QEX

ARRL Experimenter's Exchange

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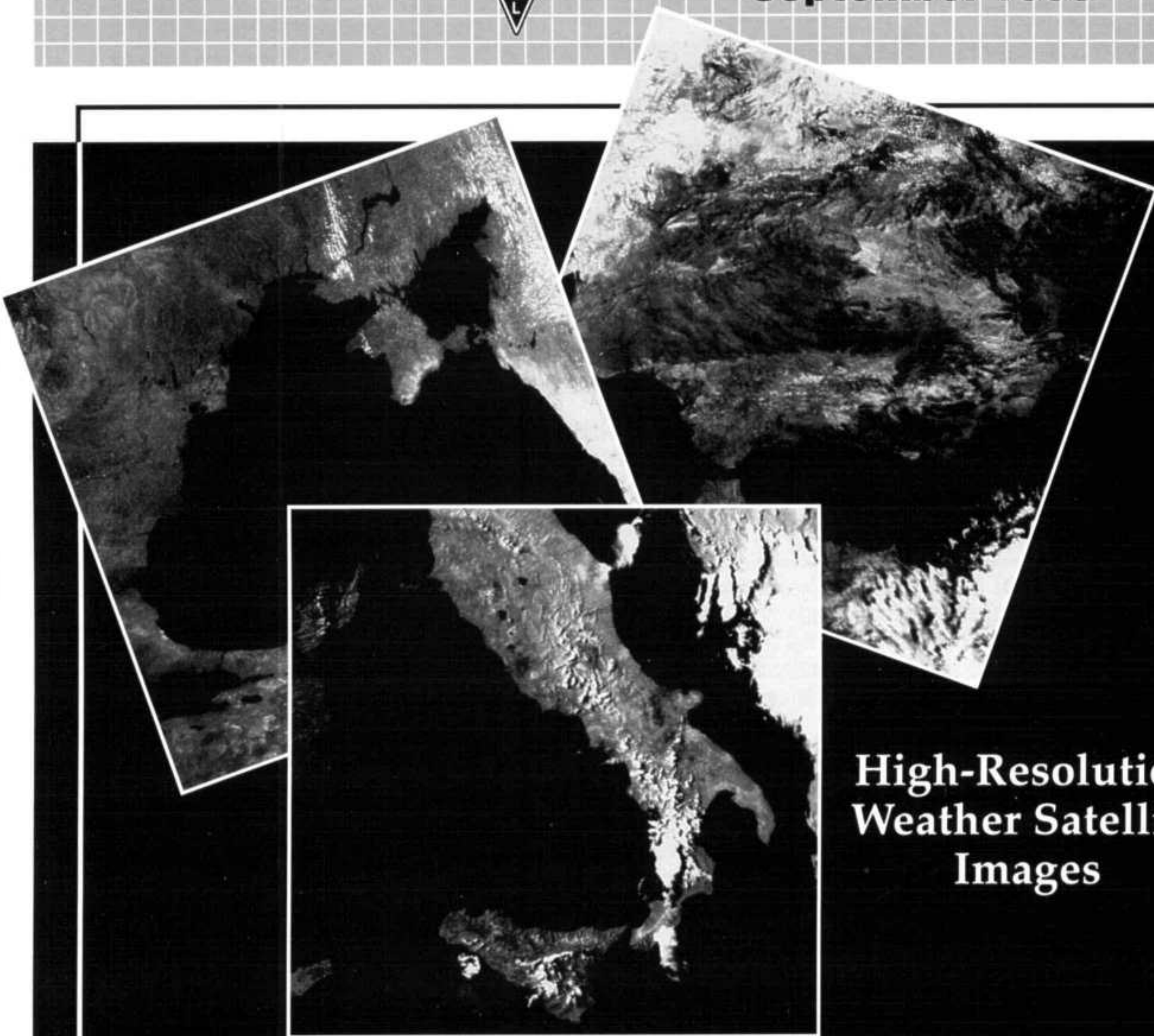
September 1995



High-Resolution Weather Satellite Images

QEX: The ARRL
Experimenter's Exchange
American Radio Relay League
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QEX

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David Sumner, K1ZZ
Publisher

Jon Bloom, KE3Z
Editor

Lori Weinberg
Assistant Editor

Harold Price, NK6K
Zack Lau, KH6CP
Contributing Editors

Production Department

Mark J. Wilson, AA2Z
Publications Manager

Michelle Bloom, WB1ENT
Production Supervisor

Sue Fagan
Graphic Design Supervisor

Joseph Costa
Technical Illustrator

Joe Shea
Production Assistant

Advertising Information Contact:

Brad Thomas, KC1EX, *Advertising Manager*
American Radio Relay League
860-667-2494 direct
860-594-0200 ARRL
860-594-0259 fax

Circulation Department

Debra Jahnke, *Manager*
Kathy Fay, N1GZO, *Deputy Manager*
Cathy Stepina, *QEX Circulation*

Offices

225 Main St, Newington, CT 06111-1494 USA
Telephone: 860-594-0200
Telex: 650215-5052 MCI
FAX: 860-594-0259 (24 hour direct line)
Electronic Mail: MCIMAILID: 215-5052
Internet:qex@arrl.org

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About the Cover

Like weather satellite images? APT is nice; HRPT is better, and I4GU, I4MY and IW4CSG show how to do it.

ISSUE NO. 163

Features

Columns

September 1995 QEX Advertising Index

Purpose of *QEX*:

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

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

Both theoretical and practical technical articles are welcomed. Manuscripts should be typed and doubled spaced. Please use the standard ARRL abbreviations found in recent editions of *The ARRL Handbook*. Photos should be glossy, black and white positive prints of good definition and contrast, and should be the same size or larger than the size that is to appear in *QEX*.

Empirically Speaking

DSP versus GP

Those who read Usenet news and who've stumbled across the comp.dsp newsgroup will recall the occasional outburst of discussion of the question: With the current fast crop of general-purpose CPUs, is there still a need for dedicated DSP chips? (The discussion is similar to the deathless "code versus no-code" debate on rec.radio.amateur.policy.misc, except the comp.dsp discussions are a bit less, um, contentious.) The argument never quite gets settled, of course, but the two major camps resolve into 1) I can do all the DSP I need to do on a Pentium/PowerPC/Alpha/what-have-you, and 2) then you aren't doing leading-edge DSP.

Whatever your view of that issue, it raises questions relevant to the amateur community. What about the DSP *we* do? Could it as easily be performed using general-purpose processor systems? Consider that most of the DSP presently done in Amateur Radio is performed on narrowband signals—often at baseband. Can a modern general-purpose computer serve to do that processing? The answer is almost certainly "yes," in all but some extreme cases.

What sparks this line of thought is a talk given by Phil Karn, KA9Q, at the 1995 TAPR annual meeting. Phil reported some results of his efforts to see just how well a Pentium can calculate the fast Fourier transform (FFT) of a sequence. Phil's results are quite encouraging and lead to the conclusion that, yes, you *can* do a lot of DSP on a modern processor. And, mind you, using floating point math, which is not a feature of the lowest-cost DSP evaluation kits that we've reported on here in *QEX*. (See last month's "Empirically Speaking.")

One of the things that makes PC-based DSP particularly attractive is the availability of low-cost "sound cards" for sample acquisition and reconstruction. Heck, most of the PCs we see advertised at the retailers these days come with a sound card. So there are a lot of potential DSP platforms out there, awaiting appropriate software. (On the down side, such systems don't include a ready-to-use interface to a radio's control signals. Plus, directly connecting the computer and the ham rig may lead to problems with conducted RFI—although such problems are treatable.)

Another advantage of PCs is the readily availability of development tools. Just code your application in the high-level language of your choice, possibly with some optimized assembly language parts for the critical code, using your existing compilers. And DSP design tools continue to become available for PCs, too, although the best of them are still out of price range of the amateur purse.

There probably is still some role for dedicated DSP chips in Amateur Radio. After all, if you optimize a processor design to do a particular job, likely it can do the job better than a general-purpose processor. But most of the applications of DSP to amateur technology probably don't require such optimized processors—today, at least. Still, there's something to be said for a stand-alone DSP board. If nothing else, it allows the main computer more time to do the non-DSP part of whatever application you have in mind. But in focusing on DSP evaluation boards and other low-cost DSP platforms, we may be overlooking the most ubiquitous DSP platform: the PCs on our desktops.

This Month in *QEX*

If you like APT weather satellite images, you'll *love* HRPT. But how do you decode and display such images? Guido Emiliani, I4GU; Marciano Righini, I4MY; and Giampaolo Rossini, IW4CSG, present their approach to "HRPT The Easy Way."

Dynamic resistance, the *V/I* characteristic of an active device, causes confusion to a lot of amateurs—and engineers, for that matter. It shouldn't. As William E. Sabin, WØIYH, shows us, "Dynamic Resistance in RF Design" is an understandable phenomenon.

Loop antennas for radio direction finding have been around for many years. Ever wonder how they work? George Brown, G1VCY, explains the basics in "The Principles of HF Radio Direction Finding Loops."

Finally, in his "RF" column, Zack Lau, KH6CP/1, presents a design for a 6-meter transverter—with a 2-meter IF. Zack's unique design is composed of several building blocks from which you can extract design ideas, as well as being a fully integrated transverter design.—*KE3Z, email: jbloom@arrl.org*

HRPT The Easy Way

Reprinted from Radiokit Elettronica, *April and May 1995.*

By Guido Emiliani, I4GU
Marciano Righini, I4MY
Giampaolo Rossini, IW4CSG

Say you've been receiving the NOAA weather satellite APT signals on 137.5 and 137.62 MHz for years, and you're proud of the fine pictures that you can display on your PC monitor. Analog APT has been a very successful system of weather image dissemination—it's simple to receive and convert into pictures.

But APT transmits only one third of the lines generated by the AVHRR, the instrument that scans the Earth surface and produces the images. All the NOAA instrument data are included in the HRPT digital transmission, but you thought that displaying all the lines would go beyond your technical or economical capabilities. This was probably true a few years ago, but now—well, let's see if it's still true.

Via Colombo Lolli 8
48100 Ravenna
Italy

We'll examine the difficulties of homebrewing an HRPT receiving station, but we'll also discover how easy it is if you have a good signal at your disposal. In fact, the greatest difficulty is acquiring a strong and steady signal from the satellite. If you can do this, the rest of the system is within the capacity of a homebrewer, and the software necessary to complete the system is available free to everybody.

Radio Frequency Section

These are the components of our RF section:

a) A 1.4-meter parabolic dish with an f/d ratio of 0.38.

b) A cylindrical horn, 17 cm long and 12 cm in diameter, at the focus of the dish. Inside the horn, two quarter-wave monopoles (probes), each 4 cm long, are placed at 4.5 cm from the horn's back end, 90° apart. To obtain circular polarization, one monopole must be fed 90° out of phase with respect to the other. The delay is obtained with two lengths of RG-59: one 11.65 cm long (λ), and the other 14.55-cm long ($\lambda+\lambda/4$). The proper polarization is left-handed (the dish inverts the right-hand polarization from the satellite); thus, circling counterclockwise, the first probe you meet should be connected to the longer cable. The ends of the two cables are connected together and go to

c) A low-noise amplifier (LNA) for 1700 MHz, to be placed at the rear of the horn.

The system also includes:

d) Yaesu G-5400B dish azimuth and elevation rotators.

e) The Kansas City Tracker (which is optional, because manual tracking is easy with this system).

f) A downconverter that converts the HRPT signal to a "transportable" frequency. (If your receiver can tune the S band and is close to the dish, you may not need a converter.)

g) An ICOM IC-R7000 receiver, with which we receive the converter's output frequencies or (if we don't use a converter) the NOAA satellite's frequencies (1698 and 1707 MHz). The R7000's rear panel includes a jack that outputs a 10.7-MHz 2nd-IF signal. We use this signal to drive our HRPT decoder. (Be careful: this jack also sources 9 V dc, so you may need to use a blocking capacitor.)

There are many more possible arrangements, but we agree on one point: If you cannot install the "external" part of the system (items a, b, c, d), this project is not for you. Again, the secret for success is *a very good RF signal from the satellite.*

Fig 1 block-diagrams the "internal" part of the system—the hardware described in this article.

10.7-MHz Amplifier and Band-Pass Filter

See Fig 2. Because the signal level from the 10.7-MHz output of the R7000 is very weak, we need amplification to drive the following stage. A white dot marks the input of the MAR-6 MMIC, and the output is the pin on the opposite side. The other two pins must be grounded.

The filter, consisting of two FM2/3 IF transformers, should have a bandwidth of about 1.5 MHz at –3 dB. This is a critical section of the system; if you do not have the proper test equipment (as is our case), you will have to adjust it on the satellite signal for a noise-free picture.

Phase Demodulator and Split-Phase-Low to Nonreturn-to-Zero Decoder

See Fig 3. This section performs the following functions:

a) It demodulates the 10.7-MHz RF signal (in IC1, a Plessey SL1451) and supplies a split-phase-low (Manchester) coded signal.

Table 1—HRPT Parameters

Transmit frequencies	1698 MHz for the NOAAs transmitting APT on 137.5 MHz		
	1707 MHz for the NOAAs transmitting APT on 137.62 MHz		
Polarization	Right hand circular		
Carrier modulation	Digital split phase, phase modulated		
Lines per frame	1		
Line rate	6 per second		
Number of words	11090 per line		
Word rate	66540 per second		
Number of bits	10 per word; bit 1 = MSB (transmitted first)		
Bit rate	665400 per second		
Words per image	2048 per line		
Spectral channels	5:		
	Channel 1	0.58 - 0.68	µm
	Channel 2	0.725 - 1.1	µm
	Channel 3	3.55 - 3.93	µm
	Channel 4	10.3 - 11.3	µm
	Channel 5	11.5 - 12.5	µm

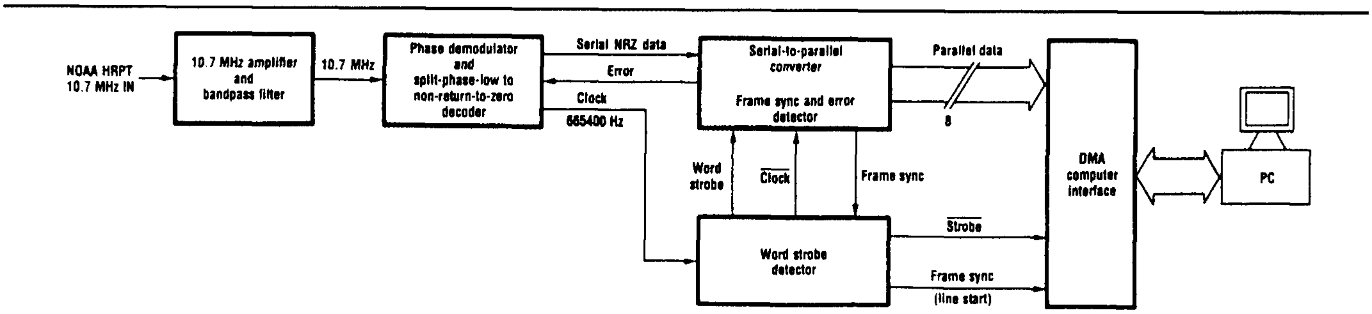


Fig 1—Block diagram of the HRPT system.

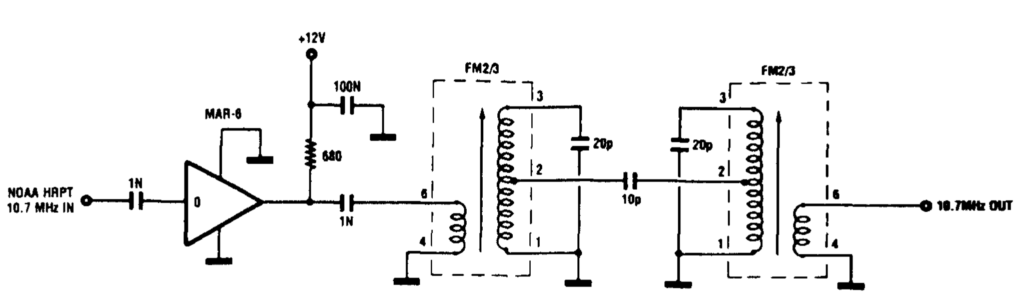


Fig 2—The 10.7-MHz amplifier and band-pass filter.

