

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.

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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 spec-

¹Notes appear on [page 26](#).

trum 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 needs in the future.

In addition to reducing harmonic products, the filter input should present the correct $50 + j0 \Omega$ 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 homebrew MOSFET power amplifier.

Another consideration is freedom from oscillations and significant regeneration. Oscillations can be free run-

ning; 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

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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 pre-

ferred 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

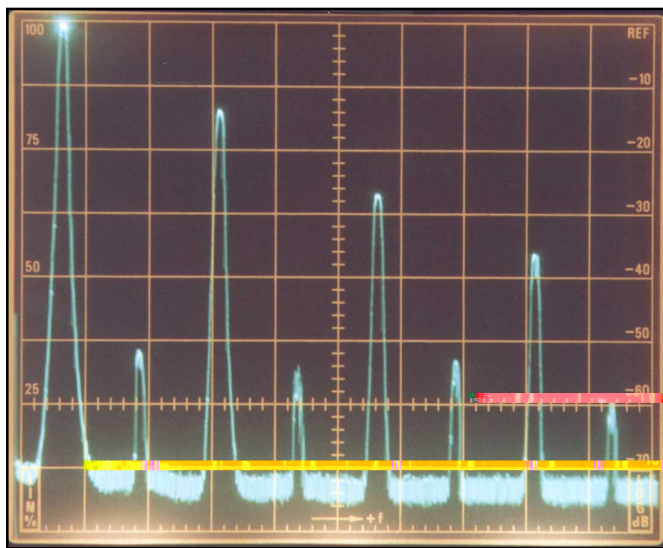


Fig 1—Spectrum-analyzer plot of a 120-W amplifier with no low-pass filtering. Fundamental frequency is 3.5 MHz and the sweep is 1.8 to 30 MHz, 10 dB/division.

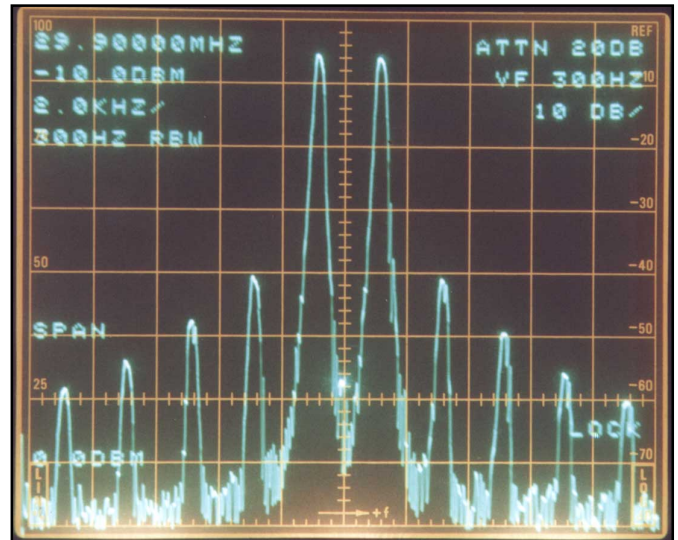


Fig 2—Two-tone intermodulation products at 29.9 MHz, 120 W PEP.

