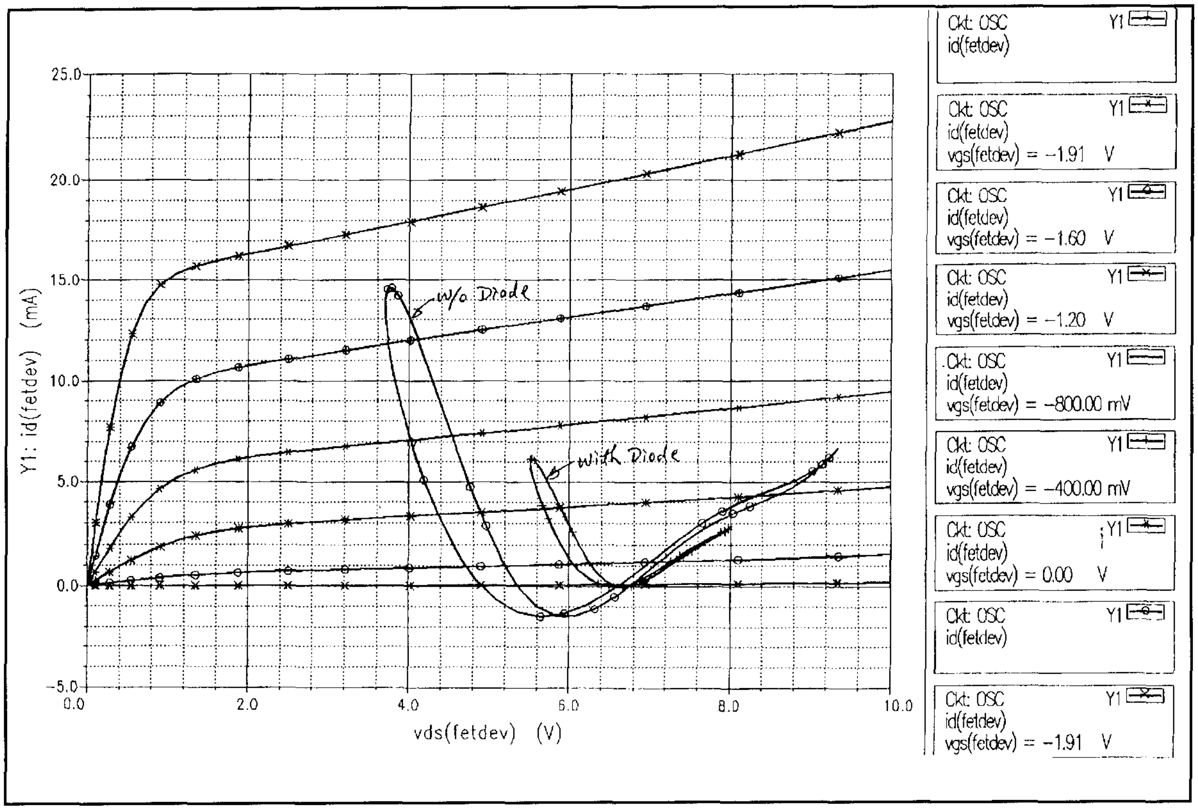

Figure 5. Output DC/IV curves with load line showing case with and without the diode.

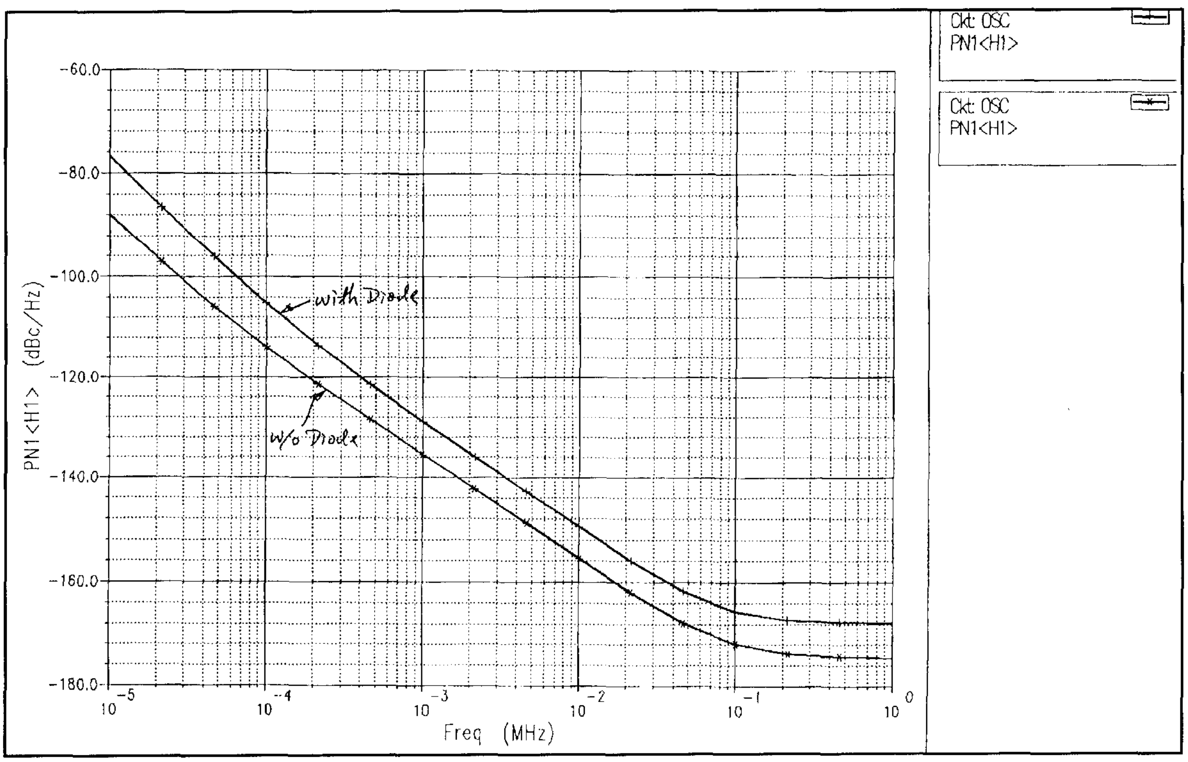

Figure 6. Predicted phase noise of the oscillator in *Figure 4* with and without the diode.

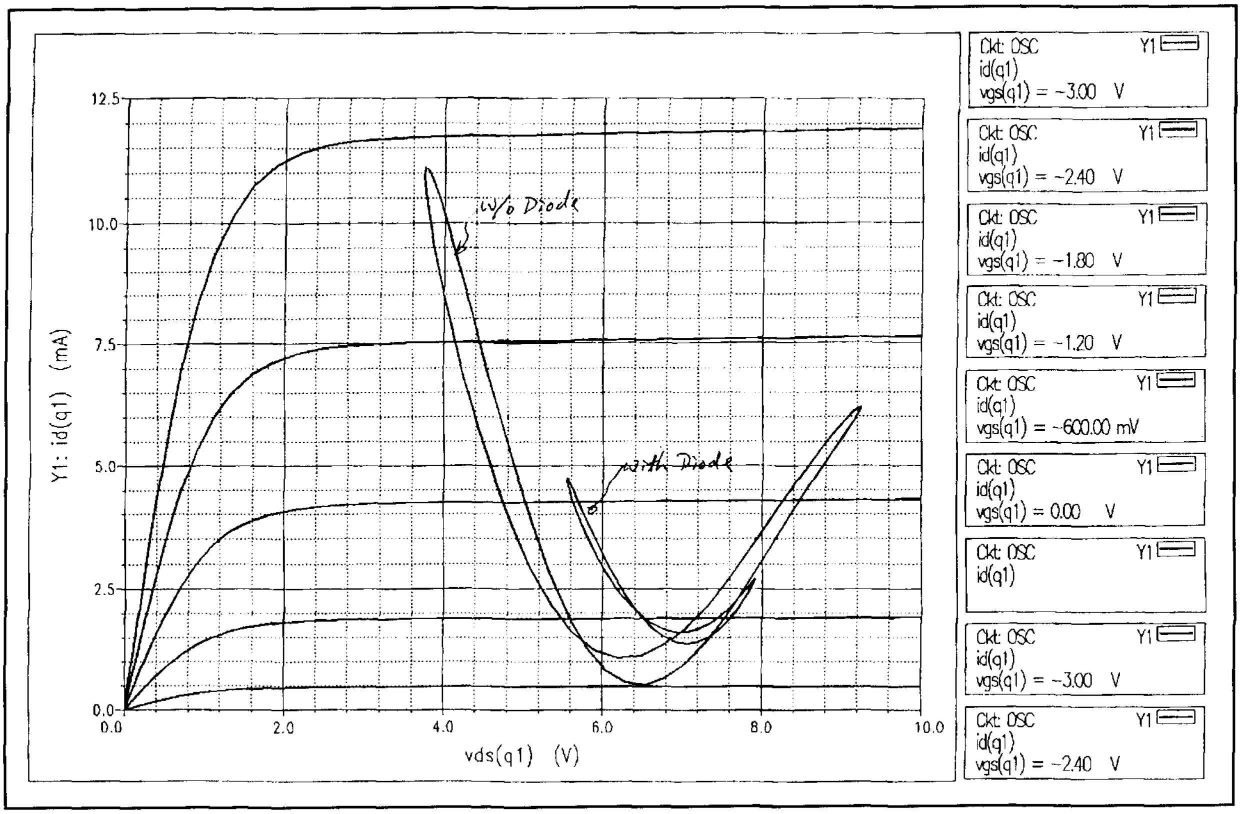

Figure 7. DC/IV output and load lines with and without the diode for a 2N3819 (Library Model).

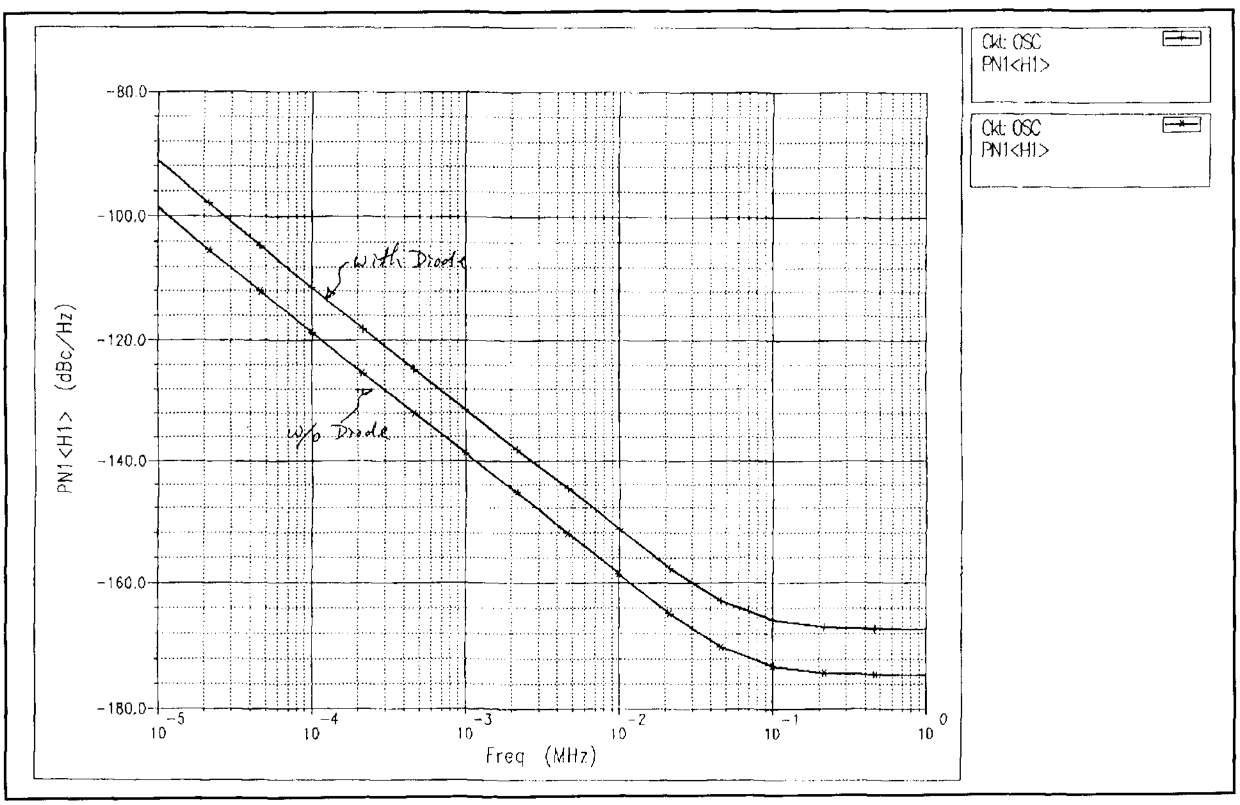

Figure 8. Predicted phase noise with and without the diode for the oscillator in *Figure 4* using the 2N3819 Library Model. Note the absence of flicker noise.

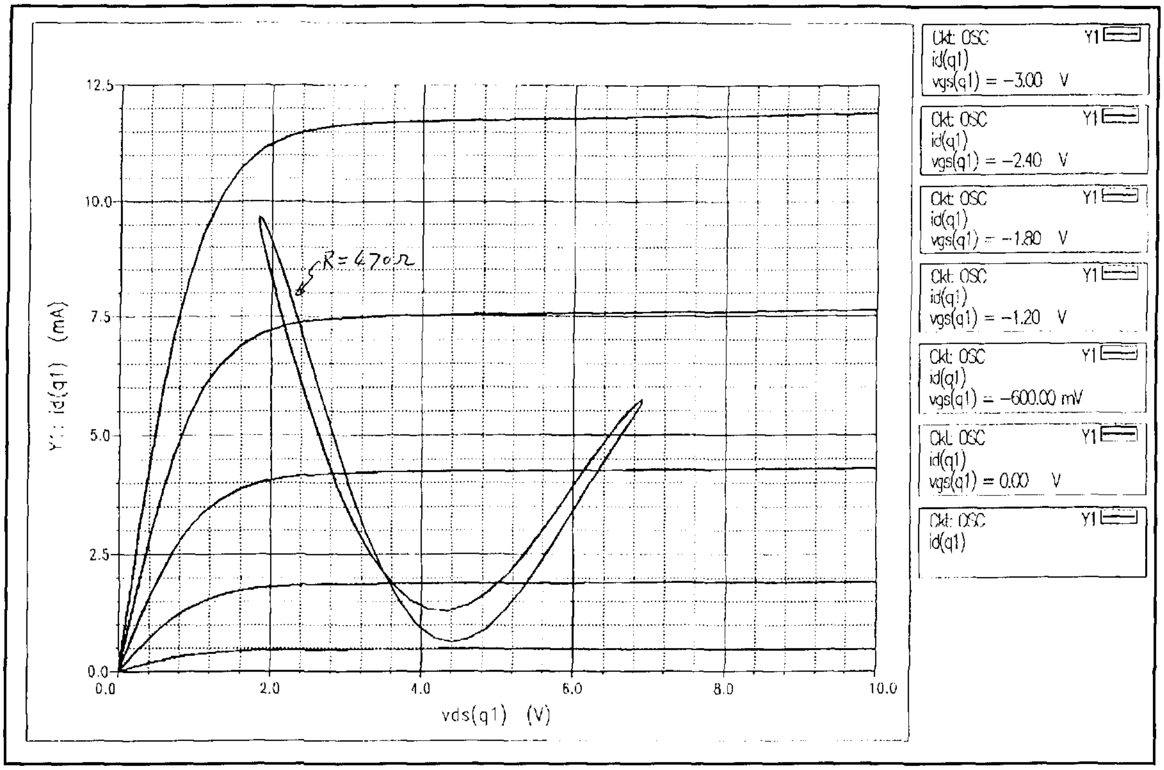

Figure 9. DC/IV output curve with load line with 470-ohm source resistor instead of the diode.

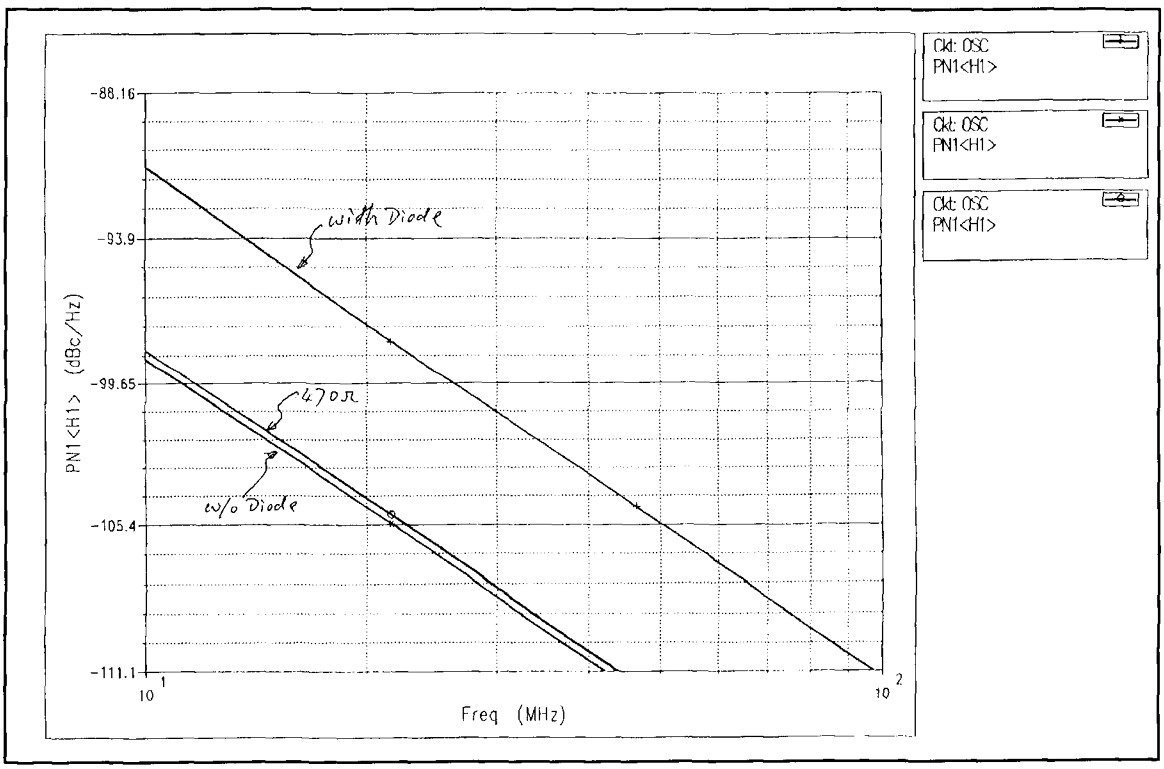

Figure 10. Predicted phase noise of the oscillator in *Figure 4* with no diode source resistor, with resistor and diode, and source resistor.

tor at the same bias point. Specifically, the S-parameters obtained from modeling and measured data *must* agree. Our experience in this area is based on a number of subcontracts we have received. Among these are our activities with Siemens Semiconductors in Munich where

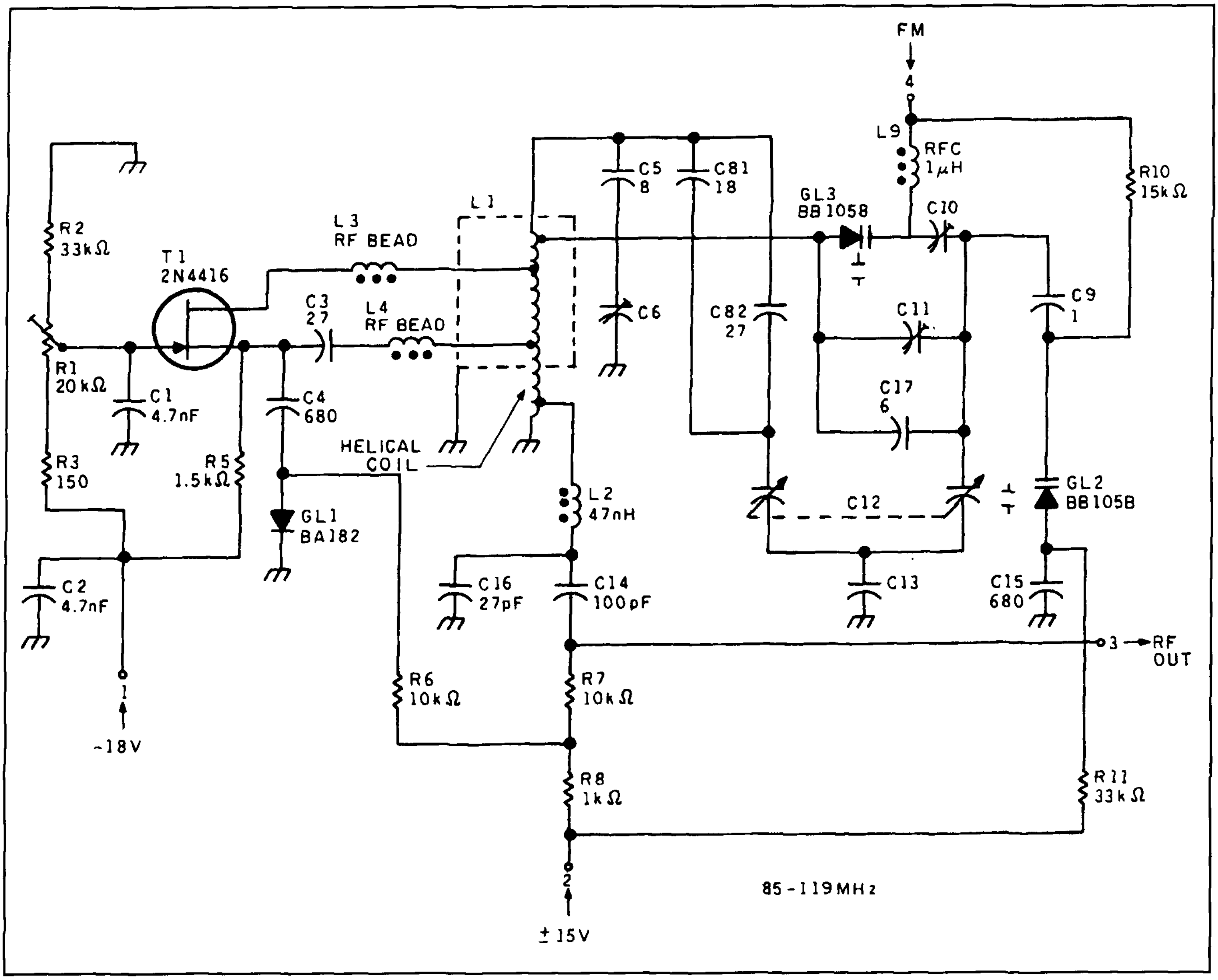

Figure 11. Circuit diagram for the 85 to 119-MHz FET oscillator from the Rohde & Schwarz SMDU signal generator.

CSI modeled their new 25-GHz f_t transistors, as well as GaAsFET lines including the new power FETs.

The FET model

In our opinion, the best known N junction FET is the 2N4416 or members of its family. **Table 1** shows the elements circuit file used for optimization. **Figures 1** and **2** show the agreement of S-parameters (measured versus predicted).

For the purpose of this paper, we took measured Y-parameters covering the frequency range up to 450 MHz and forced an optimization against the complex equivalent circuit for the linear FET (**Figure 3**).

We won't go into the equivalent procedure for nonlinear parameters, but Compact Software's *Scout* program (using the modified Materka model or the Statz or Curtice models) can generate an equivalent nonlinear circuit even for junction FETs, which are perfect for this procedure.

FET oscillator

The first FET oscillators looked fairly similar to those that came from the tube technology. **Figure 4** shows an oscillator taken from **Reference 1**.

This oscillator was built, measured, and modeled. Published reports on this in **References 7** and **8** caused considerable interest because no other work had ever mentioned the noise contribution of this diode. The basic oscillator is a modified Colpitts type that has a clamping or rectification diode in parallel at the input. The

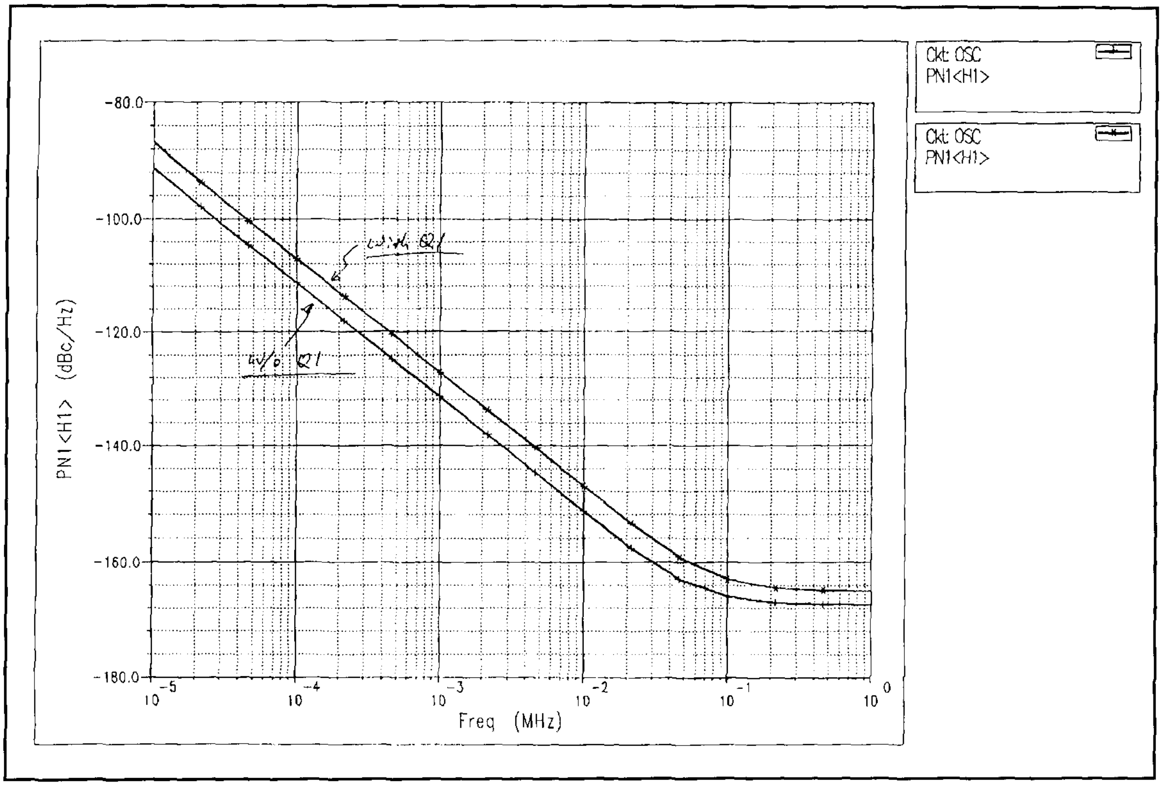

Figure 12. Predicted phase noise of the circuit in *Figure 4* with loaded Q of 135 and Q = 500, which is the same as Q = ∞

noise performance and output power of this transistor oscillator depends greatly on the nonlinear parameters of the transistor and the diode is claimed to provide more uniform performance—such as constant output power on tolerances of the transistor or temperature.[7,8] The diode had been said to increase the stability; however, it only improves thermal drift—not the SSB noise.

To better understand the circuit of **Reference 7**, we performed a phase noise evaluation for three cases: no diode, a clamping diode, and an RF-bypassed source resistor. We started with essentially the same transistor circuit shown in **Figure 4**; however, we initially eliminated the diode and the source resistor and then exchanged or added components. Consistent with the first statement, we used a specific nonlinear model, as implemented in the nonlinear program *Microwave Harmonica* with phase noise analysis capabilities, and added the parameters that described the transistor's noise performance.

Note that in the previously mentioned publication, both phase noise with and without the diode had been presented and compared to measured data.[7] The diode, which was connected in parallel between the gate and source, is activated only in an RF sense to look at the noise contribution without affecting the DC-bias point. Note that the junction FET already has a diode from gate to source and, therefore, one effectively puts two diodes with different threshold levels in parallel. **Figure 5** shows the output phase plane of the oscillator for both of these cases.

The next thing we did was to look at the noise prediction with or without the diode. Considering the fact that the diode at this time was fully active and changed the DC operating point, we knew the output power would drop by 8 dB down. Because the phase noise is the ratio of two power levels, the absolute power output of the oscillator is irrelevant for close-in phase noise considerations. By looking at **Figure 6**, we obtain the same phase noise as reported in **Reference 7,** and we also show the flicker noise contribution of the FET is the same when the diode is removed. With the diode, the output power is –6 dBm; without the diode, the output power is +2 dBm.

The next step was to evaluate the quality of the *SPICE* library models. **Figure 7** shows the simulation of the load line using the parameters supplied by a third party *SPICE* library (*PSPICE*). Close examination of the DC IV curve reveals an unrealistic characteristic of the FET in the saturation mode. This early method of (simple) parameter extraction is insufficient for accurate high frequency applications. There is no provision for parasitic elements.

Having exchanged the two transistor models for the same circuit, we recalculated the noise.

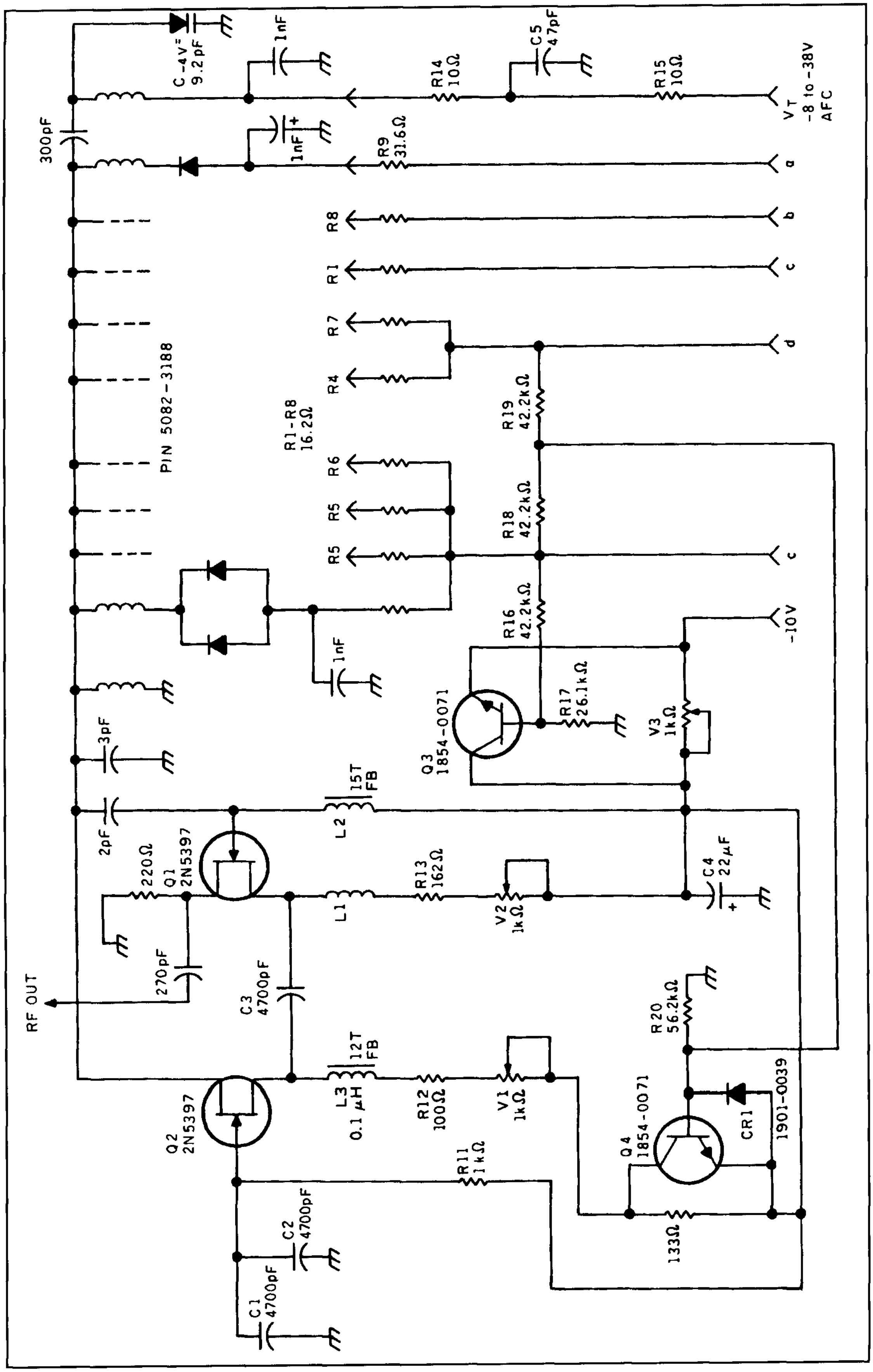

Figure 13. Schematic of the HP 8662A VCO operating from 260 to 520 MHz.

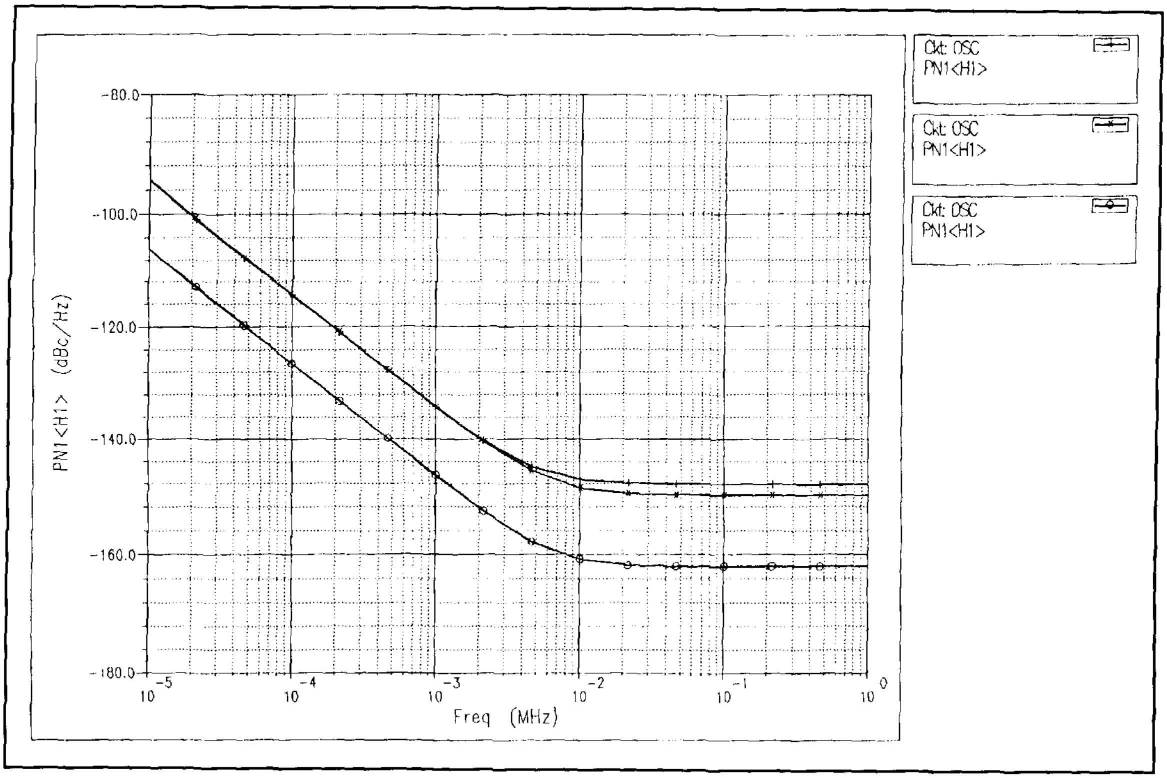

Figure 14. Predicted phase noise of an FET differential limiter oscillator for low current. Top curve and low current of 10 mA (lower curve).

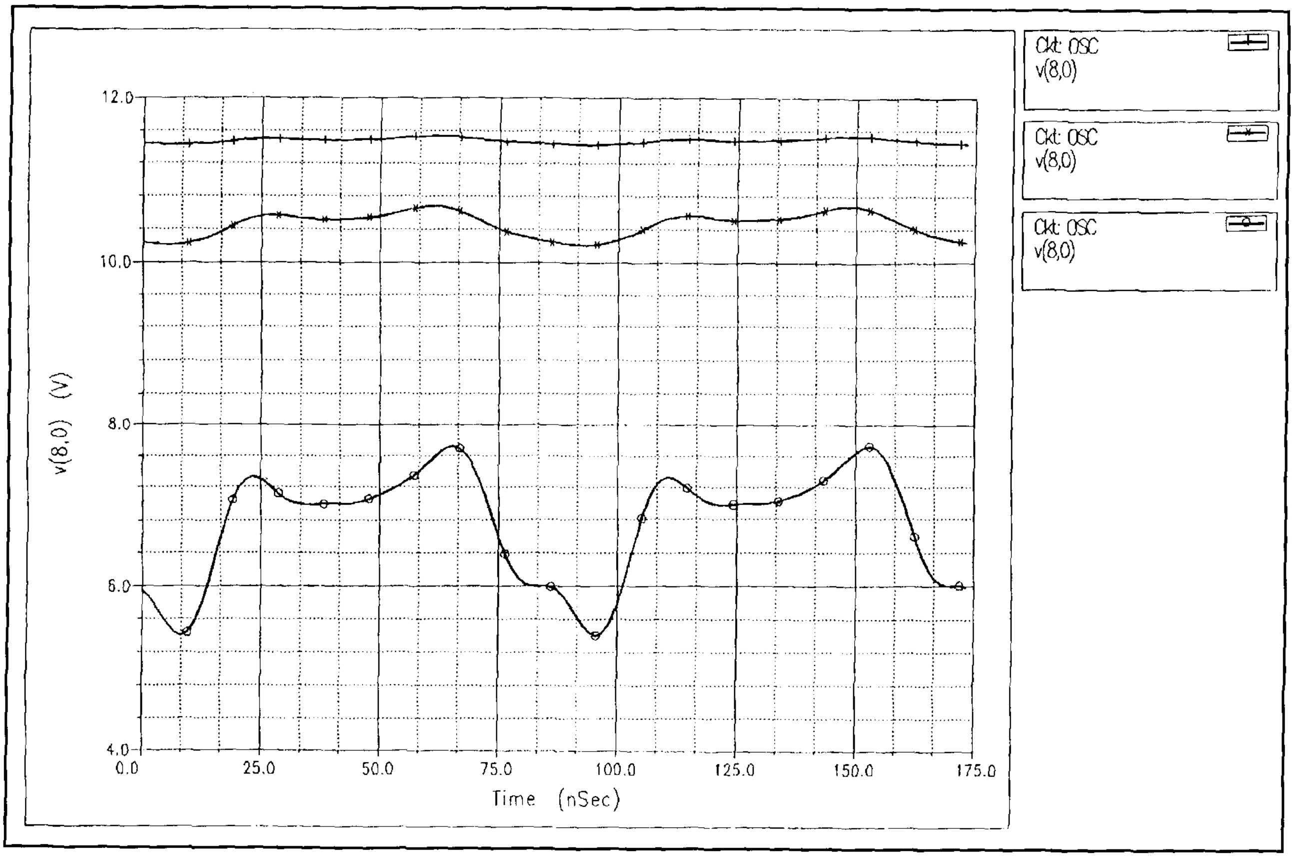

Figure 15. Output waveform of an FET differential oscillator for three different bias points.