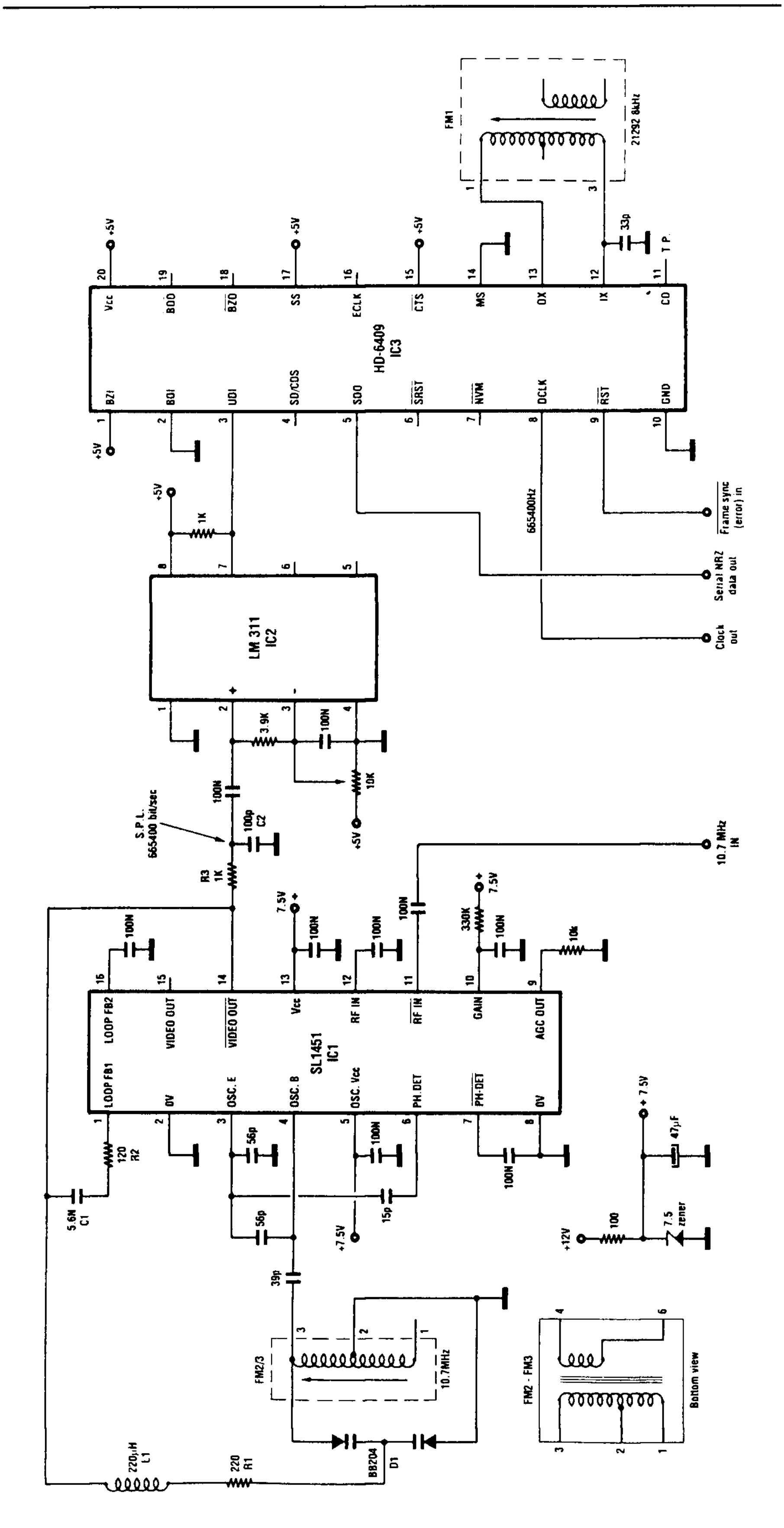


Fig 3—The phase demodulator and split-phase-low to nonreturn-to-zero decoder.

b) It decodes the SPL data into nonreturn-to-zero (in IC3, a Harris HD-6409).

IC1 is a Plessey PLL FM detector intended for satellite TV reception.[1] Inside the chip is a transistor (pin 3 emitter, pin 4 base, pin 5 collector) that must be connected to an external coil (in our case, a small IF transformer) to act as an oscillator at 10.7 MHz. The HRPT signal from the filter of Fig 2 is fed into pin 11. The video output (pin 14) drives a dual tuning diode (D1, a BB204) via isolating network L1-R1. It also drives the SL1451's loop feedback 1 input (pin 1) via the filter consisting of C1 and R2. The gain of the SL1451's internal input RF amplifier is programmable by varying the voltage on pin 10; the gain is maximum with pin 10 connected to V_{cc} via a 330-kΩ resistor.

The 665.4 kbit/s split-phase-low signal passes through a post-detection low-pass filter (R3-C2), which is designed to cut off above twice the bit rate (1330.8 kHz), and on to IC2, which outputs it at a 5 V level.

The decoding process is based on an HD-6409 Manchester encoder-decoder (Harris Corporation), a chip that converts Manchester code into NRZ code and provides clock recovery. The SPL data is fed into the decoder's Unipolar Data Input (UDI) pin. The HD-4609's SDO (serial data out) provides decoded serial NRZ data synchronous with the decoder clock (DCLK). The decoder requires an oscillator with a frequency 16× or 32× the bit rate. The HD-6409's speed selector (SS) pin (17) sets this, with SS low producing a 16× clock and SS high a 32× clock. We use the 32× mode, so the frequency of the free-running oscillator should therefore be 21292.8 kHz for clock synchronization with the incoming data (665.4 kHz × 32 = 21292.8 kHz). A frequency counter connected to the clock output (DCLK) via a 1-kΩ resistor is very useful in showing when the internal clock is locked to the incoming data rate (665.4 kHz) (Doppler shift causes the clock to shift over 30 Hz between AOS and LOS.)

IC3's reset ($\overline{RST}$) pin provides control of the decoder outputs. When $\overline{RST}$ is low, SDO and DCLK are forced low. $\overline{RST}$ is connected to the error detector in order to resolve the ambiguity

[1]GEC Plessey Semiconductors, "Designing with the SL-1451 Phase Locked Loop," *Consumer IC Handbook*, Sep 1991, pp 8-23 to 8-25.

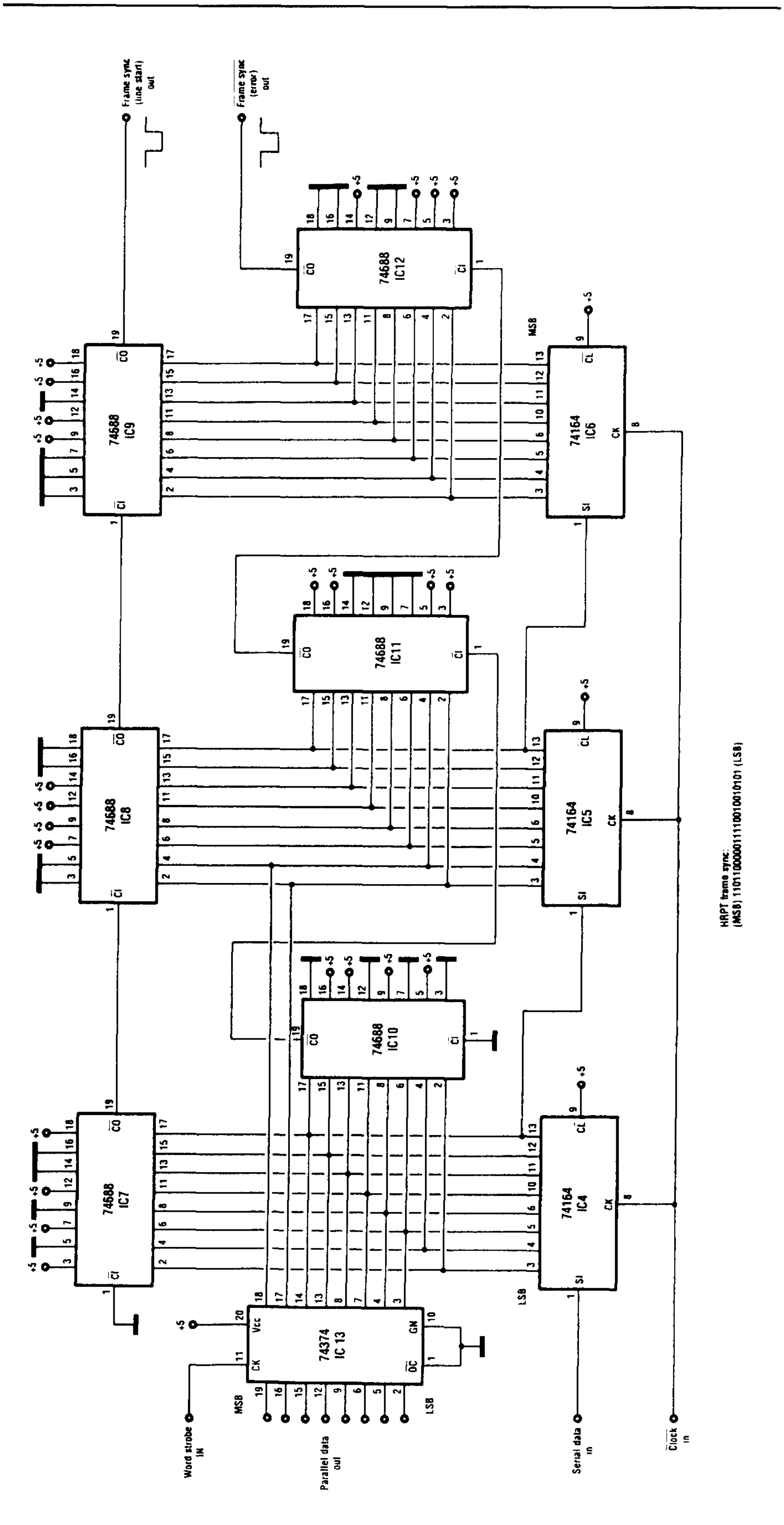


Fig 4—The serial-to-parallel converter, frame sync and error detector.

problem. If $\overline{\text{frame sync}}$ is detected by IC10-12, the clock error generated is fed to $\overline{\text{RST}}$ so that the phase of the clock is changed by 180°.

IC3's $\overline{\text{RST}}$ input is normally kept high by IC12. If you test this circuit subsection without connecting it to the following section (Fig 4), temporarily tie $\overline{\text{RST}}$ high.

Serial-to-Parallel Converter, Frame Sync and Error Detector

See Fig 4. This section:

a) Performs the serial-to-parallel conversion.

b) Generates the $\overline{\text{frame sync}}$ (line start) at the last bit of the 60-bit sequence transmitted at the beginning of each line.

c) Generates the $\overline{\text{frame sync}}$ (error) at the last bit of the 60-bit sequence if it detects a phase error (that is, if the value of the bits is inverted).

d) Feeds the parallel data into the DMA computer interface (Fig 6) subdivided into bytes corresponding to the data words.

The serial-to-parallel conversion is performed by passing the serial data through a 24-bit shift register (IC4-6). Our system makes use of the eight most significant bits of the 10-bit word. The eight data lines are derived from the last six outputs of the first register (IC4, a 74164) and the first two of the second shift register (IC5, another 74164). The bus is fed into a D-type register (IC13, a 74374) that is clocked by the word strobe (a high at the end of each word). The eight parallel lines are connected to the DMA computer interface.

Only the last 24 bits of the 60-bit frame sync sequence transmitted at the beginning of each frame (line) are detected. When the 24th bit enters the first register (IC4), a low (frame sync flag) is generated at pin 19 of IC9 (a 74688); this is the line start flag.

This section also detects the inverted sequence in order to resolve the ambiguity problem. The effect of a phase error is that the data appears inverted (0 instead of 1 and vice versa). This situation is detected by checking the serial data for frame sync as well as $\overline{\text{frame sync}}$ and correcting the clock if $\overline{\text{frame sync}}$ is found. This signal, generated at pin 19 of IC12 (another 74688), is fed into the $\overline{\text{RST}}$ input of the HD-6409 decoder (IC3).

Word Strobe Detector

See Fig 5. This section performs these functions:

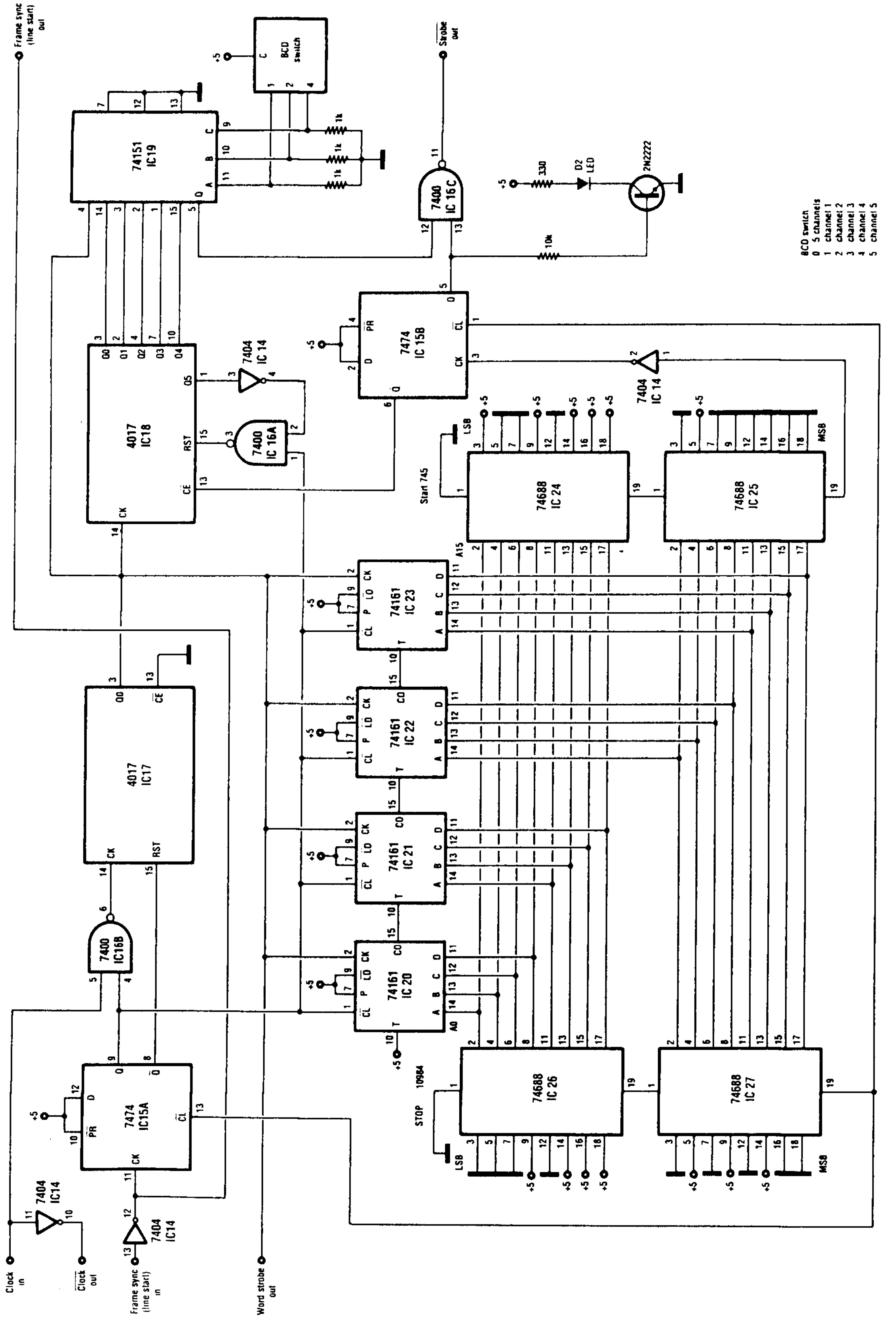


Fig 5—The word strobe detector.

a) It divides the flow of clock pulses into blocks of 10 bits that correspond to the bytes or words of the data, and provides a pulse (word strobe) at the last (tenth) bit of each word.

b) It divides the word strobe by 5 (in IC18, a 4017) so that you can choose one of the five channels transmitted by the satellite. IC18's five output lines carry the data of five multiplexed channels, each of which represents a view of the same area of the Earth as viewed by five bands of the light spectrum. A front-panel-mounted BCD switch selects the channel outputted by IC19, a 74151. (It's often practical to choose one channel and ignore the other four. This allows you to display a real-time image and avoids filling up your hard disk with data you won't use.)

c) It supplies the computer with the data of all the five channels if the divide by 5 (IC18) is bypassed. This option is implemented by the BCD switch. Storing five channels implies the creation of a *50-Mbyte* file for a 15-minute pass, which is not always feasible. Moreover, after the pass it is necessary to explode the five channels, an operation which requires another 50 Mbytes of hard-disk space.

d) It blocks the data not belonging to the images that are present on the line, ie, the first 750 bytes and the last 100 bytes of the frame.

e) It signals with D2, a front-panel-mounted LED, that the system recognizes the frame sync (line start); that is, the repetitive sequence situated at the beginning of each frame (line).

When the frame sync pulse (line start) is fed into the IC15A's CK input, Q goes high and therefore the clock pulses pass through IC16B and reach IC17, which divides the frequency by 10 thus generating the word strobe (66.54 kHz)—a pulse at the end of each 10-bit word. The word strobe:

1) Drives IC13 of Fig 4 to signal that the 8 bits present at its output are an image word.

