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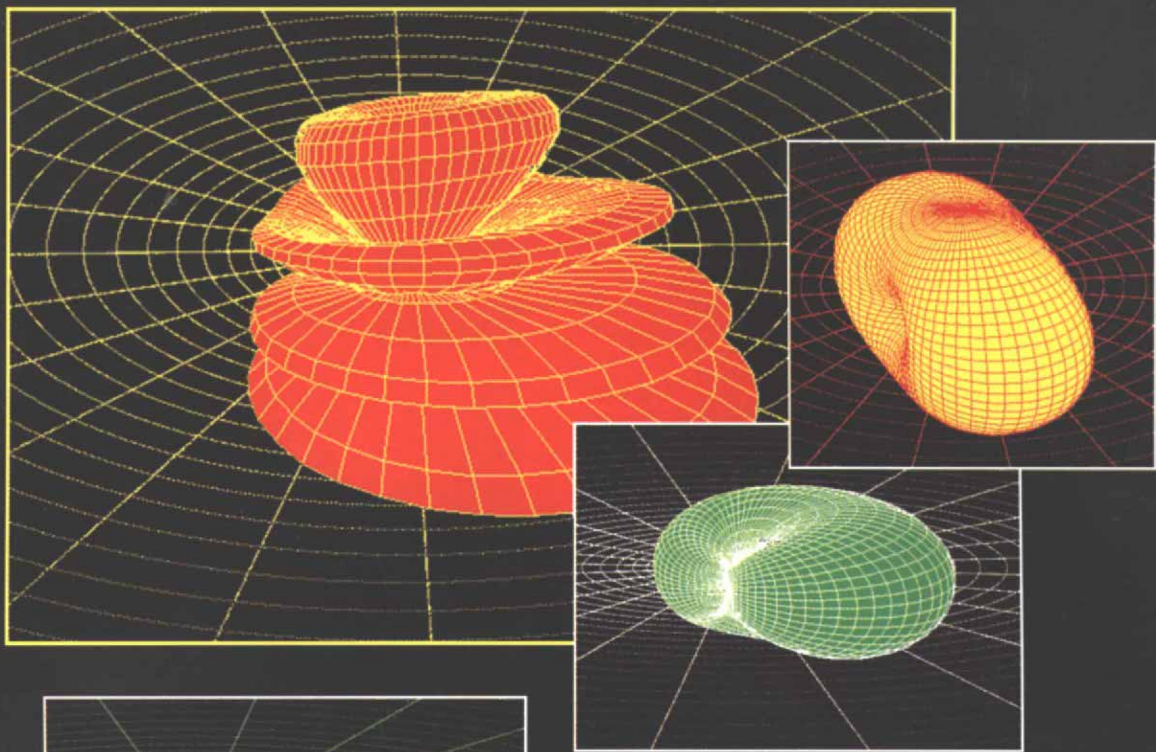
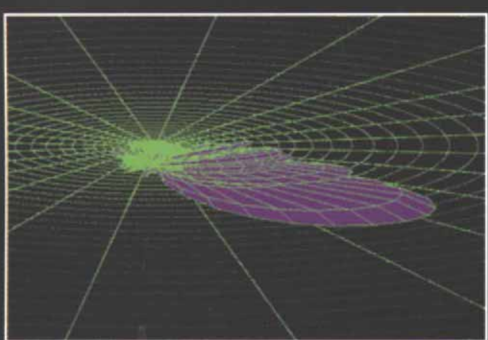
July/August 1999

Forum for Communications Experimenters





KF6DX Explains How to Render Three-Dimensional Radiation Patterns for Modeled Antennas on Your Computer Screen

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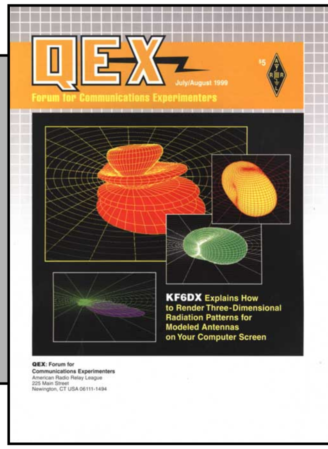

About the Cover

Editor Doug Smith describes how he plots radiation patterns with *QuickBasic* 4.5.

Features

Columns

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Empirically Speaking

Let's consider reducing interference on our crowded bands as a top technological goal. PSK31 sets a good example, since it shrinks occupied bandwidth to the scale of the information rate. Attempts have been made to achieve similar bandwidth reduction for phone modes, but not much has revolutionized voice modulation since SSB became popular. In the early '80s, amplitude compan-dored single sideband (ACSSB) was hailed as an efficient replacement for narrow-band FM (NBFM) systems, especially in the Land Mobile Service. Today, it is little more than a footnote in discussions of modulation formats. It wasn't compatible with contemporary systems, and suffered from a problem its NBFM cousin largely avoided—co-channel interference. Further, ACSSB didn't really bring substantial new technology to the table.

DSP and new algorithms now make *frequency companding* practical. This technique shrinks transmitted bandwidth by compressing the range of input frequencies, then expands them when received. The concept has now been reduced to practice, and before long, I hope to show how it's done. It is compatible with existing AM, FM and SSB phone transceivers in the form of an external audio processor, and it effectively removes the restriction placed by transceiver bandwidths on audio fidelity and frequency response. It allows greater spectral occupancy while maintaining excellent audio quality, but is more susceptible to selective fading and frequency errors than some other modes.

In addition to this angle, I see some of us are ready to tackle the challenge of an affordable, extremely high-dynamic-range receiver for HF. Gird your loins, because the obstacles are manifold and the stakes are high. The reward, however, may well be the continued usefulness of our cherished resources.

In This *QEX*

PSK31 has inspired fresh enthusiasm in the Amateur Radio-teletype world, especially among us rag-chewers. It won't be long before contesters are motivated to exploit the mode. It will be fascinating to hear and see great combs of warbling signals across the HF RTTY segments! We bring you the complete text of Peter Martinez, G3PLX's ground-breaking article. Special thanks to our fellow IARU members at the RSGB for their kind assistance.

Ray Mack, WD5IFS, has come through with a simple forward converter for those who want to get started with switched-mode power supplies. This unit might make a good companion to that Class-E or envelope elimination and restoral (EER) system you've been thinking about.

Grant Bingeman, KM5KG, has some observations about center-loaded whip losses. What is the radiation cost of those slim whips with miniscule center-loading coils?

Bill Sabin, W0IYH, adds another section to his solid-state amplifier design with some diplexer LPFs. Bill is after the best possible linearity, and he explains why this approach is necessary to achieve that end. Ed Wetherhold, W3NQN, has done a lot more work with elliptical filters, with an eye to minimizing interference between stations operating on adjacent bands, such as in a "multi-multi" contest environment.

Bob Zavrel, Jr, W7SX, contributes his further efforts on the extended double Zepp and its variations. The $\lambda/4$-stub linear end loading and trap technique is retained. Modeling with *EZNEC* 2.0 reveals the radiation-pattern effects of Bob's design mutations. This program offers 3-D pattern plotting. Some work at KF6DX has been done on drawing 3-D radiation patterns, and I give you those generalized results, along with a sample program in *BASIC*.

Josef Maier, OE3JIS, is ready for P3D, and has some pertinent suggestions for those of you who aren't, but want to be. Thanks to *AMSAT-NA Journal* and Russ Tillman, KC5JVB, for reprint permission. Al Williams, VE6AXW, gives us his design of a regulated high-voltage supply, and explains how it will improve your transmitter's envelope. Parker Cope, W2GOM, returns with some notes on current regulators along with some circuit examples. RF presents a 70-cm power divider. Keep those projects going!—73, *Doug Smith, KF6DX*, **kf6dx@arrl.org**.

PSK31: A New Radio-Teletype Mode

Many error-correcting data modes are well suited to file transfers, yet most hams still prefer error-prone Baudot for everyday chats. PSK31 should fix that. It requires very little spectrum and borrows some characteristics from Morse code. Equipment? Free software, an HF transceiver and a PC with Windows *and a sound card will get you on the air.*

By Peter Martinez, G3PLX

[Thanks to the Radio Society of Great Britain for permission to reprint this article. It originally appeared in the December '98 and January '99 issues of their journal, RadCom. *This article includes February 1999 updates from Peter Martinez.*—Ed.]

I've been active on RTTY since the 1960s, and was instrumental in introducing AMTOR to Amateur Radio at the end of the '70s. This improved the reliability of the HF radio link and paved the way to further developments that have taken this side of the hobby more into data transfer, message handling and computer linking, but further away from the rest of Amateur Radio, which is based on two-way contacts between operators.

There is now a gap opening between the data-transfer enthusiasts using the latest techniques and the two-way contact fans who are still using the traditional RTTY mode of the '60s, although of course using keyboard and screen rather than teleprinter. There is scope for applying the new techniques now available to bring RTTY into the 21st century.

This article discusses the specific needs of "live QSO" operating—as opposed to just transferring chunks of error-free data—and describes the PSK31 mode I have developed specifically for live contacts. PSK31 is now becoming popular using low-cost DSP kits. The mode could become even cheaper as the art of using PC sound cards is developed by Amateur Radio enthusiasts.

What is Needed?

I believe that it is the error-correcting process used in modern data modes make them unsuitable for live contacts. I have identified several factors; the first revolves around the fact that all error-correcting systems introduce a time-delay into the link. In the case of an ARQ link like AMTOR or PACTOR, there is a fixed transmission cycle of 450 ms or 1.25 s or more. This delays any key press by as much as one cycle period, and by more if there are errors. With forward-error-correction systems, there is also an inevitable delay, because the information is spread over time. In a live two-way contact, the delay is doubled at the point where the transmission is handed over. I believe that these delays make such systems unpleasant to use in a two-way conversation.

This is not so much a technical problem as a human one. Another factor in this category concerns the way that quality of information content varies as the quality of the radio link varies. In an analogue transmission system such as SSB or CW, there is a linear relationship between the two. The operators are aware of this all the time and take account of it subconsciously: They change the speed and tone of voice instinctively and even choose the conversation topic to suit the conditions. In a digital mode, the relationship between the signal-to-noise ratio (S/N) on the air and the error-rate on

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England

Table 1

The Varicode alphabet. The codes are transmitted left bit first, with “0” representing a phase reversal on BPSK and “1” representing a steady carrier. A minimum of two zeros is inserted between characters. Some implementations may not handle all the codes below 32.

*ASCII**	*Varicode*	*ASCII**	*Varicode*	*ASCII**	*Varicode*
0 (NUL)	1010101011	+	111011111	V	110110101
1 (SOH)	1011011011	,	1110101	W	101011101
2 (STX)	1011101101	-	110101	X	101110101
3 (ETX)	1101110111	.	1010111	Y	101111011
4 (EOT)	1011101011	/	110101111	Z	1010101101
5 (ENQ)	1101011111	0	10110111	[	111110111
6 (ACK)	1011101111	1	10111101	\	111101111
7 (BEL)	1011111101	2	11101101	]	111111011
8 (BS)	1011111111	3	11111111	^	1010111111
9 (HT)	11101111	4	101110111	_	101101101
10 (LF)	11101	5	101011011	‘	1011011111
11 (VT)	1101101111	6	101101011	a	1011
12 (FF)	1011011101	7	110101101	b	1011111
13 (CR)	11111	8	110101011	c	101111
14 (SO)	1101110101	9	110110111	d	101101
15 (SI)	1110101011	:	11110101	e	11
16 (DLE)	1011110111	;	110111101	f	111101
17 (DC1)	1011110101	<	111101101	g	1011011
18 (DC2)	1110101101	=	1010101	h	101011
19 (DC3)	1110101111	>	111010111	i	1101
20 (DC4)	1101011011	?	1010101111	j	111101011
21 (NAK)	1101101011	@	1010111101	k	10111111
22 (SYN)	1101101101	A	1111101	l	11011
23 (ETB)	1101010111	B	11101011	m	111011
24 (CAN)	1101111011	C	10101101	n	1111
25 (EM)	1101111101	D	10110101	o	111
26 (SUB)	1110110111	E	1110111	p	111111
27 (ESC)	1101010101	F	11011011	q	110111111
28 (FS)	1101011101	G	11111101	r	10101
29 (GS)	1110111011	H	101010101	s	10111
30 (RS)	1011111011	I	1111111	t	101
31 (US)	1101111111	J	111111101	u	110111
32 (SP)	1	K	101111101	v	1111011
!	111111111	L	11010111	w	1101011
“	101011111	M	10111011	x	11011111
#	111110101	N	11011101	y	1011101
$	111011011	O	10101011	z	111010101
%	1011010101	P	11010101	{	1010110111
&	1010111011	Q	111011101	|	110111011
‘	101111111	R	10101111	}	1010110101
(	11111011	S	1101111	~	1011010111
)	11110111	T	1101101	127	1110110101
*	101101111	U	101010111		

*ASCII characters 0 through 31 are control codes. Their abbreviations are shown here in parentheses. For the meanings of the abbreviations, refer to any recent *ARRL Handbook.*

the screen is not so smooth. The modern error-correcting digital modes are particularly bad at this, with copy being almost perfect while the SNR is above a certain level and stopping completely when the SNR drops below this level. The effect is of no consequence in an automatic mailbox-forwarding link, but can badly inhibit the flow of a conversation.

A third factor is a social one: with error-correcting modes, you only get good copy when you are linked to one other station. The copy is decidedly worse when stations are not linked, such as when calling CQ or listening to others. This makes it difficult to meet other people on the air, and there is a tendency to limit contacts to a few close friends or just mailboxes.

These factors lead me to suggest that there is a case for a transmission system that is *not* based on the use of error-correcting codes, when the specific application is that of live contacts. The continued popularity of traditional RTTY using the start-stop system is proof of this hypothesis: There is minimal delay (150 mS), the flow of conversation is continuous, the error-rate is tolerable, and it is easy to listen-in and join-in.

Improving on RTTY

How, then, do we go about using modern techniques that were not available in the '60s, to improve on traditional RTTY? First, since we are talking about live contacts, there is no need to discuss any system that transmits text any faster than can be typed by hand. Second, modern transceivers are far more frequency stable than those of the '60s. We should be able to use much narrower bandwidths than in those days. Third, digital processors are much more powerful than the rotating cams and levers of mechanical teleprinters, so we could use better coding. The drift-tolerant technique of frequency-shift keying, and the fixed-length five-unit start-stop code still used today for RTTY are a legacy of 30-year-old technology limits. We can do better now.

PSK31 Alphabet

The method I have devised for using modern digital processing to improve on the start-stop code, without introducing extra delays due to the error-correcting or synchronization processes, is based firmly on another tradition, namely that of Morse code. Because Morse uses short codes for the more common letters, it is actually very efficient in terms of the average duration of a character. In addition, if we think of it in terms that we normally use for digital modes, Morse code is self-synchronizing: We don't need to use a separate process to tell us where one character ends and the next begins. This means that Morse code doesn't suffer from the "error-cascade" problem that results in the start-stop method getting badly out of step if a start or stop-bit is corrupt. This is because the pattern used to code a gap between two characters *never* occurs inside a character.

The code I have devised is therefore a logical extension of Morse code, using not just one-bit or three-bit code-elements (dots and dashes), but any length. The letter-gap can also be shortened to two bits. If we represent key-up by 0 and key-down by 1, then the shortest code is a single one by itself. The next is 11, then 101 and 111, then 1011, 1101, 1111, but not 1001 since we must not have two or more consecutive zeros inside a code. A few minutes with pencil and paper will generate more. We can do the 128-character ASCII set with 10 bits.

I analyzed lots of English-language text to find out how common was each of the ASCII characters, then allocated shorter codes to the more-common characters. The result is shown in Table 1, and I call it the *Varicode* alphabet. With English text, Varicode has an average code length—including the "00" letter gap—of 6.5 bits per character. By simulating random bit errors and counting the number of corrupted characters, I find that Varicode is 50% better than start-stop code, thus verifying that its self-synchronizing properties work well.

The shortest code in Morse is the most-common letter: "e", but in Varicode the shortest code is allocated to the word space. When idle, the transmitter sends a continuous string of zeros. Fig 1 compares the coding of the same word in ASCII, RTTY, Morse and Varicode.

PSK31 Modulation and Demodulation

To transmit Varicode at a reasonable typing speed of about 50 words per minute needs a bit-rate of about 32 per sec. I have chosen 31.25, because it can be easily derived from the 8-kHz sample-rate used in many DSP systems. In theory, we only need a bandwidth of 31.25 Hz to send this as binary data, and the frequency stability that this implies can be achieved with modern radio equipment on HF.

The method chosen was first used on the amateur bands, to my knowledge, by SP9VRC. Instead of frequency-shifting the carrier, which is wasteful of spectrum, or turning the carrier on and off, which is wasteful of transmitter power capability, the "dots" of the code are signaled by reversing the polarity of the carrier. You can think of this as equivalent to transposing the wires to your antenna feeder. This uses the transmitted signal more efficiently since we are comparing a positive signal before the reversal to a negative signal after it, rather than comparing the signal present in the dot to no-signal in the gap. But if we keyed the transmitter in this way at 31.25 baud, it would generate terrible key clicks, so we need to filter it.

If we take a string of dots in Morse code, and low-pass filter it to the theoretical minimum bandwidth, it will look the same as a carrier that is 100% amplitude-modulated by a sine wave at the dot rate. The spectrum is a central carrier and two sidebands at 6dB down on either side. A signal that is sending continuous reversals, filtered to the minimum bandwidth, is equivalent to a double-sideband suppressed-carrier emission, that is, to two tones either side of a suppressed carrier. The improvement in the performance of this polarity-reversal keying over on-off keying is thus equivalent to the textbook improvement in changing from amplitude-modulation telephony with full carrier to double-sideband with suppressed carrier. I have called

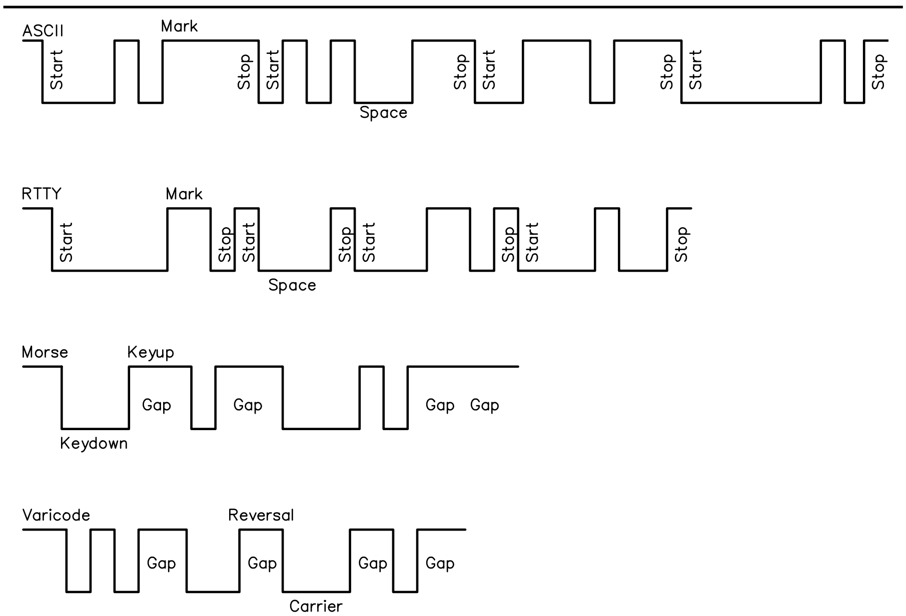

Fig 1—The word "ten" in ASCII, RTTY, Morse and Varicode.

this technique "polarity-reversal keying" so far, but everybody else calls it "binary phase-shift keying," or BPSK. Fig 2 shows the envelope of BPSK modulation and the detail of the polarity reversal.

To generate BPSK in its simplest form, we could convert our data stream to levels of ±1 V, for example, take it through a low-pass filter and feed it into a balanced modulator.

The other input to the balanced modulator is the desired carrier frequency. When sending continuous reversals, this looks like a 1 V (P-P) sine wave going into a DSB modulator, so the output is a pure two-tone signal. In practice we use a standard SSB transceiver and perform the modulation at audio frequencies or carry out the equivalent process in a DSP chip. We could signal logic zero by continuous carrier and signal logic one by a reversal, but I do it the other way round for reasons that will become clear shortly.

There are many ways to demodulate BPSK, but they all start with a bandpass filter. For the speed chosen for PSK31, this filter can be as narrow as 31.25 Hz in theory. A brick-wall filter of precisely this width would be costly, however, not only in monetary terms but also in the delay time through the filter, and we want to avoid delays. A practical filter might be twice that width (62.5 Hz) at the 60-dB-down points with a delay-time of two bits (64 ms).

For the demodulation itself, since BPSK is equivalent to double sideband, the textbook method for demodulating DSB can be used. However, it can also be demodulated by delaying the signal by one bit period and comparing it to the signal with no delay in a phase comparator. The output is negative when the signal reverses polarity and positive when it doesn't.

We could extract the information from the demodulated signal by measuring the lengths of the "dots" and "dashes," as we do by ear with Morse code. It helps to pick the data out of the noise, however, if we know when to expect signal changes. We can easily transmit the data at an accurately timed rate, so it should be possible to predict when to sample the demodulator output. This process is known as synchronous reception, although the term "coherent" is sometimes wrongly used.

To synchronize the receiver to the transmitter, we can use the fact that a BPSK signal has an amplitude-modulation component. Although the modulation varies with the data pattern, it always contains a pure-tone component at the baud rate. This can be extracted using a narrow filter, a PLL or the DSP equivalent, and fed to the decoder to sample the demodulated data. Fig 3 shows block diagrams of a typical BPSK modulator and demodulator.

For the synchronization to work we need to make sure that there are no long gaps in the pattern of reversals. A completely steady carrier has no modulation, so we could never predict when the next reversal was due. Fortunately, Varicode is just what we need, provided we choose the logic levels so that zero corresponds to a reversal and one to a steady carrier. The idle signal of continuous zeros thus generates continuous reversals, giving us a strong 31.25-Hz modulation. Even with continuous keying, there will always be two reversals in the gaps between characters. The average number of reversals will therefore be more than two in every 6.5 bits, and there will never be more than 12 bits with no reversal at all. If we make sure that the transmission always starts with an idle period, then the timing will pull into sync quickly. By making the transmitter end a transmission with a "tail" of unmodulated carrier, it is then possible to use the presence or absence of reversals to squelch the decoder. Hence, the screen doesn't fill with noise when there is no signal.

Getting Going

So much for the philosophy and the theory, but how do you get on the air with this mode? In the first experiments on this mode in early 1996, the route to getting on PSK31 was to obtain one of several DSP starter kits. These are printed-circuit cards, usually with a serial interface to a PC, marketed by DSP processor manufacturers at low cost to help engineers and students become familiar with DSP programming. Some radio amateurs have started to write software for these, not just for RTTY but also for SSTV, packet, satellite and digital-voice experiments. They have audio input and output and some general-purpose digital input/output. The construction work needed is limited to wiring up cables, building a power supply and putting the card into a screened box. The DSP software is freely available, as is the software that runs in the PC to interface to the keyboard and screen, and can be obtained most easily via the Internet. It would certainly be possible to construct a PSK31 modem in hardware, although I know of no one who has done this yet.

However, it became clear late in 1998 that soundcards now common in personal computers are capable of performing the audio input/output function needed for PSK31, with the DSP software running in the PC. At Christmas 1998, I completed a basic *Windows*-based PSK31 program to use the soundcard. The availability of this program has dramatically increased the level of PSK31 activity worldwide. (It's available on the Web: **http://aintel.bi.ehu.es/psk31.html**—*Ed.*)

PSK31 Operating

Since PSK31 performance is the same when calling, listening or in contact, it's easy to progress from listen-

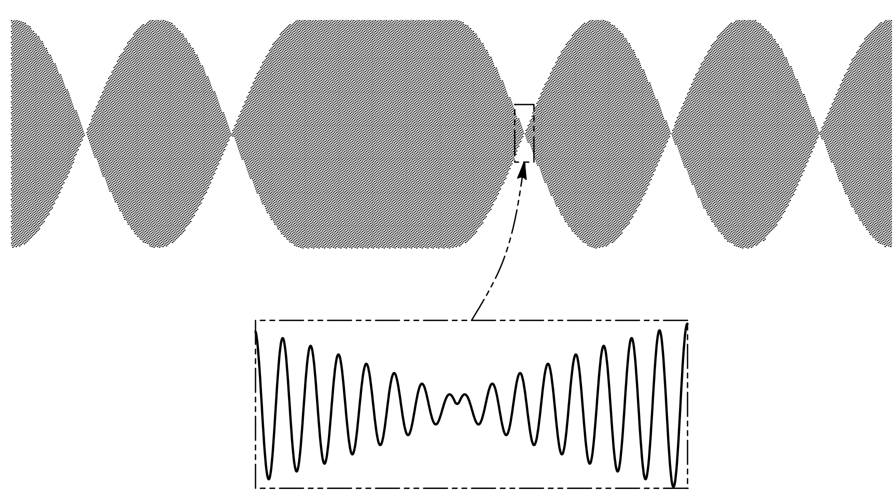

Fig 2—The waveform of BPSK sending the Varicode space symbol., with a close-up of the detail during a phase reversal.

ing to others, to calling CQ, two-way contacts and multi-station nets. The narrow bandwidth and good weak-signal performance do mean learning a few new tricks: First, set the radio dial on one spot. Then fine-tune the audio frequency, while listening through the narrow audio filter rather than the transceiver's loudspeaker, while using an on-screen phase-shift display to center the incoming signal within a few hertz. On transmit, since the envelope of the PSK31 signal is not constant (as is the case for FSK), it is important to keep the transmitter linear throughout. However, since the PSK31 idle is identical to a standard two-tone test signal, it is easy to set up. The worst distortion products will be at ±45 Hz at (typically) 36 dB below PEP.

So far, we've looked at requirements for a live-contact, keyboard/screen communication system, and proposed the narrow-band PSK31 mode as a candidate for a modern equivalent to traditional RTTY. This mode has now been in use on the HF bands by a small but growing band of enthusiasts for about two years. Now, let's look at two recent additions to PSK31.

A Second Look at Error Correction

After getting PSK31 going with BPSK modulation and the Varicode alphabet, several people urged me to add error correction to it in the belief that it would improve it still further. I resisted for the reasons that I gave earlier, namely that the delays in transmission, the discontinuous traffic flow and the inability to listen-in, all make error correction unattractive for live contacts. There is another reason. All error-correcting systems work by adding redundant data bits. Suppose I devise an error-correcting system that doubles the number of transmitted bits. If I wanted to maintain traffic throughput, I would need to double the bit rate. With BPSK that means doubling the bandwidth, so I lose 3dB of S/N and get more errors. The error correction system will have to work twice as hard just to break even! It is no longer obvious that error correction wins. It is interesting to note that with FSK, where the bandwidth is already much wider than the information content, you *can* double the bit-rate without doubling the bandwidth, and error correction *does* work. Computer simulation with BPSK in white noise shows that when the S/N is good, the error-correction system does win, reducing the low error rate to very low levels. At the S/N levels that are acceptable in live amateur contacts, it's better to transmit the raw data slowly in the narrowest bandwidth. It also takes up less spectrum space!

However, there was the suggestion that error correction could give useful results for bursts of noise, which cannot be simulated on the bench, so I decided to try it and do some comparison tests. The automatic repeat (ARQ) method of correcting errors was ruled out. Forward error correction (FEC) seemed to deserve a second look, provided the transmission delay was not too long.

I realized that comparing two systems with different bandwidths and speeds on the air would be difficult. Adjacent-channel interference would be different, as would the effects of multipath. There is, however, another way to double the information capacity of a BPSK channel without doubling its bandwidth and speed. By adding a second, 90° phase-shifted BPSK carrier at the transmitter and a second demodulator in the receiver, we can do the same trick that is used to transmit two color-difference signals in PAL and NTSC television. I call this quadrature polarity-reversal keying, but everybody else calls it quaternary phase-shift keying, QPSK.

There is a 3-dB S/N penalty with QPSK, because we must split the transmitter power equally between the two channels. This is the same penalty as doubling the bandwidth, so we are no worse off. QPSK is therefore ideal for my planned comparison experiment: The adjacent-channel interference, the S/N and the multipath performance would be the same for both.

In the next section, I will think of QPSK not as two channels of binary data, but as a single-channel that can be switched to any of four 90° phase-shift values. By the way, the clock-recovery idea used for BPSK works just as well for QPSK, because the envelope still has a modulation component at the bit-rate.

QPSK and the Convolutional Code

There is a vast amount of available knowledge about correcting errors in data that are organized in blocks of constant length (such as ASCII codes) by transmitting longer blocks. I know of nothing that covers error correction of variable-length blocks like Varicode. There are ways of reducing errors in continuous streams of data with no block structure. (This seems a natural choice for a radio link, since its errors don't have any block structure either.) These are called convolutional codes. One of the simplest forms does actually double the number of data bits; it is therefore a natural choice for a QPSK channel carrying a variable-length code.

The convolutional encoder generates one of the four phase shifts, not from each data bit to be sent, but from a sequence of them. This means that each bit is effectively spread out in time, intertwined with earlier and later bits in a precise way. The more we spread it out, the better will be the ability of the code to correct bursts of noise, but we must not go too far or we will introduce too much transmission delay. I chose a time spread of five bits. The table that determines the phase shift for each pattern of five successive bits is given in the sidebar "The Convolutional Code." The logic behind this table is beyond the scope of this article.

In the receiver, a device called a

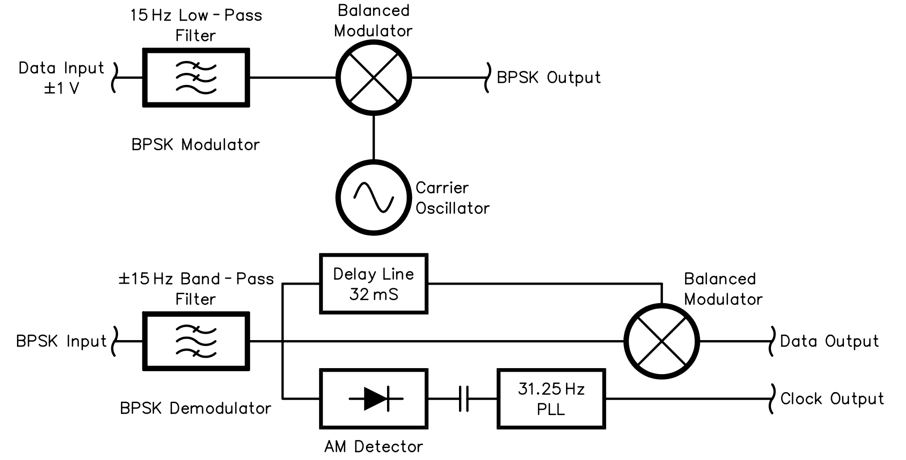

Fig 3—Block diagram of analog BPSK modem.

Viterbi decoder is used. This is not so much a decoder as a whole family of encoders playing a guessing game. Each one makes a different "guess" at what the last five transmitted data bits might have been. There are 32 different patterns of five bits and thus 32 encoders. At each step the phase-shift value predicted by the bit-pattern-guess from each encoder is compared with the actual received phase-shift value, and the 32 encoders are given "marks out of ten" for accuracy. Just as in a knockout competition, the worst 16 are eliminated and the best 16 go on to the next round, taking their previous scores with them. Each surviving encoder then gives birth to two "children," one guessing that the next transmitted bit will be a zero and the other guessing that it will be a one. They all do their encoding to guess what the next phase shift will be and receive scores again, which are added on to their earlier scores. The worst 16 encoders are killed-off again and the cycle repeats.

It's a bit like Darwin's theory of evolution, and eventually all the descendants of the encoders that made the right guesses earlier will be among the survivors and will all carry the same "ancestral genes." If we record the family tree (the bit-guess sequence) of each survivor, we can trace it back to find the transmitted bit-stream. We must wait at least five generations (bit periods), however, before all survivors have the same great great grandmother (who guessed right five bits ago). The whole point is that the scoring system based on the running total ensures that the decoder always gives the most-accurate guess, *even if the received pattern is corrupted.* Although we may need to wait a bit longer than five bit periods for the best answer to become clear. In other words, the Viterbi decoder corrects errors.

The longer we wait, the more accurate it is. I chose a decoder delay of four times the time spread, or 20 bits. We now have a 25-bit delay from one end to the other (800 ms), giving a round-trip delay to a two-way contact of 1.6 seconds. I think this is about the limit before it becomes a nuisance. In any case, the decoder could change to trade performance for delay without incompatibility.

QPSK on the Air

PSK31 operators find QPSK can be very good, but it is sometimes disappointing. In bench tests with white noise, it is actually *worse* than BPSK, confirming the simulation work mentioned earlier, but in conditions of burst noise, improvements of up to five times the character error-rate have been recorded. This performance doesn't come free, however. Apart from the transmission delay, which can be a bit annoying, QPSK is twice as critical in tuning as BPSK. A QPSK signal will start to decode wrong when the phase shift is greater than 45°, and that will be the case when the tuning error is only 3.9 Hz. This could be a problem with some older radios. What tends to happen is that contacts start on BPSK and change to QPSK if it is worth doing and there is no drift. There is one aspect of QPSK that must be kept in mind—it is important for both stations to use the correct sideband—on BPSK it doesn't matter.

Extending the Alphabet

In English-speaking countries, virtually all the characters and symbols that are needed for day-to-day written communications are present in the 128-character ASCII set. However, many other languages have accents, umlauts, tildes and other signs and symbols that are not in the ASCII set,

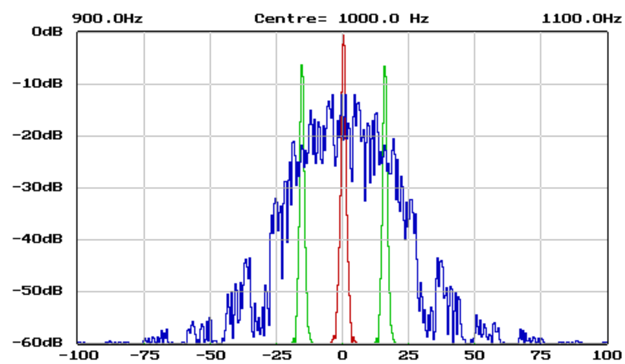

Fig 4—The spectrum of a BPSK signal, idling and sending data, compared with an unmodulated carrier at the same signal level. The carrier is the center pip; the smaller pips are the PSK31 reversals, and the large, ragged hump is noise shaped by the filter.

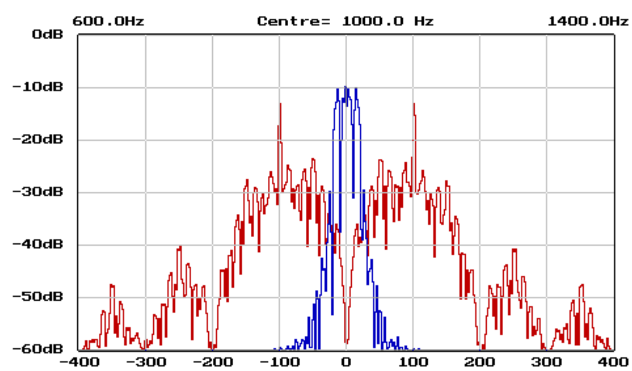

Fig 5—Comparison of the PSK31 spectrum with 100-baud, 200-Hz-shift FSK (AMTOR/PACTOR). The taller, three-hump signal at center is PSK31. The smaller, double-peak (±100 Hz) signal is FSK.

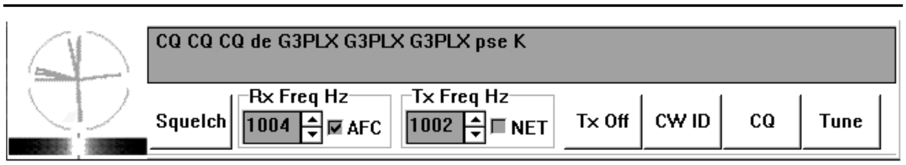

Fig 6—A screenshot of the *PSK31 Windows* program control panel, receiving a slightly noisy QPSK signal (notice the scope display at left). Fine-tuning controls for receive and transmit audio tones are near the bottom-center of the panel.

Is PSK31 Legal?

Some armchair lawyers have questioned the legality of PSK31 since its Varicode is not specifically mentioned as "legal" digital code in Part 97. Some confusion is understandable, give the wording of 97.309(a). However, the FCC clarified the meaning of the rules in an Order released October 11, 1995 (December 1995 *QST*, p 84). The Order (DA95-2106) reads in part: "This Order amends Section 97.304(a) of the Commission's Rules …to clarify that amateur stations may use any digital code that has its technical characteristics publicly documented. This action was initiated by a letter from the American Radio Relay League, Inc. (ARRL)."

The Order goes on to note that "The technical characteristics of CLOVER, G-TOR and PACTOR have been documented publicly for use by amateur operators, and commercial products are readily available that facilitate the transmission and reception of communications incorporating these codes. Including CLOVER, G-TOR and PACTOR in the rules will not conflict with our objective of preventing the use of codes or ciphers intended to obscure the meaning of the communication. We agree, therefore, that it would be helpful to the amateur service community for the rules to specifically authorize amateur stations to transmit messages and data using these and similar digital codes"

Given that PSK31 is in the public domain for amateur use, that software is readily and freely available and that its emission characteristics clearly meet the standards of Section 97.307 for RTTY/data, there is little doubt that its use by FCC-licensed amateur stations is legal.

However, just to complete the documentation, in a letter to the FCC dated January 27, 1999, ARRL General Counsel Christopher D. Imlay, W3KD, documented the technical characteristics of PSK31 in a manner similar to how CLOVER, G-TOR and PACTOR were previously documented. There is no need for PSK31 to be mentioned specifically in the rules, because CLOVER, G-TOR and PACTOR are simply given as examples.—*Dave Sumner, K1ZZ*

but are now used in everyday written text generated on computers. These extra symbols are now standardized worldwide in the ANSI alphabet, the first 128 characters of which are identical to ASCII, and the second 128 contain all the special symbols. Since the *WINDOWS* operating system uses ANSI, and most PC programs are now written for *WINDOWS*, I have recently extended the PSK31 alphabet in a *WINDOWS* version.

It is very easy to add extra characters to the Varicode alphabet without backwards-compatibility problems. In the early PSK31 decoders, if there was no "00" pattern received 10-bits after the last "00", it would simply be ignored as a corruption. In the extended alphabet, I let the transmitter legally send codes longer than 10 bits. The old decoders will just ignore them and the extended decoder can interpret them as extra characters. To get another 128 Varicodes means adding more 10-bit codes, all 11-bit and some 12-bit codes. There seemed little reason to be clever with shorter common characters so I chose to allocate them in numerical order, with code number 128 being 1110111101 and code number 255 being 101101011011. The vast majority of these will never be used, so it hardly slows the transmission rate at all, but it would not be a good idea to transmit binary files this way!

The Convolutional Code

The left-most numbers in each column contain the 32 combinations of a run of five Varicode bits, transmitted left bit first. The right-most number is the corresponding phase shift to be applied to the carrier, with "0" meaning no shift, "1" meaning advance by 90°, "2" meaning polarity reversal and "3" meaning retard by 90°. A signal that is advancing in phase continuously is higher in radio frequency than the carrier.