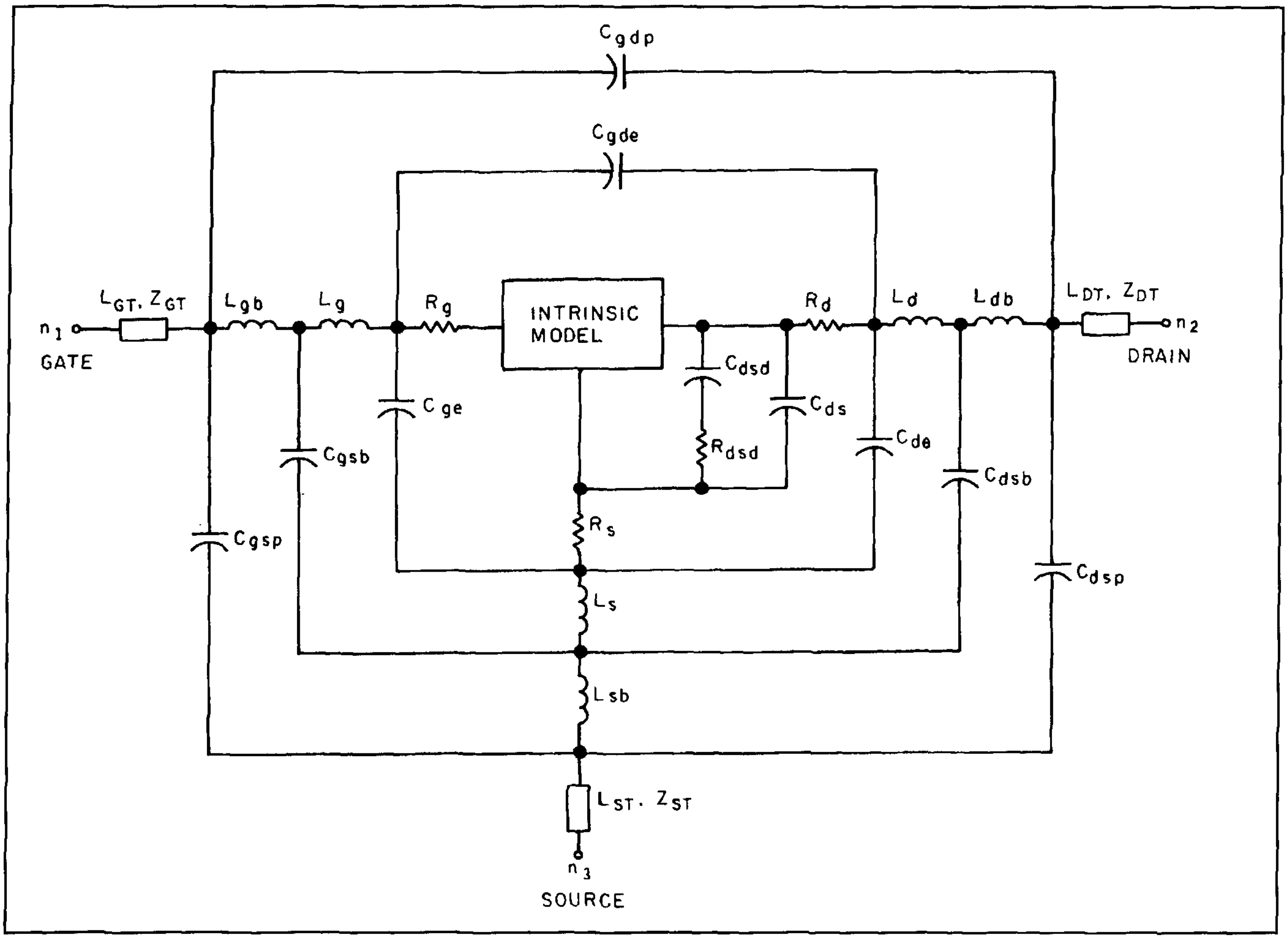

Figure 16. Extrinsic parasitics used to accurately model high frequency effects of FETs.

Once again, it showed a significant difference with and without the diode. This was consistent with the measured data previously reported, which is shown in **Figure 8**.

The output power for this oscillator was the same as the simulation with our model. This indicates that the calculation of the output power is less sensitive to the parameters that make the difference in the noise calculation.

When looking at the noise calculation, the two most relevant quantities are the loaded Q under operation, which the simulator predicts, and the noise contribution of the semiconductor. It's interesting to note that the *Spice* library totally underestimated the transistor noise contribution and provided unrealistic low close-in phase noise measurements.

The next logical step was to evaluate the circuit without the diode, but with a resistor in the source. **Figure 9** shows the output load line/DC IV curves for the same circuit with a 470-ohm resistor. What results in this case is +1 dBm output that compares favorably with the example without the diode. We now encounter the exciting question, "What will the phase noise do?"

Figure 10 illustrates all three cases: phase noise without the diode, AGC action with a 470-ohm source resistor, and with the diode, as previously mentioned. It becomes clear that the phase noise with the diode still remains the worse case scenario, while the DC feedback is the best case. This is also consistent with reports in **Reference 3**.

For further reduction of the undesired "warm-up" effect of the FET at high bias point, one must go to emitter resistors as large as 1 k or higher and may even need to bias the gate positive.

Figure 11 shows the circuit diagram for the 85 to 119 MHz FET oscillator from the Rohde & Schwarz SMDU signal generator. This signal generator, which is no longer in production, had the same specifications as the HP 8640 and obtained its extremely low level of phase noise with the use of a helical resonator. It reflects the situation above where the DC bias of the transistor is provided by a fairly large source resistor of 1.5 k and an adjustable resistor at the

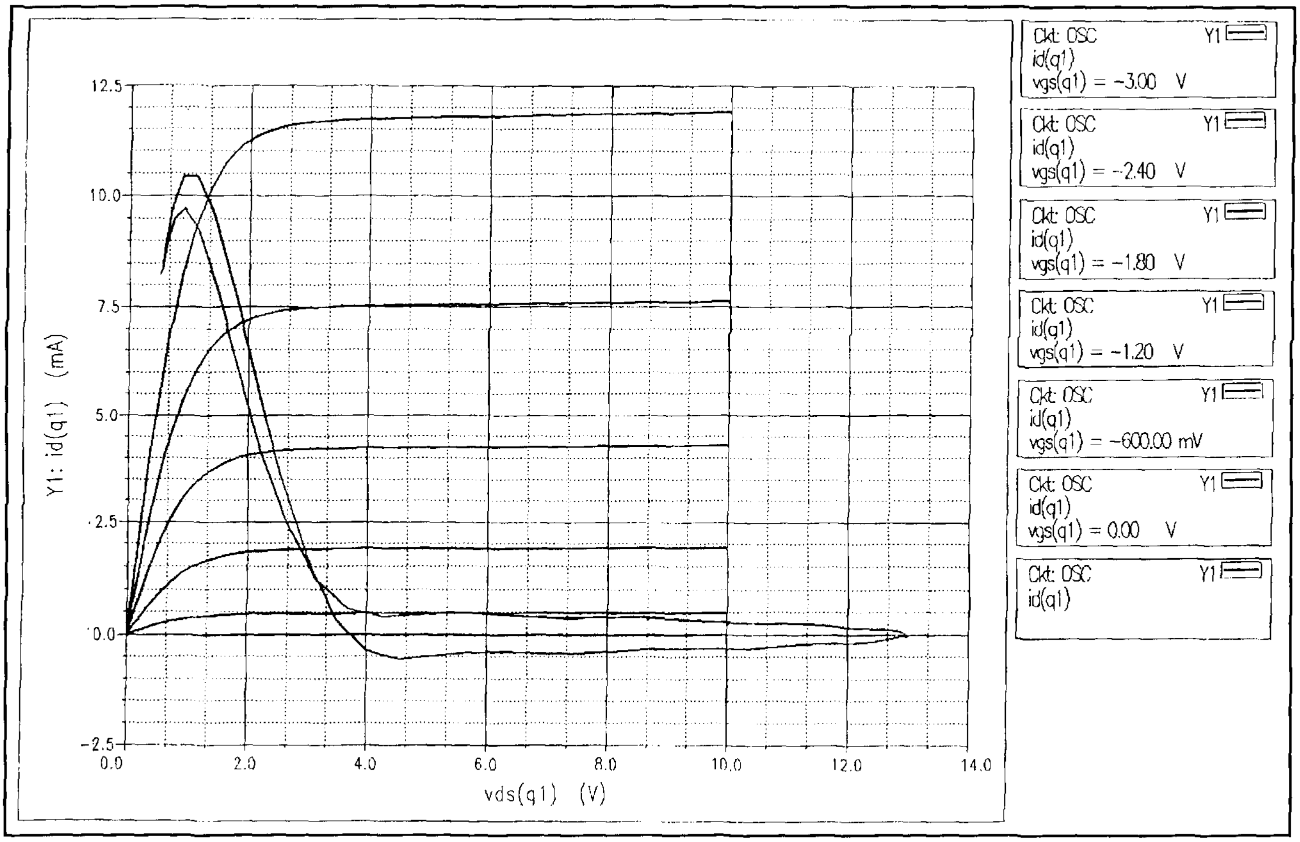

Figure 17. DC/IV curve of the FET oscillator with the external parasitic effects removed.

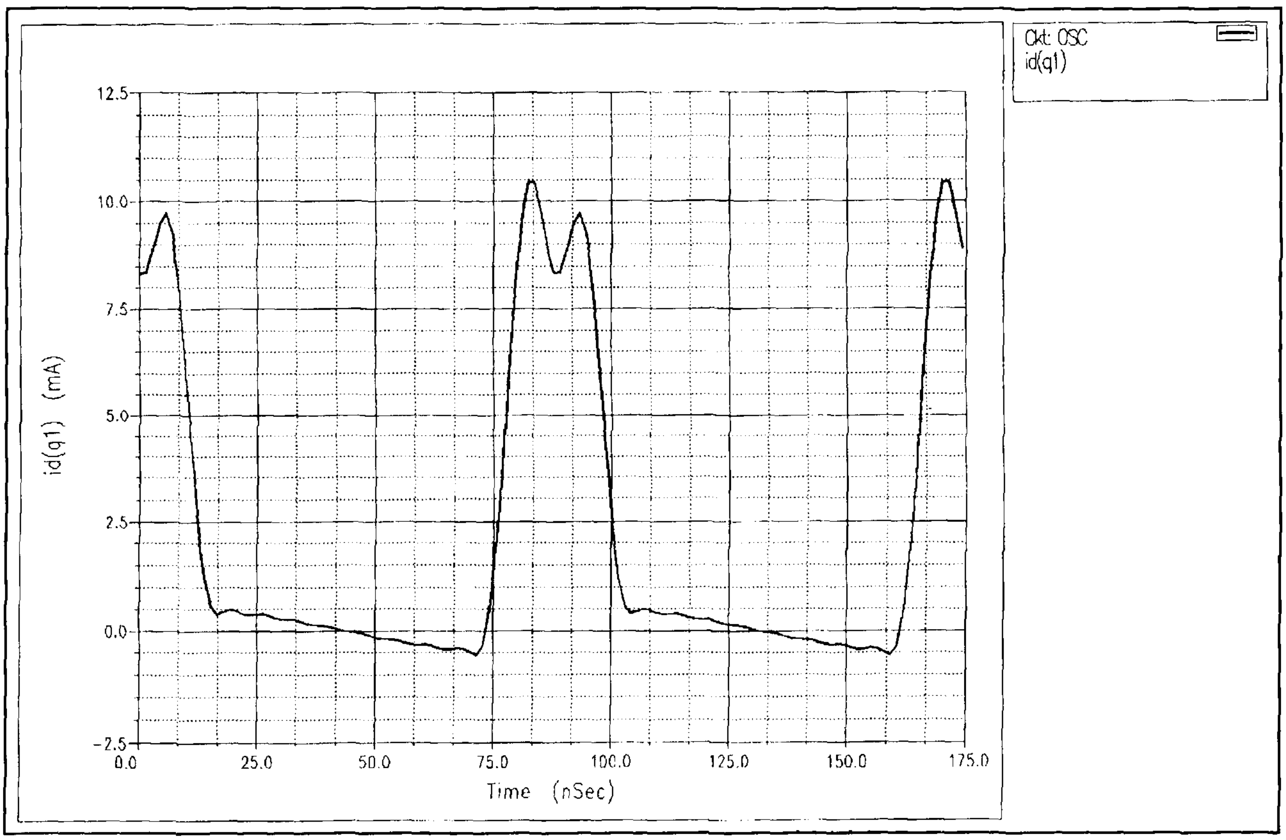

Figure 18. Drain current of the FET oscillator. Please note the short duty cycle.

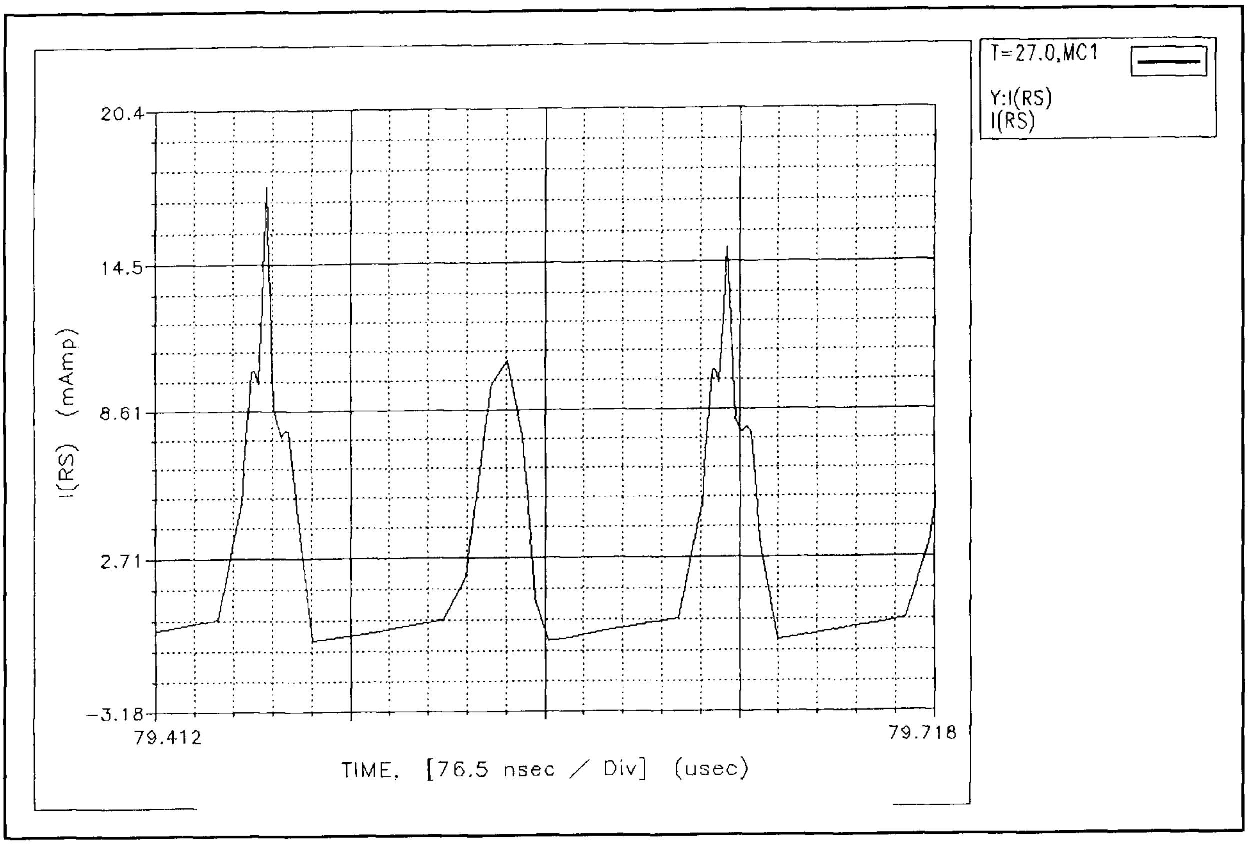

Figure 19. ***SPICE*** **analysis of the FET oscillator showing the drain current. Please note the numerical instability due to the limited numeric accuracy inherent in** ***SPICE.***

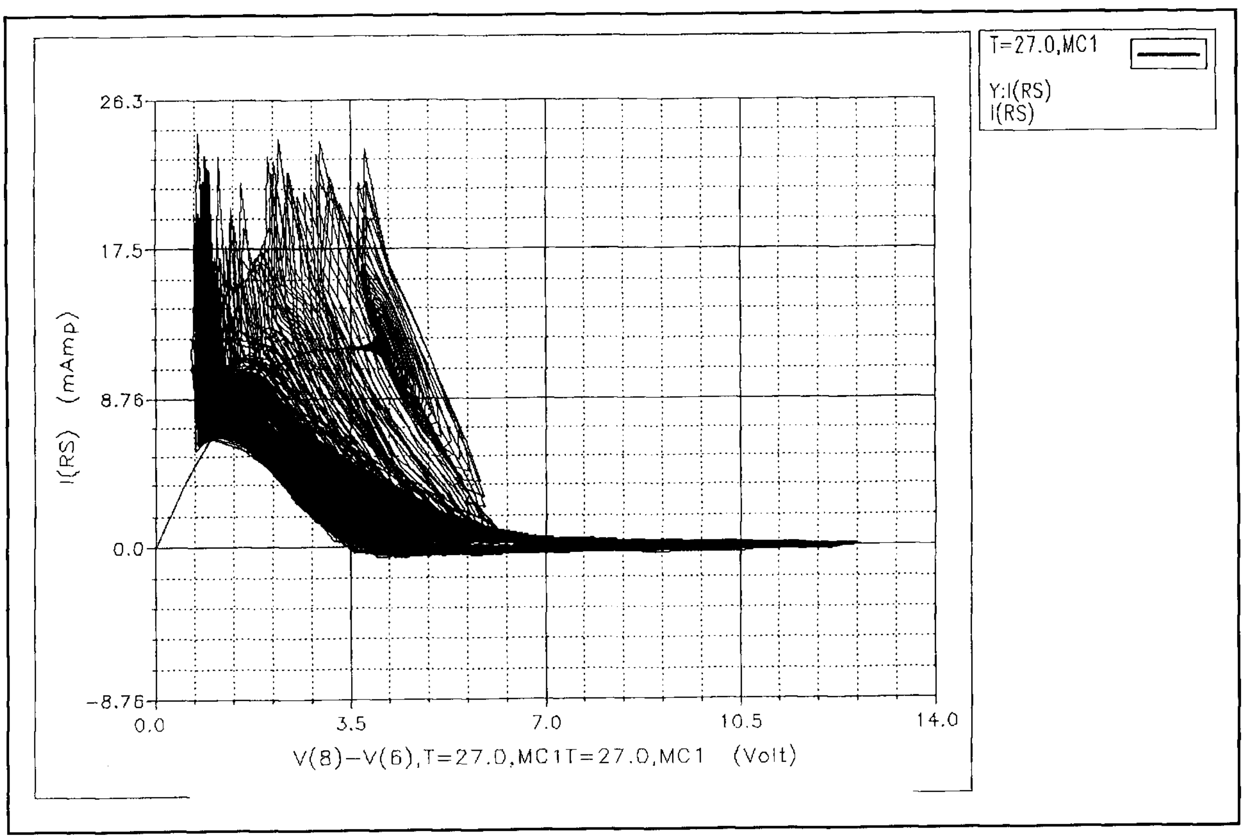

Figure 20. DC/IV curve starting condition of FET oscillator using ***SPICE.*** **DC curve starts at the left-hand corner, works its way up and then moves back to the left with a resulting DC operating point shift.**

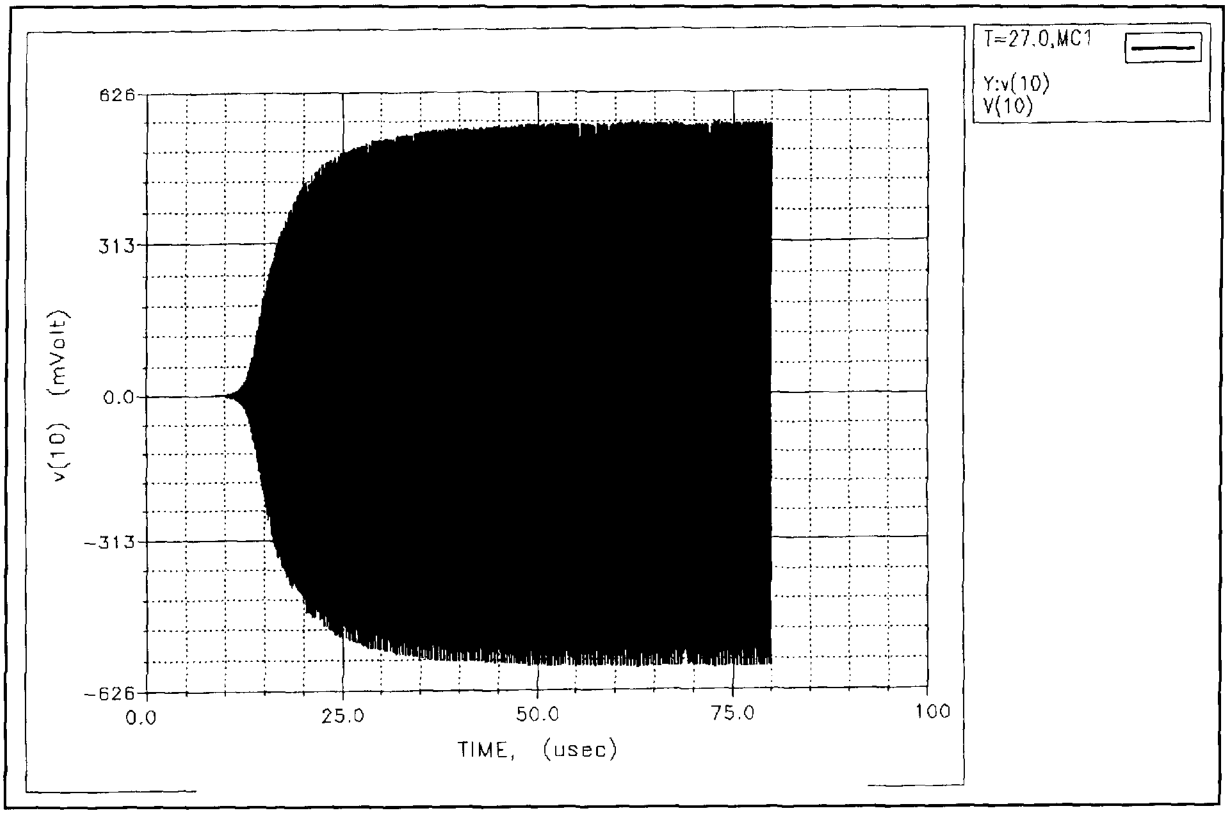

Figure 21. Transient accuracy of the FET oscillator. Takes about 50 µS before having a stable amplitude.

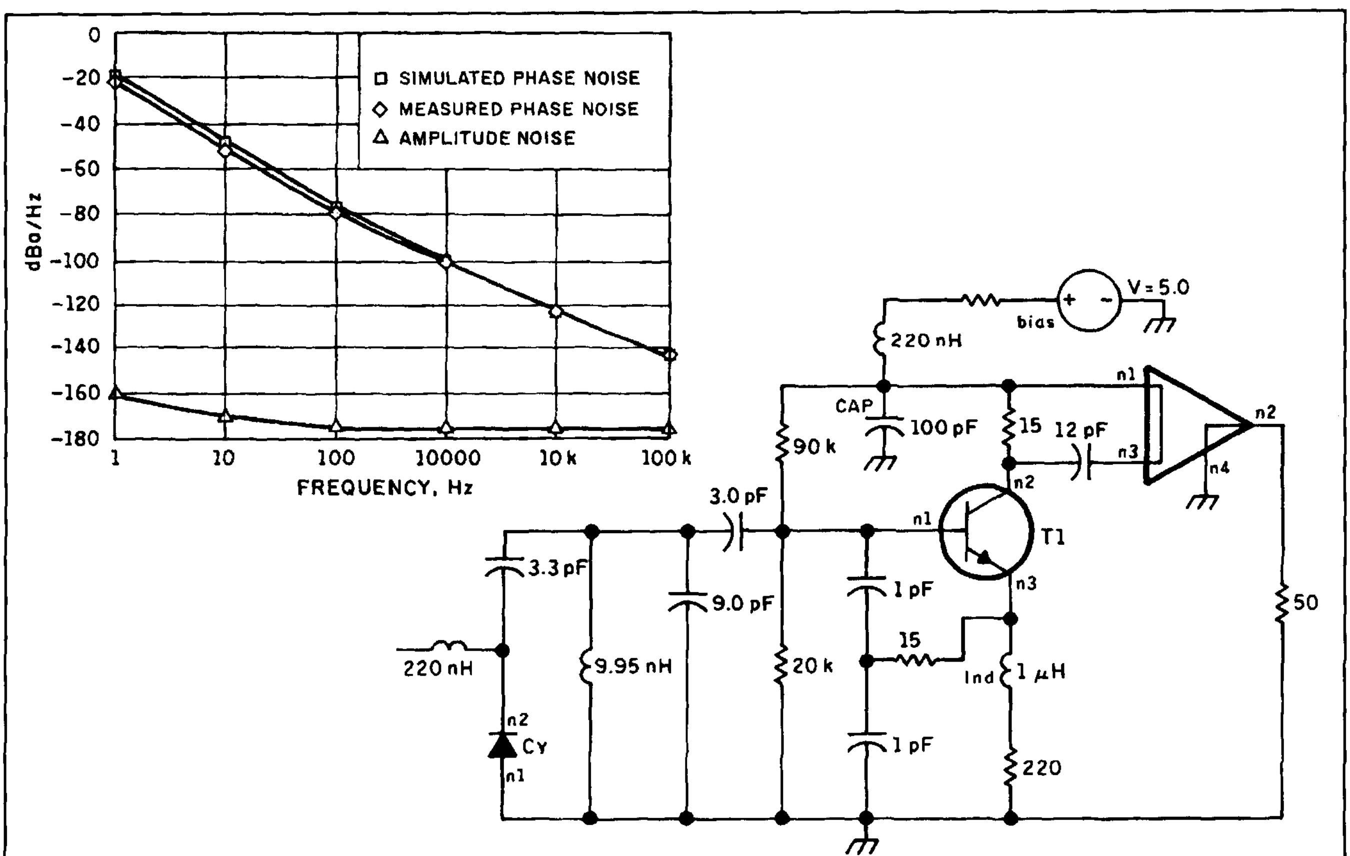

Figure 22. (A) Schematic of the Motorola 800-MHz VCO. (B) Comparison between the predicted and measured phase noise. The resistor between the emitter and the capacitive feedback divider reduces the flicker noise contribution.

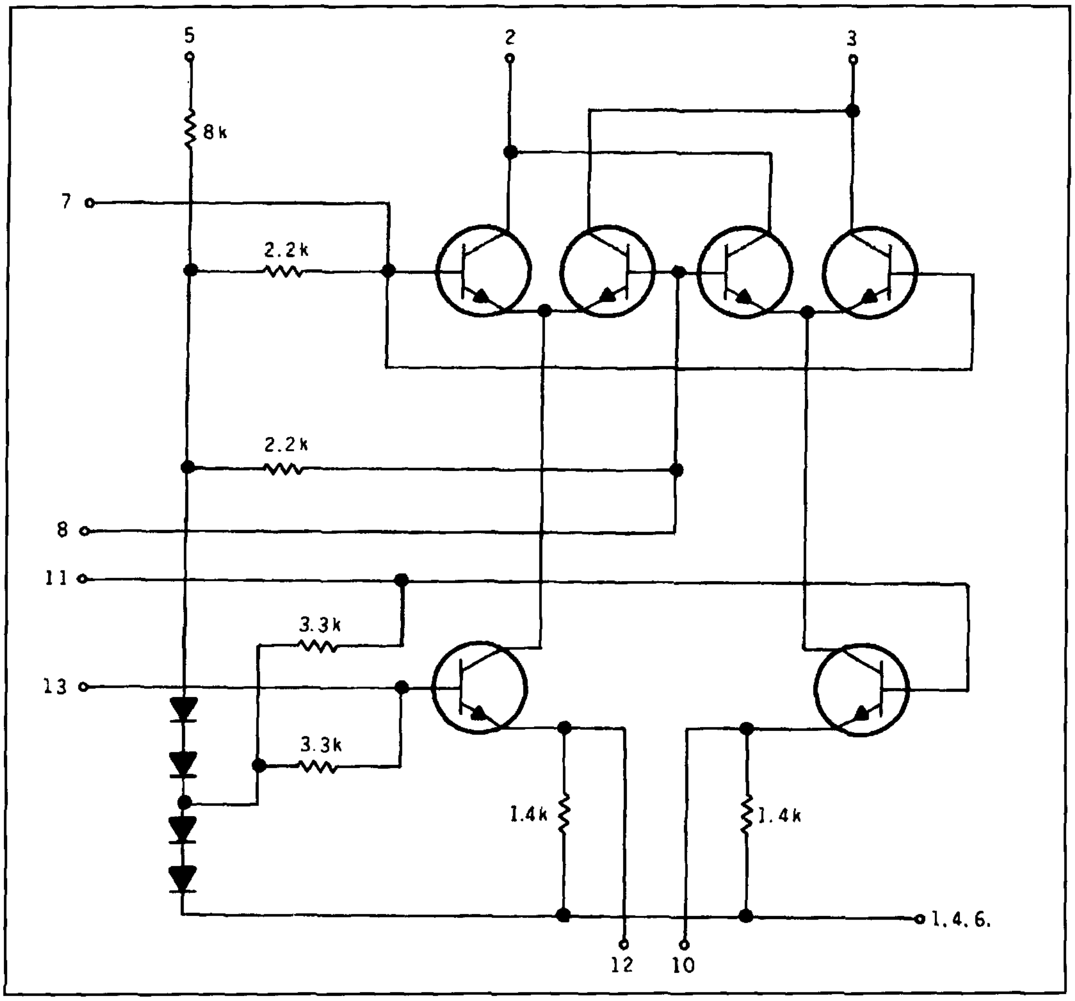

Figure 23. Pair of differential transistor amplifiers that are found in most modern IC designs.

gate offers the optimum bias condition. The large DC feedback compensates for both tolerance in the device and temperature effects. This is the correct method of operating the transistor rather than using an AGC diode.

Other questions arise in the modeling area. How much does the actual Q of the inductor affect the performance in high Q cases like that of the 10-MHz oscillator? Is it permissible to allow the Q to be simulated at Q = ∞ where Q = 500, which is obtainable at these frequencies, is used as a point of reference. If we examine a more modest Q, like 135, we get approximately 4-dB deterioration of phase noise, which shows up at both close in and far out. For high frequency applications above 30 MHz, it's certainly important to properly model the Q of the tuned circuit before it is connected to the transistor. **Figure 12** shows the noise for differently loaded Q.

When using integrated circuits for DC coupling, one must resort to symmetrical circuitry using a differential limiter type oscillator circuit, as shown in **Reference 3**. Unfortunately, the outputs of these circuits were distorted and we had questions about their phase noise performance. We took the same circuit, and essentially using FETs, modeled a differential oscillator circuit. This circuit is consistent with that in **Reference 3**, but operates at 10 MHz and has a less elaborate DC circuitry. (See **Figure 13**, which shows the differential FET oscillator that has become the heart of the HP 8662—the Hewlett-Packard signal generator).

The initial starting condition was set by having the common source resistor at 470 ohms, which produced the phase noise prediction of the first plot shown in **Figure 14**. By playing with the output load, the phase noise far out can be improved slightly. We finally decided to take more drastic steps and reduce the source resistor by a factor of 10. As can be seen, this decreased the output phase noise significantly. The phase noise improvement is approximately 12 dB across the board. This noise calculation is still somewhat unrealistically optimistic

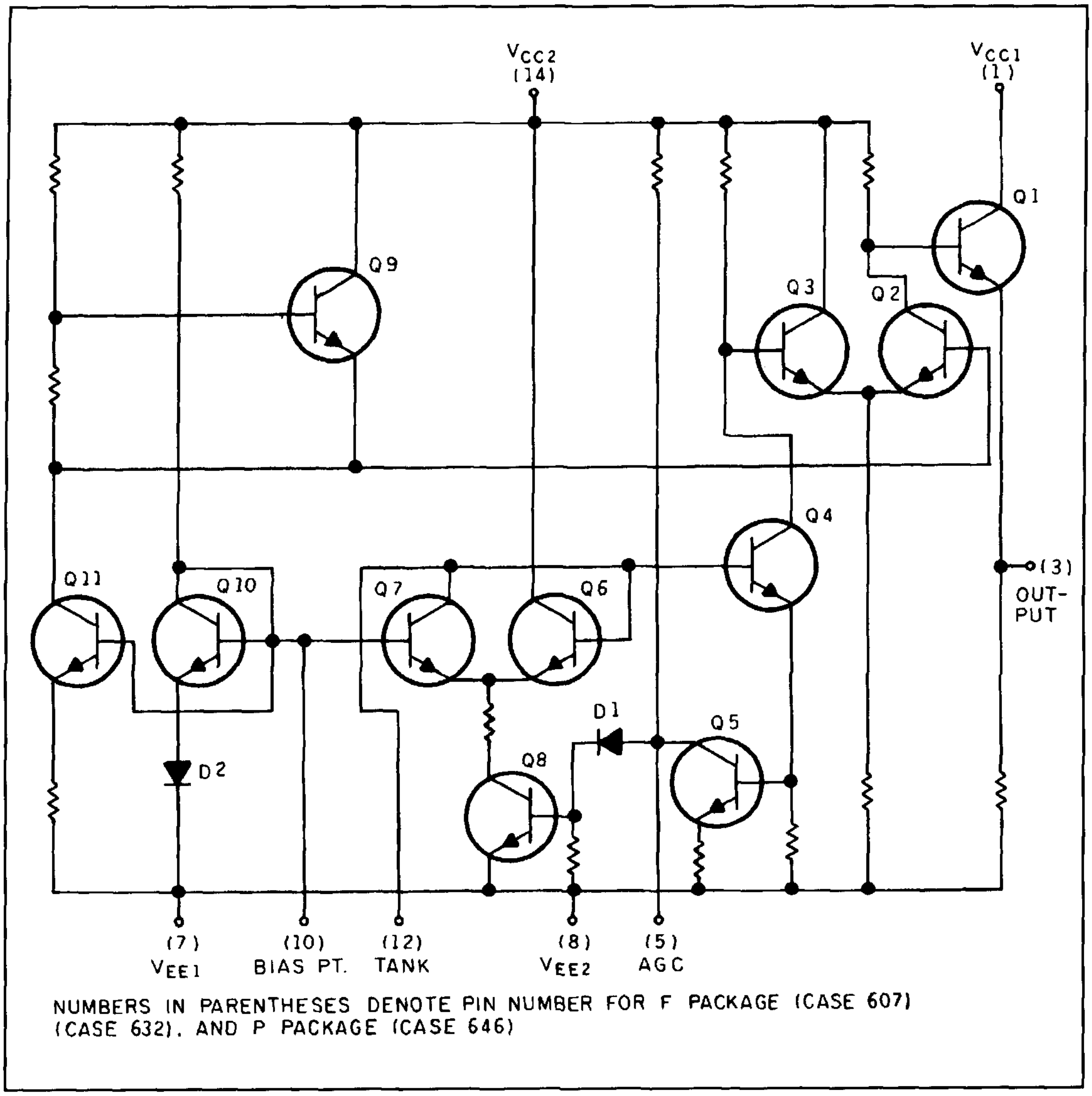

Figure 24. Schematic for the Motorola MC1648 oscillator chip.

because the FET model supplied is unrealistic in its flicker noise contribution; however, it clearly illustrates the trend.

Figure 15 shows the output waveform as a function of DC bias when the absolute output of the differential oscillator is –6 dBm.

The N junction FET model in *SPICE* doesn't consider any parasitics, which limits its practical high-frequency use. **Table 1** indicates that the first criteria for successful modeling involves a match between predicted and measured S-parameters.

Figure 16 depicts the equivalent circuit for the packaged nonlinear FET. To model energy traps in the transistor, a series combination of R and C is identified as parasitics labeled CDSD and RDSD. By setting those two values to zero, the DC/IV curve shifts looks as shown in **Figure 17**. The fly-wheel effect to the right of the transistor has disappeared. This is the case with no diode connected and no emitter resistance. **Figure 18** shows the time domain display of the drain current of the device. As one looks at the this figure, it becomes obvious that this has a fairly narrow duty cycle and, therefore, a high harmonic content. For the sake of completeness, one should now look at a *SPICE* modeling of the same circuit. Remember that the *SPICE* models are less complete for high frequencies. Also, the inherent dynamic range is less than that of our harmonic balance simulator. *SPICE* typically is lucky to get 60-dB dynamic range. Our harmonic balance simulator has 180 dB dynamic range. By inspecting **Figure 19**, it's apparent that there is no difference in resistance, as the drain current is numerically unstable compared to the calculation in harmonic balance.

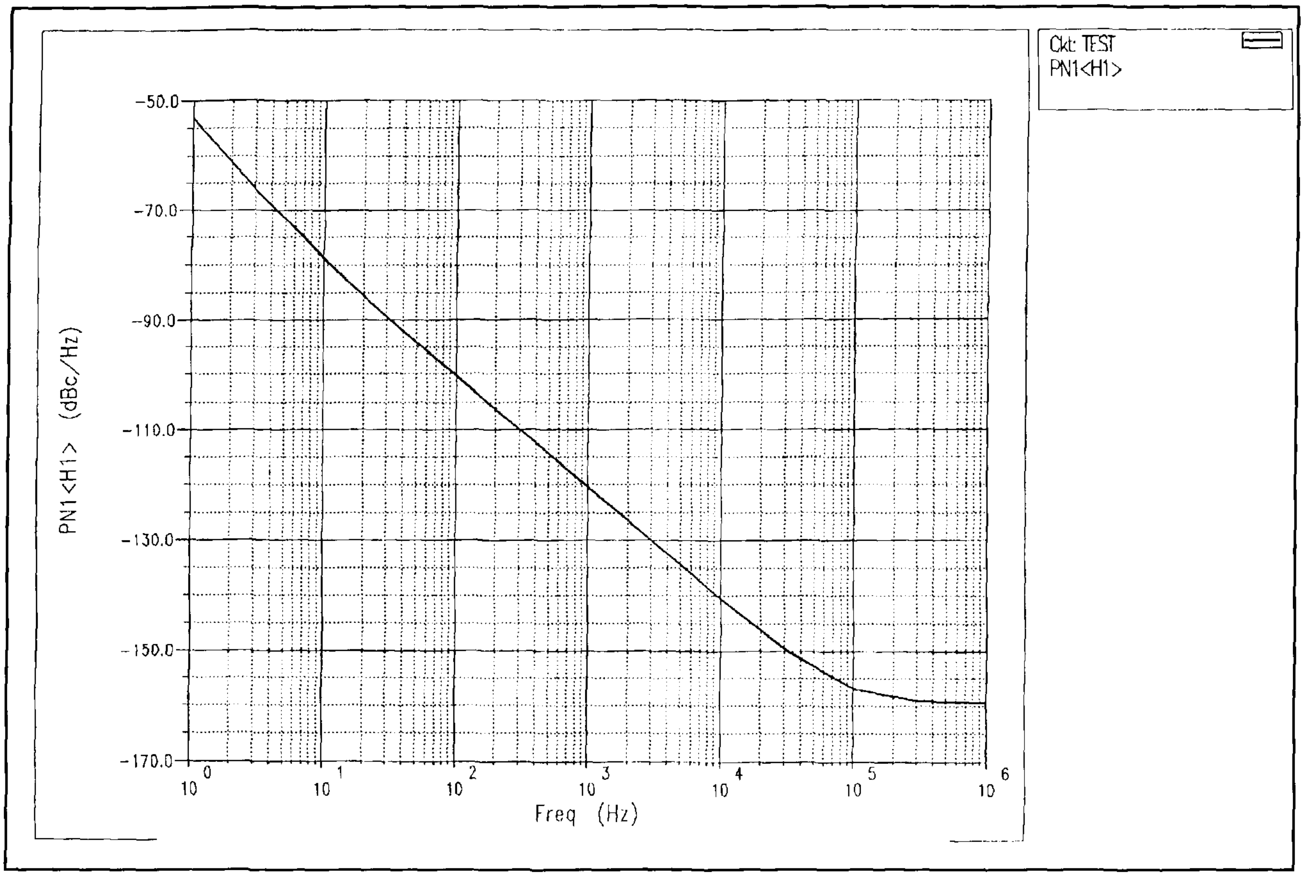

Figure 25. Simulated phase noise for the oscillator circuitry shown in *Figure 15* at 10 MHz. By reducing the collector current of the transistor's Q6 and 7 down to 500 μA, the flicker noise would be improved from 30 dB at 1 Hz to 52 dB at 1 Hz.

The big advantage *SPICE* offers is that it lets one look at the transient response and not just at the steady state supplied by the harmonic balance simulation. **Figure 20** shows the "start up" condition for the same oscillator. This "start up" condition begins at the lower left-hand corner with current and voltage at the zero level. By the time the voltage increases to 3.5 volts and higher, the current begins to saturate and oscillation starts. The DC bias point moves to the left and then automatically settles down. This curve becomes similar to the DC/IV curve previously shown in **Figure 19**. Finally, it's interesting to measure the time it takes the oscillation to actually start. This is done by inspecting the load voltage. The oscillation hesitates 10 μS, then builds up to reach most of its amplitude after 25 μS (and certainly after 50 μS), before it achieves stable amplitude. This is shown in **Figure 21**.

At this point we ended our investigation of FETs and decided to look at the bipolar transistor circuits. The reason for our interest is that most hand-held radios or wireless applications will be built with silicon bipolar technology in the future and, therefore, the noise of those circuits will become an issue. The single transistor oscillator operating at 800 MHz had already been successfully modeled and reported on in **Reference 4**. **Figure 22** provides a plot of the predicted and measured phase noise, as well as the oscillator's schematic. Oscillators for highly integrated circuits make intensive use of differential types of transistor pairs. **Figure 23** shows the standard two differential transistor pairs as offered by many suppliers. This has become the de facto standard for the pairs in either mixers or oscillators.