2) Drives the DMA computer interface (via IC19 and IC16C) if you want

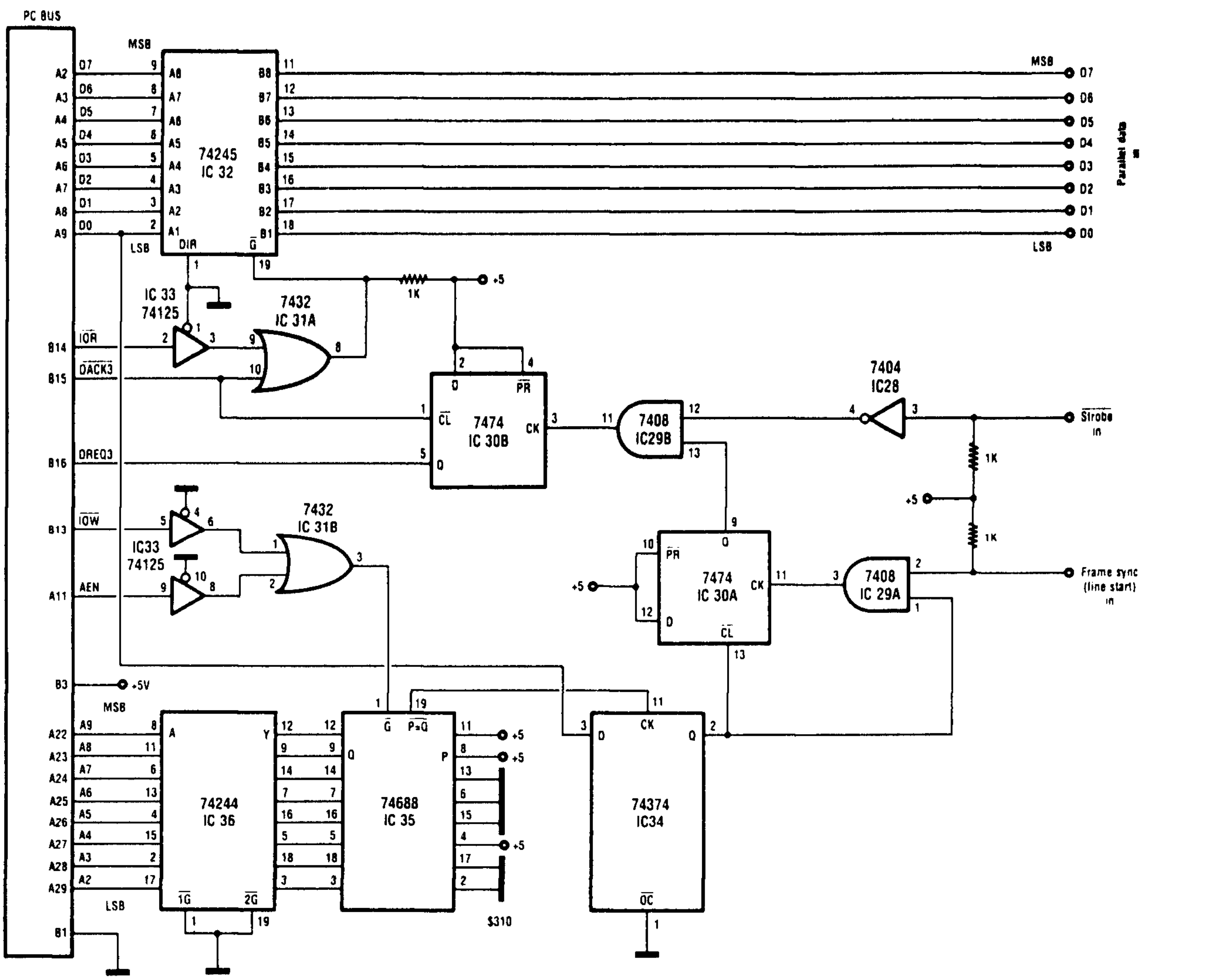


Fig 6—The DMA interface.

to acquire all 5 channels. If you want to receive only one channel, IC18 divides the word strobe by 5, generating the sample strobe—a pulse that occurs every 50 bits, at the end of all the words of one of the five multiplexed channels. The desired channel can be selected by means of the BCD switch.

3) Drives counters IC20-23, the outputs of which are connected to the inputs of both the start comparators (IC24-25), and the stop comparators (IC26-27). The start pulse (a low at pin 19 of IC25) is generated at word 745 of the line (the first word of earth data is 751 minus 6 frame sync words); the stop pulse (a low at pin 19 of IC27) is generated at word 10984 (the last word of earth data is 10990 minus 6 frame sync words).

When the start signal is applied to IC15B's clock input, Q goes high, IC16C allows the strobe to pass, and LED D2 blinks 6 times per second, signaling that the system works. At the same time, $\overline{Q}$ goes low, thus enabling divide-by-5 IC18. IC18 is reset after the fifth count. The stop signal resets both flip-flops IC15A and B, and, via outputs Q and $\overline{Q}$ of IC15A, the counters and the dividers.

Frame sync and strobe are sent to the direct memory access (DMA) computer interface by means of separate shielded cables.

DMA Computer Interface

See Fig 6. This board, which installs in an expansion slot on the PC system board, performs these functions:

a) It activates the data input when requested by the receive software; then it resets the DMA request.

b) It recognizes the line start pulse generated by IC9.

c) It makes a direct memory access request for the PC whenever a data item is acquired.

The three buffers IC32, IC33 and IC36 isolate the PC bus from the rest of the board. IC32 feeds the data into the PC bus. IC35 acts as memory address decode logic (310_{16}). IC34 is a software-driven switch.

When the software keeps output Q of IC34 (pin 2) high, the line-start signal is applied to IC30A which, in turn, activates IC29B. Under this condition, the strobe signal reaches IC30B, whose output Q (pin 5) generates a Dma REQuest (DREQ). Then the computer responds by activating the Dma ACKnowledge (DACK) line, which resets IC30B and enables IC32 to feed data (D0-D7) into the data bus.

PC

Our computer is a 486DX, 33-MHz machine with 4 Mbyte of RAM. The hard disk is a 340-Mbytes Western Digital with a VESA controller; it has 15-ms access time. The video adapter card is a Cirrus 5428 with 1 Mbyte of memory. Slower machines may not be able to perform all the options of the program (eg, storing five channels, displaying and storing at the same time), but they should be able to do all the basic functions.

Software

We used *Turbo Pascal* as our programming tool. The program, produced by Giampaolo, IW4CSG, covers HRPT as well as APT/WEFAX and HR (digital) Meteosat, but this article does not deal with the hardware interfaces necessary for the latter. The pictures can be dispayed in various formats, stored on disk and processed (zoomed, enhanced and cleaned if single pixels

Table 2—HRPT Line Format

Function	*Number of Words*	*Word Position*	*Word Code & Meaning*
			MSB LSB
Frame sync	6	1	1010000100
		2	0101101111
		3	1101011100
		4	0110011101
		5	1000001111
		6	0010010101
Telemetry, TIP, etc	744	7 to 750	
Earth data	10240	751	Ch 1 - Sample 1
		752	Ch 2 - Sample 1
		753	Ch 3 - Sample 1
		754	Ch 4 - Sample 1
		755	Ch 5 - Sample 1
		756	Ch 1 - Sample 2
		757	Ch 2 - Sample 2
		758	Ch 3 - Sample 2
		759	Ch 4 - Sample 2
		760	Ch 5 - Sample 2
		761	Ch 1 - Sample 3
		–	–
		–	–
		–	–
		10985	Ch 5 - Sample 2047
		10986	Ch 1 - Sample 2048
		10987	Ch 2 - Sample 2048
		10988	Ch 3 - Sample 2048
		10989	Ch 4 - Sample 2048
		10990	Ch 5 - Sample 2048
Auxiliary sync	100	10991 to 11090	

NOAA Satellite Status

NOAA 9	Downlink: APT 137.62 MHz; HRPT 1707 MHz Passes: morning (descending), early evening (ascending)
NOAA 10	Downlink: APT 137.5 MHz; HRPT 1698 MHz Passes: early morning (descending), afternoon (ascending)
NOAA 12	Downlink: APT 137.5 MHz; HRPT 1698 MHz Passes: morning (descending), afternoon (ascending)
NOAA 14	Downlink: APT 137.62 MHz; HRPT 1707 MHz Passes: early afternoon (ascending), night (descending)

Photo 1—NOAA-10, 27 January 1991, 17:50-18:05 UTC, Ch 4 IR. The sky over central Europe is overcast. Only the snow-capped Alps emerge from the fog. A huge vortex spins over the Mediterranean.

Photo 2—NOAA-11, 29 August 1992, 12:53-13:08 UTC, Ch 2 VIS. Two great rivers, the Dnieper and the Danube, flow into the Black Sea. The picture also shows the Marmara Sea between the Straits of Bosphorus and Dardanelles.

Photo 3—NOAA-12, 1 June 1994, 07:37-07:52 UTC, Ch 2 VIS. The picture shows southeastern England, the Strait of Dover, Holland, Belgium and northern France. London is barely visible because of a veil of mist; Paris is a large dark spot. Many more smaller spots mark other cities and towns.

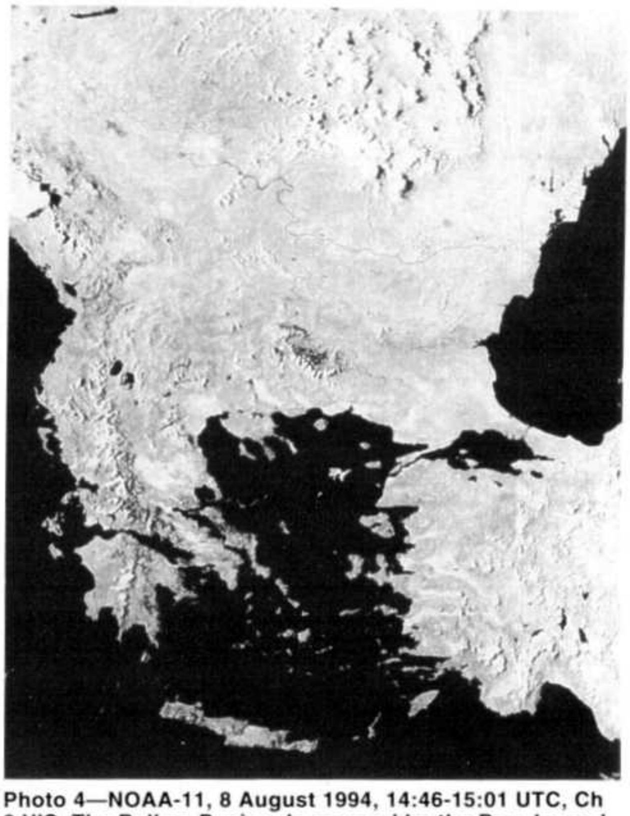


Photo 4—NOAA-11, 8 August 1994, 14:46-15:01 UTC, Ch 2 VIS. The Balkan Peninsula crossed by the Danube and the Aegean Sea between Greece and Turkey.

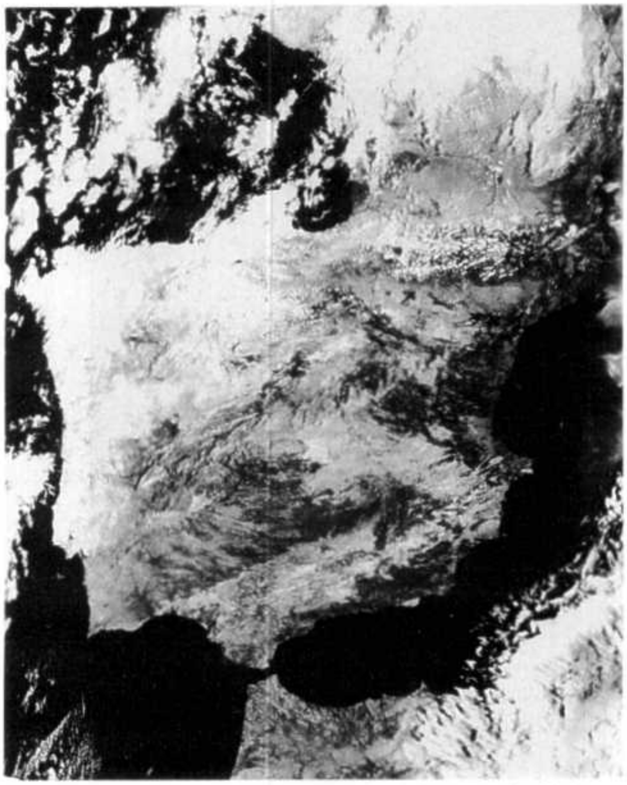


Photo 5—NOAA-9, 3 June 1994, 09:22-09:37 UTC, Ch 2 VIS. Spain and Portugal.

Photo 6—NOAA-11, 4 August 1994, 15:36-15:51 UTC, Ch 2 VIS. A summer picture of Italy.

Glossary

AOS—Acquisition Of Signal.

APT—Automatic Picture Transmission. Analog system for the transmission of environmental images used by the NOAA weather satellites. The APT signal derived from the AVHRR (see below) consists of a multiplexed output of two selected channels of this instrument. The APT signal modulates a 2400-Hz subcarrier, which is then frequency-modulated on a VHF RF carrier and transmitted to ground stations.

AVHRR—Advanced Very High Resolution Radiometer. The instrument, which scans the Earth's surface by means of a rotating mirror, is sensitive in five spectral regions. The scan of the AVHRR is converted to a digital format, which is then phase-modulated on an S-band carrier and transmitted to ground stations.

Frame—Major frame corresponds to three successive lines during which the same TIP (TIROS Information Processor) data is transmitted. Minor frame means single line. In this article, the word *frame* is used with the meaning of scan line. The HRPT frame rate is 6 per second. In a frame there are 11090 10-bit words.

Frame sync—A sequence of 60 bits (six words) at the beginning of each frame. In this article, *frame sync* means also the line start pulse generated at the 60th bit of the sequence.

HRPT—High Resolution Picture Transmission. Digital system for the transmission of environmental images generated by the AVHRR. The HRPT is provided in a split-phase format to the S-band transmitter.

LOS—Loss Of Signal.

NRZ—Nonreturn-to-Zero. A code representing the binary values (logic 0 and logic 1) with a static level maintained throughout the data cell. Consequently, there is no transition between two successive bits of the same binary value.

Sample strobe—Word strobe divided by 5; therefore a pulse at the end of all the words of one of the 5 multiplexed images.

Split-Phase-Low—Also known as Biphase-Low or Biphase Manchester code, represents data with a level transition in the middle of the data cells. The direction of the transition indicates the binary value of data. A logic 0 is defined as a low-to-high transition in the middle of the data cell, and a logic 1 as a high-to-low mid-bit transition. In Manchester, the serial data stream contains both the clock and the data, with the position of the mid-bit transition representing the clock and the direction of the transition representing data. Therefore, there is no phase variation between the clock and the data.

Word strobe—A pulse generated at the last bit of each word.

Photo 7—Southern Italy and Sicily on the monitor of our weather satellite receiving station.

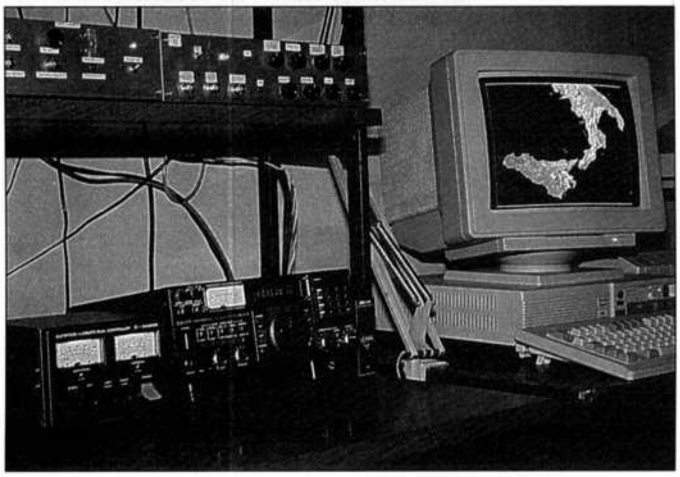


Photo 8—The entire HRPT receiving/control setup.

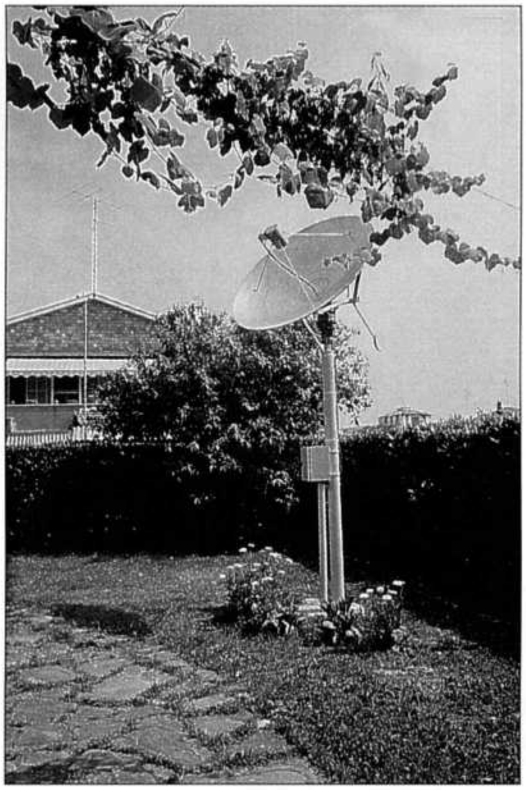


Photo 9—The NOAA HRPT dish antenna.

or lines have gotten lost because of noise).

The software is free, is not copy-protected, and can be downloaded as METEO.ZIP from the ARRL BBS at 860-594-0306 and meteo.zip from the *QEX* FTP site (**ftp://arrl.org/pub/qex/**).

The Pictures

All the pictures have been received and printed by the authors. Our fax machine is a vintage Muirhead-Jarvis Picture Receiver, type D-356D, It produces 8 × 10-inch images on photographic paper. The software for printing the pictures was developedby Cludio Pagnani; this article does not describe the related hardware.

Dynamic Resistance in RF Design

A straightforward introduction to an oft-misunderstood phenomenon.

By William E. Sabin, WØIYH

In this article we will discuss a concept that is very commonly found in almost every aspect of RF active circuit design. Electronic devices of various kinds routinely exhibit relationships between voltage and current that are not explained by the rules of ordinary dc or ac circuit principles as found in elementary texts. We will review some additional aspects of resistance and hopefully get some perspectives that will help us to better understand the behavior of many of the RF circuits that we encounter.

