THE BUTTERWORTH FILTER COOKBOOK

BY DONALD LANCASTER*

Part I

Below are some recipes for painless filter design based on Butterworth polynomials. By choosing the basic circuit and following the directions for multiplying and dividing, most low pass, high pass and band pass filter problems can be solved rather easily.

THE designing of filters for any electronic circuit always seems to involve a bit of deep mystery and perhaps require a hand or two dabbling in the black arts. A powerful new class of filter circuits with remarkable properties is easily designed and applicable to the majority of electronic filter problems. By choosing one of the basic circuits given here, followed by a simple division and multiplication, the majority of low pass, band pass, and high pass filters can be immediately solved. And the results are in actual component values, ready for immediate breadboarding and circuit verification. Further, the results are always in good agreement with the original design and are exactly predictable.

These filters are called *Maximally Flat Amplitude* or *Butterworth* filters. Maximally flat simply means that as much of the filter's passband is made as flat as possible. Since the math behind these filters uses *Butterworth Polynomials*, the name is carried over.

There is no such things as a perfect filter. Even one with an infinite number of perfect inductors and capacitors would have a significant drawback: It would take infinitely long for a signal to pass from one end of the filter to another! Thus, any filter, however good, must result in some compromise. It will be ideally suited for some applications and misapplied in others. Here's how the Butterworth filter performs:

Advantages

1. The passband of the filter is always defined to points at which the response is precisely 3 decibels attenuated. This is the *cutoff frequency* or frequencies of the filter.
2. The passband is very smooth and flat with no lumps, ripple, or overshoot.
3. The further one ventures into the stopbands, the higher the attenuation becomes.
4. The attenuation at *any* frequency is accurately predictable from a simple curve.
5. Lossy (finite "Q" and "D") components do not drastically affect performance, but show up mainly as insertion loss and changes in skirt steepness.

Disadvantages

1. For extremely steep skirts, other forms of filters (Tschebyscheff) will do the same job with fewer parts.
2. There are no points of infinite rejection near the cutoff frequency as might be required if a signal near the passband must be sharply rejected. Thus Butterworth filters are not well suited to single sideband filters and similar applications.

Low Pass Filter Types

Let's see how a Butterworth filter's amplitude response will compare with an ideal filter, as in fig. 1. An ideal low pass filter passes all frequencies from d.c. to a specified cutoff frequency without any loss or attenuation. Beyond cutoff, all signals are infinitely rejected. The Butterworth response will be uniform over most of the passband, but will be attenuated exactly 3 decibels (0.707 voltage or 0.500 power) at the cutoff frequency. The response then continues to decrease as frequency increases.

The Butterworth low pass filter takes the form of a *ladder* network consist of *series* inductors and *shunt* capacitors. At high frequencies, each shunt capacitor provides a very low reactance to ground and each series inductor presents a very

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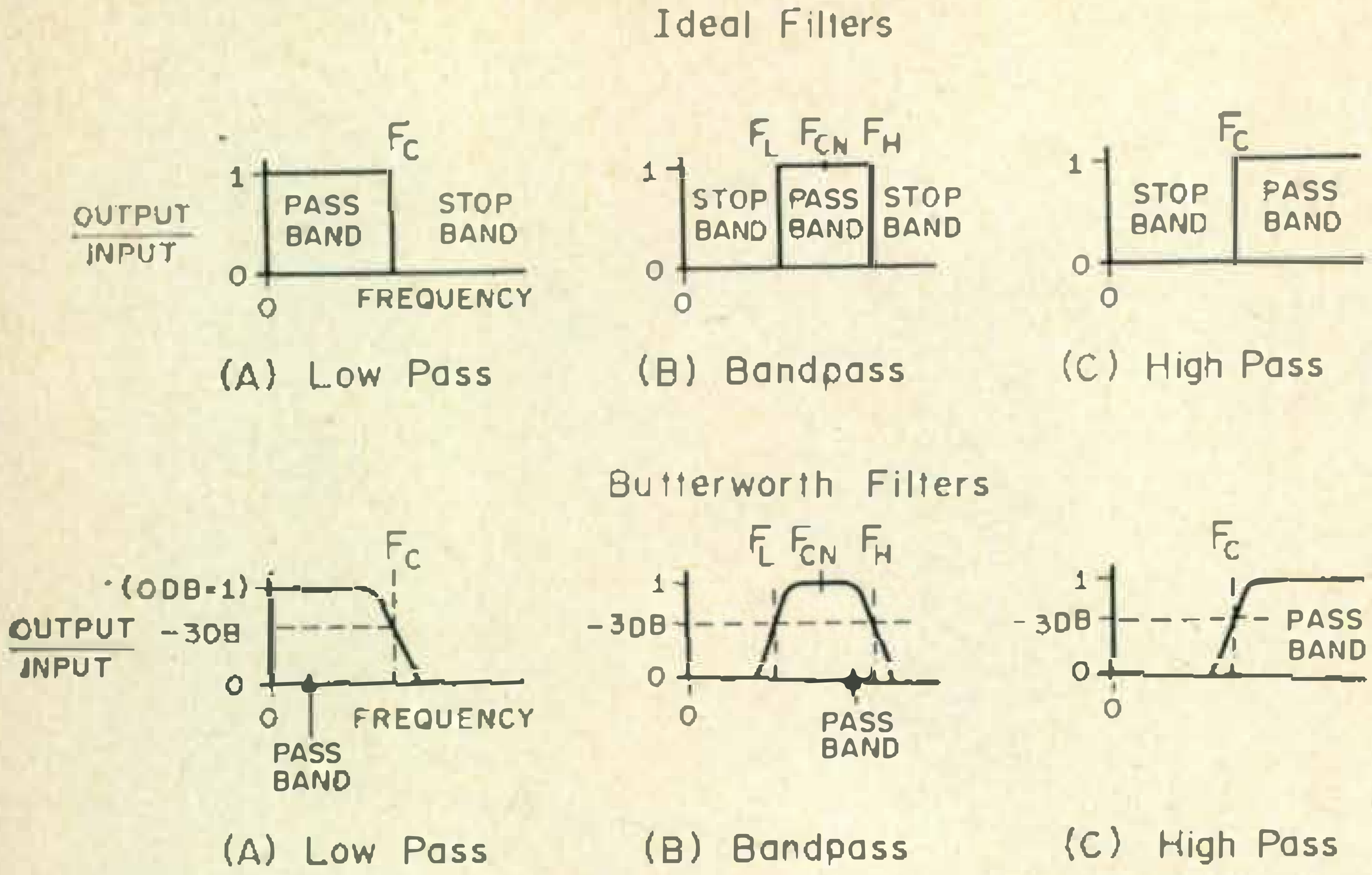