00000	2	01000	0	10000	1	11000	3
00001	1	01001	3	10001	2	11001	0
00010	3	01010	1	10010	0	11010	2
00011	0	01011	2	10011	3	11011	1
00100	3	01100	1	10100	0	11100	2
00101	0	01101	2	1010	3	11101	1
00110	2	01110	0	10110	1	11110	3
00111	1	01111	3	10111	2	11111	0

As an example, the "space" symbol, a single 1 preceded and followed by zeros, would be represented by successive run-of-five groups 00000, 00001, 00010, 00100, 01000, 10000, 00000, which results in the transmitter sending the QPSK pattern 2,1,3,3,0,1,2.

Note that a continuous sequence of zeros (the Varicode idle sequence) gives continuous reversals, the same as BPSK.

Summary

This article has identified some of the characteristics of modern HF data-transmission modes that have contributed to the decline in live QSO operation on these modes, while traditional RTTY is still widely used. By concentrating on the special nature of live-QSO operation, a new RTTY mode (I don't call it a "data" mode) has been devised, which uses modern DSP techniques and uses the frequency stability of today's HF radios. The bandwidth is much narrower than any other telegraphy mode. Fig 4 shows the spectrum occupied by PSK31 and Fig 5 compares this to the bandwidth of a PACTOR signal.

At the time of writing (February 1999) PSK31 is available for the Texas TMS320C50DSK with software written by G0TJZ, the Analog Devices ADSP21061 "SHARC" kit with software by DL6IAK and my own software for the Motorola DSP56002EVM. For the SoundBlaster card, DL9RDZ has written a *LINUX*-based program for the PC. Some commercially available DSP-based multimode controllers have already been upgraded to include PSK31 and more will follow. However, the most popular implementation of PSK31 so far is the *WINDOWS*-based soundcard program, which I have written for the soundcard. The DSP algorithms for PSK31 are being made available free-of-charge to bona-fide amateur programmers, so there should be a wide choice of PSK31 systems in the future.

News of the latest PSK31 developments and activity can be found at **http://aintel.bi.ehu.es/psk31.html**. The site also contains a link to information for those who want to implement their own PSK31 modem. □□

A Switching Power Supply for Beginners

Many hams are intimidated by switching power supplies because minor wiring errors can result in spectacularly violent failures when full power is first applied. This relatively simple supply includes instructions for low-voltage tests, so problems can be fixed before applying full power. Come try your hand with this safer 13.6-V, 15-A power supply.

By Raymond Mack, WD5IFS

Switching-mode power supplies, or just "switchers," have been commercially available for about 20 years, but they are just now seeing wide use in Amateur Radio. I attribute this delay to worry about interference and complexity of design. The primary advantage of switchers is that they can be quite compact, since the transformer, filter choke and final filter capacitor can be very small compared to those required for linear supplies.

There has been very little information available to amateurs on switchers with the exception of a few years of the *ARRL Handbook* earlier this decade. Chapter 6 of the *1992 Handbook* has a very good description of the electronics involved in creating a switcher, but there are no projects or descriptions of how to actually build such a supply. It is my aim to provide information that allows you to create a simple, but useful, switching power supply.

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Why Build This Supply?

Unless you are very good at scrounging parts, this supply could easily cost $100. Some commercial sources have better supplies for $79. This design is presented as a means to learn about the subject and still have something useful at the finish.

The Off-Line Switcher

Switchers come in an incredible number of flavors. We are going to examine a forward-mode, off-line switcher. It is called an off-line switcher because it gets its power from the ac power lines. For our purposes, an off-line switcher is composed of four sections. The first is an EMI filter to keep the RF it generates from affecting other electronics. The second is an ordinary full-wave capacitor-input power supply. The third is a transformer-isolated dc/dc converter, and the fourth is a voltage regulator. Our example power supply will be a 15-A, 13.6 V dc supply suitable for powering a typical 100-W amateur transceiver. This design will provide about 10 A CCS or 15 A ICAS.

Safety First

It is very important to remember that everything on the primary side of the transformer has *no* isolation from the power line. When troubleshooting the primary side of the power supply, *always* use an isolation transformer for safety. A ground-fault-interrupter outlet is also a good investment for your workbench, if you are working on a switcher. During initial checkout, I use back-to-back filament transformers as an isolation transformer, as shown in Fig 1. I use a 10-Ω power resistor as a load for initial testing.

DC/DC Converter

This section puts the "switch" in switch-mode. Those of us who have experience with class-C finals in transmitters are already familiar with this circuit. It is the same thing as a master-oscillator/power-amplifier transmitter. The only difference is that we rectify the output of the final amplifier instead of sending it to an antenna.

Fig 2 is a block diagram of the dc/dc converter. Three blocks make up this section. The first block is simply a square-wave oscillator with a 50% duty cycle. The oscillator drives the next block which is the switch transistor, output transformer, and rectifier. The third block is the output filter.

All of the magic in a switcher is in the switch transistor and transformer. To understand how the forward converter works, let's start with the first pulse at time T_0 that turns on the transistor (see Fig 3). Current begins to flow through the primary of the transformer, which causes current to flow through the secondary and output rectifier into the load. When the pulse finishes at time T_1, the transistor turns off. The transistor turns off while a considerable current is flowing in the parasitic inductances of the transformer windings. There is no way for the current to continue in the primary because the transistor is off, and no current can flow in the secondary because the output rectifier is reverse-biased. We need some way to dissipate the residual flux in the core so we can start the process over. Power supply engineers call this "resetting" the core. It is imperative that the flux in the core is reset after every pulse. If the flux is not reset, the flux will eventually build up to the point where the core becomes saturated. At that point, the transformer will have very little magnetizing inductance and the current through the switch transistor can rise to destructive values.

There are two common ways to reset the flux stored in the transformer. The simplest is to add a third winding to the transformer, called a reset winding. It is connected in the reverse direction from the primary and secondary (see Figs 4A and 6). D3 only conducts when the transistor is off. The energy stored in the transformer is returned to the input supply. The second way to reset the core looks just like the circuit for a relay or motor driver (see Fig 4B). We use the resistor to dissipate the energy stored in the transformer. Again, D3 only conducts while the transistor is off. The capacitor helps to smooth the pulses and provide a constant voltage for predictable resetting of the core. Higher voltages across R reset the core faster. R and C could be replaced with a high-voltage, high-wattage Zener diode and have the same effect (see Fig 4C). You probably don't see this commercially because the cost of R and C is probably $^1/_{10}$ that of a Zener. The voltage on the drain of Q4 will rise above the supply voltage by the amount required to reset the core. In our supply, it rises to twice the input supply voltage. The reset voltage is set by the ratio of the turns in the reset winding to those in the primary winding. It is possible to adjust the maximum drain voltage by constraining the duty cycle and adjusting the turns ratio. Lower drain voltage requires longer core reset times.

The drain current has a trapezoidal shape. It jumps up to the value that corresponds to the current flowing in L5 when the transistor turns on. It increases linearly as we build up field in L5. When the gate drive pulse ends, the drain current drops almost immediately to zero. The L5 current then decreases linearly for the rest of the cycle.

The waveforms in Fig 3 show ringing when Q4 and the 3524 regulator chip switch on and off. A substantial amount of that ringing is because of the wiring method I used in the prototype. A PCB with nice wide traces would eliminate much of it.

The output filter is a standard choke-input power supply filter, with one exception. Q2 acts as a commutating diode so that when the transistor stops conducting, the current in the filter choke flows through Q2 instead of the transformer secondary. Without Q2, the transformer core would not reset properly. An additional advantage is that it reduces the maximum reverse voltage for Q3 to the input voltage rather than twice the input voltage.

Selection of the filter capacitor is important. All capacitors have an equivalent series resistance (ESR) in series with the capacitance. This ESR models the heat losses inherent in the conductors of the plates, as well as the dielectric loss. Normal aluminum electrolytics have very large ESRs that increase with frequency. When you select an output filter capacitor, be sure to look for the words "low-impedance" and "high-frequency." The Pioneer HFS series is an appropriate selection. A normal electrolytic has the potential to get explosively hot.

Voltage Regulator

The average value of the input waveform to a choke input filter is the same as the filter output's average value (less the filter's parasitic losses). This is true regardless of the input wave

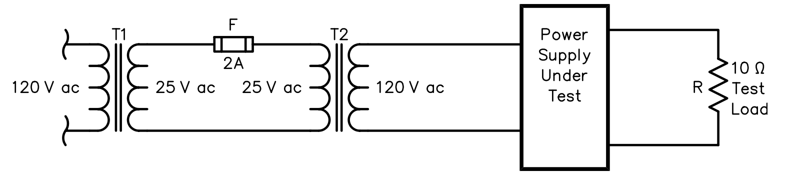

Fig 1—Back-to-back filament transformers used for isolation during initial tests.

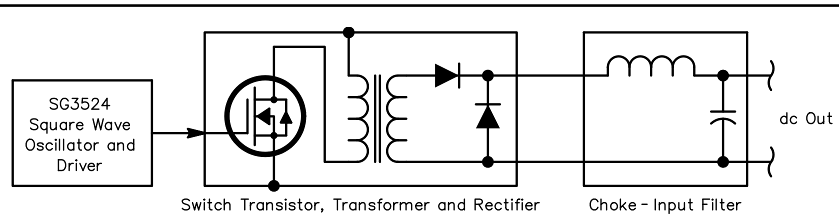

Fig 2—The Block Diagram.

shape. The voltage regulator takes advantage of this property to keep the output voltage constant. Our voltage regulator changes the average value of the input waveform by keeping the frequency constant and changing the on time of the switch. The average is simply the input voltage times the on time divided by the period of the oscillator:

$$V_{out} = V_{in}\left(\frac{t}{T}\right) \quad \text{(Eq 1)}$$

The math and the control circuitry are simple because we are working with rectangular waveforms. During testing, I used an 80-V dc lab supply instead of the input ac supply. This gave an input voltage of 10 V to the filter. When the duty cycle was 30% (t/T), the output voltage was 2.97 V.

Three things require the regulator to adjust the pulse width. These are changes in load current (load regulation), changes in line voltage (line regulation) and input-supply ripple voltage. It is not necessary to design a low-ripple input supply since the regulator adjusts the output voltage 833 times during each cycle of input ripple.

The voltage regulator is a circuit of logic gates and comparators between the oscillator and the switch transistor. The voltage-regulator circuit reduces the duty cycle of the drive to some value less than 50% as needed to maintain a constant output voltage. The dc/dc converter must be designed so that a 50% duty cycle will provide the desired output voltage at the largest current load and the lowest input line voltage. As input line-voltage rises or current output decreases, less duty cycle will be needed.

One of the comparators in the 3524 regulator provides current limiting for the supply. It does this by limiting the peak current to the transistor switch. This has the benefit of providing a soft start for the output as well.

Circuit Description

Figs 5, 6 and 7 comprise the schematic for our power supply. Since we are going to use this supply for a very sensitive receiver, we want to make very sure that the strong RF generated by this supply does not reach the power lines. We place a multistage EMI filter on the input to block the RF. Note that we have filtering from line to line but none to ground. C1 and L1 need to be mounted as close as possible to the power-entry connector to limit the energy conducted out of the power supply. You must use power-line safety-rated capacitors for C1, C2, C4 and C5. The Panasonic TS series is appropriate.

We use a bridge rectifier for 240-V ac (Fig 6) and a doubler for 120-V ac input (Fig 5). Either circuit converts the power mains voltage to a nominal 340 V dc. This ensures that the dc/dc converter is the same for either line voltage. The input voltage is chosen by simply changing the connections to the diodes and filter capacitors.

It is important to select the proper diodes for the input supply. Unlike a supply that uses a transformer, there is minimal inherent resistance ahead of the diodes. We must assume that the power switch will be activated at the peak of the input waveform. The resistance of L1, L2 and R1 are all that serve to limit the inrush current. The resistance of the two inductors is approximately 1 Ω. We select R1 (1 Ω) so the maximum inrush current is approximately 180 A for the 240-V ac input and 90 A for the 120-V ac input.

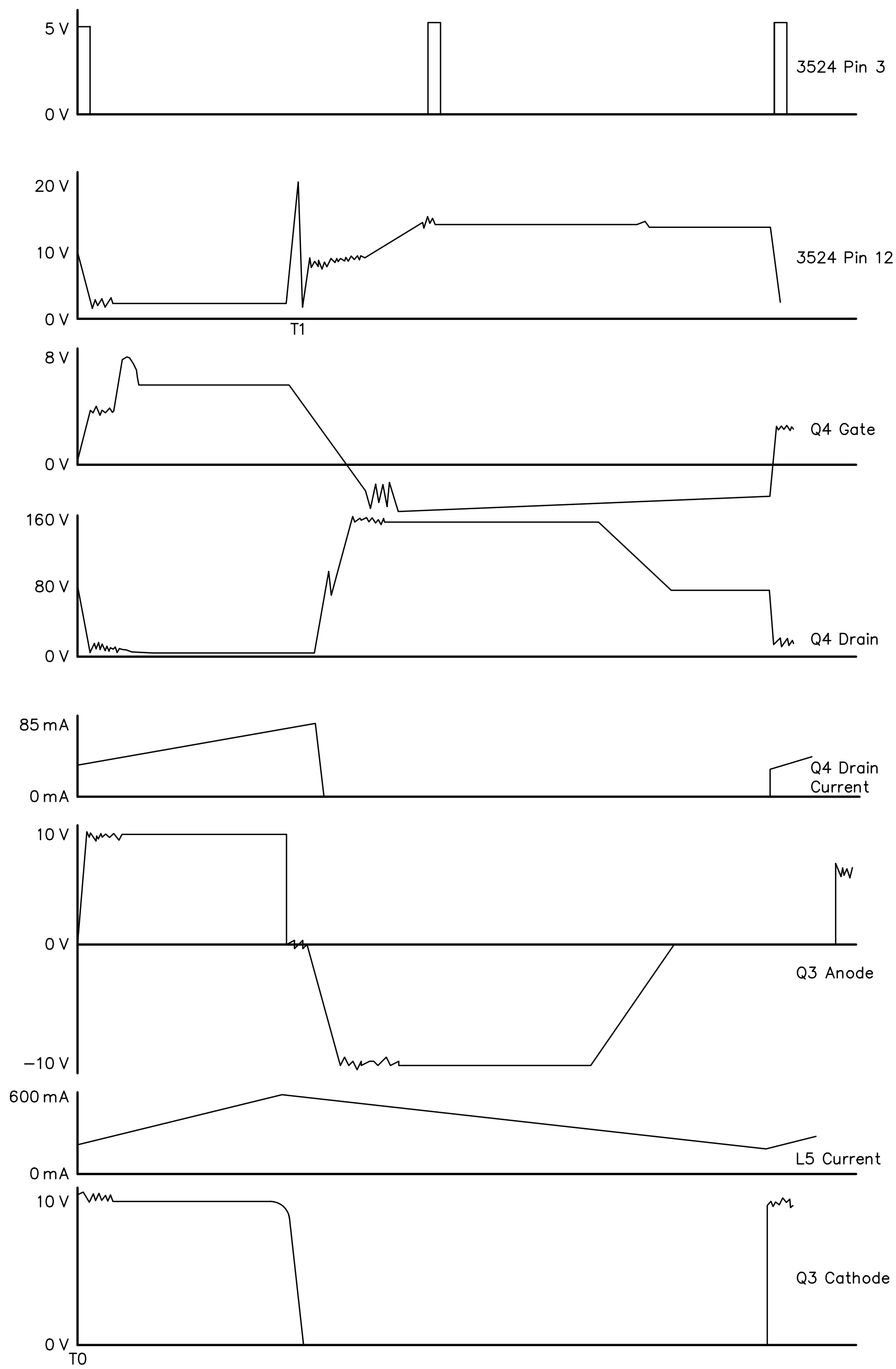

Fig 3—Switcher waveforms.

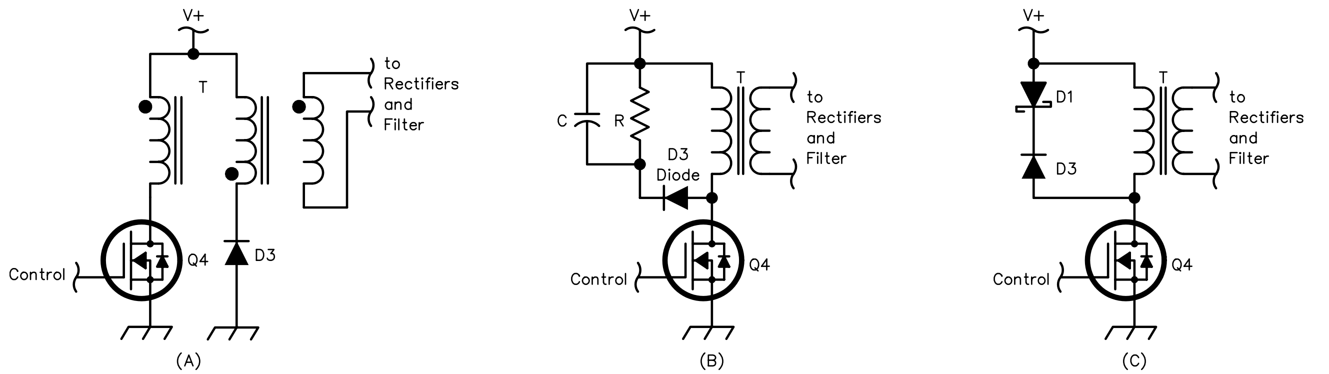

Fig 4–Three circuits to reset the flux stored in the transformer.

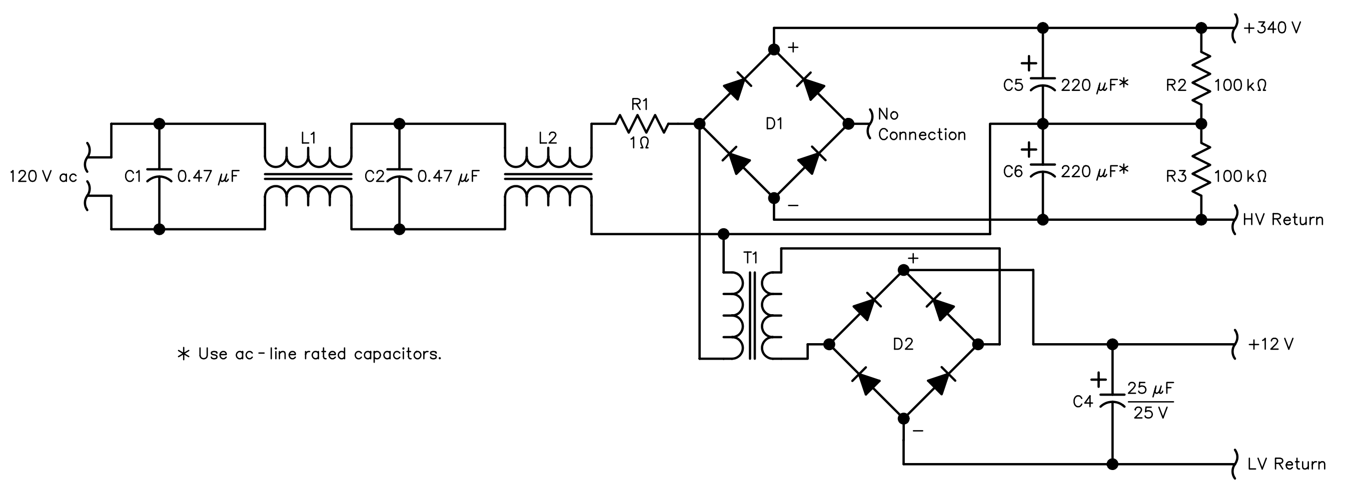

Fig 5—The 120-V ac input supply.

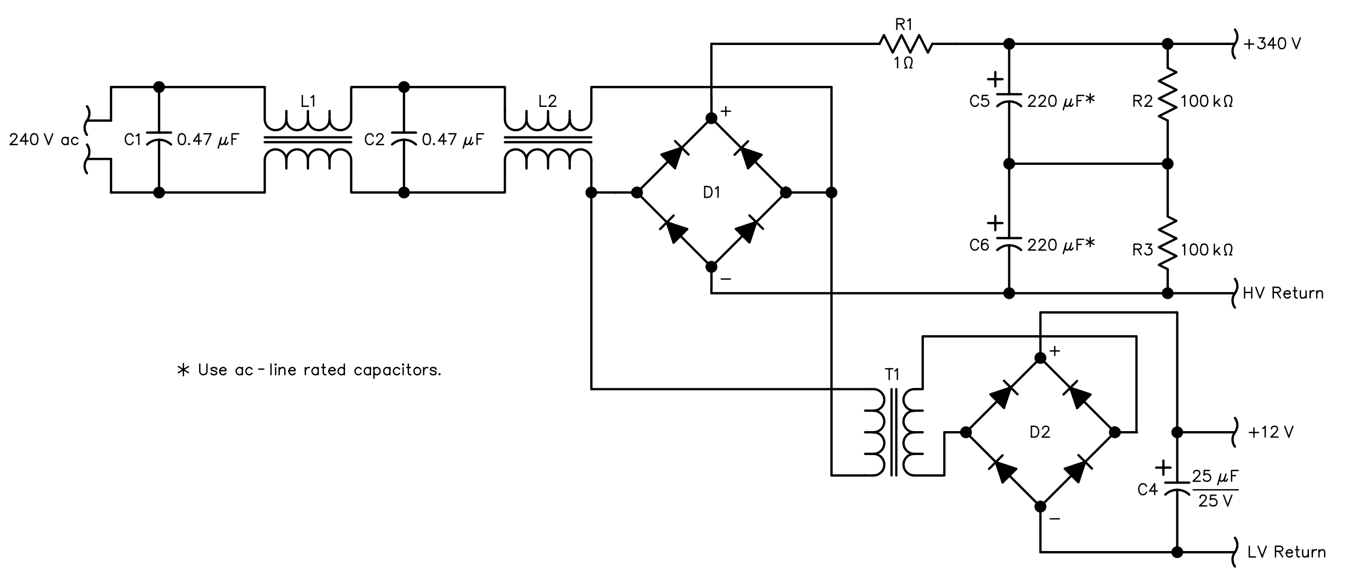

Fig 6—The 240-V ac input supply.

We looked at the ESR of the output capacitor earlier, but ESR in the input capacitance can be a problem as well, since we allow a moderate ripple voltage and ripple current. This is the equivalent of a fair ac current flowing through the input capacitors. We look for a series like Panasonic EB that is specified for high-ripple applications.

The switch transistor can see double the input voltage while the core resets, so we need a transistor rated for twice the absolute maximum input-supply voltage. The power line can be 120% of the nominal. This means 240 V ac could really be 288 V ac. 288 V × 1.414 × 2 gives a maximum transistor voltage of 814 V dc. To be safe, we pick a transistor with a 900-V rating.

The output rectifiers are the next components to select. It takes a finite amount of time for the capacitance of a rectifier to discharge when it becomes reverse biased. This is called the reverse recovery time. The faster we run the switcher, the more important this time becomes, because a large current spike gets sent back to the transistor during reverse recovery. Schottky diodes have the shortest reverse recovery times, but they are only available in voltages up to 100 PRV. They also have the added benefit of a low forward-voltage drop, so they dissipate very little power. Unfortunately, they are also expensive. Another alternative is to use the inherent diode of a power MOSFET. The IRF540 can handle 27 A of forward current and has a typical reverse recovery of 150 nanoseconds. The forward voltage drop at 15 A for the IRF540 is 1.0 V. These parts will need a heat sink since they will have a peak power dissipation of 15 W.

Probably the biggest roadblock for amateurs building switchers has been the lack of readily available transformer cores. Fortunately, our old friend Amidon supplies a line of E-shaped cores designed for use in switchers. We need their largest core, which is designed for use in 200-W supplies. The transformer turns ratio must be selected so that we supply enough voltage to the filter at maximum output current and lowest input voltage. We choose to let the ripple voltage drop all the way to 200-V dc for this worst-case situation. This will require a duty cycle of 50%. The secondary voltage is then (13.6 / 0.50 + 1.0) V. The turns ratio is then 200 / 28.2 or 7.1 / 1. This value also determines the minimum PRV rating of the output diodes. The maximum input voltage is 407 V, so the diodes can see 407 / 7.1 or 57 V.

A variety of IC regulators are designed for use in switchers. The SG3524 series is manufactured by most companies doing linear ICs, including Motorola, SGS, National and LTC. They are available from sources such as Jameco, Digi-Key and Future Active. This is an old part (about 20 years old), but it's still readily available. This part contains the master oscillator, the regulator circuitry and the driver.

Fig 8 is a block diagram of the circuitry inside the 3524. The oscillator is a circuit very much like the 555 timer. The comparators, gates and voltage reference are used to modify the 50% duty cycle of the oscillator to something shorter. The two drive transistors are driven 180° out of phase, so this circuit can be used in one-or two-transistor switchers. Each transistor is capable of providing 100 mA of drive. The flip-flop inside the 3524 divides the oscillator by two, so we need to set the oscillator for 200 kHz to get operation at 100 kHz. Many application notes for the 3524 show the transistors in parallel for single-ended circuits like ours, but this does not ensure a maximum 50% duty cycle. We can only use one transistor to drive the FET.

We need a small auxiliary supply for the 3524. I just grabbed a small "wall-wart" supply from the junk box and cut the transformer out of the plastic. This supplies about 12 V dc to the regulator circuit. The isolation from the power line to the output is maintained by using a pulse transformer to drive the gate of Q4.

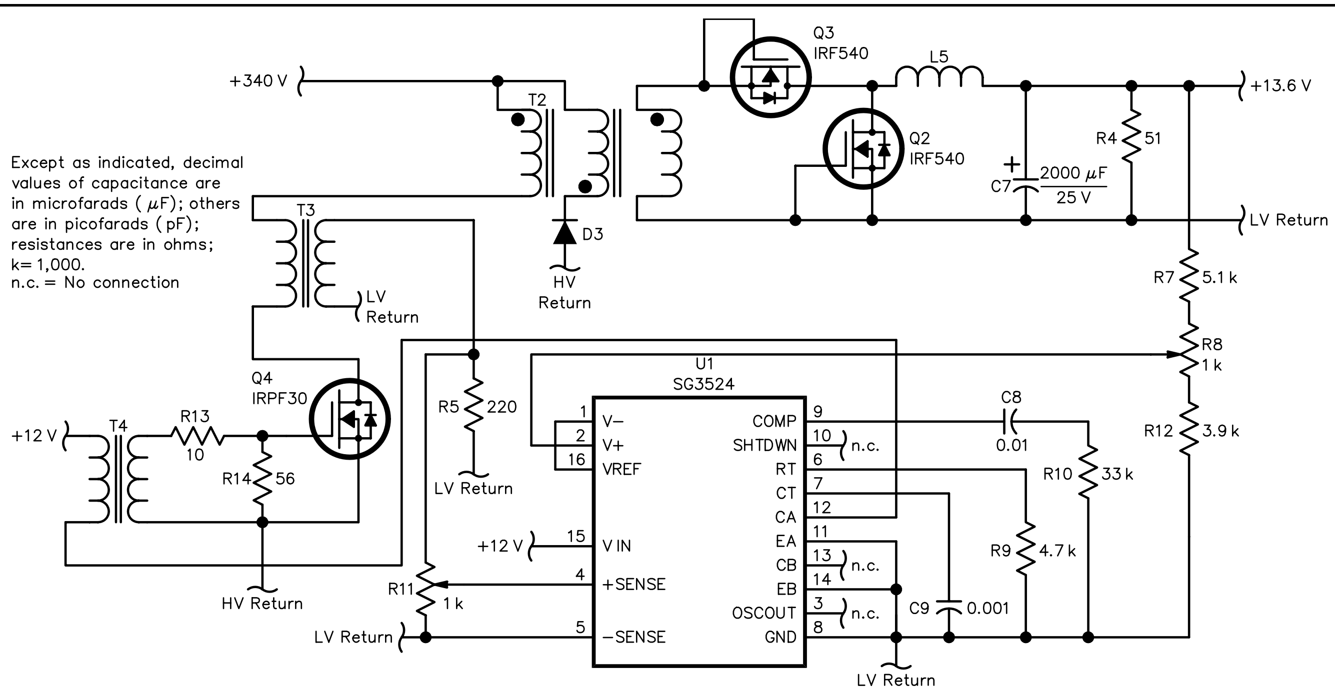

Fig 7—Schematic of the dc-to-dc converter plus regulator and output filter.

Switching regulators require a small amount of output current—at all times—to work properly. The output duty cycle can be made quite small, but it cannot be reduced to zero. We use a 51-Ω resistor to assure there is always a load for the supply.

Construction

The Amidon core comes in two pieces. You should wind the primary around the center post of one piece, and the secondary around the other piece. Be sure to use Teflon tape around the core for additional power-line voltage isolation. You can also build or buy a plastic bobbin for easier winding. After you have made the windings, you will need to assemble the transformer. In commercial supplies, the core halves are held together with a very thin layer of epoxy. You can use any method that holds the cores together and doesn't create a shorted turn.

A printed circuit is not necessary for this project. Normal protoboard construction for the regulator is quite acceptable. The input supply, dc/dc converter and filter can be built on circuit board material using the razor-knife approach. The most important part of construction is to keep the primary voltages and the secondary voltages physically isolated as much as possible.

You will need a small heat sink for Q4. It will dissipate from 1 W up to perhaps 20 W at full load and low line voltage. It is very important to remember that this transistor has 400 V dc on the drain. Be sure to use a heat-sink insulator and silicone grease between Q4 and the heat sink. Also, remember that the primary side is connected directly to the power line! Diodes Q2 and Q3 also need heat sinks.

The whole supply should be enclosed in a metal compartment so that we don't radiate RF into the radio. Conversely, it will keep the transmitter's RF from affecting the regulator circuits. Be sure to provide adequate ventilation for the heat generated by R1 and Q4. A small fan may be necessary to remove the excess heat. The power supply can then be mounted inside a convenient cabinet with associated power switch, fuse and output connectors.

Troubleshooting

Regulated power supplies are closed-loop systems, so it can be tough to troubleshoot problems. I start by breaking the connection to R7 and connecting it to the 12-V supply for the 3524. Then you can adjust the voltage pot (R8) to get approximately 50% duty cycle to the FET. Then I hook up a low-voltage supply to the transformer circuitry. One of the 25-V ac filament transformers connected to the input supply will work, or you can use a high-voltage lab supply. That will allow you to investigate the various circuits without the full stress of 340-V dc on the transformer.

One additional word of caution: I blew several fuses in the lab supply before I realized it didn't like the energy returned to it by the reset winding. They source current just fine, but most lab supplies won't sink it. I attached the lab supply to the negative output on one side, and the positive of the lab supply to the hot ac connection.

With 40-V dc on the transformer, you should get about 2-V dc out. The waveforms should be pretty close to those in Fig 3, just smaller. Once you have verified that the waveforms are correct and that you have all the transformer windings phased correctly, you can reconnect R7 and hook up the isolation transformer.

Parts and Suppliers List

Parts

C1, C2—Panasonic 470 nF (Digi-Key P4614-ND)
D3—Liteon ultrafast rectifier (Digi-Key UF1007DICT-ND)
L1, L2—Magnetek Triad 2.37 mH or Panasonic 2.7 mH (Digi-Key PLK1006-ND)
L3—2 stacked 2-inch, mix #26 cores (Amidon T200-26), 25 turns (t) #14 enameled wire
D1—Motorola MDA990-3
T2—Amidon EA77-625 core. Winding 1 is 30 t #22 enameled wire. Winding 2 is 30 t #22 enameled wire. Winding 3 is 4 t #14 enameled wire.
T3—Mix 77 toroid 0.5 inches in diameter (Amidon FT50-77). Primary winding: pass T2's winding-1 lead through T3 (1 turn). Secondary: 50 t #28 enameled wire.
T4—Magnetek-Triad 1:1 gate-drive transformer (Digi-Key 10552-ND)

Suppliers

Farnell (Newark)	tel 800-718-1997	http://www.farnell.com
Bytemark (Amidon)	tel 800-679-3184	http://www.bytemark.com
Jameco	tel 800-831-4242	http://www.jameco.com
Digi-Key	tel 800-344-4539	http://www.digikey.com
Future-Active	tel 800-655-0006	http://www.future-active.com

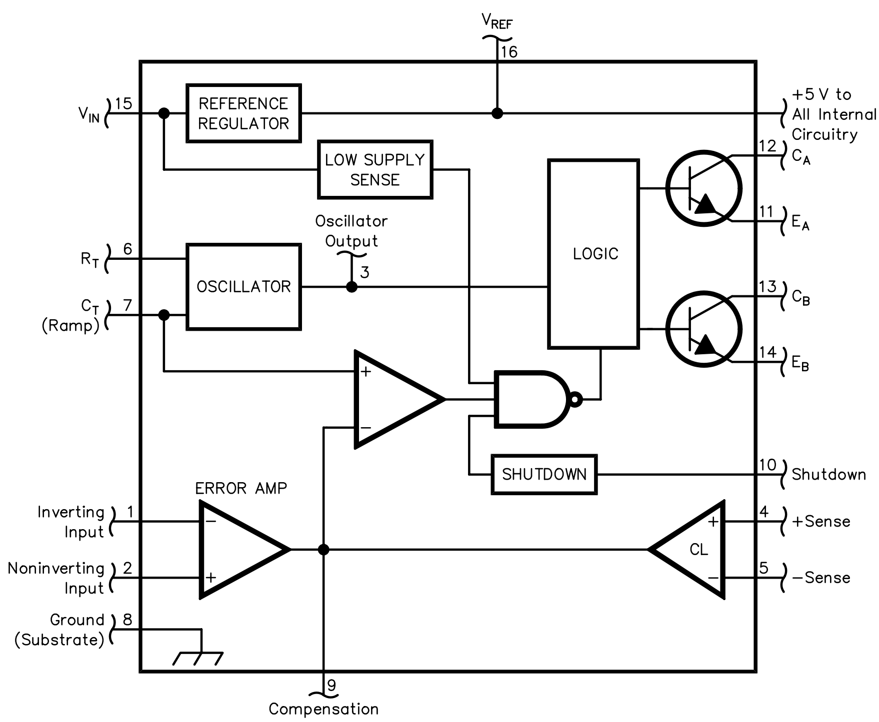

Fig 8—Block diagram of the SG3524 regulator IC.

Alternate Parts Sources

Defunct PC power supplies are a reasonable source for some of the parts. Just about any company PC support group or computer repair shop will have a few lying around. The input supply capacitors, bridge diodes, EMI filter and the heat sinks for Q4, Q2 and Q3 are all likely candidates. R1 can also be replaced with the inrush thermistor from a PC power supply.

It is also possible to use the pulse transformers, main power transformer and filter cores from most of these supplies. I have had some success using a cold chisel to knock the core halves apart. One sharp (but not very forceful) rap at the joint usually breaks them apart. For E-cores, I usually end up with three or four big pieces. This isn't a problem. Just glue them back together with cyanoacrylate (such as Krazy Glue—*Ed.*). *Do not* introduce a significant gap with the glue, though. I usually end up removing all the windings and making my own, so I know how many turns are on the core. The biggest problem is that the cores are small for my windings. The #14 wire on this project was not usable for most PC cores. Copper foil as used in the January 1999 *QST* article[1] is an alternative for use on a PC supply core.

The prototype in the photos uses Ferroxcube 4229 pot cores. I tried 3B7 and 3C8 materials. These old materials were originally used for 20-kHz supplies. I found that both materials worked okay in this application without undue heating. Additionally, you can buy the latest Ferroxcube high frequency, low-loss cores (3F3, 3C80 and 3C85 materials) from Farnell, Newark Electronics in the USA. When choosing a substitute for the power-transformer core, look for adequate cross-sectional area on the center leg of the core. A core with an area of 1.8 cm^2 (0.28 $inch^2$—*Ed.*) or larger is adequate for this power level.

The output-filter cores are color-coded. Find the largest cores that are coded yellow/white. I used two cores about two inches in diameter stacked side by side, and wound 25 turns on them. □□

[1] M. Mornhinweg, XQ2FOD, "A 13.8-V, 40-A Switching Power Supply," *QST*, Part 1 Dec 1998, pp 37-41; Part 2 Jan 1999, pp 41-44.

Center-Loaded Whip Antenna Loss

Aren't those skinny little HF mobile antennas great? They decrease the "QRMS" (spousal interference) and lessen the nerdly appearance of HF mobile hams, but what is the cost?

By Grant Bingeman, KM5KG

It is a common practice among mobile-whip antenna manufacturers to compromise the efficiency of their antennas by using very "lossy" center-loading coils to offer a shorter whip. This means they can build a shorter antenna that still provides an input impedance reasonably close to 50 Ω. Manufacturers only rarely tell you that what you get for your money is a much lesser signal strength, since much of your transmitter power is wasted in the coil and never gets radiated.

Simple mobile-whip antennas have practical height limits governed by trees and bridges. At frequencies below the 15-meter band, a quarter-wave whip is physically too tall for most mobile applications. The nice thing about a quarter-wave whip is that its input resistance is naturally close to 50 Ω, and the reactance is small. Thus, no special impedance manipulation is required to produce a low SWR for the transmitter if one has the room. For bands below 15 meters, mobile antenna designers must deal with the low resistance and high negative reactance typical of an electrically short whip.

One approach for dealing with a short vertical is to place a series coil at the base of the whip to tune out the capacitive reactance. An alternate coil placement near the center of a 0.1 λ whip would typically need be twice as large to achieve the same tuning effect as a coil at the base. However, a coil near the center of the whip has the advantage of changing the current distribution along the whip so that the base input resistance is raised. When you add the loss resistance of the coil, it is easy to produce a net input resistance of 50 Ω—hence a low SWR—at the base of the whip.

For example, a seven-foot whip mounted on a ground plane has an input impedance of about 12 –*j*430 Ω at 14.25 MHz. You would need a base inductor of 4.8 μH to tune out this reactance and a matching network to bring the 12-Ω resistance to 50 Ω; or you could use a very lossy coil having a resistance of 38 Ω, or a Q of 11. You might as well just operate into a dummy load, because you would lose

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76% of your power in the coil before it ever radiated.

We will see later that this kind of compromise is not as severe if we mount the coil near the center of the whip. Yet, a significant loss of power and field intensity occurs. You should consider that when shopping for a mobile antenna. In general, get the tallest whip you can stand. First, some background.

Analysis

A quarter-wave whip antenna mounted over an aluminum ground plane atop a van roof, as in Fig 1, produces an input impedance of $55 - j36\ \Omega$ at 14.25 MHz above 5 mS/m earth having a dielectric constant of 13. Clearly, this vertical whip is too tall to serve as a practical mobile antenna. It produces a reference forward gain of about −0.3 dBi at an elevation angle of 24°, as shown in Fig 2. Note that the azimuthal radiation pattern from this van is somewhat directional, favoring the forward direction. The front-to-back ratio is only about 1 dB, so the overall circularity is ±0.5 dB—essentially omnidirectional.

The van is a typical full-size, short-wheel-base version. Its wire-grid model consists of one-inch diameter wires, segmented every foot. NEC2 was used to analyze the antennas described in this article. All calculations were done at 14.25 MHz, in the SSB portion of the 20-meter Amateur Radio band.