Resistance, Static and Dynamic

The resistance of an ordinary resistor that is driven by a sine wave of voltage is the ratio of the voltage across the device at any instant to the current through the device at the same instant. This value of resistance is constant (*static*) at each point of the sine wave. The power dissipated in the resistor at each instant is the product of voltage and current at that instant. The average power is the total energy that is dissipated over one complete sine wave, divided by the time duration of the sine wave.

The *dynamic* or *ac* resistance of this same resistor is the ratio of a very small *change* in the instantaneous voltage at any point in the sine wave to the corresponding *change* in the instantaneous current. In other words, dynamic resistance is the *slope* of the curve of voltage versus current. Fig 1 shows a plot of voltage versus current for an ordinary resistor. The slope of this plot (the dynamic resistance) is seen to be identical to the static value. That is, for the ordinary resistor:

$$R_{\text{static}} = R_{\text{dynamic}} = \frac{V}{I} = \frac{\Delta V}{\Delta I} \qquad \text{Eq 1}$$

On the other hand, Fig 2 shows the voltage versus current of a generic "device" of some kind. The static resistance at point (a) is the ratio of V to I. The dynamic resistance, the slope, does not have the same value that it had in Fig 1; the ratio of V to I is not the same as the ratio of ΔV to ΔI.

Consider the following examples of dynamic resistance:

1. In a vacuum tube the *plate resistance* is a measure of the ability of a change in plate voltage to produce a change in plate current, when the grid voltage is held constant. Fig 3 shows the plate voltage versus plate current characteristic of a triode for various values of constant grid voltage. The slope of each curve is the reciprocal of r_p, the plate resistance. Note that the slope varies somewhat for different values of the grid voltage. At very low

1400 Harold Dr SE
Cedar Rapids, IA 52403
email: sabinw@crpl.cedar-rapids.lib.ia.us

plate voltage the plate resistance becomes very low.

2. In a transistor, either BJT, JFET or MOSFET, a similar effect exists. It is called the Early (name of a person) effect. For example in a BJT, for a constant base current, the collector current is affected by a change in collector voltage.

3. A forward-biased diode has an ac resistance that is much different from its dc resistance. So do Zener diodes, tunnel diodes, SCRs, MOVs, etc.

Power Dissipation Due to Dynamic Resistance

Return to Fig 2. Suppose that a sine wave of voltage of amplitude V_s, centered on point (a) and varying between points (b) and (c), is applied to the device through a coupling capacitor. A sine wave (almost, because of the almost constant slope) of current I_s flows through the device. We want to know the total average power dissipated by the device. Using the common method of calculating this we proceed as follows:

$$P_{ave} = \frac{1}{2\pi}\int_0^{2\pi}(V + V_s \sin\theta)(I + I_s \sin\theta)\,d\theta$$

$$= \frac{1}{2\pi}\int_0^{2\pi}(V + v_s \sin\theta)\left(I + \frac{V_s}{R_d}\sin\theta\right)d\theta$$

$$= \frac{1}{2\pi}\int_0^{2\pi}\left[VI + \underbrace{\frac{VV_s}{R_d}\sin\theta}_{\text{integral=0}} + \underbrace{V_s I \sin\theta}_{\text{integral=0}} + \underbrace{\frac{V_s^2}{R_d}\sin^2\theta}_{\text{use trig identity}}\right]d\theta$$

$$= VI + \frac{V_s^2}{2R_d}$$

$$= VI + \frac{V_s(\text{rms})^2}{R_d} \qquad \text{Eq 2}$$

The total power is the dc VI product plus the additional ac power that would be dissipated by the superimposed sine wave in an actual physical resistor whose value is R_d. It is very important to this discussion to note, however, that the ac power is not dissipated in a "thing" called R_d, but is dissipated in the device itself. This point is sometimes not well understood and it is not a trivial observation, as we shall soon see. In order to keep things straight we must be precise in our language.

Fig 2 also tells us how to *measure* the value of R_d. If we capacitively couple an ac signal generator to the device and measure the ac current into the device, the value of R_d is easily calculated. In other words, the device acts as a *load* for the ac generator. This is exactly how a vector impedance meter (for example, my HP4815A) works.

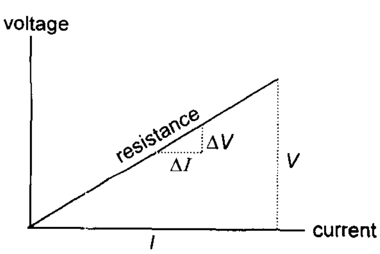


Fig 1—Static and dynamic resistance of an ordinary resistor.

Efficiency of a Class-A Amplifier

To get further perspective on dynamic resistance, consider the simplified, idealized, perfectly linear class-A amplifier circuit in Fig 4. The ac sine-wave grid drive is sufficient to drive the instantaneous plate voltage from zero to twice the dc plate supply voltage. The load resistance is such that plate current varies from zero to twice the dc plate current. Assume initially that the tube has no plate resistance *effect*. That is, plate voltage changes do not by themselves affect plate current. The efficiency of this amplifier is 50%. It is calculated in the following manner and assumes only that plate voltage excursions and plate current excursions are as large as theoretically possible:

$$P_{dc} = V_{bb} I_{bb}$$

$$P_{load} = V_{ac} I_{ac} = \frac{V_{bb}}{\sqrt{2}} \cdot \frac{I_{bb}}{\sqrt{2}} = \frac{V_{bb} I_{bb}}{2} = 0.5 P_{dc} = 50\% P_{dc} \qquad \text{Eq 3}$$

Now suppose that an appreciable plate resistance effect occurs in the tube. The following actions occur:

1. When the plate voltage is increasing, the plate current wants to decrease, as shown in the sine wave diagrams. (The plate voltage and plate current excursions are 180 degrees out of phase).

2. The plate resistance effect tries to increase the plate current. In other words, the decrease of plate current is retarded somewhat.

3. When the plate voltage is decreasing the plate current wants to increase.

4. The plate resistance effect tries to decrease the plate current. In other words, the increase of plate current is retarded somewhat.

5. As a result of this action, the ac plate current, and therefore the ac load voltage, and therefore the ac load power, are smaller.

6. Since the dc power is constant the efficiency of the amplifier is reduced by the presence of the plate resistance effect.

7. But now suppose we increase the ac voltage on the grid of the tube so that the plate voltage and plate current variations return to their full amplitudes. Assume enough "head

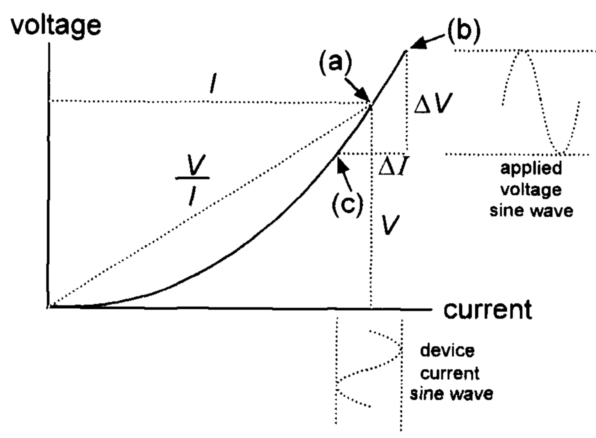


Fig 2—Static and dynamic resistance of a generic "device."

room" is available at the grid that we can do this.

8. The ac load power returns to its maximum value and therefore the efficiency returns to it original 50% value.

9. Whereas before in Fig 2 we attributed the additional power dissipation to the dynamic resistance of the device, in this case the power dissipation attributed to plate resistance has been reduced to precisely *zero*.

10. This experiment can be repeated at lower values of efficiency, say 25% or so. This value can be restored, after the plate resistance effect is enabled, just as we have described.

This example illustrates forcefully the importance of keeping dynamic resistance in the proper perspective. Another way to get a handle on this example is to consider the linear approximation to the triode tube using small ac signal quantities:

$$i_p = g_m e_g + \frac{e_p}{r_p} \qquad \text{Eq 4}$$

In words, an increase in plate current can be caused either by an increase in grid voltage, multiplied by the trans-conductance, or by an increase in plate voltage, divided by the plate resistance. But from Fig 4:

$$i_p = -\frac{e_p}{R_L} \qquad \text{Eq 5}$$

If we combine these two equations we get the following result:

$$i_p = \underbrace{\left[\frac{g_m}{1+\frac{R_L}{r_p}}\right]}_{g_{m'}} e_g \qquad \text{Eq 6}$$

The term in brackets is the *effective* transconductance of the tube, in the presence of the plate resistance effect and a load resistance R_L. It is made smaller if the plate resistance is comparable to the value of the load resistance. And if this effective transconductance is reduced, an increase in grid voltage restores the plate excursion to its full value.

If we compare the experiments that we performed in Figs 2 and 4 we see something interesting. In Fig 2 the device was used as a *load* (dumped power into it) in order to measure its dynamic resistance. In Fig 4 the device is a *source* of power (an amplifier) and a somewhat different conclusion regarding the effect of dynamic resistance was made. In Fig 4 we concluded that the introduction of dynamic resistance has no bearing on the efficiency of the amplifier circuit, if the grid drive can be increased sufficiently. A transistor would behave the same way.

There is one other thing about dynamic resistance. It is always associated with an actual resistance of some kind. In Fig 2 it is related to the "device" at point (a). In Figs 3 and 4 it is related to a vacuum tube or transistor. It could also be one of the other devices that we mentioned. Somewhere in the circuit, at least in actual

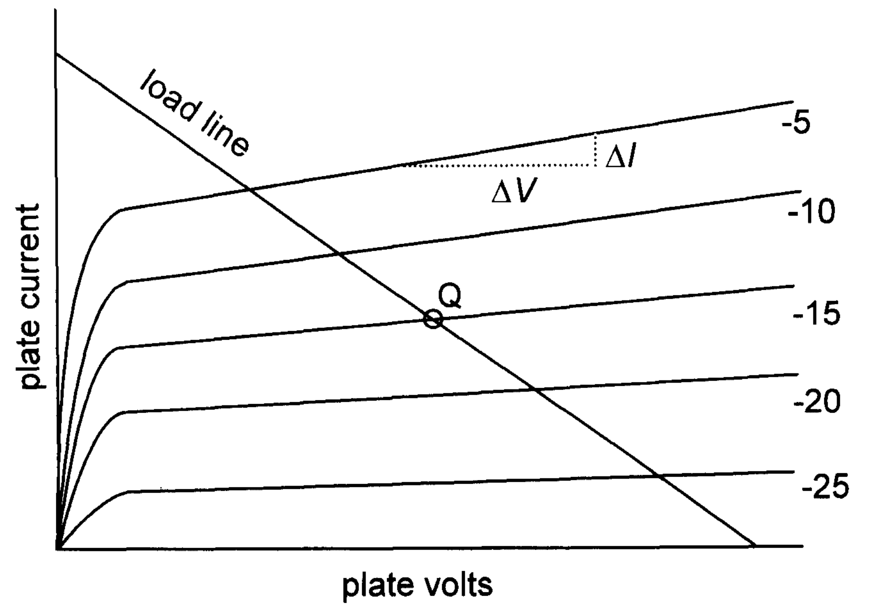


Fig 3—Typical plot of plate voltage versus plate current for a triode vacuum tube. Each curve is for a constant value of grid voltage. Also shown is a resistive load line and an operating point, Q. Plate resistance varies slightly with grid voltage.

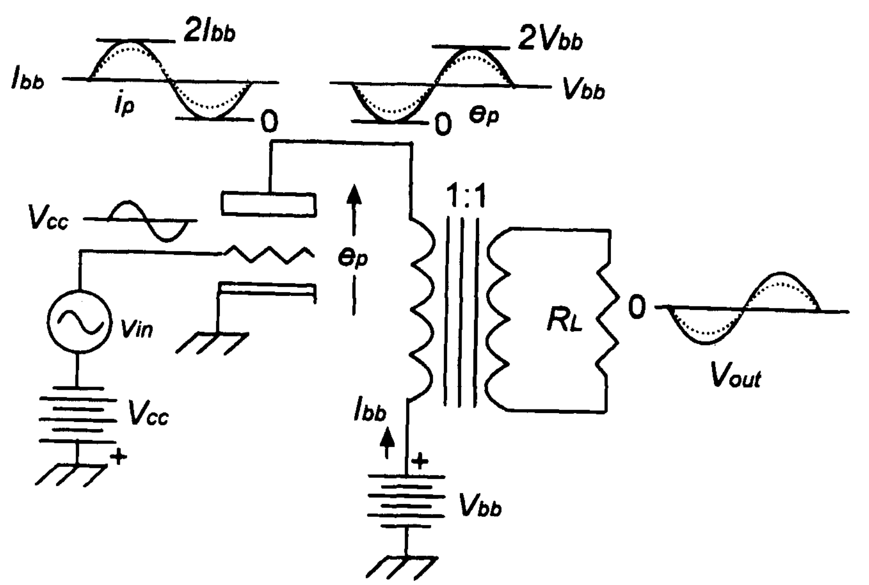


Fig 4—Simple triode amplifier circuit to demonstrate the efficiency with and without plate resistance.

practice, is a power-dissipating element of some kind. The relationship is not always obvious. In addition to the tube/transistor or other "device" it may involve driver stage output variations or even such things as bias resistors, power-supply regulation or line-voltage fluctuations.

Feedback and Dynamic Resistance

Fig 5 shows an amplifier circuit that incorporates negative feedback. A portion, ß, of the output voltage V_o is combined with the input V_{in} to produce the signal voltage V_s. The output resistance of the amplifier can be measured by connecting a signal generator at the output and measuring the current that is "dumped" into the amplifier output. The drive signal, V_{in}, is placed at zero for this measurement. An elementary circuit analysis shows that the small-signal output resistance with feedback is given by:

$$R_{OUT} = \frac{V_t}{I_t} = \frac{1}{\frac{1}{r_p} + g_m\beta} = \frac{r_p}{1+\mu\beta} \text{ where } g_m = \frac{\mu}{r_p} \qquad \text{Eq 7}$$

The constant μ will be recalled as the amplification factor of the tube. A second way to measure the output resistance is to connect a load resistor as shown in Fig 5 and apply an input signal V_{in} to the amplifier. We then vary the value of the load resistor a little and observe the change of output voltage. The output resistance is then:

$$R_{OUT} = \frac{V_1 - V_2}{\frac{V_2}{R_2} - \frac{V_1}{R_1}} = \frac{V_1 - V_2}{I_2 - I_1} \qquad \text{Eq 8}$$

where V_1 and V_2 are the values of V_o for each load. From this approach we see that the dynamic output resistance is related to the concept of "regulation," just as in a power supply. In the limit, if R_{OUT} is zero (that is, if V_1=V_2) the circuit is a constant voltage source. If R_{OUT} is infinite (that is, if I_1=I_2) the circuit is a perfect current source. This method may have a small error if a tuned circuit is slightly detuned by the small change in load resistance (some retuning may be needed). We must also be sure that the plate/collector voltage and current excursions (max and min) do not get beyond their acceptable limits.

Refer now to Fig 6, a tube (or transistor) shown as an ac current source with a dynamic resistance R_{OUT} (due to plate/collector resistance and feedback) and a load resistance R_L in shunt. The dc supply to the device is also shown. If R_L is equal to R_{OUT} then the circuit is impedance matched and the power in R_L is maximum. Consider also two other situations:

1. R_L is much smaller than R_{OUT}. Most of the RF current flows into R_L, but because R_L is small the power output is also small.
2. R_L is much larger than R_{OUT}. In this case, the RF plate voltage is limited by the value of R_{OUT}. Also, the peak RF plate voltage is limited by the B+ dc voltage. Therefore the RF power output is again small.

But there are very often times when we do not want the maximum output. Instead, we are adjusting the load and the drive level for best intermodulation performance, harmonic levels, tube/transistor dissipation or power supply capability. This adjustment may, or may not, coincide with impedance match. Maximum output and "best" output are frequently not the same adjustment (or they may be; we cannot rule it out).

And yet, as we have demonstrated, the *efficiency* of the plate/collector circuit, which is the ratio of RF output power to dc plate/collector power, maximum or otherwise, is not impaired by the value of dynamic output resistance. The real "movers and shakers" are:

1. The amplitude and the wave shape of the RF voltage (as a percentage of the dc supply voltage) as influenced by tuned circuit flywheel effect (loaded Q).
2. The amplitude and wave shape of the RF current (as a percentage of the dc supply current) as influenced by the value of R_L and by the tube's plate current conduction angle.

These are the things that determine the efficiency. The power loss due to the dynamic resistance is compensated as previously discussed by an increase in the drive level from the previous stage. The output power is not reduced, as it would be if R_{OUT} were a real resistor. The class-A amplifier (Figs 4 and 5) can then achieve its theoretical 50% efficiency and we can see how classes AB, B and C amplifiers can achieve real-world efficiencies greater than 50%.