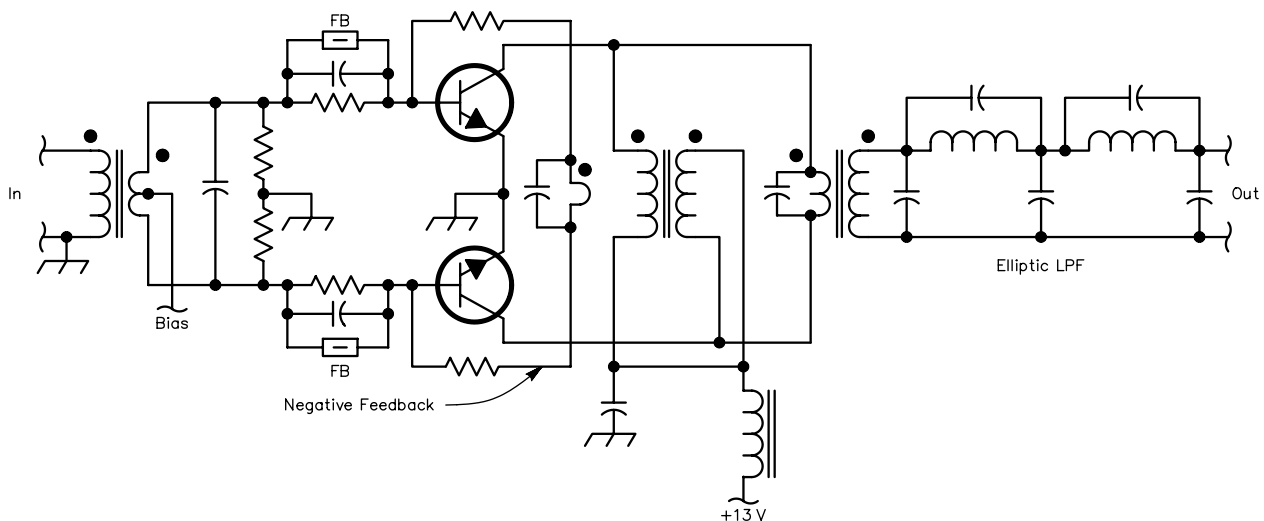


Fig 3—Typical transistor power amplifier circuit arrangement with conventional transformers and feedback.

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

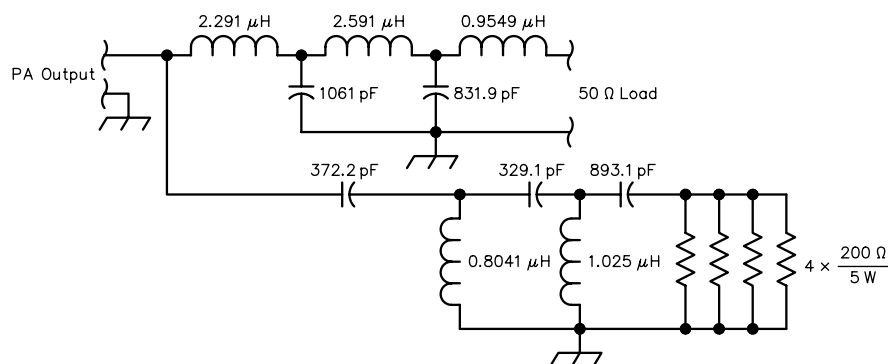


Fig 4—Diplexer filter for the 80-meter band with exact values.

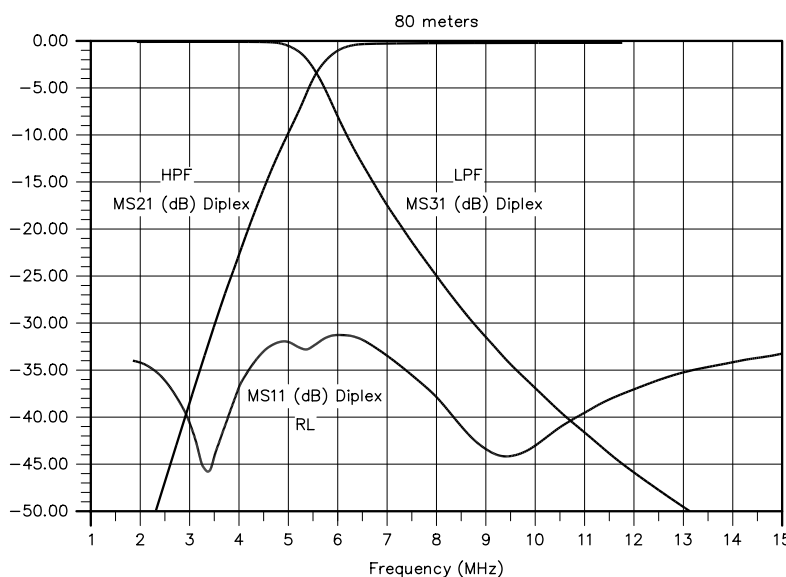


Fig 5—Computer simulation of the 80-meter diplexer.