It is common knowledge that mobile whip antennas can be made shorter with the addition of a coil near the center. Ideally, this coil changes the current distribution on the whip to increase the base input resistance, a reduction of the capacitive reactance and perhaps a slight improvement in gain.[1] This center-mounted coil should not be confused with the commonly-seen base-mounted coil, which is primarily designed to tune out the capacitive reactance of an electrically short antenna and provide an L network (when the coil is tapped) to match a low resistance to 50 Ω. Nor should the center-loaded short whip antenna be confused with electrically longer dual-element collinear arrays that use a center coil for phasing. Collinear antennas are normally used only at VHF and UHF because they would be too tall for mobile installation at lower frequencies.

The original premise of the center-loaded whip-antenna design assumes a high-Q coil located high up the antenna. Unfortunately high-Q means a big coil with lots of wind resistance and weight. The high Q is necessary to avoid excessive coil losses and thereby to provide maximum gain. As we shall see shortly, you can easily lose 3 dB of gain in a low-Q center-located loading coil.

Many whip antenna manufacturers provide a very lossy center-located loading coil, close-wound with #22, or smaller, wire. This lossy coil may be wound on a fiberglass pole and covered with shrink-wrap, or in very short piano-wire whips, it might look like a plastic ball or cylinder. The manufacturers like the fact that the loss resistance of the coil increases the feed-point resistance of the antenna, so it approaches 50 Ω and provides a low SWR. In fact, the gain from this type of center-loaded whip is sometimes 6 dB below that of a properly designed whip antenna. (The resistance also "swamps" the antenna reactance, in-

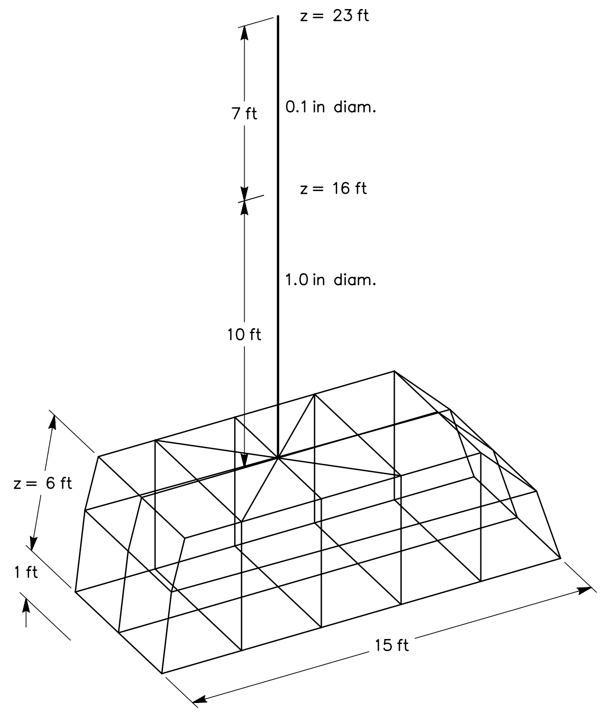

Fig 1—20-meter, quarter-wavelength whip mounted atop a full-size van.

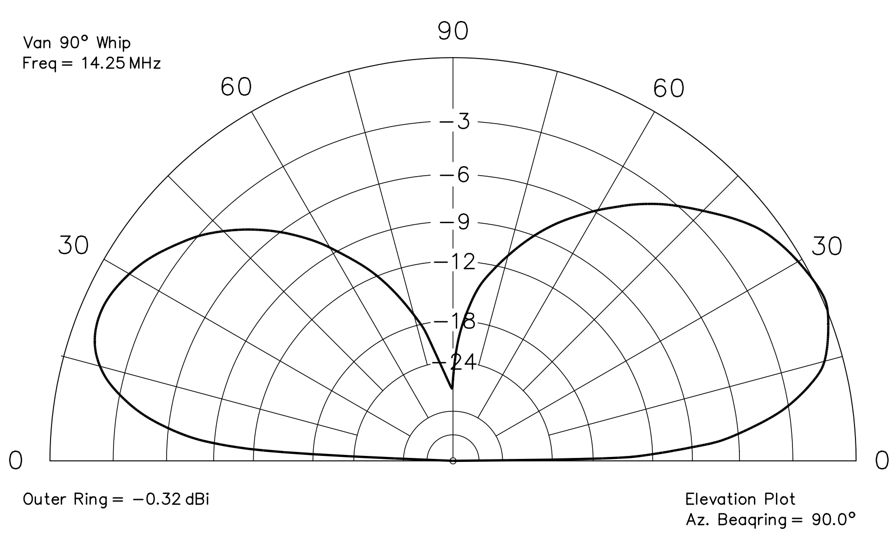

Fig 2—Radiation pattern of the van mounted 20-meter whip.

[1] B. Brown, "Optimum Design of Short Coil-loaded High-frequency Mobile Antennas," *The ARRL Antenna Compendium*, Vol 1 (Newington, Connecticut: ARRL, 1985).

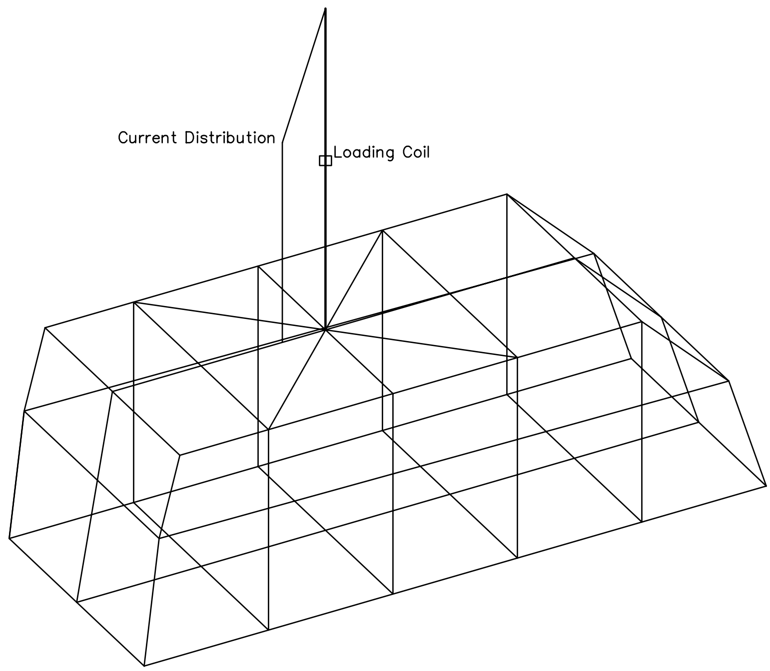

Fig 3—A typical center-loaded whip installation.

Table 1

Input Z	*Coil Z*	*Coil Q*	*Max Gain*
53 +*j*890 Ω	0 +*j*1200 Ω	∞	–0.2 dB
23 +*j*17	0 +*j*800	∞	–0.3
49 +*j*15	24 +*j*800	33	–3.6
12 –*j*430	0 +*j*0	short circuit	–0.4

creasing the operating bandwidth. In many cases, narrow SWR bandwidth indicates less loss and consequently more-efficient antennas—*Ed*.)

I have always felt that an antenna should be a radiator, not a dummy load. You might be surprised at how many low-efficiency ham antennas exist in the industry. The lossy whip is just one of many types of compact, but poor, antennas available to the unsuspecting ham. As a general rule, smaller antennas provide less gain at a particular frequency.

Let's look at how short we can make our 14.25 MHz whip if we cheat by inserting a lossy coil at its center. What is the price we pay in gain to get a reasonable input impedance and low-profile antenna? A typical 20-meter-band, center-loaded whip is about seven feet tall (Fig 3). The first four feet might consist of a slow spiral of #14 AWG wire wrapped on a 0.6-inch-diameter fiberglass pole. The top three feet of the antenna is typically strong, flexible piano wire. The length of the piano wire is adjustable with setscrews at the top of the fiberglass pole. Just below the setscrew's location, is an inductor of perhaps 90 turns in three inches. The coil does not have a very high Q, but does have a very high reactance.

A center-located loading coil does two things for us. First, it allows us to tune out the negative reactance that normally appears at the base input of an electrically short whip. Second, it increases the radiation resistance as seen at the base of the whip. And a lossy coil adds a third parameter by allowing us to increase the feed-point resistance; we add some loss resistance to the equation at the expense of gain.

As you can see in Table 1, an ideal 8.9-µH, 800-Ω coil just about doubles the input resistance at resonance. However, SWR is still greater than 2.0, so an external matching network would be desirable. If our coil Q were only 33, the coil impedance would be 24 +*j*800 Ω, which produces a much lower SWR at the base. No external impedance matching is necessary. Now we have two new problems: (1) Our radiated signal just dropped 3 dB. (2) Our coil may burn up if we hold the key down too long, because half of the transmitter power is converted to heat in the coil. If you have one of these antennas, have you ever performed a "feel test" of the coil after transmitting for a few minutes? Perhaps now you know why they paint those coils black.

So is it cheating to use a lossy center located loading coil in a whip antenna? Yes, if the coil is lossy, and the manufacturer doesn't tell you how much power is lost in the coil. Especially if they claim the advantages normally reserved for a loss-less coil, such as improved current distribution, etc. Yes, the current distribution on the antenna is improved in the lossy case, but the magnitude of the current is substantially reduced over the loss-less, "big-coil" case. The radiation pattern shape of the lossy case is the same as in Fig 1, but the pattern size is a lot smaller—an important difference.

Conclusion

If you see a ham radio whip antenna with a big open coil at its center—and perhaps a couple of nylon strings guying it in place—you know the owner cares about antenna efficiency. On the other hand, if you can't see a high-Q coil in the middle of a short whip, you know the owner either has chosen to match impedance inside the vehicle, or has settled for an antenna with a lossy, low-profile center located loading coil. In either case, it is a good idea to mount the whip in the center of the roof to reduce directional effects, and, if possible, to install a plate of aluminum under the base of the whip. The aluminum plate can be as small as a one-foot square. It will prevent the thin steel roof material from cracking and failing because of metal fatigue. The plate also reduces RF losses by shielding the lossy steel at the base of the antenna, where the near field is strongest. A copper, brass or aluminum base plate will improve gain.

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Diplexer Filters for an HF MOSFET Power Amplifier

A filter with wide-band matching cuts spurious emissions, while avoiding problems caused by out-of-band mismatch: reduced output, instability, poor linearity or efficiency. This filter suits a 120-W amplifier for the HF bands.

By William E. Sabin, W0IYH[1]

Filtering at the output of a solid-state linear HF SSB power amplifier (PA) is an important design problem for two reasons:

- Strong harmonics must be attenuated to acceptable levels.
- Interactions and reflections between the filter and the power amplifier affect the power level, efficiency, stability and linearity.

The third harmonic, in particular, is typically only 10 to 15 dB below the fundamental. In well-balanced push-pull amplifiers, the second harmonic is typically down 40 dB or can be improved to that level. Fig 1 shows a spectrum analyzer plot of a 120-W push-pull amplifier with no filtering. At the 120-W level, the FCC presently requires a minimum of 40 dB, and 55 dB is a desirable (and sufficient) goal for our expected needxs in the future.

In addition to reducing harmonic products, the filter input should present the correct 50 +*j*0 Ω load resistance at the operating frequency, for which the power amplifier was designed. The amplifier then has the desired output level and linearity, as normally determined by two-tone intermodulation tests. Fig 2 shows the worst-case two-tone products of my 120-W, 1.8 through 29.7 MHz home-brew MOSFET power amplifier.

Another consideration is freedom from oscillations and significant regeneration. Oscillations can be free running; or they can be triggered by the desired signal, by switching the B+ off and on or adjusting the B+ and bias levels up and down. There are many insidious ways for instabilities to occur, but the ones that we will consider involve the wide-band (especially stop-band) impedance that the filter input presents to the transistors.

Fig 3 shows a typical power-amplifier output driving an LC low-pass filter. Outside the passband, especially *just* outside, the input impedance of the filter is highly reactive. This reactance can be transformed in complicated ways by the transformers, reactances and transmission lines that lie between the transistors and the filter input.

The impedance presented to the transistors in this stop band can be of such a high magnitude and so reactive

[1]Notes appear on page 26.

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at certain frequencies that instability problems are encouraged by feedback mechanisms inside the transistors. At the higher frequencies, and particularly with MOSFETs designed for these higher frequencies, problems can occur, especially when the output-load impedance of the filter is moved around a 2:1 SWR circle. One specific problem is that a harmonic of an in-band signal can land on a frequency in the stop band where a response anomaly is located. A spectrum analyzer with a tracking generator, sweeping at various power levels, gate bias and drain voltage, is a very valuable asset for detecting these anomalies.

Also, my experimental experience has verified that the harmonics, especially the third, are quite often reflected by the filter and returned to the transistors at such an amplitude and phase that intermodulation distortion (IMD) products are degraded in unpredictable ways, and are therefore difficult to specify.

The methods commonly used to stabilize the power amplifier are negative feedback and resistive loading, as shown in Fig 3. As one example, the conventional output transformers get hot, and are therefore a constant resistive loading on fundamental and harmonic frequency products. My preferred design uses negative feedback, but uses a transmission line output transformer (1:4 impedance) that dissipates almost no power, and virtually runs at room temperature. This approach is simple with MOSFETs operating at 40 to 50 V dc and at the 120-W level. That is, the peak-to-peak RF voltage is a sufficiently small fraction of the dc supply voltage that the class-AB operation is highly linear. This assumes that linearity has a higher priority than power efficiency and maximum output.

Solid-state-power-amplifier design can be a tough game (see Note 1) unless we copy a well-established design from

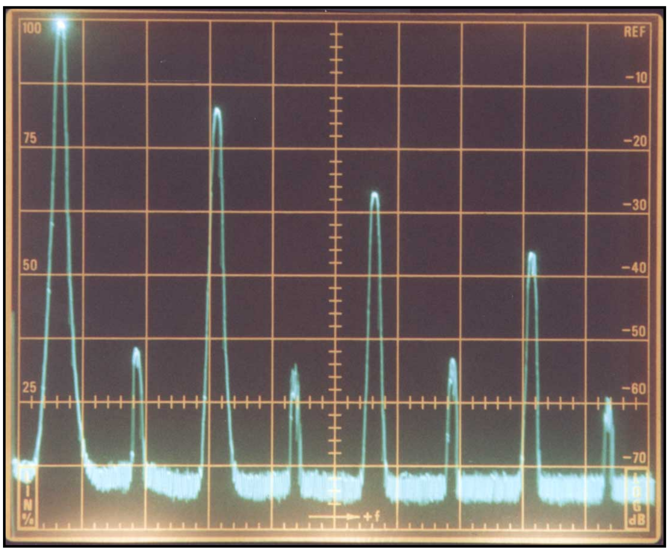


Fig 1—Spectrum-analyzer plot of a 120-W amplifier with no low-pass filtering. Fundamental frequency is 3.5 MHz and the sweep is 1.8 to 30 MHz, 10 dB/division.

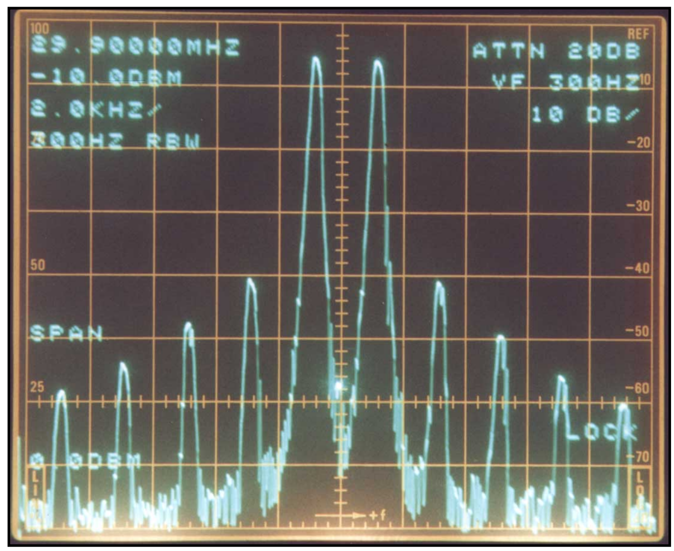


Fig 2—Two-tone intermodulation products at 29.9 MHz, 120 W PEP.

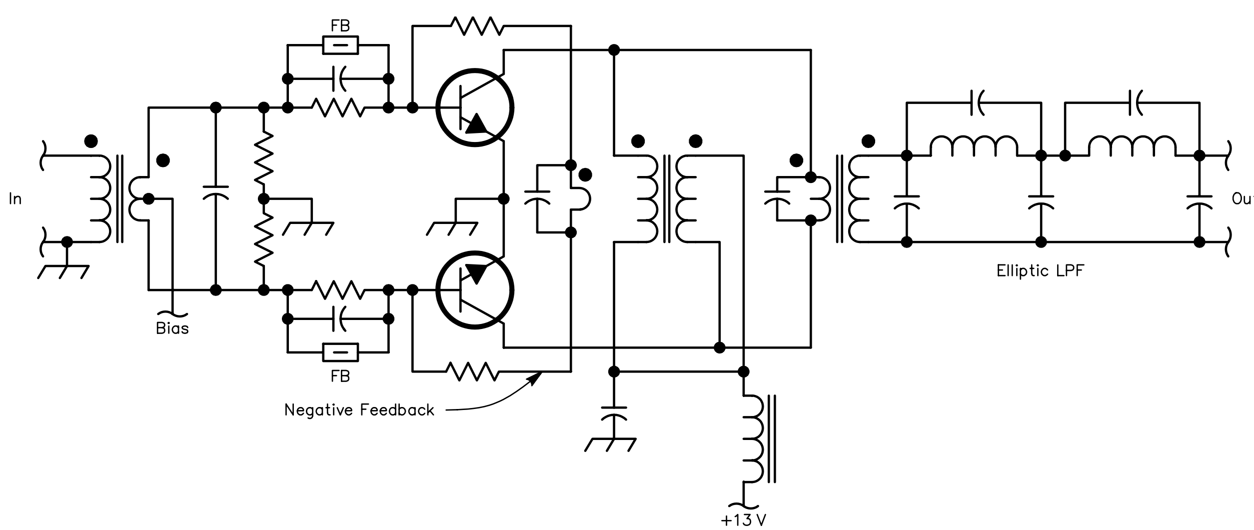


Fig 3—Typical transistor power amplifier circuit arrangement with conventional transformers and feedback.

a publication or kit. My experience with homebrew power-amplifier design efforts has been that if the amplifier is stable with a broadband 50-Ω load, stability and linearity with filtering installed are much easier to get with the diplexer filter method described in this article. It also improves confidence that transistors will not be zapped by a large oscillation of some kind. The restriction is that this filter method is most feasible in narrow-frequency bands, such as the HF ham bands.[2] The success of this method also requires that the second harmonic be reduced at least 40 dB prior to filtering by a well-balanced, push-pull amplifier, which I have found easy to achieve in a MOSFET amplifier using matched-pair of MRF150 SSB transistors.

The Diplexer Filter

The diplexer filter presents a load impedance to the power amplifier that is essentially 50 Ω, with a return loss (RL) of better than 25 dB (in principle), from dc to well beyond 50 MHz. Fig 4 shows a filter of this type for 80 meters, and Fig 5 is an idealized computer simulation, using *ARRL Radio Designer*, that shows its low-pass and high-pass frequency responses. A more realistic discussion is presented later. The return loss within the 80-meter band is better than 35 dB, which is quite good. The worst return loss is in the region of the crossover frequency (5.45 MHz). Figs 6A and 6B show the same spectrum as Fig 1 with the frequency components separated. The harmonics are dissipated in the 50-Ω "dump" resistor. We also see that this resistor dissipates a small amount of

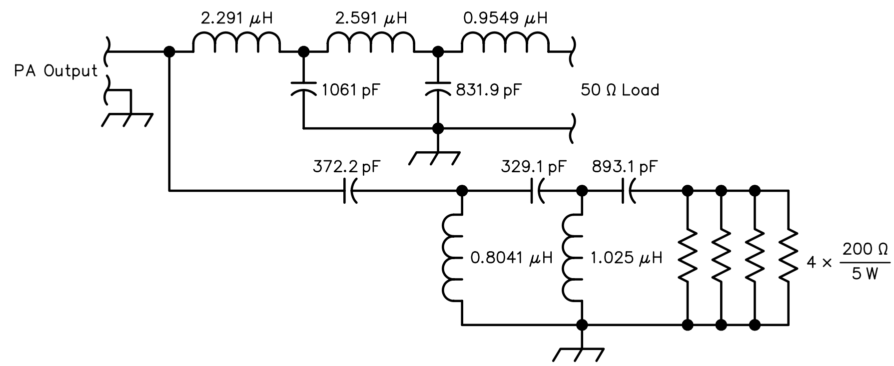


Fig 4—Diplexer filter for the 80-meter band with exact values.

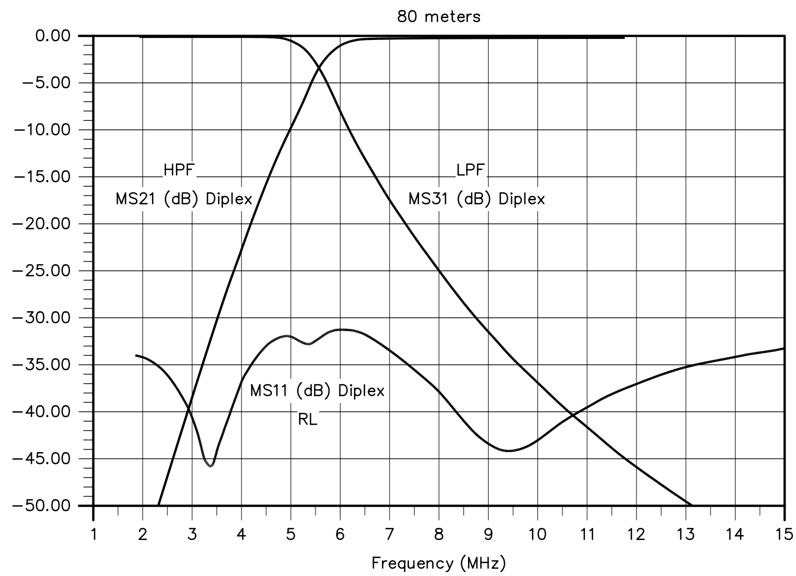


Fig 5—Computer simulation of the 80-meter diplexer.

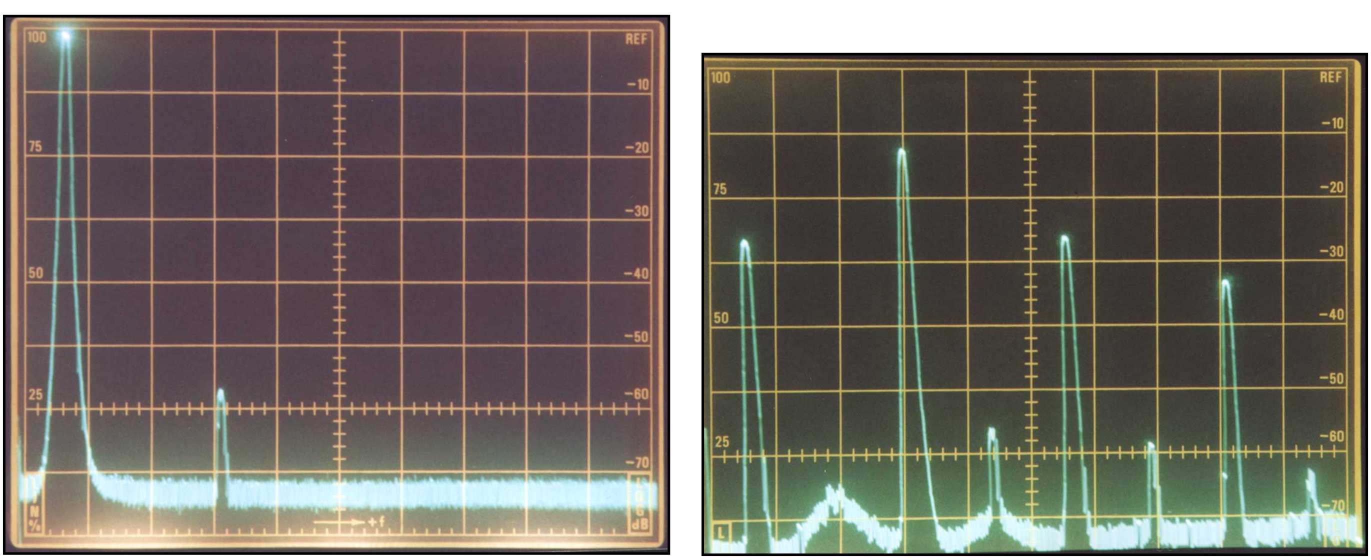


Fig 6—Same spectrum as Fig 1 except: (A) at output of low-pass filter, (B) at output of high-pass filter.

the fundamental frequency, and that the low-band filter output contains small amounts of the harmonics. This is also seen in Fig 5, where the high-pass response is down 22 dB at 4.0 MHz, and the low-pass is down 18 dB at 7.0 MHz, and 40 dB at 10.5 MHz. This 7.0 MHz item points out why we need a push-pull, balanced amplifier to get adequate attenuation of the second harmonic. Once that reduction is achieved, I have found it reliable.

Fig 5 also shows why the filter applies to a narrow band such as 3.5 to 4.0 MHz. For example, suppose the fundamental is moved to 5.0 MHz. The amount of fundamental lost in the dump resistor would increase quite a lot. If the fundamental is reduced to 3.0 MHz, the second (6.0 MHz) and third (9.0 MHz) harmonic reductions may not be good enough. However, we will see that three diplexers cover the 40/30 meter bands, the 17/15 meter bands, and the 12/10 meter bands. We can then cover all nine HF bands with six diplexers.

Designing the Diplexer

The diplexer is derived from the low-pass prototype of Fig 7 (upper part), which shows a zero-resistance voltage source. This important point is treated in the following paragraph. The filter is a five-element, low-pass filter with a series-inductor input (also important), a 1.0-radian-per-second (0.1592 Hz) cutoff frequency, and a 1-Ω load resistor. This filter may be a Butterworth, Chebyshev or Bessel type. I chose the 0.1 dB Chebyshev because of its steeper roll off. These normalized prototype element values are easily found in various tables[3,4] and are shown on the figure. The high-pass filter, Fig 7's lower part, is found by:

1. Replacing a series L (low pass) with a series C (high pass) whose value is 1/L, and

2. Replacing a shunt C (low pass) with a shunt L (high pass) whose value is 1/C.

The high-pass prototype values are also shown.

When the two filters of Fig 7 are combined as illustrated, the input resistance at the crossover frequency, and all other frequencies, is close to 1.0 Ω, even though each filter is 3 dB down at the crossover point. For the Bessel or Butterworth, this would be almost exactly 1.0 Ω. For the Chebyshev, there is a slightly larger error; but if we multiply each of the LPF values by some constant and divide the HPF values by the same constant, the return loss at the crossover can be improved several decibels.[5] I used the number 1.005, which was experimentally determined by simulation for filters whose inductors have a Q of 160. As we see from the return loss in Fig 5, the two filters terminate each other quite well at and near the crossover frequency because their input susceptances are complex conjugates. For this reason, the transfer characteristic of the diplexer is pretty much insensitive to the impedance of the generator,[6] which works in our favor for a solid-state power amplifier whose dynamic output impedance is not usually known or specified.

Having identified the low-pass (LPF) and high-pass (HPF) prototypes, the next step is to find the L and C values of the final LPF section:

$$L_{LP} = \frac{K\, L_{P(LP)} R}{2\pi\, f_{co}} \qquad C_{LP} = \frac{K\, C_{P(LP)}}{2\pi\, f_{co} R} \qquad \text{(Eq 1)}$$

where $LP_{(LP)}$ and $CP_{(LP)}$ are the prototype LPF values in Fig 7, $K = 1.005$, $R = 50\ \Omega$, and f_{co} is the cutoff (crossover) frequency in Fig 5 (5.45 MHz in this example). For the final HPF section:

$$L_{HP} = \frac{L_{P(HP)} R}{2\pi\, f_{co} K} \qquad C_{HP} = \frac{C_{P(HP)}}{2\pi\, f_{co} K\, R} \qquad \text{(Eq 2)}$$

Where $L_{P(HP)}$ and $C_{P(HP)}$ are the HPF prototype values in Fig 7. An important decision is the choice of f_{co} (see Fig 5). It must be such that the desired ham band is well within the passband of the LPF. The response of the HPF should be down at least 20 dB so that the dump resistor does not waste a lot of desired signal power, eg, 1.2 W for a 120-W PA. The response of the LPF must be adequate at the second and third harmonics. Some experimentation, using *Radio Designer* simulation, is very helpful for this.

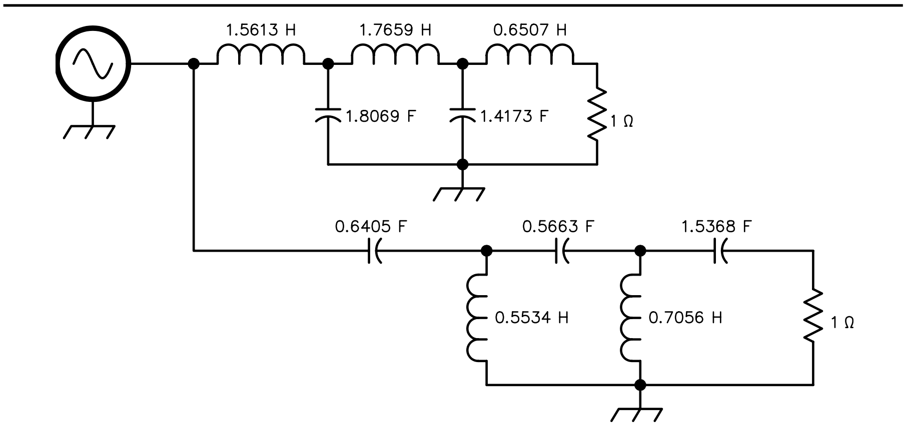


Fig 7—Low-pass and high-pass prototype diplexer filter.

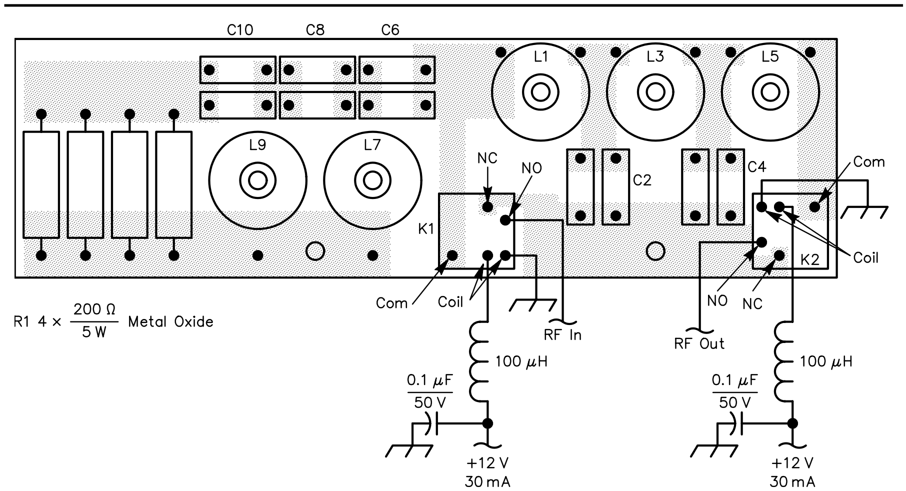


Fig 8—Diplexer PC-board layout.

Note that the shape of Fig 5 is the same at any part of the HF spectrum, and only f_{co} moves. I will suggest values of f_{co} for the six HF diplexers. Especially noteworthy in Fig 5 is the way that the LPF and HPF collaborate to maintain an almost perfect 50-Ω input resistance within the 3.5 to 4.0 MHz band. They do this by sending to the dump resistor the small amount of power that would otherwise be "reflected" by the LPF. Another item of interest is that the inductors—especially the first—in the LPF may become parallel resonant at some high frequency, but the HPF bypasses quite well any problems that this might cause for the PA.

The five-element low-pass prototype was selected as a compromise between complexity, cost and performance. The harmonic attenuation and the power dissipated in the dump resistor, including a small amount of fundamental, have proved to be quite reasonable, in my opinion. However, to accommodate the 40- and 30-meter bands with a single filter, a few extra watts of dissipation on 30 meters had to be accepted.

Diplexer Construction

Fig 8 shows the PC-board layout of a diplexer filter. It is built on a 2×8 3/8-inch one-sided PC board. The copper, shown as shaded areas, is on the opposite side, and the components are all on the near side. The LPF is on the right, and the HPF is on the left. The four resistors are 200-Ω, 5-W, 5% metal-oxide types that have low L and C and excellent stability. Twenty watts is overkill for these resistors, but I believe that it was a good decision that will provide plenty of safety margin. The two relays are Radio Shack 275-248 with 10-A contacts and measured stray C and L values that are plenty small enough for this application. They work quite well, but should not be "hot-switched" to assure long life. Space is provided for two capacitors in parallel at each location so that the correct C values can be closely approximated. The HPF and LPF ground surfaces are separated on the board and connected to the chassis to minimize cross talk, which can distort the frequency response and the RF IN return loss. The very short RF IN/OUT wires and the 12-V wires are brought out through small holes in the chassis.

Fig 9 shows the complete assembly of the experimental model. Each filter is mounted to the chassis with two 3/8-inch lengths of aluminum angle stock and #4 hardware. To prevent cross talk between filters, maintain the distance between them as shown. The band-select toggle switches can be replaced by programmable switching that sources +12 V at 60 mA. The RF INs and RF OUTs are joined together underneath the chassis with short lengths of 50-Ω miniature coax. Each of these lengths is grounded at both ends to provide a uniform 50-Ω Z_0. The 160-meter filter is closest to the BNC connectors, and the 12/10-meter filter is at the far end. That way the 12/10-meter filter does not have open-circuited stubs appended that could cause complications. Because the coaxes are in segments and have small values of stray L and C at the filter connection points, I found—experimentally—that a 10-pF capacitor across each BNC connector improved the return loss in the 10 to 60-MHz region. Each filter has an insertion loss of 0.2 to 0.3 dB in its ham band(s). The set of six identical PC boards is available from FAR Circuits.[7]

Test and Tweak

As one might expect, the actual filters to do not conform exactly to the idealized computer example of Fig 5 because of inaccuracies in coil and capacitor measurements, stray Ls and Cs, lead lengths, and so on. When the stray Cs of the coils—especially the HPF coils—are included in the simulation, the return loss looks much like Fig 10. That figure is a spectrum analyzer photo that shows the composite LPF, HPF and return loss (RL) for the 40/30-meter filter. The problem is that the L values must be correct at the crossover frequency f_{co}, but their "effective" values at much higher frequencies are a little larger because of their stray Cs. So to get adequate return loss—and therefore excellent LPF and HPF behavior—some experimental tweaking was necessary. To do this, I used the setup in Fig 11A, which shows a high-quality spectrum analyzer with built-in tracking generator and a dual directional coupler. Connecting lead A to point B sets a reference at the top of the screen. Connecting lead A to point C then registers the return loss, in decibels, on a 5-dB/div scale. The procedure is to tweak the Ls and Cs to get close to the response in Fig 10. It is important that the load resistor be an accurate 50 Ω up to 60 MHz.

The complete and detailed data sheets for the six filters that I have

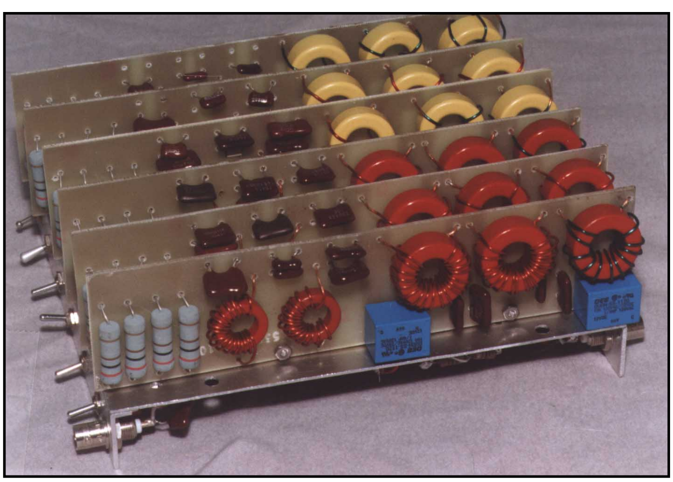


Fig 9—Photo of complete diplexer filter assembly for all nine HF amateur bands.

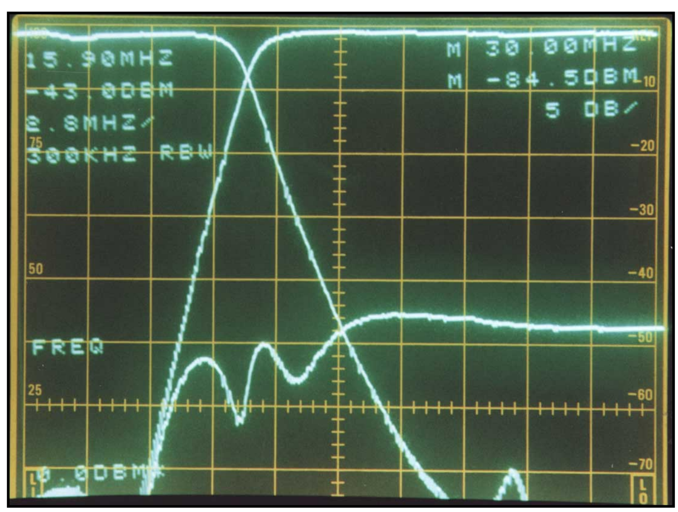


Fig 10—Composite spectrum analyzer plot, 1.8 to 30 MHz, of LPF response, HPF response and return loss. The scale is 5 dB/div, and the filters are for the 40/30-meter band.

prepared[8] give the exact measured L and C values and the coil details at which I arrived using the tweak procedure. Capacitors should be measured with a good digital C meter (better than 2% accuracy), and the meter readings should be within 2% of the desired values. The 500-V dipped mica capacitors, types CMO5 and CMO6, are recommended because of their very low dissipation factor 1/Q and small size. Inductor values can be closely approximated using the cores and winding instructions given in the data-sheet package. It will most likely be necessary to do some additional spreading or compressing of the toroid coil turns to fine tune the return-loss values. On the 160-, 80- and 40/30-meter filters, small values of capacitance were placed across the output inductor (L5 in Fig 8) that improved third-harmonic reduction by 4 to 5 dB, with no degradations in diplexer performance.

Observing return loss is a very sensitive indicator of coil adjustment. The spectrum analyzer photos in the data sheets also indicate certain important frequencies (as indicated by "•") that can be tweaked individually with a signal generator and a receiver as shown in the setup of Fig 11B. This method could use the harmonics of a 100-kHz or 1-MHz oscillator as a signal source. Establish a reference using point B. Then, connect to point C and reduce the attenuator from 25 dB to 0 dB in a ham band, or from 20 dB to 0 dB otherwise and work to get the same reference level to the receiver. The receiver must be in AGC-Off mode and must be operating linearly in that mode. Turn down the RF gain. Work back and forth over the frequency range until the best result is obtained. A little experience will show that this goes fairly smoothly.

Fig 12 shows two homebrew devices that aid in the testing and tweaking procedures. At the bottom is a high-impedance probe that connects to a 50-Ω load. It is used to look at the LPF or HPF output, as shown in Figs 6 and 10. The seven 470-Ω resistors in series, mounted as shown in mid-air, greatly reduce stray capacitance to ground. At the coax end, place a 47-Ω resistor to ground, and connect the coax to a 50-Ω measuring device. The probe response is down 1 dB at 30 MHz. At the top of Fig 12 is a 50-Ω, 150-W load resistor consisting of five 250-Ω carborundum resistors in parallel. It is used in place of the four 5-W metal-oxide resistors when I want to sweep from 1.8 MHz to as high as 60 MHz at the full 120-W power to check for power-amplifier instabilities. This device has been very helpful, and it reassures that there are no hidden problems.