Motorola built the first successful medium-scale integrated circuit oscillator, the MC1648M oscillator chip. We were particularly interested in modeling this circuit because initial inspection of the circuit indicates there is no DC voltage difference between the chip and collector of the transistors on the left. **Figure 24** is the schematic of the internal oscillator as published by Motorola.

We were pleased to be able to successfully model the Motorola circuit, but the resulting phase noise was quite surprising. **Figure 25** shows the phase noise of the bipolar transistor IC. It is well understood that the phase noise of these oscillators is significantly higher than the phase noise of the FETs. It was quite surprising to find that the flicker noise contribution worsened the phase noise to such a degree. We did a

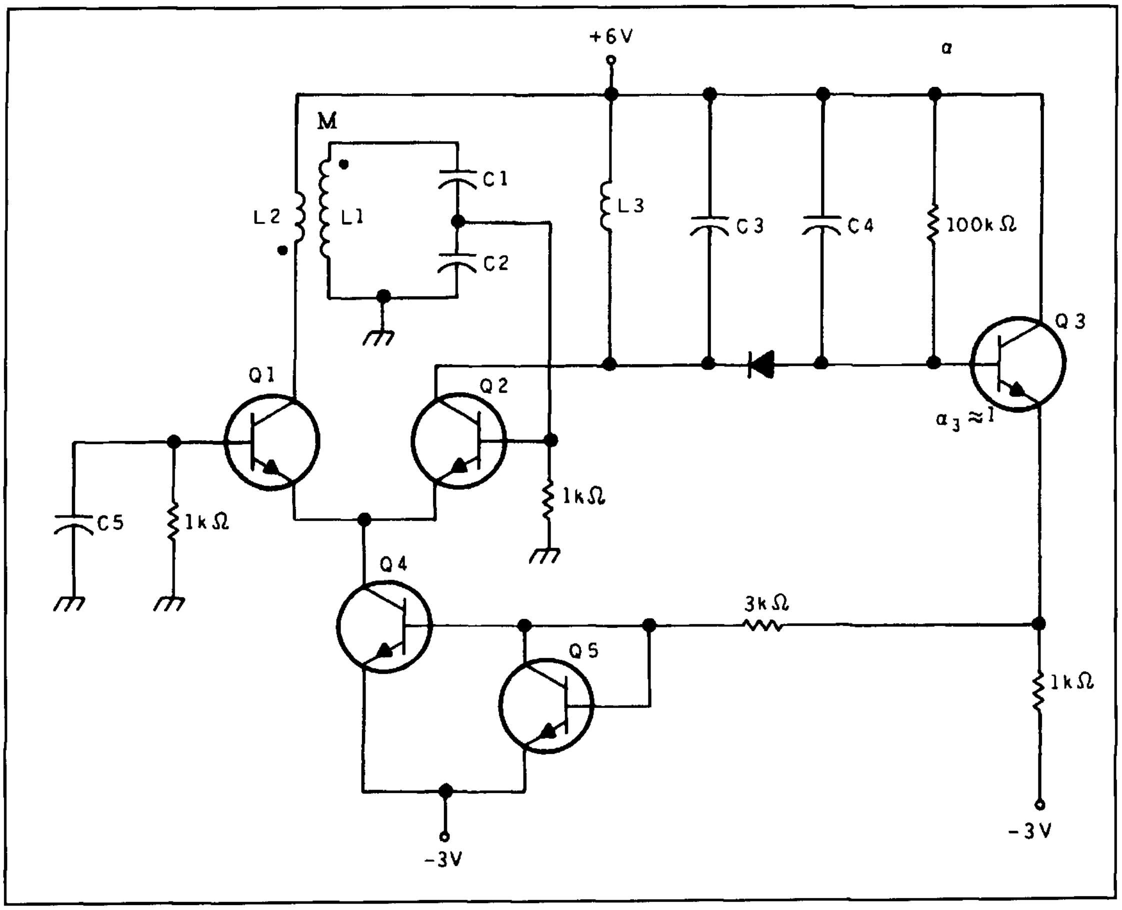

Figure 26. Oscillator using a differential amplifier and a biasing scheme. This should be used as an analysis example for modern CAD tools.

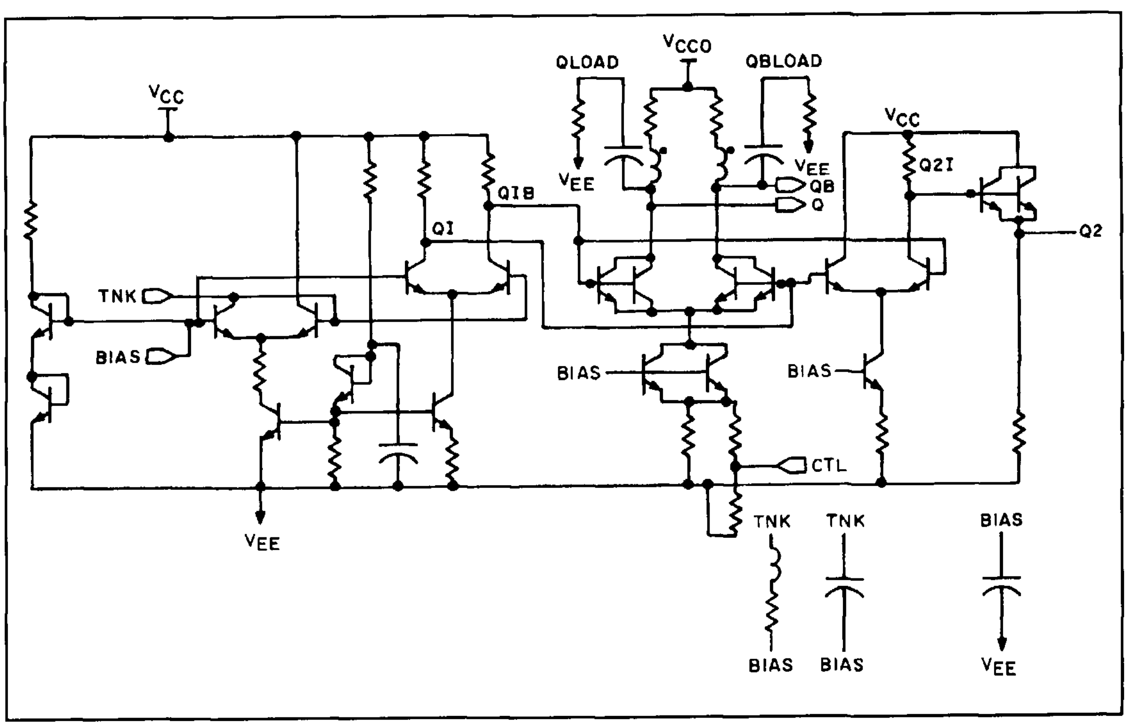

Figure 27. Schematic of the Motorola VCO shown in *Photo A*.

little experiment where we changed the bias of the oscillator and reduced the loop gain. The DC current dropped from 3 mA to approximately 500 µA. Because the flicker noise is extremely bias-dependent, the phase noise improved; however, the output power dropped by about 10 dB. Simulation of this circuit also showed a nonsinusoidal waveform at the output, which, due to space constraints, is not reproduced here. A circuit typically used for LSI application is shown in **Figure 26** and is taken from **Reference 1**. These types of oscillators are actually being used for various applications and the interested user should use modern CAD tools to investigate their performance. Due to the heavy feedback in those oscillators, the simulation shows that they are not as sensitive to parameter variation of the devices as one might expect. A particularly nice application of this type is a Motorola VCO done entirely on silicon material. **Photo A** shows its layout and **Figure 27** shows its schematic representation. The monolithic inductors used in the VCO were designed using Compact Software's *Microwave Explorer* program to model the effect of various ways of building those inductors. Measured and predicted Qs of the inductors show extremely good agreement as comparisons. **Figure 28** shows a discrete VCO for complexity and size. One can see that considerable progress has been made here in high integration and first-hand layout of these devices, when compared with the older version of **Figure 29**.

As mentioned, these types of circuits are found in large integrated circuits, and we decided to model the circuit of **Figure 30** shown in **Reference 6**. What's different between this circuit and others previously reported in this paper is that the tuned circuit is coupled very loosely to the transistor. We also made a modification to the circuit by taking the feedback from the second transistor rather than the first and placing the second transistor in the loop. **Figure 31** shows a proper presentation with modern ICs.

When comparing these two cases with the modified one, we found an improvement of approximately 6 dB of the SSB phase noise with the same output power. **Figure 32** shows the predicted phase noise and **Figure 33** shows the output waveform available at the collector. According to our simulation analysis, both circuits only provide +4 dBm output. We weren't able to confirm the +17 dBm measurement reported by the authors.

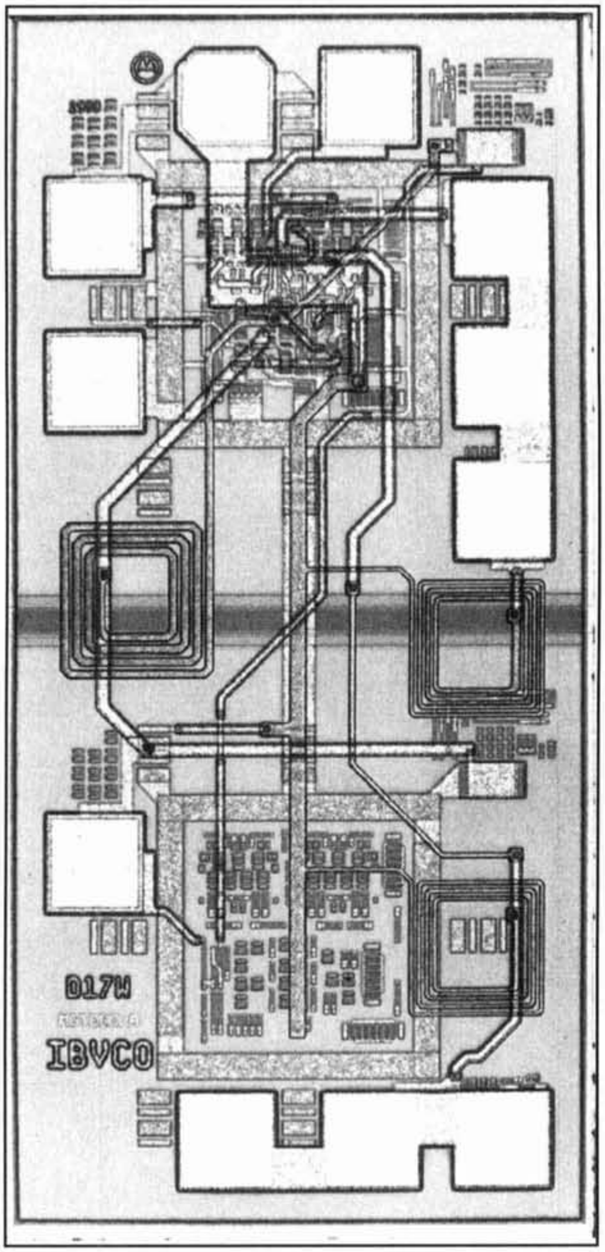

Photo A. Layout of a Motorola VCO for VHF handie-talkies, which is built around silicon substrate material.

Circuit optimization

The oscillator circuit itself provides several degrees of freedom over which the designer has control. One of the three key issues to consider is the DC operating point of the transistor. While it varies the output power, we have shown that it more drastically changes the flicker noise. Secondly, the feedback circuit, as such, is responsible for the loop gain and the loaded Q. Third, some negative rather than positive feedback—like resistive feedback between the emitter of the bipolar transistor and the capacitive voltage divider—is responsible for the phase noise reduction to the flicker noise contribution. The automated optimization as implemented in *Microwave Harmonica* can significantly improve the circuit.

Figure 34 shows the phase noise—before

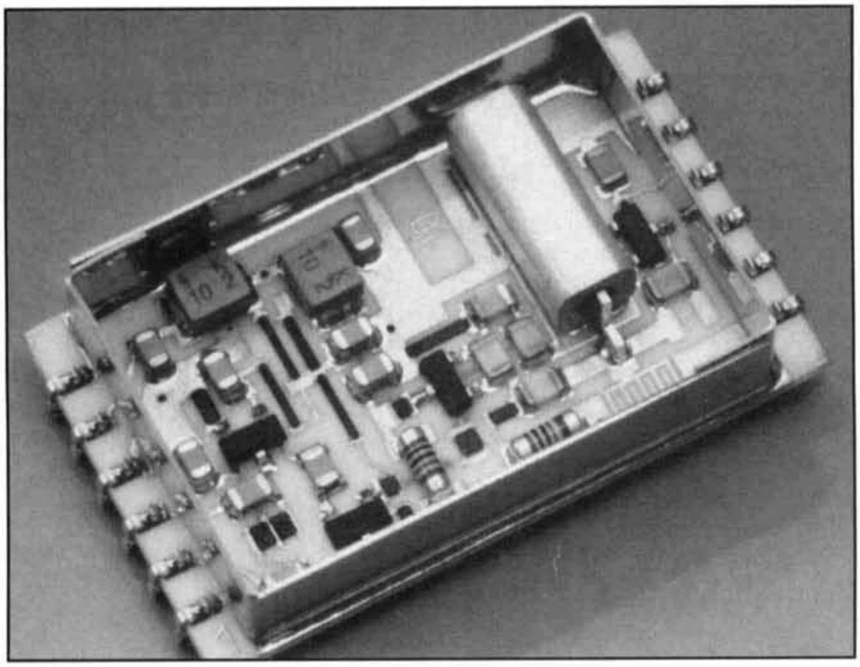

Photo B. Modern ceramic resonator oscillator for hand-held cellular telephones.

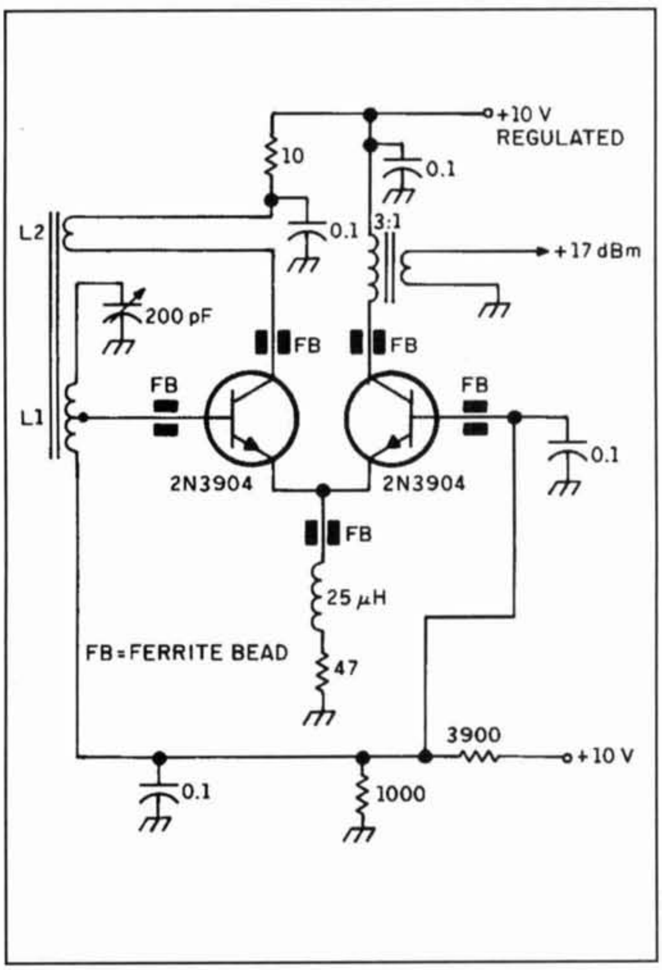

Figure 28. Circuit of the K7HFD low-noise oscillator. L1 is 1.2 µH and uses 17 turns of wire on a T68-6 toroid core. The tap is at 1 turn. Q at 10 MHz is 250. L2 is a 2-turn link over L1.

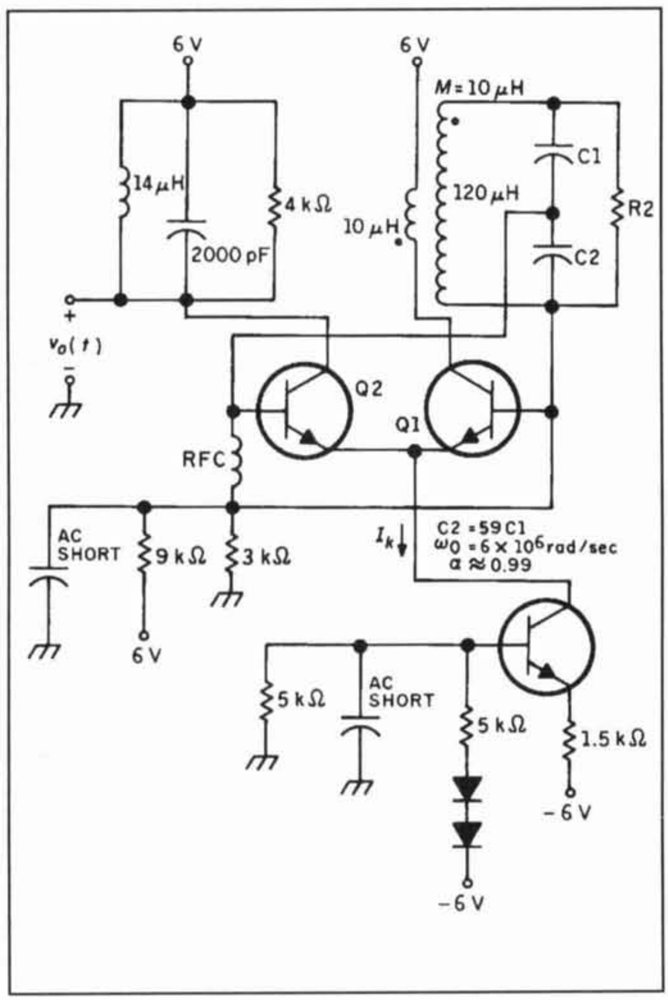

Figure 29. Implementation of the oscillator shown in *Figure 28* using an IC approach and combining of the feedback of both transistors.

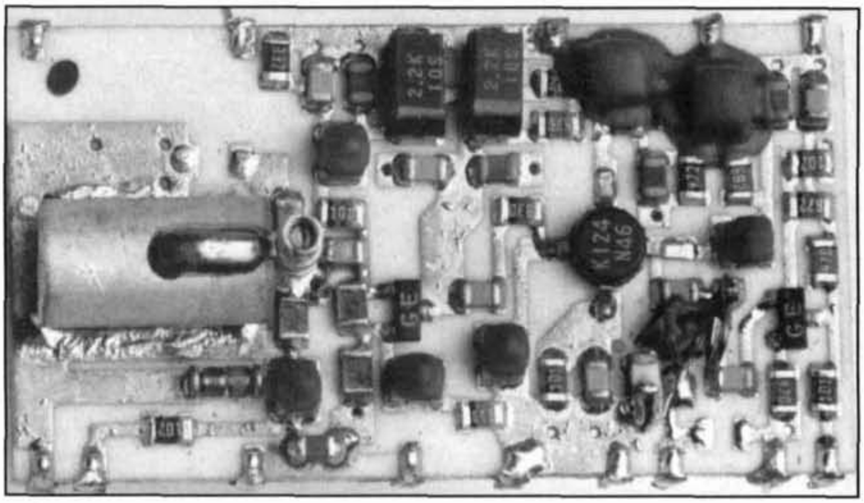

Photo C. Test oscillator with conventional SMD design for 900-MHz hand-held cellular telephone.

and after optimization—of the Motorola 800-MHz transistor oscillator for **Figure 16**. The close-in phase noise at 1 Hz has been improved by approximately 32 dB and even the phase noise of approximately 1 . . . 10 MHz off the carrier was improved. Because of the reduction in amplitude and the resistive feedback, the phase noise at 20 MHz and further away is set at –160 dBc per Hz, while the original circuit predicted a slightly better performance. This technique is applicable for arbitrary topologies of oscillators and active devices for which we have a good understanding of the relationship between the flicker noise and the bias point.

Some other useful oscillator circuits

Figure 35 shows a bipolar transistor oscillator with a frequency range of 2.75 to 3.75

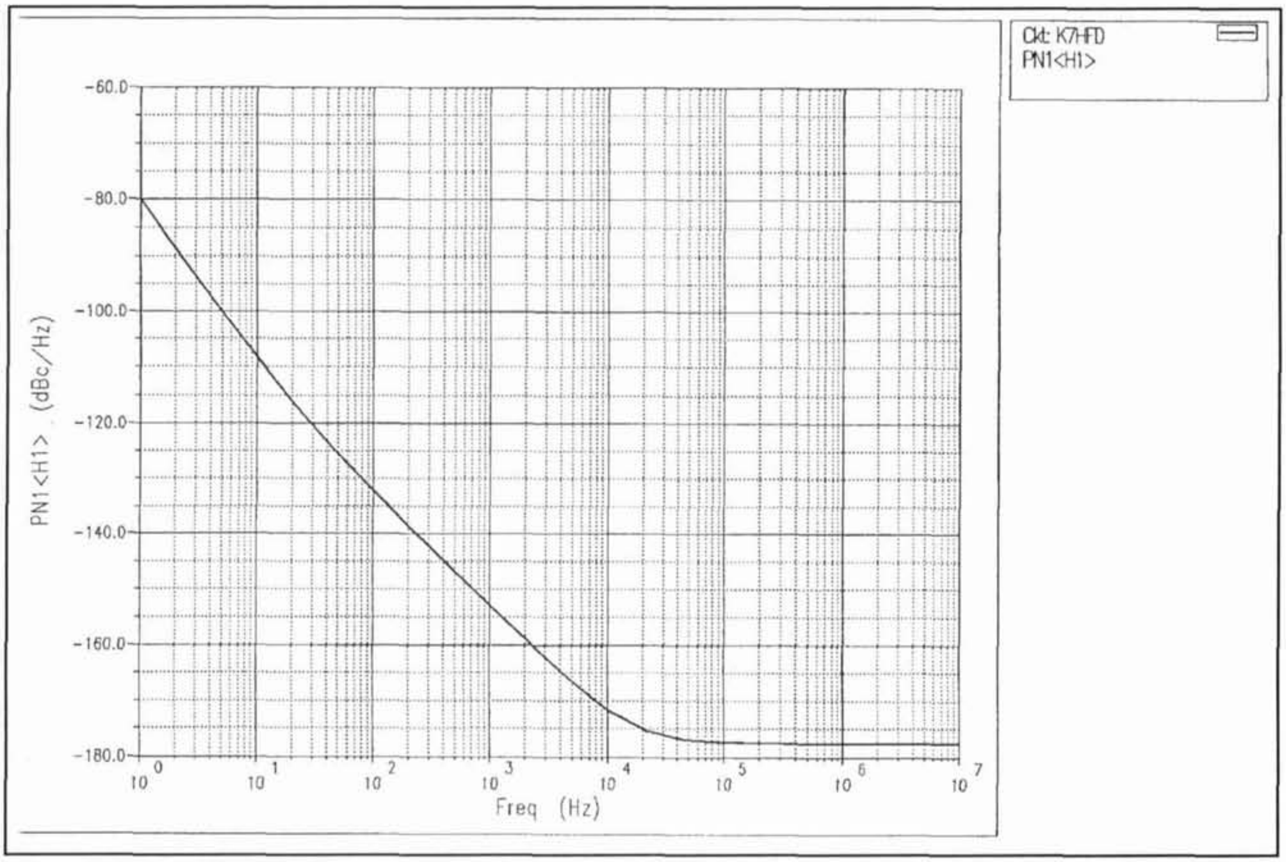

Figure 30. Simulated phase noise of the oscillator shown in *Reference 5*.

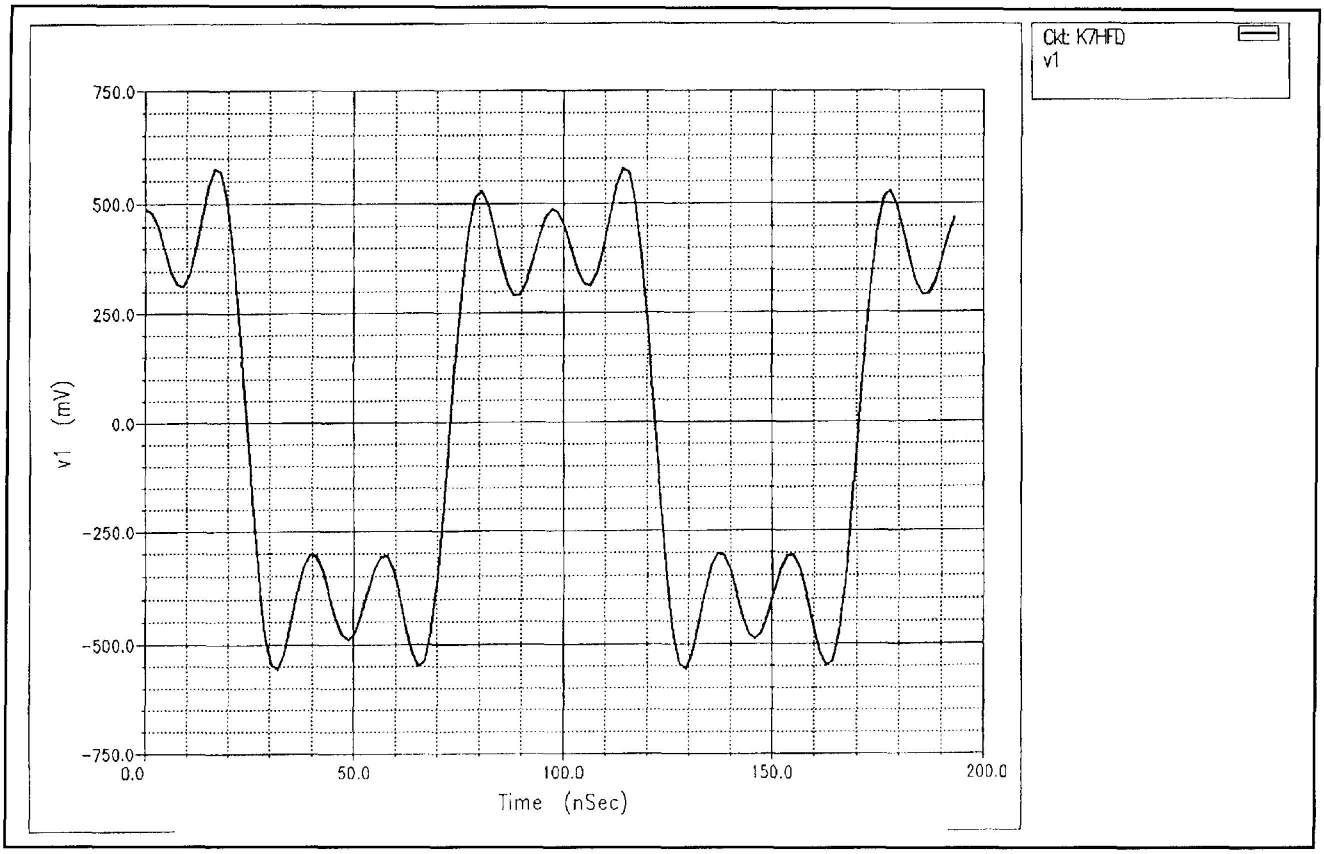

Figure 31. Output waveform for the oscillator shown in *Reference 5*. In both cases, note that the circuit had been modified to incorporate the second transistor in the feedback loop. This improved the signal to noise ratio.

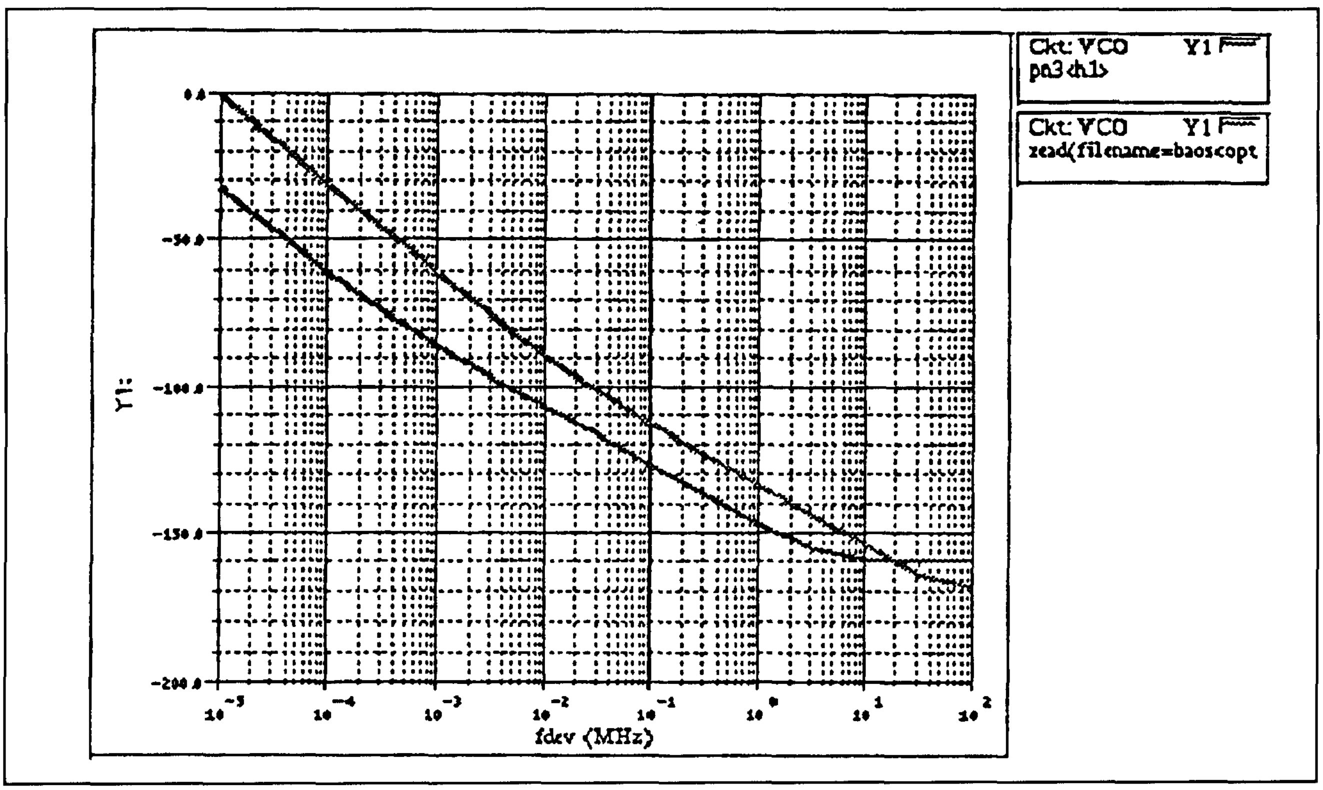

Figure 32. Phase noise of the oscillator in *Figure 16*, before and after optimization.

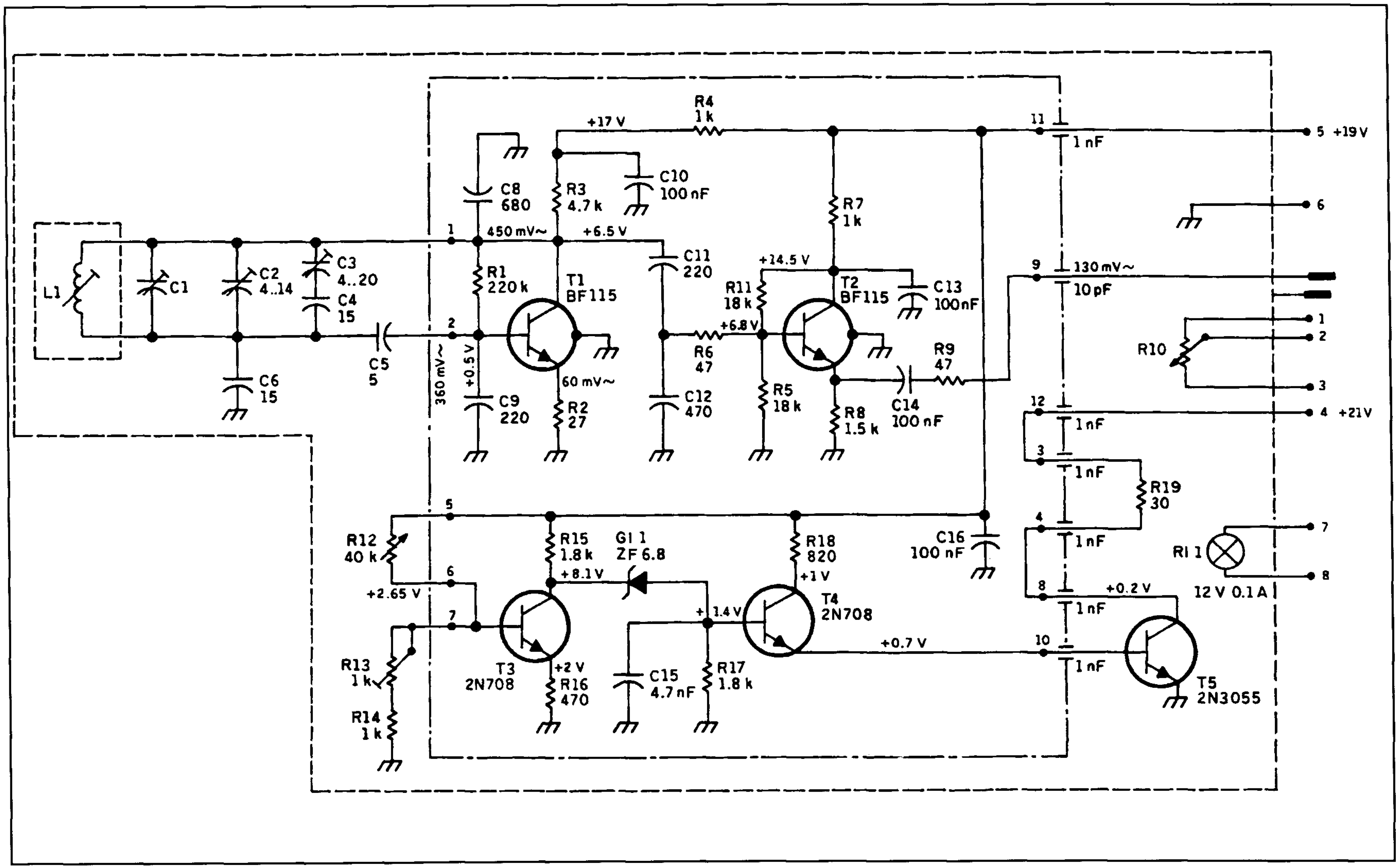

Figure 33. 2.75 to 3.75-MHz high-stability oscillator with proportional controlled oven.

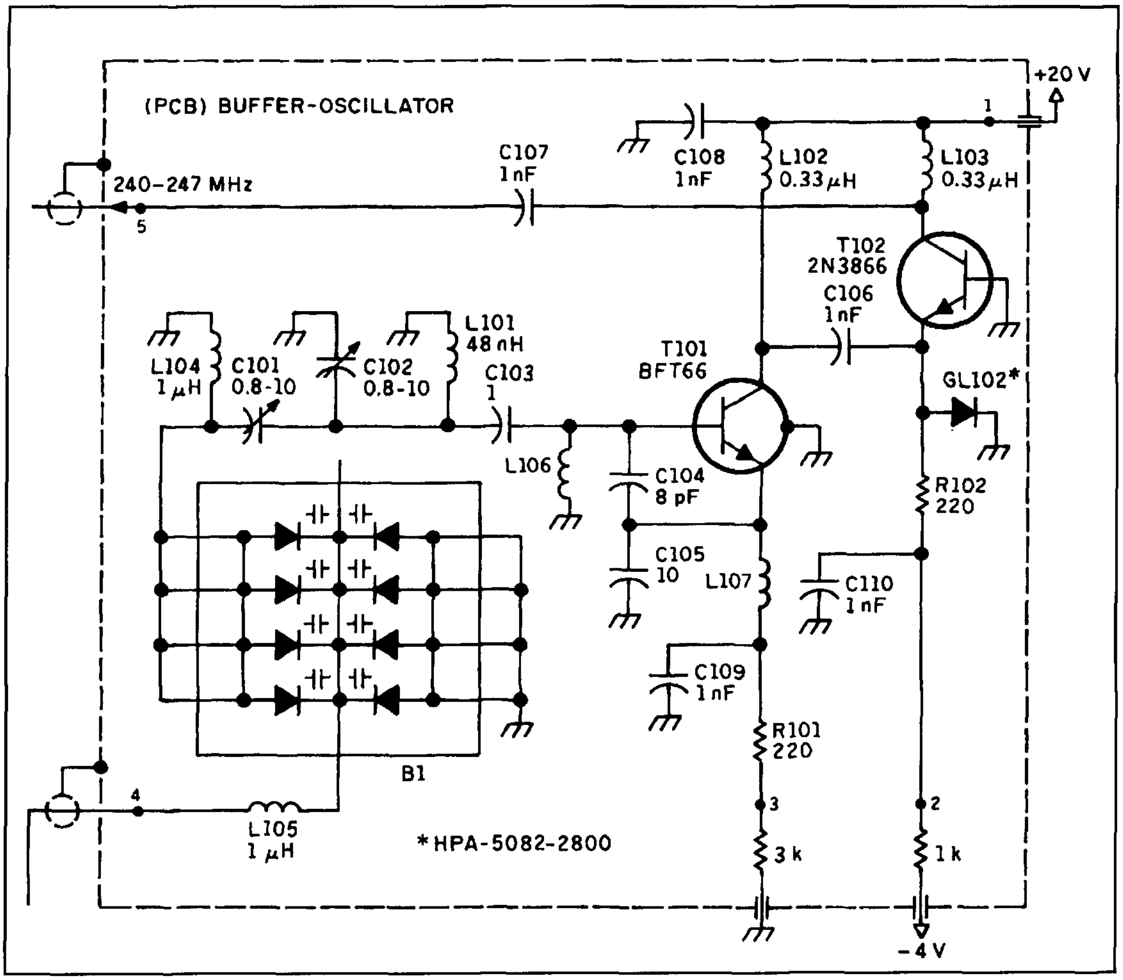

Figure 34. Low phase noise VCO from the Rohde & Schwarz XPC synthesizer, which can be modified for the 220-MHz ham band. It offers extremely low phase noise of 150 dB per Hz, 25 kHz off the carrier.

MHz. This range can be shifted by adjusting L1 and C1. A good application is 5 to 5.5 MHz. This bipolar oscillator inside its oven-controlled environment has a frequency shift of less than 50 Hz per day and, even without the electric heating, won't drive more than 200 Hz per day after 15 minutes warm up.

For those attempting to build low phase noise VCOs for the 200-MHz ham band as shown in **Figure 36**, here's a unique low phase noise oscillator that should be included in the synthesizer. Its phase noise at 200 MHz is 150 dB per Hz, 25 kHz off the carrier.

Finally, for those experiencing noise in their UHF repeater oscillator, the cavity resonator-based low phase noise VCO in **Figure 37** may eliminate some headaches caused by some noisy designs.