Nonlinear Dynamic Resistance

In Fig 2, if the dc operating point is at point (a), a much larger superim-

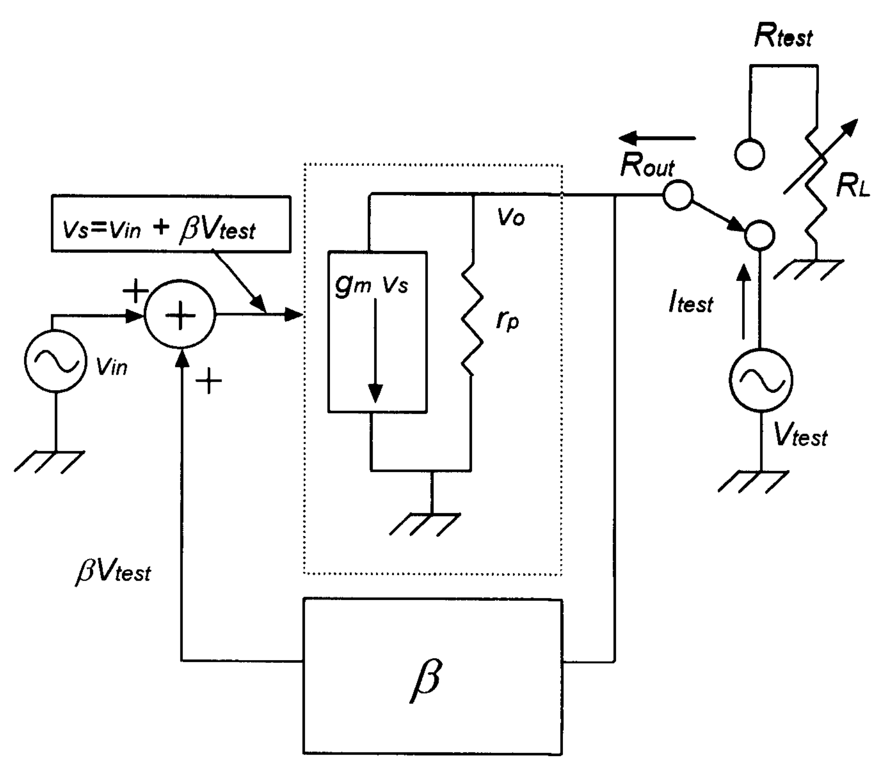


Fig 5—Measuring the output resistance of a class-A amplifier with negative feedback.

posed sine wave of voltage, extending into the highly nonlinear regions, will produce a highly distorted wave of current in the device. The dynamic resistance of the device (the slope) also varies considerably over a cycle and *its* waveform might look something like Fig 7. An amplifier that is overdriven and/or improperly loaded might show large variations like this. The average value can become quite small, compared to a normally driven and properly loaded amplifier. These instantaneous variations in dynamic resistance constitute *one* of the causes of intermodulation distortion that must be made as small as possible. Negative feedback is very helpful in this respect. That is, for a given voltage swing, distortion due to nonlinearity in dynamic resistance is reduced as compared with the absence of feedback.

On the other hand, if there is a high-Q tuned circuit operating, its flywheel effect will be very helpful in averaging out the effects of instantaneous variations in dynamic resistance over many cycles of RF. But as the peak-to-peak amplitude changes with low-frequency modulation the average value can also change. This is also a distortion producing mechanism that is helped by feedback.

The Class-AB, -B or -C Amplifier

In the previous example a constant plate current flowed, even without an ac drive signal into the grid. We could then measure dynamic resistance very easily. In a class-AB, -B or -C amplifier we can measure the dynamic resistance only when the amplifier is being driven to the input and output levels at which it is supposed to operate. The measurement setup previously described would destroy a sensitive impedance measuring instrument. (RF lab personnel of my acquaintance are unhappily familiar with this phenomenon). A good way to describe it is to say that the amplifier must be "pumped" with signal to get it operational, just as a mixer must be pumped by a local oscillator to make it work.

In the class-B or -C amplifier, the plate current flows for less than the full cycle. The plate resistance effect is *defined* only during the plate conduction interval, and the full-cycle average value is calculated using only those values. Plate resistance is not "infinite" during the off time.

In order to limit the scope and size of this article, we will avoid getting

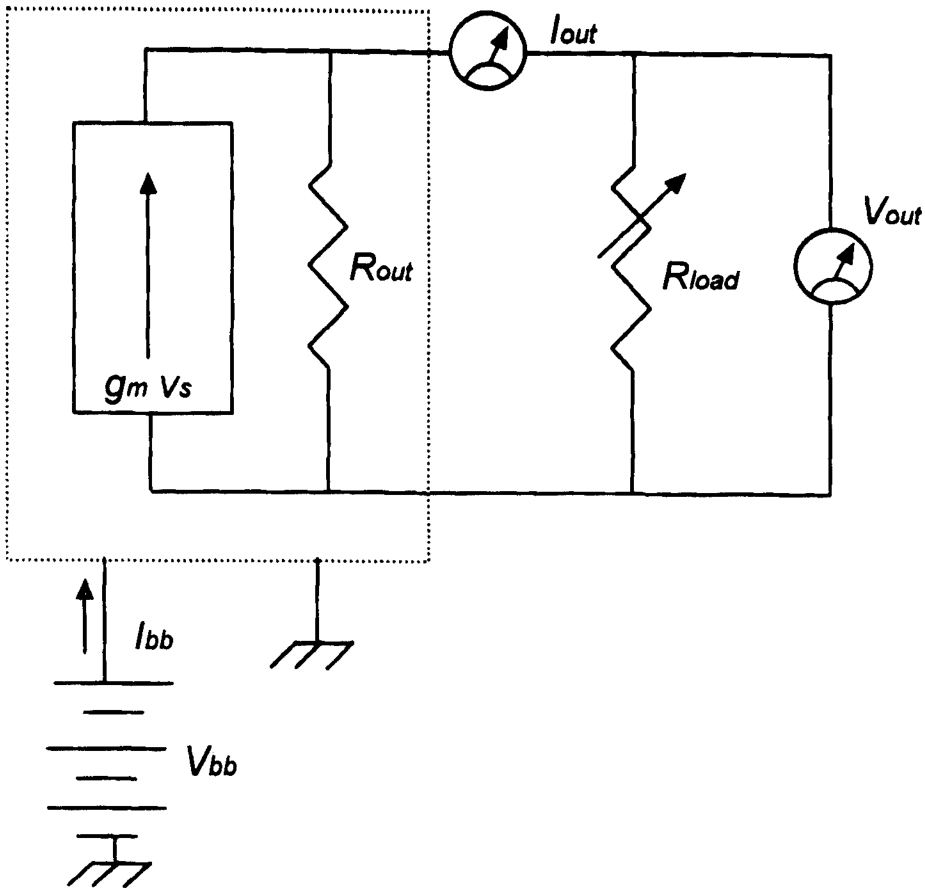


Fig 6—A current generator is impedance matched when the load resistance is equal to the output resistance of the source.

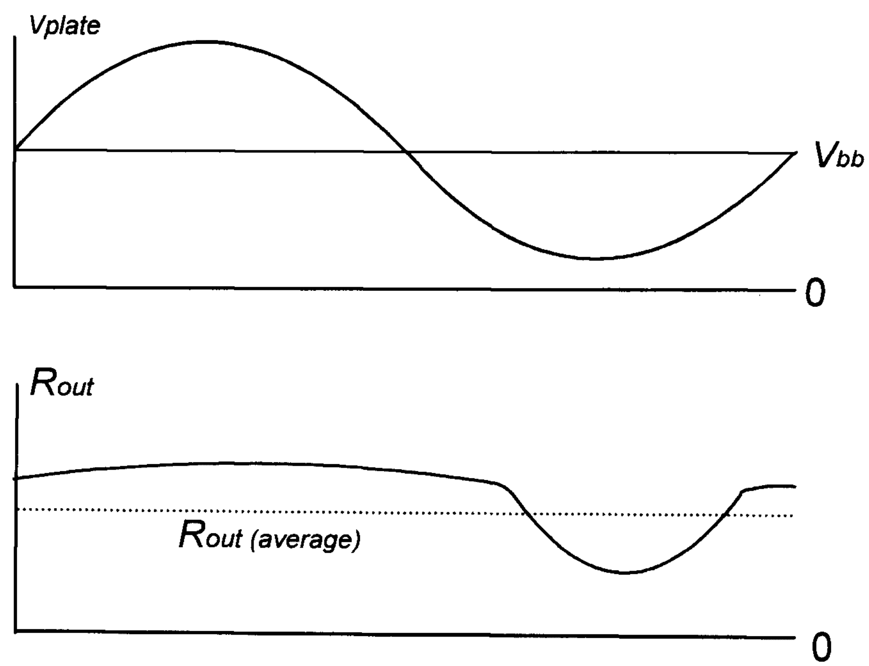


Fig 7—Variations of dynamic resistance over one cycle of the RF voltage across a device.

involved with the details of measuring the output resistance of these types of amplifiers, but instead suggest Note 1 and the additional references found there. It will be sufficient to say at this point that the value of output resistance R_{OUT} is important for the following reasons, among others:

1. R_{OUT} can reduce the selectivity of a tuned circuit *in the neighborhood* of the resonance frequency, not because it is a dissipater of RF power, but because either the plate resistance effect or feedback, or both, operate in a way that tends to maintain a somewhat more constant RF voltage as the frequency departs from center.
2. There are some situations in which we want the load resistance to be the same as the dynamic output resistance R_{OUT}. In this case we find that small variations in load resistance produce very little reduction in output power. In other words the curve of output power versus load resistance is flat (slope=0) in this particular region. This concept has been used from time to time. Negative feedback is often used to achieve this goal.
3. If R_{OUT} is reduced by feedback the linearity of the amplifier is greatly improved.

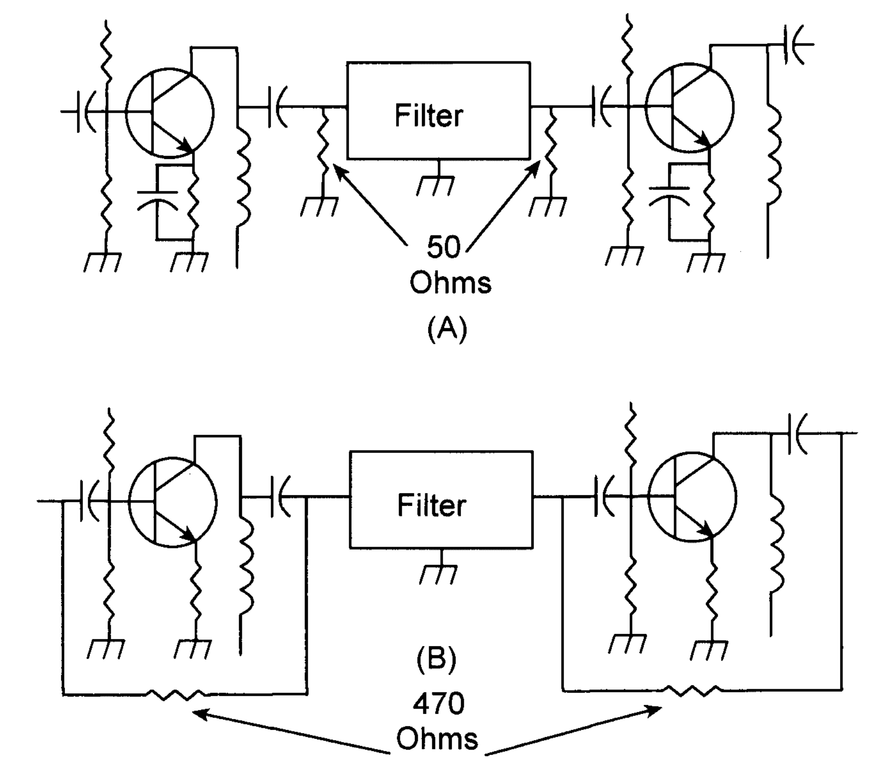


Fig 8—Showing how feedback reduces the dynamic input and output resistances of two amplifier stages, thus reducing loss of signal power.

Plate Modulation

Plate modulation (AM) continues to be a very important application of plate resistance. This effect allows the audio-frequency plate voltage variations to vary the amplitude and also to some extent the conduction angle of the plate current pulse of the class-C tube. This is an almost perfect *multiplication* process in the time domain that produces high-quality sidebands in the frequency domain. The tubes that are used like to have a fairly low value of plate resistance. Zero-bias class-B tubes like the 811A, for example, are not preferred for plate modulation. Instead, use a pair of 812As, and use the 811As in the class-B modulator. This was a popular ham rig in the AM days of Amateur Radio.

Filter Terminations

Fig 8 shows an interesting application of dynamic resistance, a receiver IF filter that must be properly terminated with 50 Ω at each end. In part (a) the resistors burn up valuable signal power and make it more difficult to keep the noise figure as low as possible. In part (b) negative feedback is used to create dynamic resistances that terminate the filter properly and recover much of the lost signal power.

Techniques roughly similar to this are widely used in receiver design at frequencies up to 1 GHz.

Also, the dynamic resistance in this case, and in nearly all cases, is not a source of *thermal* noise, and this should be remembered in computer noise modeling.

Conclusions

There are additional aspects to this subject that are beyond the scope of this article.

The limited objective has been to deal briefly with some of the more basic ideas. The hope is that we have helped to put these ideas into a sharper focus that will be useful.

[1]Sabin, W. E. and Schoenike, E. O., Editors, *Single-Sideband Systems and Circuits*, Second Edition, Chapter 14, "High Power Amplifiers," by W. B. Bruene, New York, McGraw-Hill, 1995. This chapter also contains an excellent discussion of negative feedback in RF power amplifiers.

The Principles of HF Radio Direction Finding Loops

The loop is the standard HF DF antenna. Here's why.

By George Brown, BSc, PhD, CEng, FIEE, G1VCY

Introduction

Many radio amateurs derive much pleasure from radio direction finding (RDF)—mostly through participation in the so-called "fox hunts." But it is a technique that may be put to more serious purposes, such as in tracking down an interfering transmitter. While most people will know how to put together a workable system, the theory upon which it depends is not so widely known. This article sets out to introduce the subject to the beginner in such a way as to make the design of RDF systems less of a hit-and-miss affair.

1, Langford Crescent
Benfleet, Essex SS7 3JP
England

The Aerial

Before the advent of the ferrite rod, the loop aerial was the favorite design; it remains so with many enthusiasts. Fortunately, the principle behind the RDF aerial remains substantially the same for both types. The loop aerial produces a signal null when broadside on to the transmitter, while the ferrite rod aerial does so when in line with the transmitter. In both cases, the null occurs in a direction perpendicular to the plane of the looped antenna conductor, for reasons we shall see. Because the geometry of the loop is more palatable than that of the rod, the loop aerial will be used as the subject of this description.

First, it is necessary to understand the basic fact that a vertical aerial of length h meters, placed in a vertical electric field of strength $E\ Vm^{-1}$, will develop an EMF (V) between its ends given by:

$$V = Eh \qquad \text{Eq 1}$$

Consider the loop aerial, Fig 1. PS and QR act as vertical aerials. If the plane of the loop is perpendicular to the signal direction, then the EMFs (V_1 and V_2) induced in PS and QR are equal in phase and in magnitude, but act in opposite directions, as shown, and thus cancel. The horizontal sections can be ignored when receiving a vertical wave, because no EMF is induced.

If the loop is turned until its plane is parallel to the wave, PS and QR now have different path lengths to the transmitter. The EMFs induced in the two vertical sections have the same magnitude as before, but have different phases. This is seen in Fig 2, where the phase difference (ϕ) is given by:

$$\phi = (d / \lambda)360 \qquad \text{Eq 2}$$

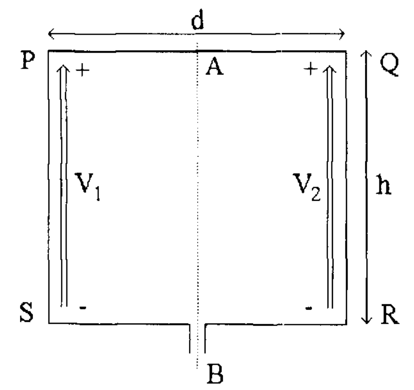


Fig 1—The loop aerial.

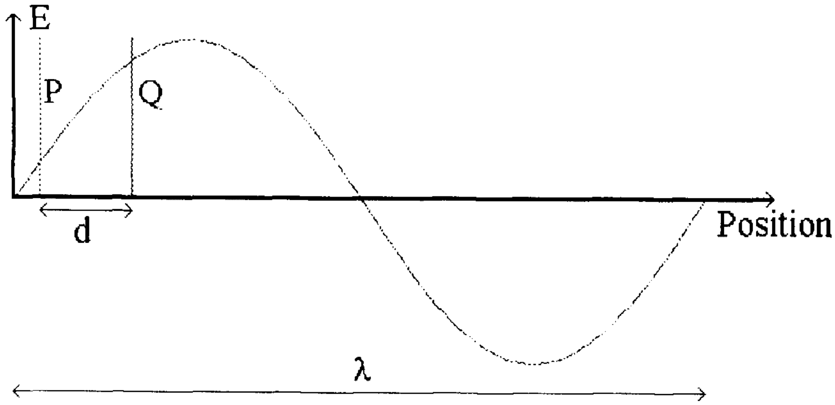


Fig 2—A phase difference exists between the two vertical sections.