The responses shown on the data sheets, especially the return loss, have been verified to be sufficient for excellent and stable performance of my MOSFET PA. There is no need for perfectionism in order to get a satisfactory

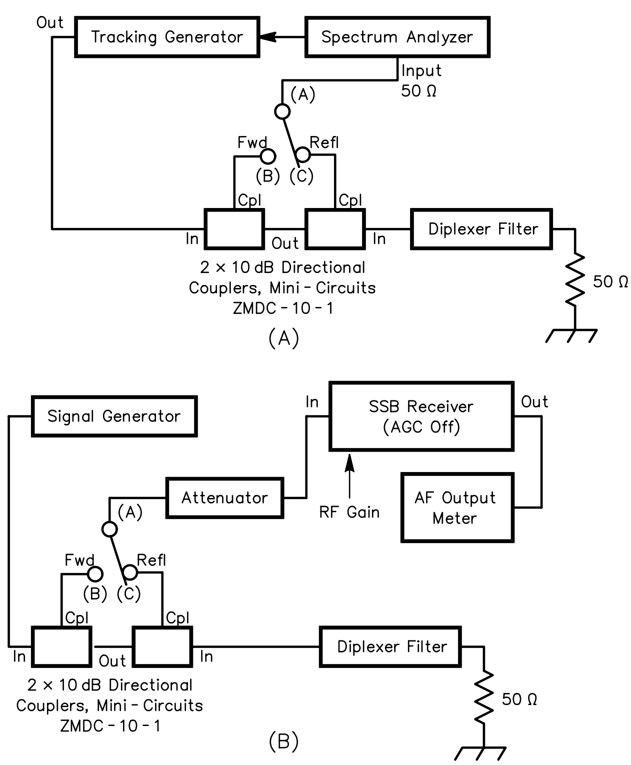


Fig 11—(A) Spectrum analyzer and tracking generator to measure and adjust diplexer return loss. (B) Signal generator and SSB receiver to measure adjust diplexer return loss.

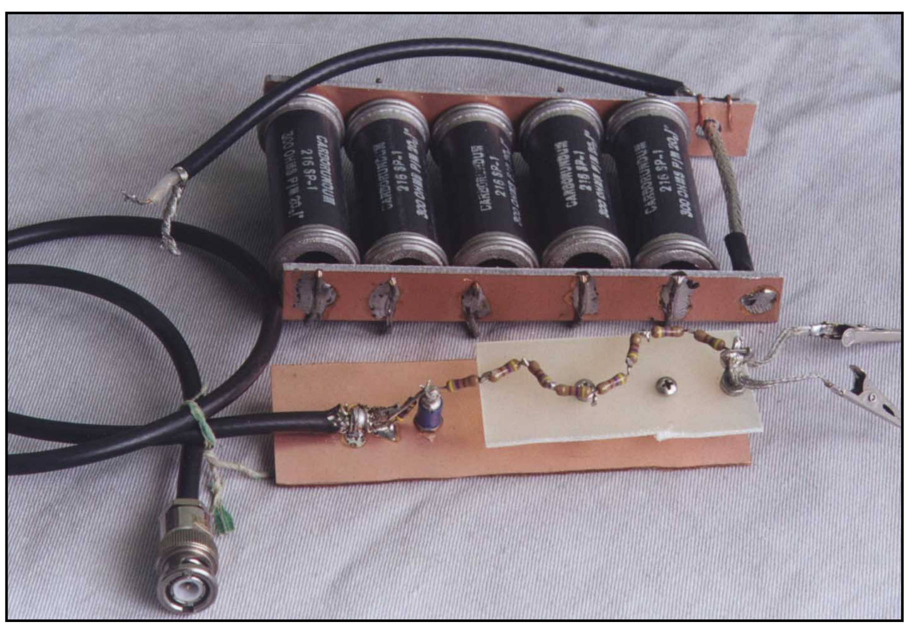


Fig 12—Two devices that help to test the diplexer filters. At the top is a 150-W dummy load that is mentioned in the text. At the bottom is a high-impedance probe for testing the LPF and HPF filter outputs.

result, but a return loss of 25 dB in the ham bands and 20 dB at other frequencies is a good goal to pursue. The HPF response should be down 20 dB or more at the upper end of the ham band. If carried to extremes, this tweak process can be rather tedious, and it is not at all necessary. If the gain of the power amplifier rolls off above 30 MHz, as it should, this helps render tweaking of the 12/10-meter filter (the most difficult one) much less critical, as shown on its data sheet.

Conclusions

The diplexer filter is larger, more expensive, a little more effort to build, and requires more tweaking than the usual cookbook-type LPF. The test setup in Fig 11B and the devices in Fig 12 are simple, inexpensive and very helpful in tweaking the filters and verifying their correct operation prior to connecting to the PA.

The motivation for the additional work is that it makes clean and spurious-free performance of the transistor power amplifier a lot easier to get. That has been my experience with my homebrew efforts, after a lot of experimentation using more conventional and rather frustrating approaches, and it is the reason for this article. Dye and Granberg (see Note 2) liked this approach when it was feasible. The home-basement-lab equipment that I used is of unusual quality for its environment, but the results can be duplicated with simpler gear using the data I have provided.

Notes

[1]W. E. Sabin and E. O. Schoenike, Editors, *Single Sideband Systems and Circuits* (New York: McGraw-Hill, 1995) and *HF Radio Systems and Circuits*, (Tucker, Georgia: Noble Publishing **http://www.noblepu.com**, 1998); Chapter 12 by Rod Blocksome. *HF Radio Systems and Circuits* is also available as ARRL Order No. 7253. ARRL publications are available from your local ARRL dealer or directly from the ARRL. Mail orders to Pub Sales Dept, ARRL, 225 Main St, Newington, CT 06111-1494. You can call us toll-free at tel 888-277-5289; fax your order to 860-594-0303; or send e-mail to **pubsales@arrl.org**. Check out the full ARRL publications line on the World Wide Web at **http://www.arrl.org/catalog/**.

[2]N. Dye and H.Granberg, *Radio Frequency Transistors, Principles and Applications,* p 151, Butterworth-Heinemann, 1993.

[3]A.Williams and F. Taylor, *Electronic Filter Design Handbook,* third edition, McGraw-Hill, 1995.

[4]A. Zverev, *Handbook of Filter Synthesis*, Wiley, 1967

[5]F. Methot, "Constant Impedance Bandpass and Diplexer Filters," *RF Design Magazine*, November 1986, pp 104-109.

[6]J. E. Storer, *Passive Network Synthesis,* McGraw-Hill 1957, pp 168-170. The constant-resistance diplexer is derived from a modified Darlington synthesis procedure (no transformers). The required high-pass section can be synthesized if a voltage source is assumed. This leads to the values in Fig 7 of this article. Ideally, the transfer properties are then independent of the actual generator resistance.

[7]A package of six PC boards is available from FAR Circuits for $30 plus $2 shipping. Single boards are $7.50 each. Contact Far Circuits, 18N640 Field Ct, Dundee, IL 60118; tel/fax 847-836-9148; **farcir@ais.net**; **http://www.cl.ais.net/farcir/**. Orders by mail and phone only.

[8]You can download a package of filter values and coil-winding instructions from the ARRL Web site (**http://www.arrl.org/files /qex/**). Look for the file SABIN799.ZIP. □□

Receiver Band-Pass Filters Having Maximum Attenuation in Adjacent Bands

Third-order Cauer filters can boost performance of multi-transmitter, multi-operator contest stations to the "next level." The filters are practical and you don't need expensive test equipment to align them.

By Ed Wetherhold, W3NQN
ARRL Technical Advisor

In a recent *QST* article,[1] I explained how to design, assemble and test six three-resonator band-pass filters (BPFs) for attenuating the phase noise and harmonics of the typical 150-W transceiver, both on transmitting and receiving. In this article, I will explain how to design, assemble and tune smaller-sized four-resonator BPFs having maximum attenuation in the two ham bands adjacent to the band being received. The BPFs are intended for connection to the 50-Ω RF input terminals of a receiver. They are especially useful for the multi-multi contester, where six receivers and six high-power transmitters are in simultaneous operation, and the receivers need preselection filtering to prevent front-end overload.

Several receiving band-pass filter designs are in current use by the multi-multi contest fraternity, but they are either difficult to assemble, have insufficient attenuation, or lack design information so the interested reader can confirm the correctness of the design or try a different design. For example, an article in *CQ CONTEST Magazine*[2] described a group of band-pass filters for the multi-multi operator station. Although the BPFs had exceptional stop-band attenuation (on the order of 80 dB in adjacent bands), the number of components (seven inductors and seven capacitors) was more than really needed, and the construction was difficult. The author made a passing reference to Zverev's *Handbook of Filter Synthesis*[3] as a source of the designs, but no explanation of the design procedure was included; consequently, none of the BPF designs could be confirmed. Other receiver BPF designs used over the past 15 years by the better-known multi-multi operators used a capacitively coupled three-resonator design with capacitor input and output. The attenuation of low-frequency signals was very good because of the capacitive coupling, but the high-frequency performance was poor. In addition, the tuning procedure was difficult unless you used a network analyzer.

In comparison, the new four-resonator receiver BPFs described below

[1]Notes appear on page 33.

1426 Catlyn Pl
Annapolis, MD 21401
tel 410-268-0916

need only four inductors and four capacitors for each BPF. Preliminary tuning of the four resonators requires a signal generator and detector, with final tuning using the return-loss test described in a previous article.[4] Stop-band attenuation of 60 to 80 dB is obtained in the center of the bands adjacent to the passband. The four inductors and capacitors of each BPF can be mounted on a piece of 1×1½-inch perfboard and installed in a 2⅛×1⅝×3½-inch aluminum Minibox. The design procedure is fully explained. Anyone having a computer can duplicate the designs and confirm the correctness of each design by means of free software that is available.

Whether you want to update your receiver BPFs for better selectivity, or design different BPFs, this article will show you how to do it.

Background

Some previously published designs used two or three top-coupled resonators, such as the N1AL designs,[5] or the W3LPL designs.[6] K4VX used three-resonator Butterworth designs for his BPFs,[7] and the most-selective BPFs by N6AW used seven resonators in a series-parallel configuration (See Note 2). I elected to base my BPF designs on the four-resonator, third-order Cauer. The input and output shunt resonators are tuned to the center frequency of the passband and the two series-connected resonators are tuned to the center frequencies of the adjacent bands. The intent is to have one more resonator than used in the simpler designs while getting maximum attenuation in the adjacent bands by having two of the resonators tuned to the frequencies where maximum attenuation is needed.

Although the stop-band attenuation of the third-order Cauer may be less than that of the N6AW seven-resonator design, the less-complex Cauer has less passband insertion loss, and is easier to assemble and tune. The design procedure to be explained shows how to confirm each BPF design and how to calculate other designs having different center frequencies or bandwidths. The computer used for designing needs only a DOS operating system. The computer I used has a 386SX microprocessor operating at 20 MHz with MS-DOS Ver. 4.01.

In addition to a computer, you need filter-design and analysis software. Normally, such software would cost more than $100, but for you to design, analyze and plot the responses of any third-order filter, the software costs nothing! This unusual offer to the Amateur Radio fraternity is made by Jim Tonne, president of Trinity Software.[8] Jim's intent is that those seriously interested in filter design and analysis—either for amateur or professional purposes—should have the opportunity to become familiar with his *ELSIE* (LC) filter-design and analysis software. He is therefore offering a demo disk of his DOS-based *ELSIE* software to anyone who asks. Although the program on the demo disk is limited to filters of the third order only, all options of *ELSIE* are available for use. These include plots and tables of all parameters, *ELSIE* can design filters, and tune the designs.

Those interested in either duplicating these third-order Cauer BPF designs or designing other third-order BPFs for different bands may obtain *ELSIE* software on a 3½-inch floppy disk by writing to Jim. In your letter, please include a description of your intended application and your filter-design background.

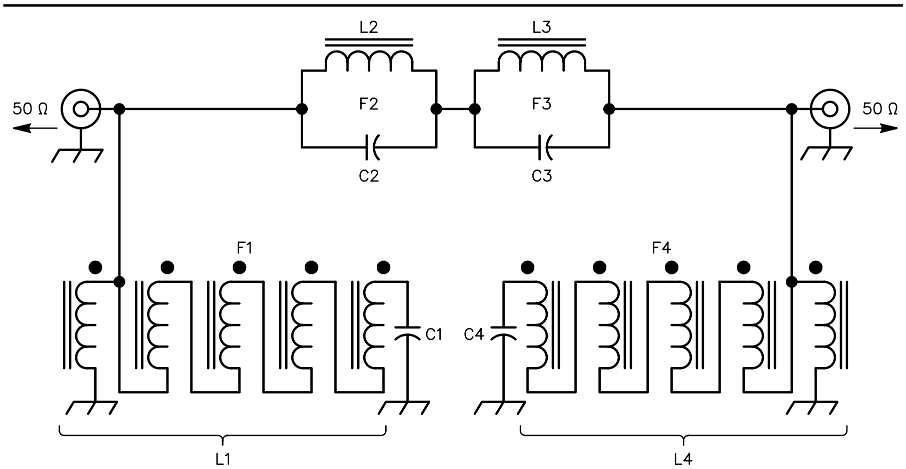
BPF Design and Confirmation Procedure

The design of these third-order Cauer BPFs involves discovering the optimum values of many parameters, such as passband and stop-band widths, center frequency, stop-band attenuation, passband return loss, and impedances of the input and output resonators. Finding the optimum values of all these would have been impossible without the help of *ELSIE*. My *ELSIE* designs for the 160, 80, 40, 20 and 15-meter BPFs are shown in Table 1. With these data, you can assemble and tune a set of BPFs with confidence.

Fig 1 shows the schematic diagram and the L, C and frequency values of the 40-meter BPF. As specified in rows 4, 5 and 6 of the 40-meter data in Table 1, inductors L1 and L4 each have five quintifilar turns of #18 and #20 magnet wire wound on T94-6 powdered-iron cores. A tap at the fifth turn above ground serves as the input and output connection to a 50-Ω source and load. The other BPFs are wired in a similar manner, except the 160 and 80-meter BPFs use quadrifilar windings for L1 and L4 instead of quintifilar windings.

Using quadrifilar or quintifilar windings on L1 and L4 results in an interleaving of all turns, with a correspondingly greater coupling between turns than that obtained with the more-customary single continuous winding. The inter-winding coupling reduces leakage inductance while optimizing the filter performance. This same winding technique was used in the wiring of the input and output resonators in the transmit BPFs discussed in an earlier article (see Note 1).

It was necessary to connect resonators 2 and 3 of the BPFs to taps on L1 and L4 at ¼ or ⅕ of the total turns so that the component values of resonators 2 and 3 would be practical. For example, the inductive reactances of L2 and L3 in the 40-meter BPF design are 413 Ω and 391 Ω at 14.287 and 3.734 MHz, respectively. These reasonable reactances can be achieved



C1, 4 = 100 pF L1, L4 = 4.926 µH F1, F4 = 7.17 MHz
C2 = 27 pF L2 = 4.60 µH F2 = 14.29 MHz
C3 = 110 pF L3 = 16.5 µH F3 = 3.734 MHz

Fig 1—Schematic diagram and component values of the 40-meter receiver band-pass filter. The diagram is representative of all receiver BPFs, except for the 160 and 80-meter BPFs, which have quadrifilar windings for L1 and L4. See Table 1 for the component values and coil-winding details.

Table 1—Design Parameters For 160, 80, 40, 20, 15-Meter Receiver Band-Pass Filters

Parameters	*160-Meter (1.8 to 2.0)*	*80-Meter (3.5 to 4.0)*	*40-Meter (7.0 to 7.3)*	*20-Meter (14.0 to 14.4)*	*15-Meter (21.0 to 21.45)*
F_c, BW, Stop-Band Width (MHz)	1.897, 0.222, 2.22	3.74, 0.82385, 5.2256	7.17, 0.779, 10.127	14.2, 0.7137, 11.6377	21.2, 1.83, 19.484
As (dB), RL (dB), Z (Ω)	60.2, 23.84, 800	52.0, 20.10, 800	64.0, 26.89, 1250	64.0, 32.8, 1250	60.0, 25.67, 1250
L1, L4 (μH); Qu & XL (Ω) @ F_c	11.17, 120, 133	8.965, 122, 211	4.926, 200, 222	1.570, 150, 140	1.281, 120, 171
Core & A_L (nH/N^2)	T94-2 (Red), 8.4	T94-2 (Red), 8.4	T94-6 (Yel), 7.0	T94-17 (Blu/Yel), 2.9	T80-17 (Blu/Yel), 2.9
Turns	9 Quadrifilar	8 Quadrifilar	5 Quintifilar	4 Quintifilar	4 Quintifilar
Turns, Wire Length & AWG	36: 9T, 13" #18; 27T, 32" #20	32: 8T, 12" #18; 24T, 29" #22	25: 5T, 8.3" #18; 20T, 26" #20	20: 4T, 7" #18 16T, 21" #18	20: 4T, 5" #18 16T,16" #20
L2 (μH), F2 (MHz)	14.17, 3.60	4.72, 7.085	4.596, 14.287	2.11, 21.1	0.9228, 28.84
XL (Ω) @ F2 (MHz), Qu	321, 210	210, 240	413, 80	280, 150	167, 200
Core (Color) & A_L	T94-2 (Red), 8.4	T94-6 (Yel), 7.0	T94-6 (Yel), 7.0	T94-17 (Blu/Yel), 2.9	T80-17 (Blu/Yel), 2.2
Turns, Length & AWG	40T, 45", #22	26T, 29", #20	24T, 28" #21	25T, 29", #20	18T, 20" #18
L3 (μH), F3 (MHz)	53.39, 0.984	19.4, 1.880	16.51, 3.734	11.0, 7.23	2.46, 14.21
XL (Ω) @ F3 (MHz), Qu	330, 170	229, 250	387, 200	500, 140	220, 115
Core (Color) & A_L	T94-2 (Red), 8.4	T94-2 (Red), 8.4	T94-2 (Red), 8.4	T94-2 (Red), 8.4	T80-6 (Yel), 7.0
Turns, Length & AWG	38.5 Bifilar, 45",#24	47T (2-layer), 56",#20	44T, 49" #23	35T, 41", #20	22T, 24" #18
C1, C4 (pF)	630 = 620 + 10	202 = 180 + 22	100	80 = 33 + 47	44 = 22 + 22
C2 (pF)	138 = 82 + 56	107 = 68 + 39	27	27	33
C3 (pF)	490 = 470 (+20 interwinding)	370 = 270 + 100	110 = 100 + 10	44 = 22 + 22	51 = 33 + 18

TABLE NOTES:

1. The first two rows list *ELSIE* parameters of center frequency, ripple bandwidth, bandwidth between upper and lower stop-band frequencies, attenuation depth in the stop band, minimum passband return loss and the impedance level of resonators 1 and 4, respectively. See the article text for explanation of the 160 meter C3 capacitance value of 490 pF.
2. Most of the capacitors are obtained from the PHILIPS 680 Series because of its low K (high Q), 2% tolerance and 100-V dc rating. See the FARNELL/NEWARK electronic components catalog (March/September 1998, p 62). The 630-pF value of C1 and C4 in the 160-meter BPF design is realized with a paralleled 620-pF dipped silver-mica cap and a Philips ceramic cap, both selected to realize the design value within one percent. The 620-pF, 5% mica capacitor is available from Hosfelt Electronics, 2700 Sunset Blvd, Steubenville, OH 43952; tel 800-524-6464, fax 800-524-5414; **hosfelt@clover.net**; **http://www.hosfelt.com/**.
3. MICROMETALS cores (for RF applications) are used in all the BPFs.
4. The odd-numbered wire sizes are AWG equivalents of SWG wire sizes obtained from FARNELL/NEWARK. See the March/September 1998 FARNELL catalog, p 865.
5. The minimum return-loss values listed above were obtained from the computer-generated data of the *ELSIE* filter design/analysis software. The return loss of the assembled BPF as measured with a network analyzer may be different.
6. For frequencies less than 3 MHz, resistive considerations outweigh capacitive considerations; consequently, multiple-layer windings are acceptable to reduce resistive losses and improve Q. Above 3 MHz, single-layer windings provide maximum Q.

with toroidal powdered-iron cores. If these resonators had been connected to the tops of resonators 1 and 4, the reactances would have been impractical at 25 times greater; that is, at 10.3 kΩ and 9.7 kΩ.

If resonators 1 and 4 had been designed for an impedance of 50 Ω at the start, this would have eliminated the need for taps, but then the inductance and reactance of L1 and L4 would have been much too low at 0.197 µH and 8.88 Ω to realize these inductance values with acceptable Q. The procedure to obtain optimum component values for all resonators is to design resonators 1 and 4 for an impedance equal to the square of 2, 3, 4 or 5 times 50 Ω, and then to connect the center resonators between 50-Ω taps on L1 and L4.

Whether a quadrifilar or quintifilar winding is used for L1 and L4 depends on the BPF percentage bandwidth. For example, the percentage bandwidth of the 80-meter BPF is $(100 \times BW)/F_c$ = 82.385 / 3.74 = 22%. This is a relatively broad bandwidth, and a quadrifilar winding is satisfactory. In comparison, the percentage bandwidths of the 40, 20 and 15-meter BPFs are 11.3%, 5.0% and 6.8%, and quintifilar windings are more appropriate. The 160-meter BPF has a relative percentage bandwidth of 11.7%, and either quadrifilar or quintifilar windings could be used.

The BPF designs listed in Table 1 may be confirmed in two ways. The simplest way is to use the analysis option of *ELSIE* wherein the listed component values are entered at the *ELSIE* prompts, and the insertion loss and return-loss response plots are viewed to confirm that the design is satisfactory. However, a minor correction to the tabular data for resonators 2 and 3 must be made before entering their component values at the *ELSIE* prompts because *ELSIE* is not capable of evaluating tapped inductors. Consequently, the two series-connected resonators 2 and 3 must be moved to the tops of resonators 1 and 4, and the component values of resonators 2 and 3 corrected to account for the change in impedance level. This is accomplished by multiplying and dividing the tabular inductance and capacitance values, respectively, of resonators 2 and 3 by a factor equal to the impedance of resonators 1 and 4 divided by 50, or 1250 / 50 = 25. Fig 2 shows the schematic diagram and component values of the 40-meter BPF in a corrected form suitable for *ELSIE* to analyze the design and plot the attenuation and return-loss responses.

The attenuation peaks should fall in the center of the 80 and 20-meter bands, and the minimum passband return loss should be 30 dB.

The BPF designs may also be confirmed by letting *ELSIE* assist you in designing a BPF. At the *ELSIE* prompts, enter the width of the passband, the center frequency, the stop-band width, the depth of the stop-band attenuation and the impedance. *ELSIE* will then design a BPF to meet these requirements. The design values to use are listed in the first two rows of each column in Table 1, except for the passband return loss, which is not needed by *ELSIE*, and is included only for reference.

After reviewing the attenuation and return-loss response plots, you then manually tune the design so the attenuation peaks fall at the center of the bands adjacent to the passband. This is done by varying the C and L values of resonators 2 and 3 while maintaining a passband return loss greater than 20 dB. Additional minor adjustments can be made to the center frequency so that the values of C1 and C4 are convenient. For example, the center frequency of the 40-meter BPF was increased slightly from 7.15 MHz to 7.17 MHz so C1 and C4 would become exactly 100 pF instead of the original nonstandard value.

When you are satisfied with the tuned design, the impractical component values of resonators 2 and 3 are scaled from the design impedance to 50 Ω. The 50-Ω taps on resonators 1 and 4 serve as the BPF input and output connections. Fig 1 shows the schematic diagram of the completed design of the 40-meter BPF.

The third-order BPFs designed by *ELSIE* originated as classic Cauer designs, where the minimum stop-band attenuation both below and above the passband are identical. However, after the modifications, the lower and upper-frequency minimum-attenuation levels are no longer identical, thus showing that the design is no longer a legitimate Cauer. For this reason, these modified Cauer designs cannot be duplicated using the published Zverev tables. For our purposes, this is of no concern as long as the attenuation peaks are in the center of the adjacent ham bands, and the computer-calculated minimum passband return loss is greater than 20 dB. By using *ELSIE* to design these third-order Cauers, what before was impossible now becomes simple!

BPF Assembly and Tuning

Fig 3 shows the 40-meter BPF assembled on a piece of perfboard installed in an LMB 873 aluminum Minibox. The toroidal inductors are secured to the perfboard by passing their leads through the holes in the perfboard, then sharply bending the leads sideways. All capacitors are connected to the inductor leads under the perfboard. A cardboard strip insulates the capacitor and inductor leads (under the perfboard) from the bottom of the aluminum box. The #18 wire leads of L1 and L4 connect at each end of the assembly to the center pins and ground lugs of the phono connectors. These four #18 leads provide sufficient support to hold the assembly in place. The other BPFs are assembled in a similar manner.

The assembly of the BPF components is greatly simplified by the omission of shielding partitions between stages. The lack of any shielding apparently had no effect on the BPF stop-band performance, since attenuation levels greater than 80 dB were noted in the upper frequencies in all the BPF tests.

Because resonators 1 and 4 must be

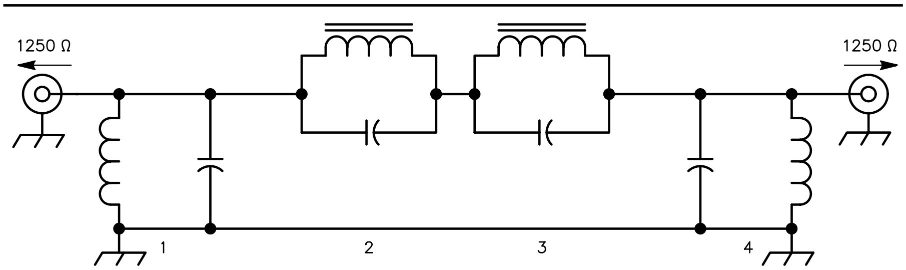


Fig 2—Schematic diagram of the prototype third-order Cauer 40-meter BPF before resonators 2 and 3 are moved to the 50-Ω taps on L1 and L4. Use these component values if you want *ELSIE* to analyze the BPF performance.

tuned to the same center frequency, successful tuning depends on using precisely matched capacitors, preferably both having the same value, and within one percent of the design value. For the 40-meter BPF, this frequency was 7.17 MHz, so C1 and C4 could be standard values of 100 pF. Capacitors 2 and 3 can be within two percent of the design values. Sufficient room should be left on the T94 cores so the windings can be squeezed or spread to fine tune each resonator. This is important so that all resonators can be tuned either to the center of the BPF passband, or to the center frequency of the adjacent amateur bands.

Initially, tune each resonator before installation on the perfboard. First, pass a single-turn wire loop from a signal generator through the center of the inductor. Then put a second loop through the inductor, and connect it to a sensitive wide-band detector. Vary the signal-generator frequency until you see a voltage peak on the detector output meter; that indicates circuit resonance. Measure the generator frequency with a frequency counter, then squeeze or spread the inductor turns until a resonance peak is obtained at the design frequency. After this, install the resonator on the perfboard without disturbing the turns on the core. A final check on the BPF tuning is made with the return-loss response test as explained in the referent of Note 4. After the final check, the inductor turns may be secured with a coating of polystyrene Q-dope.[9]

Special Tuning Considerations

To find the optimum parameters for the 160-meter BPF, I used *ELSIE's* "tune" mode to find convenient capacitance values while keeping the upper-frequency attenuation peak in the center of the 80-meter band and while keeping the minimum passband return loss greater than 20 dB. The placement of the lower-frequency attenuation peak, established by resonator #3, was not critical, and the C3 and L3 values were varied until a convenient C3 value of 470 pF was obtained with a computer-calculated minimum return loss of more than 20 dB. However, when the design was assembled and final tuning adjusted by observing the measured return-loss response as seen on an oscilloscope, I discovered that the optimum return-loss response occurred when resonator #3 was tuned to 0.984 MHz, *not* to the original frequency. The measured return-loss response of the assembled BPF could be duplicated with an *ELSIE* analysis only when C3 was made equal to 490 pF instead of the 470-pF value, and F3 was 0.984 MHz.

The actual value of C3 is 20 pF greater than the 470-pF capacitor installed on the perfboard because of the inter-winding capacity of L3. Con-

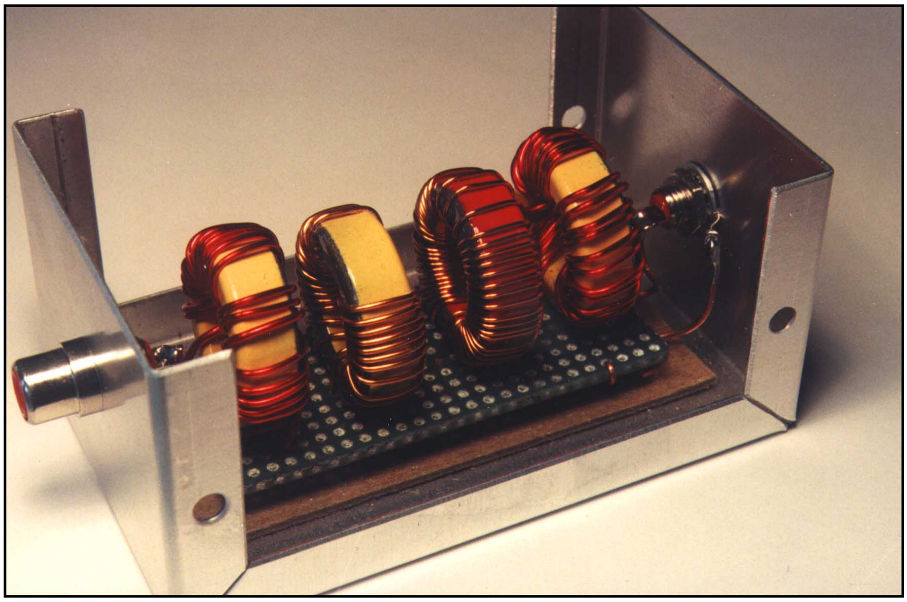


Fig 3—The photo shows the 40-meter BPF assembled in an aluminum Minibox $3^1/_2 \times 2^1/_8 \times 1^5/_8$ inches (LMB 873). The T94 cores are installed on the top of a piece of perfboard (1×2.6 inches) with the prepunched holes on a 0.1-inch grid. All capacitors are mounted under the perfboard. A strip of cardboard glued to the inside of the box bottom provides insulation between the BPF leads and the aluminum box. The BPF input and output leads that are connected at each end to the phono-connector center pins and ground lugs are stiff enough to hold the assembly in place.

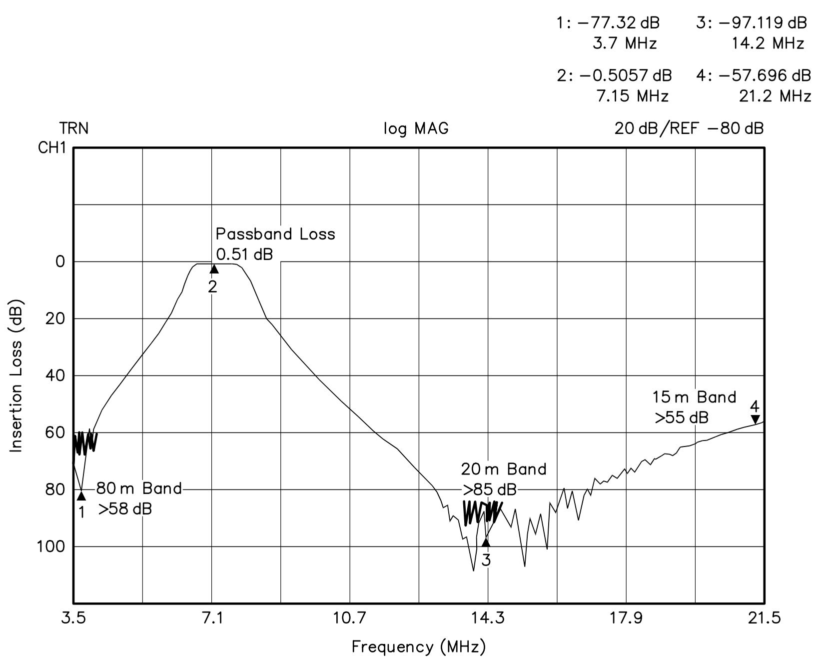


Fig 4—The plot shows the insertion-loss response of the 40-meter BPF as measured with a network analyzer. The attenuation of signals in the adjacent 80 and 20-meter bands is greater than 58 and 85 dB, respectively. The passband loss is about 0.5 dB and the passband return loss (not shown) is greater than 20 dB.

sequently, when assembling the 160-meter BPF, use a 470-pF capacitor for C3; when analyzing the design with *ELSIE*, use a 490-pF value. The additional 20 pF caused by the L3 winding capacitance is indicated by the designation "+20 interwinding" in the 160-meter BPF column for C3 in Table 1.

Because of the unknown effects of stray variables associated with these BPFs, it is important to final tune each BPF using the return-loss-measurement procedure so you have assurance that the BPF is correctly tuned.

Insertion Loss and Return-Loss Performance

Figs 4 and 5 show the measured responses of the 40-meter BPF insertion loss and return loss after the BPF tuning was completed. The insertion-loss response, obtained with a network analyzer and plotter by Tim Duffy, K3LR, shows that the stop-band attenuation is maximum at 75 and 85 dB at the centers of the 80 and 20-meter bands, respectively. The passband loss is about 0.5 dB. The stop-band and passband losses of the other BPFs are similar to those of the 40-meter BPF, with the greatest passband loss being 0.55 dB in the 15-meter BPF.

The decrease in insertion loss above 15 MHz is attributed to imperfect coupling in the windings of L1 and L4, and this same anomaly was noted in the transmitter BPFs described in the referent of Note 1. Although the insertion loss decreases above 15 MHz, this should cause no problem because the attenuation in the 15-meter band is still substantial, at more than 55 dB.

The return-loss response of the 40-meter BPF is shown in Fig 5. This figure is a photograph of the scope waveform obtained with the return-loss test setup described in the referent of Note 1. Two important additions to the equipment shown in Note 1's Fig A3 should be noted, however. (1) A seven-element, 50-Ω low-pass filter with a cutoff frequency equal to about 1.3 times the upper cut-off frequency of the BPF being tested should be connected directly to the 50-Ω output of the voltage-controlled oscillator (VCO). (2) A 50-Ω, 6-to-10-dB pad should be connected between the low-pass filter output and the RF-IN port of the return-loss bridge.

The 50-Ω pad provides a well-defined 50-Ω source impedance for the return-loss bridge and eliminates any minor impedance variations that might be present at the filter output port. The low-pass filter prevents harmonics of the VCO from distorting the waveform of the return-loss response. If the VCO harmonics are not sufficiently attenuated, the sharp peaks of the return-loss response may be completely missing, and it may be difficult to determine when the BPF is correctly tuned. BPF tuning is accomplished by squeezing or spreading the turns on inductors L2 and L3 until the three peaks are obtained on the return-loss response and the minimum return-loss levels are identical. Sometimes it may be necessary to remove or add a turn to L2 or L3. You can see whether this is necessary by the effect on the return-loss response when squeezing or spreading the turns of L2 and L3. Do not touch the turns on L1 and L4 since they need no adjustment.

Correct tuning of the BPF is indicated when you get three distinct peaks in the return-loss response and the two minimum return-loss levels between the peaks are identical. In the 40-meter BPF, the measured minimum return-loss level was a few decibels above 20. A 20-dB reference level was established on the oscilloscope screen by replacing the 50-Ω terminated BPF with a 61-Ω resistive load. A 61-Ω load on the return-loss-bridge "load" port produces a straight line on the display that is equivalent to an SWR of 61/50 = 1.22, which is equal to a return loss of 20 dB. For the 160 and 80-meter BPFs, the minimum return loss was a decibel or so below the 20-dB reference level.

The 40-meter passband width can be measured from the return-loss response by subtracting the lower frequency from the upper frequency on the return-loss response curve where the curve crosses the minimum return-loss level. In the case of the 40-meter BPF, the measured passband width was about 0.79 MHz, which closely approximates the design value listed in Table 1.

As viewed on a network analyzer or in the *ELSIE* plot of return loss, the return-loss response increases in a downward direction. However, I prefer to see return loss increasing upward.I accomplished this reversal of return-loss direction by using the scope **INVERT** switch on the Y-channel input.

160-meter BPF Performance Under Operating Conditions

An indication of the usefulness of these receiver BPFs under actual operating conditions was provided by Tony Kazmakites, N2TK/V26AK. He used both the transmit and receive BPFs in the operation of V26B during the 1998 CQ Worldwide Sideband Contest, in the multi-multi category. Tony reports that originally, with the old-style three-resonator capacitive-coupled BPF on the 160-meter re-

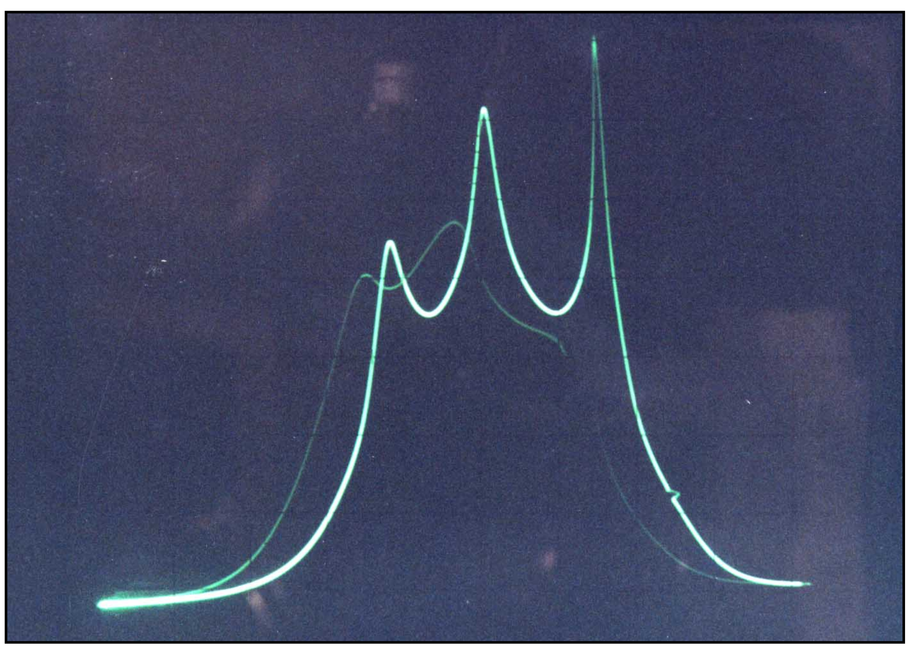


Fig 5—The photo shows the return-loss response of the 40-meter BPF obtained with the test equipment described in the text. The center peak of the return-loss response is at 7.17 MHz, which is the center frequency of the tuned 40-meter BPF. The frequencies at the beginning and end of the 40-meter passband (7.0 and 7.3 MHz) are to the left and right of the center peak in the valleys of the response curve. The BPF minimum return-loss level is greater than 20 dB.

ceiver, nothing on the 160-meter band could be heard because of noise from the 75-meter station. After the old-style BPF was replaced with the new third-order Cauer 160-meter receive BPF, the entire problem on 160 meters was solved. The 160-meter operator wanted to buy the filter on the spot! Tony further reports that "...the W3NQN transmit and receive filters have taken us to another level of improvement in multi-transmitter operations." Tony says he will be using these BPFs again at V26B for the 1999 CQ Worldwide Sideband Contest.