Fig. 1—Comparison between the responses of the ideal filter and Butterworth filters. The cutoff frequency of the Butterworth filter is at the —3 db point which is 0.707 of the maximum voltage or 0.5 of the maximum power (the half power point).

F_l = Low f. F_c = Cutoff f. F_{cn} = Center f. F_h = High f.

high reactance between input and output, highly attenuating any input signals. At low frequencies, the opposite is true, allowing an input signal to reach the output with no attenuation.

We might rightly suppose that as the number of inductors and capacitors (each is called an *element*) is increased, the skirt would become steeper and the passband response would become flatter. This is plotted in fig. 2 which shows the improving filter response as the number of elements are increased. At high frequencies, the attenuation improves at the rat ef —6n decibels/octave. As an example, a five element filter (three inductors; two capacitors) gives a *rolloff* of —30 decibels per octave well above cutoff. Going up one octave *doubles* frequency. A —30 db/octave slope means that only 1/1000 the power at some frequency will be passed for a signal of double that frequency. It can be seen that this is a powerful filter technique.

Note that the cutoff frequency is *always* 3 db down and at the same frequency, regardless of the number of elements.

The High Pass Filter

A *high pass* filter is nothing but an "insided out" low pass filter. The ideal filter passes all frequencies above a cutoff frequency and infinitely rejects all those below. (See fig. 1 (B).) The Butterworth filters are, again, flat over most of the pass band and down 3 decibels at cutoff. This time, the filter consists of *series* capacitors and *shunt* inductors. At high frequencies, the signal sees a low impedance path between source and load. At low frequencies the input signal sees a very low impedance to ground and a very high one to the load, providing the required attenuation improves at the rate of —6n decibels/wards and are shown in fig. 3. Outside of the fact that they are opposites, the performance of low pass and high pass filters are quite similar, requiring the same number of elements to do a similar job.

The Band Pass Filter

A *band pass* filter only passes a certain range of frequencies, called the *bandwidth*. Again, the ideal filter is "square", while the Butterworth is down 3 decibels at the cutoff frequencies. (See fig. 1 (C).) The bandwidth is the difference between the upper and lower cutoff frequencies or

$$\Delta F = F_h - F_l$$

where F_h is the upper cutoff frequency and F_l is the lower. The *center frequency* of the passband is the *geometrical* mean of the cutoff frequencies, or

$$F_{cn} = \sqrt{F_h F_l}$$

where F_{cn} is the center frequency. Note that this is *not* the "average" of F_h and F_l.

Figure 4 shows how bandpass circuits take the form of resonant *LC pairs* that form a ladder network. The *shunt* pairs parallel resonate; the *series* pairs series resonate. In the passband, a low impedance path is formed from input to output providing signal transmission since the series resonant pairs are *low* impedances and the shunt pairs are *high* impedances. Well above the passband, the filter rejects just as a low pass one does. Well below the passband, the filter behaves just as a high pass filter.

Each *LC* pair is called a *section*. The skirt steepness is not defined as so many decibels/octave of center frequency, but *as so many decibels/octave of the bandwidth*. The skirt steepness for a given number of sections will be identical as the skirt steepness of a low pass filter equal to the bandwidth having as many elements as the bandpass filter has sections.