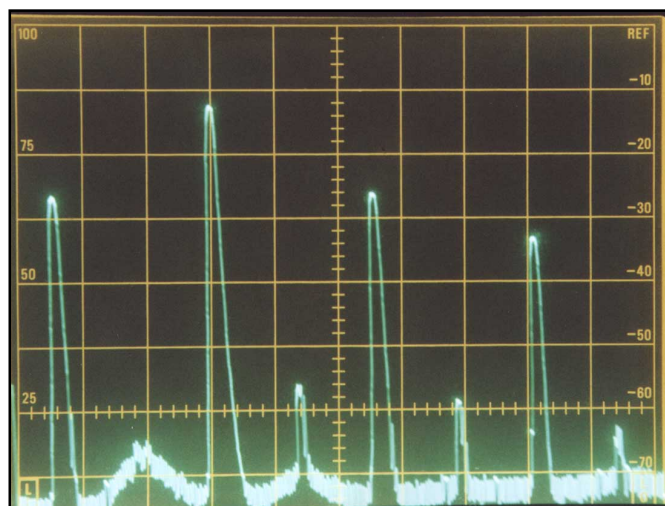
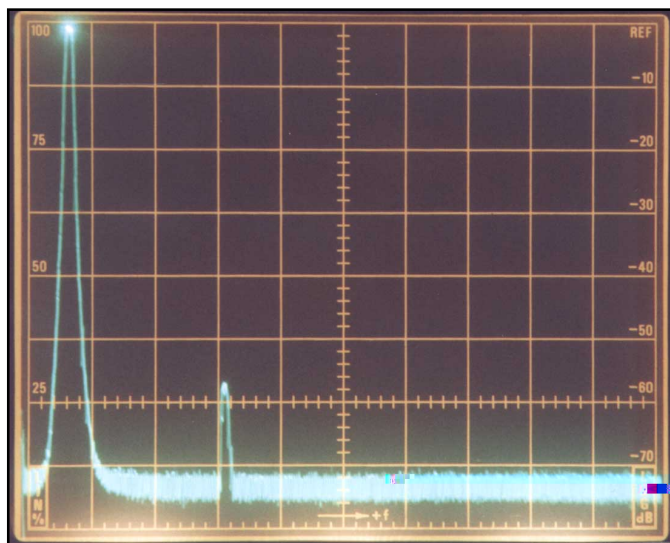


Fig 6—Same spectrum as Fig 1 except: (A) at output of low-pass filter, (B) at output of high-pass filter.

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:

$$L_{LP} = \frac{K L_{P(LP)} R}{2\pi f_{co}} \quad (Eq 1)$$

$$C_{LP} = \frac{K C_{P(LP)}}{2\pi f_{co} R}$$

where $L_{P(LP)}$ and $C_{P(LP)}$ 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:

$$L_{HP} = \frac{L_{P(HP)} R}{2\pi f_{co} K}$$

$$C_{HP} = \frac{C_{P(HP)}}{2\pi f_{co} K R} \quad (Eq 2)$$

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.

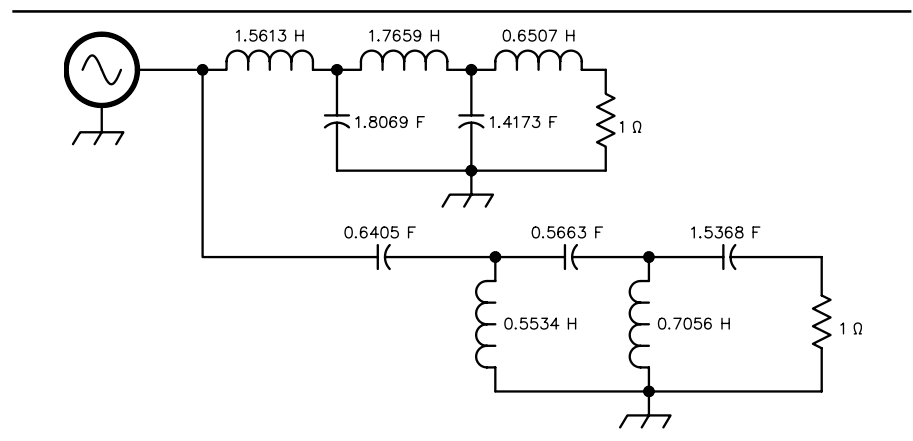


Fig 7—Low-pass and high-pass prototype diplexer filter.