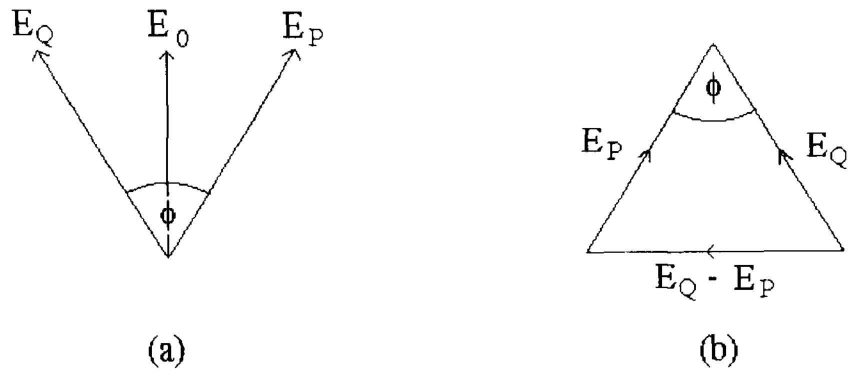


Fig 3—(a) Vector positions relative to E_0. (B) The loop EMF is 90° out of phase with the signal voltage E_0.

where
ϕ = phase difference in degrees
d = width of the loop
λ = wavelength of the signal

If E_{max} is the maximum EMF, and E_θ the EMF when θ is the angle between the loop plane and the signal direction, then $E_\theta = E_{max}$ when $\theta = 0°$, and $E_\theta = 0$ when $\theta = 90°$.

Or,

$$E_\theta = E_{max} \times \cos(\theta) \qquad \text{Eq 3}$$

Thus, E_{max} is dependent on three parameters:

(a) the received signal strength;
(b) the vertical height (h) of the loop; and
(c) the loop width (d) over the range $0 < d \leq \lambda/2$

From this, it is clear that in order to maximize E_{max}, h and d should be as large as possible. This is equivalent to saying that the loop area (A) should be as large as possible.

The loop width controls the phase angle, as shown in Eq 2. The RMS vectors are shown in Fig 3(a). E_o is the relative field direction at the loop center, thus making E_P and E_Q symmetrical about it. Fig 3(b) illustrates two important facts:

(a) $E = E_Q - E_P \propto \phi$
(b) E is perpendicular to E_o.

Therefore, although the EMFs induced in PS and QR are both in phase with the incoming signal (albeit at different times), the loop EMF is 90° out of phase with it.

Under most circumstances, the loop size is very much less than one wavelength of the signal, ie, $d << \lambda$, which means that ϕ is small and hence cancellation of the components from the two vertical sides is almost complete. To overcome this sensitivity problem, there are two principal approaches, both of which involve adding more turns to the loop. A multiturn loop can be achieved two ways: by winding more turns on the same form in a helical configuration, or by winding more turns in a pancake or spiral configuration. Whichever is chosen, the EMF developed by the aerial is N times as large as with a single turn, if N is the number of turns. (Whether the aerial is constructed as a loop or a rectangle does not affect the argument here.)

The helical and spiral approaches are not equivalent. The helical form (known as the *box* form when using a frame aerial) has three dimensions and thus will never give a true null when perpendicular to the transmitter. The spiral form retains its two-dimensional character and continues to produce a null. The compromise between the number of turns and the area is usually struck with a large area coil with relatively few turns.

Finally, there is the *vertical effect* (or antenna effect) to contend with. This is the situation when the loop is not exactly balanced electrostatically and the effects of the two vertical sections are not equivalent. This can be overcome (where it is a significant problem) by enclosing the loop in a Faraday cage, which makes the loop sensitive to magnetic fields but not to electric fields—hence the name "magnetic loop." The vertical effect can also be overcome (in most cases) by using a balanced input circuit (described later).

The Polar Diagram

Eq 3 shows that the output of the loop was given by:

$E_\theta = E_{max} \times \cos(\theta)$

E_{max} is the output signal when the plane of the loop is parallel to the signal direction. Although the $\cos(\theta)$ term implies that E_θ varies both positively and negatively about zero, this is not detected in practice because the receiver is not sensitive to the phase of the incoming signal. Consequently, the behavior is not proportional to

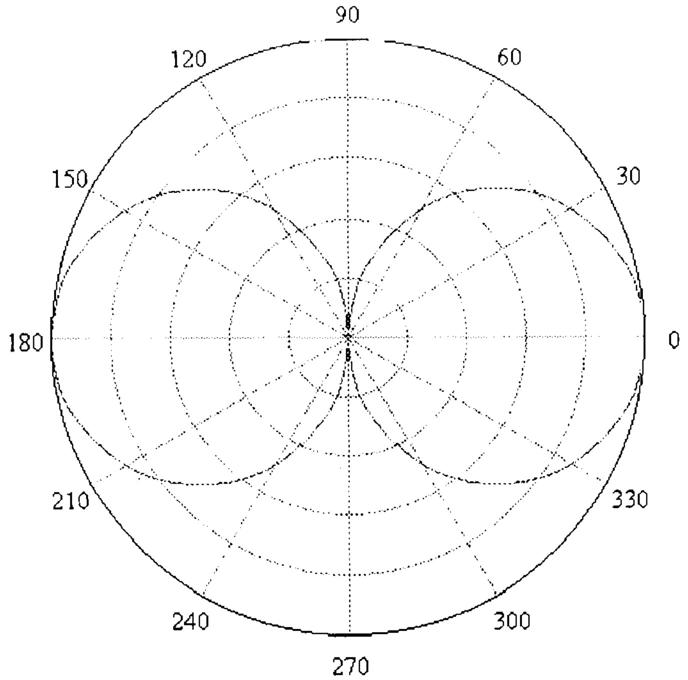


Fig 4—The polar diagram in the horizontal plane.

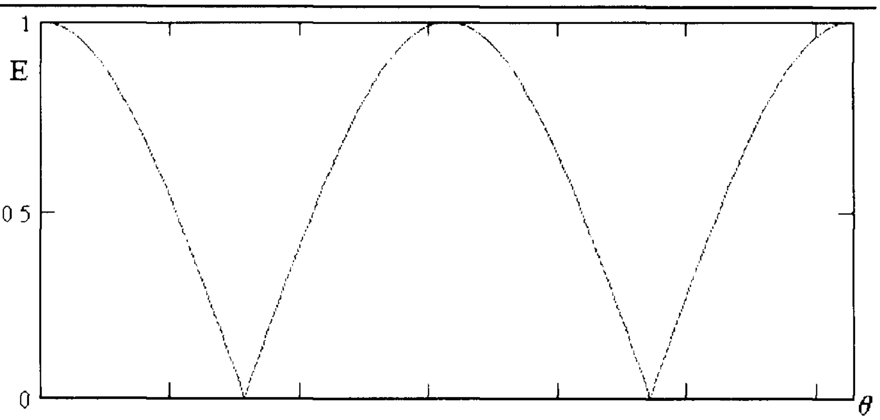


Fig 5—The polar diagram on Cartesian axes.

cos(θ) but to |cos(θ)|, the modulus signs meaning that the negative signs generated when θ lies between 90° and 270° are ignored. This produces the well-known figure-eight pattern shown in Fig 4. It is symmetrical mathematically because of the modulus function and intuitively because the loop has no way of knowing whether the signal is approaching from behind or from in front.

Used as it has been described so far, the loop aerial can detect, using any reputable receiver, the line along which a transmitter lies. At first sight, this is determined by turning the aerial to the position of maximum signal strength and stating that the transmitter lies in the plane containing the loop. This is not the most accurate method, however. A careful glance at Fig 4 will show that the polar diagram is very well-rounded in the positions of maximum sensitivity. As a result, the signal on the S-meter will remain constant for several degrees of rotation of the loop. However, all is not lost! Consider the two nulls; these are *very* sharp. If the loop is turned *until a null is found* (and this can best be done aurally rather than using the S-meter), this position will be found to be very well localized and the angle can be determined to a degree or so! Having found the null, the transmitter will lie along a line perpendicular to the plane of the loop. Fig 5 shows the same polar diagram plotted, this time, on Cartesian axes. The nulls show as cusps on this graph, indicating how much better defined they are than the maxima.

Sense Determination

From the polar diagram of Fig 4 it can be seen that there is a 180° ambiguity in the indicated direction to a distant station. It is necessary to find the correct bearing from the two indicated positions. This is called *sense finding*.

To understand how the sense of a signal may be deduced, it is necessary to consider the signals induced in the sections PS and QR of the loop shown in Fig 1. Suppose that, at a given instant, when the plane of the loop lies in the signal direction, $V_1 > V_2$ for a signal coming from the left of the diagram. (Note: This is just an assumption—it must not be taken to imply that V_1 is always greater than V_2 when the transmitter is to the left!) This would induce a clockwise current in the loop. An identical signal coming from the right (known as the *reciprocal bearing*) would make $V_2 > V_1$, inducing an anticlockwise current in the loop. Although the receiver cannot discriminate between these, the signals received from opposite directions *are* different, being 180° out of phase. To make this phase difference apparent to the receiver is the function of the *sense finder*.

A single vertical aerial has an omnidirectional polar diagram in the horizontal plane, ie, it responds equally to signals from all directions (all values of θ). If the induced signal in such an aerial is in phase with the maximum signal induced in the loop then, when the loop is rotated 180° from this position about AB (Fig 1), the signals will be out of phase. The addition of these signals in the receiver would thus produce a null on the reciprocal bearing (if the two received signals were equal in magnitude) *and nowhere else*. This is shown in Fig 6.

This represents an ideal case, because it has been assumed that the maximum voltage (E_{max}) developed by the loop is equal to the signal (E_{vert}) from the vertical aerial; this produces a true null on the reciprocal bearing. In general, this perfect nulling will not be the case. If, for example, $E_{max} < E_{vert}$, then the polar diagram of Fig 7 results, showing incomplete cancellation of the two signals when the loop is on the reciprocal bearing. This is not too critical as a pronounced minimum of sound in the loudspeaker is usually

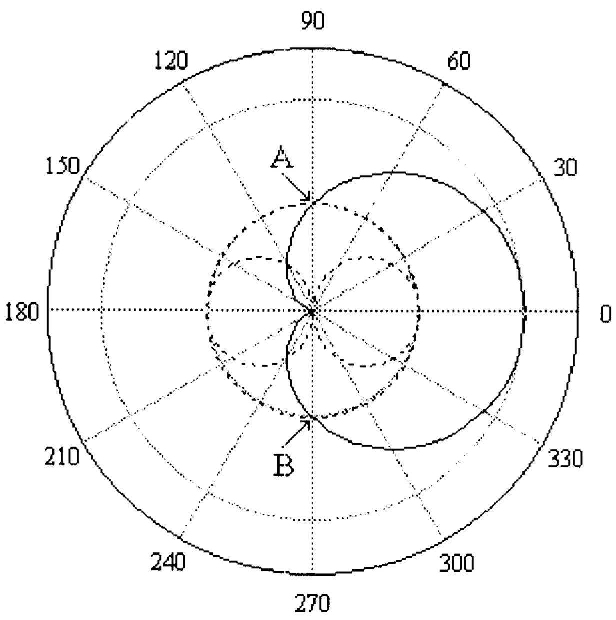


Fig 6—The cardioid combined response of a vertical aerial (omnidirectional) and the loop (figure-eight) produced when both aerials generate the same magnitude of signal at the receiver input.

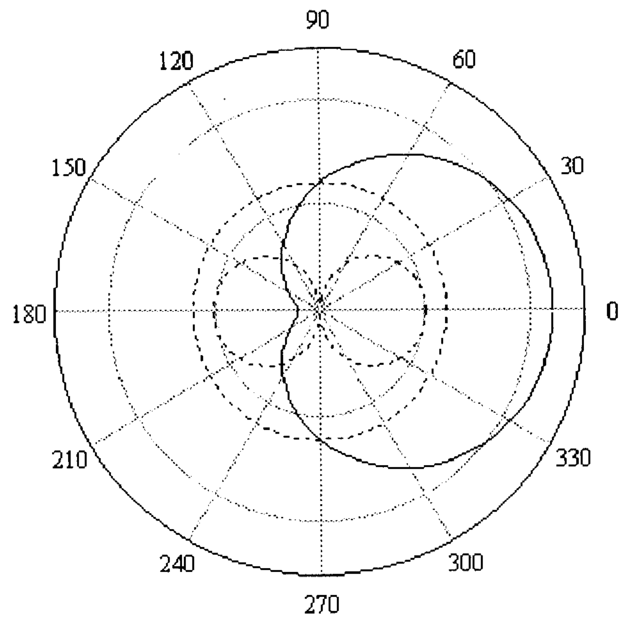


Fig 7—The limaçon combined response of a vertical aerial and the loop, produced when the signal from the loop is less than that from the vertical aerial at the receiver input.

apparent. This does not occur if $E_{max} > E_{vert}$, as Fig 8 shows. *Two* nulls are produced symmetrically about the reciprocal bearing. The angle subtended by the two nulls will increase as E_{vert} decreases with respect to E_{max}. Generally, therefore, the presence of a null does *not* necessarily indicate the true reciprocal bearing.

In Practice

It should be evident by now that the combination of the signals from the two aerials is one of the most significant practical aspects of accurate direction finding. The other (the construction of the loop or frame) depends much upon the application and whether the system is for fixed-station or mobile/portable use. Different solutions also apply for RDF at higher frequencies, such as at VHF.

For portable use, a small loop may be used, usually of diameter 30 cm or less. There is a rule of thumb that says the total conductor length should be less than 0.08 λ for a good cardioid polar diagram.[1] The loop can be screened if necessary. The use of a coil or coils wound on a ferrite rod has found much favor, not least because of its inherent compactness. Used with a sense aerial, it performs in the same way as the loop, except that the figure-eight pattern of the rod is perpendicular to that of the loop in the horizontal plane. A novel HF RDF system is described by W6QYT et al, which devotees may find interesting.[2]

The connection of a loop and sense aerial to a receiver is relatively simple, as Fig 9 shows, although there are some important features here. First, note the way the loop is balanced about ground, thus reducing the vertical effect. Second, the sensing aerial is not connected permanently, but via a switch. The system is not normally used with the sensing aerial in circuit, but as follows.

Consider Fig 6. Suppose the transmitter is on a bearing of 90°, using the angular scale provided. A null is first found with the loop alone. The pointer on the loop's angular scale indicates 90° although, at this time, the absolute value could be 90° or 270°. (The pointer on the loop's angular scale is usually set broadside on to the plane of the loop, because a null indicates the bearing or the reciprocal bearing to the transmitter.) Then the sensing aerial is switched in, producing the cardioid pattern shown in Fig 6. The signal is being received at point A of the cardioid. If the loop is turned clockwise, the signal strength will fall if the transmitter is at the indicated angle of 90°; if it rises, the signal must be coming in at B, and the transmitter must be on the reciprocal bearing of 270°. The important thing is that the null is found with the loop alone because it is a much sharper null than that of the true cardioid, and is very much sharper than that of the generic limaçon, produced when the signals from the two aerials are not equalized. Operators have their own favorite methods of resolving ambiguities, but this method is probably the simplest of all, requiring only *one* precise reading followed by a quick observation of the S-meter as the loop is rotated slightly clockwise.

You may have noticed that there is potentially an omission from Fig 9. No means is provided for equalizing the signals from the loop and the sensing aerial. This is where pragmatism scores over theory; the sensing aerial is usually a simple telescopic whip, the length of which can be varied *in situ* to match the output of the loop, a solution which does not lend itself to insertion in a circuit diagram!

Neither has mention been made of how the 90° phase shift is introduced to enable the loop output and the vertical aerial output to be added in phase. This is produced by the inductive coupling between the loop and the tuned circuit of Fig 9. The current in

[1]Notes appear on page 23

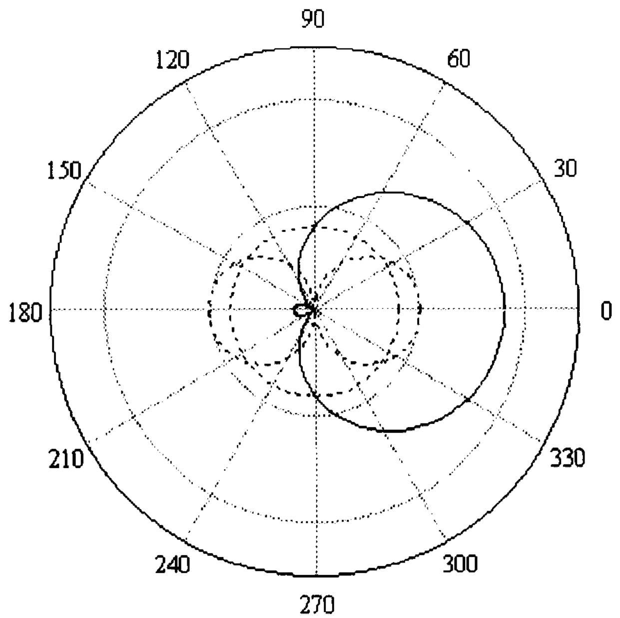


Fig 8—The limaçon combined response of a vertical aerial and the loop produced when the signal from the loop is greater than that from the vertical aerial at the receiver input.

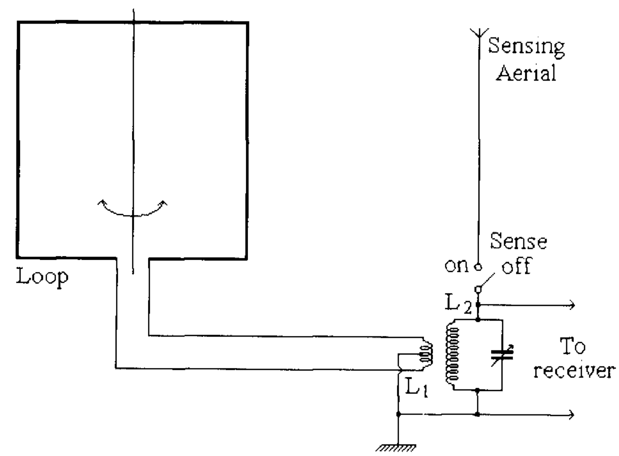


Fig 9—A typical input circuit for combining the loop and the sense aerials in order to produce the 90° phase shift required by the theory.