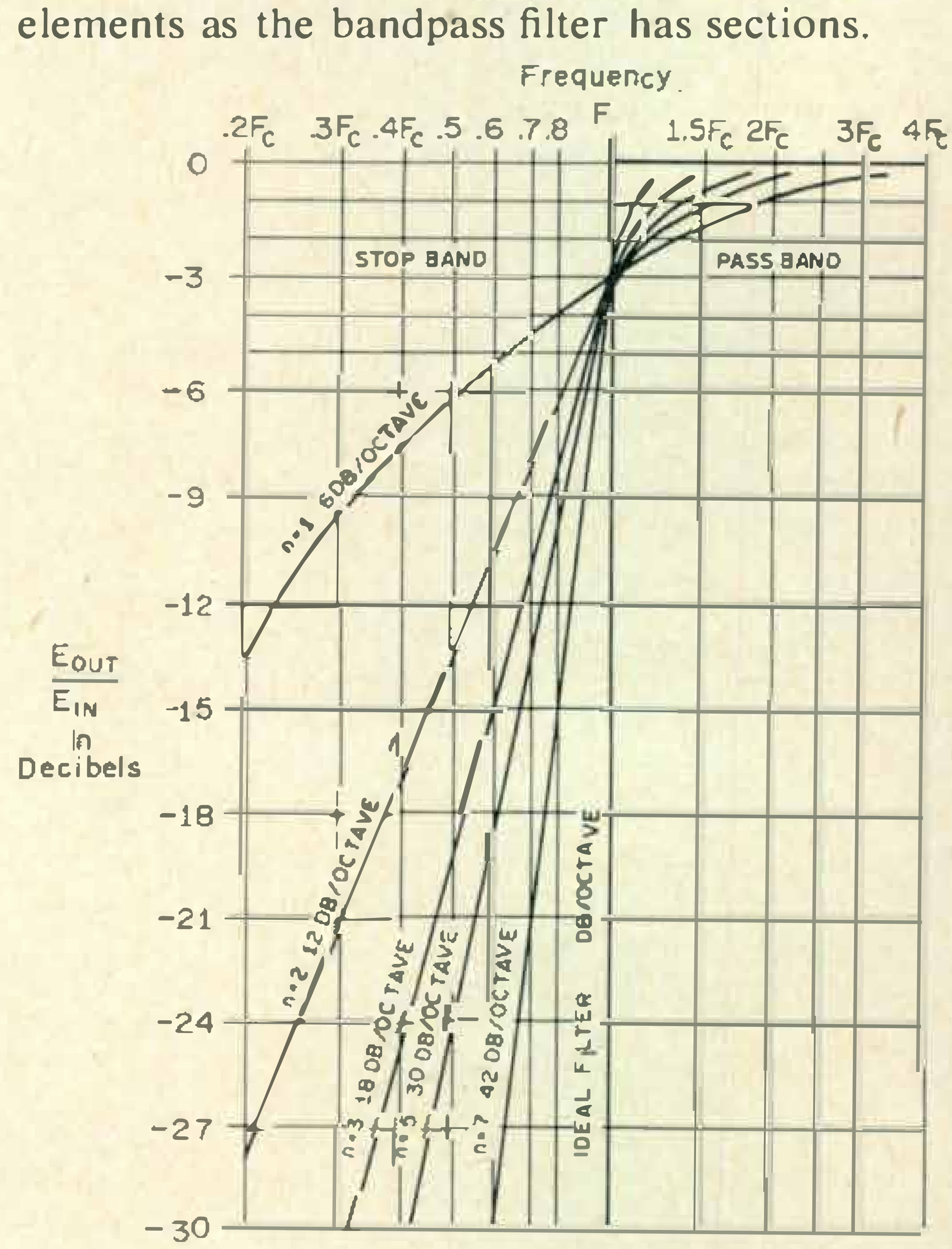

Fig. 2—Frequency response of Butterworth low pass filters. F_c is the cutoff frequency and n indicates the number of filter elements.

Thus fig. 2 gives the skirt steepness of a bandpass filter in terms of so many decibels/octave of the bandwidth. The *n* required will equal the number of *sections* (half the number of elements) to allow the bandpass filter to meet a certain skirt requirement. The skirt on the low side will be the geometric mirror image of the high side skirt. Two frequencies X and Y will have identical attenuation if $XY = F^2_{co}$.

Normalization

Normalization is one of those strange filter terms but its meaning is quite simple. The only difference between a 1 kilocycle filter and a 10 kilocycle filter is that all the inductors and capacitors are ten times as large on the 1 kc model. Similarly, the only difference between a 50 ohm filter and a 100 ohm one is a factor of two on all the parts values.

As we raise frequency, we lower the values of L and C required, but the resistance values are not affected. To raise the impedance level of a circuit, we must increase all resistance and reactance. To increase reactance, we must increase all inductors and decrease all capacitors, since a capacitors reactance decreases with increasing capacity.

Since only constants are invloved in moving around in frequency or impedance level, we only need consider circuits working into an *one ohm load* with a *one cycle per second* cut off frequency. This is called a basic, or *normalized* filter. To scale frequency, we divide all L and C