Summary

This paper has analyzed both FET and bipolar oscillators, including circuit combinations adapted for high-level integration. The general theory seems to be that FETs require more current, but the f_t of the available N junction FETs isn't sufficient to build oscillators above 400 MHz and the flicker noise contribution is very small. On the other hand, bipolar transistors are ideally suited for oscillator circuits for high integration, and their bias point must be chosen very carefully to avoid too much flicker noise contribution. When these types of oscillators are designed as one terminal oscillator circuits, they do not provide an attractive choice of topologies and, therefore, a good compromise between performance and complexity must be sought. Most hand-held two-way radios or radio telephones with digital signal processing don't require very high performance oscillators. This paper has sought to provide some guidelines for obtaining high performance circuits. We have also addressed AGC action by self-limiting the use of DC feedback and diode

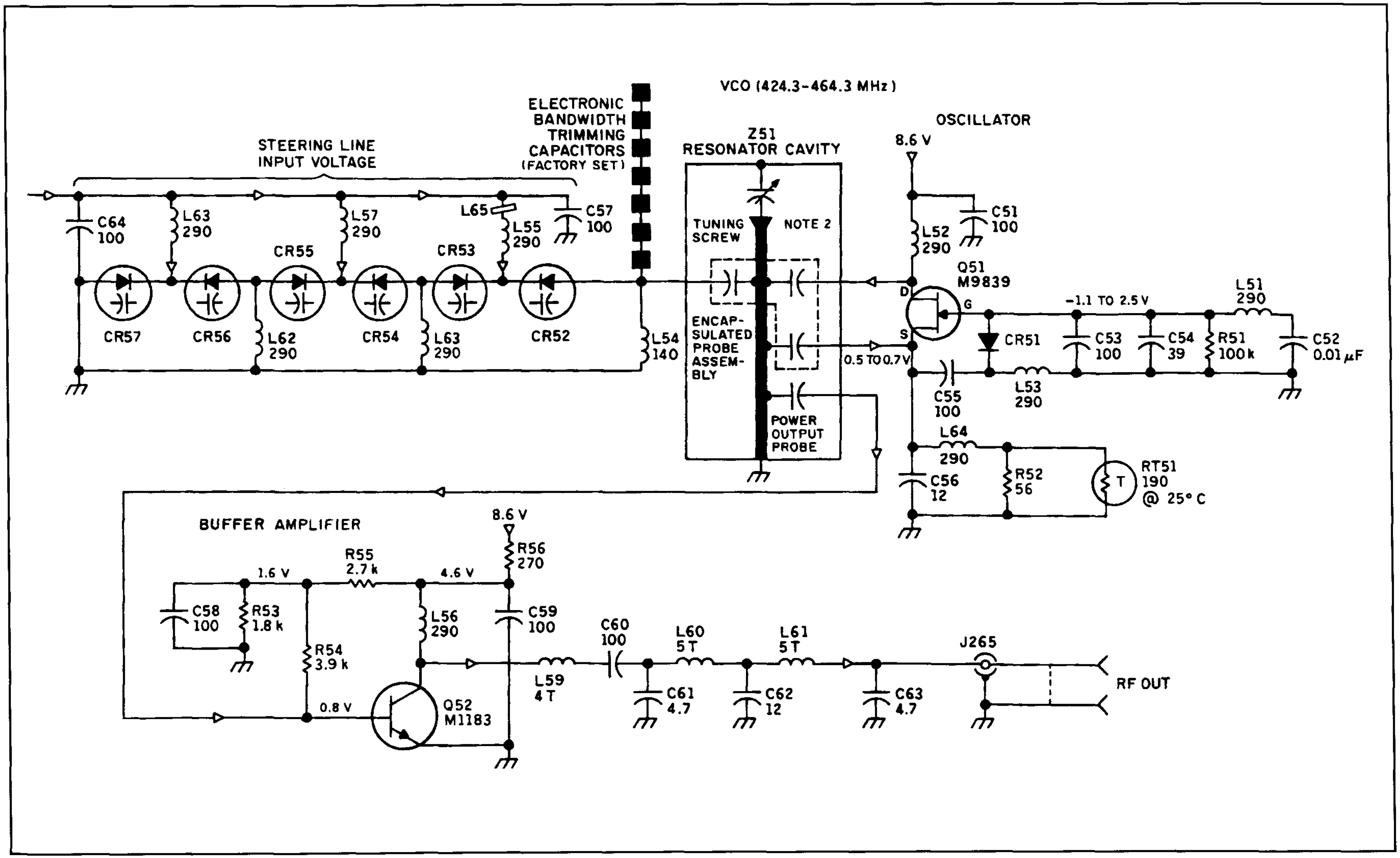

Figure 35. Cavity stabilized VCO for application in UHF repeaters.

clamping. There has been some discussion on these three topics within the engineering community and we have used our tools to provide insight on these sensitive circuits. ■

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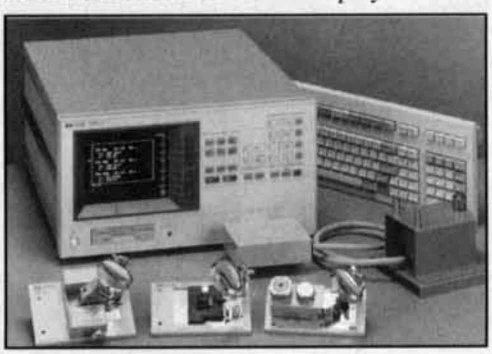
6. Wes Hayward and Doug DeMaw, *Solid-State Design for the Radio Amateur*, American Radio Relay League, Newington, Connecticut, page 126, **Figure 26**.
7. Ulrich L. Rohde, "All About Phase Noise in Oscillators," *QEX*, December 1993, January 1994, and February 1994.
8. Ulrich L. Rohde, "Key Components of Modern Receiver Design, Parts I, II, and III," *QST*, May 1994, June 1994, July 1994, pages 29–32, pages 27–30, and pages 43 and 45, respectively.

PRODUCT INFORMATION

New 1-GHz LCR Meter Applies New Measurement Technique

Hewlett-Packard Company has announced their RF LCR meter with a frequency range of 1 GHz. An RF LCR meter is an instrument that measures inductance, capacitance and resistance. This meter applies a new measurement technique to test components at their actual operating conditions, allowing circuit designers to select the right components for their designs. The HP 4286A builds on and expands HP's precision LCR meter family, the HP 4284A and HP 4285A. The newest meter provides a component test solution for the fast-growing RF industry. The LCR meter offers 1 MHz to 1 GHz with 10-kHz resolution; a test signal of 10 mV to 1V; a measurement speed of 15 msec; and has a monochrome CRT display.



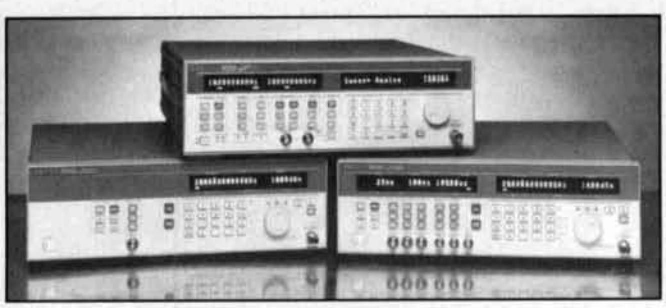
The HP 4286A RF LCR meter is available now with delivery six weeks from receipt of order; and can be purchased for $27,600. For more information, contact Hewlett-Packard Company, Direct Marketing Organization, P.O. Box 58059, MS51L-SJ, Santa Clara, CA 95051-8059.

HP Enhances Synthesized Microwave Sources

Hewlett-Packard Company has added options to the HP 8370 family of synthesized microwave sources and improved the standard instrument performance at no increase in price. Enhancements include optional analog-phase-modulation capability; optional internal modulation generator for amplitude modulation (AM), frequency modulation (FM) and phase modulation; optional linear AM; and increased output-power range from –90 dB to –110 dB for the optional step attenuator. This product family now provides improved phase noise and seven convenient FM sensitivity ranges.

The HP 8370B product family can be configured with options that deliver powerful signal simulation tailored for specific applications in the aerospace, defense, satellite, microwave-communications and intelligent-transportation-system industries. The products include the HP 83710B series of CW generators for frequency-converter and exciter applications, and the HP 83730B series of synthesized signal generators with signal-simulation capability for receiver-test applications. In addition, HP offers the HP 83750B series of synthesized sweepers designed for component test.

For more information, write to Hewlett-Packard Company, Direct Marketing Organization, P.O. Box 58059, MS51L-SJ, Santa Clara, CA 95051-8059.



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QUARTERLY REVIEW

NEC-WIN BASIC For Windows

Paragon Technology has adapted its engineering experience with NEC and the optimization of antenna design to the Windows platform. In the process, it has developed a near-entry-level program for antenna modelers, NEC-Win Basic, priced within reach of students and amateur radio operators.

The Windows platform offers users some facilities not easily implemented in DOS. For example, one can produce presentation-quality graphic pattern plots as a matter of course (see **Figure 1**). Those who prefer pull-down menus

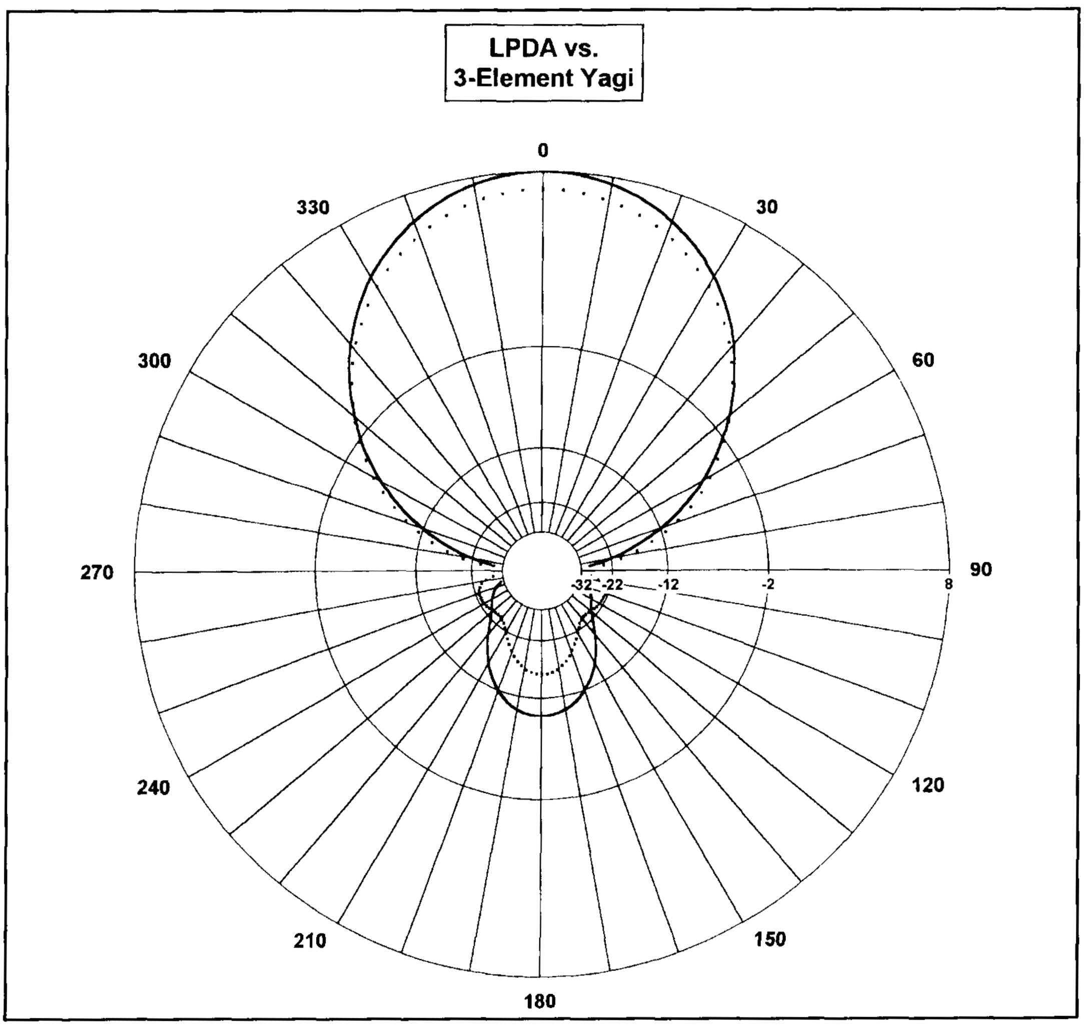

Figure 1. A sample NEC-Win Basic plot comparing free-space azimuth patterns for a log-periodic antenna and a 3-element Yagi beam on a log scale.

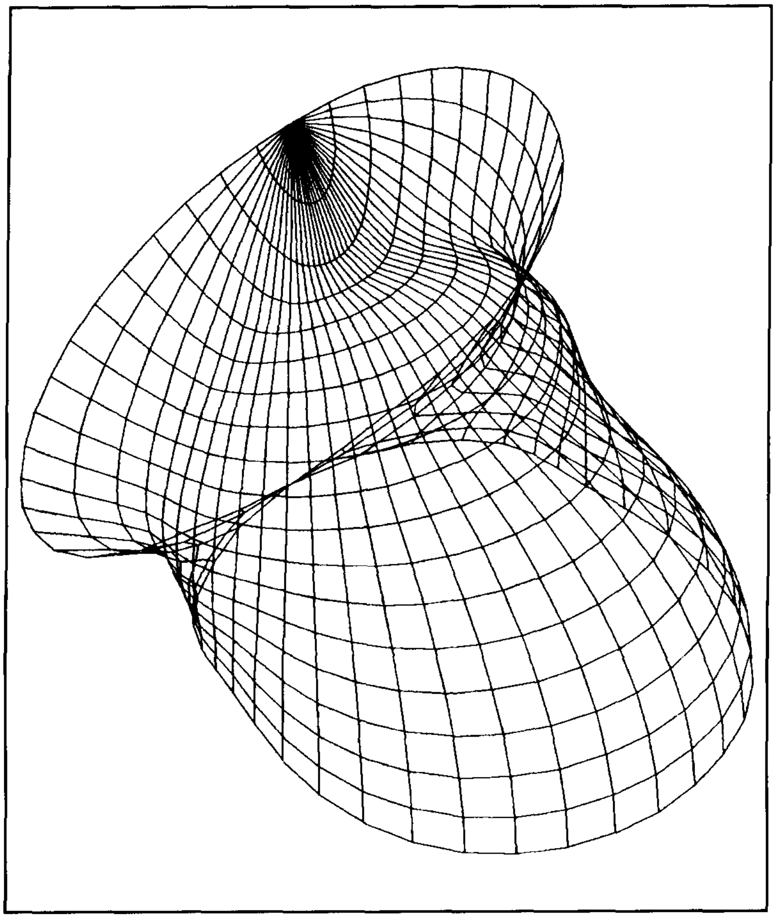

Figure 2. The upper half of the rear lobe of a log periodic antenna as viewed from below, with potentially confusing lines from the opposite half removed for clarity of detail.

(including cut, copy, paste, rotate, translate, and scale operations) along with mouse navigation through the program elements will find that the program has similarities to a spreadsheet with graphic outputs—interrupted, of course, by the engagement of the NEC-2 calculation engine.

NEC-Win Basic tries to retain, in a fundamental program, much of the flexibility of design decision to which engineers are accustomed. For example, there are 9 preset choices of antenna materials and two spaces for user specifications. There are some 13 categories of printable output data, much of which is in engineering notation. The user can specify loads as either a set of RLC values or as a combination of resistance and reactance. Even the range of azimuth and elevation coverage can be specified, both for the calculations and for pattern plots.

Such flexibility allows the user to develop some precise tools for analysis. For example, NEC-Win Basic offers a graphical surface plot of far-field gain values (which, like the graphical view of the antenna, appears in DOS-based graphics). By selectively setting the limits of the patterns for both azimuth and elevation, the user can focus in on small portions of the surface plot, freed from the confusion created by lines on the less interesting opposing surface (see **Figure 2**). Likewise, one can judiciously select for screen display or printing from within the program the most useful data from the calculations.

The wide range of open-ended options offered by the program places responsibilities on the user. Once the input file is created by a combination of spread-sheet-style wire entry and pull-down-menu entry of ground, frequency, element conductor, transmission line(s), source, and load values, nothing happens without user decision. The user must decide what he or she wants by way of result, and this requires some understanding in advance of what is most likely to be of significance. Paragon has produced a well-constructed manual combining introductory and advanced walk-through exercises with a feature-by-feature reference section. However, to use the program comfortably to its fullest capabilities, the user should combine his or her growing experience with a wider reading into the NEC-2 program and into antenna modeling in general.

Although Paragon recommends a 4 MB RAM minimum to run the program, excessive TSR storage in upper RAM may make 8 MB a better minimum to meet the NEC-2 need for expanded memory pages. Laser printers should have 1 MB or more of memory (standard for the present and immediate past generations of printers) for handling both the word and pictorial elements of plots. The calculation portion of the program is highly interactive with the hard drive: hence, both hard disk and CPU speeds are relevant to using the program efficiently. Available Pentium and DX CPUs and current generation drives are well up to the task.

NEC-Win Basic is available from Paragon Technologies, 200 Innovation Blvd., State College, PA 16803. The price is $75 plus shipping and handling. ■

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Fort Ethan Allen
Colchester, Vermont 05446

SIMPLE AND INEXPENSIVE HIGH-EFFICIENCY POWER AMPLIFIER

For 160 to 40 meters

I've been designing RF power amplifiers and transmitters since shortly after receiving my novice license in 1961. As a high-school student, my budget was rather limited and the industry-standard 6146 power tube at $5 was out of the question. Fortunately for me, government surplus 1625 tubes (75 watt) were available for about 19 cents each.

Unfortunately, modern RF-power MOSFETs cost about $1 per watt. The RF-power transistors in a typical transmitter can cost over $100 and can easily be damaged by static or over-

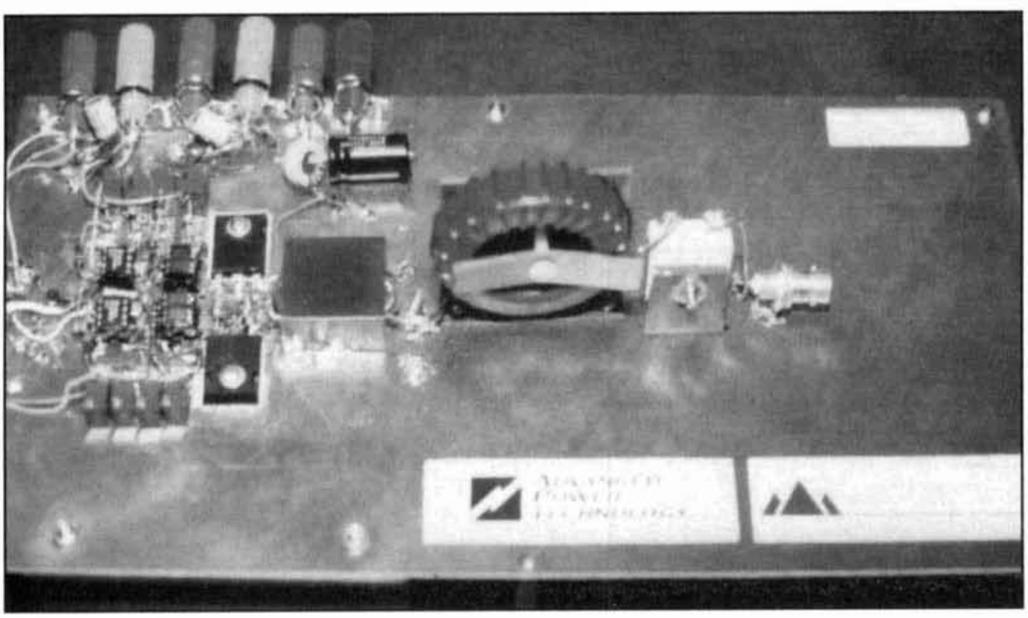

Photo A. Breadboard circuit. Note: Breadboard includes additional complementary MOSFET gate drivers

load. It's not hard to see why few of today's hams design their own power amplifiers.

I'll describe a simple, inexpensive, and efficient 250-watt transmitter for 160 to 40 meters (**Photo A**). Its key features are:

- driver based on two low-cost ICs,
- final amplifier based on two low-cost RF-power MOSFETS,
- simple output transformer, and
- Class D operation.

The PA and driver operate from +12 and +50 volts and require an RF input of only 10 mW. The APT RF-power MOSFETs cost only about $15 each and the whole breadboard circuit only about $150. This should make it possible for today's hams to once again experiment with RF-power circuits.

Basic design considerations

Class D power amplifiers use two transistors in a push-pull configuration. The transistors are driven to act as switches and generate a square-wave voltage. The fundamental frequency component of the square wave is passed to the load through a filter. Power output is controlled by varying the supply voltage.

A Class D PA is ideally 100 percent efficient at all amplitudes and in spite of load reactance. In practice, it is significantly more efficient than a similar Class B PA—especially for lower amplitudes and reactive loads.

Drain-load line and turns ratio

The power output of a Class D PA[1,2] is:

$$P_0 = \frac{8}{\pi^2} \frac{V_{\mathrm{eff}}}{R} \tag{1}$$

where the effective supply voltage (for MOSFETs) is:

$$V_{\mathrm{eff}} = V_{DD} \frac{R}{R + R_{\mathrm{on}}} \tag{2}$$

Above, R is the drain-load line (seen by one drain with the other open) and R_{on} is the on-state drain-source resistance. For the ARF440 and the ARF441, $R_{\mathrm{on}} \approx 0.8$ ohms. Thus, a 250-watt output with a 50-volt supply requires $R \leq 6.4$ ohms.

For a simple transformer, the drain-load line is related to the load R_o by:

$$R = (m/n)^2 R_0 \tag{3}$$

where m and n are the numbers of turns in the primary and secondary windings, respectively. For an RF transformer, m and n must be small integers. For a 50-ohm load, a bit of iteration yields $m = 1$, $n = 3$, and $R = 5.56$ ohms.

In the absence of transformer, switching, and capacitance losses, the efficiency of the PA is:

$$\eta = V_{\mathrm{eff}} / V_{DD} = 0.874 \tag{4}$$

The effects of switching and drain capacitance can be predicted by the equations given in **Reference 2**. These calculations yield an 82-percent efficiency for operation at 7 MHz, excluding losses in the transformer and output filter. With typical losses in those components, an efficiency of 70 percent is expected

Gate voltage and current

The peak drain current[1,2] is:

$$i_{D\max} = \frac{4}{\pi} \frac{V_{\mathrm{eff}}}{R} = 10.01 \text{ A} \tag{5}$$

which corresponds to $I_{dc} = 6.37$ A. The data sheets for the ARF440/ARF441 show that a gate-source voltage of 9 to 10 volts should be sufficient for minimum R_{on} at $I_{D\max} = 10$ A. Because the threshold voltage is about 3.5 volts, the RF-drive voltage must be about 6.5 volts or slightly more.

The data sheets provide typical small-signal active-region gate-source and gate-drain capacitances of 730 and 77 pF, respectively. The input capacitance is, however, the sum of small-signal capacitances only while the MOSFET is in the cut-off region. As the MOSFET enters the active (velocity-saturation) region, the gate-drain capacitance is Miller-multiplied by the voltage gain (plus one). Finally, as the MOSFET enters saturation (resistive region), the gate-drain capacitance inflates by a factor of five, or so. As a result, the effective gate-drain capacitance is close to 2600 pF during the switching process.[3]

The total gate charge is 37 nC at 10 volts. Transitioning the gate voltage in one tenth of the RF period (73.7 ns at 13.56 MHz) requires a peak gate current of 5 A. A low driving resistance and low inductance are clearly required.

Circuit

Figure 1 shows the power amplifier and driver circuit; parts are identified in **Tables 1** and **2**. All stages are ac coupled. Consequently, the unexpected absence of an input signal results

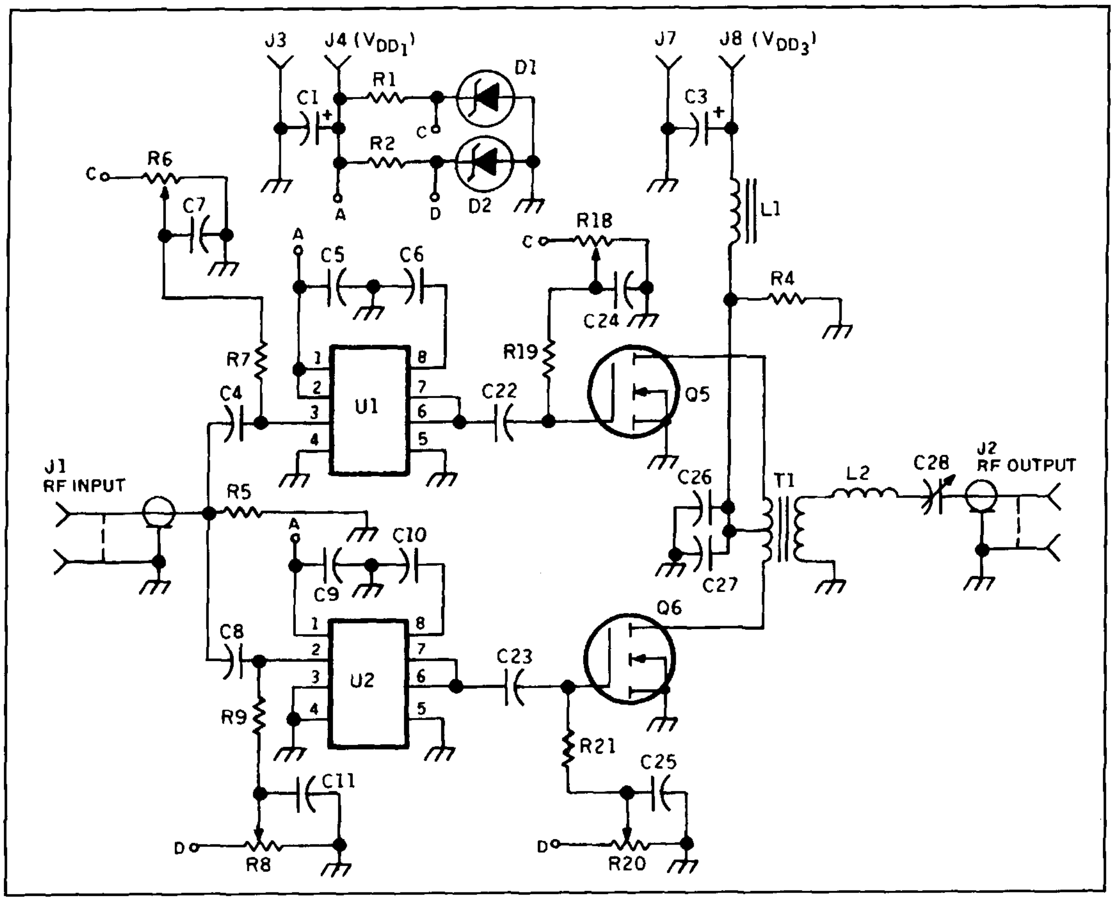

Figure 1. Circuit of power amplifier and driver.

only in an inactive circuit with minimal current flow and power consumption.

Driver

The driver uses a pair of low-cost ICs (U1 and U2 in **Figure 1**) rather than the conventional RF transformer to provide out-of-phase driving signals for the two final MOSFETs. It also provides hard limiting (sine-wave to square-wave conversion) of the input signal.

The Elantec EL7144C is intended for use as a gate driver. The internal Schmidt trigger allows it to serve as hard limiter, and the presence of both inverting and noninverting inputs allows a pair to serve as a phase splitter.

The RF input is ac coupled to the noninverting input of U1 and the inverting input of U2. Adjustment of the biases via R6 and R8 allows the transition points to be selected to produce the desired duty ratio (50:50). The phase error between the two EL7144s is about 0.5 ns. (If an oscilloscope isn't available, use a voltmeter with a 1-k series resistor and set the average output voltage to 6 volts.)

The EL7144s have high input impedances, so R5 provides a 50-ohm input impedance for the signal source. Input signals in the range of 10

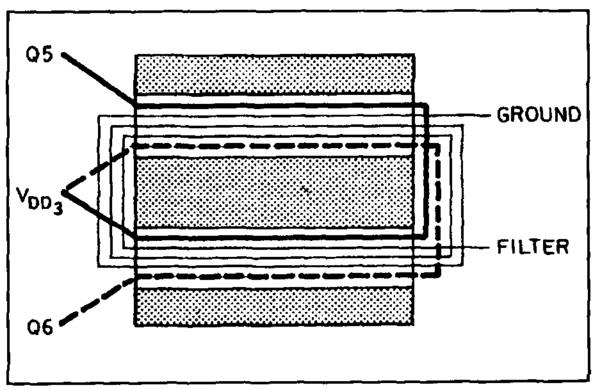

Figure 2. Construction of output transformer T1.

to 100 mW are satisfactory, allowing this PA to be driven directly from a laboratory signal generator or oscillator with buffer.

The best switching speed is obtained with $V_{DD1} \approx 12$ volts.

Final amplifier

The final amplifier is based upon the ARF440 and ARF441 symmetric-pair MOSFETs. The quiescent currents are individually adjustable via R18 and R20 and should be

Reference Designator	Part Description
C1	33-µF, 50-V electrolytic, Mallory
SKR330M1HE11V	
C3	20-µF, 250-V electrolytic, Mallory TT250M20A
C4-C27	0.1-µF, 50-WV chip, ATC 200B104NP50X
C28	See Table 2.
D1, D2	5.1-V, 0.25-W Zener, 1N751A
J1, J2	BNC femal bulkhead
	Recommended: Amphenol 31-5538
	Used: RF Industries RFB-1116S/UG
J3, J4, J7, J8	European-style binding post, Johnson 111-0104
L1	3.5-µH, 7 turns #24 enameled wire on Ferroxcube
L2	768XT188 4C4 toroid. See Table 2.
Q5	*n*-channel MOSFET, APT ARF440
Q6	*n*-channel MOSFET, APT ARF441
R1, R2	330-Ω RC07
R4	10-kΩ RC07
R5	51-Ω RC07
R6, R8, R18, R20	1-kΩ trimpot, Bournes 3299X-1-102
R7, R9, R19, R21	4.7-kΩ RC07
T1	One Ceramic Magnetics 3000-4-CMD5005 block of CMD5005 ferrite wired per Figure 3
U1, U2	Schmidt trigger/limiter, Elantec EL7144C
Heatsink for DIP IC	Aavid 5802 clip-on
Heatsink for Q5 and Q6	Aavid 61475, 6.5-in wide by 4-in long
IC socket, 8-pin DIP (4)	Augat 208-AG190C
PC board	Approximately 6.5-in wide by 8-in long
Plastic bracket for L2	Cut from plastic L
Plastic bracket for C28	Cut from plastic L
Feet (6)	Aluminum threaded spacer, 4-40 x 2 in (Keystone 2205)

Table 1. Components other than tuning

set to about 0.1 A each (i.e., just on the verge of conduction). This results in gate bias of about 3.5 to 3.8 volts.

The variable gate biases provide a convenient means of checking final MOSFETs Q5 and Q6 during development. The ac coupling also provides some isolation of the driver MOSFETs in case there's a short-circuit failure of the final MOSFETs. It is, however, possible to simplify the circuit by direct-coupling the output of the driver to the gate of the final.

Output transformer T1 is made by winding no. 22 insulated wire through one block of CMD5005 ferrite as shown in **Figure 2** The use of different colors facilitates identification of the leads. The VSWR and overrating factor are less than 1.9 and 1.5, respectively, from 1 to 14 MHz. Capacitors C26 and C27 ensure a good RF ground at the center of the primary winding. Inductor L1 keeps RF out of the power-supply line; its value isn't critical.

The prototype uses simple series-tuned circuits (**Table 2**) with $Q = 5$ at the frequency of operation. Tuning is accomplished by adjusting padder C28 for maximum output power or maximum dc-input current. Other output filters or matching networks can be used provided they include a series inductor on the transformer side to prevent current from flowing at the harmonic frequencies.

Layout

The general layout of the principal components is shown in **Figure 3**. MOSFETs Q5 and Q6 are separated by about 0.8 inch, so their drain leads are aligned with the leads from T1. The "symmetric-pair" packaging conveniently places both drain leads on the right and both gate leads on the left. The driver ICs are installed roughly in line with the leads from Q5 and Q6.

The prototype breadboard is constructed on a 6.5 by 8-inch piece of pc board. All wiring is done on or above the surface so the lower ground plane is unbroken.

For convenience in experimentation, I socketed U1 and U2. Socketing is not, however, recommended for a final layout as it adds inductance between the devices.

The input to U1 and U2 is high-impedance and therefore not especially critical. Install chip capacitors C5, C6, C9, and C10 as close as possible to the power-supply leads. Place a clip-on heat sink over each IC to provide adequate heat dissipation. Connect the driver outputs to the final gates through short, wide (low-inductance) traces.

Mount the heatsink flush with the bottom of the pc board. Cut holes for Q5 and Q6, but otherwise *do not* interrupt the integrity of the ground plane. Cut connection pads near the gates and drains. Trim the leads of Q5 and Q6 at the point where they narrow and then bend them downward slightly to contact the pc board. Use thermal grease on the bottoms of Q5 and Q6 to ensure adequate conduction of dissipated power into the heatsink.

Output transformer T1 lies flat on the pc board and is held in place adequately by its leads. Tuning elements L2 and C28 are supported on plastic L brackets.

Performance

The waveforms for operation at 1.8 MHz with $V_{DD} = 20$ volts are shown in **Figure 4**. They include the final gate v_{GS}, final drain v_{DS}, transformer output v_s, and filter output v_o.

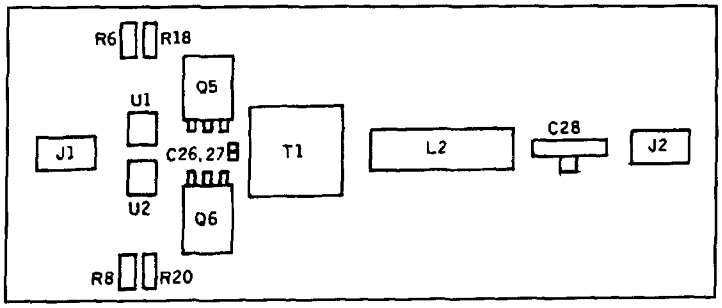

Figure 3. General layout of amplifier.

Turn on begins shortly after the gate voltage begins to rise, and flattens the driving waveform because of the Miller effect and inflation of the gate-drain capacitance. The transients in v_{GS} are probably due to the length of wire connecting the driver and gate, and/or the voltage that is developed across the internal inductances of the MOSFETs.

The transients in the drain voltage are the normal consequence of the MOSFET switching faster than the rise time of the transformer. The transient frequency of 56 MHz corresponds roughly to the resonance of the 193-pF typical drain capacitance and the measured 40 nH of transformer inductance. The peak voltage remains well within the ratings of the ARF440 and the ARF441. The transient is prevented from reaching the load by the output filter.

The measured efficiency and power output are shown in **Figure 5** and **Table 3** as functions of frequency. The efficiency is based upon measured RF output and dc input to the final amplifier (Q5 and Q6), and includes *all* losses. For 160 meters, the power output and efficiency are in excellent agreement with the predictions—especially given the addition of a few percent of loss in the transformer and filter. As frequency increases, the efficiency drops off faster than expected. For 40-meter operation, V_{DD} should be reduced slightly to maintain safe drain currents ($i_{D\max} \leq 11$ A, $I_{dc} \leq 7$ A).

1.8 MHz	L2:	22 µH, Micrometals T200-6 toroid (2-in O.D.) with 52 turns of 24-AWG enameled wire
	C28:	354 pF, 2.5-kV padder, FW Capacitors AP091HV
3.5 MHz	L2:	11.4 µH, Micrometals T200-6 toroid (2-in O.D.) with 32 turns of 24-AWG enameled wire
	C28:	180 pF, 2.5-kV padder, FW Capacitors AP051HV
7 MHz	L2:	5.7 µH, Micrometals T200-2 toroid (2-in O.D.) with 20 turns of 24-AWG enameled wire
	C28:	90 pF, 2.5-kV padder, FW Capacitors AP031HV

Table 2. Tuning components.

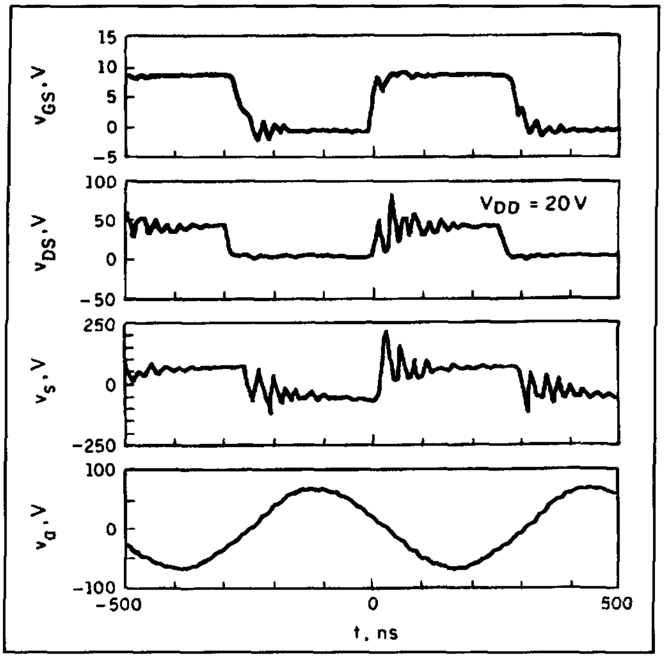

Figure 4. Waveforms for operation on 160 meters.

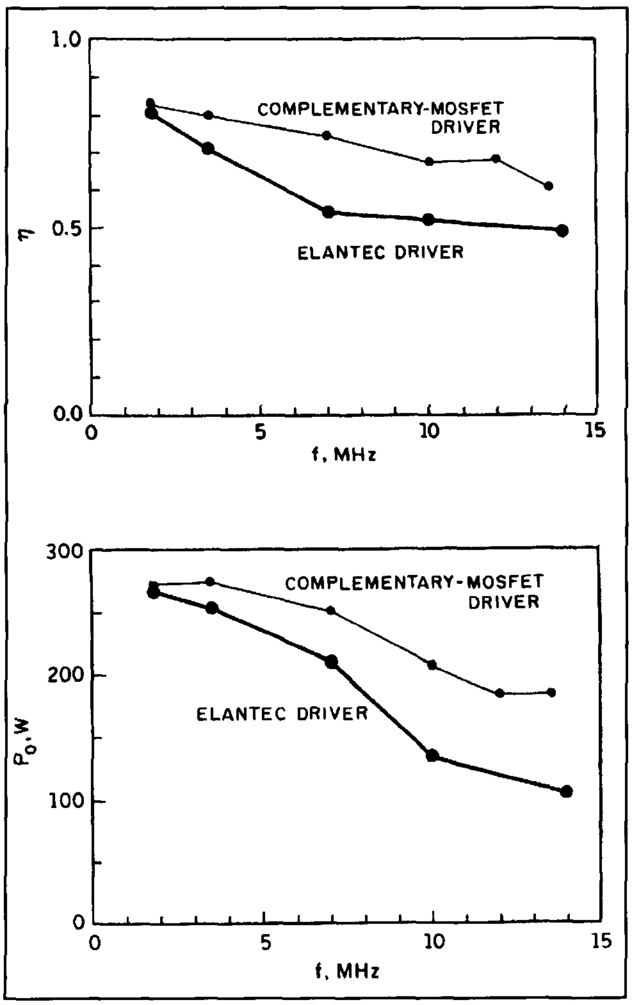

Figure 5. Efficiency and output versus frequency.

The faster-than-expected decrease in efficiency occurs because both MOSFETs are on simultaneously for a short period of time; this, in turn, is because the turn-off time is greater than the turn-on time.[4] This problem can be eliminated by adding a complementary MOSFET driver as discussed in the next section.

The variations of efficiency and output voltage with supply voltage are shown in **Figure 6** and **Table 3**. In contrast to Class A and Class B PAs, the efficiency remains relatively constant and high for all output levels. The highest efficiency generally occurs at mid-range supply voltages because the drain capacitance is reduced but the current isn't yet high enough to increase R_{on}. The linearity for control and modulation is quite good.

The total driver current varies from 110 mA on 160 meters to 270 mA on 40 meters. This means the total driver input power varies from 1.3 to 3.2 watts.

Completing the transmitter

To make the power amplifier and driver into a transmitter, add the following:

- power supply,
- signal source (VFO),
- keying and/or modulation, and,
- output filter.

Since the Elantecs can be driven by a +10 dBm input signal, almost any VFO with an output buffer can serve as the signal source. A variety of suitable VFO and power supply circuits can be found in **Reference 5**.

The power amplifier and driver can be keyed for CW transmission by interruption of either the RF drive or the dc supply. Amplitude modulation is very linear (**Figure 6**) and can be accomplished efficiently by a Class S modulator.[6,7] Production of a single sideband and other complex signals can be implemented using the Kahn envelope-elimination-and-restoration (EER) technique.[8,9]

On-the-air use of the PA requires more harmonic suppression than the simple series-tuned circuits of **Figure 1** can provide. A three-section bandpass filter or five-section T-type should suffice. A number of designs are given in **Reference 5**. Remember that Class D operation requires the input side of the filter to include a series inductor or series-resonant tank to prevent harmonic currents.

Operation can be extended to 30 and 20 meters by inserting complementary MOSFET gate drivers between the Elantecs and the APT MOSFETs.[4,10] Because the turn-off delay of the APT MOSFET is slightly larger than the turn-on time, operation at these frequencies requires the

V_{DD}, V	1.8 MHz		3.5 MHz		7.0 MHz	
	V_{om}, V	η	V_{om}, V	η	V_{om}, V	η
10	37.9	0.802	38.2	0.723	36.0	0.574
20	68.1	0.820	68.4	0.744	68.4	0.581
30	101.6	0.825	101.4	0.741	99.6	0.604
40	133.9	0.825	133.0	0.748	126.8	0.593
45	149.3	0.820	147.0	0.734		
50	164.1	0.813	159.7	0.716	145.3	0.545

V_{DD}, V	10 MHz		14 MHz	
	V_{om}, V	η	V_{om}, V	η
50	136.3	0.526	107.5	0.500

Table 3. Output voltage and efficiency as functions of supply voltage.

Elantecs to be biased to produce slightly less than a 50:50 duty ratio. This addition to the circuit also improves the power output and efficiency on 80 and 40 meters (**Figure 5**).

Use of the ARF442 and ARF443 MOSFETs and a 100-volt supply should allow the output to be increased to about 400 watts.