L1 is in phase with the EMF in the loop, and hence is in phase with the incoming wave. The EMF induced in L2 lags (or leads, depending on the winding sense of L2) the current in L1 by 90°. This required phase shift enables the sensing aerial to be connected directly to L2.

When using the RDF system at HF, it should be noted that ground-wave signals give the best results for the sharpness of the null and hence the most accurate bearings. Sky-waves are subject to multipath distortion and scatter, and can give rise to very poor (or even missing) nulls.

Summary

There is nothing inherently difficult about direction-finding. The aerial requirements are minimal, and the combining circuit is equally simple. Construction of the equipment can follow the operator's preference, as can the operation. I've set out to explain some of the basic principles underlying radio direction finding, and hope to have explained the two fundamental concepts of practical RDF: searching for a null with the loop only (not with the loop and sensing aerial together); switching in the sensing aerial after the null has been located, and obtaining an indication of whether the signal increases of decreases when the loop is turned clockwise.

The use of RDF systems at HF has been touched upon; at HF the use of ground-wave signals is to be preferred. There is always plenty of scope for originality, and it is hoped that this article has pointed the way to some more of these innovations.

Good hunting!

Notes

[1] *The ARRL Handbook*, 1989, ARRL, p 39-41.

[2] Villard, O. G., Jr, W6QYT, et al, "Simple Equipment for HF Fox Hunting," *QST*, August 1994, pp 33-35.

RF

By Zack Lau, KH6CP/1

6-Meter Transverter Design

This month I present a design for a 6-meter transverter. You might say that this design is meant for illustrative purposes only—I expect anyone who builds a project like this to customize it to their needs. All the building blocks shown in Fig 1 feature 50-Ω input and output impedances. You can pick the parts of this design you think are best and integrate them into your own design. I don't intend to develop a set of boards for this project. If you are looking for something of that sort, check out Ten-Tec's popular kit with its 14-MHz IF.

The question every nonengineer asks about the Ten-Tec kit is, "why 14 MHz?" Very simply, it makes it practical to market an inexpensive kit that someone can put together and have a reasonable chance of meeting FCC spectral-purity requirements. A 28-MHz IF is a bit of a challenge—the second harmonic of the IF falls just outside the band where it can interfere with broadcast TV. Not that a 28-MHz IF can't be done, particularly if you narrow up the band-pass filters to cover just the first 1 MHz of the band. But the good layout and shielding needed are impractical if you want the project to be inexpensive.

One of the reasons hams like the idea of a 28-MHz IF is the availability of low-cost 10-meter SSB/CW radios. While such radios work—sort of—as an IF, you'll notice markedly better performance with a better radio. Unlike in satellite operation, where the bird limits the dynamic range of the signals you hear, the range of received VHF signals can be from extremely loud to extremely weak—at the same time. A prime example of this occurs with multihop E-skip. Single-hop signals are often ridiculously loud, while stations requiring multiple hops or scatter paths will often be right at the noise floor. It's not just dynamic range that counts, either; filtering also makes a difference. Expensive HF transceivers often have filters at two or more IFs. This gets around the problem of signals leaking past the filters, since you can typically add the rejection obtained at each frequency. As a lot of people have discovered the hard way, putting two 50-dB filters in series at one frequency rarely adds up to 100 dB of rejection.

Fig 1 shows the block diagram of the transverter. I chose a 144 to 148-MHz IF since I normally carry several 2-meter radios when I operate QRP portable. Turns out that this also works well in terms of spurious trans-

225 Main Street
Newington, CT 06111
email: zlau@arrl.org

mitter products. Since the QRP power limit is 10-W PEP, this was my design-goal power output. As for receiver performance, according to Ray Rector, WA4NJP, a 3-dB NF is adequate for 6-meter EME due to the high amount of sky noise.[1] He shows a plot that indicates minimum noise levels to be around 4000 kelvins. On the other hand, if you have a really bad location that looks into hills in every direction, it's certainly possible to have a much lower background noise temperature, possibly approaching the temperature of the ground. Of course, real locations often have man-made noise, so I doubt that anyone ever sees anything below 500 kelvins. I decided that the measured 2.9-dB NF and 10-dB insertion gain of this transverter is adequate for use with a sensitive 2-meter IF radio.

Fig 2 shows the T/R relay and the output low-pass filter, a straightforward design.

6-meter band-pass filters are used in the receive converter, the transmit converter and the common circuitry. Since these filters affect both the transverter performance and its ease of adjustment, some thought is required in selecting an appropriate design. Unlike in narrowband microwave applications, there is a distinct advantage to using more than two resonators in a 50 to 54-MHz band-pass filter. Not only are the filter skirts steeper, but you can make the passband flatter. In other words, as you add filter elements you can make the filter closer to the ideal rectangular filter. Such a filter has a flat response across the passband and infinite rejection everywhere else. There is also a distinct disadvantage: multiresonator filters can be difficult to adjust. Most people use a sweep network, which is ideally a tracking generator and a spectrum analyzer. It's also possible to use a sweep generator and a log detector, although spurious signals and responses are more of a problem with this technique. I've provided two band-pass filter designs (Figs 3 and 4). You can use either design for any of the BPF blocks of Fig 1. There are tradeoffs, though.

A software simulator, such as *ARRL Radio Designer*, is convenient for modeling real filters. You can also get more details by studying Anatol I. Zverev's *Handbook of Filter Synthesis*. (This classic, published by Wiley, will set you back over $100.) In these designs, I assumed Qs of 80 for the inductors and 300 for the capacitors. Fig 3 shows the 3-pole filter design that resulted from the design process. (An interesting exercise is to compare two- and three-pole filters—the three-pole filter is markedly better.)

The three-pole filter is, of course, more difficult to adjust than a two-pole design. So, why not just cascade a pair of two-pole filters? I did just that, using the filter circuit shown in Fig 4. One of the cascaded filters is in the circuitry common to both transmit and receive, while the separate transmit and receive chains each have a BPF within them. For transmit, this has the advantage of lowering the amount of broadband transmitted noise as the second filter is placed after several amplifier stages. The disadvantage is that cascading filters makes their nonideal passband responses even worse. True, you could cleverly design and tune up the cascaded filters as a unit to avoid this problem, but I don't think this is a realistic approach, par-

[1]Notes appear on page 30.

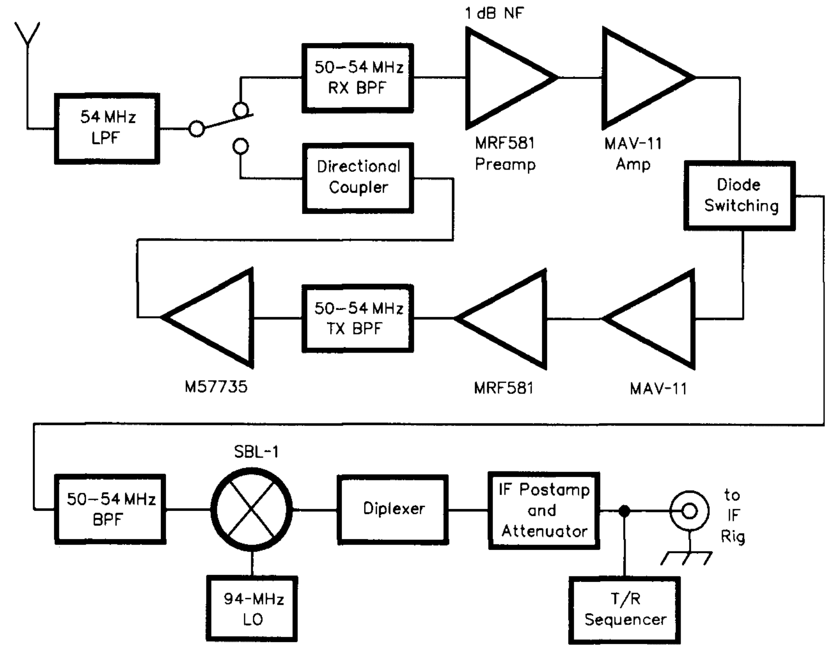


Fig 1—Block diagram of the transverter.

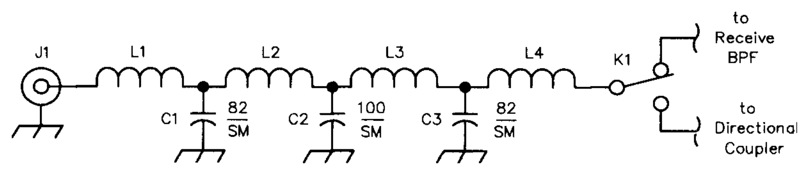
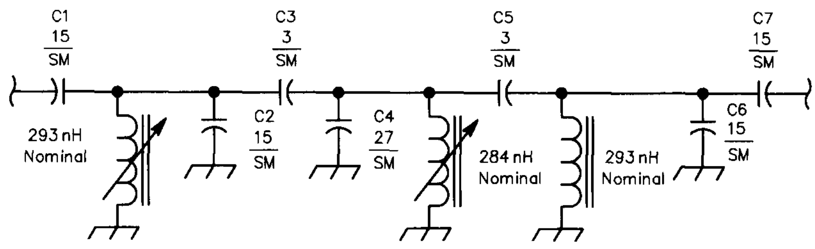


Fig 2—54-MHz low-pass filter and T/R relay. The filter has a 62-MHz –3dB cut off and 0.2 dB of insertion loss.

J1—UG-58 Type-N panel jack.
K1— Omron G5Y-1-DC12 relay (Digi-Key part Z24-ND).
L1, L4—90 nH, 5 turns no. 20 closewound, 0.19-inch inside diameter.
L2, L3—0.22 µH. 10 turns no. 20 closewound, 0.19-inch inside diameter.



Measured filter response relative to 1.2-dB insertion loss

Response (dB)	*Frequency (MHz)*
-0.4	50.0, 54.0
-1	49.4, 54.6
-3	48.8, 55.4
-10	47.6, 57.6
-20	46.0, 61.8
-30	43.8, 69.6
-40	41.2

Fig 3—A 3-pole band-pass filter design.
L1, L2, L3—0.3-µH adjustable inductors with a Q of 80. Toko America, Inc. type MC120 with case, part No. E526HNA-100301 (Digi-Key part TK-2708-ND).

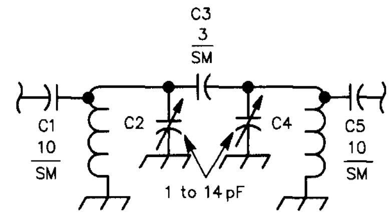


Fig 4—The low-loss, 2-pole band-pass filter.
C2, C4—1 to 14-pF Johanson piston trimmer capacitors.
L1, L2—8.5 turns no. 14 wire, space wound, 0.90-inch length. Tap 1/2 turn from top. Center-to-center spacing between coils is 1.0 inches. Coils are wound in the opposite sense from each other. Substitution is *not* recommended.

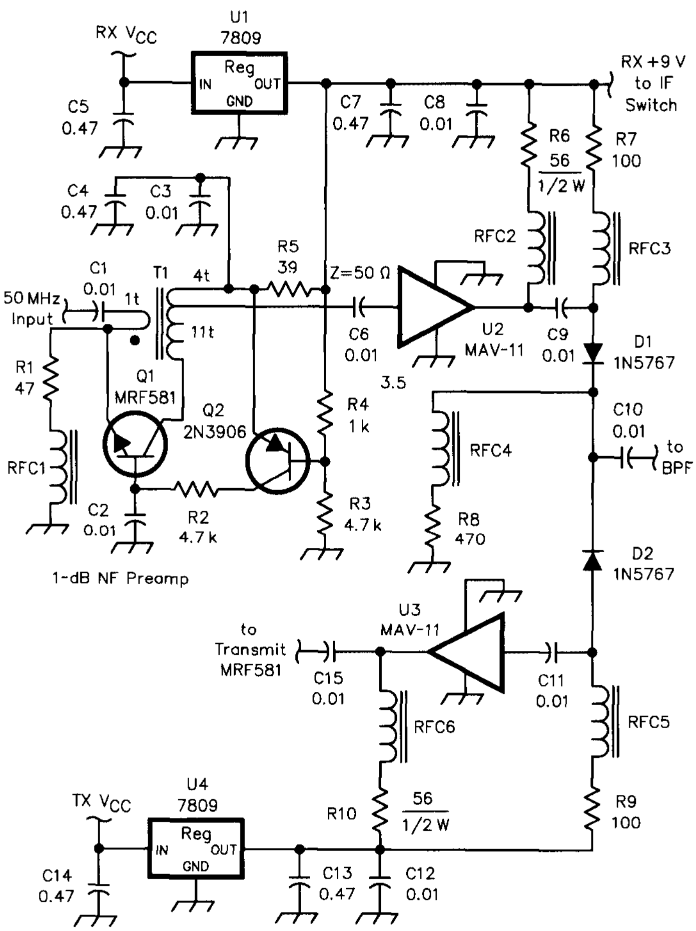


Fig 5—Small-signal 6-meter amplifiers.
D1, D2—1N5767 PIN diodes. Exact type not critical.
Q1—MRF581 low-noise transistor.
RFC1-6—10 turns no. 28 enam on FT-23-43 toroid core.
T1—15-turn primary on FT-37-67 toroid, with core material selected for low loss. Tap 11 turns from collector. 1-turn secondary. Use no. 28 enam. wire for both windings. *Phasing is critical for stable operation.*

ticularly if your reason for avoiding three-pole filters in the first place is the difficulty of tuning them up! The filter of Fig 4 is very low-loss, about 0.3 or 0.4 dB, but rather costly to construct, both in time and materials. It is quite similar to a 1.1-MHz-wide design I published previously.[2] Of course, if you have the appropriate test equipment, go ahead and use the 3-pole design.

I built the needed three filters using the design of Fig 4 and chose the best of them to put in front of the receiver preamplifier. At least for me, this made cascading filters a practical approach. This results in good immunity to interference from other transmitters. It would work even better at rejecting TV transmitter interference if the filter was just 1-MHz wide, but then it couldn't cover the entire 6-meter band. Of course, to maintain the low loss it's necessary to use even lower loss components. I'd consider a helical filter based on the design procedure in chapter 16 of the *ARRL Handbook*.[3] When tuning one of these up, it helps to understand the filter concepts discussed by Hayward.[4]

The preamplifier of Fig 5 is an optimized Norton design. This lossless feedback amplifier has been around a long time—so long that the patent has run out. Using type-67 ferrite to construct a low-loss transformer gives measured noise figures just under 1 dB with a gain of 10.5 dB. However, I found the output intercept at low bias current (under 10 mA) to be only +20 dBm. I was able to raise the output intercept to +31 dBm, while sacrificing 0.06 dB of noise figure, by raising the bias current to 16 mA. A variable resistor in series with R5 can be used as a bias-current adjustment. The use of an MRF 581 device also helps, as it performs a bit better than the MRF 586. (The latter device is no longer listed in the new Motorola data book—they are phasing out devices in that package.) Due to the low permeability (40) of type-67 material, the preamp isn't as broadband as one might expect. It will cover the 6-meter band just fine, but the noise figure degrades at 30 and 60 MHz.

There are several potential problems with this design. First, it is easy to convert into an oscillator. Switching the sense of the one-turn loop is enough to make it oscillate. Insufficient bypassing of the power supply also may result in oscillation. And the input-to-output isolation is poor. Thus, cascading a pair of these is ill-advised. On the other hand, you do get a very good output intercept and noise figure without any tweaking on expensive test equipment. I decided that using an MAV-11 as the second stage was an acceptable compromise, though someone building a high-performance system would use something more exotic. An MAV-11 typically has a 3.5-dB NF and 12 dB of gain at low frequencies.

Improved Local Oscillator Design

A complaint I've heard about the 222-MHz no-tune design I published in the July 1993 *QEX* is that there aren't enough tuning adjustments! People want to be able to tweak that crystal to a precise frequency. In the design shown in Fig 6, I was able to move the frequency upward with the following procedure. First, adjust L1 for parallel resonance with the holder capacitance. This can be done with a sweep setup that looks at the insertion loss of the device under test. At resonance, the shunting effect of the stray reactance is minimized, resulting in maximum attenuation. Next, try different capacitors for C4. The higher the capacitance, the lower the frequency. A piston trimmer isn't needed for C1, but it's an easy way to adjust the circuit inside a shielded box. It also serves as a rugged tie point for ground-plane construction.

I originally planned to use a Mini-Circuits TAK-1H high-level mixer in this design (Fig 6) but had difficulty getting the expected input intercept when integrating the system. I did get the expected performance from an SBL-1 mixer. A low-pass filter between the LO and mixer is needed for adequate transmit spectral purity. Otherwise the second harmonic of the local oscillator would mix with the RF signal and generate a large spurious signal around 44 MHz. Adding the filter dropped the spur from −11 to −38 dB, relative to the desired signal. Unfortunately, the SBL-1 is the weak link of the transverter since its input intercept is around +15 dBm while the output intercept of the MAV-11 is +30 dBm. I didn't worry too much about that since the weak link of the *system* is the 2-meter IF radio. It's difficult to find even HF transceivers with low noise figures and input intercepts in excess of +20 dBm, never mind getting it out of a 2-meter multimode radio.

The post-mixer IF amplifier of Fig 7 is another MAV-11, which is switched out of the path during transmit.