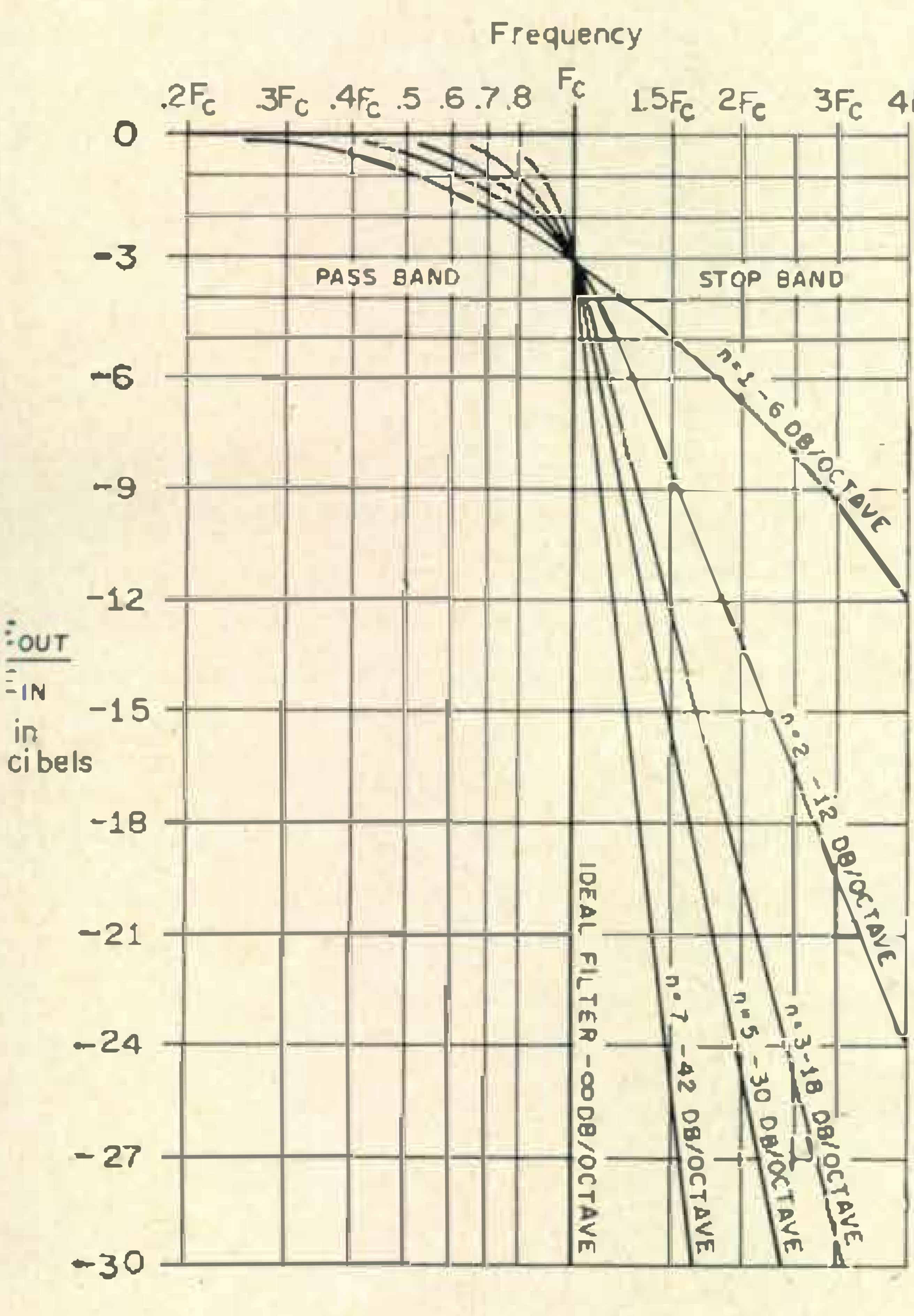

Fig. 3—Frequency response of Butterworth high pass filters. F_c is the cutoff frequency and n indicates the number of filter elements. Note that these curves are simply a reversal of the low pass curves shown in fig. 2.

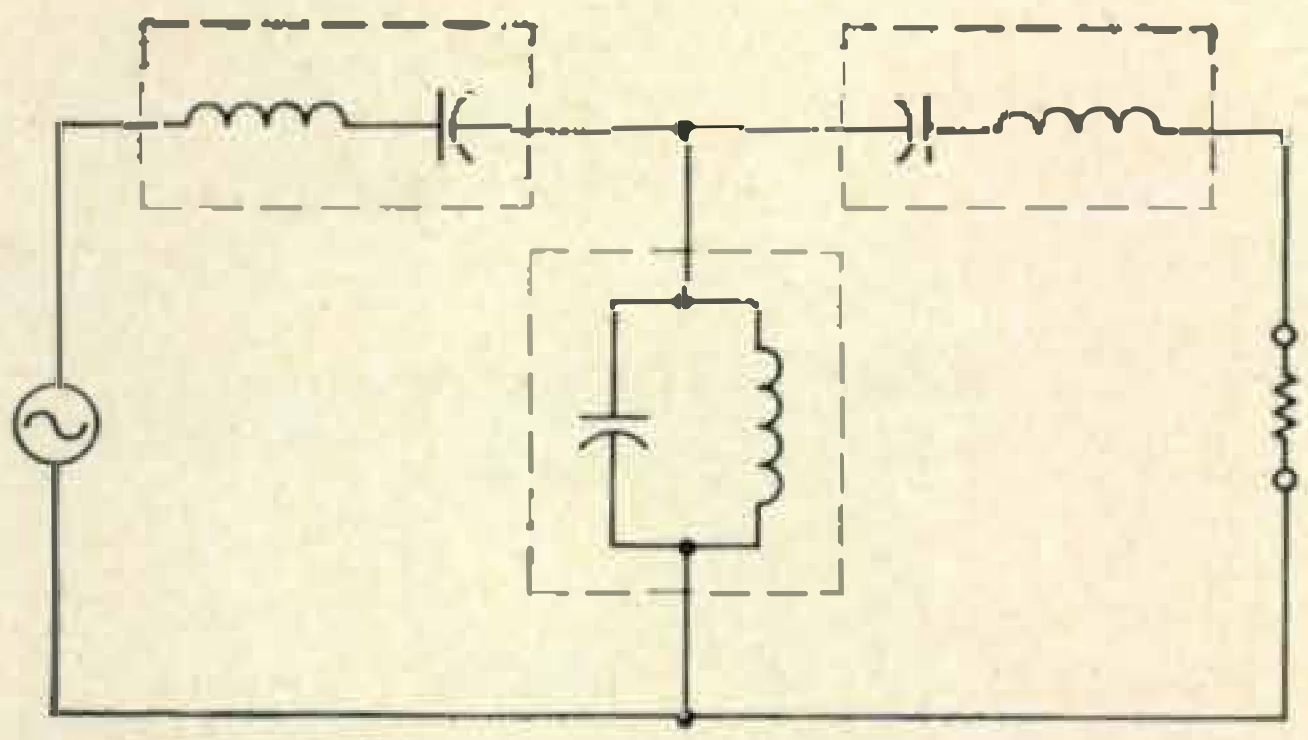

Fig. 4—Basic band pass circuit makes use of series and parallel resonant pairs. Each LC pair is called a section.

by the new frequency. To scale impedance, we multiply all R and L and divide all C by the new impedance level.