Summary

The deficiencies of LC BPFs currently being used to prevent receiver overload were discussed. A previously unused BPF type—the third-order Cauer—was introduced. The Cauer is easier to assemble and tune than other filter types, and it provides maximum attenuation in the adjacent amateur bands. This filter type was not previously considered because it was impossible to calculate the component values for the special stop-band response that was desired. However, free filter-design and analysis software makes it possible for anyone with a computer to design and analyze any type of third-order passive LC filter.

Those interested in only building new third-order Cauer BPFs, can do so from a table of values for the 160, 80, 40, 20 or 15-meter bands and the tuning procedure described. To demonstrate the performance typical of all the BPFs, the insertion and return loss of a 40-meter BPF was evaluated and its response curves are shown in two figures. An explanation was included for those who wish to confirm the tabulated designs or design BPFs having different parameters. Those wishing to obtain any of the assembled and tested BPFs should send an SASE (business-sized envelope) to the author for details.

Radio amateurs now have access to free LC filter software that allows them to design and analyze any type of third-order passive LC filter. This new and powerful capability should help to advance the state-of-the-art in Amateur Radio filter design.

Acknowledgements

The author is grateful to Jim Tonne, WB6BLD, for making his *ELSIE* filter-design and analysis software available for use in this project. Thanks also to Tim Duffy, K3LR, for his comments on the receiver BPF performance and for plotting the insertion and return-loss responses using a network analyzer. The report provided by Tony Kazmakites, N2TK, on his experiences in using the receiver BPFs is gratefully acknowledged.

Notes

[1]E. Wetherhold, W3NQN, "Clean Up Your Signals with Band-Pass Filters," *QST*, 1998, Part 1 May , pp 44-48 and Part 2 Jun, pp 39-42.

[2]J. Perkins, N6AW, "Band-pass Filters for the Serious Multi-Operator Station," *CQ Contest*, Jan 1996, pp 14-17.

[3]A. I. Zverev, *Handbook of Filter Synthesis* (New York: John Wiley and Sons, 1967).

[4]See Note 1: Appendix of Part 2, pp 41, 42 and Fig A3.

[5]A. Bloom, N1AL, "Inexpensive Interference Filters," *QST*, Jun 1994, pp 32-36.

[6]Description of W3LPL receiver band-pass filters from notes of K3ND, provided to the author by Tony Kazmakites, N2TK.

[7]L. Gordon, K4VX, "Band-Pass Filters for HF Transceivers," *QST* Sep 1988, pp 17-28.

[8]Trinity Software, 7801 Rice Dr, Rowlett, TX 75088, Jim Tonne, WB6BLD, President, tel 972-475-7132.

[9]Q-dope, 2 oz bottle with brush, part #10-3702, $3.75; Ocean State Electronics, PO Box 1458, 6 Industrial Dr, Westerly, RI 02891; tel 1-800-866-6626, 401-596-3080, fax 401-596-3590; **http://www.oselectronics.com/**.

Ed Wetherhold received a degree in Radio Engineering from Tri-State University, Angola, Indiana, in 1956. From 1962 to 1992, he was employed at the Annapolis Signal Analysis Center of Alliant Techsystems, Inc (Alliant Techsystems was formerly the Defense Division of Honeywell, Inc), as a communication systems test engineer and as a certified TEMPEST Professional Level II.

Ed has written many articles on simplified filter design that have been published in the electronics trade and Amateur Radio journals, such as Interference Technology Engineers' Master (ITEM), QST, QEX, CQ *and* Practical Wireless*, and in professional EMC journals and* Amateur Radio Handbooks. *For example,* The 1998 ARRL Handbook *contains Ed's SVC filter design tables and an explanation of how to design passive LC filters.*

Ed obtained his Amateur Radio license in 1947, while serving in the Air Force as a radio mechanic instructor at Scott AFB, in Illinois. Since 1977 he has been a Technical Advisor to the ARRL on passive LC filters. He may be contacted at his home at 1426 Catlyn Pl, Annapolis, MD 21401, or by telephone at 410-268-0916.

The Multiband Extended Double Zepp and Derivative Designs

W7SX's search for simple multiband antennas with bidirectional patterns continues. Here are several new examples to cover 14, 18, 21, 24, 28 and 50 MHz, 7, 10, 14 and 21 MHz, 21 and 50 MHz and a W8JK derivative for 14 through 50 MHz. Grab your rope and wire cutters!

By Robert J. Zavrel Jr, W7SX

The great reader response to "Multiple-Octave Bidirectional Wire Antennas" in the July/August 1998, issue of *QEX* prompted me to look more closely at the λ/4 stub technique. This stub provides both linear end loading on lower frequencies and an effective trap at the frequency where the stub is λ/4. The original antennas were built to provide multiband bidirectional operation from a single design. Two configurations were constructed: one for 80/40/20-meter operation and one for 40/30/20/10-meter operation. Further analysis has yielded some interesting interrelationships in this design. This paper reports these findings.

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Fig 1 shows the fundamental design of this antenna. The highest frequency (f_h) of operation (for a bidirectional pattern) is defined where L_i is chosen as an equivalent length for a double extended Zepp. The length of the end-shorted stub (L_s) is λ/4 at this same frequency. Therefore, the total antenna length is 1.75 λ (free space) at f_h. At f_h, there is a current maximum at the very ends of the antenna. However, the equal currents flowing in either side of the stub are out of phase, thus the radiation pattern contribution from these currents cancel. These current maximums therefore do not contribute to the pattern of the array at this frequency, where L_s = λ/4. Consequently, the only portion of the antenna that contributes to the pattern is the single-wire section, which corresponds to an extended double Zepp at f_h. In effect, a λ/4 shorted stub located at the end of an antenna element will act like a trap, with the effective trap location being the open end of the stub. This is true for any element configuration, and can be used in a variety of schemes. To this date, I have not found a configuration that will permit a multiband stub trap, or use of the stub anywhere but the end of an antenna element.

Above and below f_h, the currents do not cancel, and currents in the stub contribute to the radiation pattern of the antenna overall. This results in the "four-leaf" pattern, similar to that of simple long wires. As frequency is reduced further, the four-leaf lobes gradually decrease and bidirectional lobes gradually increase, until an extended double Zepp pattern returns. I define this as f_i, or the intermediate bidirectional frequency. The ratio of f_h / f_i for wire antennas in the HF range is about 1.61. Therefore, this antenna will exhibit the equivalent extended double Zepp gain and bidirectional pattern at two frequencies whose ratio is about 1.61.

As the frequency of operation is reduced below f_i, a bidirectional pattern is maintained, but with lesser broadside gain. This is exactly as we expect from a "normal" extended double Zepp. This gain gradually decreases to that of a simple dipole. I define this frequency as f_d. The ratio f_h / f_d is about 3.57.

Figs 2 and 3 plot gain versus f_h, f_i, and f_d for the Antennas 1 and 2. A 0 dBd (λ/2 dipole reference broadside gain) line is shown for convenience of comparison. All plots are free-space diagrams; the wire elements are #12 copper wire. Stub wire spacing is 0.1 feet, similar to the standard 450-Ω ladder-line spacing that was used in the original arrays. Two graphs show the relationship between broadside gain and frequency for the simple, in-line antenna elements.

Modeling Notes

I used *EZNEC* Version 2.0 for all these configurations. I noticed that the antenna patterns and feed-point impedances changed radically when changing the number of segments used in the models. For more-accurate results, Roy Lewallen, W7EL, suggests using *many* more segments than would normally be used in arrays of these dimensions.[1] See the sidebar "Modeling Closely Spaced Wires." The current-canceling characteristics of the stub seem to be the main problem. In addition, the number and alignment of segments used in the λ/4 stub wire should be parallel with the segments in the opposite wire—the respective portions of L_t in Fig 1. (That is, the numbers of segments in the upper and lower wires of the stub should be equal, and the segment-end points should align vertically. See Fig 1 inset—Ed.)

I used five wires to model these arrays in *EZNEC*. Wire 1 corresponds to L_t in Fig 1. Wires 2 and 3 represent the 0.1-foot shorting wires at the antenna's ends (use one segment each for these short wires). Wires 4 and 5 represent the parallel stub wires. The number of segments used is listed for both L_t and L_s in the two sample antennas. The points on Figs 2 and 3 are actual broadside gain numbers taken from the numerous simulation runs.

Some Practical Single-Wire Designs

A quick analysis of the dual-peak broadside-gain characteristics opens some very interesting possibilities for amateur multiband applications. Three separate wire-antenna dimensions are presented for multiband use:

Antenna 1: 14/18/21/25/28/50 MHz

L_t = 34.4 ft
L_s = 4.9 ft
f_h = 50 MHz
f_i = 31 MHz
f_d = 14 MHz
85 segments used for L_t (Fig 1)
12 segments used for each L_s (Fig 1)

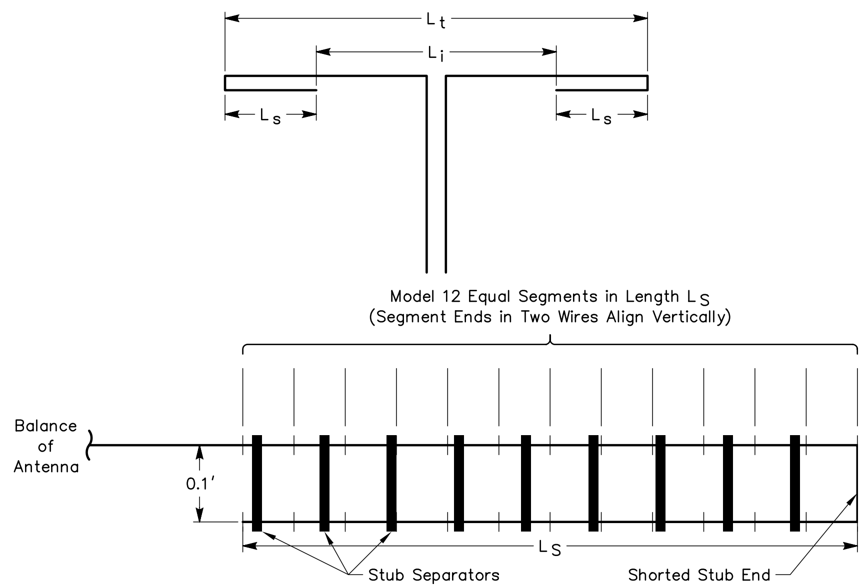

Fig 1—Antennas 1 and 2 are built from this plan. L_s is the length of the end-loading stubs, which act as shorted λ/4 transmission-line stubs at f_h. At f_h, L_i is 1.25 λ; L_s is λ/4; L_t is 1.75 λ. Inset shows details of stub model and construction. Dimensions for each antenna are given in the text. Drawing not to scale.

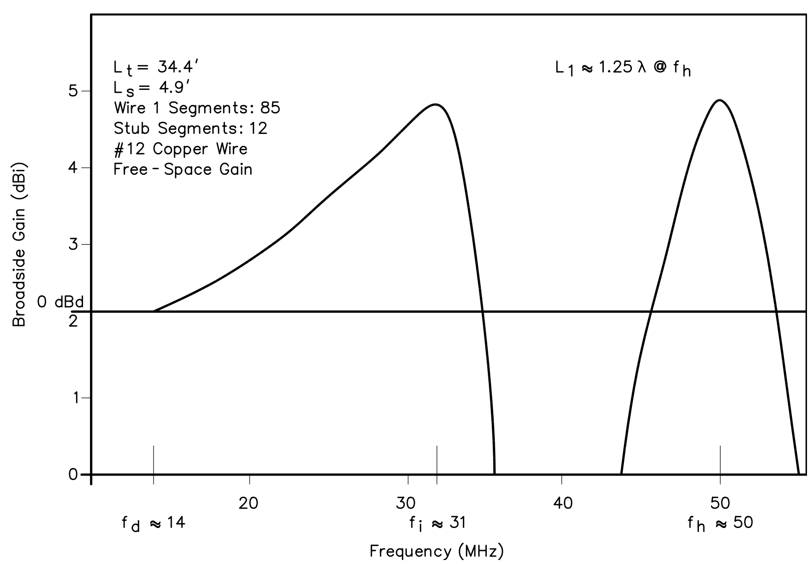

Fig 2—Antenna 1 broadside gain (dBi) versus frequency.

[1]Notes appear on page 39.

This antenna can provide significant bidirectional gain over six amateur bands, with the gain peaks appearing at six and just above 10 meters. (See Fig 4.) The actual ratio of 50.1/28.1 MHz is about 1.78, a bit more than the 1.61 for this antenna. However, the 10-meter gain is only about 0.5 dB below that of an extended double Zepp, and gain on all amateur frequencies above 14 and below 51 MHz is better than a dipole. This might be an attractive array for the sunspot-cycle peak, with optimum response on six and 10 meters.

Antenna 2: 7/10/14/21 MHz

L_t = 74.0 ft
L_s = 12.46 ft
f_h = 21 MHz
f_i = 14 MHz
f_d = 7 MHz
85 segments used for L_t (Fig 1)
14 segments used for each L_s (Fig 1)

This interesting antenna provides an extended-double-Zepp response at both the 14 and 21 MHz bands, as well as some broadside gain at 10.1 and 7 MHz. (See Fig 5.) At 18.1 MHz, the antenna pattern resembles a four-leaf long wire, the broadside gain being just about 0 dBi.

When the desired frequency ratio for the two gain peaks is less than 1.61, as is the case for 21.1 and 14.1 MHz, L_i can be decreased. However, I noticed that about 5 dBi gain could be maintained by a slight lengthening of L_s. By experimenting with exact lengths, the desired dual-peak frequency ratio of 1.5 can be obtained, with L_i about 1 λ rather than 1.25 λ. This antenna has a response equivalent to an extended double Zepp at both 15 and 20 meters! It also has gain at 10.1 MHz and has a simple dipole response on 40 meters.

Antenna 3: 21/50 MHz

L_t = 18.1 ft
Ls = 5.1 ft
f_h = 50 MHz
f_d = 21 MHz
85 segments used for L_t
24 segments used for L_s

Here, L_i is chosen to be near λ/2 rather than 1.25 λ at f_h and L_s is chosen to be the predefined λ/4 stub. In this case, the bidirectional gain at both f_h and f_d will approximate a simple dipole. However, the feed-point impedance at both f_h and f_d will also approximate a simple dipole (in practice between about 50 and 100 Ω). For these dimensions, the ratio will be about 2.38, which is exactly the correct ratio for 50 and 21 MHz!

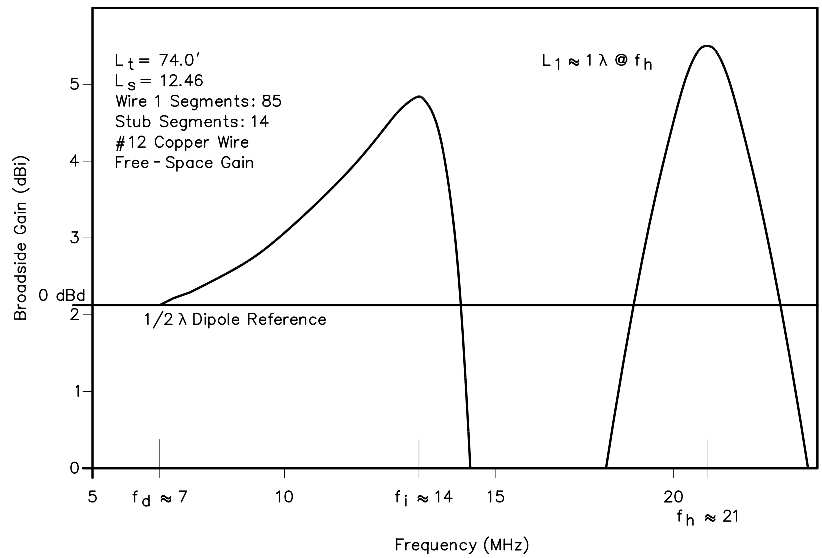

Fig 3—Antenna 2 broadside gain (dBi) versus frequency.

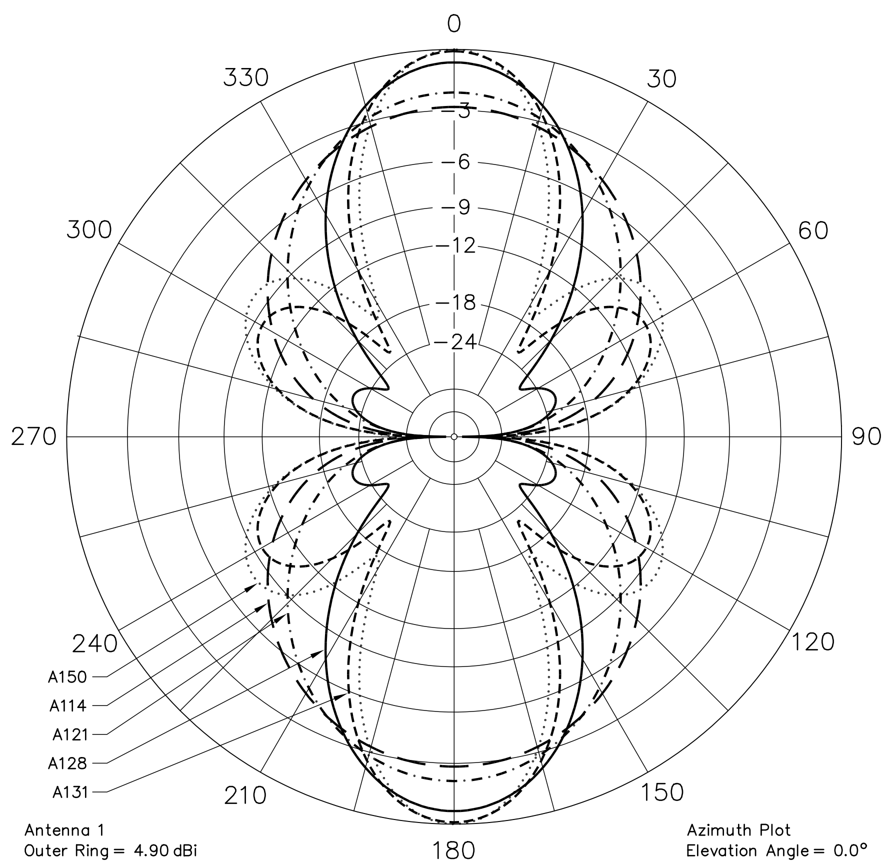

Fig 4—Antenna 1 free-space azimuthal radiation-pattern plots at 0° elevation for several frequencies. A114 indicates Antenna 1 at 14 MHz; A121 indicates Antenna 1 at 21 MHz, etc. 14.15 MHz = f_t; f_d < 21 MHz < f_i; 28.1 MHz < f_i; 31 MHz = f_i; 50 MHz = f_h.

The exact ratio was achieved by experimenting (as with Antenna 2) with small derivations of L_t and L_s. The best compromise was found to be a feed-point impedance of about 100 Ω at 50 MHz. This impedance was "lowered" by adding two parasitic elements and thus forming a three-element, 6-meter Yagi. (See Fig 6.) A 15-meter reflector could also be added, but the SWR will increase to about 2:1. Left as a dipole, it is near 1:1 on 15 meters. With the actual feed-point impedance near 50 Ω on both 6 and 15 meters, in effect, a "free" 15-meter rotatable dipole can be built into a "normal" 6-meter Yagi. The driven element is simply fed through a 1:1 balun and 50-Ω feedline.

This antenna was modeled using #12 copper wire, so an aluminum-tube Yagi's dimensions and taper schedule would have to be optimized. Fig 7 shows radiation patterns for 21 and 50 MHz.

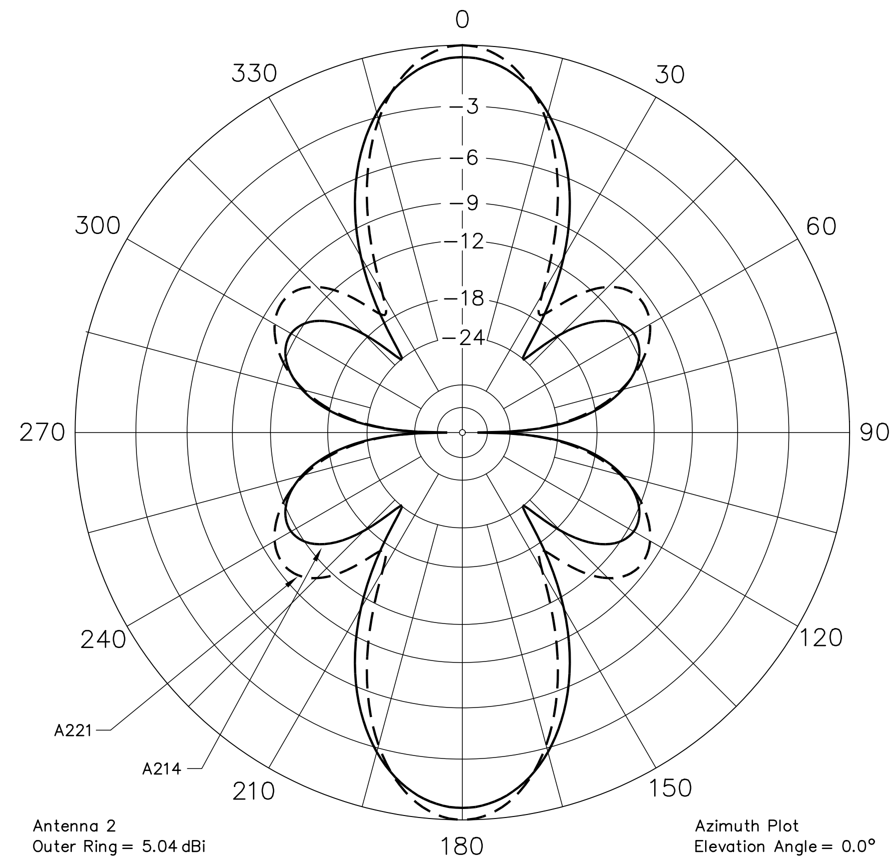

Fig 5—Antenna 2 free-space azimuthal radiation-pattern plots at 0° elevation for two frequencies. A214 indicates Antenna 2 at 14 MHz (f_l); A221 indicates Antenna 2 at 21 MHz (f_h).

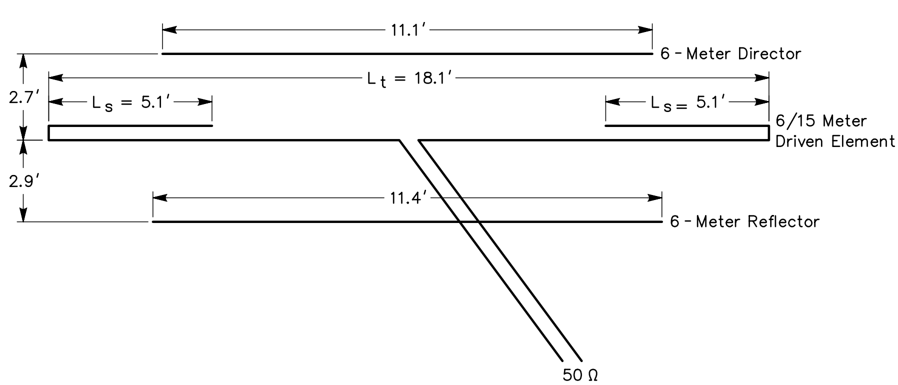

Fig 6—Antenna 3 is built from this schematic, a bottom view. It acts as a three-element beam at 6 meters, but functions as a simple dipole at 15 meters.

W8JK and Yagi Configurations

We can make a very effective two-element array using two extended double Zepps—like Antenna 1—configured as a W8JK array. (See Fig 8.) The bidirectional gain is roughly equivalent to a three-element Yagi in

Modeling Closely Spaced Wires with NEC-2

Closely spaced wires carrying very different currents are a bit tricky to model with *EZNEC* and other NEC-2 based programs. One important thing is to make the segment junctions on the parallel wires align with each other as closely as possible. As sent, the model reports a source impedance of 1263 –*j*2335Ω. I changed wire 1 to 43 segments to more closely match its segment junctions with those of the loading wires. The result was 254 –*j*1372 Ω. Actually, 42 segments is closer, but an odd number is required so that the source can be at the antenna center (unless a split source is used), so I also tried 41 for wire 1. The result was then 275 –*j*1415 Ω, not a large difference. Then I doubled the number of segments, keeping them aligned on the parallel wires. With 85 segments for wire 1 and 12 for the loading wires, it reported 325 –*j*1481 Ω. Approximately doubling again to 169 and 24 segments gave 366 –*j*1535 Ω. I'd say from this that wire 1 should have at least 85 segments for good results. More than that will improve accuracy some, but not a great deal. Now here's the interesting part: With wire 1 held at 169 segments, I changed the number of segments on the loading wires from 24 to 36, greatly disturbing the alignment of the segment junctions. The impedance changed to 366 –*j*1536 Ω—almost no change at all from the value with 24 segments. This shows that it's much more important to line up the segment junctions when the segment lengths are comparable to the wire spacing than when the segments are shorter.—*Roy Lewallen, W7EL* *both* directions at f_h and f_i. (See Fig 9.) Therefore, a rotatable two-element W8JK using Antenna 1's element dimensions (modified for tapered tubing elements) could provide quite impressive gain over six amateur bands. The array modeling here used six-foot spacing. However, this spacing drops the feed-point impedance to a low value at 14 and 18 MHz. This problem can be solved by placing a relay at the element feed points. By using a 61-pF capacitor inserted in line at the element's center, a reflector element is created at 14 MHz, and the feed-point impedance is changed to a more acceptable value. There is a common misconception that the driven element in a Yagi array must be a resonant dipole. In fact, the only critical electrical lengths are those of the parasitic elements and their spacing. Even the parasitic-element lengths can be tuned by appropriate reactive loads, as the example above shows.

Here is an interesting possibility: Because the two elements are identical, either element may switched into the reflector mode. Thus, the main pattern lobe may be easily and instantaneously reversed (180°) at 14 MHz by placing a relay and 61-pFcapacitor at the center of each element. This is very useful because one can instantly check whether a signal is arriving via short or long path without physically rotating the Yagi. Therefore, it is quite practical to configure the two-element antenna to provide a switchable pattern (0° and 180° azimuth for 20 meters) *and* a W8JK bidirectional pattern for bands from 14 through 50 MHz. Of course, with added switching complexity, a unidirectional pattern could be configured at 18 and perhaps 21 MHz. This feature would be very useful for working gray-line DX, nets and contests.

Note that getting the exact required reactive values from NEC and NEC-based modeling programs is difficult. Some tweaking will be required to optimize front-to-back ratios and/or maximize forward gain.

The combinations seem endless. The essential point is that this antenna is the same size as, and has comparable hardware to, a standard two-element, 20-meter Yagi, yet it can show excellent gain performance on all bands from 20 up to 6 meters. Two Antenna-2 elements could be fashioned into a W8JK as well, to provide similar gain characteristics on 40, 30, 20 and 15 meters. It is important that for short spacings in the W8JK configuration, the feed-point impedance can become difficult to

What about Ground Effects?

The antennas in this article are modeled in free space. That's fine for general comparisons of the major-lobe signal strength among several antennas. It's a good idea, however, to look at the big picture as well. To do that, we need to put the antenna over real earth and view the elevation plot as well as the azimuth plot. As an example, I've taken W7SX's Antenna 1, placed it 35 feet above real ground (0.005 S/m, ε = 13) and screen captured three-dimensional plots for *EZNEC2* high-accuracy ground analyses. The antenna placed l/2 above real ground exhibits take-off angles greater than 0°, and the nulls are largely filled by ground reflections (Fig A). As the height increases—with respect to λ—the antenna may show lesser lobes above the main lobe. At some frequencies and heights, there is significant vertical radiation (Fig B). Similar effects remain, even when the antenna is 125 feet high.—*Bob Schetgen, KU7G*

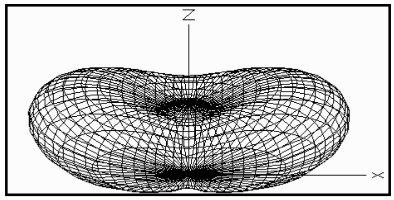

Fig A—Antenna 1 at 14 MHz, modeled 35 feet (λ/2) above real ground. The takeoff angle is about 25°. We can consider this the best-case ground effect for this antenna.

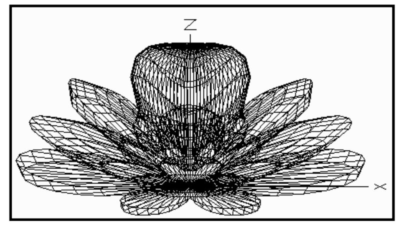

Fig B—Antenna 1 at 50 MHz, modeled 35 feet (1.77 λ) above real ground. The takeoff angle is about 10°. We can consider this the worst-case ground effect for this antenna.

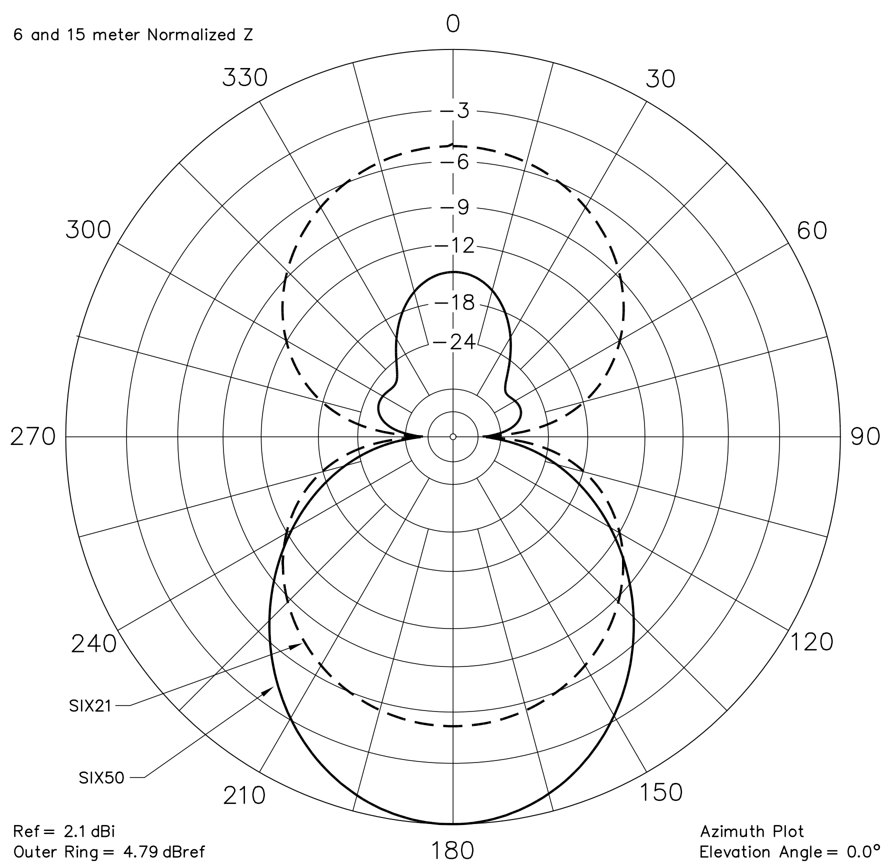

Fig 7—Antenna 3 free-space azimuthal radiation-pattern plots at 0° elevation for two frequencies. SIX21 indicates Antenna 3 at 28.1 MHz; SIX50 indicates Antenna 3 at 50 MHz.

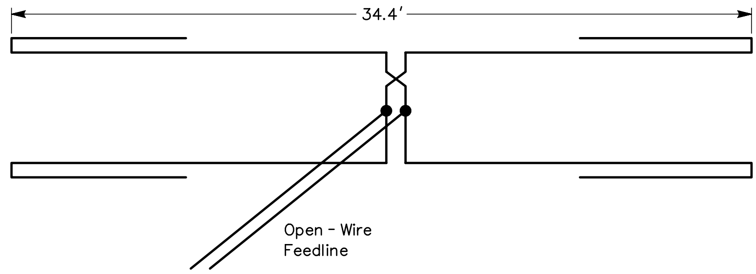

Fig 8—This schematic shows a W8JK-style array (I call it W8JK-EDZ) made from two elements, each with dimensions as shown for Antenna 1; it's a bottom view. Both elements lie in a horizontal plane.

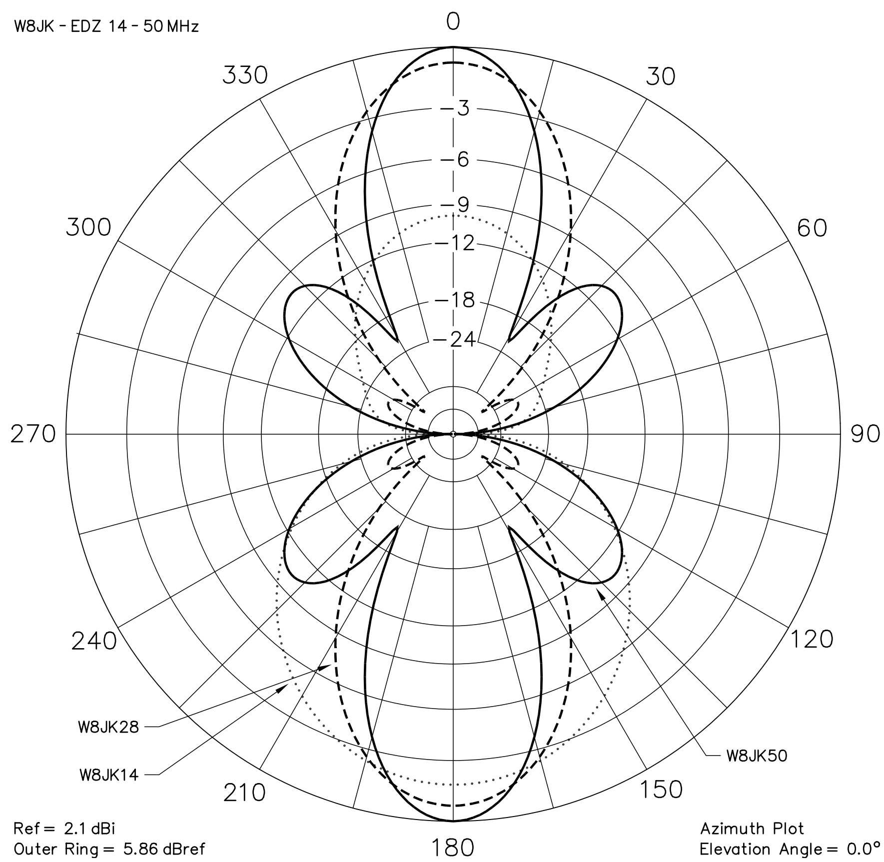

Fig 9—W8JK-EDZ free-space azimuthal radiation-pattern plots at 0° elevation for two frequencies. W8JK28 indicates W8JK-EDZ at 28 MHz (28.1 MHz = f_l); W8JK50 indicates W8JK-EDZ at 50 MHz (50 MHz = f_h). W8JK14 indicates a W8JK-EDZ operated as a 14.1-MHz, two-element Yagi. The Yagi is just as shown in Fig 8, except only one element is fed and the other loaded with a series connected 61 pF capacitor at its center.

match as operating frequency approaches f_d. This is the motivation for switching to a parasitic array at and near f_d, not to mention greater gain at these lower frequencies!

Except for Antenna 3, I've assumed that these antennas have open-line feeds, with a tuner placed in some convenient location. I used 450-Ω ladder line in the original arrays I built, and tuned to the operating frequency in the shack. However, more intensive modeling with a Smith Chart software tool—such as *MicroSmith*—might reveal some clever multiband scheme that matches to 50 Ω.

Conclusion

Similar gain numbers are possible for other frequencies by appropriately scaling lengths and spacing. By simply replacing λ/2-dipole wires with this element, many of the familiar array configurations shown in the *ARRL Antenna Book* and other texts take on new dimensions of gain versus frequency. Yet, this basic antenna element is lightweight, inexpensive and easy to build. In the battle to squeeze a few more decibels of gain out of a given space, I hope this technique will be considered as an interesting decibel/dollar option.

Note

[1]"Convergence testing consists of increasing the number of segments per unit of length equally throughout the antenna structure and observing changes in the output data for parameters significant to the modeling exercise." This definition is taken from "NEC-4.1: Limitations of Importance to Hams," by L. B. Cebik, W4RNL (*QEX*, May/June 1998, pp 3-16). When increasing the number of segments significantly changes some result, say feed impedance, it indicates that the previous results are in error because there were too few segments. When the results change little with added segments, they are said to converge and are considered accurate. Readers can learn more about modeling limitations from Mr. Cebik's article and "Wire Modeling Limitations of *NEC* and *MININEC* for Windows" by John Rockway and James Logan, N6BRF (*QEX*, May/June 1998, pp 17-21).

Creating 3-D Antenna Radiation-Pattern Plots

We've needed this for a long time. Learn how a QuickBasic 4.5 program can convert tabulated antenna-pattern files into three-dimensional radiation-pattern plots. It could even let you vary your point of view.

By Doug Smith, KF6DX

Graphical depictions of antenna radiation patterns yield a great deal of information at a glance. While the two-dimensional projections we're used to seeing allow us to read gain versus angle in a single plane immediately, many plots would be required to get an idea of performance over an entire hemisphere. Three-D plots are useful because they display the whole pattern in a single illustration. They let the antenna experimenter see more quickly what he or she has produced, and to some, they are aesthetically pleasing.

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As I discovered, producing 3-D renderings on a computer is more difficult than the pretty results might suggest. I'll explain how I did it and provide examples of programming constructs along with some of the mathematics involved.

NEC Plot Files

I'm using *NEC4WIN* V1.91[1] running under *Windows* 3.1. Sorry, but I can't abide the upgrades to operating systems that require concurrent and continual upgrades in hardware and the associated expenses! My computer is a 486DX2-50 MHz machine. In that most antenna-analysis software produces output files that adhere to a standard format, the following ought to apply whether you're using a *MININEC* program or *NEC4*.

My plotting software, written in and running under *QuickBasic* 4.5,[2] assumes that pattern data are contained in an ASCII file. This file must be generated by the antenna-analysis software prior to running my program. For antennas plotted above "real" ground, only zenith angles from 0 to 90° make sense. Antennas in free space can use zenith angles from –90 to 90°. When generating the data file, attention must be paid to these limitations, or anomalous plates will be generated. Each line in the file is delimited by a CR/LF combination and bears fields separated by spaces representing (in left-to-right order)

[1]Notes appear on page 43.

the zenith angle, azimuth angle, vertical gain, horizontal gain and total gain. The first few lines of a file are shown in Table 1.

Of course, the file can be rather large, since it typically includes every zenith and azimuth-angle combination in increments of, say, 2.5° to 5°. The number of points plotted, therefore, is approximately $36 \times 72 = 2592$. With one line per point and about 60 characters per line, the file size is approximately $60 \times 2592 \approx 152$ kB.