T/R switching

T/R switching is done with switch-

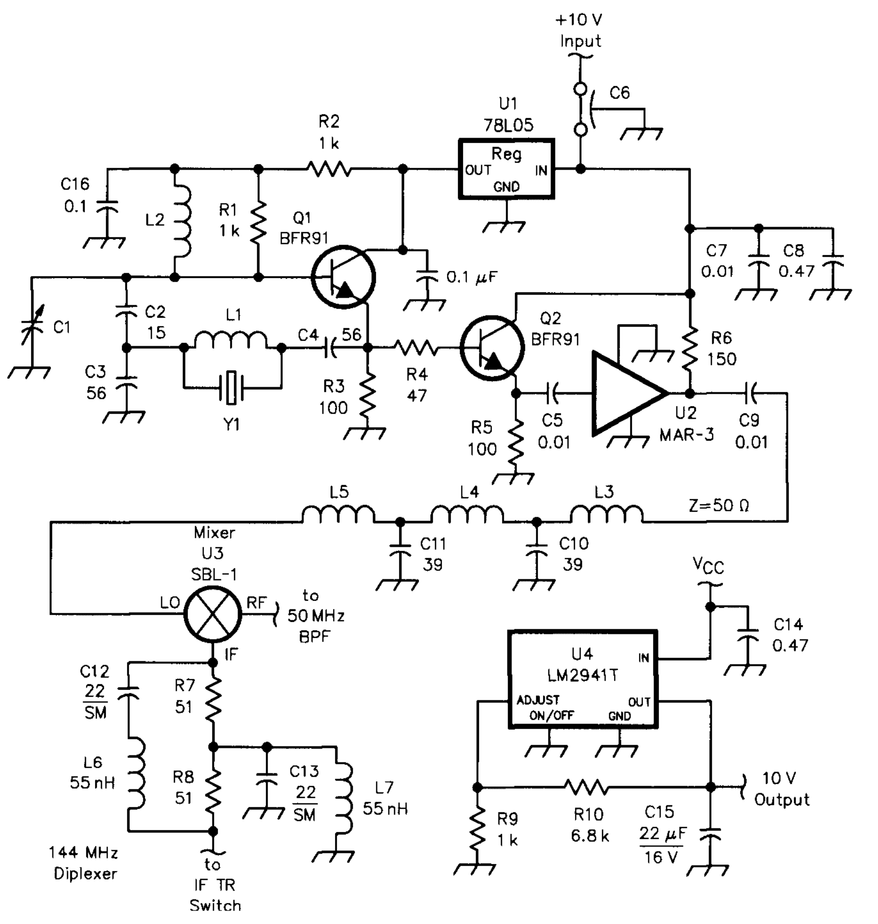


Fig 6—The mixer and 94-MHz local oscillator.

C1—Johanson 5459 1 to 14-pF piston trimmer. (See text.)
C4—Select for desired frequency. (See text.)
C6—Feedthrough capacitor. Value not critical. 100 pF to 0.01 µF works fine.
C15—22-µF tantalum capacitor, 16 V. *Absolutely required* for stable regulator operation. May be replaced with 100-µF electrolytic.
L1—Approximately 15 turns no. 28 enam on T-25-6 core. Inductance resonates with holder capacitance of crystal. (See text.)
L2—0.13 µH, 9 turns no. 28 close wound. Length is 0.14 inches. 0.10-inch inside diameter.
L3, L5—57 nH, 4 turns no. 26 enam space wound so that length is 0.25 inches. 0.19-inch inside diameter.
L4—0.12 µH, 7 turns no. 26 enam close wound. Length is 0.375 inches. 0.19-inch inside diameter.
L6, L7—55 nH, 5 turns no. 24 enam close wound. 0.125-inch inside diameter.
U4—LM2941T low-dropout regulator.
Y1—94-MHz, 5th-overtone crystal. International Crystal Manufacturing 473390.

ing diodes, PIN diodes and a relay. The switching diodes are used in the transmit signal path. While PIN diodes do offer a bit better performance than switching diodes, who needs high performance in a lossy transmit path where the idea is to dump excessive transmit power? The Omron RF relay is used for the high-power switching and seems to work well at the 10-W level, particularly if you use a sequencer, like the one by Chip Angle, N6CA, on page 22.57 of the 1995 *ARRL Handbook*. (Sequencer boards are available from FAR Circuits.[5]) I added a few switching transistors so that the sequencer switches the +12-V supplies. (See Fig 8.)

I modified my IF radios so they put a +5 to +12-V signal on the antenna line during transmit. The sequencer senses this voltage and switches the voltages appropriately. For the input attenuator, 1- and 2-W metal-oxide resistors work just fine at 144 MHz, so it isn't necessary to track down carbon composition resistors. The latter do make a better overload indicator, though. (You can smell 'em.) The attenuator and switching circuitry of Fig 7 is needed because the radios put out 2 W of RF. Much of this circuit can be omitted if a milliwatt-level IF radio is available. I use red LEDs to indicate transmit and green LEDs to indicate receive as a matter of personal preference.

The M57735 power module (Fig 9) is the easiest way to generate 10 W on this band. Tests show that only +15 dBm of input is needed to get 10 W out. This input level is easily provided with the +20-dBm MRF 581 design published in the March 1994 issue of *QEX*. Normally, people mount the power module directly to a heat sink. Instead, I mounted the module to a

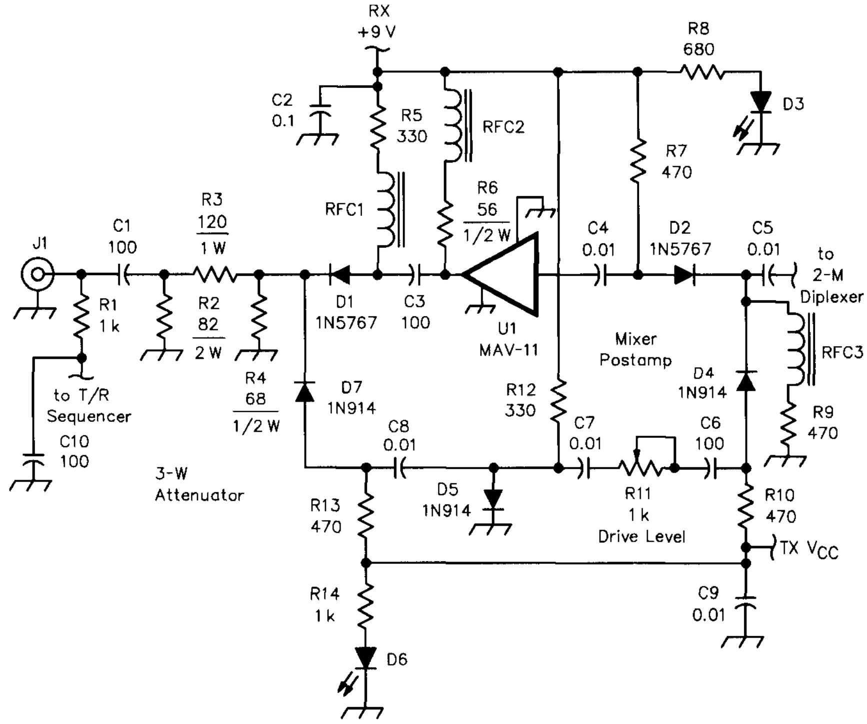
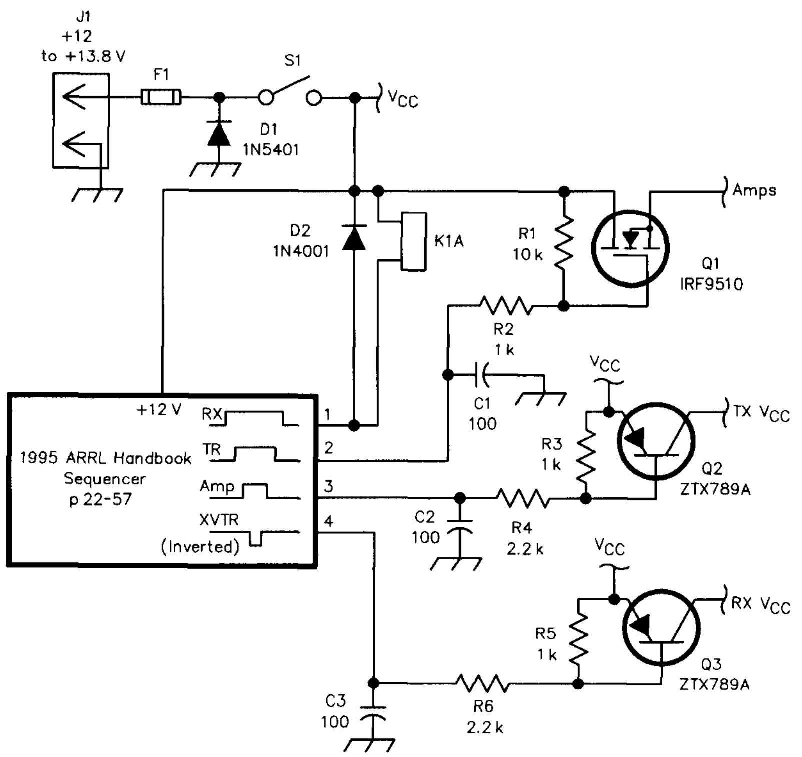


Fig 7—2-meter IF circuitry.
D1, D2—1N5767 PIN diode. Exact type not important.
D3—Green LED.
D6—Red LED.
J1—UG-290A/U BNC Jack.
RFC1-3—10 turns no. 28 enam on FT-23-43 toroid core.

Fig 8—Sequencer circuitry.
F1—5-A fuse used in prototype. 3-A fuse may be more appropriate.
J1—Panel-mounted ARRL recommended 12-V power connector (MOLEX 1545), as specified on page 22.6 of the *ARRL Handbook*. Female housing with male pins; + line is indexed (Radio Shack part RS 274-222).
K1A— Omron G5Y-1-DC12 relay (Digi-Key part Z24-ND).
Q1—IRF9510 P-channel HEXFET.
Q2, Q3—3-A plastic cased PNP transistor (very high current) (Digi-Key part ZTX-789A-ND).
S1—25-A, 12-V dc power switch (Radio Shack part RS 275-708).



2 × 3 × 1/4-inch piece of 6061-T6 aluminum bar stock and attached the metal block to a heat sink. This is similar to the approach that I've seen used in some microwave power amplifiers. The sheet stock is much easier to machine than most heat sink extrusions, yet thick enough to not bend easily.

SWR Coupler

This circuit board design has an unusually high directivity for a microstrip design using 1/16-inch FR-4 circuit board. Two samples showed directivities of 26 and 31 dB. I've talked with several people who found it exceedingly difficult to come up with a similar design and gave up. I found it necessary to optimize the design by hand with the assistance of *Microwave Harmonica*, an expensive simulation program sold by Compact Software. This is the big brother to *ARRL Radio Designer* that has the sophisticated microstrip models that engineers spend big money to get. (Don't bother asking about a 2-meter design—I'm still working on it.)

The coupler shown in Fig 10 is quite simple—it's just two 135-mil-wide microstrips with a spacing of 20 mils, along with some right-angle transitions to get signals in and out of the coupler. The length of the closely coupled lines is 3000 mils. The expected dielectric thickness is 56 mils—this is perhaps the most variable aspect of working with glass epoxy board. A significantly different thickness will upset the impedance of the lines. The dielectric constant is 4.8. I wouldn't worry about this too much, unless you use a different type of board material. I got better directivity by drilling holes through the board and directly connecting the center conductors of small coax to the traces and the shields to the ground plane than by trying to make end-launch BNC connections.

The insertion loss of the coupler is too low to measure easily, perhaps 0.05 dB. For best accuracy, it may be worthwhile to place another low-pass filter between the transmit amplifier and the coupler since the coupler will respond to harmonics. Omitting the filter between the coupler and the antenna will increase the likelihood of interference to the SWR measurement from other transmitters as the broadband diode detector has little selectivity.

A single coupler can measure both forward and reverse power. People have encountered problems with

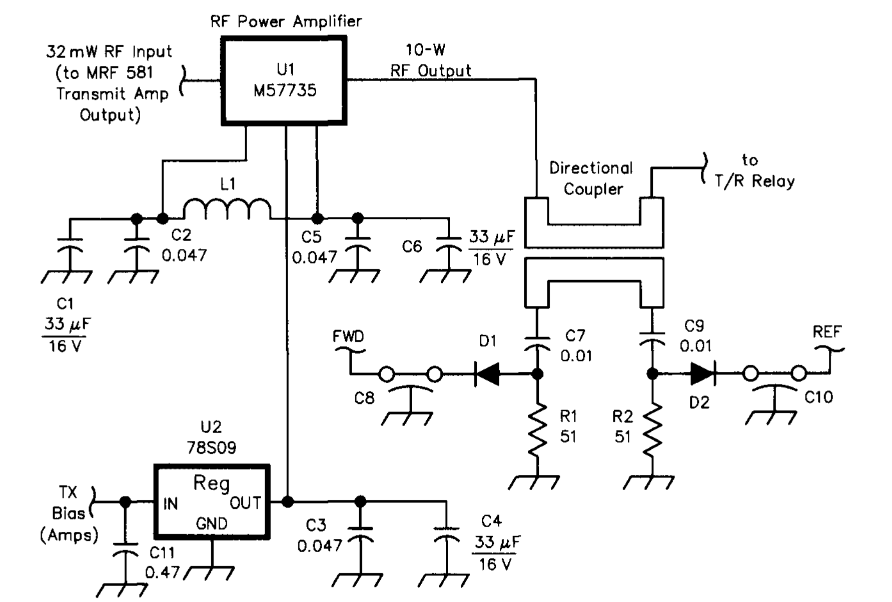


Fig 9—The 10-W power amplifier and directional coupler.

C8, C10—100 pF to 0.01 µF feedthrough capacitors.
D1, D2—Detector diodes. 1N34As may work best, though Shottky types will also work.
L1—7 turns no. 22 enam closewound. 0.19-inch inside diameter. Exact inductance shouldn't be critical.
U1—Mitsubishi M57735 RF power module.
U2—78S09 9-V regulator.

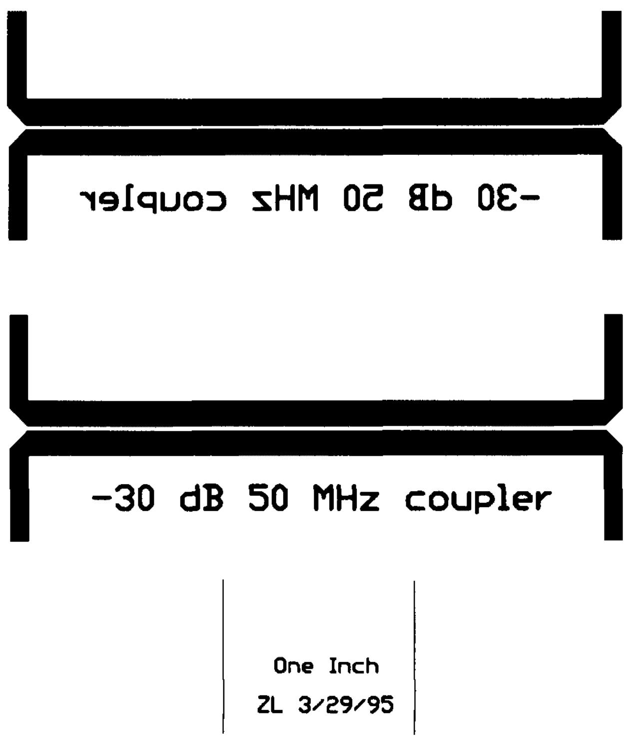


Fig 10—Etching pattern for a directional coupler on 1/16-inch FR-4 circuit board.

detector diodes generating dc bias, upsetting the reverse-power measurement, leading them to use dual couplers. Instead, I used dc blocking capacitors. The coupling is –30 dB. Thus, 10 W through the detector, assuming 50-Ω terminations, results in 10 mW of forward power and 10 μW of unwanted reflected power. Ideally, there would be no reflected power. Chapter 22 of the 1995 *ARRL Handbook* has a good low-power wattmeter circuit that could take advantage of the high directivity of this coupler. I was able to buy new 50 μA meters—which are getting rare—from Ocean State Electronics.[6]

For the most part, the components used in these circuits are pretty easily found from Ocean State, your local Radio Shack or Digi-Key.[7]

If you'd like copies of the proceedings papers mentioned here, contact the ARRL Technical Department Secretary at ARRL HQ. Copies are $3 per article.

Notes

[1]Rector, R., WA4NJP, "6 Meter EME," *Proceedings of the Central States VHF Society*, 1988, ARRL.

[2]Lau, Z., KH6CP, "A Collection of VHF Filters," *Proceedings of the 20th Eastern VHF/UHF Conference*, 1994, ARRL.

[3]Schetgen, R., KU7G, ed, *The ARRL Handbook for Radio Amateurs*, Newington: 1995, ARRL.

[4]Hayward, W., W7ZOI, "The Double-Tuned Circuit: An Experimenter's Tutorial," *QST*, Dec 1991, pp 29-34.

[5]FAR Circuits, 18N640 Field Ct, Dundee, IL 60118-9269, fax/voice-mail: 708-836-9148.

[6]Ocean State Electronics, PO Box 1458, 6 Industrial Drive, Westerly, RI 02891, tel: 401-596-3080, 800-866-6626.

[7]Digi-Key Corporation, 701 Brooks Ave South, PO Box 677, Thief River Falls, MN 56701-0677, tel: 800-344-4539, fax: 218-681-3380.