One distressing sidelight of this normalization is that the component values will initially show up in henrys and *farads*. These somewhat ridiculous capacitance values become entirely reasonable as impedance and frequency are scaled.

Impedance Matching

We are here presenting only two types of basic filters. One type operates only between a "stiff" voltage source and a load. Typical of such sources are the a.c. power line, emitter and cathode followers, and the outputs of audio amplifiers with high feedback and damping.

The second filter type operates between an impedance matched source and load. Some typical situations are a 50 ohm r.f. signal generator and its 50 ohm load; a 72 or 300 ohm antenna and its transmission line, or that same line and the input to a matched r.f. stage; a 600 ohm audio line and its 600 ohm driver transformer. These two types of filter will cover the majority of practical applications.

One fact cannot be overemphasized: *A filter, any filter, will only operate properly when driven from its proper source and into its design load.* Changing either source or load will drastically affect performance, invariably for the worse. The filters given here only perform as predicted when source and load are those of the original design. It is usually best to terminate a filter with a load resistor directly at the load terminals. In high frequency circuits, cables and interconnections can add serious capacity to the apparent load if this is not done. Also, there is a tendency to forget that a specific load is required if it is remote from the filter; an inadvertant change in cable length or an attempt to change system gain can then completely ruin system response.

In cases where non-linear or variable loads must be driven, it is usually best to terminate the filter in a known, fixed load and insert an emitter follower or similar isolation between the filter and the ultimate load.

Part II, will present the actual design procedures and several examples for each type of filter.

[*To be Continued*]

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Part II

The design concepts of the Butterworth filters were presented in Part I. Presented below are the design techniques and several typical examples.

To design a Butterworth filter, just follow the step by step instructions given in the tables. Table I covers low pass designs, Table II the bandpass types, and Table III covers the high pass designs. The procedure for the low and high pass designs is quite simple. First the problem is specified in terms of load, source, cutoff frequency, and skirt steepness. From fig. 2 or fig. 3, the number of elements is determined. This defines a basic filter circuit which is selected from one of the tables, observing the proper source. This basic filter is then scaled in frequency and impedance by a simple multiplication and division. That is all there is to it.

The bandpass design is a bit more involved, but still quite straightforward. Again the problem is defined in terms of load, source, upper and lower cutoff frequencies, and skirt steepness. The bandwidth and center frequency are then calculated using the formulas given.

The next step may seem strange. After deciding which basic filter is to be used, we in effect form a low pass filter equal to the *bandwidth* and then resonate each element with a new L or C about the center frequency. This is the result of a mathematical technique called the transformation of a complex variable. In this case, we have designed a low pass filter and then *transformed* it into a bandpass one. Scaling impedances then completes the design.

To obtain these tables, we started with the Butterworth polynomial coefficients and realized ladder networks designed for a given number of elements and a specific type of source. We then normalized the results to one cycle per second filters with a one ohm load that perform as per figs. 2 and 3. The details of all this are in any good synthesis book. Fortunately, this need only be done once; the results in the tables are all that is needed for any practical design.

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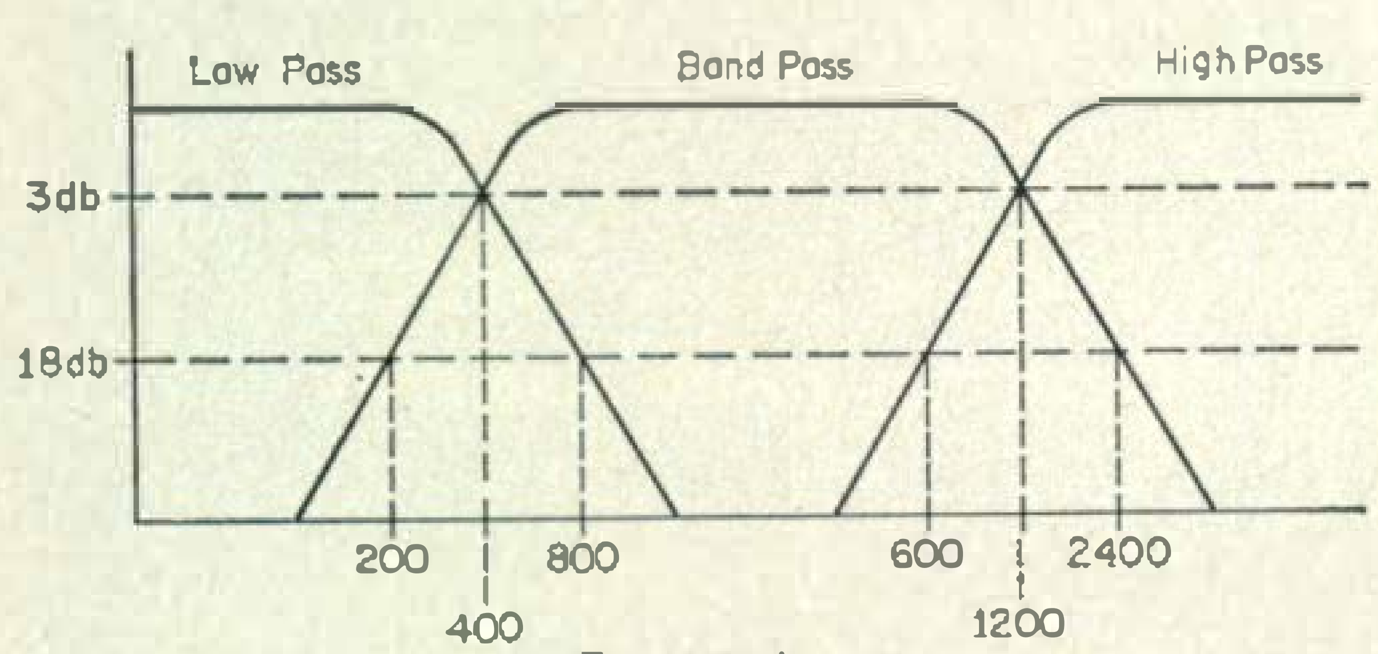