Acknowledgment

The PA and driver were designed under contract 409-1215 from Advanced Power Devices.

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7. F.H. Raab and D.J. Rupp, "High-efficiency Amplitude Modulator," *RF Expo East '94*, Orlando, Florida, pages 1–9, November 15–17, 1994.
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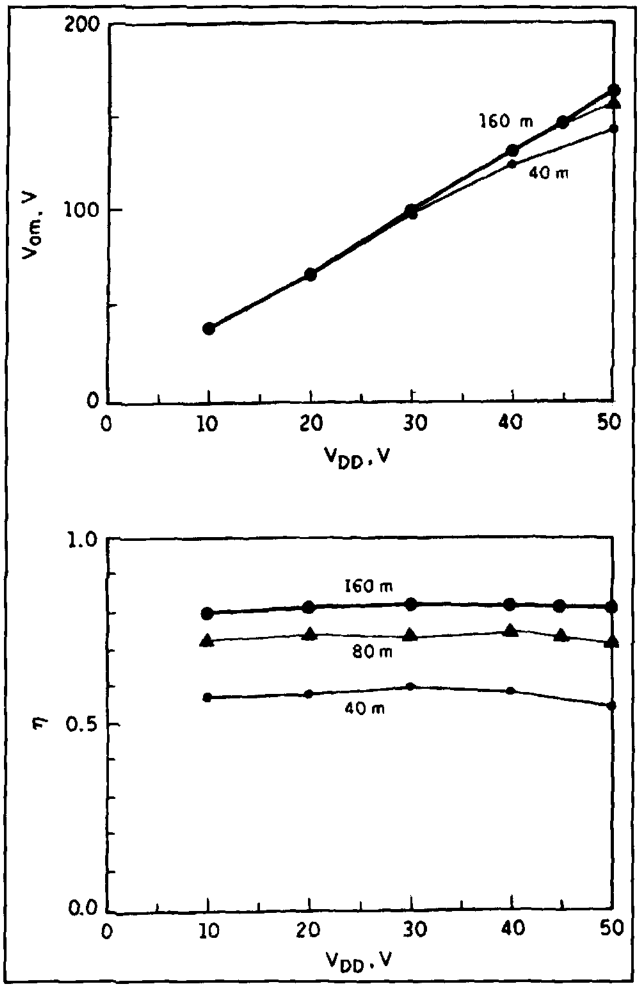

Figure 6. Efficiency and output versus supply voltage.

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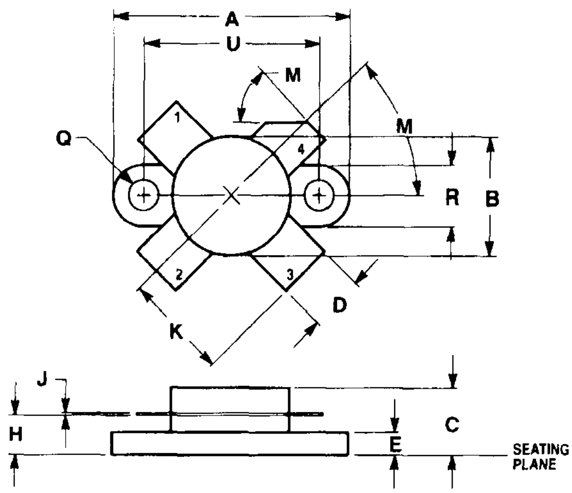
QUARTERLY DEVICES

The MRF-255 RF Power Field-Effect Transistor and Digi-Key's Panasonic Multilayer Ceramic Chip Capacitor Kits

RF-power MOSFETs first appeared as an alternative to bipolar junction transistors (BJTs) in the mid-1970s. My first introduction to Motorola TMOS devices dates back nearly 12 years, when I built a small 30-watt SSB transceiver around a MRF-138.[1] However, despite many inherent advantages, the 28-volt Vdd requirement for early TMOS devices prevented them from gaining popularity in the 12-volt world of amateur and land-mobile applications. For years, unofficial rumors persisted that a low-voltage linear device would be forthcoming, but none appeared. As it turned out, this probably had more to do with the small size of the U.S. manufacturing market than with any particular roadblock in MOSFET technology.

In the end, the Japanese amateur-radio manufacturing industry provided the stimulus needed to bring a viable 12.5-volt FET to market. According to Mike Civiello, Motorola's sales representative for RF products in Japan, the MRF-255 was developed especially for ICOM to replace BJTs traditionally used in the RF-power stages of amateur multi-band transceivers. MOSFETS generally perform better in extremely wideband amplifier designs, and this characteristic fit in well with ICOM's plans to build extended frequency-coverage multimode transceivers. The first radio to use the Motorola



NOTES:
1. DIMENSIONING AND TOLERANCING PER ANSI Y14.5M, 1982.
2. CONTROLLING DIMENSION: INCH.

DIM	INCHES MIN	INCHES MAX	MILLIMETERS MIN	MILLIMETERS MAX
A	0.960	0.990	24.39	25.14
B	0.465	0.510	11.82	12.95
C	0.229	0.275	5.82	6.98
D	0.216	0.235	5.49	5.96
E	0.084	0.110	2.14	2.79
H	0.144	0.178	3.66	4.52
J	0.003	0.007	0.08	0.17
K	0.435	—	11.05	—
M	45 °NOM		45 °NOM	
Q	0.115	0.130	2.93	3.30
R	0.246	0.255	6.25	6.47
U	0.720	0.730	18.29	18.54

STYLE 2:
PIN 1. SOURCE
2. GATE
3. SOURCE
4. DRAIN

Figure 1. Case dimensions for the Motorola MRF255.

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Drain–Source Voltage	V_{DSS}	36	Vdc
Drain–Gate Voltage (R_{GS} = 1.0 MΩ)	V_{DGR}	36	Vdc
Gate–Source Voltage	V_{GS}	±20	Vdc
Drain Current — Continuous	I_D	22	Adc
Total Device Dissipation @ T_C = 25°C Derate above 25°C	P_D	175 1.0	Watts W/°C
Storage Temperature Range	T_{stg}	−65 to +150	°C
Operating Junction Temperature	T_J	200	°C

THERMAL CHARACTERISTICS

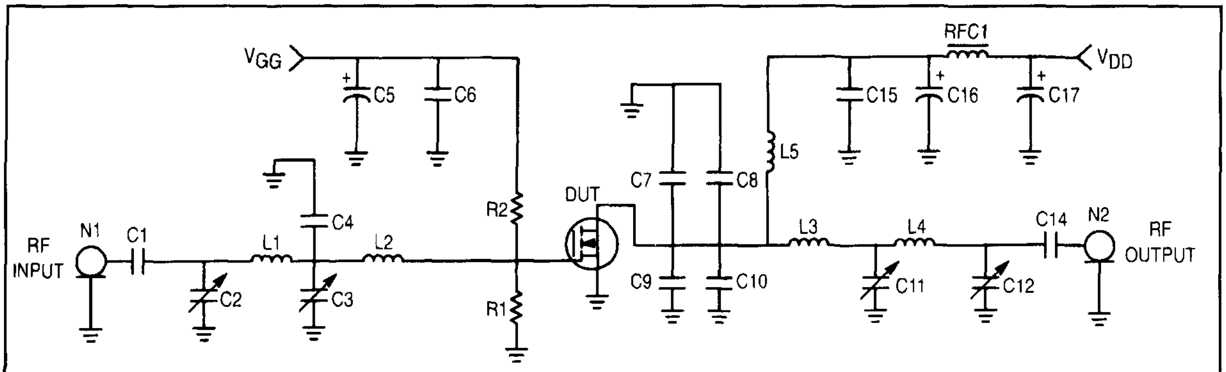
Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	1.0	°C/W

ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted)

Characteristic	Symbol	Min	Typ	Max	Unit
OFF CHARACTERISTICS					
Drain–Source Breakdown Voltage (V_{GS} = 0, I_D = 20 mAdc)	$V_{(BR)DSS}$	36	—	—	Vdc
Zero Gate Voltage Drain Current (V_{DS} = 15 Vdc, V_{GS} = 0)	I_{DSS}	—	—	5.0	mAdc
Gate–Source Leakage Current (V_{GS} = 20 Vdc, V_{DS} = 0)	I_{GSS}	—	—	5.0	μAdc
ON CHARACTERISTICS					
Gate Threshold Voltage (V_{DS} = 10 Vdc, I_D = 25 mAdc)	$V_{GS(th)}$	1.25	2.3	3.5	Vdc
Drain–Source On–Voltage (V_{GS} = 10 Vdc, I_D = 4.0 Adc)	$V_{DS(on)}$	—	—	0.4	Vdc
Forward Transconductance (V_{DS} = 10 Vdc, I_D = 3.0 Adc)	g_{fs}	4.2	—	—	S
DYNAMIC CHARACTERISTICS					
Input Capacitance (V_{DS} = 12.5 Vdc, V_{GS} = 0, f = 1.0 MHz)	C_{iss}	—	140	—	pF
Output Capacitance (V_{DS} = 12.5 Vdc, V_{GS} = 0, f = 1.0 MHz)	C_{oss}	—	285	—	pF
Reverse Transfer Capacitance (V_{DS} = 12.5 Vdc, V_{GS} = 0, f = 1.0 MHz)	C_{rss}	—	38	44	pF
FUNCTIONAL TESTS (In Motorola Test Fixture.)					
Common Source Amplifier Power Gain, f_1 = 54, f_2 = 54.001 MHz (V_{DD} = 12.5 Vdc, P_{out} = 55 W (PEP), I_{DQ} = 400 mA)	G_{ps}	13	16	—	dB
Intermodulation Distortion (1), f_1 = 54.000 MHz, f_2 = 54.001 MHz (V_{DD} = 12.5 Vdc, P_{out} = 55 W (PEP), I_{DQ} = 400 mA)	$IMD_{(d3,d5)}$	—	−30	−25	dBc
Drain Efficiency, f_1 = 54; f_2 = 54.001 MHz (V_{DD} = 12.5 Vdc, P_{out} = 55 W (PEP), I_{DQ} = 400 mA)	η	40	45	—	%
Drain Efficiency, f = 54 MHz (V_{DD} = 12.5 Vdc, P_{out} = 55 W CW, I_{DQ} = 400 mA)	η	—	60	—	%
Output Mismatch Stress, f_1 = 54; f_2 = 54.001 MHz (V_{DD} = 12.5 Vdc, P_{out} = 55 W (PEP), I_{DQ} = 400 mA, VSWR = 20:1, at all phase angles)	ψ	No Degradation in Output Power Before and After Test			

(1) To MIL–STD–1311 Version A, Test Method 2204B, Two Tone, Reference Each Tone.

Table 1. Ratings and performance specifications.



C1 — 470 pF, Chip Capacitor
C2, C3, C11, C12 — 20–200 pF, Trimmer, ARCO #464
C4 — 100 pF, Chip Capacitor
C5, C17 — 100 μF, 15 V, Electrolytic
C6 — 0.001 μF, Disc Ceramic
C7, C8, C9, C10 — 330 pF, Chip Capacitor
C14 — 1200 pF, ATC Chip Capacitor
C15 — 910 pF, 500 V, Dipped Mica
C16 — 47 μF, 16 V, Electrolytic

L1 — 8 Turns, #20 AWG, 0.126″ ID
L2 — 5 Turns, #18 AWG, 0.142″ ID
L3 — 3 Turns, #20 AWG, 0.102″ ID
L4 — 7 Turns, #24 AWG, 0.070″ ID
L5 — 6.5 Turns, #18 AWG, 0.230″ ID, 0.5″ Long
N1, N2 — Type N Flange Mount
RFC1 — Ferroxcube VK-200-19/4B
R1 — 39 kΩ, 1/4 W Carbon
R2 — 150 Ω, 1/4 W Carbon
Board — G-10 .060″

Figure 2. Motorola 54-MHz test circuit.

MRF255 is ICOM's new IC-706, where a pair of these devices provide 100 watts-plus output over a 1.8 to 54 MHz range from a single compact amplifier deck. High stage gain and simplified PA biasing contribute to the transceiver's uncommonly small size. According to a Motorola engineer, no one earth-shattering breakthrough was responsible for the development of a successful 12.5-volt device. Rather, development was a function of many smaller refinements and optimizations of existing Motorola MOSFET technology.

MRF255 specifications

The MRF255 is a 55-watt 12.5-volts DC N-channel broadband power-FET designed for large-signal common-source linear amplifier applications at frequencies to 54 MHz. Like its TMOS predecessors, the MRF255 is a vertical-channel enhancement mode device. Minimum gain at 54 MHz is guaranteed at 13 dB, with 16 dB gain being the norm (expect to provide about 1.2 watts of drive for 55-watts CW output). In SSB service, typical IMD/3 performance is -30 dB or better at 55 watts PEP output. Zin at 54 MHz is specified at 6.5=j7.96 and Zout specified at 1.27+j1.45 (S-parameter data from 1 to 260 MHz is provided in the data sheet to assist engineers with amplifier design). For linear operation, Idq (idle current) is specified at 400 mA. The device is packaged in a 211-11 style-2 case with gold mounting tabs and an aluminum nitride package insulator. For more complete ratings and performance specifications, refer to **Table 1**. Case dimensions are provided in **Figure 1.**

FET advantages

The MRF255 offers all the fundamental advantages provided by other RF power FETS. For one thing, it has very high gain—16 dB being typical at 54 MHz—with higher gain at lower frequencies. This, in turn, can downscale your driver transistor requirements into the "small-signal device" category—saving a lot in packaging volume, excess heat, and cost. FET amplifiers also have a simple bias supply requirement. While BJT bias circuits use large components and consume extra power, FET bias circuits accomplish the same function on milliwatts using only a couple 1/4-watt carbon resistors and a small trimpot. The FET bias circuit can also function as a control gain, and may be adapted to provide a AGC or remote-MGC function.

In addition to biasing easily, FETS normally don't require thermal sensing in their bias circuit to prevent runaway. Unlike BJTs, the hotter they get, the less current they draw—offering a measure of inherent built-in overcurrent protection. And, unlike BJTs, FETs can be extremely forgiving of mismatched loads.

Generally speaking, as long as you don't abuse the gate or exceed rated Vdd, it's nearly impossible to kill a FET in a properly designed and well laid-out amplifier circuit! Finally, the FET offers immunity from some types of instability that can plague BJT amplifiers—especially half-f_O mode parasitics.

FET precautions and techniques

The most important thing to avoid when working with FETs is gate damage (sometimes called "punch through"). Gate damage usually occurs when an electrostatic charge or overvoltage input ruptures and destroys a portion of the thin oxide layer that forms the device's gate. To prevent this from happening, always store MOSFETs in conductive-type protective materials. Also, use a grounded wrist strap and grounded-tip iron when working with these devices on the bench. In addition, be sure to design a measure of gate protection into your amplifier circuits. To see how this is done, refer to the bias network in the 54-MHz Motorola test-circuit shown in **Figure 2**. While the voltage divider provides a high DC resistance path to ground through R1, the AC impedance through R2, C5-C6 is low across a broad frequency range. This limits the potential for gate damage from input transients—without increasing the bias network's DC current consumption. If necessary, you may install an external zener on the gate for added transient protection (the MRF255 is not internally protected). Finally, avoid introducing large voltage spikes into the amplifier's drain circuit. These can couple through to the gate via internal parasitic capacitance within the device.

The high gain exhibited by most RF MOSFETs is generally considered an advantage. However, gain can also become a liability on a poorly designed board. Think "UHF," and work carefully to eliminate all un-necessary sources of parasitic inductance—especially in the area of the source-to-ground path (parasitic inductance in the source path can reduce gain *and* increase the potential for instability). Double-sided pc board is essential, and Motorola recommends solidifying the top and bottom groundplane surfaces with plate-throughs or eyelets every 0.1 inch in the area immediately around the device. Be sure to use multiple bypassing on Vgg and Vdd lines to provide a low-Z path to ground across the broadest frequency range possible. For low frequency designs, consider using an isolation resistor in series with each gate to reduce high-frequency gain (a value of 10-ohms or less is probably sufficient for the MRF255).

Finally, most MOSFET amplifiers require some external circuitry to optimize IMD performance in SSB service. You'll note that in **Figure 2**, four 330 pF drain shunts (C7-C10) are installed to ground. These provide harmonic termination, which lowers amplifier's IMD. In some applications, introducing a small amount of negative feedback via a simple RC network may also reduce IMD without significantly reducing stage gain.

For a more complete discussion of MOSFET RF power amplifier design methods, obtain a copy of Motorola Application Note AN211A, "FETs in Theory and Practice," from the Motorola Literature Center. Also, *Radio Frequency Transistors—Principles and Practical Applications* by Norm Dye and Helge Granberg[2] provides an excellent hands-on treatment of both MOSFET and BJT amplifier design. This book is also available through the Motorola Literature Center.

Ordering the MRF255

The MRF255 is currently available in the U.S. from Motorola distributors. Single lot (or low-volume) pricing is approximately $47 per unit. Samples from stock—and requests for production-quantity pricing—should be directed to your area Motorola sales office or Motorola distributor. Literature may be ordered directly from the Motorola Literature Distribution Center at 1-800-411-2447. For data sheets, order MRF255/D; for a circuit board photomaster, order MRF255PHT/D. Motorola Application Note AN211 is also recommended for designers not familiar with design and construction practices used in conjunction with high-power FET power amplifiers.

Conclusion

MOSFET RF-power amplifiers provide designers and builders of HF and VHF equipment with some interesting options and possibilities not offered by BJTs. Up until now, the primary drawback to using FETs in low-voltage portable or mobile SSB transceivers has been the requirement for 28 or 50 volts Vdd. The MRF255 now opens the door for designers of 12.5-volt equipment to give FETs a serious try. I, for one, have been waiting a long time for this opportunity!

A "Quarterly Devices" Bonus Device

In this issue, "Quarterly Devices" brings you two for the price of one. Peter Bertini, K1ZJH, our senior technical editor has information on

PCC9-KIT-ND

Size Code	Type	WV DC	Cap. pF	Tolerance
0603	NPO*	50	0.5	±0.25 pF
		50	1	±0.25 pF
		50	1.5	±0.25 pF
		50	2	±0.25 pF
		50	3	±0.25 pF
		50	4	±0.25 pF
		50	5	±0.25 pF
		50	6	±0.5 pF
		50	7	±0.5 pF
		50	8	±0.5 pF
		50	9	±1 pF
		50	10	±10 percent
		50	12	±10 percent
		50	15	±10 percent
		50	18	±10 percent
		50	22	±10 percent
		50	27	±10 percent
		50	33	±10 percent
		50	39	±10 percent
		50	47	±10 percent
		50	56	±10 percent
		50	68	±10 percent
		50	82	±10 percent
		50	100	±10 percent
		50	120	±10 percent
		50	150	±10 percent
		50	180	±10 percent
		50	220	±10 percent
		50	270	±10 percent
		50	330	±10 percent
		50	390	±10 percent

*12 pF not included

PCC10-KIT-ND

Size Code	Type DC	WV pF	Cap.	Tolerance
0603	X7R	50	220	±10 percent
		50	270	±10 percent
		50	330	±10 percent
		50	390	±10 percent
		50	470	±10 percent
		50	560	±10 percent
		50	680	±10 percent
		50	820	±10 percent
		50	1000	±10 percent
		50	1200	±10 percent
		50	1500	±10 percent
		50	1800	±10 percent
		50	2200	±10 percent
		50	2700	±10 percent
		50	3300	±10 percent
		50	3900	±10 percent
		50	4700	±10 percent
		50	5600	±10 percent
		50	6800	±10 percent
		50	8200	±10 percent
		50	10000	±10 percent
		50	12000	±10 percent
		50	18000	±10 percent
		50	22000	±10 percent

PCC7-KIT-ND

Size Code	Type DC	WV pF	Cap.	Tolerance
0805	NPO	50	0.5	±0.25 pF
		50	1.0	±0.25 pF
		50	1.5	±0.25 pF
		50	2.0	±0.25 pF
		50	3.0	±0.25 pF
		50	4.0	±0.25 pF
		50	5.0	±0.25 pF
		50	6.0	±0.5 pF
		50	7.0	±0.5 pF
		50	8.0	±0.5 pF
		50	9.0	±0.5 pF
		50	10	±0.5 pF
		50	12	±5 percent
		50	15	±5 percent
		50	18	±5 percent
		50	22	±5 percent
		50	27	±5 percent
		50	33	±5 percent
		50	39	±5 percent
		50	47	±5 percent
		50	56	±5 percent
		50	68	±5 percent
		50	82	±5 percent
		50	100	±5 percent
		50	120	±5 percent
		50	150	±5 percent
		50	180	±5 percent
		50	220	±5 percent
		50	270	±5 percent
		50	330	±5 percent
		50	390	±5 percent
		50	470	±5 percent
		50	560	±5 percent
		50	680	±5 percent
		50	820	±5 percent
		50	1000	±5 percent

Table 2. Capacitor values for the 0603, 0805, and 1206 capacitors in both X7R and NPO.

PCC8-KIT-ND

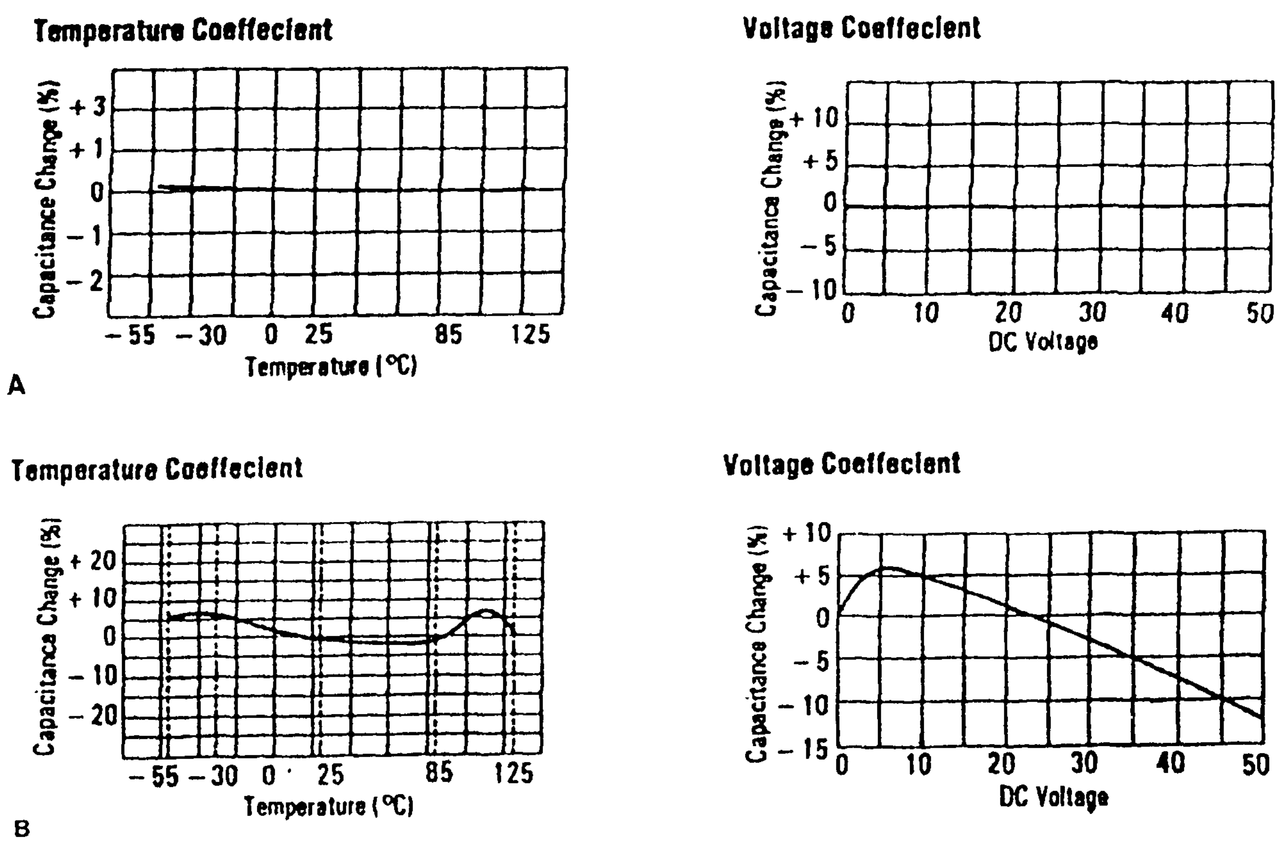
Size Code	Type DC	WV pF	Cap.	Tolerance
0806	X7R	50	220	±10 percent
	50	270		±10 percent
	50	330		±10 percent
	50	390		±10 percent
	50	470		±10 percent
	50	560		±10 percent
	50	680		±10 percent
	50	820		±10 perceent
	50	1000		±10 percent
	50	1200		±10 percent
	50	1500		±10 percent
	50	1800		±10 percent
	50	2200		±10 percent
	50	2700		±10 percent
	50	3300		±10 percent
	50	3900		±10 percent
	50	4700		±10 percent
	50	5600		±10 percent
	50	6800		±10 percent
	50	8200		±10 percent
	50	10000		±10 percent
	50	12000		±10 percent
	50	15000		±10 percent
	50	18000		±10 percent
	50	20000		±10 percent
	50	22000		±10 percent
	50	27000		±10 percent

PCC5-KIT-ND

Size Code	Type DC	WV pF	Cap.	Tolerance
1206	NPO	50	0.5	±0.25 pF
		50	1.0	±0.25 pF
		50	1.5	±0.25 pF
		50	1.8	±0.25 pF
		50	2.2	±0.25 pF
		50	2.7	±0.25 pF
		50	3.3	±0.25 pF
		50	3.9	±0.25 pF
		50	4.7	±0.25 pF
		50	5.1	±0.25 pF
		50	5.6	±0.25 pF
		50	6.8	±0.5 pF
		50	7.5	±0.5 pF
		50	8.2	±0.5 pF
		50	9.1	±0.5 pF
		50	10	±0.5 pF
		50	12	±5 percent
		50	15	±5 percent
		50	18	±5 percent
		50	20	±5 percent
		50	22	±5 percent
		50	27	±5 percent
		50	33	±5 percent
		50	39	±5 percent
		50	47	±5 percent
		50	51	±5 percent
		50	56	±5 percent
		50	68	±5 percent
		50	100	±5 percent
		50	120	±5 percent
		50	1000	±5 percent

PCC6-KIT-ND

Size Code	Type DC	WV pF	Cap.	Tolerance
1206	X7R	50	220	±10 percent
		50	270	±10 percent
		50	330	±10 percent
		50	390	±10 percent
		50	470	±10 percent
		50	560	±10 percent
		50	680	±10 percent
		50	820	±10 percent
		50	1000	±10 percent
		50	1200	±10 percent
		50	1500	±10 percent
		50	1800	±10 percent
		50	2200	±10 percent
		50	2700	±10 percent
		50	3300	±10 percent
		50	3900	±10 percent
		50	4700	±10 percent
		50	5600	±10 percent
		50	6800	±10 percent
		50	8200	±10 percent
		50	10000	±10 percent
		50	12000	±10 percent
		50	15000	±10 percent
		50	18000	±10 percent
		50	22000	±10 percent
		50	27000	±10 percent
		50	33000	±10 percent
		50	39000	±10 percent
		50	47000	±10 percent
		50	1000000	±10 percent



(A) NPO dielectric type specifications: Operating range: –55 degrees C ~ +125 degrees C. Capacitance range: 1.0 pF ~ 1000 pF. Voltage: 50 volts DC. Q factor (1 MHz, 0.5 ~ 5 volts rms, 25 degrees C). Capacitance 30 pF: 400+20C min. 30 pF. Capacitance: 1,000 pF: 1,000 min. Dissipation factor (1 kHz, 0.5 ~ 5 v rms, 25 degrees C). Capacitance: 1,000 pF...0.2 percent max. Insulation resistance (rated voltage applied at 25 degrees C): 10,000 megs min.
(B) X7R dielectric type specifications: Operating temperature range: –55 degrees C ~ +125 degrees C. Capacitance range (1 kHz, 1.0 volts rms, 25 degrees C): 220 pF ~ 47,000 pF. Voltage: 50 volts DC. Dissipation factor (1 kHz, 1.0 volts rms, 25 degrees C): 2.5 percent max. Insulation resistance (rated voltage applied at 25 degrees C): 10,000 megs min. or 500 megs x µF min., whichever is smaller.

Figure 3. Temperature and voltage shifts for (A) NPO and (B) X7R capacitors.

some surface-mount-technology kits available from Digi-Key.

Digi-Key's Panasonic Multilayer Ceramic Chip Capacitor Kits

Surface-mount-technology (SMT) components are used in most new products, and RF design engineers and service technicians need to have an assortment of these devices. Unfortunately, unlike most other discrete components, SMTs are often sold in production lot quantities, with minimum orders set at 100 to 1000 devices of one value. Recently, I needed several surface-mount capacitors of different values to assemble a Motorola engineering evaluation board. Alas, the average price of these items ran about $14 per hundred for each value I required—a substantial outlay considering I only needed about 70 cents worth of components! On average, about 60 or 70 different value capacitors are needed to cover the most popular ranges, and most of these are available in several different package types and temperature coefficients, as well.

Fortunately, Digi-Key[3] has introduced a series of parts kits encompassing many of the items offered in their rather extensive SMT parts line. Of immediate interest to me were the Digi-Key kits for Panasonic's line of multilayer ceramic chip capacitors. With some exceptions at the higher capacitance values that are available in X7R only, most of these capacitors can be ordered with either X7R or NPO temperature coefficients (see **Figure 3**).

Table 2 lists the capacitor values for the NPO and X7R versions of the Panasonic 1206, 0805, 0603 SMT capacitors. **Figure 4** shows these three SMT package sizes. All capacitors are rated for 50 volts DC. The Panasonic 1206, 0805, and 0603 SMT capacitors are currently offered as Digi-Key assortment kits in both X7R and NPO versions. The SMT 0402 package-style capacitors will soon be featured in kit assortments found in upcoming Digi-Key catalog releases. Ten capacitors of each value are supplied in each kit (see **Table 3**).

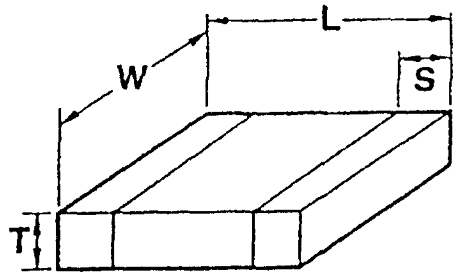
Careful handling required

The kits come in handsome white vinyl-clad notebooks that may be stored vertically on a bookshelf. The kit part number is noted on the spine of the notebook. Inside the notebook, individual plastic bins are provided for each value. Velcro closures secure the capacitors in their individual bins when the cover is shut. Each group of ten capacitors is supplied on a small segment of reel tape in a labeled plastic bag. The front inside flap of the notebook has a diagram keyed to the bin locations and gives capacitor values and parts numbers. it takes about five minutes to sort and place the capacitors into their assigned bins.

I *strongly* suggest that you keep each value in its labeled shipping bag and each value together on the tape segment! The bags will fit into the bins if folded. The worst experience of my life was dropping a CoilCraft assortment kit. Despite the color coding, I still spent hours sorting out the different values and getting them back in order. I can't imaging trying to sort out 300 chip capacitors on the loose! The parts values stamped on the SMT packages are impossible to read without a strong magnifying glass. If you do remove the caps from the bags, I would advise that you at least consider storing the tape segments under the foam rubber supplies in each bin. The foam is a snug fit, and will survive the trauma of a small drop. Having all of the parts loose in the bins is handy, but courts disaster.

It would be nice to be able to have one each of these kits on hand, but this is my hobby and not my full-time job. Fortunately, most circuit board layouts will generally accommodate the use of capacitors with either 0603, 0805, or 1206 SMT footprints. I elected to start with the Digi-key PCC9-Kit-ND and the Digi-Key PCC6-KIT-ND SMT capacitor assortments. and fill in the remainder as the need and finances permit.

I chose the PCC9 because it yields a good assortment of 30 NPO capacitor values over the 0.5 to 390 pF range. These capacitors would most likely end up in critical RF VFO or tuned circuits, and I wanted the stability offered by the NPO temperature coefficient. The 0.5 to 390 pF range also seemed to cover the majority of my needs for both NPO and small value capacitors. The biggest drawback of this assortment of handling the tiny 0603-style package.



Size Code	L	W	T	S
	±0.2	±0.2	max.	±0.3
0603	1.6	0.8	0.8	0.3
0805	2.0	1.25	0.65	0.5
1206	3.2	1.6	0.65	0.6

Figure 4. Length, width, thickness, and solder area data for 0603, 0805, and 1206 packages.

By the way, the 0603 NPOs come in 31 values, but the 12-pF capacitor is omitted from the PCC9 kit. There was a small error in the early 1995 Digi-Key catalogs that listed 25 values of 10 capacitors, each for the PCC9 assortment, for a total of 250 instead of the actual 300 capacitors in the PCC9 kit. Likewise, the information for the PCC10 incorrectly listed 30 values of 10 capacitors instead of the actual 25

Panasonic 1206 NPO, 10 ea. of 30 values, 300 pieces. Digi-Key part number PCC5-KIT-ND, $39.95.
Panasonic 1206 X7R, 10 ea. of 30 values, 300 pieces. Digi-Key part number PCC6-KIT-ND, $49.95.
Panasonic 0805 NPO, 10 ea. of 30 values, 300 pieces. Digi-Key part number PCC7-KIT-ND, $39.95.
Panasonic 0805 X7R, 10 ea. of 26 values, 260 pieces. Digi-Key part number PCC8-KIT-ND, $29.95.
Panasonic 0603 NPO, 10 ea. of 30 values, 300 pieces. Digi-Key part number PCC10-KIT-ND, $34.95.
Panasonic 0603 X7R, 10 ea. of 25 values, 250 pieces. Digi-Key part number PCC10-KIT-ND, $34.95.

Table 3. Catalog listing of Digi-Key's Panasonic chip kits, from the 951 1995 Digi-Key catalog. Prices subject to change.

values for the X7R capacitors. This means that the actual parts count will be 250, and not 300 as stated in the catalog.

I also felt that the PCC6 offered me the widest and best assortment (30 values) of higher capacitance values, spanning from 220 to 100,000 pF (0.1 µF). These X7R capacitors are suitable for RF bypassing or coupling where some temperature drift is easily tolerated. The PCC6 capacitors are the larger 1206-style SMT package and most easily handled. I am well pleased with the Digi-Key Panasonic capacitor kits; the price is fair and the assortment selections are large and varied. The only additional thing one could wish for is a small pair of tweezers to help handle these little fellows!

Comments or questions?

As usual, if you design and build an interesting circuit using a quarterly device, we would like to share it with our readers as a construction article or "Tech Notes" entry in the Quarterly. Also, if you have comments or questions concerning "Quarterly Devices," please direct them to the *Communications Quarterly* editorial office at Box 465, Barrington, New Hampshire 03825, or via e-mail to k1bqt@aol.com. ■

CORRECTIONS

Two articles were inadvertently omitted from the Fall 1995 *Communications Quarterly* Article Index. Under the heading "Antennas and Related Topics," please add:

• **Designing the Long Wire Antenna System**
Bill Sabin, W0IYH
Summer 1994, page 75

Also, add the following under "Filters" and "Quarterly Devices:"

• **Quarterly Devices: New Narrowband 10.7-MHz Ceramic Filters from Murata**
Rick Littlefield, K1BQT
Summer 1994, page 92

PRODUCT INFORMATION

New Software For Radio Antenna Design

A new computer-aided antenna design and analysis package, based on a Penn State engineer's core technology promises to replace trial-and-error methods of antenna design and optimization.

The package, called NEC-WIN, can be run at expert, intermediate, or novice skill levels and can be used to optimize an amateur radio antenna or to model advanced applications for a professional designer. Users can input their requirements and have the software generate a 3-D graphic view of an appropriate antenna. The software can also generate 2-D and 3-D plots of the antenna's output pattern.

NEC-WIN is packaged and marketed by Paragon Technology, Inc., of State College, Pennsylvania and is based on modeling techniques developed by Dr. James K. Breakall, associate professor of electrical engineering at Penn State.

For more information, contact Barbara Hale or Vicki Fong at 814-865-9481; or e-mail either bah@psu.edu or vyfl@psu.edu.

New Brochure from Hewlett Packard

Hewlett-Packard Company has released a brochure that describes test and measurement products that evaluate wireless-communications-systems performance. Product families included in the brochure are vector signal analyzers, modulation domain analyzers, RF-channel simulators, peak-power analyzers, signal generators, LCR meters, vector and scalar network analyzers, spectrum analyzers, board test systems, mixed-signal integrated-circuit testers, modulation/audio analyzers and transceiver test sets. Also included is dedicated software for evaluating transmitter performance according to the requirements of various wireless-communications standards. These product families are capable of evaluating the performance of products including GSM 900, DCS 1900, TDMA (NADC), DECT, PHS, CT2, CT3, DCA, AMPS, NAMPS, TACS/ETACS and NTACS.

The 18-page booklet, "Test and Measurement Solutions for Wireless Communications," (Literature 5953-9471E), is available free of charge from Hewlett-Packard Company by calling 1-800-452-4844, ext. 9149; or write to Hewlett-Packard Company, Direct Marketing Organization, P.O. Box 58059, MS51L-SJ, Santa Clara, CA 95051-8059.

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INSTRUMENTS FOR ANTENNA DESIGN DEVELOPMENT AND MAINTENANCE PART 4:

Field strength meters, grid dip oscillators, and some mechanical devices

Most amateurs are surprised to learn they own an unused field strength indicator—their SWR meter. With a telescoping antenna attached to one connector and the sensitivity control turned to high, this meter is helpful in many situations. Some small indicators sold for CB use even have a special jack for just such an antenna.

A better design (see **Figure 1**) uses a tuning coil, which increases the sensitivity and reduces the effect of interfering signals. If you live near a broadcast station, a tuned unit is a necessity. The handbooks provide information on variations in design, including some for field intensity measurement. Commercial versions are fairly common at hamfests.

A field strength indicator can be calibrated by measuring the strength of a high-end broadcast station at one or more known locations, and using the strength curves the station filed with its FCC application. With care, a calibration can be obtained using a short vertical fed with a known amount of power.

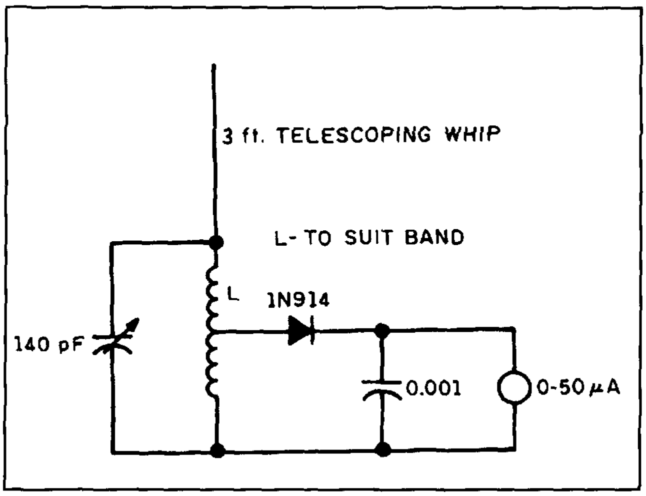

Figure 1. Basic field strength meter. Shown are the essential elements of a field strength meter measuring the electric field. A magnetic field measurement type omits the pickup whip, and enlarges the coil into a small loop or a ferrite "loop-stick." Low sensitivity types omit the tuned circuit, but become sensitive to stray fields. An amplifier may be added to increase sensitivity. See handbooks for design detail.

The signal strength meter indicates the RF field—the entire purpose of the antenna. Such a meter is mandatory for proper adjustment of a low-frequency phased vertical array. During development of a new directional antenna, the pattern is the best indicator of gain performance. One way to measure it is with the field strength meter.