The first task is to read the data points from the file and translate them from their spherical coordinates to Cartesian coordinates. Since the goal is to plot them on a two-dimensional screen, we must finish with coordinates indicating dimensions along two orthogonal axes of the three in the Cartesian system, x, y and z. The file's spherical coordinates are taken to represent the zenith angle ϕ, the azimuth angle θ and the radius ρ. The relationship between these coordinate systems is shown in Fig 1. Data in the file are in ASCII format, so I use the *BASIC* "VAL" function to convert the fields to numeric values. Angles in degrees are converted to radians, since that's what the *BASIC* trigonometric functions require as arguments. The radius in this case is in dBi. I scale this variable to compress the dynamic range of the plot. In addition, I wanted to be able to rotate the final view through all possible zenith and azimuth angles to be able to see all sides of the pattern. First I'll describe the spherical-to-Cartesian transposition, then I'll add rotation of axes.

Transposing the Coordinates and Rotating the Axes

Below are the equations for spherical-to-Cartesian transposition:

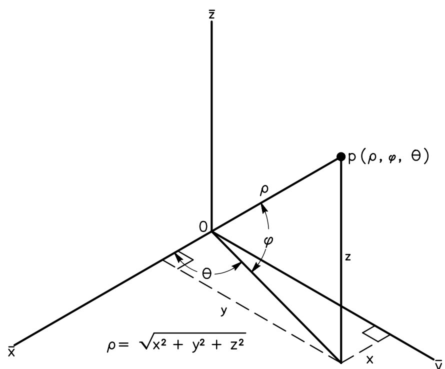

Fig 1—Relationship between spherical and Cartesian-coordinate systems.

$$x = \rho \cos\theta \cos\phi \qquad \text{(Eq 1)}$$

$$y = \rho \sin\theta \cos\phi \qquad \text{(Eq 2)}$$

$$z = \rho \sin\theta \qquad \text{(Eq 3)}$$

Note that to produce a radiation-pattern plot as viewed by a person standing on the ground some distance from the antenna, coordinates y and z would be used, not x and y. Variable x represents the axis perpendicular to the screen (or page).

Allowing the observer to walk in a circle around the antenna is equivalent to adding an offset to the azimuth angle, θ. I'll call the arc through which the observer has walked $d\theta$. Now we have:

$$x' = \rho \cos(\theta + d\theta)\cos\phi \qquad \text{(Eq 4)}$$

$$y' = \rho \sin(\theta + d\theta)\cos\phi \qquad \text{(Eq 5)}$$

$$z' = \rho \sin\phi \qquad \text{(Eq 6)}$$

Now put the observer in a cherry picker on wheels so that they cannot only go around the antenna, but can also increase their elevation above ground. This is equivalent to altering the zenith angle by an amount $d\phi$, but this rotation affects x and z (not y) depending on the offset $d\theta$. So the

Table 1

File : B10M3EL.N4W
FAR FIELD in dBi; Power : 66.36 Watts
From Zenith : 0° to 90° in 2.5° Increments
From Azimuth : 0° to 360° in 5° Increments
Far Field Pattern DATA

Zenith Angle	*Azimuth Angle*	*Vertical Pattern (dB)*	*Horizontal Pattern (dB)*	*Total Pattern (dB)*
0	0	–999.00	3.21	3.21
2.5	0	–999.00	3.62	3.62
5	0	–999.00	4.22	4.22
7.5	0	–999.00	4.97	4.97
10	0	–999.00	5.82	5.82
12.5	0	–999.00	6.70	6.70
15	0	–999.00	7.53	7.53
17.5	0	–999.00	8.26	8.26
20	0	–999.00	8.81	8.81
22.5	0	–999.00	9.09	9.09
25	0	–999.00	9.01	9.01
27.5	0	–999.00	8.42	8.42
30	0	–999.00	7.07	7.07
32.5	0	–999.00	4.43	4.43
35	0	–999.00	–1.25	–1.25
37.5	0	–999.00	–20.07	–20.07
40	0	–999.00	1.19	1.19
42.5	0	–999.00	6.67	6.67
45	0	–999.00	9.46	9.46
47.5	0	–999.00	10.73	10.73
50	0	–999.00	10.69	10.69
52.5	0	–999.00	9.11	9.11
55	0	–999.00	4.92	4.92
57.5	0	–999.00	–10.77	–10.77
60	0	–999.00	2.36	2.36
62.5	0	–999.00	8.69	8.69
65	0	–999.00	11.15	11.15
67.5	0	–999.00	11.45	11.45
70	0	–999.00	9.59	9.59
72.5	0	–999.00	3.90	3.90
75	0	–999.00	–10.54	–10.54
77.5	0	–999.00	6.56	6.56
80	0	–999.00	10.76	10.76
82.5	0	–999.00	11.85	11.85
85	0	–999.00	10.55	10.55
87.5	0	–999.00	5.76	5.76
90	0	–999.00	–111.53	–111.53
0	5	–17.98	3.18	3.21
2.5	5	–17.59	3.58	3.62
5	5	–17.01	4.18	4.21
7.5	5	–16.30	4.93	4.96
10	5	–15.51	5.78	5.81
12.5	5	–14.71	6.66	6.69
15	5	–13.97	7.49	7.52
17.5	5	–13.35	8.22	8.25
20	5	–12.94	8.77	8.79
22.5	5	–12.80	9.05	9.08
25	5	–13.05	8.97	8.99
27.5	5	–13.83	8.38	8.40
30	5	–15.38	7.03	7.05
32.5	5	–18.26	4.38	4.41
35	5	–24.19	–1.30	–1.27
37.5	5	–43.29	–20.12	–20.10
40	5	–22.33	1.14	1.16
42.5	5	–17.19	6.62	6.64

transposition is applied to the modified coordinates of Eqs 4, 5 and 6, above, to produce:

$$x'' = x' \cos d\phi - z' \sin d\phi \qquad \text{(Eq 7)}$$

$$y'' = y' \qquad \text{(Eq 8)}$$

$$z'' = x' \sin d\phi + z' \cos d\phi \qquad \text{(Eq 9)}$$

Connecting the Dots

Each of these data points is stored in a three-dimensional array, $d(x,y,z)$. We know the data appear in m groups in order of increasing zenith angle, and within each group are n points in order of increasing azimuth angle. In our example, $m = 36$, and $n = 72$. For every point where index $k > 36$ and $(k - 1)/36$ is not an integer, point d_k's nearest neighbors are points d_{k-1}, d_{k-36} and d_{k-37}. These points are shown in Fig 2, with dotted lines connecting successively computed points.

These four points define a rhombus, or four-sided figure. The idea here is that each group of four points may be connected with lines, colored in, and treated as a panel with which to build up a 3-D projection. (See Fig 3.) Note that while three points always lie in some single plane, four points may not; the plates aren't necessarily flat, therefore. We plow through the data, compute an interior point for each plate, paint it and build up the pattern on the computer screen. The *BASIC* "PAINT" function is used to do this. I use dark lines to outline the plates and a lighter color for the paint. The only trouble is that we want the plates closest to the observer to cover those that would be hidden from view. The data points must be sorted in order of distance from the observer.

Sorting the Data

Much effort has been expended refining sorting algorithms over the years. As stated above, the X axis is the one perpendicular to the screen or paper, so we use only that coordinate during sorting. The method I used is far from the best, I'm sure, but it employs one of the underlying principles of fast sorting—the binary search. It makes no assumptions about the data order at the outset.

Theorem: If we compare the datum to be sorted (at index k) with the datum centermost in the data list, we can decide whether that datum is greater than or less than the centermost datum. The next datum used for comparison resides at the midpoint of the previously selected half of the list. That half is again divided, and so on, until no further division is possible. The sorted datum is moved to that position in the list. The process is repeated for the datum at index $k + 1$, and so on, until all data have been repositioned once. Please don't ask me to prove why this works, because I can't, but work it does. The slowest part of it seems to be rearranging the data when the correct insertion point is found. For a list of length mn, the maximum number of comparisons is $\log_2 mn + 1$ It seems to me that a second—or even third—array could be used to store the final results with a further increase in speed, but I found I didn't need it to be that fast. My old dinosaur PC sorts the 2592 points in a minute or two. The antenna-analysis program, on the other hand, takes 20 to 30 minutes to write the data file for an antenna of 225 segments.

If you've been following all this rigmarole, you may be saying "Hey, now that you've sorted the data in order of the X coordinate, you don't know which points are adjacent any more, so you can no longer build those nice plates..." or "Who's on first, What's on second, I don't know's on third." The snappy comeback is this: During sorting, I didn't really move the data, but simply used another one-dimensional array of length mn to store the indices of the data in the sorted order. So I still have the original matrix intact for forming the plates.

Plotting the Data

Starting with the fourth datum as indexed by the sorted array, dark lines are drawn between that point and its three neighbors using the y coordinate for horizontal and the z coordinate for vertical. The first three points are the rearmost points, and so do not have rearward neighbors. When the matrix index is greater than 36, however, the dark lines are drawn, and a seed point is computed. The seed point is the place painting begins for the plate.

The seed point is found by calculating the mean of the four points. Eg, the y coordinate is found using $y = (y1 + y2 + y3 + y4)/4$, and the z coordinate by $z = (z1 + z2 + z3 + z4)/4$. This works fine as long as the plate has sides that don't cross one another along the field of view; ie, as long as the enclosed area doesn't resemble an hourglass. I found it necessary to draw the plates in an unoccupied area of the screen (the lower left corner) and then move them into place on the main figure using binary GET and PUT commands. Trying to paint over already-painted areas with my

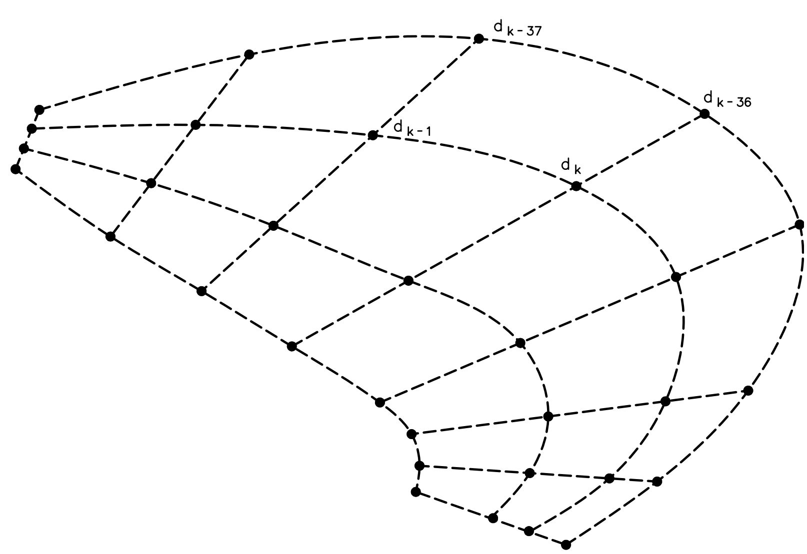

Fig 2—Four adjacent points in the data set.

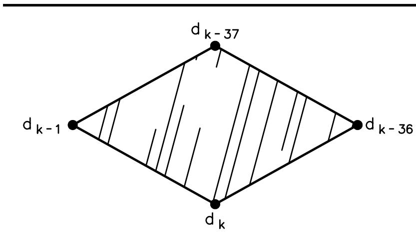

Fig 3—The four adjacent points form a "plate."

version of BASIC gave unexpected results. I further found I had to draw rectangles around the plates and repaint the background inside the rectangles to eliminate spurious effects. It was even necessary to prohibit painting when the enclosed rhombus was too skinny. I attribute the need for these work-around solutions to *BASIC*'s (and my own) statistical errors in integer math.

After the figure has been drawn, the program adds a reference grid consisting of ellipses and radial lines. This grid isn't calibrated to any particular scale, although it could be. On the cover are examples of the program output. The images were captured using *Pizazz Plus*[3] in .TIF format. This is what the ARRL production folks like to see for easy layout.

Program Details

The first line in the program relates to the way *BASIC* allocates memory for variables. The "$DYNAMIC" metacommand allows *BASIC* to allocate memory at run time "on the fly." *BASIC* should be invoked using the "/AH" command-line option to allow numeric data to occupy more than 64 kB. A fair bit of variable memory is required, as we have at least five variables of size mn. Other versions of BASIC may not be able to handle large arrays, so the number of points plotted may have to be reduced. The variable "MAXSIZE" should be changed if an "Out of Memory" error is encountered.

Defining large arrays as integers saves a great deal of memory space, and speeds execution of the program, especially during sorting. Variables beginning with letters "*C*", "*D*" and "*K*" through "*N*" default to the integer data type.

The screen resolution is fixed at 640 × 480 pixels. This requires a VGA display, or better. Variable *KMOD* represents the number of zenith steps, m, used in the input file for each azimuth step. *SFACTOR* is a linear scaling factor for the radius, ρ, of the plotted points. It can be altered to vary the size of the plotted image on the screen. *PROT* represents $d\phi$, and *TROT* corresponds to $d\theta$, both in radians.

Variable *LIMIT* sets the radius scale's limit in decibels. *B* is computed to compress the range of radii plotted for a more serviceable visual result. Linear arrays *DX*, *DY* and *DZ* hold the data points, while array *DA* is used to hold the sorted array indices. Array *DR* stores the value of ρ for each data point so that plate color can be altered at plot time according to radius. At the beginning of the plotting section, ten colors from a palette of 262,143 may be selected for use as plate paint. The formula for calculating the color number is given in *BASIC* as:

$$color = 2^{16}\ B + 2^{8} G + R \qquad \text{(Eq 10)}$$

where B, G and R are numbers from 0 to 63 indicating the intensity of blue, green and red in the color selection. Since there are gaps in the sequence, the formula should be used rather than just picking a number.

Reference grid ellipses are calculated point by point, because *BASIC*'s "CIRCLE" statement doesn't support rotation of axes. This is also done so that the reference grid doesn't overwrite the pattern.

Conclusion

I understand some of the better antenna analysis programs now have 3-D pattern plotting, but many still do not. I hope this encourages you to write your own programs to do it. Anyone interested can download my *BASIC* source code.[4] Comments and suggestions are welcome!

Appendix

Experience compiling and running my BASIC program under *QuickBasic* 4.5 and *VisualBasic* 1.x prompted me to add a few notes about the limitations of the program, as well as some peculiarities of the *MININEC*-generated data files.

To avoid gaps in memory allocation, QB 4.5 indicates that integer arrays ought to be dimensioned to integral-power-of-two sizes, eg, 16,384. In at least one instance, we were stopped by a "subscript out of range" error as soon as we added buffers for other functions, such as disk navigation for the input file. Various versions of *BASIC*, especially those running under *Windows* or other multitasking operating systems, may react differently. I suggest redimensioning the large arrays to the minimum size necessary for your applications.

Some antenna modeling software may not write data files that contain all the information required to produce a 3-D plot. Others may not format the data precisely as shown in my example. You programmers will see that my program is not particularly flexible in handling alternate data file formats, but you will also see that the parsing routines are easily modified.

Most modeling software writes a zenith angle that is 90° minus the actual zenith angle. My program assumes this. Also note that the data file representing an entire above-ground pattern can be generated in two ways: (1) Use a zenith angle range of 0-90° and an azimuth angle range of 0-360°, or (2) use a zenith angle range of 0-180° and an azimuth angle range of 0-180°. The same idea applies to free-space patterns. Unexpected things occur when the data file contains overlapping pattern segments, or when the data are ambiguous because of weird angle-range settings.

So please view my program as a sort of example, and not as a finished product. Thanks to Managing Editor Bob Schetgen, KU7G, an accomplished programmer, for valuable feedback and assistance.

Notes

[1] *NEC4WIN*, Orion Microsystems, Madjid Boukri, VE2GMI, on the Web at **http://www.cam.org/~mboukri**. *NEC4WIN95* is now available with more features that are powerful.

[2] *QuickBasic* 4.5, Microsoft Corporation, 1988.

[3] *Pizazz Plus*, Application Techniques, Inc, Pepperell, Massachusetts, on the Web at **http://www.screencapture.com**.

[4] You can download this package from the ARRL Web **http://www.arrl.org/files/qex/**. Look for SMITH3D.ZIP.

Preparing to Receive Phase 3D's 10.4-GHz Downlink: A Project Study Report

Here's a project to keep you busy while awaiting Phase 3D's ascent. The dishes described are compact and easy to disguise.

By Josef Maier, OE3JIS

Editor's Note: This article also appears in the AMSAT-NA Journal *(Sep/Oct, 1998). Thanks to AMSAT-NA for permission to reprint it here.*

This project was carried out in my preparation for my ground station to receive 3-cm downlink signals from the Phase 3D satellite. I was curious enough about new X-band technology to identify components that are already available on the market for Phase 3D use. These components are relatively small in size and are so compact that in my case, I was able to install the components in an existing transceiver.

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Layout Basics

According to various AMSAT publications and previous studies, we can expect the Phase 3D frequencies shown in Table 1 for Mode X. Note that below 1 GHz, transmission and reception of space microwave signals are disturbed by cosmic-noise signals and above 15 MHz through absorption from atmospheric water vapor, water content and oxygen. However, the 10.4-GHz band is relatively free from these effects. Also, a lot of experience in the 11-12 GHz range exists from European satellite television technology for similar bands and also for the necessary equipment.

Microwave Antennas

The reception of 10.4 GHz signals requires high-gain antenna dishes, which usually must be home constructed. The following dish-construction options are available:

A) Parabolic dishes with central reception/transmission feeds.

B) Parabolic dishes with offset reception/transmission feeds

C) Dishes with indirect feed constructions (ie, Cassegrain, Gregorian and other backfire systems)

For amateur reception purposes, the first two (A and B) options are most interesting, and these are the ones that I have realized, tested and report in this article. Fig 1 shows the approximate relationship between power gain, dish diameter (Option A) and 3-dB beamwidth in degrees for 10.4 GHz.

Bigger dishes yield higher gains and narrower beamwidths, so they require more-precise directional control to track a satellite's position. Therefore the dish should be as small as possible for a good signal reception. In addition, smaller dishes have lesser wind loads. (See Table 2.)

Table 1—Preliminary Phase 3D 10-GHz Specifications

Downlink Frequencies
Analog: 10,451.025 MHz to 10,451.275 MHz with center at 10,451.150 MHz
Digital: 10,451.450 MHz to 10,451.750 MHz

Satellite Transponder
PEP transponder: 50 W
Satellite antenna gain: 20 dBi
Satellite EIRP: 37 dBWi
PEP per QSO: 24 dBWi
Path loss: 207 dB
Ground station EIRP: –183 dBWi

Ground Station Options
Station 1: 60 cm (2ft) dish
Gain: 33 dBi
Signal Power/QSO: –150 dBWi
Noise Temp 150K: 1.5 dB NF
Noise Power in SSB: –173 dBW
Signal-to-Noise Ratio: 23 dB
Station 2: 30 cm (1 ft) dish
Gain: 27 dBi
Signal Power/QSO: –156 dBWi
Noise Power in SSB: –173 dBwi
Signal-to-Noise Ratio: 17 dB

Meanwhile an offset dish uses a section of the parabolic shape that is not symmetric about the centerline. Fig 2 shows the center cross-section view of such a dish. This construction has several advantages:

- No feed shadow on the dish
- No reduction in gain
- Easier to construct and t adjust the exact focus point
- Flexibility for future installation of a second feed for other bands.

The offset of the focal point is less critical for adequate performance. This makes it possible to place a second feed (for another band) on the

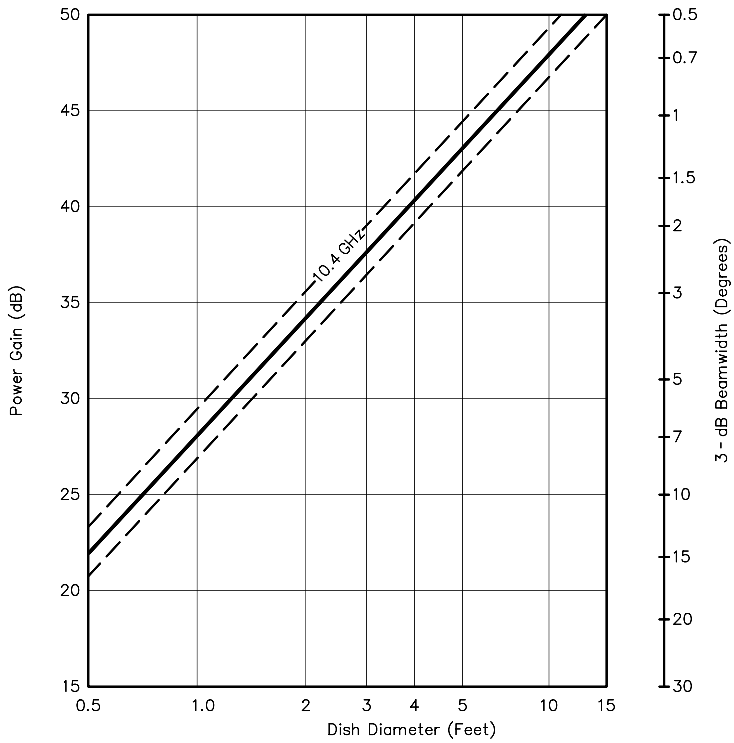

Fig 1—Relation between gain, dish diameter and 3-dB beamwidth (in degrees) for 10.4 GHz.

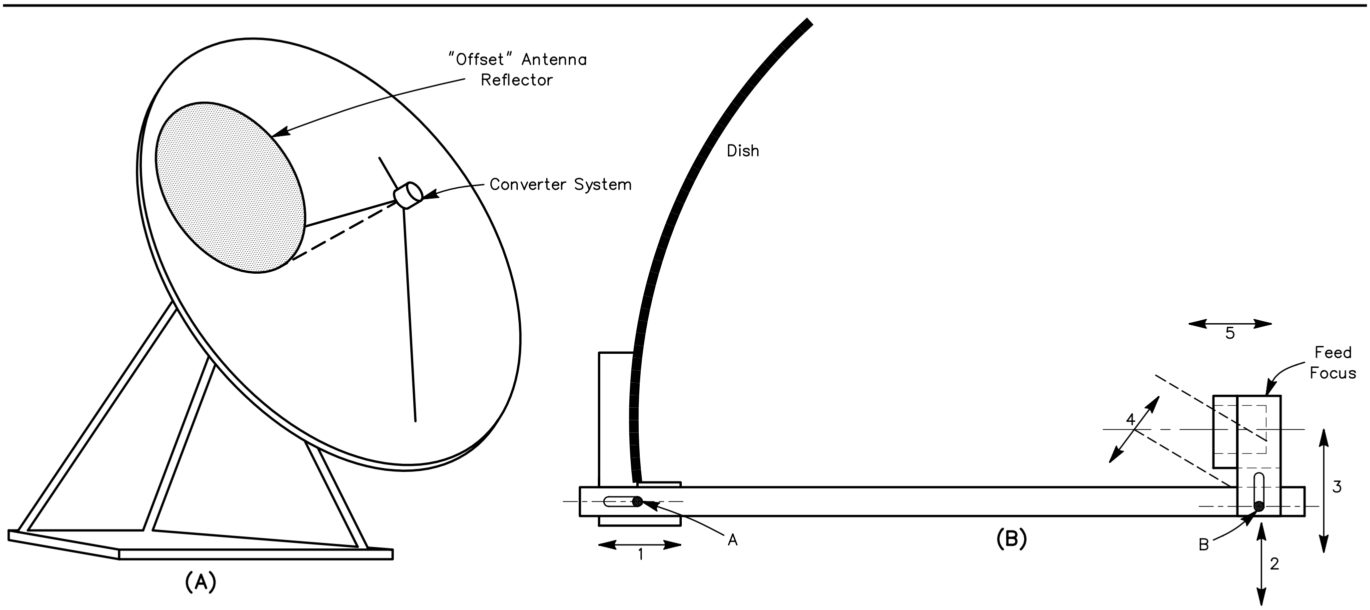

Fig 2—A shows the derivation of an offset-feed dish from a center-fed dish. B shows how an offset-feed may be adjusted. Arrows labeled 1 through 5 indicate possible focal point adjustments. Joints A and B swivel in the plane of the drawing.

same dish. In addition, small surface irregularities (holes or bolt heads) have no noticeable effect on the efficiency of the dish. The relation of focal length and dish diameter is essential for the feed construction. The feed must have a beamwidth that illuminates the dish from edge to edge, typically at the –10-dB points. If the beamwidth is too narrow, not all of the dish area is illuminated.

There are dishes made of metal-mesh constructions suitable for the frequency in use. For the 10.4-GHz band, there is no advantage for such dishes. For my experiments, I used industrially prefabricated aluminum dishes that are relatively inexpensive and available on the market.

Remarks about Microwave Transmission Lines

In the microwave bands, energy transport is done on the surface of conductors in thin layers. This is well known as the *skin effect* of metal conductors. Coax cables have extremely high losses in microwave bands; that is the reason why waveguides are used. The cross section of this guide can be rectangular, round or other shapes. The section surface is essential for the way that energy propagates down the waveguide. These different kinds of propagation are referred to as *modes*. A mode describes a pattern in which the field strength varies across the transmission line. In a well-designed line, only one mode exists; it is called the *dominant* mode. If a cross-sectional dimension of the waveguide is inappropriate (usually larger than the $\lambda/2$), a variety of propagation patterns occur—making performance unpredictable. The wave-guide should also be a good conductor with a high surface quality. For maximum performance, gold-plated constructions are used. For example, steel waveguide losses (in decibels) in the 10-GHz region are 2.5 times the losses in copper waveguide. (For copper waveguide with 1-dB loss, a similar steel waveguide would have 2.5 dB loss.—*Ed.*) This in a waveguide 25.4 ×12.7 mm in external, rectangular dimensions (Waveguide Nu.16 dimension). As shown in Table 3, you can also use copper pipes as waveguides.

To couple a signal from a waveguide to a downlink signal converter a *waveguide transition* is necessary. With this item, the signal is transferred to a coax cable and connector from the converter. Such transitions require some research using test procedures in a microwave laboratory and precision manufacturing. Fig 3 shows a waveguide transition for 10.4 GHz.

Downconverter

During my experiments, I used the 10-GHz super-low-noise MKU 10 OSCAR opt. 01 that downconverts signals from 10.451 GHz to 432 MHz. The gain is more than 30 dB and my test results show 42 dB. Its noise figure is 1.15 dB at 18°C. The MKU 10 downconverter is small (30×56×74 mm) and weighs only 95 grams. Power consumption is 220 mA, from 12-15 V dc. Fig 4 shows the circuit diagram of this downconverter that was designed by Michael Kühne, DB6NT. It is available at Kühne Electronic, BRD.[1] (In the United States this downconverter is available from SSB Electronics and Downeast Microwave.—*Ed.*)[2,3]

[1]Notes appear on page 49.

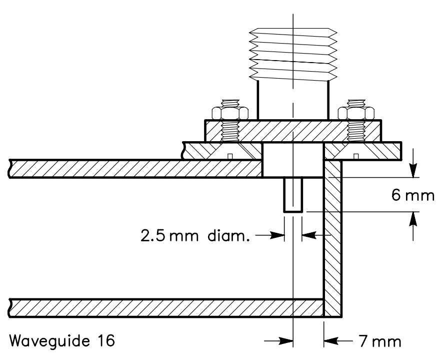

Fig 3—Cross-section view of a waveguide transition.

I used a 70-cm portion of an all-mode Kenwood TR-851 to receive the downconverted signal and with an ICOM IC-R7000 for backup. The converter case is not completely weatherproof. Later, I plan to place the MKU 10 in an airtight, soldered metal box for the outdoor installation. This weatherproofing is necessary to overcome the danger of corrosion from condensation, which I have experienced in earlier S-band installations.

10.451150-GHz Test Beacon

Unfortunately, Phase 3D is now on the ground and not in orbit. Therefore, a test beacon is necessary to test how well the assembly works. I ordered the MKU 10 from Kühne Electronics with several modifications (Fig 5). The power was reduced to the minimum level of 10 mW, and the oscillator crystal was trimmed to 10.41150 GHz. That frequency resides in the middle of Phase 3D's 10.4-GHz analog band. The size of the beacon is 111×55×30 mm, and its weight is 160 g. Laboratory test results concluded:

- Output power 10 mW
- Spurious and harmonics less than 40 dB
- DC current 220 mA, 12 to 15 V

Even at this reduced power level, a simulation of the expected Phase-3D signal at ground level is not possible because of the closeness between the transmitter and receiver. Calculations show that the beacon is much too strong, but there is a good possibility to control the function of the installation, and I have an S-band beacon for future tests.

Feedhorn

Fig 6 shows a cross-section drawing of the feed. The beamwidth of the feed is 140° at the –10-dB points—sufficient for the offset-dish solution I have identified. This feed was tested in the laboratory, and its losses are very low. The material is aluminum and the resonant monopole is gold-plated brass. Such feeds can be readily bought on the market.[4]

Table 2—Estimated Dish Wind Loads

Dish Diam.	Projected Surface Area (m^2)	Max. Force (Newtons, wind at 100 kph)
30 cm	0.072	85 N
60 cm	0.28	335 N
90 cm	0.64	770 N

Table 3—Properties of Copper Pipe used as Waveguide

OD (mm)	Wall Thickness (mm)	F1 Min (Cutoff, MHz)	F2 Max (Atten. Freq. MHz)
15	0.5	12,557	16,404
22	0.6	8452	11,041—3-cm guide
28	0.6	6560	8569

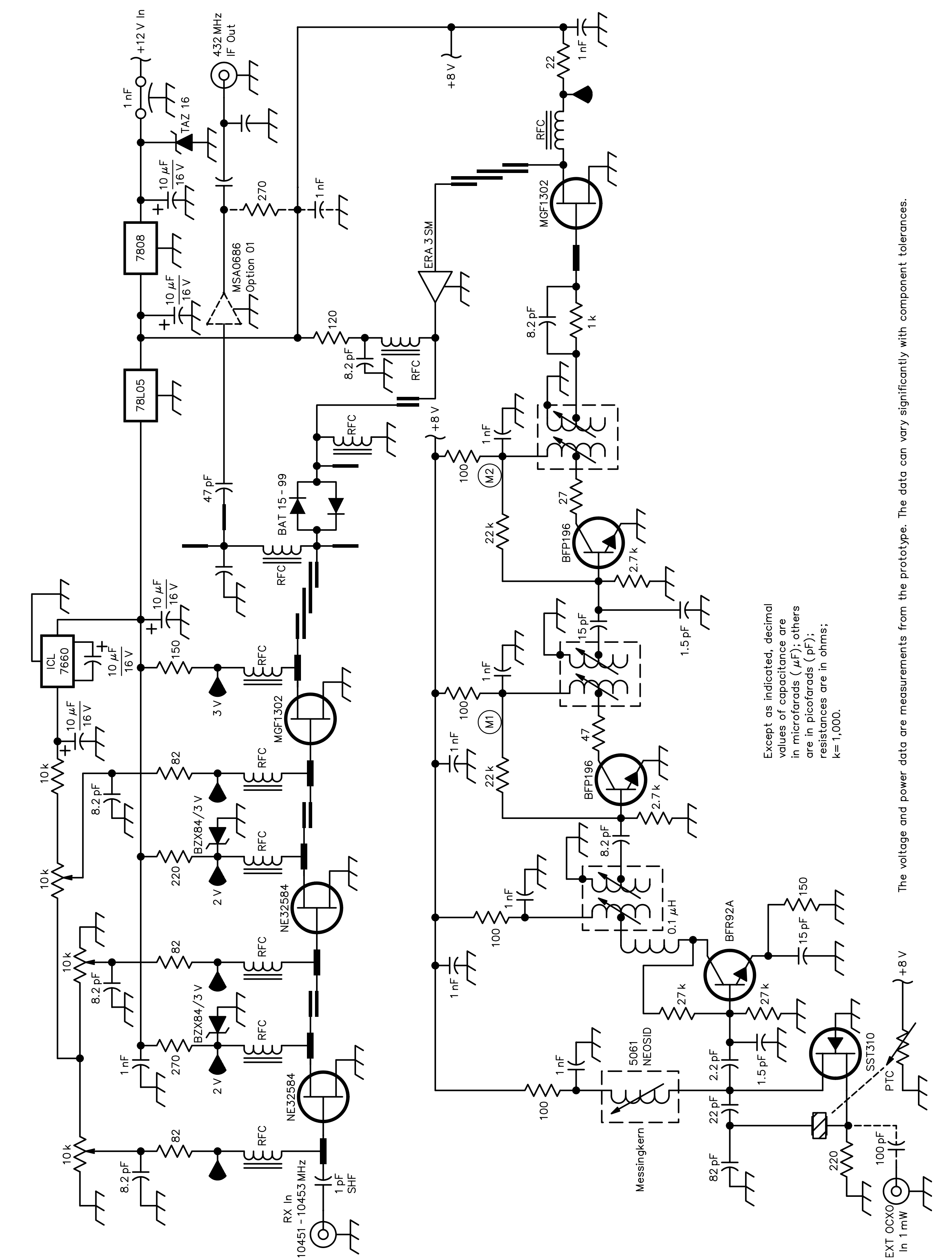

Fig 4—Downconverter MKU 10 OSCAR (option 1) circuit diagram.

Option A: Dish with Central Feed

I can buy this dish ready-made from Eisch Electronics (see Note 2). The dish is constructed by PROCOM-Denmark.[5] Table 4 shows its specifications.

At 10.4 GHz, the SWR is about 1.6:1 (see Fig 7). Meanwhile, Fig 8 shows the dish from the front side, with the reflector soldered to the central waveguide. Fig 9 is a view from the backside with the waveguide transition, converter and the other test arrangements. The waveguide and the transition are gold plated.

Option B: Offset-Dish Construction

This dish type is characterized by its elliptical circumference shape (Fig 10). The outside dimensions are 40 cm and 36 cm. The manufacturer is unknown to me, but it is a dish for digital satellite television that was inexpensively purchased. The feed (see the earlier description) is fixed with clamps, and the whole feed position can be adjusted in many ways (See Fig 2).

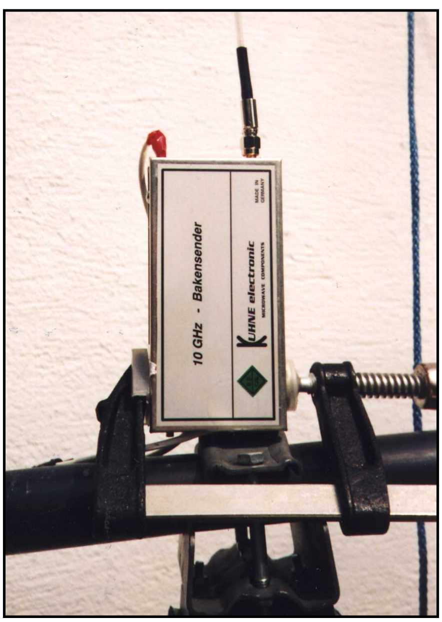

Fig 5—The MKU 10 test beacon.

Table 4—Procom Dish Specifications

Technical Data	
Diameter	48 cm
Gain	27 dBd
F/D	0.4
Beamwidth	6°

Testing These Dishes

First, I must state that these dishes were not made with exact scientific methods. I started the beacon and after a warm-up period of five minutes, and I heard a roaring S9+ signal on the all-mode Kenwood TR-851E at 432.150MHz. The high noise level was because of the high gain (42 dB) of the converter (at about S7) and was to be expected. The signal strength was so high, however, that the S-meter indicator was pinned. This first trial shows only that the system works on both dish types. Of course, the distance of five meters is too small and

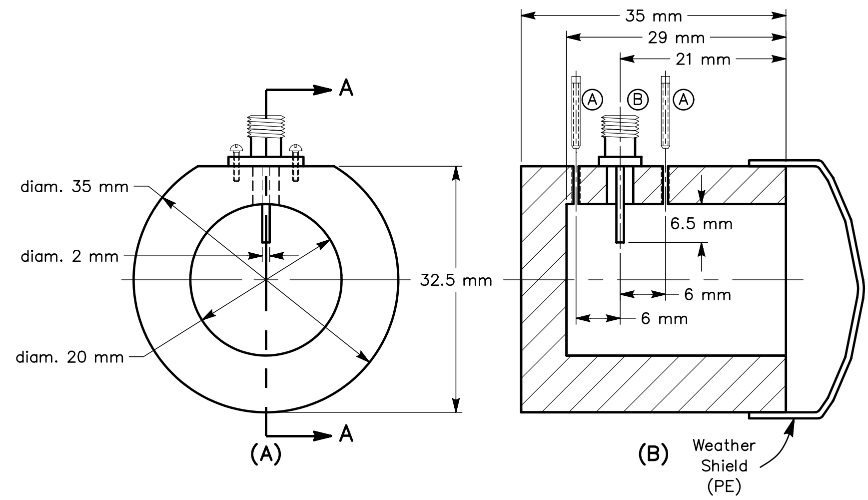

Fig 6—Cross section of 10.4-GHz feed. A is an end view; B is a cross section through AA. Circles A indicate M2 (metric) trim screws. Circle B indicates the SMA connector.

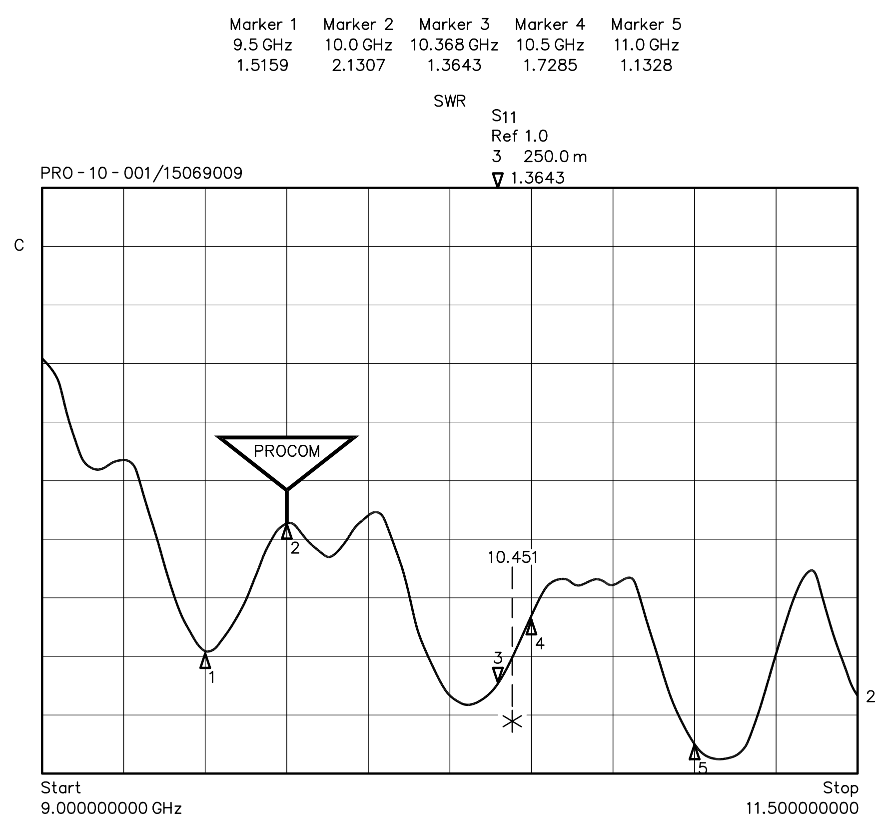

Fig 7—SWR diagram for the Procom dish.

Fig 8—Center-fed Procom dish front view.