Fig. 5—Required curves for the audio crossover network design discussed in Problem #1.

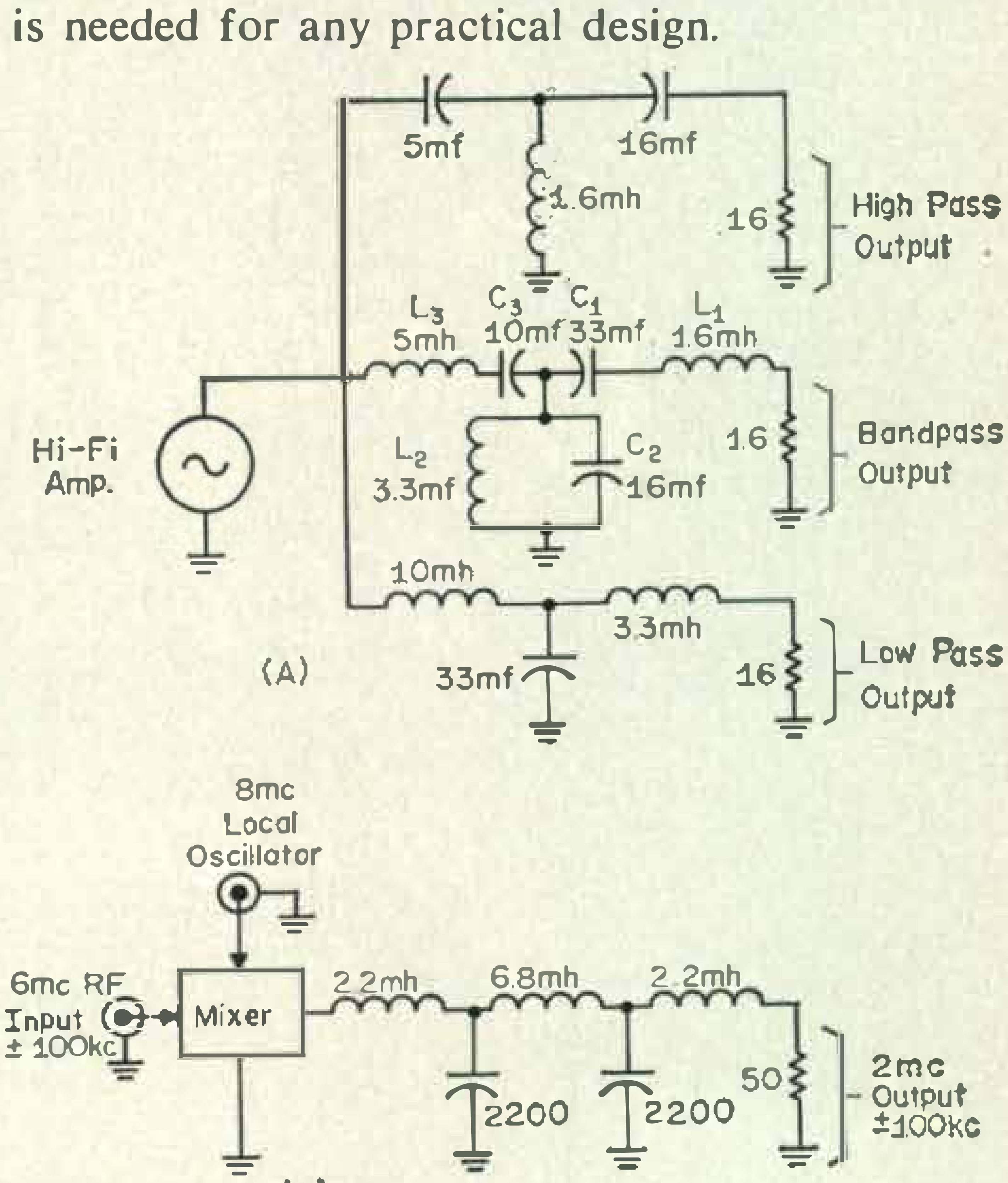


Fig. 6—Circuits and component values arrived at for the two problems illustrated in the text. Circuit (A), the audio crossover network and (B) the mixer output filter component values have been rounded off to available stockroom values.