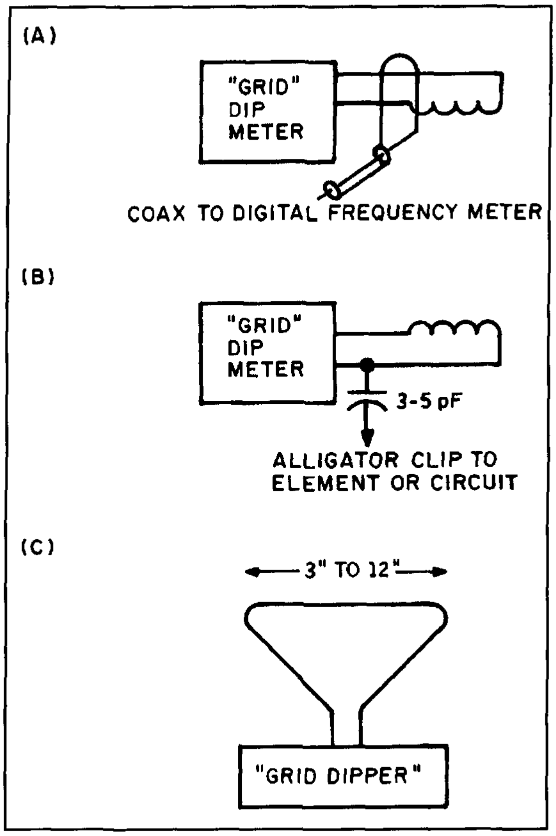

Figure 2. Dipper modifications. (A) Addition of a pickup loop for magnetic coupling to a digital frequency meter. The loop should be permanently mounted, together with its associated cable, after determining that the pickup level is adequate for the meter in use. The dipper is now a precision resonance indicator. (B) Addition of a coupling capacitor to permit capacitive coupling to a resonant circuit, for example an antenna. The coil connection can be one or two turns of the lead around the coil pin, but an internal connection to the tip jack is better. This should be connected to a high-voltage point. In an antenna, this is the element end. In a trap antenna, it's the inside end of the trap resonant on the band being investigated. Traps are best checked when separated from the rest of the element. A shallow dip is reason to suspect a poor trap. This is an alternative to (A) for the frequency meter. (C) Coil shape to increase magnetic coupling to a linear circuit, such as an antenna element. Several turns are needed for the lower frequencies. Large coils tend to damage the coil receptacle, so mechanical support should be added for these. A pair of hooks can be built into this, to hold the dipper at a constant position to the element. Coupling should be to a high-current point. In dipoles and trap dipoles, this is the element center. Using this with (A) allows determination of the exact resonant frequency of the element, especially important in parasitic beams. The assembly is very useful in checking for tower, guy, and boom resonances. It's also useful in TVI elimination, for searching for unwanted resonances in the transmitter and TV antenna, and as an interfering signal source.

The grid dip oscillator

The grid dip oscillator (GDO) is one of the most neglected antenna measuring devices. The GDO is used to obtain the resonant frequency of an antenna element, a trap, a guy, or a boom. It provides an indication of the usable bandwidth and also the loss. For a good dip indication, simply adjust the dipper's sensitivity control. A narrow, deep dip occurs if the element being checked is high Q, which also means low loss. A narrow, shallow dip usually means low coupling. A wide-band reduction of reading with no pronounced dip indicates loss.

Figure 2 shows three additions or modifications the dipper needs to be really useful. The first at (A) allows use of a digital frequency meter and makes the dipper a precision resonance indicator. At the dip, the frequency of the dip oscillator is determined by whether the coupled circuit is high Q, and is the frequency of the coupled circuit at the exact dip—even for lower Q circuits.

A temporary loop or a pickup antenna attached to the frequency meter is adequate for occasional work. But for regular use it's necessary to add a connector and mount the loop permanently. Alternatively, the antenna may be connected to some internal pickup point. In the tube-type unit, there's a capacitor to the cathode. In tunnel diode and transistor models, it may be necessary to add an FET isolation stage coupled to the tuning circuit. These should really be built into the unit by the designer/manufacturer.

The second addition, **Figure 2B**, makes provision for capacitance coupling to the measured element. In antennas, coupling is to the end of an element or to the inside end of a trap. The capacitor can be a 2 to 5 pF ceramic with one end lead wrapped around one coil-plug pin and the other ending in a small alligator or battery clip. This is the connection to the circuit being measured. The frequency scale calibration is affected, especially at the high-frequency end of the dial. For repeat work, this capacitor should be built-in and connected to a separate jack. Again, these modifications should be provided by the dipper's designer.

The third change, shown in **Figure 2C**, involves the construction of a set of special coils. These coils are shaped to provide good magnetic coupling to a linear element, the antenna wire, or tube, and resemble wire coat hangers. You could try to make the inductance match that of the regular coils, but the frequency meter connection makes this unnecessary. In use, the straight section is brought close to the antenna.

With this change in place, the dipper allows direct measurement of the element's resonant frequency. In the Yagi, the reflector should resonate a few percent below a design frequency, and the director a few percent above. The design frequency depends on whether the antenna is optimized for maximum gain or maximum F/B ratio. The percentage depends on the desired bandwidth. Usually the radiator is made resonant, but it doesn't need to be. Specification sheets will sometimes provide the resonant frequencies.

This dipper measurement is helpful because it includes the effect of element taper, clamps, and boom shortening. It's a valuable check on design details. Use this measurement to check the calculated resonant frequency.

Other instruments

Until now, this series has dealt with electrical measurements. However, there are several useful mechanical devices. The most basic is a steel tape calibrated in meters and centimeters. This basic tool saves no end of conversions. With just a little practice, it's easy to think in meters rather than in feet.

A spring scale is another helpful mechanical device for antenna work. Not only is a scale handy when checking the weight of elements, but it's even more important for proof testing element and clamping strength. The projected element area times the wind loading (50 lbs. per square foot typically) gives the element load. The center clamp must withstand this load as a pull along the boom. The distributed load can be converted to an equivalent end load, and the scale can be used to place this at the element end. The boom-to-mast fitting must withstand the entire load of the antenna.

A torque wrench is useful when working on larger beams and on towers. Most amateur mechanics pride themselves on their ability to tighten fittings correctly. Perhaps so; however, a torque wrench is better.

An ohmmeter, which indicates fractions of an ohm, is a handy device. It indicates if you have an open trap or a badly corroded connection. If you're doing a lot of antenna work, it's worthwhile to modify a pair of large battery clips so the clamp point is semi-circular instead of nearly square. This modification provides better contact to tubing elements.

The antenna range

An antenna range can be a place for continuous antenna development or simply a place to check out "that new beam." Results depend on several fundamentals common to all ranges.

The most important consideration is the distance between the antenna and the measuring device. This distance must be great enough to allow the emitted signal to be essentially a plane wave at the measuring point. For a point source (or detector pickup) Kraus gives this as d=k*a*a/lambda, where *lambda* is the wavelength and *a* is the maximum antenna dimension, in the same units as wavelength. The *k* is determined by the accuracy needed. A value of 2 is satisfactory for work on the main lobe for gain. For adequate resolution of the sidelobes, a value of 4 will do. For very low-sidelobe designs, a value around 9 is needed.

For a typical small triband yagi, with a turning radius of 23 feet, *a* is equal to 14 meters. The minimum separation is very nearly 20 meters; that's 66 feet for gain measurement, or 130 feet for sidelobe checking. On 10 meters, it would approach twice as much. This data is for a point source or pickup. If the pickup were a half wave long, approximately another 33 or 66 feet would be required on 20 meters. Try the values for a large moonbounce antenna on 2.

Few amateurs have space for an adequate range. For individual stations, an interested, friendly neighbor is an important factor in building a personal antenna range.

Even if you have plenty of room, there are points to watch. A major one is ground reflection. The effect of ground reflection decreases with height of the antennas. It also decreases if the antennas at both ends of the path are directional; but this also means that greater separation is needed. Pattern distortion is caused by reflections like those from a roof or the side of the building. Moving objects are a delaying nuisance. These factors are the reason pattern measurements are often neglected. If measurements are made, they're usually done with the cooperation of another ham. This is sometimes a local some miles away, and sometimes DX. Neither outcome is really good: the first because of stray reflections, the second because of fading.

There are several possibilities for radio club activity here. Some groups maintain an antenna trailer for Field Day and emergency use. The addition of a good field strength meter or a low-power remotely controlled transmitter makes such a setup into a portable measuring unit. Temporary parking in a low traffic area, well away from the antenna under test, shouldn't be any problem; but warning cones and lights are indicated, and a review of plans with the local police may be in order.

Two well-separated flat roofed buildings can offer space for a good range, if there's no reflecting traffic between them. An individual, better still a club, might be able to make arrangements for use of such rooftops as an antenna range. Schools, especially high school level or above, may be more cooperative if there's an arrangement for student participation. This provides good publicity for the club and a source for new hams, as well.

Many computerized repeaters have a signal strength measuring/reporting routine. Unfortunately, this doesn't seem to attract the attention it deserves—either from the station manager or repeater users. With a little work on equipment stability and software revision for easy use, it's possible to develop a real antenna range that's available to all users. Control can be over the air, or via a special telephone line.

This idea can be extended. There are many relatively low-cost transmitters, receivers, and transceivers that cover all amateur bands (and more) under computer control. Such a radio at the repeater site, or even a special site, would make a remote antenna range a multi-band activity. The largest problem would be antennas; however, high efficiency isn't needed, so small loaded crossed loops or dipoles would serve. A really good installation would have a choice of antennas for both polarizations. ■

PRODUCT INFORMATION

New GPS Firmware For AEA Products

Advanced Electronic Applications' PK-96 dual speed 9600/1200 bps Packet controller is now shipping with GPS firmware. The PK-96 now also includes in the packaging: AEA's PC PakRatt Lite™, the packet-only, DOS TNC terminal control software and the Automatic Packet Reporting System (APRS™) software developed by Bob Bruninga, WB4APR, for GPS use, as does AEA's PK-12 1200 bps packet TNC.

The GPS firmware incorporated in both the PK-12 and PK-96 automatically detects if there is a GPS receiver connected to the TNC upon power-up. If a GPS receiver is detected, an initialization string will be sent and the TNC will be ready for GPS work; if no GPS receiver is detected, the TNC will be ready for traditional packet data work.

One new feature of the PK-96 and PK-12 is that GPS commands can be remotely programmed, so in Stand Alone Tracking applications, the unit does not need to be removed and connected to a computer to change GPS parameters—it is all done remotely. PK-96s and PK-12s automatically transmit their position information at user defined intervals and now can also be remotely polled for GPS location information at any time. Other new GPS firmware features include time and date setting from the GPS receiver, remote programming of the GPS receiver itself, and the ability to operate as a WIDE and RELAY digipeater. Both TNCs work with AEA's APRS Adapter Cable which saves a communication port on the computer. This cable allows the TNC and GPS receiver to connect to a single COM port.

GPS firmware upgrades for early PK-96s and PK-12s are available directly through AEA for $10.00 (free shipping). Call the AEA Upgrade Hotline at 206-774-1722 to order the GPS upgrade. The PK-96 and PK-12 packet controllers and AEA APRS Adapter Cable, are available from you local amateur radio dealer. GPS Firmware has also been added to AEA's PK-232MBX and PK-900 multi-mode data controllers. For more information, call AEA's Literature Request Line at 1-800-432-8873; or fax 206-775-2340.

New 80-watt Triple-Output Power Supply

Hewlett-Packard Company has introduced the HP E3631A. This new power supply provides multiple outputs and allows users to read and program current and voltage remotely through both interfaces using SCPI (Standard Commands for Programmable Instruments).

The front panel of the HP E3631A power supply utilizes a bright vacuum-fluorescent display and a traditional knob allowing users to change settings in one motion. The HP E3631A powers up components, circuits or assemblies. For systems not requiring high throughput, the power supply is a high-performance, programmable lab-bench supply that can be used to build simple systems.

The HP E3631A offers triple output (0 to –25V, 1A; 0 to +25V, 1A; and 0 to +6V, 5A); HP-IB or RS-232 programmability with SCPI via a controller or a personal computer; low noise with 0.01 percent load and line regulation; separate digital panel meters; and software calibration.

The HP E3631A programmable power supply is available from Hewlett-Packard for $995. The instrument is shipped with an operating manual (available in multiple languages), a service manual, certificate of calibration and power cord. The power supply comes with a standard three-year limited warranty.

For more information, please contact Hewlett-Packard Company, Direct Marketing Organization, P.O. Box 58059, MS51L-SJ, Santa Clara, CA 95051-8059.

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Fractal Loops and the Small Loop Approximation

By Nathan "Chip" Cohen, N1IR

2 Ledgewood Place
Belmont, MA 02178

Summary: The article discusses fractal resonance in small loop antenna designs.

Winter 1996 issue, pages 77-81.

Please contact the author for additional information.

PRODUCT INFORMATION

New CMOS Frequency Counter Chip from Radio Adventures Corp.

Radio Adventures Corp. has announced the availability of the C5 advanced CMOS Frequency Counter chip. The C5 offers many counting features in a 28-pin DIP package. The C5 along with a standard 74HC02 gate and three driver transistors drives a six-digit seven-segment LED display to 100 Hz resolution. Frequency range is DC to beyond 50 MHz and the update rate is approximately 40 times per second. Some of the advanced features include "anti-jitter code" which reduces last digit jitter, sixteen non-volatile programmable offsets, five of which are pushbutton selectable when used with optional EEPROM, selectable reverse counting for reverse tuning VFOs, selectable direct frequency readout, selectable 100 Hz digit blanking, selectable automatic display blanking for power conservation, automatic display enable when frequency changes and leading zero blanking of MHz digits.

The C5 CMOS Frequency Counter chip is available from Radio Adventures Corp. for $14.95. To receive a free data sheet, contact Lee M. Richey, Radio Adventures Corp., P.O. Box 339, Seneca, PA 16346; phone 814-677-7221; fax 814-677-6456; or e-mail rac@USA.net.

QUALCOMM Releases New Master Selection Guide

QUALCOMM Incorporated VLSI Products has announced the newly revised Master Selection Guide. Their Master Selection Guide features QUALCOMM's complete line of synthesizer, forward error correction and voice compression products for advanced communication systems.

The 37-page Master Selection Guide details the three product areas with photos, product applications, block diagrams, and tables. This resource provides background information, technical data and product summaries for DDS, Digital to Analog Converters (DAC), Phase-Locked Loop (PLL) Frequency Synthesizers, Viterbi Decoders, Trellis Codecs, Variable-Rate Vocoders, and evaluation circuit boards.

For additional information, contact QUALCOMM, 6455 Lusk Blvd., San Diego, CA 92121; fax 619-658-1556; or e-mail vlsi-products@qualcomm.com.

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CompuServe: 72510,2302
MCI Mail ID: 407-8117

QUARTERLY COMPUTING

Software that's good enough to use

Recently, a friend sent me several packages containing dusty but mechanically intact 1930s-vintage issues of *QST* magazine. As I browsed through several equipment construction articles, I gradually became aware of an interesting omission.

Almost without exception, these stories concentrated heavily on the how-to aspects of rig construction and offered very little in the way of theoretical explanations of why circuits worked, or of the compromises the unsophisticated technology of the times forced on its users.

For example, one experimenter reported instability in an exciter stage that stabilized when he lowered the screen grid-supply voltage. Chances are, the vacuum tube in question generated parasitic oscillations at a frequency far beyond the experimenter's measurement capabilities. Lowering the screen voltage probably decreased the tube's gain at VHF and suppressed the parasitics.

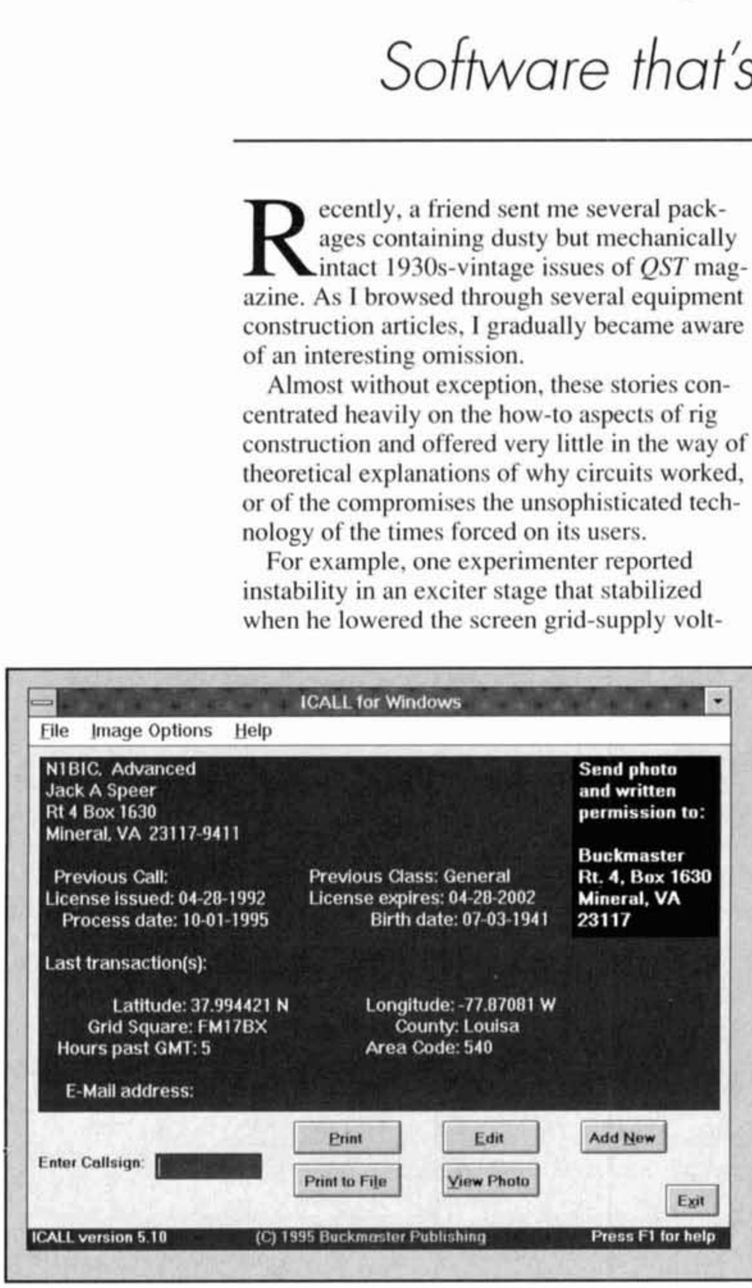


Figure 1. Enter a callsign in the lower left-hand data block of Buckmaster's ICALLW working window.

Before we allow ourselves the luxuries of condescension or amusement at the foibles of early experimenters, remember that we enjoy the twin benefits of hindsight and, in many cases, sophisticated technical knowledge that far surpasses the wildest imaginings of 1930s-era research engineers.

Sixty years hence, will anyone remember—let alone use—any of today's software? Windows 95 will no doubt occupy a place in marketing and economics textbooks, either as a footnote or a cautionary tale, but an average working software package will no doubt be long forgotten.

If you, OM or XYL, are perusing these pages in the year 2056, remember that like our predecessors of 1935 who built their equipment around balky vacuum tubes and managed to enjoy amateur radio nonetheless, these are the software tools—the type 53s and 47s and 203As—we have to work with today, and they're good enough to use.

In this issue, we'll review three widely different applications that analyze antenna designs, draw technical diagrams, contain callsign data, and much more.

Hamcall CD-ROM

Released twice a year (in October and April), the $50 Buckmaster Hamcall CD-ROM represents a work-in-progress that shows continual

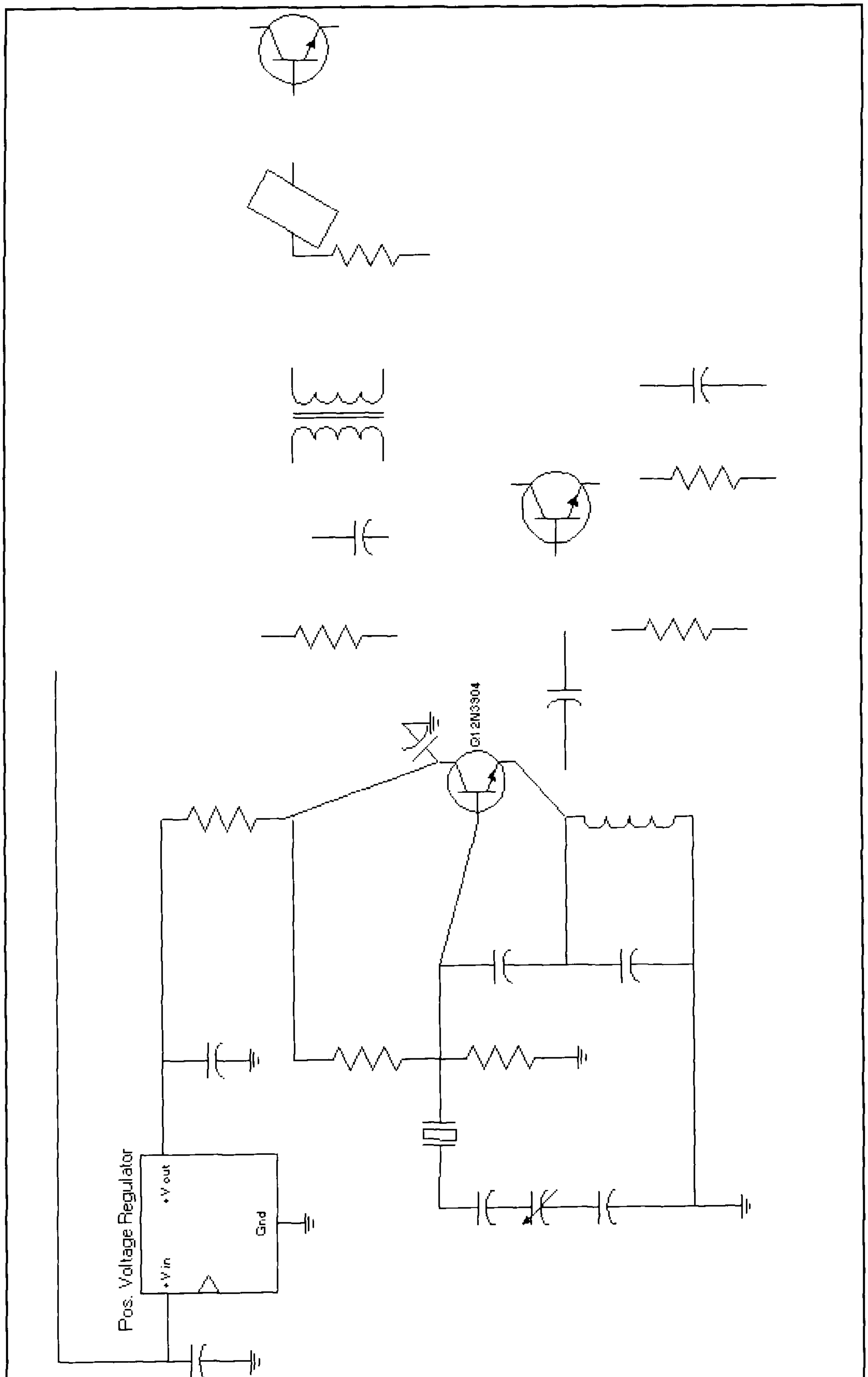


Figure 2. Schematic prepared with Visio Technical includes user-designed three-terminal voltage regulator symbol at upper left. Note rubber-banding of connections to transistor Q1.

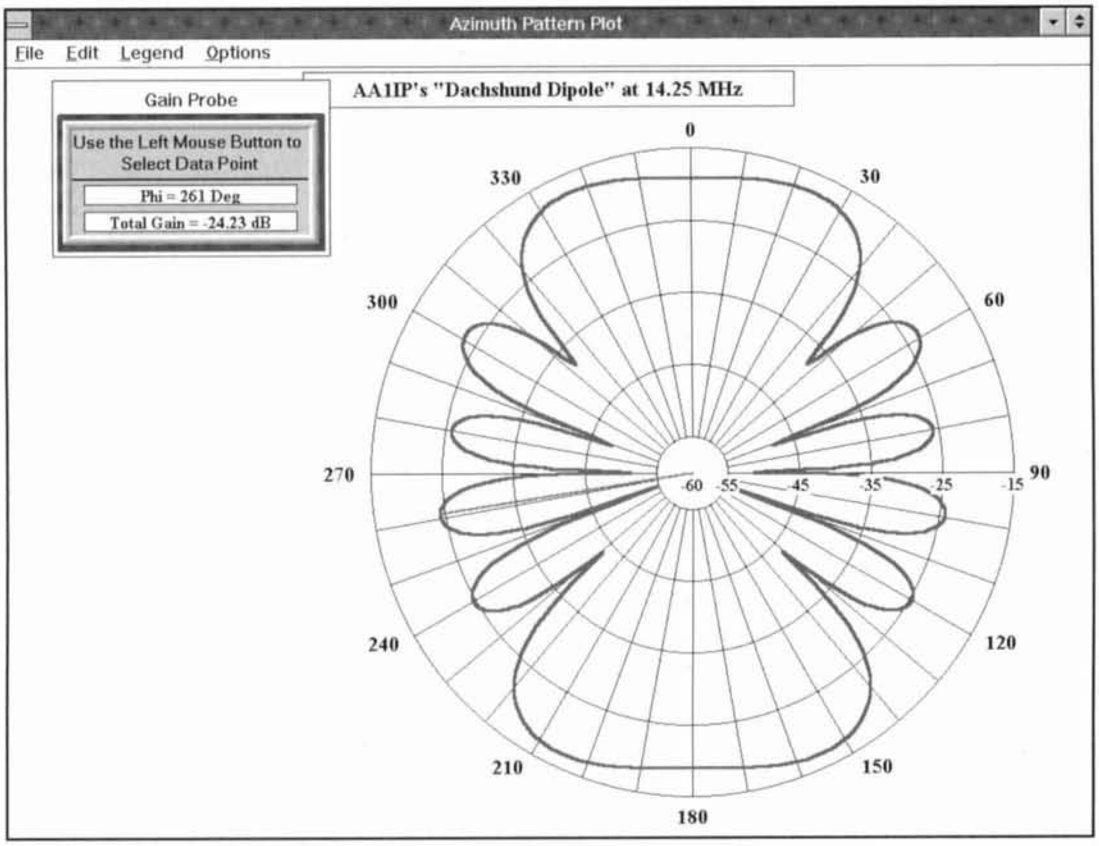


Figure 3. Azimuthal radiation plot of AA1IP's too-low, too-long, off-center-fed dipole at 14.25 MHz as compiled by NEC-Win Basic for Windows.

improvement—sometimes in increments, and sometimes in major jumps.

In its October 1995 issue, Hamcall features significant improvements. For starters, most U.S. listings now include precise location data in the form of ZIP+4 postal codes, latitude and longitude values carried to five or six decimal places, and grid-square coordinates.

According to the user's notes, the latitude/longitude data is accurate to within a few hundred feet. Now, you can easily locate your QTH on a topographical map and obtain a better idea of how local geography may affect your antenna patterns.

Actually, Hamcall contains more than one search engine—if you're attempting to locate a callsign, you can choose from ICALL (for DOS users) or ICALLW (for Windows users). **Figure 1** shows an ICALLW screen, with Buckmaster's publisher's callsign entered in the lower left-hand callsign window.

There's also a TSR (terminate-and-stay-resident) routine called LOOKUP that keeps callsign data available at a keystroke or two—a convenience if you're using a logging program.

For those of us who have difficulty remembering callsigns, Buckmaster provides a program called HAM. Built around the Folio search engine, HAM features structured searches that can help you locate any or all radio amateurs named Jack who live in, say, Topeka, Kansas. This program will find favor among those of us who operate mobile-in-motion and retain names and places longer than calls.

Other interesting features new to Hamcall include user-editable records, and photos of licensees. You can look up a friend's callsign and add his or her telephone number to the record's area code field. While writing new data to a CD-ROM disc is obviously impossible, Hamcall creates a file on your PC's hard disk drive to retain new records. When you purchase a later revision of Hamcall, the software is able to retain your existing edited records.

If you wish, you can submit a personal photo for inclusion in future releases of Hamcall. So far, over 375 hams have done so—the images range from professional to basic driver's license in quality. Under DOS, ICALL's viewing feature worked fine, but ICALLW's Windows viewer refused to display any images on the review system.

Label printing rates have improved over earlier editions. You can look up callsigns and create a queue of mailing labels (which don't include callsigns—good camouflage for the TVI-prone) that won't print until you issue a page-eject command to the printer. Labels emerge in one-up format—ideal for dot-matrix printout, but not well-suited to laser or page-oriented printers.

As a workaround, you can print the list of labels to a file, and then use an editor or word processor to reformat the labels into page layouts. Given the large number of permutations of label formats and printer models, it's unsurprising that mailing-list creation remains something of a challenge, not only for Hamcall, but for all callsign programs.

Beyond its primary mission of callsign identification, Hamcall includes over 42 MB and 320 files of assorted shareware, freeware and public-domain software. Highlights include YTAD.ZIP, an experimental program for exploring effects of HF antenna elevation versus terrain height.

Hamcall also includes over 10 MB in 635 files of ARRL-related and contributed material. You can research the ARRL's positions on legislation (some now past, some pending), check product reviews, and browse through indexes for *QST* magazine for the years 1989 through 1993.

Should you purchase a CD-ROM-based callsign directory instead of a paper version? Yes, if you need the ability to create mailing lists, search rapidly for a variety of callsigns, or just plain enjoy browsing. In the long run (say, 30 years or so), evolving technology may bypass our current CD-ROM format and render the discs unreadable, while we'll still be able to read our crumbling paper callbooks. Still, for now, Buckmaster's CD-ROM Hamcall offers considerable convenience and a bumper crop of usable shareware at a reasonable price.

Visio Technical

For those of us who document our homebrew projects in orderly notebooks filled with clearly drawn schematics, the availability of PC-based schematic-capture software has represented both a boon and a bane.

On the positive side, PC-schematic programs can create high-quality drawings, and more expensive programs can transfer circuit diagrams to printed-circuit layout software. However, most schematic-capture programs are limited in flexibility and sometimes present a formidably steep learning (or relearning) curve for the occasional user.

Offering an alternative approach, Visio Corporation's $299 (approximate street price) Visio Technical for Windows program uses a "stencil" metaphor to build a wide range of drawings and diagrams of interest to radio amateurs and anyone who works in technical fields ranging from architecture to computer programming. Visio Technical includes over 2,000 objects, or "SmartShapes" in Visio's parlance, contained in 94 stencils—roughly analogous to a PC layout program's component libraries.

Unlike the paper cutouts found in childrens' signmaking stencil kits, Visio's SmartShapes more closely resemble electronic modules in the way they're used. You assemble a Visio drawing by dragging individual symbols from a stencil and positioning each symbol on a page. When you've placed symbols as required, you draw interconnecting lines and add text labels as appropriate.

So far, this description could apply to any one of a number of what are known as object-oriented drawing programs. What sets Visio apart is its range of symbols and ease of use, plus a few features that most computer-aided design programs don't offer.

For starters, you can alter the shapes of stencil elements without introducing distortion. For example, dragging on the end of a bolt lengthens the bolt without distorting its dimensions. Another stencil includes pie chart segments; manipulating a "handle" on a segment alters its size and percentage. You assemble segments as needed to draw the chart.

Figure 2 shows a partially completed schematic diagram of a VXO in which one transistor has been moved off-grid to show the degree of "stickiness" with which components can remain interconnected or locked to the reference grid.

Visio's symbols include connection points that specify where you connect lines or other symbols. For example, a three-input AND gate would include three input points and one output point. Note that unlike true schematic-capture software, Visio's logic and analog-circuit symbols don't include assigned pin numbers or packaging of multiple circuits per device. It's up to you to label circuit pinouts and keep track of multiple-gate utilization.

I noted that a few useful symbols for electronic design were missing—including those for an LED/bipolar-transistor optical isolator, a ferrite-cored transformer, and three-terminal positive and negative voltage regulators. However, it took me only a few minutes to create the three-terminal voltage regulator symbol used in **Figure 2** and add it to the stencil library.

Also, Visio Technical's grouping feature lets you couple elements together, which makes it easy to synthesize an eight-pole switch from a group of four two-pole contact sets. You can link one page of a drawing to another via a hypertext feature.

Visio Technical includes several add-on or expansion programs dubbed Wizards, which apply a degree of automation to a task. Examples include: a utility for measuring perimeter and area of a closed shape, a property-plotting routine (handy for keeping antennas within your own QTH), and a Gantt chart creator.

File Format/Version	Imported?	Exported?
ABC Flowcharter 2.x-4.x	Yes	No
AutoCAD DXF, DWG	Yes	Yes
Bitmap BMP/DIB	Yes	Yes
Computer Graphics Metafile CGM	Yes	Yes
CorelDraw 3.x-5.x	Yes	No
CorelFLOW 2.0	Yes	No
Encapsulated PostScript EPS/AI	Yes	Yes
IGES	Yes	Yes
PCX	Yes	Yes
Macintosh PICT	Yes	Yes
Micrografix DRW	Yes	No
TIFF	Yes	Yes
Windows Metafile WMF	Yes	Yes

Table 1. Visio Technical 4.0 reads and writes files created in most common file formats.

Although I haven't had time to experiment, I'll speculate that Visio Technical might be useful for creating printed-circuit board layouts that don't lend themselves well to conventional layout tools. Examples might include: planar antennas and stripline couplers for the lower-microwave region, irregularly shaped copper-clad shields, and PC boards that include non-standard discrete components.

As **Table 1** shows, you can import and export a wide range of file types into and out of Visio Technical. Of special interest is AutoCAD compatibility, which allows you to convert drafting symbols available in AutoCAD's .DWG format into Visio master objects for reuse or editing. You can also export a Visio Technical drawing in AutoCAD's .DWG and .DXF formats.

As shipped, Visio Technical arrives on 17(!) 3 1/2-inch floppy disks and one CD-ROM. If you own a CD-ROM drive, the latter will save considerable time otherwise spent in shuffling floppies. Documentation consists of a clearly written 114-page introductory manual, plus a 420-page developer's manual and supplemental floppy disk containing help files and Visual Basic and C/C++ source code. Visio encourages users to delve into custom programming that takes advantage of Visio Technical's heavy reliance on OLE (object linking and embedment) technology for exchanging data with other Windows applications.

To use Visio Technical, you'll need a PC running Windows 3.1, Windows 95 or Windows NT—the software selects appropriate drivers at installation—and equipped with at least 4 MB of memory, VGA graphics and a 386 or higher processor. You'll enjoy faster performance if your PC contains a 486DX CPU and at least 8 MB of RAM. Also, note that a full installation of Visio Technical from CD-ROM consumed approximately 25 MB of hard disk space. One minor glitch occurred during installation from CD-ROM when Visio's promised quick tour appeared only as a grayed-out menu item.

Clearly, Visio isn't intended for large-scale logic designs, but would work well for knocking off a schematic containing, say, ten to 20 ICs and 50 or more discrete components. However, Visio also includes a wealth of symbols that will enable you to design a new ham-shack, plan a 10-GHz mountaintop repeater site using microwave-equipment symbols, create a Gantt chart, and design software via Booch, Chen, Martin, or Yourdon charts. Can your schematic-capture software do as much?

NEC-Win Basic For Windows

If you're interested in antennas, chances are you've heard of NEC—the Numerical Electromagnetics Code—that forms the core of much antenna-analysis software. You may have even looked over the public-domain source code, and found the prospect of compiling and using the code somewhat daunting. You may have tried a shareware version and achieved less-than-satisfactory results due to user bewilderment.

NEC-Win Basic for Windows, available for \$75 in version 1.0 from Paragon Technology, Inc. of State College, Pennsylvania, goes a long way toward simplifying NEC and actually making it available for radio amateurs to use on an everyday basis.

To use a classic version of NEC, you model your antenna as a collection of conductors whose endpoints are identified at (X, Y, Z) coordinates in space. You divide each conductor into segments and attach driving sources or load impedances to segments. You specify conductor

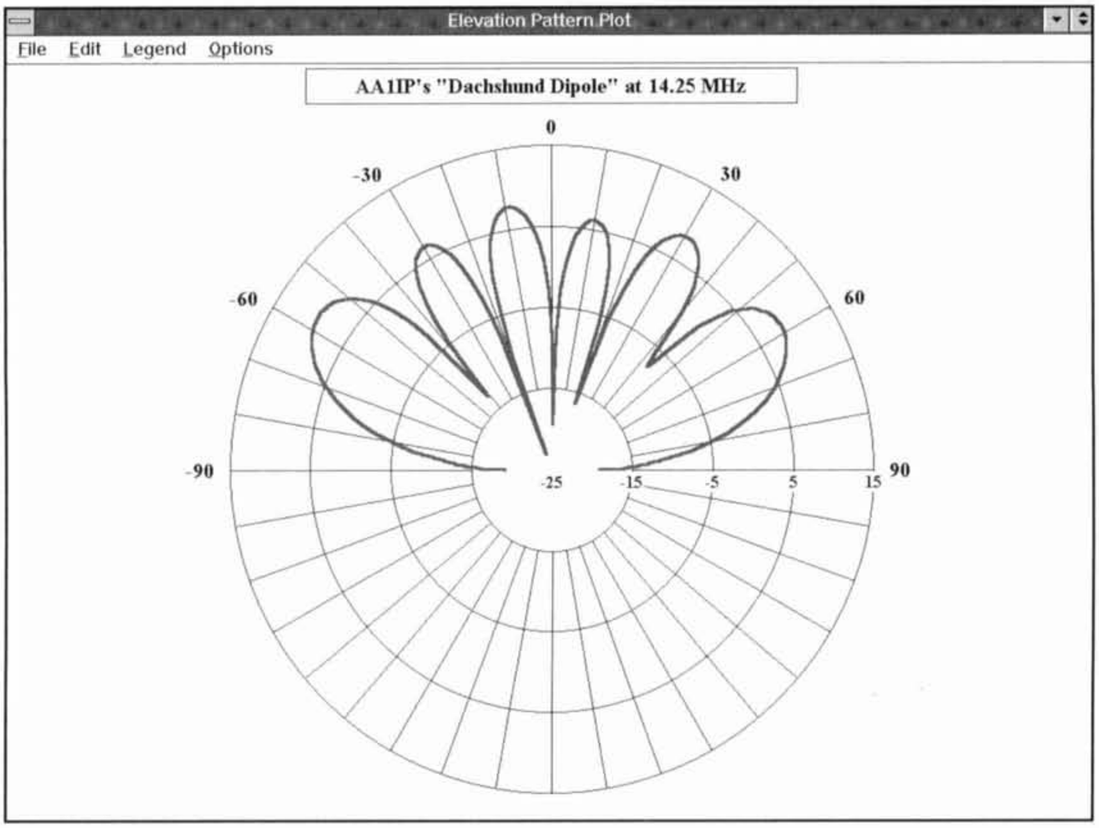


Figure 4. Elevation radiation plot of antenna in *Figure 3*.

material, frequency of operation, and run the program. In its most spartan implementations, NEC feeds you results in tabular form, forcing you to invoke an auxiliary plotting program to visualize the antenna's radiation pattern.

NEC-Win Basic neatly short-circuits the tedium by invoking a spreadsheet-like format for entry of antenna conductor lengths and properties. You specify frequency, and up to ten azimuth and elevation radiation patterns for analysis, and then click on an NecVu control. You view a three-dimensional representation of

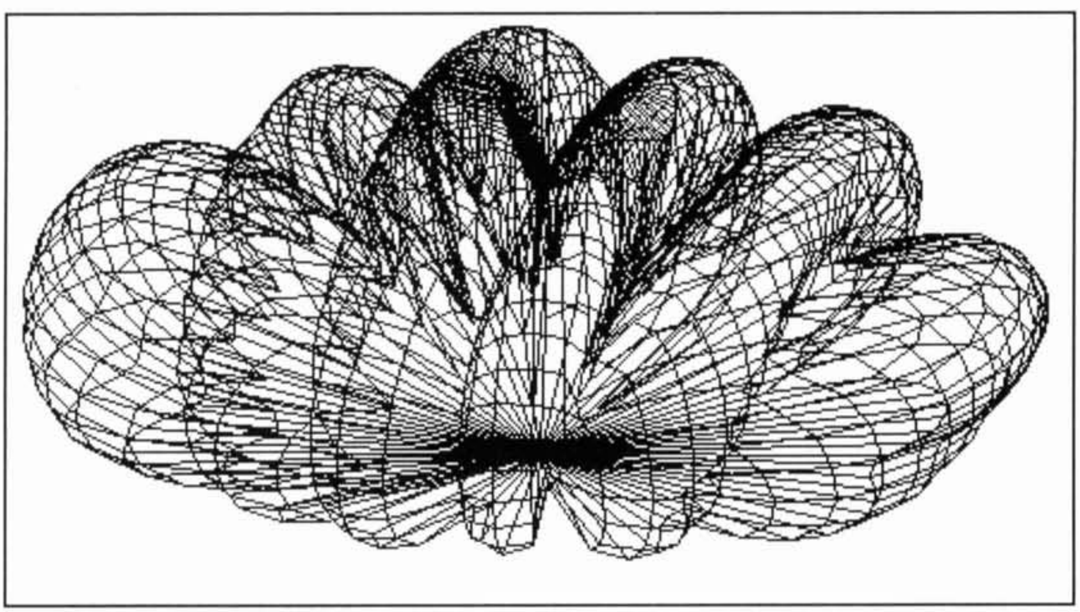


Figure 5. Three-dimensional radiation plot as calculated from azimuth and elevation data.

your antenna and its orientation to the X, Y, and Z axes and a user-specified ground plane (if present).

If the configuration looks satisfactory (e.g., you haven't confused the coordinates and created an accidental sloper), you run NEC and view another spreadsheet-like form that lists the display patterns you've selected.