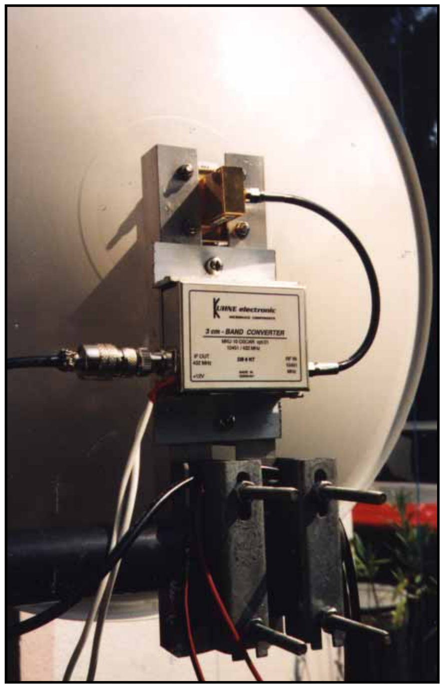

Fig 9—Center-fed Procom dish rear view.

Fig 10—Offset-feed dish view.

there were some wall reflections. The next step was an area-field test over a distance of 1 km and those tests confirmed that both the dishes work. Even this test showed strong S9+ beacon-signal readings. By moving the antennas up and down and left and right, I could make some adjustments of the offset-antenna feed.

The field tests had been repeated with an ICOM IC-R7000 as indicator. The noise readings were reduced to S1.5, and the signal indications were S9 to S9+. There was a difference of S7.5 to S8 between the noise and the USB signal. As I mentioned before, the beacon was too strong for the simulation of the Phase 3D predicted values.

For my case, tests have shown that the dish (Option 1) of the downconverter with its high gain has too much reserve and would be not necessary. The noise level is no big problem for me because I use an NF digital filter that rejects the noise to a great extent. I am very much delighted at the clear signal strength of the converter and the good reception S/N value.

Both types of antennas have a clearly defined polarization direction from the feed side. It was interesting to observe the effect of changing the polarization direction 90° relative to the beacon. In most test cases, the readings on the ICOM IC-R7000 S-meter changed by two S units. If the space signal comes down with circular polarization, a good S/N reserve will help a lot.

I intend to install the offset dish on the vertical rotating boom of my rig with an separate TV-dish rotator. I think I am ready for the Phase 3D X band. Hopefully, this satellite will soon be in orbit!

Notes

[1]Kühne Electronics, Birkenweg 15, D-95119 NAILA/Hölle BRD, Germany; tel 09288/8232, fax 09288/1768; kuhne.db6nt@hof.baynet.de.

[2]SSB Electronic, 124 Cherrywood Dr, Mountaintop, PA 18707; tel 570-868-5643; http://www.ssbusa.com.

[3]Down East Microwave Inc, 954 Rt 519, Frenchtown, NJ 08825 USA; tel 908-996-3584, fax 908-996-3702; http://www.downeastmicrowave.com/.

[4]Eisch Electronic, Abt-Ullrich-Str.16, 89079 ULM-Gögglingen BRD, Germany; tel:07305 23208, fax 07305 23306.

[5]Procom A/S, Vinkelvaenget 21-29-DK-3330 Gorlose, Denmark; tel (++45)42 27 84 84; fax (++45)42 27 85 48.

Josef Maier, OE1/OE3JIS, is married and has two sons. With an Austrian engineering degree, he has worked in several companies as a designer, project engineer and work planner. Josef is now retired. He was first licensed (CEPT 1) in 1987. Josef is very active on satellites (over 7000 QSOs—including 230 countries—on analog satellites) and holds several satellite awards. He is a member of ÖVSV, AMSAT-UK, ARRL, ISWL and a life member of AMSAT-NA. □□

A Regulated 2400-V Power Supply

Look at this great power-supply idea. SCRs on the transformer primary adjust input duty cycle based on output voltage. Output voltage drops about 3 V with a 1 kW load!

By A. R. (Al) Williams, VE6AXW

In the past, amateur-built power supplies to produce the high voltages needed for transmitters and receivers consisted of a transformer, rectifier and filter. Poor regulation caused receivers to shift frequency with line-voltage fluctuations. Since regulation did not exist, transmitters suffered because the leading edge of CW keying produced a higher RF output than during the remainder of the transmitted element. This tended to produce a leading-edge "thump" to the keying.

When commercial interests began building linear amplifiers, they followed the amateur practice of poor regulation. Some manufacturers introduced better transformer technology and added swinging chokes. They produced some improvement, but their regulation was usually still poor. This basic design has not changed since the early days of radio.

The power supply described in this article changes all of that. It produces 2400 no-load volts and 2397 V at the one-kilowatt load level. This is a drop of only three volts! Measurements were made using a Fluke Model 87 DMM with a high-voltage probe.

A look at Fig 1 reveals this supply is little different from any other, with one exception: It contains a pair of inverse-parallel connected SCRs in the primary of the main power transformer.

No ground is used at the negative terminal. This is in keeping with modern technology, wherein grid-current metering is done with the negative terminal of the supply "floated" from ground. My supply is intended to provide power for a cathode-driven (grounded-grid) linear RF amplifier. The plate-current meter, which reads 1 A at full scale in this case, is connected in the negative lead of the supply.

Two 813s are used. The grid-current meter is connected between the center tap of the filament transformer and ground. This measures grid current independently of plate current.

A control board regulates this supply. Very little detail is shown in Fig 1, but Fig 2 shows the complete board. Fig 1 is intended to show only external connections made to that board. It gives an overview of the system.

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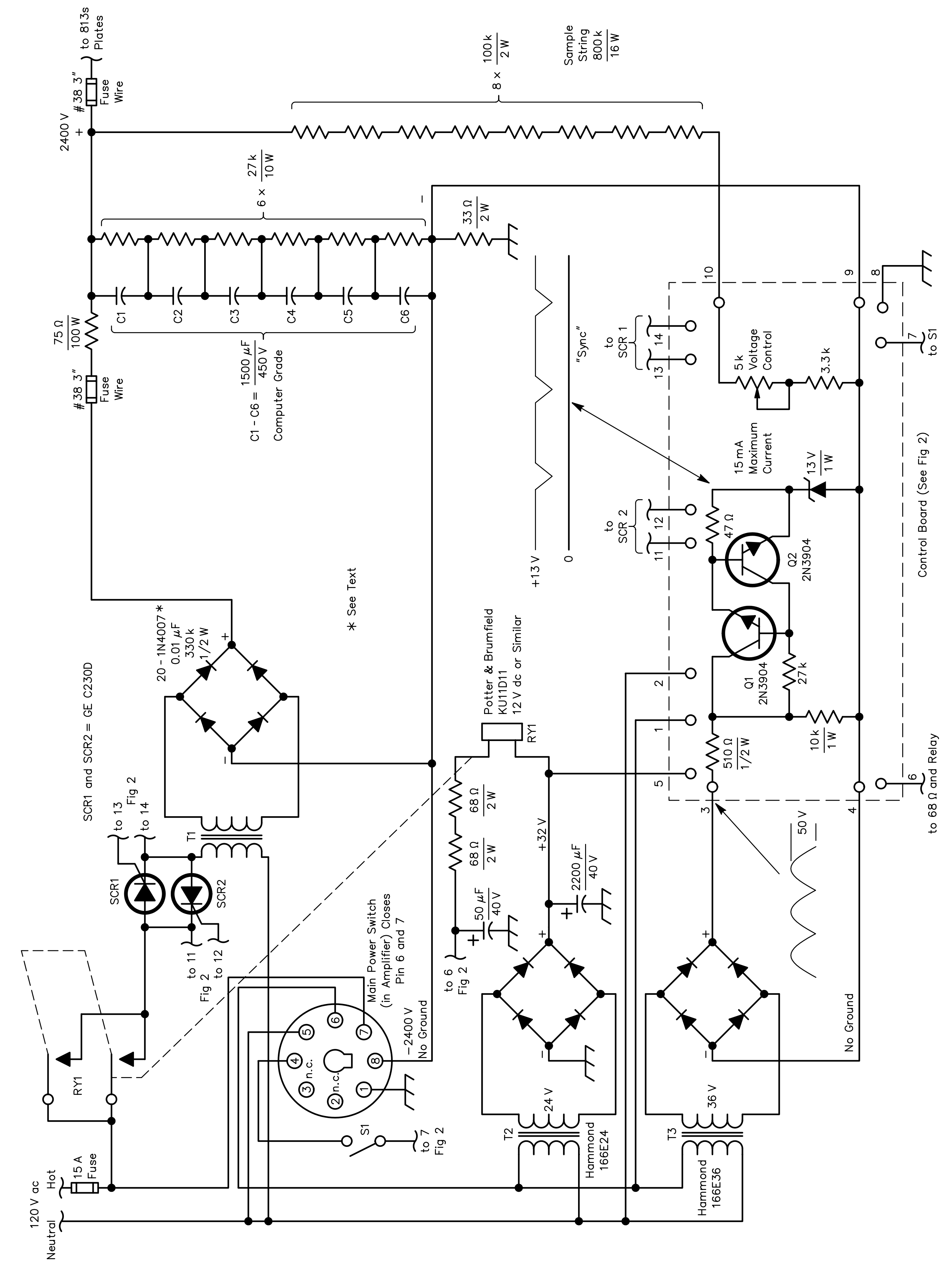

Fig 1—High-voltage power supply.

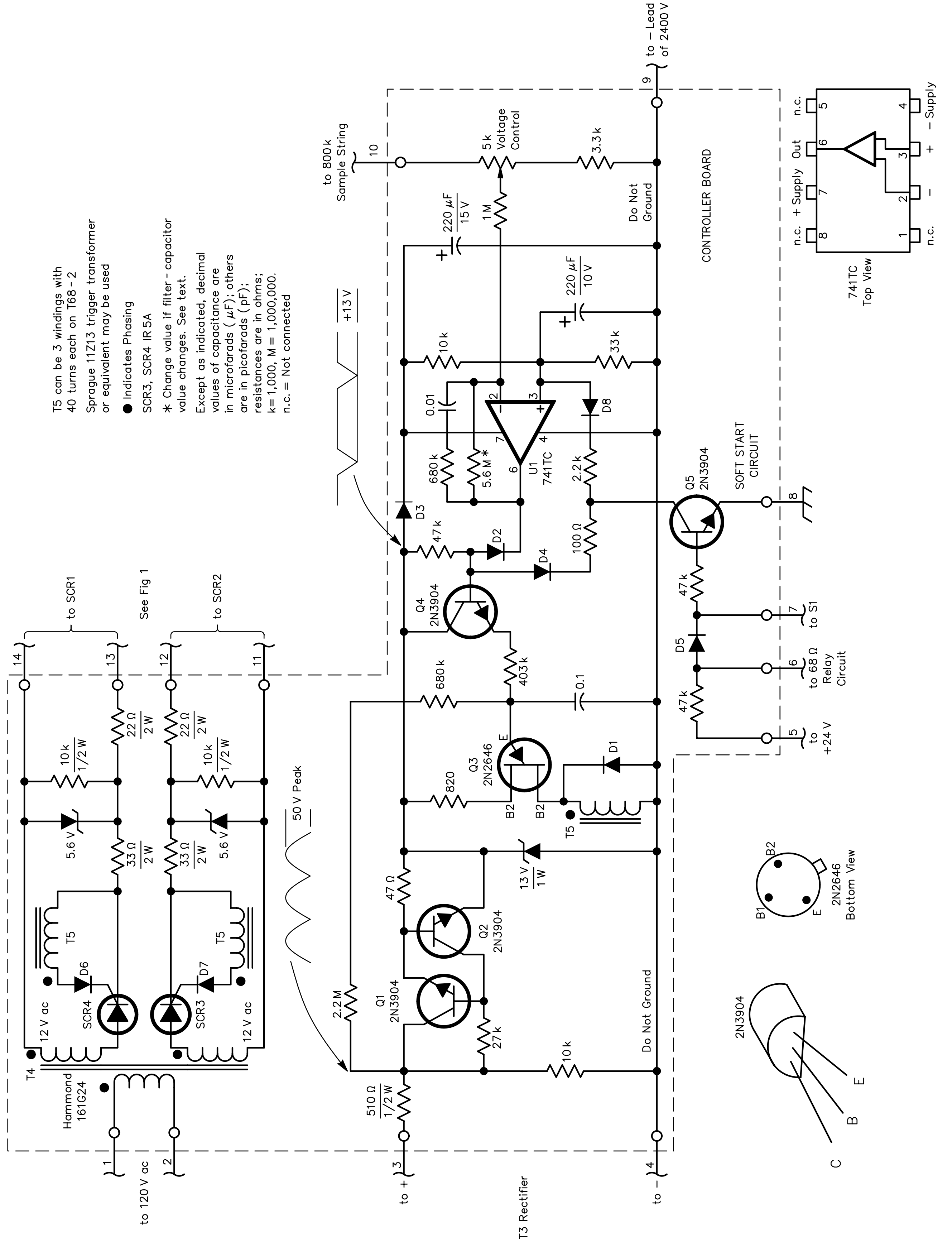

Fig 2—Regulator circuit. 741 op amp U1 and 2N3904 transistor Q4 control the height of the pedestal to the emitter of 2N2646 unijunction transistor Q3, which acts to control the firing (phase) angle of the SCRs. The response is in direct proportion to the supply's load current.

Fig 2 shows the "soft-start" features of this supply. When initially switched on, no voltage is applied to the controller via the "sample string." The controller is prevented from switching the SCRs on fully because Q5 turns on early and keeps pin 3 of the 741 at 0 V. The 220-µF capacitor at pin 3 is discharged at this time and must charge positively before phase-controlled firing of the SCRs can begin. The voltage at pin 3 must rise slowly because of the time constant of the 220-µF capacitor and associated resistance. This prevents the SCRs from turning on fully and limits inrush current. This protects the diodes in the bridge rectifier as well as the filter capacitors. The controlling mechanism operates according to the "Modified Cosine Ramp and Pedestal Generator" fully described in the General Electric (GE) *SCR Handbook*.

Whenever the output voltage tends to sag because of increased load or decreased power-line voltage, this droop is instantly impressed upon the input of the controller. The decrease in voltage is detected, and the controller responds by advancing the phase of the firing voltages applied to the SCRs to maintain the output voltage. If the output voltage tends to soar, the controller likewise retards the firing angle of the SCRs. My supply was designed for 2400 V, but it could be built for any voltage, providing a suitable sample-string resistance is used.

The bridge rectifier uses five 1N4007s in each leg. A 0.01-µF, 1-kV capacitor is paralleled with a 330-kΩ resistor, and the combination connected across each diode for voltage equalization. (This long-standing equalization practice is no longer neccessary. For details, look at the "Rectifiers, Strings or Stacks" section of the Power Supplies chapter of recent *ARRL Handbooks*.*—Ed.*)

Common 1-W resistors are rated at 500 V breakdown, and would be a better choice than the half-watt units used in my supply. I have had no breakdown problems nonetheless.

At first blush, it may seem rather poor engineering practice to place the 75-Ω resistance between the bridge rectifier and the filter capacitors. However, the nature of the controller is such that it can easily compensate when a large load is applied. The resistor also presents sufficient series resistance to prevent overshoot if the load is removed immediately after an SCR has fired. If the power supply had less filter capacitance, a higher resistance would be required in place of the 75-Ω, 100-W unit in order to maintain a similar time constant.

The total filter capacitance is 1500 / 6 = 250 µF. This allows a higher loop gain for better regulation. If less filter capacitance were used, the resistance of the 75-Ω should increase, along with its wattage. In addition, the 5.6-MΩ resistor near the 741 would decrease to decrease the 741's gain. Builders should choose this resistor for the highest resistance that still maintains firm voltage regulation.

Diode D3, together with suitable filtering, provides pure dc voltage for the 741, so it can function without interference from the 13-V "SYNC" signal that appears to the left of D3.

Whenever either the **HIGH VOLTAGE** switch in Fig 3 or S1 in Fig 1 is open, soft-start transistor Q5 is biased into conduction, and the base of Q4 is held low (0 V), together with pin 3 of the 741. RY1 is open. When both of these switches close, RY1 is activated. The base-drive current is diverted from Q5, and a soft start is allowed to begin. If those switches are both open, the coil of RY1 passes base current to Q5, which in turn keeps the soft-start circuit ready, but inhibited.

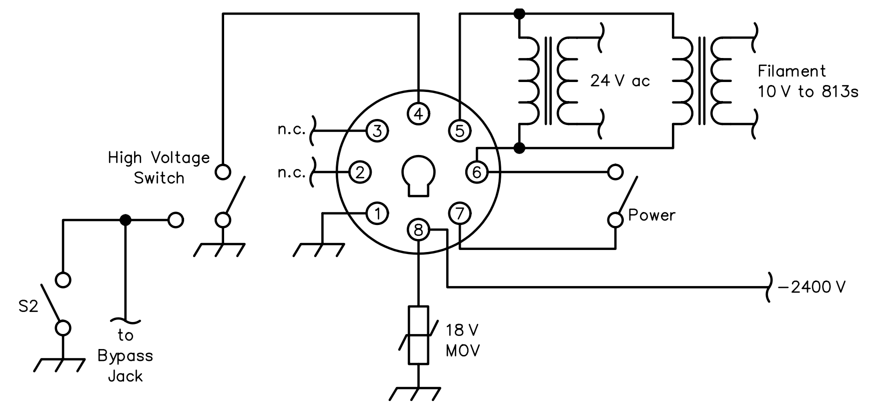

Fig 3—Amplifier end of the power-supply cable.

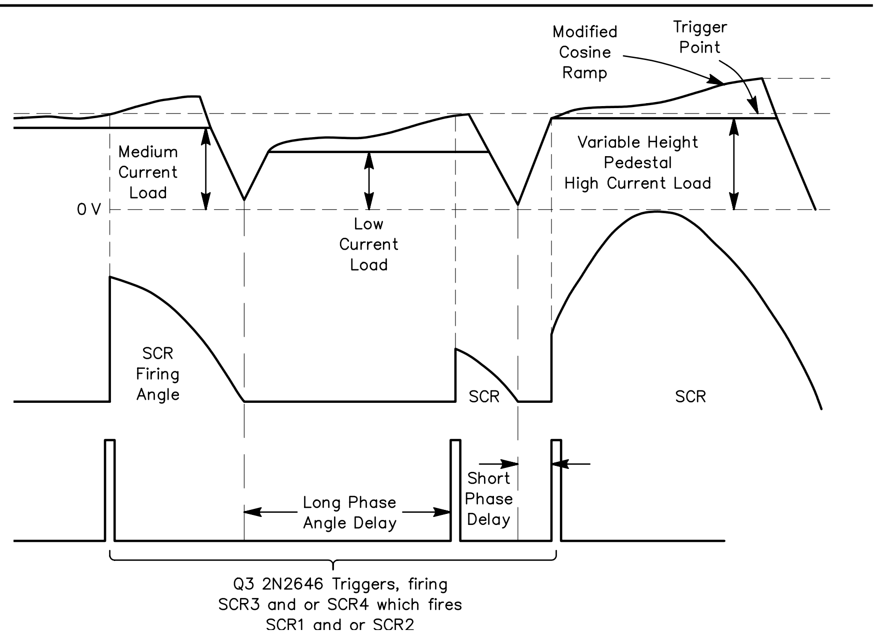

Fig 4—Modified-cosine ramp and pedestal waveforms. The modified-cosine ramp always follows the same curve, but the pedestal height changes based on supply current. U1, the 741 op amp and Q4 (2N3904) control the height of the pedestal to the emitter of Q3 (2N2646) unijunction transistor, which acts to control the firing (phase angle) of the SCRs. The response is in direct proportion to the current output demands on the power supply. T2's average primary voltage is about 26 V ac at a load current of 2.5 mA and rises to an average of about 105 V ac when 560 mA is drawn at 2400 V dc. Load regulation: 3.2 V dc!

The use of small pilot SCRs as drivers for the main SCRs has raised a few eyebrows. This is good practice, however, since it ensures proper firing of the main SCRs. Such operation comes highly recommended by the GE engineering staff, since there are some situations when one SCR will fire perhaps several times in a row, while the opposite one does not fire at all. Because of strong ac components, this often leads to "hammering" sounds, poor regulation, and it can cause damage to the transformer. Using pilot SCRs ensures uniform firing of both main SCRs in the proper sequence.

In case of unforeseen problems, I used a metal-oxide varistor (MOV) between the negative lead of the power supply and ground. It is in parallel with the 47-Ω resistor at the amplifier end of the interconnecting cable, as shown in Fig 3. Incidentally, measuring the resistance between the negative lead of the supply and ground will yield a reading of approximately 1.1 Ω. This is so because a shunt resistor parallels the meter, which has an internal resistance 150 Ω in series with the meter movement.

The resistance of my plate current meter is 0.1 Ω, and a measurement at the positive side of the plate meter to ground reads 1 Ω. I had trouble understanding this figure, but once figured out, I have never forgotten it! The MOV I used is rated at 18 V and seems to provide adequate results, but perhaps a 10-V unit might be a better choice.

Fig 4 gives information about the modified-cosine and ramp generation. Anyone interested in this subject is referred to the GE *SCR Handbook*.

Fig 5 shows the general layout of the control board. There is nothing magical about the arrangement, except that it is a good fit in the space available. Were I to build a second version of this supply, I would seriously consider using AMP or Molex connectors for the new construction. Troubleshooting and assembly would be simpler, but servicing my hard-wired system has not been necessary to date.

It is very satisfying to know that your high voltage is rock-steady, but there are some other advantages as well. The CW output from my amplifier is an exact copy of the drive signal; its envelope is a perfect replica of the keying waveform. Many stations sound different when they switch on their power amplifiers. This is so because the relatively poor regulation modifies the keying rise and fall shapes. (See Fig 6.)

A linear amplifier can be visualized as "distortionless," but poor voltage regulation produces distortion, whether CW or SSB signals are being amplified. Distortion is absolutely guaranteed if poor power-supply regulation exists. SSB voice signals will be degraded because unstable supply voltages produce nonlinear operation in any amplifier. My supply/linear-amplifier combination has no such distortion, whatever the shape of the driving signal; it can be counted on to reproduce it exactly.

Some friends have suggested that I call this supply the "Williams" supply, analogous to automobile racing's McLaren cars. A prospective amplifier buyer could then ask the vender if the gear he is considering has a conventional power supply or a Williams supply!

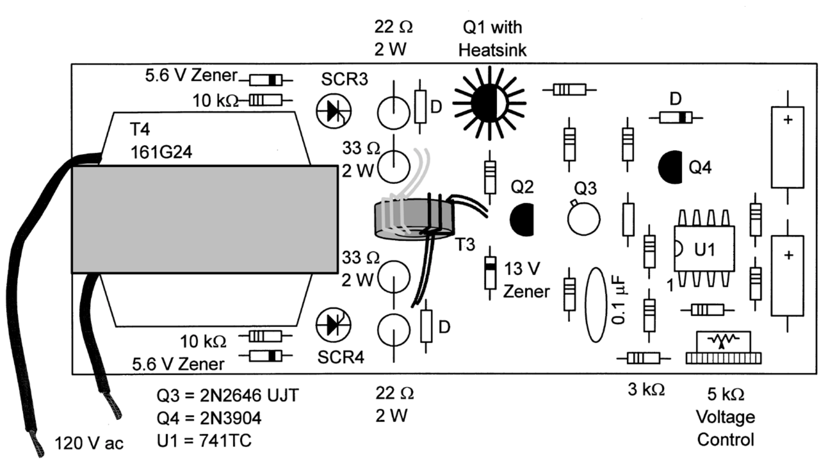

Fig 5—Top view of parts layout on perfboard. Layout should not be critical, but remember: High gain is used in the system, so keep leads reasonably short. I used a small heat sink on Q1 (TO-92 type), but it is probably not necessary. Actual size is approximately 3×7 inches.

Metering and Bias Control

As mentioned previously, my amplifier uses a pair of 813s, but the following could apply to the use of any other tube, and of course, the supply may be tailored to any voltage range.

When at rest with zero bias, 813s (operating in grounded-grid linear service) draw about 50 mA from a 2400-V plate supply. Such operating generates significant diode noise, making full-QSK operation difficult. The tubes also generate a considerable amount of heat during standby (2400 V)(0.05 A) = 120 W.

This is wasted energy. It is desirable to reduce this standing plate current to zero during periods when no RF is desired. This would be between dots in QSK CW operations, or VOX pauses during SSB operation. Fig 7 shows the circuit I use for this purpose.

The jacks in the lower right-hand corner marked "key" and "grid-blocked keying" need not be connected to a negative voltage source. I use grid-blocked keying at my station. It could be a transistor with its base to the positive-to-ground keying systems of virtually any solid-state rig via a suitable resistance, probably between 220 kΩ and 470 kΩ.

The 1-MΩ threshold pot R1 would be adjusted to the point where the V-FET is just nicely turned off with the key open. This would allow the amplifier's cathode to be grounded at the instant of key closure, and it would be ready to accept RF drive.

Fig 7 also shows my bypass/transmit-relay operation. Whenever S2 (Fig 3) *and* the "bypass" jack are grounded, Q4 is off, and the relay is in the **BYPASS** position. Whenever S2 and the **BYPASS** jack are allowed to go above ground, Q4 and Q5 turn on, and the relay is actuated into the **TRANSMIT** position; drive reaches the amplifier's input, and its output is connected to the

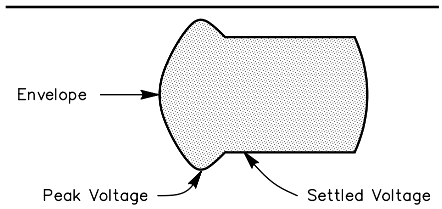

Fig 6—Typical CW output waveform.

antenna system. Note: The KRP11DG relay has a 12-V dc coil. It is supplied from a nominal 24-V source through a 300-Ω resistor. Capacitor C1, a 25-µF, 25-V unit, tends to hold this higher voltage while the relay is being pulled in. Once the relay has operated, the coil voltage will drop to 12 V, to keep the relay activated. This ensures that the relay changes from **BYPASS** to **TRANSMIT** very quickly, and live RF is never switched! This relay "thinks" it is being driven by over-voltage, and for the first few milliseconds, it really does have excess voltage on its coil.

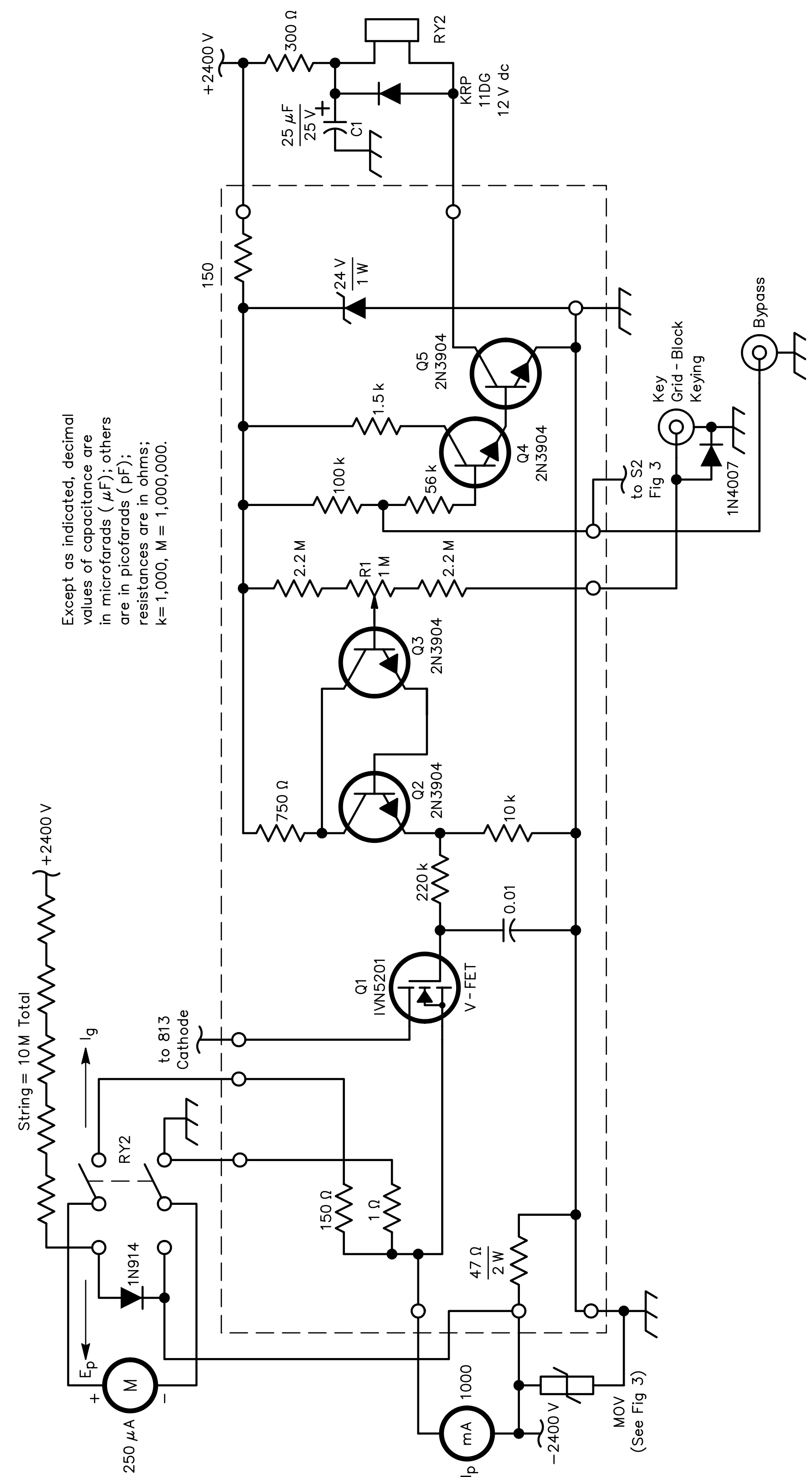

Fig 7—Bias switching, transfer-relay and metering circuit.

Many amplifiers switch so slowly that a considerable amount of RF is switched by the bypass/transmit relay. This produces wide-band clicks during VOX switching. This relay will not follow QSK keying, of course, and another arrangement is needed for QSK operations. It works beautifully for VOX operation, and I use a fast reed relay to open my (separate) receiver antenna circuit during QSK operations.

Fig 7 also shows the metering system. I recommend the use of the "key" jack of Fig 7 to effect power and heat reduction during standby periods whether QSK or just plain VOX is used. This will extend the life of the final tubes.

Several other local hams are now using this power-supply design in high-power amplifier systems, and the amplifiers have performed admirably on both CW and SSB.

Al grew up at Lac Vert, near Regina, Saskatchewan, in the years before WW2. The war interrupted his education (teachers left to fight), so he continued with the National Radio Institute and later took a job with a repair shop in Melfort. Al became VE5CP in 1947, acquired a commercial license in the summer of 1949 and became an operator/agent for Canadian Pacific Airlines.

Al left Canadian Pacific to become a district operator for the Sioux Lookout Fire District, for the Ontario Department of Lands and Forests. He later learned to fly and became a commercial bush and arctic pilot. From 1960 to 1967 he operated an electronics repair firm, Rocky Mountain Electronics and returned to flying for the summer of 1967. On his return, Al began a career in industrial electronics sales. He retired from that vocation in 1992 and became a "professional bum and author." (His book, Bush and Arctic Pilot, *has recently been published by Hancock House Publishers—ISBN 0-88839-433-0.)*

Al began designing circuits for his ham station when Canadian Pacific moved him to Sioux Lookout, Ontario. Al's work has been published in ARRL periodicals before—the last time was in QST *for November 1953. Al has designed and built many projects: a class-C RF amplifier, an automatic antenna tuner, power supplies and a keyer. Al's favorite ham activities are designing/building and CW operation.*

RF

By Zack Lau, W1VT

A 70-cm Power Divider

When stacking identical antennas, many textbook presentations suggest making a power divider out of 75-Ω coax. The math works out quite nicely—a λ/4 transforms the 50-Ω impedance of the antenna to Z_0^2 / Z_{load}, or 112.5 Ω. When two of these are parallel connected, the result is 56.25 Ω, about a 1.1:1 SWR. This is close enough for most people. I'm assuming a 50-Ω reference impedance throughout this article.

In practice, this doesn't work out so nicely on 70 cm. Here, the antennas are so large that a λ/4 just isn't long enough for proper spacing, so you need to consider a 3/4 or 5/4-λ line. Particularly since the velocity factor may shrink the electrical λ/2 to just λ/3 in actual physical spacing. However, errors in line length become more critical—a 1% change in physical length becomes a 3% or 5% change in effective electrical transformer length. One could add 50-Ω extensions, but who wants to go through the expense and reliability problems of additional connectors?

Attaching the proper connectors isn't trivial at higher frequencies and mixed impedances. Some 50 and 75-Ω connectors don't properly mate. If you are very unlucky, a fat, 50-Ω center pin will actually damage a 75-Ω socket. Almost as troublesome is a thin 75-Ω pin making intermittent contact with a 50-Ω socket. One solution is to dispense with the troublesome connectors altogether. Especially for temporary installations, like a weekend EME station designed for use only in good weather, it may make sense to just solder the coax directly together, without any coaxial connectors. The center conductors are soldered directly together. The shields can be connected with copper tape or plumbing hardware.[1]

The best solution I've seen is a power divider made out of square aluminum tubing and round brass tubing.[2] The square shape allows you to easily attach coaxial connectors—one-inch square tubing is readily available at hardware stores and is a good match for UG-58A N connectors. Square tub-

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[1]Notes appear on page 58.

ing is much easier to work with than round tubing, particularly when drilling holes in perfect alignment. Aluminum is a good choice for the outer conductor—lightweight, with excellent electrical conductivity, and it's easily worked with hand tools. The usual problem with aluminum—obtaining good soldered connections—isn't a problem since the connectors can be attached with screws. The center conductor of the custom coaxial line is a different story. I recommend attaching the center pins to the center conductor with solder. Set screws aren't really suitable in an environment where vibration and flexing might be expected. More importantly, you need to pick material dimensions that will result in the proper impedance match.

The formula for calculating the impedance of the $\lambda/4$ matching section is:

$$Z_o = \sqrt{Z_{in} \bullet Z_{out}} \quad \text{(Eq 1)}$$

If two 50-Ω loads are placed in parallel, the resulting impedance is $\sqrt{50 \bullet 25} = 35.4\Omega$

The formula for the square coaxial line is

$$Z_o = 138 \log\left(\frac{1.08D}{a}\right) \quad \text{(Eq 2)}$$

where D is the ID of the square tubing and a is the OD of the center conductor. Alternately,

$$a = \frac{1.08D}{\left(10^{\frac{Z_0}{138}}\right)} \quad \text{(Eq 3)}$$

Thus, for one-inch-square thin-wall tubing with a wall thickness of 0.055 inches, one gets a tubing diameter of 0.53 inches, or 17/32. This may be an advantage of using brass—selections of brass tubing in 1/32-inch increments aren't unusual.[3]

If you use 1/2-inch tubing instead, the impedance is 39.3 Ω. This results in an impedance transformation from 25 to 61.8 Ω, an SWR of 1.23:1. This is a bit higher than the 1.02 predicted for the 17/32 tubing, but may be acceptable. I've measured 19 dB return loss for the 1/2-inch tubing and 30 dB return loss for the 17/32 tubing, which correspond to SWRs of 1.25:1 and 1.06:1. Accurately measuring a 1.02:1 SWR at 70 cm requires precision equipment—most people don't find it necessary to go through the trouble. An advantage of thin brass tubing is the ease of soldering—relatively little heat is needed to get the metal up to the proper temperature, especially since the square aluminum tube effectively shields the operation from drafts. In addition, the thin tubing puts less stress on the soldered joints than does heavier tubing.

It may also be possible to find aluminum tubing with thicker walls, though the selection is often more limited than in the round-brass-tubing case. Nevertheless, I have made a power divider using 1/2-inch brass tubing and one-inch-square aluminum tubing with 80-mil-thick walls. The return loss measured 31 dB. At least 26 dB of return loss was measured on this divider between 410 and 473 MHz.

I strongly recommend using N connectors at 70 cm. UHF connectors aren't a good idea—they typically introduce a 1.5:1 SWR at 70 cm. While this might be tuned out with clever engineering, who needs the hassle? An obvious exception is the use of unusual or surplus cables—you may need to use whatever connector you can get.

The end caps for keeping water and bugs out are made from 3/8-inch sheet Lexan, a shatterproof plastic that is quite UV resistant. Thus, it isn't necessary to protect it from sunlight. The centers of the caps have tapped #8-32 holes—this allows me to remove the caps easily from the tubing by inserting screws and pulling them out. The

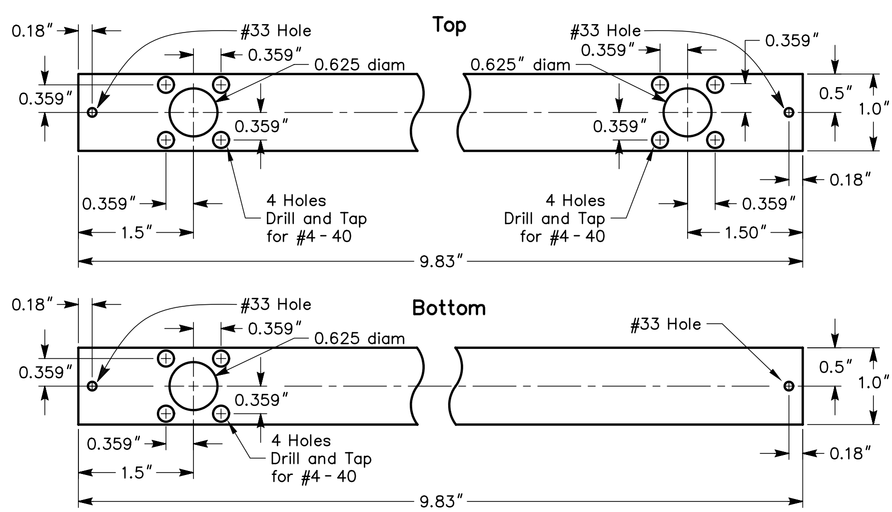


Fig 1—Drilling diagram for the square tubing.

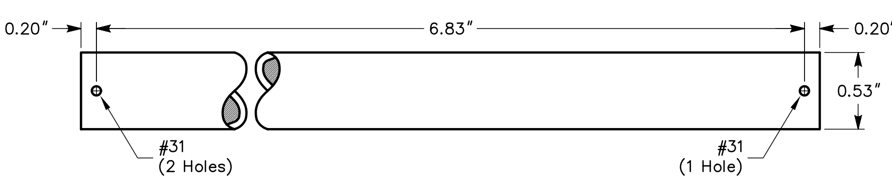


Fig 2—Drilling diagram for the brass center conductor.

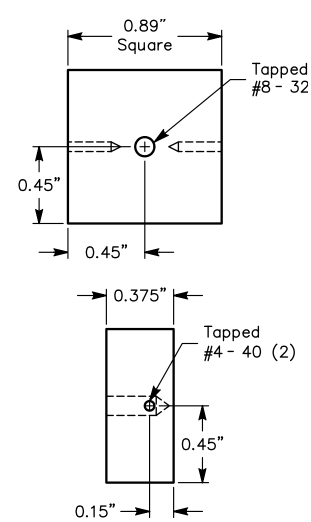


Fig 3—Lexan end plates.