Examples

Problem #1—An audio crossover network is to consist of a low pass, a bandpass, and a high pass filter with 3 decibel crossover frequencies of 400 and 1200 c.p.s. Each filter is to have at least 18 decibels of loss one octave beyond its cutoff frequency (both cutoff frequencies for the bandpass), and operate into a 16 ohm load when driven from a hi-fi amplifier with high damping. (See fig. 5.) Design the filters.

Low Pass Design—All filters will be the voltage source type. The low pass filter is to be a 16 ohm, voltage source driven type at 400 c.p.s. Looking at fig. 2 and $F = 2\,F_c$ (one octave above cutoff) we see that $n = 3$ will have 18.1 decibels of loss, meeting the attenuation requirement. We then remove basic filter #3 from Table I since it is a voltage source driven, $n = 3$, low pass filter. The inductors are scaled by dividing them by the cutoff frequency and multiplying them by the load resistance, or:

$$L_1 = 0.0795\,h \cdot \frac{16}{400} = 0.00318\,h \approx 3.3\,mh.$$

$$L_2 = 0.238\,h \cdot \frac{16}{400} = 0.00952\,h \approx 10\,mh.$$

Similarly, to scale C_1, we divide by both the frequency and the load impedance:

$$C_1 = \frac{0.212\,farads}{400 \cdot 16} = 0.000033.2\,farads \approx 33\,mf.$$

High Pass Design—This is largely similar to the low pass design except the cutoff frequency is 1200 c.p.s. Entering fig. 3 at $F = 0.5\,F_c$ (one octave below cutoff), we again see then $n = 3$ gives 18.1 decibels of loss. We might have anticipated this from the similarity of the filters and their skirt requirements. Basic filter #23, an $n = 3$ voltage source, high pass type, is removed from Table III. The elements are then scaled in frequency and impedance:

$$C_1 = \frac{0.318\,farads}{1200 \cdot 16} = 0.0000165\,farads \approx 16\,mf.$$

$$C_2 = \frac{0.106\,farads}{1200 \cdot 16} = 0.0000055\,farads \approx 5\,mf.$$

$$L_1 = \frac{0.119\,h \cdot 16}{1200} = 0.00158\,h \approx 1.6\,mh.$$

Band Pass Design—As directed in Step 1 of Table II, we can specify the following:

A—Load resistance: 16 ohms.

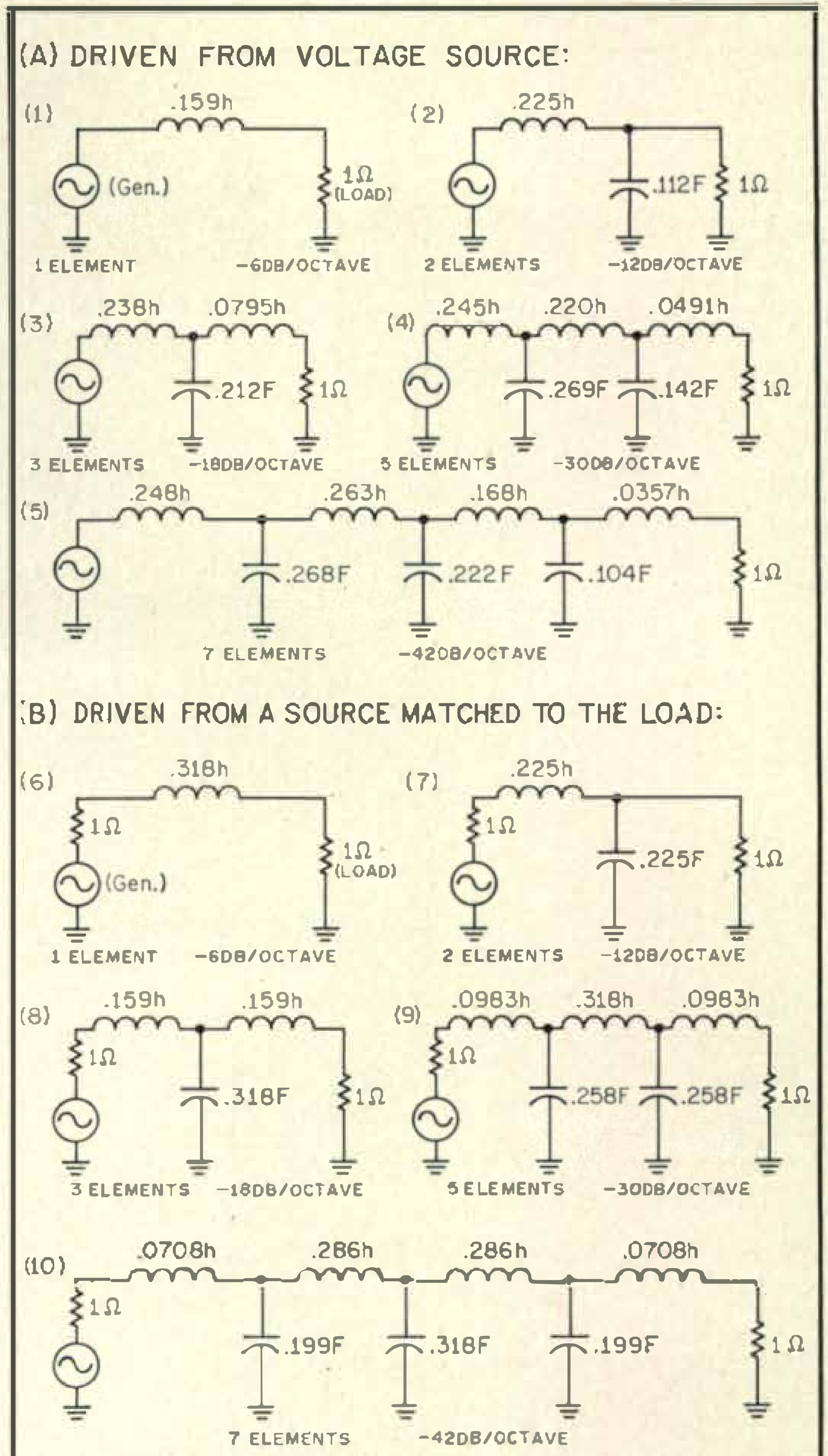
B—Type of source: voltage.