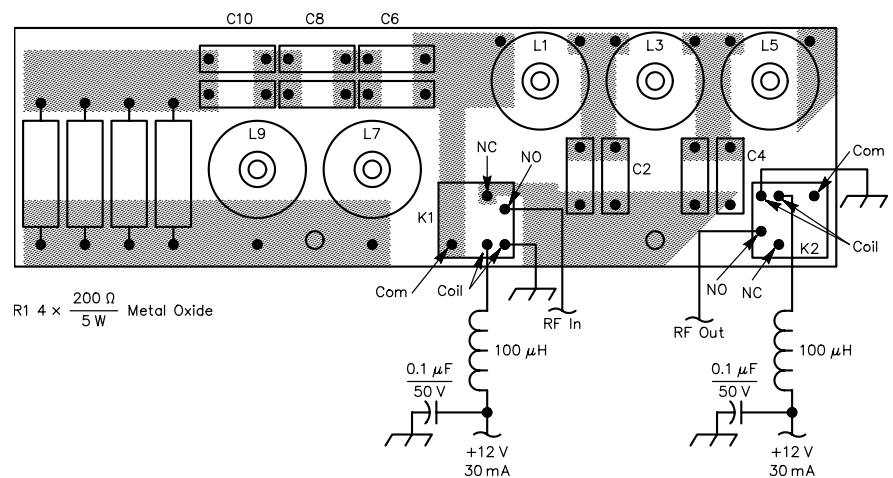


Fig 8—Diplexer PC-board layout.

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 duplexers. 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.

Duplexer Construction

Fig 8 shows the PC-board layout of a duplexer filter. It is built on a $2 \times 8\frac{3}{8}$ -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 mar-

gin. 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 $\frac{3}{8}$ -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

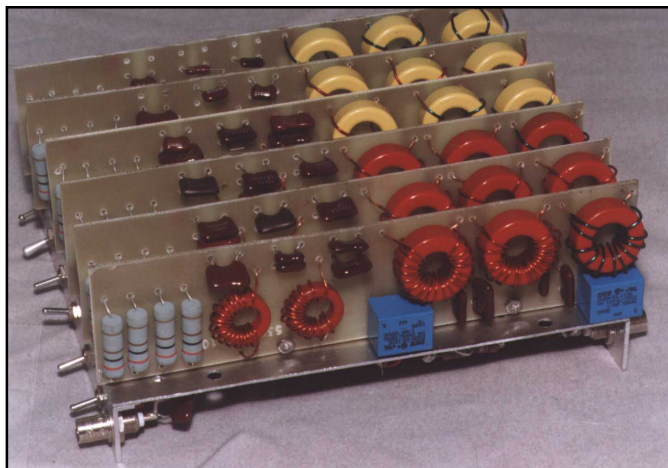


Fig 9—Photo of complete duplexer filter assembly for all nine HF amateur bands.



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.

prepared⁸ 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

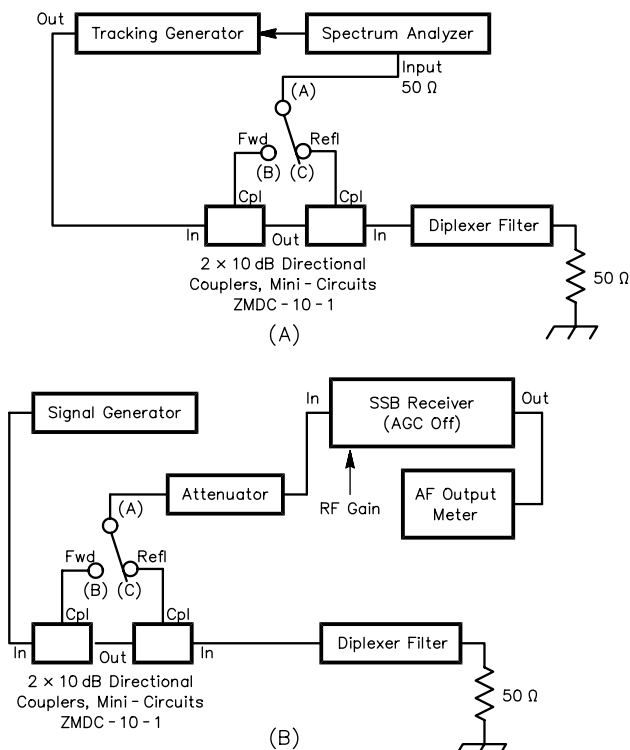


Fig 11—(A) Spectrum analyzer and tracking generator to measure and adjust diplexer return loss. (B) Signal generator and SSB receiver to measure adjust diplexer return loss.



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

¹W. 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 Blocksme. *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/>.

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

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

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

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

⁶J. 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.

⁷A 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.

⁸You 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. □□