From there, you invoke a plotting routine that generates radiation pattern plots from the data. **Figure 3** shows an azimuthal plot, and **Figure 4** an elevation plot for AA1IP's too-low, off-center-fed dipole at a frequency of 14.25 MHz. **Figure 5** shows a three-dimensional radiation plot for the antenna. You can include additional plots within a view to show variations in radiation pattern versus frequency.

If the routine listed above sounds complex, it's actually much easier to perform than to describe. NEC-Win Basic includes a well-written and clearly illustrated manual consisting of 130 8-by-11 inch pages bound with a lie-flat plastic comb binding. Chapter 2 offers a walk-through dipole antenna example, followed by examples of a tri-band auto antenna, a two-wire vertical, and a log-periodic beam in Chapter 3.

Chapter 4 is an extensive reference section that includes more detailed explanations of how to set up more complex calculations and add transmission-line and lumped-element loads to antennas. A brief appendix describes antenna-modeling basics. Aside from a few minor typos, NEC-Win Basic's manual stands as one of the better user's manuals I've encountered.

I found NEC-Win Basic a pleasure to use, both in terms of ease of entering antenna parameters and availability of easy-to-understand output data. When I encountered weird results, I typically found I'd made errors in data entry or coordinate specification—problems revealed by examining NecVu's version of what I'd entered. However, NEC-Win Basic would benefit from addition of error messages.

I somehow added a second, invisible and illegal source to an antenna—provoking a cryptic error message and a call to Paragon's technical support. My call was answered promptly, and a support person walked me through the problem, promising to examine possible improvements in user error-trapping in future versions.

To run NEC-Win Basic, you'll need a 386 or higher PC equipped with a math coprocessor, 4 MB or more of RAM, VGA graphics, MS-DOS version 5.0, and Windows 3.0 or higher. Paragon recommends a minimum of 10 MB of free hard-disk space, but the program installed in a little over 6 MB on my machine.

If you've held off exploring antenna-analysis software because of its complexity or expense, NEC-Win Basic allows you neither excuse. A few hours spent calculating your new skywire's pattern before you install it will pay off in better QSOs and less time teetering on shaky ladders.

Purchasing information

Where to buy products mentioned in this column:

You can purchase Visio Technical from most software suppliers, or from Visio Corporation, 520 Pike St., Suite 1080, Seattle, Washington, or phone (206) 521-4500.

Order Hamcall from Buckmaster Publishing, Route 4, Box 1630, Mineral, Virginia 23117, or phone (800) 282-5628.

To buy NEC-Win Basic for Windows, contact Paragon Technology, Inc., 200 Innovation Blvd., Suite 240, State College, Pennsylvania 16803, or phone (814) 234-3335. ■

PRODUCT INFORMATION

HP Accessory Provides Portable Power Choices

A new accessory for use with the HP Basic Instruments line draws operating power from the cigarette lighter of a car or from an external 12-volt battery.

The HP34397A DC-to-AC inverter, makes it possible to use instruments on a portable calibration cart that travels throughout a plant, to make measurements at remote locations where no AC line is available, and to keep instruments powered and ready to use while in transit between sites.

Specifications include: 10.5–15 volts input voltage; –115Vac, 60 Hz, ±0.06 Hz output voltage; 100 watts output power; and quasi-sine wave output waveform.

The HP34397A DC-to-AC inverter is available for $160. The unit is supplied with a manual in English, French and German and a set of battery-clip leads. The 230-volt option OE3 comes with an IEC-to-IEC power cord to connect the instrument to the inverter. The HP34397A is covered by a limited three-year warranty.

For more information, contact Hewlett-Packard Company, Direct Marketing Organization, P.O. Box 58059, MS51L-SJ, Santa Clara, CA 95051-8059; or phone 1-800-452-4844 ext. 9800.

Phil Anderson, *WØXI*
Kantronics, Inc.
1202 E. 23rd Street
Lawrence, Kansas 66046

FACTORS IN HF-ARQ SYSTEM THROUGHPUT

Frame length, protocol, and forward error correction coding

By experience, HF packeteers have learned to reduce packet length in the presence of noise or interference in order to get frames through. Like any system, 300 baud (HF) packet works well when the band is open and activity is low. However, as the reception bit error rate (BER) increases to 1 in 1,000 (0.001) or more, standard length packets (128 bytes) begin to fail and throughput of 64-byte packets is spotty, too. This fact wasn't lost on the designers of Clover, Pactor, or G-TOR.* In fact, the generally poor performance of HF packet, in the presence of a high BER, gave rise to these other modes.

While errors in HF communication are only somewhat random, we can gain some insight about throughput versus frame length by *assuming the errors are random*. Consequently, if errors are due entirely to additive white Gaussian noise (AWGN), we can calculate the probability of receiving a frame of any chosen length, given a specific BER. For example, suppose that a system with 100-bit frames exhibits a BER of 0.001. The probability (P) of receiving that frame correctly is:

$$P = (1\text{-BER})^{100} \qquad (1)$$

Figure 1 repeats and graphs this equation for *frames of 100, 300, and 500 bits for a BER* from 0 to 0.01. Note that even at a BER of 0.5 percent, the 100-bit frame has just a 50 percent chance of being received correctly. Further, the chances of receiving the longer frames at this BER are drastically reduced.

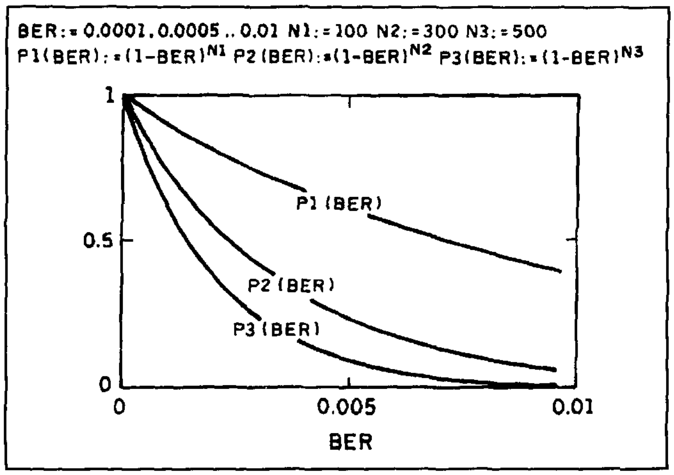
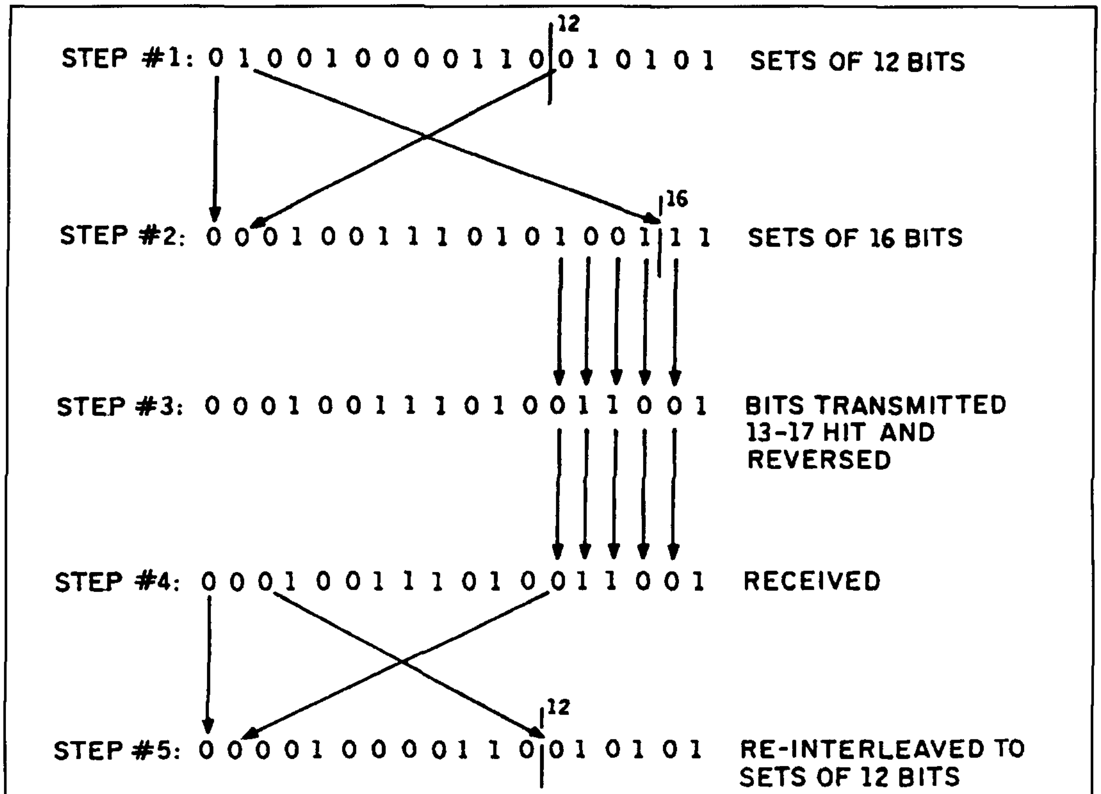


Figure 1. Probability of successful frame reception versus BER, for several frame sizes. No ARQ cycle is assumed.

No wonder HF packeteers reduce packet length (paclen) to 64 or less! With paclen set to 64, HF packets contain 512 data bits and 160 overhead bits, and this *still* exposes the frame to noise "hits" for about 2.23 seconds. For an equivalent number of characters, AMTOR would expose a "set" of 3-byte frames to hits for roughly 4.88 seconds! By combining AMTOR and packet techniques, Pactor reduces frame exposure, as compared to packet, by shortening the duration of the frame transmission to 0.96 seconds. However, AMTOR, packet, and Pactor are all Type-I ARQ systems; that is, these systems can detect errors in a frame but do not cor-

*G-TOR is a trademark of Kantronics Co., Inc., Clover is a trademark of Hal Communications, Inc., and Pactor is a trademark of SCS, Germany.



With G-TOR at 100 baud, the interleaving process starts with 192 bits per frame, or 16 sets (bins) of 12 bits each. These data are then interleaved, transmitted, and reinterleaved back to sets of 12 bits in 16 bins. However, any burst errors received during transmission are now distributed as single-bit errors throughout the 16 bins.

Step One: Data to be transmitted is divided into a set of 12 bits.

Step Two: These data are then interleaved; bit 1 goes to bin 1, bit 2 goes to bit 1 of bin 2 . . . bit 13 goes to bit 2 of bin 1, and so on.

Step Three: Transmission example, the interleaved data is transmitted, and bits 13-17 are hit and reversed.

Step Four: These data are received and stored in bins of 16 bits.

Step Five: The frame is reinterleaved; bit 1 goes to bin 1, bit 2 goes to bit 1 of bin 2 . . . bit 17 goes to bit 2 of bin 1, and so on.

Step Six: With the noise burst distributed, only one bit error now resides in five of several 12-bit bins. These single-bit errors in 12 bits are then corrected using the Golay Scheme.

Figure 2. G-TOR interleaving at 100 baud

rect errors by using error correction codes. They must continue to request a frame until it is received correctly. Pactor (Pactor-I), while not using error correction codes, does attempt to produce a good frame by adding consecutive frames together—a technique called memory-ARQ. Hence, it's clear that with a BER above 0.001, systems with long duration frames will suffer. At the same time, shorter-frame systems suffer, too, since time is wasted in switching back and forth between transmission and reception and too many bytes are used for control rather than data.

A partial solution to this dilemma is to create systems with longer frames and add error correction coding. By using data interleaving and an error correction code, bit error rates in excess of 3 percent can be tolerated. Such systems are called Type-II, or Hybrid ARQ. Clover and G-TOR are the first true instances of this type of system within the amateur community.

Adding error correction

Two good choices for an error correction

AMTOR	/7 bits/7 bits/7 bits/, 21 data bits, 210 ms frame duration, 0.45 sec cycle
PACTOR@200 baud	/flag/data/C/FCS/, /8 bits/192 data bits/8 bits/16 bits/, 192 data bits, 0.96 sec frame duration, 1.25 sec cycle
HF Packet	@ 300 baud, paclen=64 bytes: /flag/14-70 bytes address/C/PID/data/FCS/flag/, or /8 bits/112 bits/8 bits/ 8 bits/512 data bits/16 bits/8 bits/ 512 data bits, 2.23 sec frame duration, no cycle defined
G-TOR @300 baud	/data/C/FCS/ /552 bits data/8 bits/16 bits/, 552 bits data, 1.92 sec frame duration, 2.4 sec cycle

Table 1. Frame protocol, popular ARQ systems.

code for HF are the Golay and the Reed Solomon (RS) codes. G-TOR, named for the Golay code, is based on the Golay error correction scheme. Both codes are classified as block codes; that is, a fixed number of parity bits are coded for each fixed set of data bits. These codes correct errors well when the errors are random; for this reason, it's necessary to interleave the data frames for HF systems before and after transmission. By doing so, the bursty, multiple-bit hits during reception are made to look random. The interleaving process is outlined in **Figure 2**.

Frame makeup

While frame duration and error correction coding affect throughput in these systems, the number of control and data bytes within the frames also makes a difference. The frame protocol for AMTOR, Pactor at 200 baud, HF Packet (300 baud), and G-TOR at 300 baud are listed in **Table 1**.

Throughput

Assuming an ideal channel (no errors), the throughput for these systems is tabulated in **Table 2**. For example, Pactor (Pactor-I) at 200 baud, its highest rate, transmits 160 bits of information during a 1.25 second cycle; hence, its ideal throughput is 128 BPS. For packet, assuming 64 bytes of data are contained in a packet at 300 baud, its ideal rate is 146 BPS. Finally, G-TOR transmits 552 data bits during its 2.4 second cycle, so its ideal throughput is 230 BPS.

In the presence of a high BER, it's clear from **Figure 1** that systems with long frames but no error correction fail to maintain a good throughput. However, by adding error correction coding, a good throughput can be maintained—even at a BER of up to 4 percent, as indicated in **Figure 3**. Note in the graph that G-TOR's throughput falls to 50 percent of the ideal, as soon as the BER exceeds a fraction of 1 percent. However, it maintains that rate until

Mode	data bits/f	bits/f	frame efficiency	ideal throughput BPS
AMTOR	21	21	100%	46
Pactor 100 baud	64	96	66%	51
G-TOR 100 baud	168	192	87%	70
Pactor 200 baud	160	192	83%	128
Pactor 300 baud	512	672	76%	146*
G-TOR 300 baud	552	576	96%	230

*Assumes a pseudo-cycle time of 3.5 seconds.

Table 2. Throughput, HF-ARQ systems.

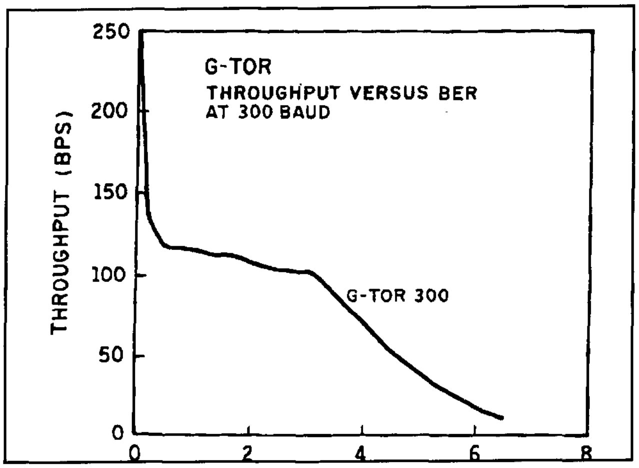


Figure 3. Throughput versus BER at 300 baud for G-TOR.

it reaches a BER of nearly 4 percent! Without error coding protection, as indicated in **Figure 1**, G-TOR's throughput with its 576-bit frame would have dropped to roughly 10 percent with a BER of as little as 0.5 percent!

Conclusion

While HF packeteers have learned to reduce paclen in order to increase the chances of getting frames through, we see that shortening frames somewhat doesn't help much as the BER exceeds 0.001. At the same time, short frame systems such as AMTOR "churn away" their time resources by switching back and forth between transmission and reception. Making ARQ system frame protocols more efficient, increasing the ratio of data bits to control bits, helps throughput some, but this scheme fails too as the BER gets heavy. As a result, the only way to overcome the time wasted by short frame systems and the hits that will surely occur with longer frame systems is to add error correction coding to a long frame system. ARQ systems without frame protection will simply fail with a BER approaching one bit per frame. At the same time, ARQ systems that incorporate forward error correction coding are capable of tolerating error rates of 4 percent or more. ■

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HF Packet: Some editorial comments

Present packet radio protocol originated in the line-of-sight propagation world. At the time, a fundamental assumption was that most everyone could receive most all of the signals. While packet radio has been reasonably successful at VHF/UHF, its extension to HF has not. W0XI's article explains some of the technical reasons why this is so, but there are sociological problems, too.

Old-time RTTY types can probably remember the courteous adherence to informal subband agreements between AMTOR and RTTY. Unfortunately, even before packet incursions started, these arrangements began to unravel with the arrival of 24-hour automatic mailboxes that took a proprietary view towards certain frequency spots.

We now have some fairly modern digital communications protocols, and they are proliferating rapidly. However, many of their relative advantages are too strongly defended with inappropriate "white noise" analysis. Meanwhile, the real concern is that while we search for excellence, we are creating a Tower of Babel.

HF digital communications such as Clover, Pactor, or G-TOR *are* advances; but this proliferation is hurtful if it further divides the amateur radio community. Maybe we need to functionally compare these systems. There are several questions we could consider. What is good for short QSOs; what is good for long data streams? Should we suggest sub-bands for functions as well as protocols? A few simple changes may make digital communications more enjoyable and user-friendly to newcomers.

HF packet's continued inefficiency and wide bandwidth casts doubt on its entitlement to continue making universal noise everywhere. RTTY is still relatively efficient. It easily fits into less that 500 Hz bandwidth and is a worldwide protocol. *But think of how many subbands of modern protocols we could fit into all those Fat Packet Bandwidths!*

Hunter Harris, *W1SI*
Communications Quarterly
Editorial Review Board

Receivers

Our fall 1994 "Tech Notes" column highlighting the commutative H-mode FET mixer by Colin Horrabin, G3SBI, generated a bit of reader interest in seeing further material on his quest for the ultimate high-dynamic range receiver. In this edition of "Tech Notes," you'll find a low-noise AGC-controlled post-mixer IF amplifier design, and a treatise on low-noise oscillators by G3SBI. Our third feature deals with designing stable LC oscillators, and is annotated with many practical examples of the circuits discussed.
Peter Bertini, K1ZJH, Senior Technical Editor

Low-noise AGC-controlled IF Amplifier

Reprinted from "Pat Hawker's Technical Topics," Radio Communication, May 1995

Several times in the past, TT has emphasized, when discussing modern approaches to the design of receivers with super-linear front-ends, the importance of the IF amplifier that immediately follows a mixer having conversion loss rather than gain. For example, in TT, February 1993, I stated: "It should be noted that with any mixer operating directly on the incoming RF signals without premixer amplification, the IF amplifier that follows the mixer, either directly or after a roofing filter, must have a low noise figure and a high intercept figure, with a diplexer arrangement often used to achieve constant input impedance over a broad band of frequencies."

	16 dB gain	**13 dB** gain
Noise figure	0.6 dB	.6 dB
Input impedance	50 ohms	50 ohms
Output impedance	50 ohms	50 ohms
Third order intercept		
Vs = +12V (max gain)	23 dBm	26 dBm
Vs = +20V (max AGC)	28 dBm	30 dBm
Input for 1 dB compression		
Vs = +12V (max gain)	0 dBm	+3 dBm
(max AGC)	7 dBm	+11 dBm
Vs = +20V (max gain)	+5 dBm	+8 dBm
(max AGC)	+11 dBm	+14 dBm

Amplifier input and output impedances are 50 ohms regardless of AGC-controlled gain. Gain range 45 dB.

Table 1. Performance of 9-MHz AGC-controlled amplifier.

A simplified diagram was included of a grounded-gate FET amplifier capable of providing a 2-dB noise figure, 9 dB gain in a 50-ohm system for a 45, 70, or 100 MHz up-converted first IF stage. It was reproduced from the IEE's *Radio Receivers* book edited by Dr. William Gosling.

Colin Horrabin, G3SBI, recently sent details

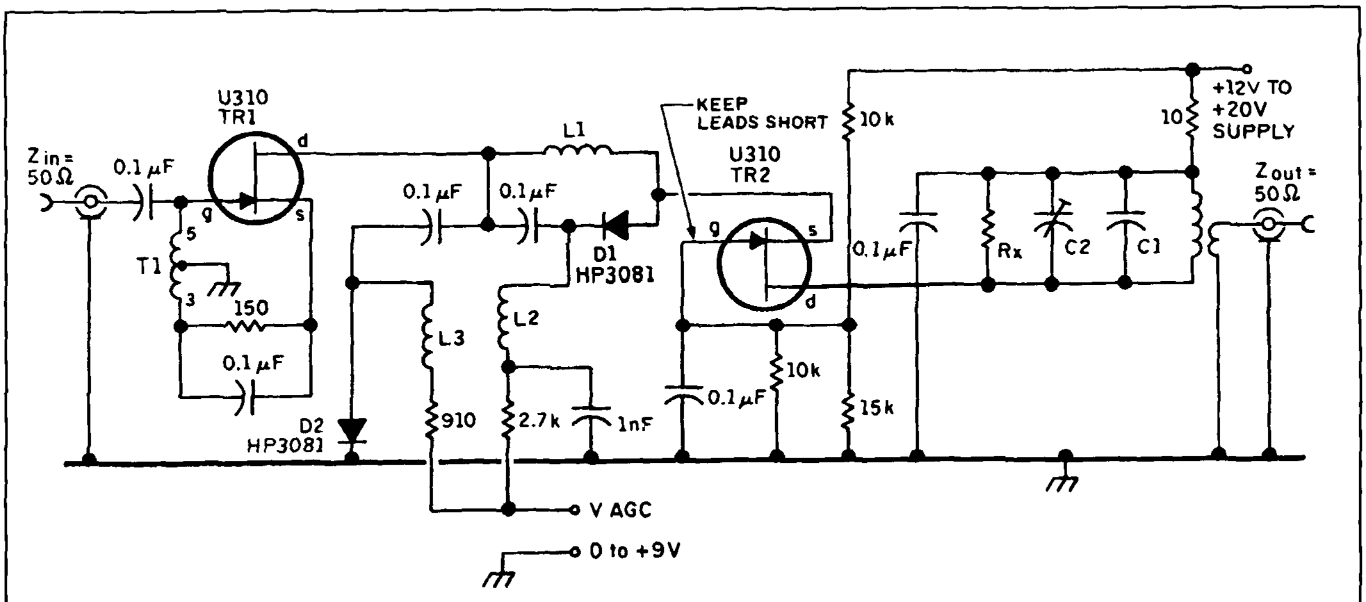


Figure 1. G3SBI's low-noise, AGC-controlled cascode IF amplifier.

of a 9-MHz cascode-FET amplifier that he proposes to use immediately behind the roofing filters in any receiver using his superlinear H-mode mixer (see TT, October 1993, the new edition of *Radio Communication Handbook*, or *Communications Quarterly*, Fall 1994). He points out that this could also be a useful approach for anyone constructing high-performance receivers because of the low noise figure, reasonable gain and intercept point, and the 50-ohm input and output impedance that remains constant over the 45-dB AGC range.

The circuit diagram of the 9-MHz version of this amplifier is given in **Figure 1**. Note that the AGC amplifier must be capable of sinking the current through D1 at 0 volts (i.e., maximum gain). The warning to keep leads short in the drain circuit of the second U310 FET arises from G3SBI's experience in finding the initial IP3 measurements were poor due to this stage oscillating around 400 MHz. Performance is given in **Table 1**.

Component notes: The two U310 are Siliconix low-noise JFETs. C1 is an 82 pF ceramic, C2 a 60 pF ceramic trimmer (Cirkit), and all other capacitors are monolithic ceramic (RS Components). Resistors are 1/8th watt metal film (RS Components). D1, D2, are HP3081 PIN diodes (Farnell). T15+3 turns of 0.224 mm diameter Bicelflux enamel on Fairite Balun core 28-43002402 (Cirkit). T2 (primary) 2.81 μH, 31 turns of 0.314 mm Bicelflux enamel on Micrometals toroid T37-6 (Cirkit). T2 (secondary) (1) for 16 dB gain 3 turns, Rx 8k2; (2) for 13 dB gain 4 turns, Rx 3k9. Note that (1) and (2) could be relay switched for use with an SSB or CW filter (loss 10 dB or 3 dB). L1, L2, and L3 7 turns 9.314 mm enamel on balun core 28-43002402 (Cirkit).

Towards The Superlinear Receiver: Low-Noise Oscillators

Reprinted from "Pat Hawker's Technical Topics," Radio Communication, July 1994

It was emphasized in the January TT (page 39) that if full advantage is to be taken for high-dynamic range receivers of the latest super-linear mixers, such as G3SBI's H-mode FET-array mixer, there remains a need to produce free-running or preferably synthesized oscillators having an extremely low phase noise of the order of 150 dBc/Hz or better, a few kHz off the tuned frequency.

Colin Horrabin, G3SBI, has progressed significantly towards this target for a free-running oscillator although he recognizes that to design and build a complete digitally synthesized local oscillator based on his preliminary results is still a major undertaking. However, he feels that information at an early stage may encourage some readers to take these developments

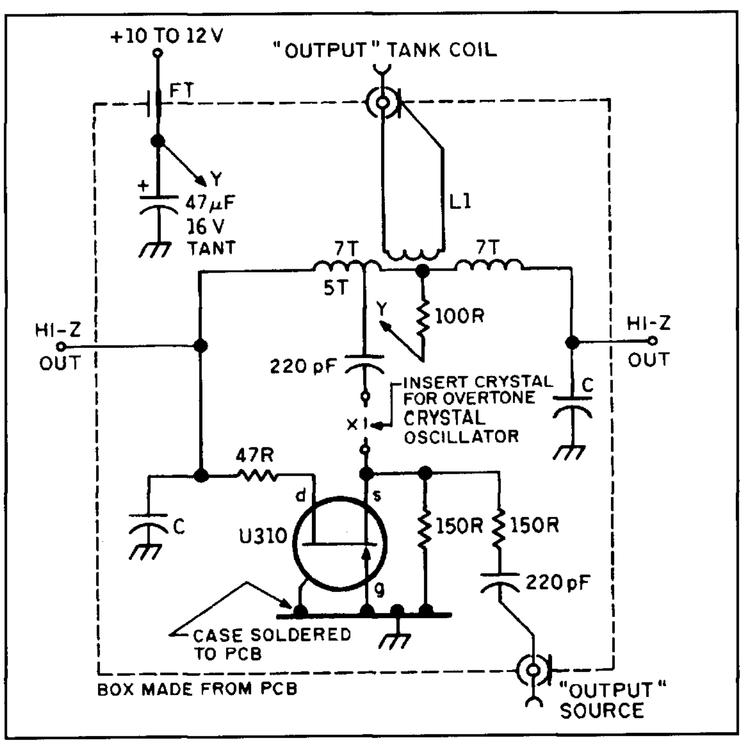
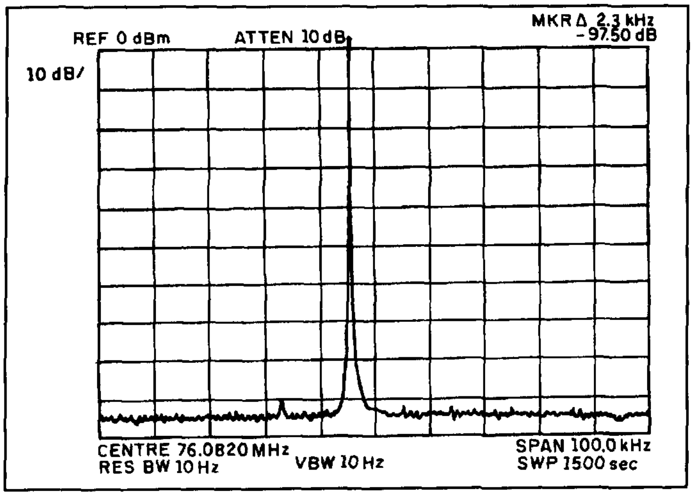


Figure 1. The initial prototype grounded-gate, low-phase-noise oscillator using lumped wound coils.

Figure 2. Typical free-running spectra of the original oscillator at 76 MHz.



further and faster than his own other commitments permit.

He writes: "These present notes cover an oscillator suitable for use as a VCO (voltage-controlled oscillator) or as an overtone crystal oscillator that has low inherent phase noise. The initial measurements were made on an oscillator operating over 20 to 80 MHz using conventionally wound inductors, but subsequently it has been found that a version using stripline inductors against a groundplane shows the most promise. This form of construction

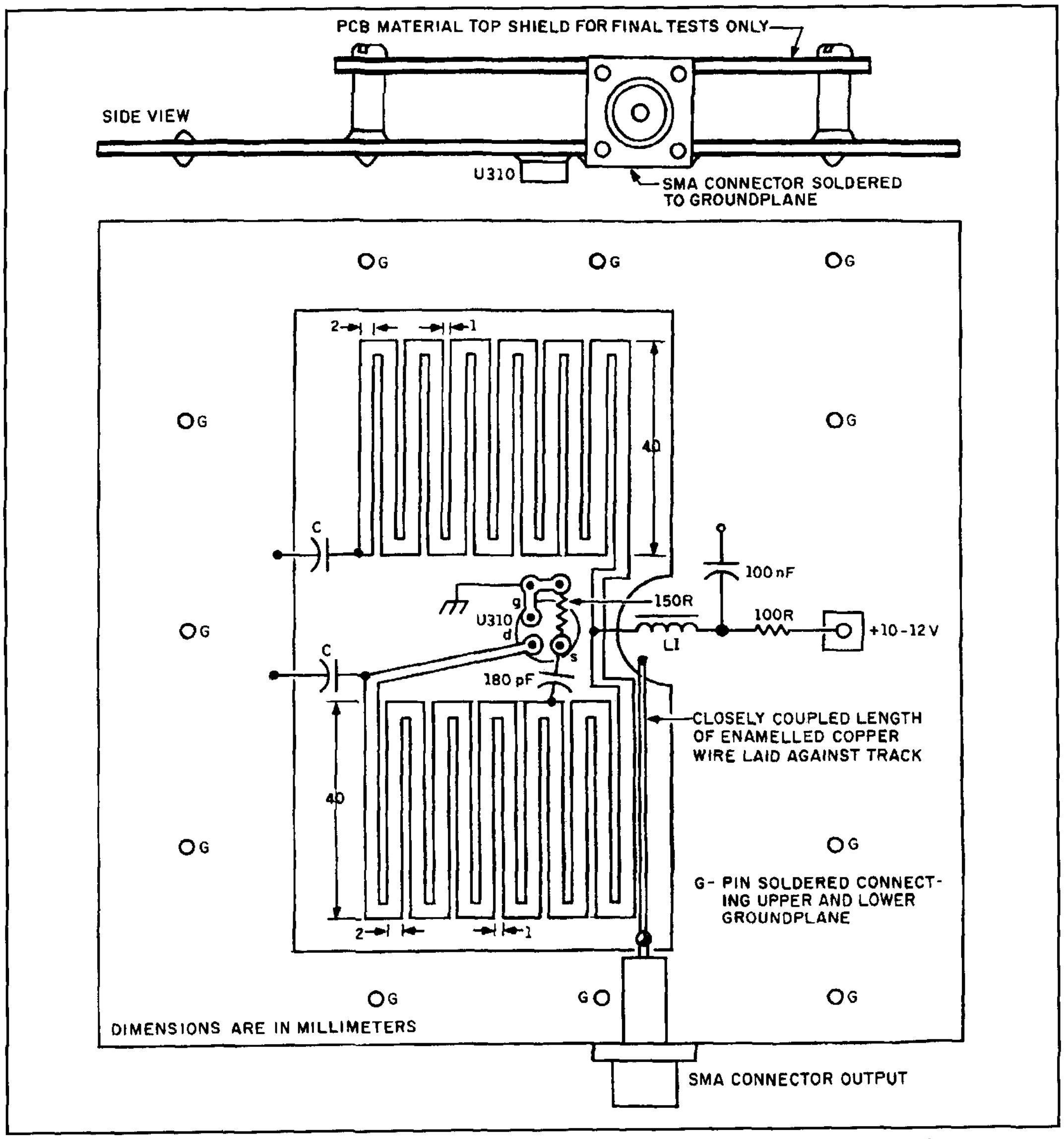


Figure 3. Prototype strip-line oscillator capable of providing an operating range of 3.5:1 before oscillation ceases. Length A: 40 mm, width B: 2 mm, width C: 1 mm. To change the frequency range, alter the length A and the length of the output coupling wire. Typical frequencies with C = 0 pF about 145 MHz, 27 pF about 80 MHz, 56 pF about 60 MHz, and 150 pF about 36 MHz (plate-ceramic capacitors). With 1/16-inch double-sided pc board, note that the bottom ground plane covers the whole of the board. Output about 0 dBm. The U310 FET is pushed through from the bottom side of the pc board until it touches the ground plane and is soldered directly to the groundplane.

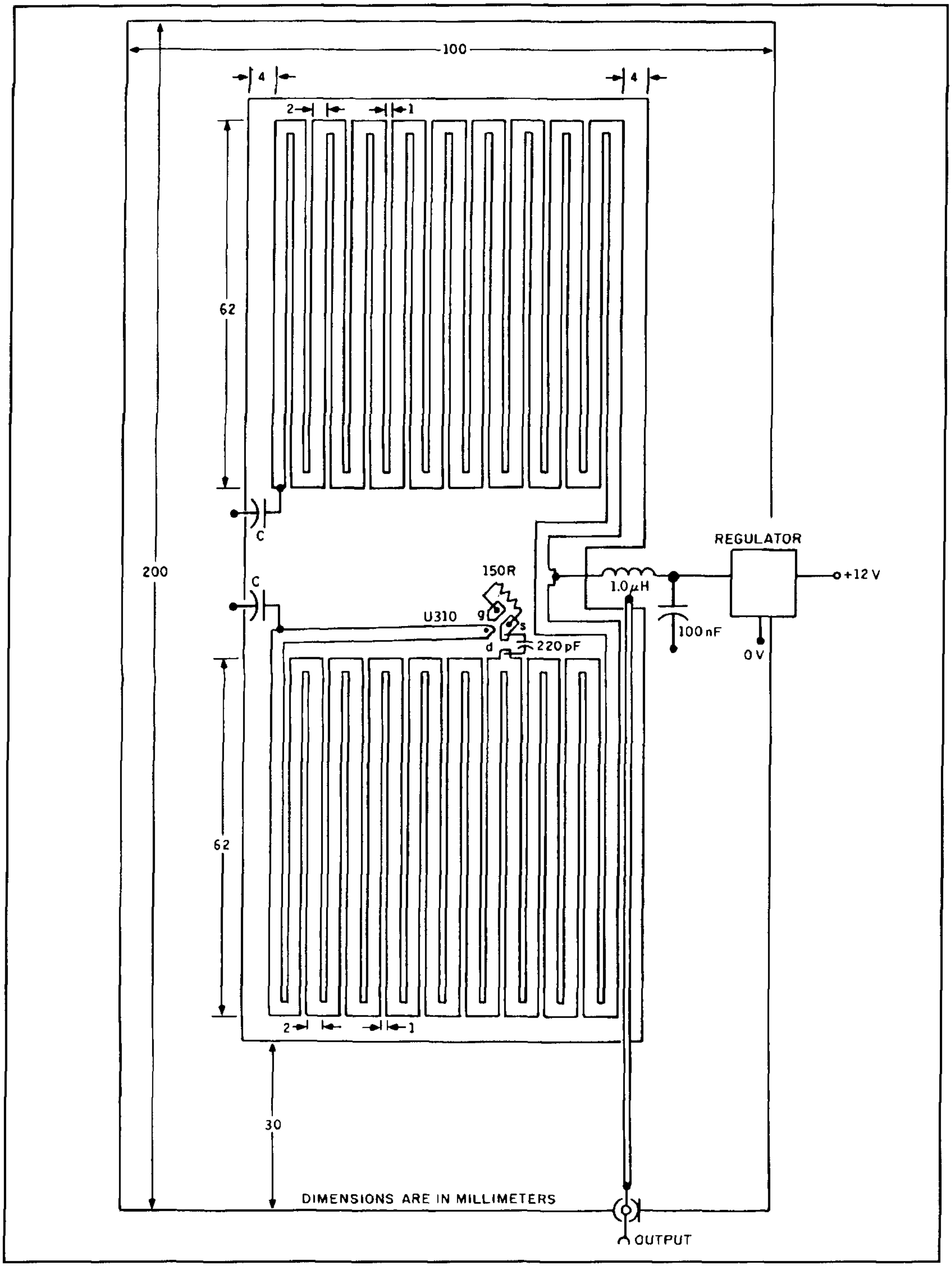


Figure 4. Later version of the stripline pc board art.

has been found to cut dramatically the sensitivity to stray capacitance fields. The prototype stripline version operates in the 36 to 150 MHz region (capacitance 0–180 pF) or 82 to 146 MHz (0–27 pF), but changes to the length of the line could give a 3.5:1 operating frequency range anywhere between 20 and 450 MHz. The stripline artworks were produced by my SERC colleague, Alex Macdonald, on an Apple PC using MacDraw software with output onto a laser printer.

"**Figure 1** shows the original design using coils. It was constructed inside a box made from double-sided pc board material. Output at 50 ohms impedance and 0 dBm signal level for a spectrum analyzer can be taken either from the FET source or via a two-turn link winding on the main tank coil, L1. These are terminated in SMA connectors. The oscillator comprises a U310 junction FET in grounded-gate configuration."

The two capacitors marked C must be of the same value. Components are as follows: C Suflex or plate-ceramic capacitors between 15 and 220 pF (Cirkit Components). L1: 7 + 7 turns 6 mm diameter enamel for about 20 to 80 MHz. When used as a crystal oscillator, the oscillator is first used free running on the crystal frequency. One of the capacitors, C, is made variable so the frequency can be tuned. The crystal is inserted after breaking line X and on the C adjusted for oscillation. Output from the source has slightly improved sidebands. For VXO use, the 220-pF capacitor can be varied to give about 5 kHz variation at 53 MHz. As a crystal or free-running oscillator, outputs of about 0 dBm are available from the source or tank. A high-impedance output of about 10 volts peak-to-peak can be taken as shown. The two Hi-Z outputs are 180 degrees out of phase.

The grounded-gate happens to be the case of the FET, so a hole is drilled in the groundplane and the FET pushed through with the case soldered to ground. A 47-ohm resistor is necessary in series with the drain lead (but not in the stripline version) to control a 500-MHz parasitic oscillation.

"L1 is a split coil; this is fundamental to the design in order to avoid the use of a bypassing capacitor to ground at its center. The other half of this coil acts as a series-tuned circuit and therefore presents a low impedance at resonance to ground at the DC feedpoint. If necessary, push-pull signals at a high impedance can be taken between the ends of L1 to ground; C1 and C2 are identical values and must be changed together. Varicap diodes, if used, must be matched, since any significant difference in value of C1 and C2 will inhibit oscillation.

"Some tests were made using a 53-MHz third-overtone crystal inserted between points. Noise sidebands were down 98 dB at 200 Hz from center frequency.

"Typical free-running spectra of the original oscillator at 76 MHz are shown in **Figure 2**, about 98 dB down at 2 kHz. Contrast this with the superior stripline circuit (**Figures 3** and **4**) at 82.9 MHz (**Figure 5**) and 201 MHz (**Figure 6**). Most of the close-in phase noise noted in a run at 62 MHz is due to FM at 50 kHz. This is presumably due to AC mains inducing fields in the tank coil causing low-level AM and FM modulation of the FET. In a closed loop of any useful bandwidth, these would be suppressed.

"So how good is the oscillator phase noise? It is visually superior to two of our professional high-grade synthesized signal generators, but without access to a Hewlett Packard phase-noise measuring system or other methods of

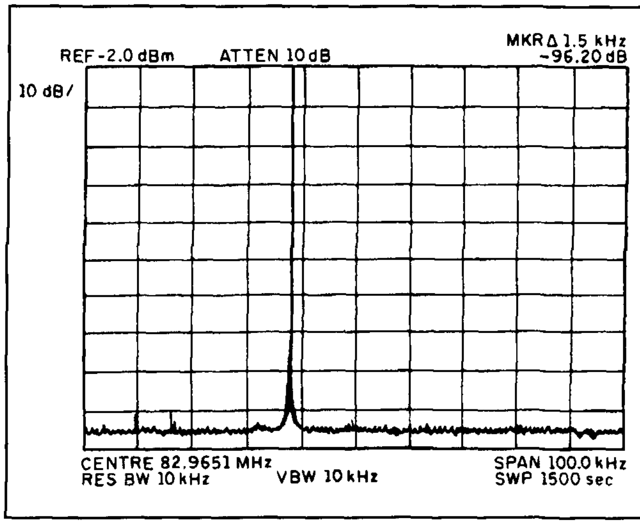


Figure 5. Spectra of stripline oscillator at 82.9 MHz.