Fig 4—The parts ready for assembly. *Do not* solder the connectors to the center conductor before assembly! Secure one connector at each end to the square tube. Then fit and solder the center conductor to the center pins of the mounted connectors. Install the remaining connector and solder its center pin to the center conductor. Press the Lexan end plates in position and secure them with #4-40 machine screws.

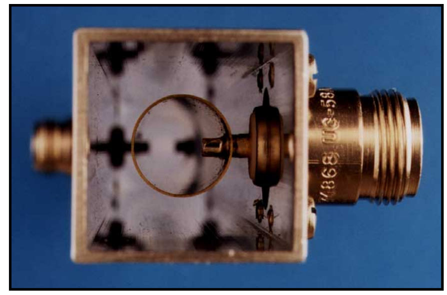


Fig 5—An end view of the completed assembly from the single-connector end. The near, single connector is at the right side of the square tube. One of the connector pair at the opposite end appears smaller and out of focus at the left side of the square tube. Notice how the N connector's center pin penetrates and supports the round center conductor.

holes don't go all the way through the plastic—though this would eliminate the possibility of the caps being attached backwards. A single cap installed backwards can be pushed out with a long stick. The caps are tapped with #4-40 threads so they can be securely attached to the aluminum tubing with stainless-steel screws and lockwashers. The aluminum tubing acts like a waveguide below cutoff. It yields about 30 dB/inch of attenuation. Thus, the end caps won't have much effect unless they get close to the connectors.

Construction

I made the divider about three inches longer than the calculated length of the matching section, so there is plenty of space for the end caps without intruding on the fields of the matching section. After squaring up the ends of the aluminum tubing, I carefully marked off the connector spacing and then the mounting holes for the three UG-58A connectors. Fig 1 is a drilling guide for the square aluminum tubing. It may be necessary to adjust the dimensions slightly to center the connectors, to compensate for slightly thicker or narrower tubing. For instance, with 1.02-inch-square tubing the centerline is 0.51 inches from either edge, not 0.50 inches.

I tried drilling the 5/8-inch holes with a 1/2-inch drill and then a Unibit step drill, but the step drill was too long to enlarge the hole without creating one on the opposite wall. A 5/8-inch Greenlee hole punch was used to finish the job, but these specialty tools are getting pricey. They list for $25.50 in the Mouser Electronics catalog.[4]

Fig 2 shows the drilling guide for the brass center conductor. Drill through both sides of the tubing on one end to accommodate the pair of connectors, while making just one hole on the other end. If the holes fit snugly over the center pins, this will ease center-conductor alignment by holding the tubing in place until it can be soldered.

The center conductor must be accurately centered—unless you wish to lower the impedance of the matching section. I used a 0.2-inch flat-bladed screwdriver tip as a gauge to judge how well the tubing is centered. It's much easier to see how well the tubing is centered when a reference is placed next to it.

Notes

[1]T. Pettis, KL7WE, "Hy-brid Hi-Power," *Proceedings of the 22nd Conference of the Central States VHF Society*, (Newington, Connecticut: ARRL, 1988), pp 149-151.

[2]"World Above 50 MHz," *QST*, Oct 1973, p 97 presents Don Hilliard, W0PW's two and four-port 50-Ω power dividers for 144, 220 and 432 MHz amateur bands.

[3]Small Parts, PO Box 4650, Miami Lakes, FL 33014-9727; 1-800-220-4242; http://www.smallparts.com; smlparts@smallparts.com; has such a selection from 1/16 to 21/32-inch OD.

[4]Mouser Electronics, 958 N Main St, Mansfield, TX 7063-4827; 1-800-346-6873; http://www.mouser.com.

Upcoming Technical Conferences

18th Annual ARRL and TAPR Digital Communications Conference

It's that time again! Time to start making your travel plans and thinking about what to publish for the upcoming 18th Annual ARRL and TAPR Digital Communications Conference (DCC), September 24-26, 1999, Phoenix, Arizona. This year's conference location is just minutes away from the Phoenix Sky Harbor International Airport (PHX).

The ARRL and TAPR DCC is an international forum for radio amateurs in digital communications, networking and related technologies to meet, publish their work and present new ideas and techniques for discussion. Presenters and attendees will have the opportunity to exchange ideas and learn about recent hardware and software advances, theories, experimental results and practical applications. The conference is not just for the digital expert, but for digitally oriented amateurs of all levels of experience.

Not only is the DCC technically stimulating, it is a weekend of fun for all who have more than a casual interest in any of the ham digital-communication modes. This includes networkers, sysops, software writers, modem designers and digital satellite-communications enthusiasts. The ARRL and TAPR DCC is a must-attend conference for those who want to become active on a national level. Now more than ever, Amateur Radio needs this great meeting of the minds. It is important that we demonstrate a continued need for the frequency allocations we now have by pushing forward and documenting our achievements. The ARRL and TAPR Digital Communications Conference is one of the few ways to record our accomplishments and challenge each other to do more.

Call for Papers

Anyone interested in digital communications is invited to submit a paper for publication in the *Conference Proceedings*. Presentation at the conference is not required for publication. The primary purpose of the conference is to communicate ideas and techniques regarding digital communications. Papers written in an informal style are welcome, as well as those written to academic standards. If you know of someone who is doing great things with digital communications, be sure to tell them about this!

Papers are due by August 9th, 1999, and should be submitted to Maty Weinberg, ARRL, 225 Main St, Newington, CT 06111 or via the Internet to **lweinberg@arrl.org**. Information on paper submission guidelines are available on the Web at **http://www.tapr.org/tapr/html/Fdccconf.paper.html**.

Fourth Annual ARRL and TAPR DCC Student Papers Award

ARRL and TAPR especially welcome papers from full-time students to compete for the first annual student papers award. Two $500 travel awards may be given, one in each of the following categories: (a) best technical/theory-oriented paper by a student and (b) best educational or community-oriented application paper by a student. The paper should relate directly to a wireless-digital-communication topic (see guidelines for more information). Papers coauthored by educators or telecommunications professionals are also eligible for this award, as long as a student is the first author.

Deadline for receipt of finished student paper manuscript: July 10th, 1999. Please note that this deadline is different than the general conference submission date. For full details and paper guidelines contact TAPR or download **ftp://ftp.tapr.org/dcc/dcc.student.paper.guideline.pdf**.

Conclusion

If you have attended a Digital Communications Conference in the past, remember how much fun it was discussing the latest developments into the wee hours! If you have never been to a Digital Communications Conference, then make your plans now to attend and find out how much fun they can be.

There are few activities where your participation can be so much fun and so important! You will be able to get together with colleagues from all over the world and bring each other up to date on your latest work. Experience all this and more for an unforgettable weekend of ham radio and digital communications. Make your travel and lodging arrangements now. We hope to see you there.

Full information on the conference and hotel can be obtained by contacting Tucson Amateur Packet Radio, 8987-309 E Tanque Verde Rd, Tucson, AZ 85749-9399; tel 940-383-0000, fax 940-566-2544; e-mail **tapr@tapr.org**; Web **www.tapr.org**.

Western States Weak-Signal Society Conference

The annual WSWSS conference is set for July 24-25 in Flagstaff, Arizona. For more information, contact NU8I at **nu8i@home.com** or check the society's Web page at **www.wswss.org**.

EME Symposium 99

Dave Halliday, K2DH, has announced that Symposium 99-A Beginner's Workshop for EME activity will be held August 20-21 at the Syracuse Marriott in East Syracuse, NY. Dave is still looking for speakers, especially on topics for newcomers. Call Dave at 716-728-9517 or e-mail him at **kb2ah@kb2ah.com** for more information and check the conference Website at **www.geocities.com/~kb2ah/symposium99.html**.

Letters to the Editor

Performance Specifications for Amateur Receivers of the Future

◊ This is generally a good article. (U. Graf, DK4SX, "Performance Specifications for Amateur Receivers of the Future," *QEX*, May 1999, pp 43-49.) However, it should be titled "Design Considerations for Today's Receivers." It does little to define dynamic-range performance standards. In fact, it introduces some confusion at times. However, it does a good job of expressing our desires for a receiver design today.

Although I am all for progress, tough dynamic-range standards already exist in sufficient detail to take care of crunch-proof receiver performance for the next hundred years—that is, if the manufacturers only followed them. This includes radios equipped with DSP implemented not only at baseband, but even at first up-conversion IFs, if so required. DSP at these frequencies is technically feasible today, but not necessarily economical for ham radios.[1] Trying to change these standards is not a good idea.

It is a little-known fact that spurious-free dynamic range (SFDR) was defined in the early 1970s at Watkins Johnson Company (CEI Division) in concert with work done at the Rome Air Development Center.[2,3] This definition includes, but is not limited to, the MDS as *a signal 3 dB higher than the noise floor*, rather than how the article defines it. The MDS is used as the sensitivity measure and affects the IP_3SFDR, IP_2SFDR and BDR.

QST adopted these *de facto* standards a long time ago. They are still with us today. What appears under the *QST* Product Review columns today comprises a very comprehensive and accurate way of measuring any good (or bad) receiver, or any nonlinear device in a signal processing chain, including DSP processors or even a "rock" like a crystal filter.

These methods of testing radios have been published extensively[4] and are still used today. They are tough and comprehensive standards that have served the industry well, and will work with any receivers for many years to come. Various articles, from time to time, have tried to alter these definitions. The real job remains to force industry to use them in light of a better-educated consumer.

One area not emphasized in the article is the *absolute necessity* today for manufacturers to use higher-level and much higher-intercept mixers in the first conversions of receivers, despite all claims of achievement in active commutating mixers. It is unacceptable in today's crowded HF environment to have class-I or II mixers (+7 or +13 dBm LO) in the first conversion of a receiver to reduce cost. Practical design implementations have proved that passive class-III mixers (+27 dBm) surpass their active counterparts despite all the claims. HF is the toughest RF environment, and Europe is the test bed. Note: According to reports, ham radios with SFDRs upward of 105 dB have generally been crushed by the tough European EMI environment.

Of course, more power (Class-A amplifiers) would be used to provide LOs to these mixers, but this is about the only area where we can gain a few more decibels in the SFDR. Ham-radio equipment manufacturers have almost never used +27 dBm Class I (triple balanced) mixers in the first conversions. Why not? Much improvement in SFDR could be gained in this area. The current IP_3SFDR commercial barrier of 105 dB set by the ham-radio equipment manufacturers could indeed be broken, satisfying even the toughest European requirements. The manufacturers could think of this challenge as creating a whole new market.

It is possible to build receivers with 120+ dB SFDR this way, and there would probably be enough hams out there, even in the US, that would appreciate the improvement and pay for it. Yes, they say that hams are cheap, but I think that there are also hams who would like to have the absolute best! Best does not necessarily mean more bells and whistles. Engineering should be simplicity and performance! Then, we could add a few more buttons for flexibility, but not too many.

To conclude, the article also propagates some false information. For instance, an interesting comment talks about designing roofing filters with ultimate bandwidths of 3 kHz at first IFs. Since today's radios are almost exclusively up-conversion types, this statement does not make much sense. IFs of 70 to 120 MHz do not allow crystal filter technology to do better than 10 kHz at the 3-dB points. Unless something changed very recently, roofing filters in today's radios are limited by the laws of physics!—*Cornell Drentea, KW7CD, 757 N Carribean Ave, Tucson, AZ 85748;* **CDrentea@aol.com**

References

1. G. Raghavan, J. F. Jensen, R. H. Walden, W. P. Possey, Hughes Research Laboratories, Malibu, California, "A Bandpass Sigma-Delta Modulator with 92 dB SNR and Center Frequency Continuously Programmable from 0 to 70 MHz," *IEEE International Solid-State Circuits Conference*, ISBN 0-7803-3721-2/97.
2. *Nonlinear System Modeling and Analysis with Applications to Communications Receivers (AD-766278)*, Signatron, Inc., Rome Air Development Center, June 1973, available from NTIS.
3. R. K. McDowell, Watkins Johnson Technical Notes, *High Dynamic Range Receiver Parameters*, Vol 7 No 2, April 1980.
4. Cornell Drentea, *Radio Communications Receivers*, McGraw Hill (#1393 TAB), 1982

Hi Cornell,

Doug Smith sent me your comments to the above article. Thank you very much for your profound objections. There were several reasons to publish this article in an American journal. When it was published in Germany last year, there were no comments on it from the Amateur-Radio industry. Since the European market is negligible compared to the US market, it seemed meaningful to make American amateurs aware of the necessity for specification improvements.

I did not intend to design new definitions for RF receivers. You are absolutely right—they have already existed for a long time, but manufacturers will only follow them when demand rises strongly. Of course, my emphasis lies solely in amateur receivers, not professional ones, and *future* meant at least into the next radio generation. Honestly, I do not know how to force Amateur Radio manufacturers to improve the RF parts of our radios. I would be more than happy if they would finally consider redesigning all this old-fashioned circuitry they have applied for more than 20 years.

Unfortunately, I had some difficulties in making relatively simple technical statements understood. MDS was described as equal to NF, and during the measurement procedure, detected by an increase of signal-to-noise ratio to S+N/N = 3 dB. This is the worldwide definition, but it is of course only one of many possible interpretations, and I don't intend to use a different one. The same is true for SFDR.

All SFDR measurements must be related to a certain bandwidth. This is part of the definition, and is essential for the correctness of the measurement, but is not evident to every user. Unfortunately, the *QST* test staff changed test bandwidth from SSB to CW some years ago. The only result: All newer numbers look a little more "friendly," and the older numbers can no longer be compared to the newer ones. Modern receivers show IP_3SFDRs of up to 105 dB only because of reduction of IF and/or audio bandwidth. They still do not succeed in surpassing a 20-year-old Drake TR-7. Because of this bandwidth reduction, amateurs are deprived of the facts. This is still worse when evaluating radios with digital noise reduction.

So there is a lot to do to make amateurs aware of the technical background of a well designed, modern radio's RF section and the way of evaluating the "specs." Just have a careful look at colorful "data" sheets; even the difference in dynamic ranges is mostly unclear or intentionally concealed. It is just not true that a super-high-level MOSFET mixer is more costly than the ancient quad-FET mixer. In addition, using small relays to choose front-end bandpass filters is even cheaper than quality PIN diodes. Still, today a DSP at reasonable cost is no substitute for quality crystal filters with respect to performance overall.

I absolutely agree with you that there would be a market for radios with ergonomic operation, ie, half as many knobs or sub-menus and superior RF performance. Hams take the "cheapies" because they are there; I think with in-depth education, everybody would like to have top RF performance.

Just some final comments: Most professional receiver designs show first IFs of more than two or three times 30 MHz. As long as you need preselection for IMD_2 reasons (even the best professional receivers have IP < +90 dBm), there is no reason to design an IF > 40 MHz for better response suppression to harmonic mixing products. One of the best receivers I know, the DASA/TELEFUNKEN E1800, with a guaranteed IP_3 of +40 dBm (typical +45 dBm) has a first IF around 40 MHz for IMD_3 reasons. This receiver makes use of discrete quartz crystals to form a first IF filter to avoid IMD_3 encountered with thin, VHF monolithic two-pole filters. So, why use such a high first IF? With preselection presupposed, a first IF around 40 MHz will grant a high IP_3 and the possibility of narrow bandwidths down to 3 kHz. Compare the on-band behavior of the KWM-380 —with *only* +15 dBm IP_3 and an 8-kHz filter in the first IF—to a modern radio.

Openly speaking, there is very little response to an article like this. This is not very encouraging. Nonetheless, as long as experienced hams with the knowledge necessary will work side by side to finally reach improvements, the amateur community will gain from the effort. In this sense, thank you again for your reply and ideas. BTW, I only recently got your book (Ref 3), and I studied it with great pleasure! Best regards—*Ulrich Graf, DK4SX, Seidlheck 19, D089081 Ulm, Germany;* **ulrich.graf@ulmail01.europe.nokia.com**

Hello Uli,

Yes, I agree with you. For the expense of a [very-high-dynamic range] transceiver design, the market just isn't there. Unfortunately, these [lower-dynamic-range] radios were designed for the US where EMI is not as big a problem as it is in Europe. You in Europe must use what is there.

About the 40-MHz versus 70-MHz first IF: The *image* kills the 40-MHz approach. As you go up in the received frequency, say towards 20 MHz, these 40-MHz IF radios will suffer interference from FM (UKW broadcast stations) even with good front-end filtering. You must trust me on this, I speak from experience.

Thanks for a good discussion. Yes, at these frequencies, relays are cheaper than PIN diodes, but maybe not cheaper than silicon diodes. You must know that the industry's flagships are still using silicon diodes in the front end. Maybe they changed that lately?—*Cornell, KW7CD*

Hi Doug,

As you predicted, I have many comments on Ulrich Graf's "Performance Specifications for Amateur Receivers of the Future." This article is a mixture of technical requirements and taste-based requirements. The former are solid, and the latter vary from person to person.

For example, Graf has no need for computer outputs from his receiver. I do need computer output from my receiver, for RTTY, PSK31, SSTV and so on. I set up my homebrew receiver to have a control panel with few controls, as Graf recommends: Four knobs, three buttons and one numeric display. This design, using a DDS for its VFO, has permitted me to experiment with a wide range of tuning rates.

I split the tuning rate into two parts: tuning steps per turn of the knob and frequency change per step. I have 512 steps/turn to simulate continuous tuning and 16 steps/turn to simulate channelized discrete tuning. I have step sizes from 1 MHz/step for band changing down to 0.1 Hz/step for precise tuning to WWV and such. The 512 steps/turn, combined with 50 Hz/step, gives 25.6 kHz/turn to simulate a traditional analog tuning knob.

In tuning the amateur HF bands, I have found that all the SSB signals are on multiples of 500 Hz and that over 95% are on multiples of 1000 Hz. Thus, I have found a channelized tuning very convenient—16 steps/turn to give discrete channels and 500 Hz/step so every SSB signal is on a channel. Note that the channels overlap, since the SSB bandwidth is wider than 500 Hz.

Mr. Graf states: "You won't carry your (future) receiver around in your jacket pocket!" due to the large size of the coils in the receiver's many filters. Once the 120-dB RF ADCs appear, you *will* carry your receiver in your pocket; the single-input low-pass anti-alias filter needs only a few large coils, and all the other filters will be implemented digitally in the DDC chip.

I agree with the Editor's sidebar, "A Better Mousetrap?" If I were in a strong-signal-interference situation, I would opt for a DDC receiver, accepting the limits of present ADCs, and use a passive, tunable preselector on the input to knock down the largest interfering signals and avoid ADC overload. Of course, this doesn't work if the desired signal is very close to a very strong interferer.

Those who use the small, high-Q tuned-loop HF antennas are enjoying the benefits of a preselector, since their antenna bandwidth is very small. *—Peter Traneus Anderson, KC1HR, 625 Main St, Apt 27, Reading, MA 01867-3006;* **peteand@vitech.com**

Hi Peter,

As long as 120-dB dynamic range of the ADC not only means 120 dB wanted signal dynamic range but

SFDR, I agree. Consider this however: An ADC is an analog component, and everything that is challenging for an analog receiver is true for this part as well. Still worse: You need considerable—and extremely linear—preamplification, and all distortion created in such a broadband ADC is *in-band* distortion. So the 120-dB SFDR receiver with a broadband RF ADC is—at this moment—not future, but pure fiction. Honestly, I'd like to find an improved receiver technology within my lifetime. With the parts to enter the market in the next decade, you must use either preselection to reduce only some of the severe distortion, or live with a very limited SFDR. For now, I still prefer an improved "classic" analog front end.

You do not need computer control for your receiver to enjoy the mentioned modes. A connection of your receiver's audio to the sound card input of your PC or modem will suffice.

Congratulations on the fine homebrew receiver with the versatile tuning system, but I question the results of your experience. You are probably able to detect an SSB signal every 500 Hz or 1 kHz, but is it correctly tuned then? What about CW? Frequency offsets of greater than 100 Hz are annoying, and manual tuning is always essential to correct discrepancies in reference oscillator frequencies. Scanning modes are ineffective in Amateur Radio.—*Uli, DK4SX*

Hi Uli,

For the highest receiver performance right now, analog is the way to go. It has been exciting to me to compare vastly different technologies to do the same receiver job. For forty years, I have been experimenting with different ways to build receivers.

I love analog-receiver design, and appreciate your articles. I also see the long-term possibilities of digital design, and I feel it is important for Amateur Radio experimenters to become aware of its possibilities.

You are quite right about "everything that is challenging for an analog receiver is true for this part [an ADC] as well." Yes, the 120-dB SFDR ADC is quite a challenge, but I would not be so bold as to say it is impossible: Analog chipmakers do amazing things. I have studied what goes into these ADCs, and considerable improvement is definitely possible.

Unfortunately, the big market—cellphone base stations—only needs 80 dB of SFDR, and ADC makers and DDC makers have stopped there for now. I vaguely recall seeing a new cell-phone system that will require more SFDR.

On preamplification before the ADC: Over the years, as the ADC dynamic range has increased and the ADC full-scale input voltage has decreased, the amount of preamplification needed has decreased. For a sampling ADC, some preamplification will always be necessary to overcome the aliased, out-of-band noise within the ADC. For a delta-sigma ADC, there is no such problem, and RF delta-sigma ADCs will eventually need no preamps.

The 500-Hz or 1000-Hz steps were specifically for SSB. For CW, much smaller steps are necessary, of course. Step size is a matter of individual preference. For myself, and my fingers on the knob, the stepped tuning works better than continuous tuning. I can't hear the difference on an SSB signal mistuned by 50 or 100 Hz. Within this tolerance, I have yet to find an SSB signal that is not on a multiple of 500 Hz. I originally used a 1000-Hz step, and found a few signals on multiples of 500 Hz.

An old or homebrew rig with a purely analog VFO, or some of the Ten-Tec rigs with analog VFOs, could transmit on odd carrier frequencies. Thanks again for sending me your comments.—*Pete, KC1HR*

Gentlemen

You have a great discussion going, and I thank you for it. We owe credit to CQDL and DARC, who ran a version of Uli's article (in German) over five segments in late 1997 and early 1998. Thanks to Michael Link, DL2EBX, for the help.—Ed.

Editor's Sidebar: A Better Mousetrap?

◊ I don't understand the rationale behind your comment at the top of page 49. I would think that articles like "Performance Specifications for Amateur Receivers of the Future" would be at or near the top of the priority list for being published in *QEX*! Regards. —*John Montague, W9JM, ARRL TA, 818 Adger Rd, Columbia, SC 29205-1912;* **john.montague@att.net**

Thanks for the note, John. The sidebar was meant to indicate we don't often receive or run speculative material, not that we did so with trepidation. Indeed, we felt Uli's piece was important to publish as a catalyst for discussion. We regret any inferred derogation.—Ed.

A Digital Commutating Filter

◊ The May/June 1999 *QEX* provoked a lot of responses from me. I liked all the articles. Mike Kossor's "A Digital Commutating Filter" covered the subject well. The design can be significantly extended: Add a second commutator connected to the same set of capacitors, and take the output from the second commutator. Run the two commutators at different frequencies, and find that the input passband is centered at a frequency set by the input commutator; the output passband is centered at a frequency set by the output commutator. Thus, we have a device that shifts the signal frequency, an SSB generator or receiver.

The frequency-shifting commutating filter is a generalization of the Weaver method of SSB generation. *QEX* published an article of mine in 1991 ("A Different Weave of SSB Exciter", *QEX*, August 1991). It describes a 75-meter SSB exciter using a frequency-shifting commutator. With two commutators, the capacitors can be replaced with multipole filters for better phase response in the passband and steeper skirts in the stopband.

To be exact, the Weaver method of SSB is a frequency-shifting commutating filter using four branches, where each branch is a multipole low-pass filter rather than a single capacitor. The four branches are +I, +Q, –I, and –Q, in order. Only two physical filters are used, as +I and –I are combined into one signal in the balanced mixers, as are +Q and –Q.

The real-output mode of the Harris HSP50016 digital down-converter (DDC) operates just this way, as a close study of the datasheet will reveal. A four-input multiplexer switches through +I, +Q, –I, –Q sequence one step each output sample. This makes the center of the real-output passband exactly one fourth of the output sample rate.—*Peter, KC1HR*

Doug,

The reference to Fig 7 in my article should be to Fig 2. Eq 12 is correct, but it is not in the simplified form I proposed in my original manuscript:

$$\left|H(j\omega)\right| = \left\{1 + \frac{2h\left[1-\cos(\omega nt)\right]}{\left(1-h^2\right)}\right\}^{-\frac{1}{2}} \qquad \text{(Eq 12)}$$

Fig 2 appears incorrect in my copy of *QEX* because all the peaks are not shown with an amplitude of 1. The plot in my copy appears to have the amplitudes of 0.3, 1.0, 0.7, 0.65 and 0.9 at frequencies 0, 500, 1000, 1500 and 2000 Hz, respectively. I would guess the plot misprint is due to the resolution capabilities of the printer and *not* the data.

One minor detail: The schematic of Fig 1 shows larger switch contact gaps between capacitors C1-C8 and C4-C5. This may incorrectly imply a longer switch time between these capacitors, which is *not* the case. The switching time from capacitor to capacitor is the same in the physically switched implementation. The switching time is nearly instantaneous in the modern implementation.—*Mike Kossor, WA2EBY, 244 N 17th St, Kenilworth, NJ 07033;* **mkossor@lucent.com**

A 300-W MOSFET Linear Amplifier for 50 MHz

◊ As soon as the article hit the street, my e-mail box started to fill. It's Monday morning and I have 27 to read. Every one of the responses so far has been: "Can I buy the kit?" I put together a FAQ sheet on the article to limit the amount of correspondence, and also made a .PDF 1:1 negative artwork for those who said that was acceptable.

My home e-mail address is wrong in the print article, and there are two Fig 5 artworks and nothing of the graph with IMD. That's not bad considering all the work done to get the copy ready.—*Dick Frey, K4XU, 405 SW Columbia St, Bend, OR 97702;* **k4xu@coinet.com**

We have considerable egg on our faces over those errors and omissions, Dick. It looks like we tried to put you in Tel Aviv with your home e-mail address. Below is the correct Fig. 6 showing the IMD performance. Thanks for your patience and understanding.—Ed.

Hi Dick,

Just a short note to say that I really enjoyed your paper in *QEX*. After reading the paper I went back and looked at the referenced paper in *Applied Microwaves and Wireless*, as well as APT9701 and 9702. Frequent mention is made of the devices being inexpensive, but how much do they cost? Of equal importance, where can one buy them? Does APT have a distributor yet? KK7B and I (plus a chapter from W7PUA) are in the process of writing an update for *Solid-State Design for the Radio Amateur*. I'd like to be able to reference your parts if hams can get them.

The biasing problem is an interesting one. While I have built several amplifiers with HEXFETs, I've not tried to extract power that would tax their thermal performance. I also wonder if there might be some interesting monolithic solutions. I was able to do some thermal compensations in some GaAs ICs, although those were not power parts. Keep up the good work.—*Wes Hayward, W7ZOI, 7700 SW Danielle Ave, Beaverton, OR 97008;* **w7zoi@mail.teleport.com**

Hi Wes,

It's been a long time since we last talked.

Parts availability: As RF parts go, they are cheap, about \$0.18/W at the single-piece level. The APT stocking distributor in the US is Richardson, **www.rell.com**. The ARF448 is about \$40 in single piece, and \$20 at 10k.

There is a new part in our stable, ARF450. Check our Web page, **www.advancedpower.com**. It is a ceramic push-pull pair of ARF449-like devices with 1 kW of P_d. Cost will probably be about \$100 at the 10k level. I am working on a test fixture right now. It should easily do 500-W out per part, indicating a fully capable 2-kW linear with just four parts.

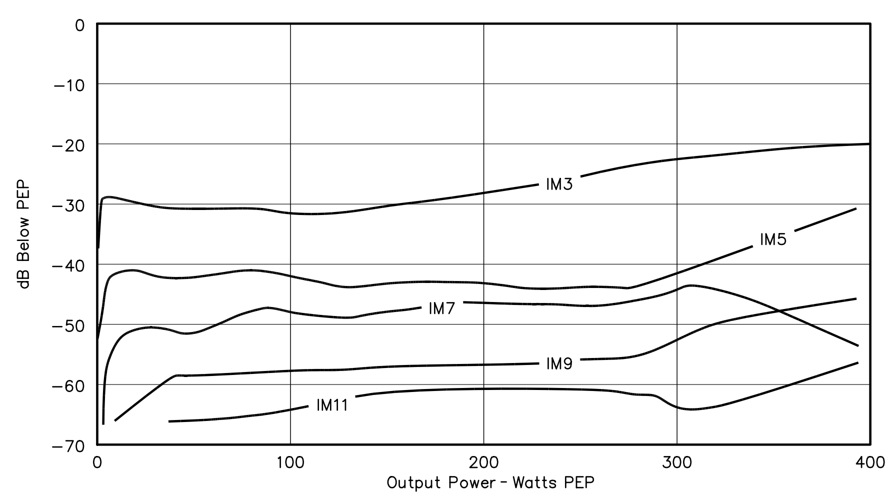
With IR or IXYS parts, the hex-cell poly-silicon gate conductors can't take the RF current needed to drive the gates at much over 14 MHz and small power. The gate fails under the bond wire junction to the top metal where the current density is highest. Due to the difference in port impedance, it's possible to have more RF current in the gate than the drain in a FET amplifier. APT RF parts use self-aligned metal gates with a multiple-comb structure. The ESR of the gate impedance is less than 0.1 Ω. That makes it harder to match and drive, but it won't fail. A perfect MOSFET would have no ESR.

The V_{th} temperature coefficient is due to the innate perversity of silicon's physical properties, and probably is why nobody has pursued high voltage RF before us. I have an experiment using special silicon "epi" material in my next engineering batch of FETs. It will take a higher implant dose to maintain the G_m, but I expect the crossover current to come way down. That's Fig 3 on one of our data sheets. Yes, I am working on the problem.

The other problem is again basic physics. You notice I reduced the drain voltage for the linear amp. It helps in an area I didn't mention—bandwidth. R_L rises with the square of voltage. C_{oss} falls only very slightly. The Q of the output rises with R_L. That makes it tough to go from 1.8 to 50 MHz in a broadband amplifier without compensating for the C_{oss} somehow. For the moment, I am hanging in at 75 V. That's cool, since it gives two times the output power for the same load line on a 50-V amp.—*Dick, K4XU*

A Calibrated Panoramic Adapter

◊ Bob Dildine uses a surplus Millen oscilloscope module in "A Calibrated Panoramic Adaptor" (May/June 1999, pp 9-22). Those who wish to build the scope from scratch should check out Antique Electronic Supply (AES). AES has 1EP1 CRTs for \$23.50, and 7- or



9-pin miniature tube sockets. The 1EP1 has an unusual 11-pin base, but the contacts from 7- or 9-pin sockets work fine. The same base is used on some ceramic-metal RF power tubes, so Johnson made a socket that will fit. The 1EP1 characteristics are in the tube tables of *The 1963 ARRL Handbook*. The 1EP1 was introduced by RCA in 1956, so it is in some RCA industrial tube manuals.—*Peter, KC1HR*

Doug,

The article on the panadapter is really neat. I'm in good company. —*Dick, K4XU*

RF: Transmission Lines and Amateur Radio Designer

◊ I found a "typo" on p 58, near the bottom of center column: "b = ID of inner conductor" should read "b = ID of outer conductor." Regards—*John, W9JM*

A Pair of 3CX800s for 6 Meters

◊ There were several errors in the schematic in my article (Jan/Feb 1999), as follows:

- K1A contacts are reversed, after a three-minute delay, the contacts should *make*
- Q2 and Q3 emitters/collectors are reversed
- K2A should be shown in the normally closed position so that +24 V dc is removed when grid trip pulls K2 in
- S3 should be shown as normally closed.

I've had quite a few calls pointing these things out.—*Dick Hanson, K5AND, 7540 Williamsburg Dr, Cumming, GA 30041;* **k5and@prestige.net**

Next Issue in QEX

We got a bunch of great feedback about Mark Mandelkern's high-performance homebrew transceiver. Mark has continued to document his rig, and presents Part 2 of the series. Come along and dive into the IF details where the emphasis remains on top performance. Additional segments will follow soon, providing more particulars about this ambitious project.

Those readers gearing up for P3D and other UHF or microwave projects will find lots of good information in Paul Wade, W1GHZ's article "Parabolic Dish Feeds—Phase and Phase Center." Paul discusses critical factors in the location of feeds for dishes, with the "focus" on gain and beamwidth.

We are fortunate to have an article about digital-radio design by Brad Brannon; it originally appeared in the November 5, 1998 issue of *EDN* magazine. This piece concentrates on dynamic-range issues confronting designers of IF-DSP and digital direct-conversion (DDC) transceivers. Brad also illustrates why his techniques are attractive in cellular base stations.

Peter Martinez, G3PLX, is a busy guy! His "Using Doppler DSP to Study HF Propagation" comprises state-of-the-art ionospheric analysis. His topic is at the leading edge of a field that many physicists are currently exploring. In various ways, DSP is being used to map the characteristics of Earth's atmosphere many hundreds of miles from its surface. These studies will undoubtedly lead to a better understanding of our environment, and perhaps they constitute a new field of research for Amateur Radio experimenters.

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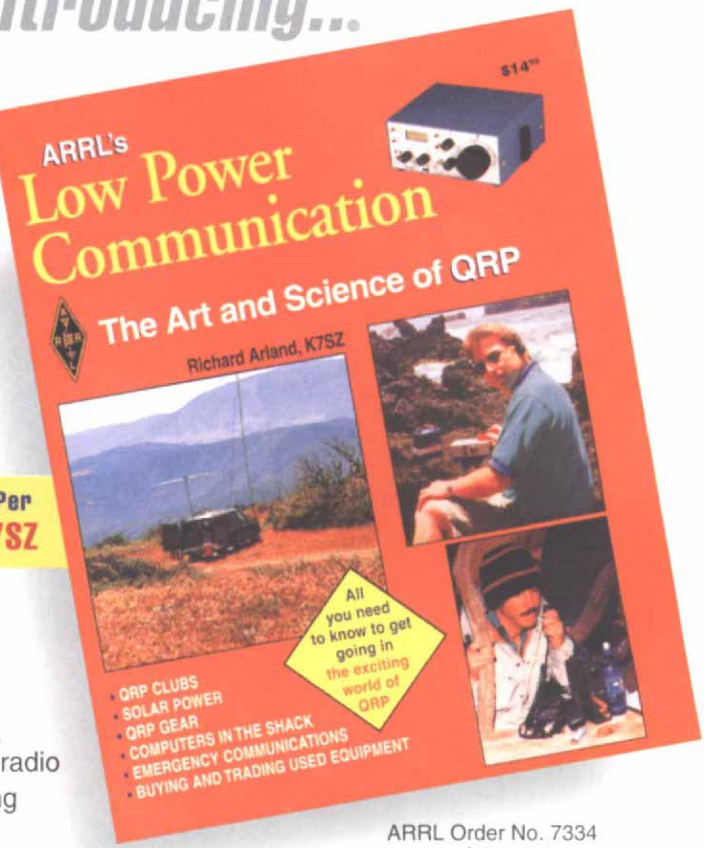

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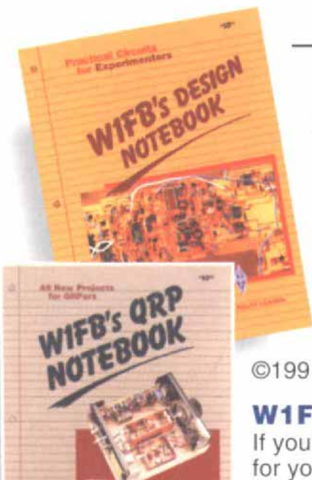

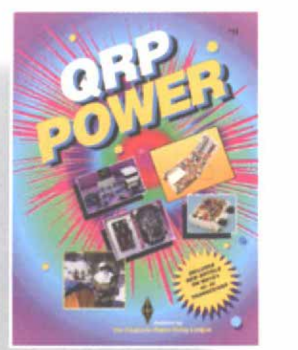

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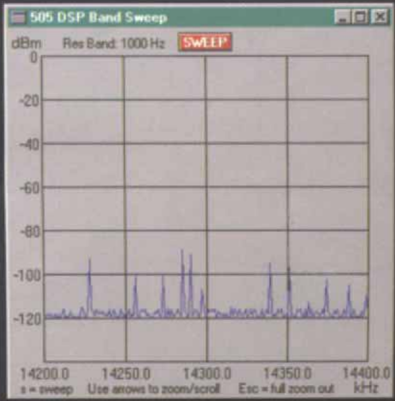
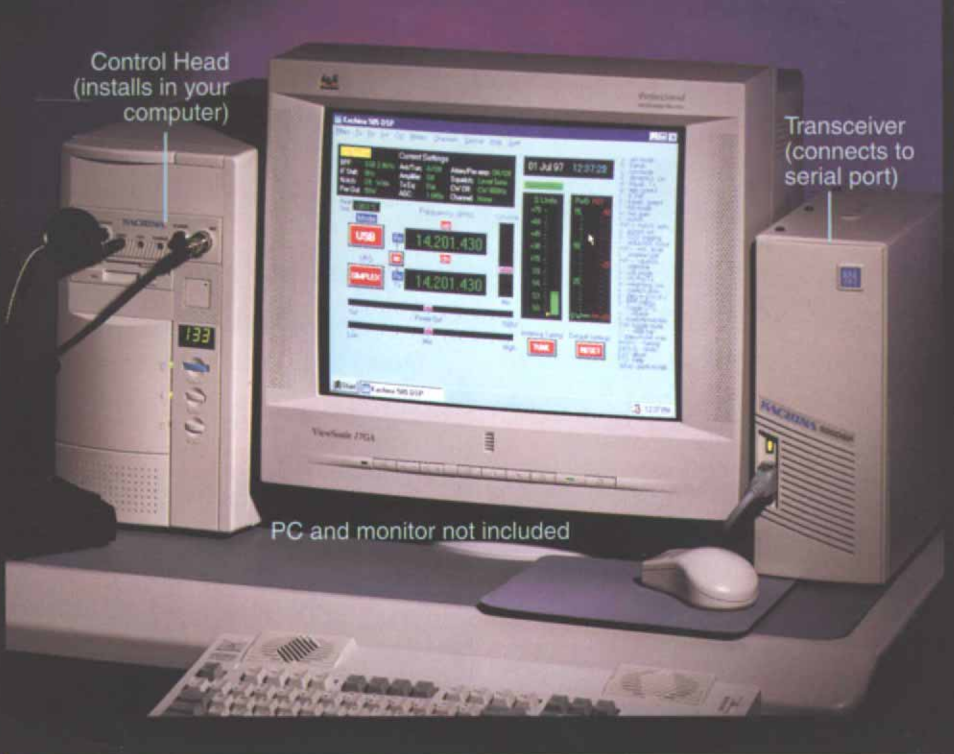
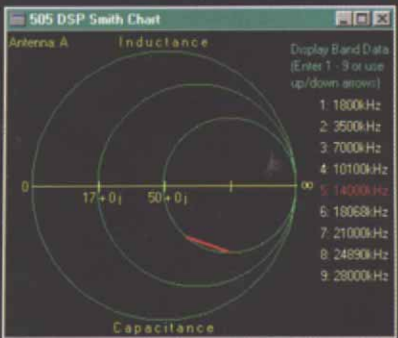


"Band Sweep" window allows you to look for the pile-ups. "Point and click" the mouse to instantly change frequency and join in.





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