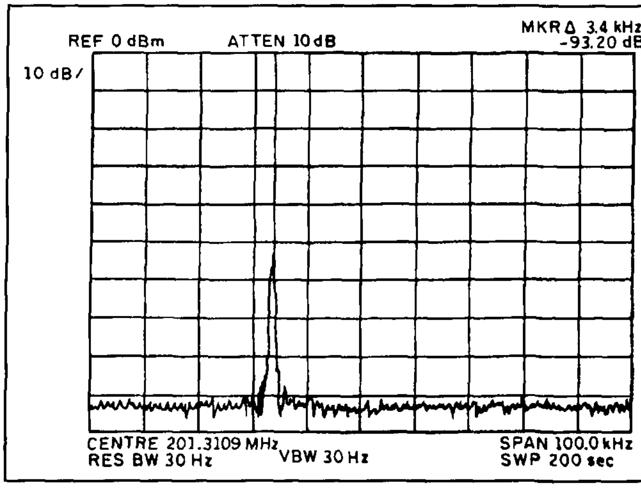


Figure 6. Spectra of stripline oscillator at 201 MHz.

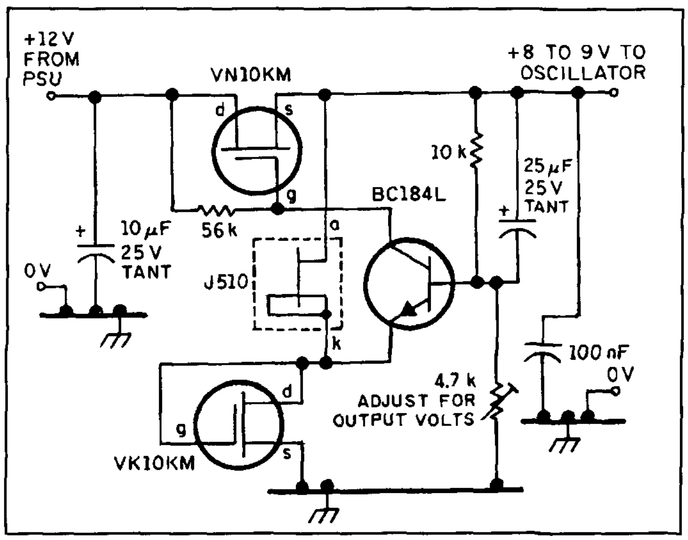


Figure 7. Low-noise voltage regulator as used for the G3SBI low-noise oscillator.

measurement more suited to home construction, exact figures can not yet be given. Nevertheless, it is clear that the oscillator shows great promise. It is intended to use one or more stripline oscillators in the design of a low-phase-noise synthesized local oscillator system when time permits.

"Two further stripline oscillator layouts have been implemented for the frequency range 20 to 50 MHz and are being tested. One of these uses thinner and closer tracks with the object of reducing the physical size of the pc board. A practical board layout for the local oscillator of a receiver would probably have the two striplines back-to-back on two separate pc boards so that a band-change wafer switch (perhaps driven by a stepper motor) would be simple to accommodate.

"Finally, it needs to be stressed that the design of the voltage regulator is important in order to obtain results as good as, or better than, a 9-volt battery as the oscillator power source. The best spectrum results have been achieved with a regulator output of 8 to 9 volts. Conventional IC regulators slightly degrade the oscillator noise floor, particularly close-in. The simple regulator circuit shown in **Figure 7** gives good results and was used during the measurements shown, but could probably be improved with a little more loop gain. In this design, the regulator voltage reference is a high gm FET type 10KM driven by a J510 JFET current source and has low-noise characteristics.

Stable LC Oscillator

Reprinted from "Pat Hawker's Technical Topics," Radio Communication, November 1994

Despite the attractions of crystal resonators in conjunction with frequency synthesis as a means of obtaining stable frequency operation, there remains a real need for free-running LC-type VFOs with stabilities approaching those of low-cost crystal oscillators. The search for improved stability of LC oscillators began in the 1920s and continued in the 1930s and 1940s. By the mid-1950s, virtually all the basic requirements needed for reasonably stable LC oscillators were understood and suitable circuits developed for use with thermionic valves. Since then, there have been few major developments with most work focused simply on adapting the proven valve circuits for use with solid-state devices, although much more attention has been paid to oscillator noise since the publication in 1953 of the book *Vacuum Tube Oscillators* by W.A. Edson.

One of the earliest oscillators that provided good stability on HF was the Franklin oscillator developed in the 1920s (**Figure 1**). This used two active devices, providing sufficient gain in the amplifier/phase inversion section to permit very loose coupling to the frequency-determining high-Q LC resonant circuit. The Franklin has several other advantages including the connection of the tuned circuit directly to earth and the use of a two-terminal coil. Curiously, the Franklin oscillator, developed for the Marconi Short Wave Beam system, has never been widely used in the U.S. where engineers continue to investigate single-valve circuits based on variations of the Hartley and Colpitts oscillators.

The early 1930s, with the coming of higher-gain tetrode and pentode valves, saw the introduction by J.R. Dow (*Proceedings of the IRE*, Vol. 19, 1931, pages 2095–2108) of electron coupling within the valve in conjunction with either Colpitts or Hartley oscillators to provide an oscillator much less affected by variations in the HT supplies. The ECO took advantage of the fact that a drop in screen voltage can compensate for a drop in anode voltage. The ECO, as adopted by amateurs, also had the advantage that the tuned anode circuit from which output is taken can be at double the frequency of the frequency-determining resonant circuit, making it easier to achieve good stability.

The introduction of neon-stabilized voltage regulator tubes by the end of the 1930s also made it possible to achieve better stability

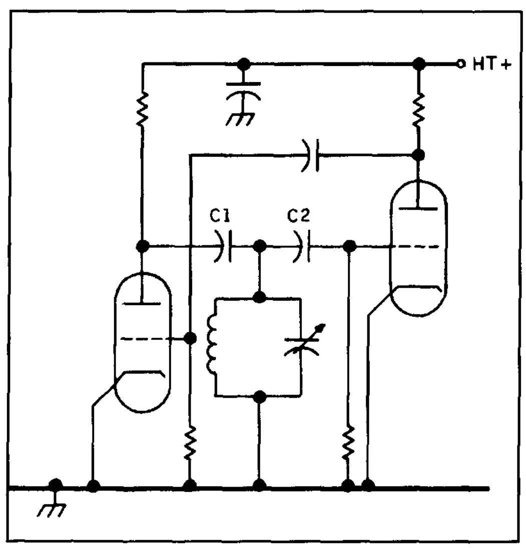


Figure 1. The basic Franklin Master Oscillator, first described in 1930, uses very low value capacitors C1, C2, (about 1 to 3 pF) imposing a very light load on the high-Q tuned circuit. Further advantages include the two-terminal inductance, and the earthing of one end of the resonant circuit. This form of oscillator can be readily adapted for use with MOSFET devices.

from basic Hartley and Colpitts oscillators without electron coupling, provided always that the components and circuit parameters were well chosen. It was soon recognized that a high-C tank circuit was particularly important, although this limited the frequency span of the oscillator; no problem for amateur bands, but a disadvantage for general-coverage receiver oscillators.

A simple, but important, modification to the basic Hartley circuit was described by A.F. Lampkin (Lampkin's Laboratories, Florida and also a radio amateur of 1924 vintage who wrote frequent article in *QST* in the 1930s) in "An improvement in constant-frequency oscillators" (*Proceedings of the IRE*, March 1939). Surprisingly little advantage has ever been taken by amateurs of his discovery that by simply tapping the grid connection down the coil, the influence of the active device can be reduced by a factor of ten or more (see **Figures 2** and **3**).

Lampkin's idea was, however, recognized by Walter Van Roberts of the RCA Patents Division, but also as W3CHO, the first person ever to describe a unidirectional close-spaced rotary Yagi beam antenna (*Radio*, January 1938, pages 19–23 and 173). In "The limits of inherent frequency stability" (*RCA Review*, April 1940), he concluded that to obtain optimum stability from an LC oscillator:

- Make the fundamental frequency as low as possible.
- Make the Q of the coil as large as possible at the fundamental frequency. This means that the coil should be as large physically as there is room for within the shield can, subject to clearance of at least half a diameter, as well as that the coil design should be good in other respects.
- Use the loosest couplings between the tuned circuit and the tube that will give the required output, and use a low enough bias resistor so that the effective transconductance in the oscillating condition is not seriously reduced.
- For the oscillator tube, choose one which has a high ratio of transconductance to capacitance fluctuations when operating at the required level.
- Keeping the oscillation strength constant, vary the ratio between the grid and plate couplings. The best ratio depends on the ratio between the capacitance variations of the grid and the plate.

These points remain as valid for solid-state devices as for valves. Walter Roberts also advocated as "tricks of the trade:" The use of temperature compensation; supporting the tuned circuit on a single rigid member to avoid bending and vibration of its parts; reducing the power taken from the oscillator as much as possible and preferably taking output at a harmonic frequency; supplying screen voltage from a voltage divider whose two portions have resistances forming the combination that best com-

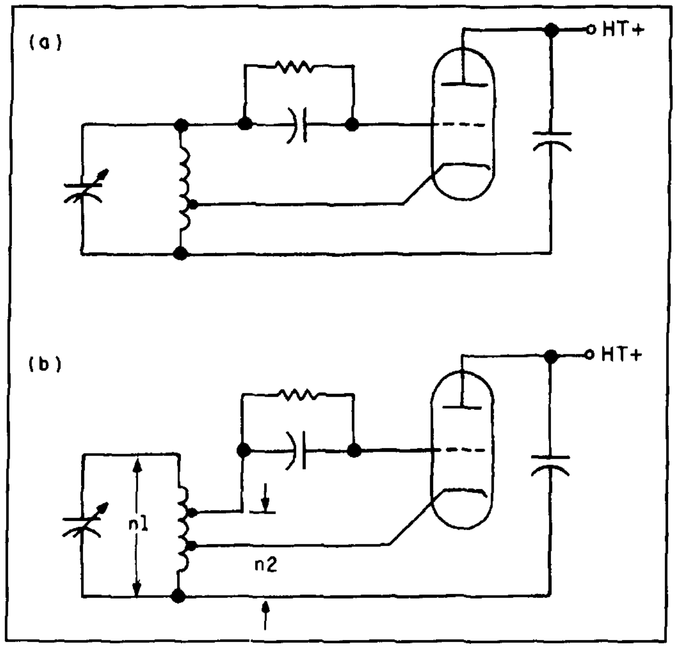


Figure 2. A.F. Lampkin in 1939 showed that a high-C Hartley oscillator (as typically used in the once popular ECO VFO) could be improved by a factor of about ten times simply by tapping the grid [gate or base] connection down the coil.

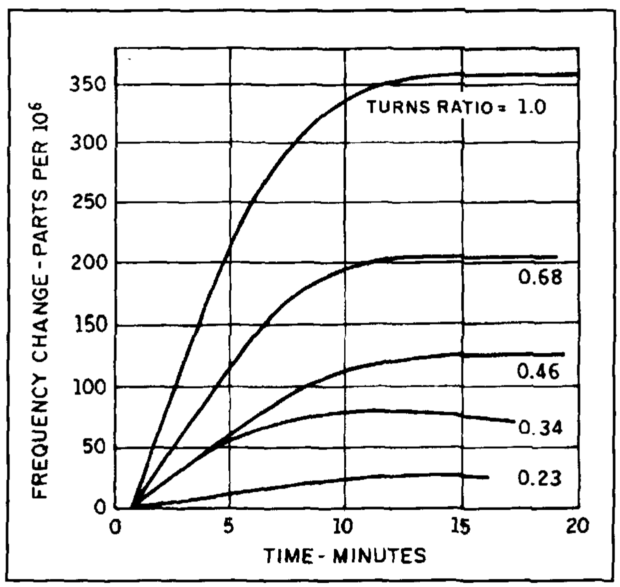


Figure 3. Showing how the long warm-up drift of a Hartley valve oscillator can be much reduced by tapping down the coil. Performance with various ratios of n1:n2 are shown.

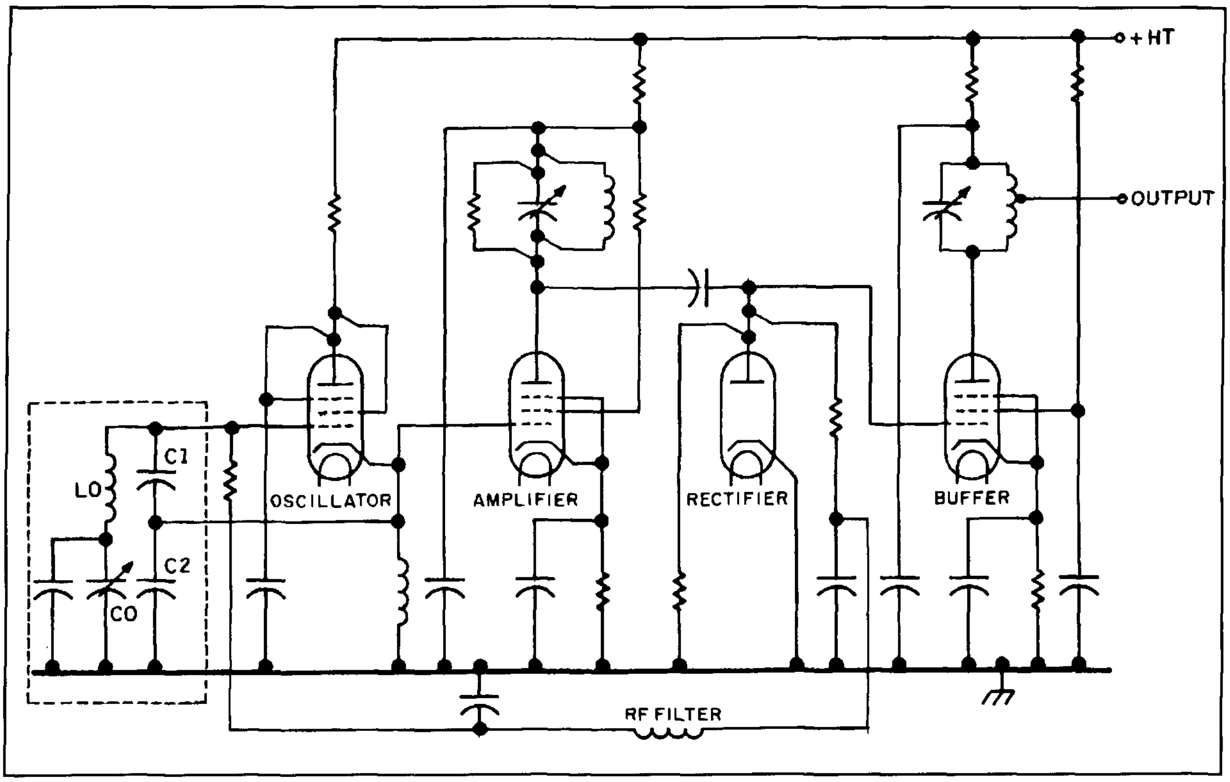


Figure 4. The BBC high-stability master oscillator developed by Geoffrey Gouriet about 1938, but not fully described by him until 1950. The stability obtained is, for all practical purposes, a function only of the parameters of the single tuned circuit at series resonance. The unit provided a one hour stability of the order of +/- one part in a million, and ten parts in a million over 24 hours. The same basic oscillator circuit was later developed independently by J.K. Clapp and published in 1948.

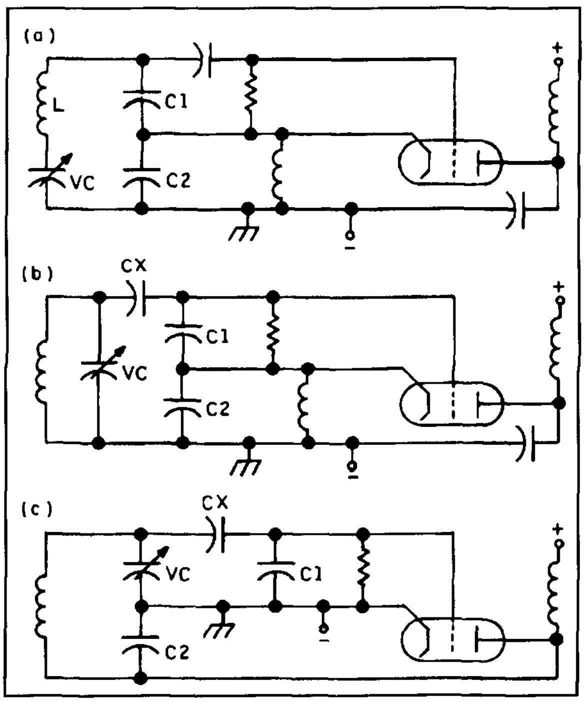


Figure 5. Three high-stability oscillator circuits showing the minor, but significant, differences between: (A) Gouriet-Clapp with series resonance; (B)Seiler low-C Colpitt's oscillator with parallel resonance. (C) The Vackar oscillator in which the ratios C2:CV and C1:Cx should both be about 1:6.

pensated for variations in supply voltage, and stabilizing the supply voltage.

About 1938, Geoffrey Gouriet of BBC Research developed a series-tuned form of oscillator (**Figure 4**) sufficiently stable to be used as a crystal-substitute for broadcast transmitters. Because of the war, full details of this were not published until J.K. Clapp of the General Radio Company had independently developed a similar circuit in 1946, details of which were published in "An inductance-capacitance oscillator of unusual frequency stability" (*Proceedings of the IRE*, March 1948). Clapp later recognized that his oscillator, quickly taken up and widely used by amateurs as the Clapp oscillator, should rightfully be called the Gouriet-Clapp oscillator. He also noted that the same form of oscillator has also been developed independently by O. Landini in Italy and described in *Radio Rivista* in 1948. A detailed paper by Gouriet was published in *Wireless Engineer*, April 1950.

E.O. Seiler, W8PK, (later W2EB) wrote in *QST* (November 1941) about a 3.5-MHz keyed VFO, which he described as a "low-C electron coupled oscillator" as an alternative to the popular high-C Colpitts oscillator and with a circuit arrangement similar to but different in

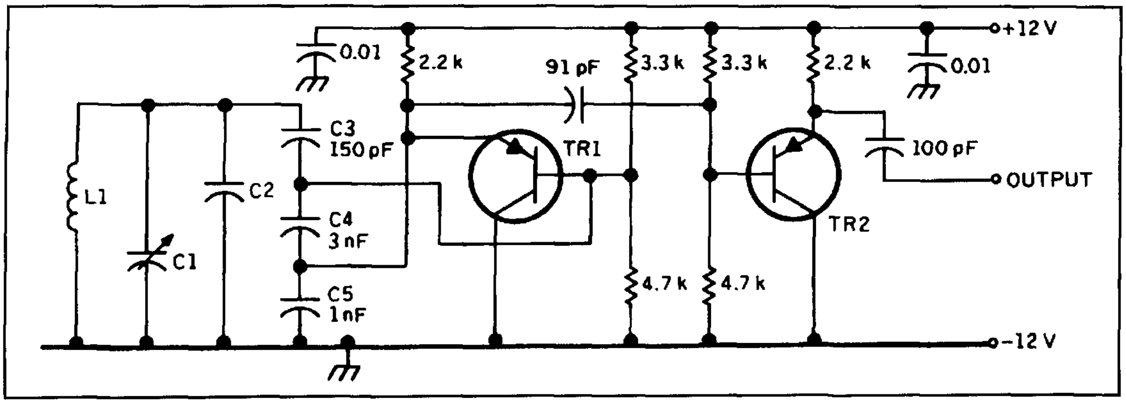


Figure 6. W3JHR's "synthetic rock" popular transistor VFO of the early 1960s was an adaptation of the Seiler oscillator. In the original design, W3JHR used high-quality components from the American surplus ARC5 equipment. Transistors were 2N384, but similar later devices can be used.

some respects from the later "Vackar" or "Tesla" oscillator developed by the Czech engineer Jiri Vackar, who worked for the state-owned Tesla organization. Vackar developed his oscillator circuit in 1945. It was described with English text in *Tesla Technical Reports*, December 1949.

The Gouriet-Clapp, the Vackar (Tesla), and W8PK's Seiler oscillators (**Figure 5**) were analyzed by J.K. Clapp in *Proceedings of the IRE*, August 1954. ("Frequency Stable LC Oscillators") suggesting that the good frequency stability range of the Vackar extends over a tuning range of 2.5:1 compared with 1.8:1 for the Seiler and 1.2:1 for the Gouriet-Clapp, thus awarding the edge (at least for receiver applications) to the Vackar. In March 1955, Tesla submitted a report on the Vackar oscillator to the CCIR SG1, Document 57E, pointing out that the high frequency stability was accompanied by low harmonic content.

The first publication in an amateur journal was by David Deacon, G3BCM, *RSGB Bulletin*, March 1956, pages 3471–2 ("The Tesla Oscillator") reproducing information on the precautions required to achieve high stability. Unfortunately, somewhere along the path from Czechoslovakia to Brussels, to the U.K., there was an unfortunate mix-up with the result that G3BCM inadvertently transposed the suggested values for Cx and C1 of **Figure 5C**. These should have a ratio of about 1:6 to provide an impedance step-down from the resonant circuit to the active device. This quite serious error, which impaired stability, was reproduced in the *RSGB Amateur Radio Handbook*

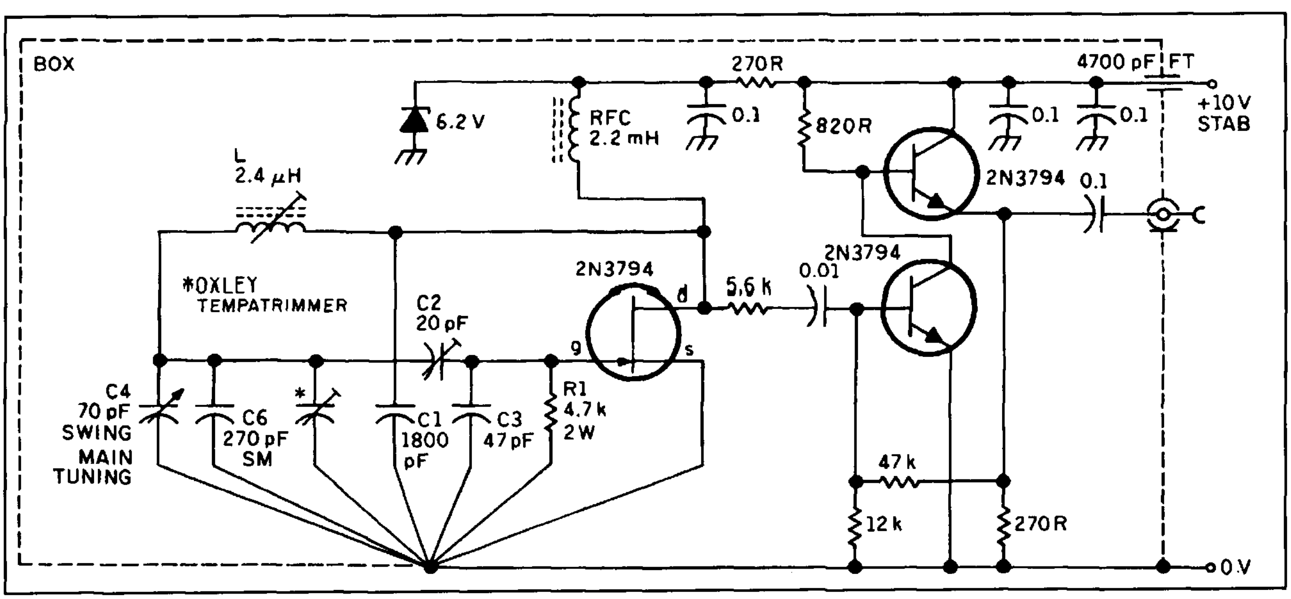


Figure 7. The high-stability FET Vackar oscillator covering 5.88 to 6.93 MHz developed by G3PDM in the late 1960s, but still a valid design for HF VFOs for receiver, transmitter, and transceiver applications.

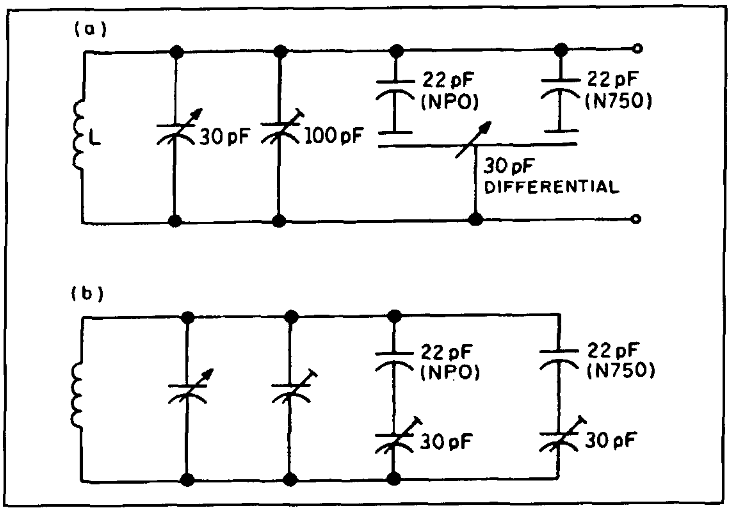


Figure 8. How adjustable temperature compensation can be achieved without the special Oxley devices. (A) With a differential capacitor; and (B) with two conventional trimmers.

(3rd edition, pages 169–70) and also in several designs published in the "Bull." It was not until 1965–66 that the error was spotted and corrected by W.H. ("Bert") Allen, G2UJ, and Lyell Herdman, G6HD, (see correspondence in the "Bull," January and March 1965).

In September 1966 in a reply to E. Chicken, G3BIK, I drew attention to the differences between the Seiler and Vackar designs. A Seiler-type oscillator using a bipolar transistor was designed by W3JHR in the early 1960s and become popular as the "synthetic rock VFO" shown in **Figure 6**.

By then bipolar transistor and FET devices were being used for Vackar type oscillators, although many designs continued (and still continue) to neglect another requirement of the Vackar to optimum stability and minimum harmonic content: the need to use a relatively high capacitance across the tuned circuit. Indeed one of the very few Vackar oscillators that meet all of the original requirements for optimum performance was the FET design by Peter Martin, G3PDM, published originally in TT (December 1969, pages 846–7), and since reproduced in a number of RSGB publications including *Amateur Radio Techniques* (7th edition, pages 166–7) and *Radio Communication Handbook* (4th edition, page 4.27 and also the new 5th edition).

The G3PDM Vackar design (**Figure 7**), although now 25 years old, remains possibly the most stable LC oscillator ever described for home construction. His design, intended as a tunable local oscillator covering 5.88 to 6.38 MHz in a double conversion hybrid valve/transistor receiver, has a switch-on drift of 500 Hz in the first 60 seconds (caused by the gate-source capacitance changing as the 2N3819 junction FET achieved thermal stability) and thereafter a drift of only about ±2Hz per 30 minutes, that is about 3 parts in 10 million! As for all LC oscillators, the mechanical stability and correct choice of good-quality components is as important no matter which basic oscillator configuration is used.

To achieve the sort of stability quoted, G3DPM listed 15 points to watch. An updated version is given below:

• Strongly recommend the genuine Vackar circuit; i.e., with C1/(C4+C6)=C3/C2 and both approximately 6:1 [C4+C6 should have sufficient capacitance to form a high-C tuned circuit. In the original Vackar prototype covering frequencies around 1.7 MHz, a 1000 pF trimmer was used across the coil. The Seiler is a low-C configuration, the Vackar high-C—G3VA].

• Use a FET rather than a valve; they are more stable, last longer, use the same circuits, and are cheaper. [But note that solid-state devices are more affected by changes in the ambient temperature than valves once these have fully warmed up—G3VA].

• Use a strong box (die-cast or better).

• Use a high-quality variable capacitor. The so-called straight-line-frequency (SLF) laws are for a tuning range of 2:1 and not useful for normal amateur use. However, Jackson Type U101 (or surplus RF-26 type) capacitors provide an almost perfect SLF law when tuning 500 kHz in this circuit.

• To reduce the heating effect of the RF currents in C2, this should be an air-spaced trimmer; this allows adjustment of feedback so that the circuit just oscillates, reducing harmonic output and drift due to interaction of harmonic energy.

• All variable capacitors should be effectively cleaned, preferably in an ultrasonic bath, before using (G3PDM stressed this really makes a difference).

• Preferably use (continuously) adjustable temperature compensation. Originally G3DPM used an Oxley "Tempatrimmer" or the lower cost Osley "Thermo Trimmer" with a more restricted range of compensation. [These appear to be no longer available. Suitable temperature coefficient capacitor(s) may have to be chosen by trial and error. Alternatively, if a differential capacitor is available, as shown in **Figure 8A** or with conventional trimmers as in **Figure 8B**—G3VA].

• C1, C3, and C6 should be silvered-mica types, isolated from surrounding solid objects (this reduces "warbling" during a "mallet test").

• The gate resistor should be a 2-watt solid

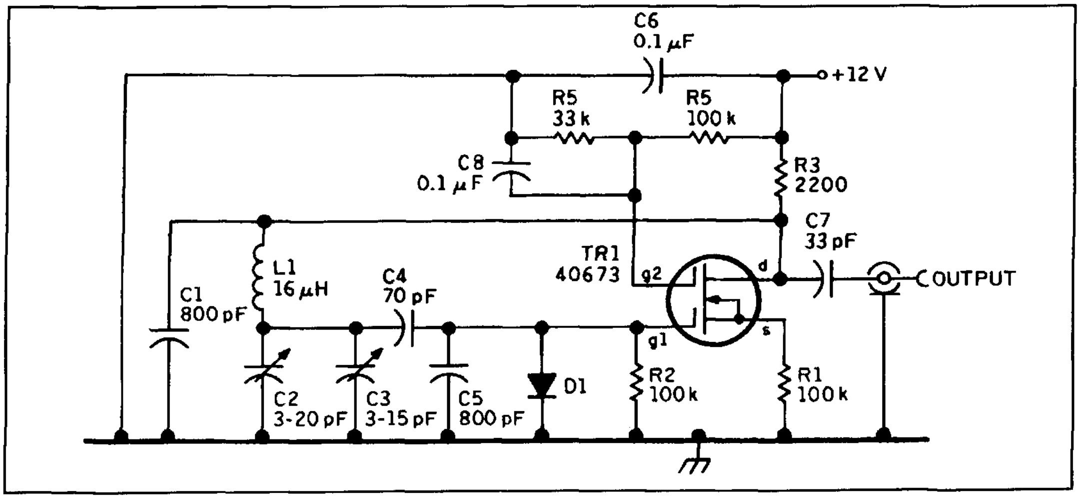


Figure 9. A recent 4.0 to 4.3 MHz oscillator design by K6BSU. It seems likely that performance could be improved by increasing the value of C2+C3 and decreasing that of L1 to make it a true Vacker oscillator. Component information recommended by K6BSU. C1, C5: 800 pF polystyrene (see text). C2: air trimmer for setting calibration. C3: air variable for main tuning. C4: 70 pF made up of NPO ceramic and silver mica to provide required temperature compensation. C7: 33 pF NPO ceramic. C6, C8: 0.1-µF 25-volt monolithic capacitor. TR1: 40673 or SK3050 dual-gate MOSFET. D1: 1N4153 or 1N914 silicon signal diode. L1: 16 µF, 34 turns no. 26 enamel on 3/4-inch diameter ceramic form (no slug core), winding length 0.6 inch. All resistors 0.25 watt, 5 percent.

carbon type for minimum heating and low inductance.

- Use of a buffer/isolating amplifier is essential. With a feedback pair, the gain is readily adjusted while negative feedback maintains low harmonic content.
- Circuits using a diode from gate to earth for rectification outside the FET appear to increase drift.
- Power supplies must be very well stabilized, and disc ceramic bypass capacitors should be liberally used to prevent unwanted feedback along the supply rails. [Note: modem IC regulators would be an improvement on the use of Zener diodes—G3VA].
- Oscillator components around the tuned circuit (L, C1, 2, 3, 4, 6, R1, and FET source) should have a single common earthing point. (This usually means using one of the fixing screws of C4.)
- Ceramic coil formers are preferred. An iron dust core facilitates VFO calibration, but ferrite cores must be avoided.
- Keep leads short, and use stiff wire (16 or 18 SWG) for interconnections in the oscillator tank circuit.

G3DPM added that performance achieved included: resetability—after switching off for 12 hours, returns to within 10 Hz of previous frequency; voltage stability (without Zener diode or other voltage regulation), 10 percent change in supply results in shift of 8 Hz; G3DPM standard mallet test results in average shift of 6 Hz; scale linearity—maximum error over 500 kHz band, 12 kHz "without any codging."

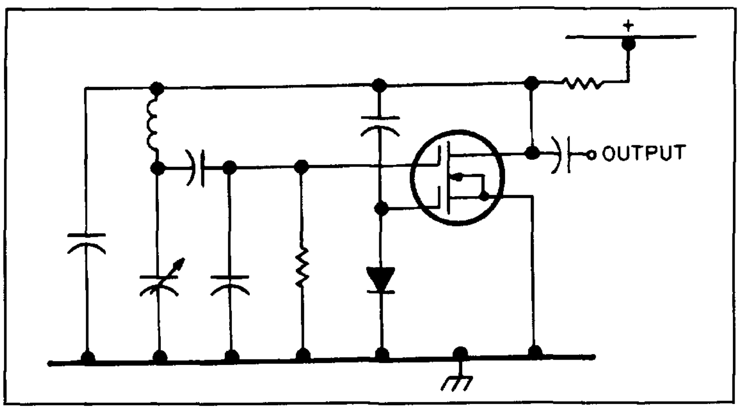


Figure 10. Possible simple AGC system for use with a Vackar oscillator which should be an improvement over a diode connected directly between gate and source. D1: any good quality RF silicon diode.

American amateurs have been relatively slow to adopt the European Vackar circuit. However, a number of articles have appeared in the American magazines since the 1970s, although few seem to have appreciated that the Vackar, unlike the Seiler, really requires a high-C tank circuit. This factor is still missing from an otherwise useful article "The Vackar High-Stability L-C Oscillator," by Floyd E. Carter, K6BSU (*CQ*, June 1994). He uses a dual-gate MOSFET (40673 or SK3050) and gives component values for tuning 4.0 to 4.3 MHz: **Figure 9**. Although he gives impressive performance figures, it seems likely that these could be improved simply by reducing the inductance of L1 and adding considerably more capacitance to the trimmer C2 and tuning capacitor

C3 to make it a true Vackar circuit, and possibly by removing D1.

Among the constructional features he recommends is the use of a high-quality double-bearing air variable capacitor for C3, with the trimmer C2 also a ceramic-base air variable. C1, 4, and 5 are made up from parallel capacitors adding up to the required value since the use of multiple capacitors reduces the RF heating of the dielectric in the individual capacitors. C1 and 5 are each implemented from two polystyrene capacitors, while C4 is a combination of an NPO ceramic and a silver mica capacitor, providing temperature compensation for the positive temperature coefficient of L1. This seems an odd way of providing temperature compensation since the purpose of C4 is to form part of a capacitive potential divider with the effect of tapping down G1.

In his prototype, he used a 62 pF NPO ceramic with 8 pF silver mica, but I feel that G3PDM's use of a trimmer adjusted to just beyond oscillation together with temperature compensation directly across the tuned circuit should prove the better approach. For supply regulation, K6BSU uses an adjustable IC regulator type LM317 to provide 12 volts, although he found that the circuit oscillated well with the 7 volt supply. He recommended following the oscillator with a high input impedance Class A buffer amplifier. The use of a diode (D1) between gate and source of the MOSFET to provide a form of AGC is common practice, but has been criticized by some designers. An effective AGC system (as used for example in the original Gouriet BBC VFO) is undoubtedly beneficial; a simpler arrangement has been proposed for use with a dual-gate MOSFET Vackar oscillator (**Figure 10**) although I have not heard of this being used in practice. ■

PRODUCT INFORMATION

Analog Devices Announces New Family of DSPs

Analog Devices, Inc. has announced the availability of the ADSP-21csp01. The ADSP-21sp01 is the first in a new family of 16-bit fixed-point DSPs from Analog Devices. The new 16-bit fixed-point core is able to power the 21csp family of concurrent signal processors. The ADSP-21csp01 addresses the challenges of many applications by integrating a highly-parallel, 50-MIPS DSP core capable of executing 550 Million Operations Per Second (MOPS); a parallel DMA port, a parallel memory port, and two multi-channel serial ports; and low latency interrupt servicing with a 20 ns task switching speed.

Based on a modified Harvard architecture, the core maintains a three-bus bandwidth on-chip through a 64-word, selective instruction cache. A three-stage instruction pipeline increases the overall instruction rate while keeping latency to a minimum. There is no arithmetic pipeline allowing all computational instructions to execute in a single cycle. The 24-bit instruction word supports multiple programs and data sets with a 16 MWord address range. Additional features include two data address generators which support eight simultaneous circular buffers. Independent arithmetic units include: 16-bit ALU; 16 x 16-bit MAC with dual 40-bit accumulators and the option to saturate every cycle; 32-bit, bi-directional barrel shifter with block floating-point support. The result is an effective processing rate of 550 MOPS.

To maximize the efficiency of interrupt processing, the ADSP-21csp01 includes a complete background register set which contributes to the device's ability to task switch in one clock cycle (20 ns). Combined with low interrupt latency, the background registers are switched to support a new task in 100 ns following an interrupt.

To minimize interruption of the core processing functions and maximize the flow of data into and out of the processor, the ADSP-21csp01 integrates serial ports running at 100 Mb/s, a 150 Mbyte/s parallel memory interface, and parallel 50 Mbyte/s DMA ports on-chip. Two bi-directional serial ports, each supporting two to 32 TDMA channels, operate at 25 Mb/s data rates. Because the serial ports are independent and bi-directional, a total of four data paths are available simultaneously, providing a total serial port transfer capacity of 100 Mb/s. The serial ports also have Direct Memory Access (DMA) capability, allowing converters and other peripherals to move data into and out of DSP memory without interrupting the DSP. The internal DMA port is a 16-bit parallel port interfacing the ADSP-21csp01 to other processors and to system buses. A DMA controller supports five DMA channels between the serial ports or IDMA port and the memory of the ADSP-21csp01.

Samples are available now in limited quantities. General sampling begins in January with volume production planned for the summer of 1996. The 10,000 unit price is $33.00.

For further information, contact Analog Devices, Inc., Three Technology Way, Norwood, MA 02062; phone 617-461-3881; fax 617-821-4273.

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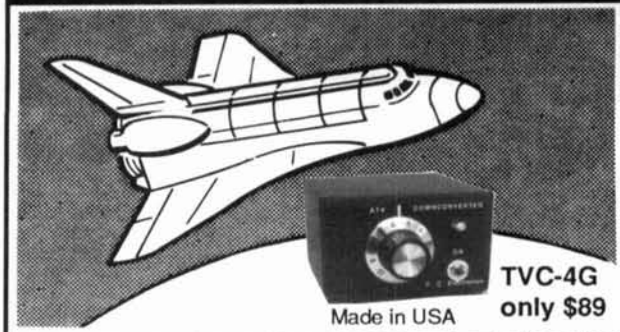

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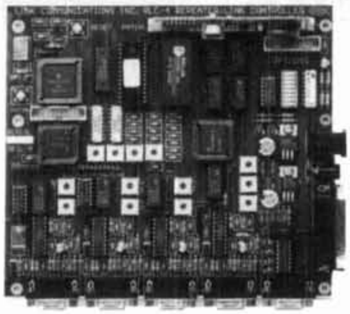

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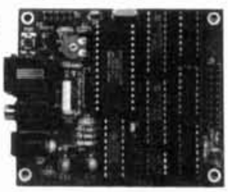

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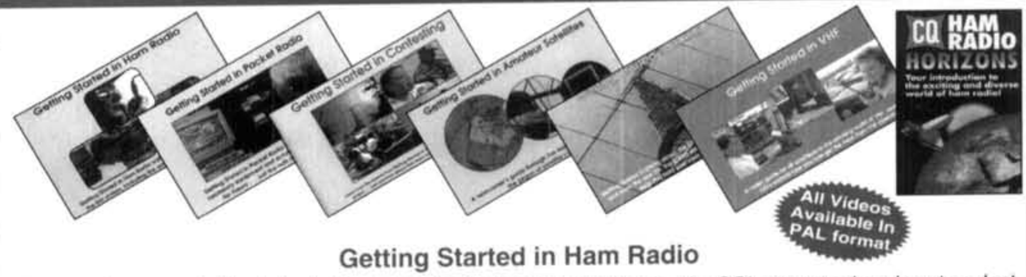

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