C—Upper and lower cutoff frequencies: 400 and 1200 c.p.s.

Step 2 indicates the calculation of the bandwidth and center frequency.

$$\Delta F = F_h - F_l = 1200 - 400 = 800 \text{ c.p.s.}$$

$$F_{cn} = \sqrt{F_h \cdot F_l} = \sqrt{1200 \cdot 400} = 695 \text{ c.p.s.}$$



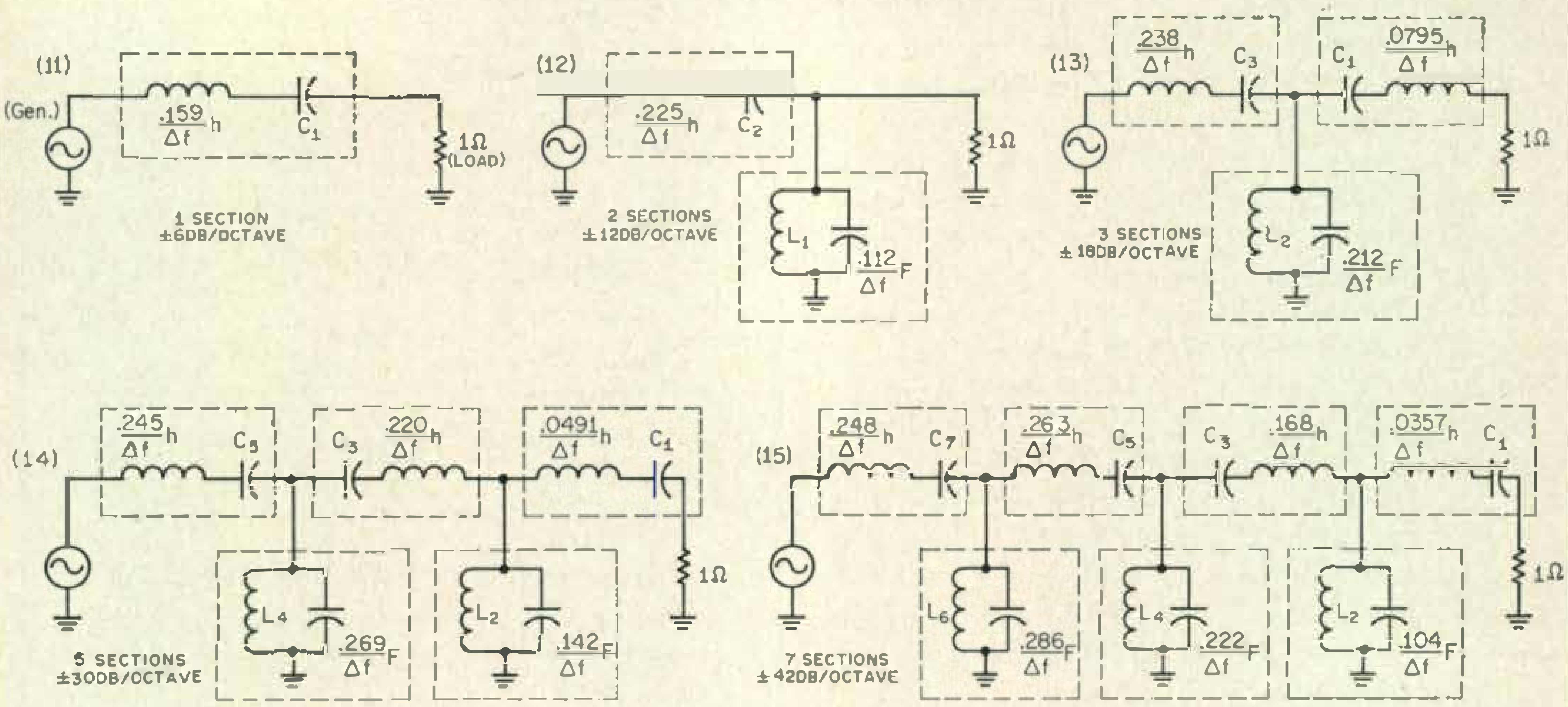
1. Specify the following:
   A. The load resistance.
   B. The type of source (voltage or matched).
   C. The cutoff frequency.
2. Establish some criterion for the steepness of the skirts required in the following form:
   The signal must be "X" decibels attenuated at a frequency of "Y" times the cutoff frequency.
3. From fig. 3, find the required number of elements. If the result lies between the curves, choose the higher number of elements.
4. From Table I, remove the basic filter having the right number of elements and the proper type of source.
5. *Multiply* each R and L by the load resistance. *Divide* each C by the load resistance.
6. *Divide* each new L and C by the cutoff frequency.

The design is now complete.

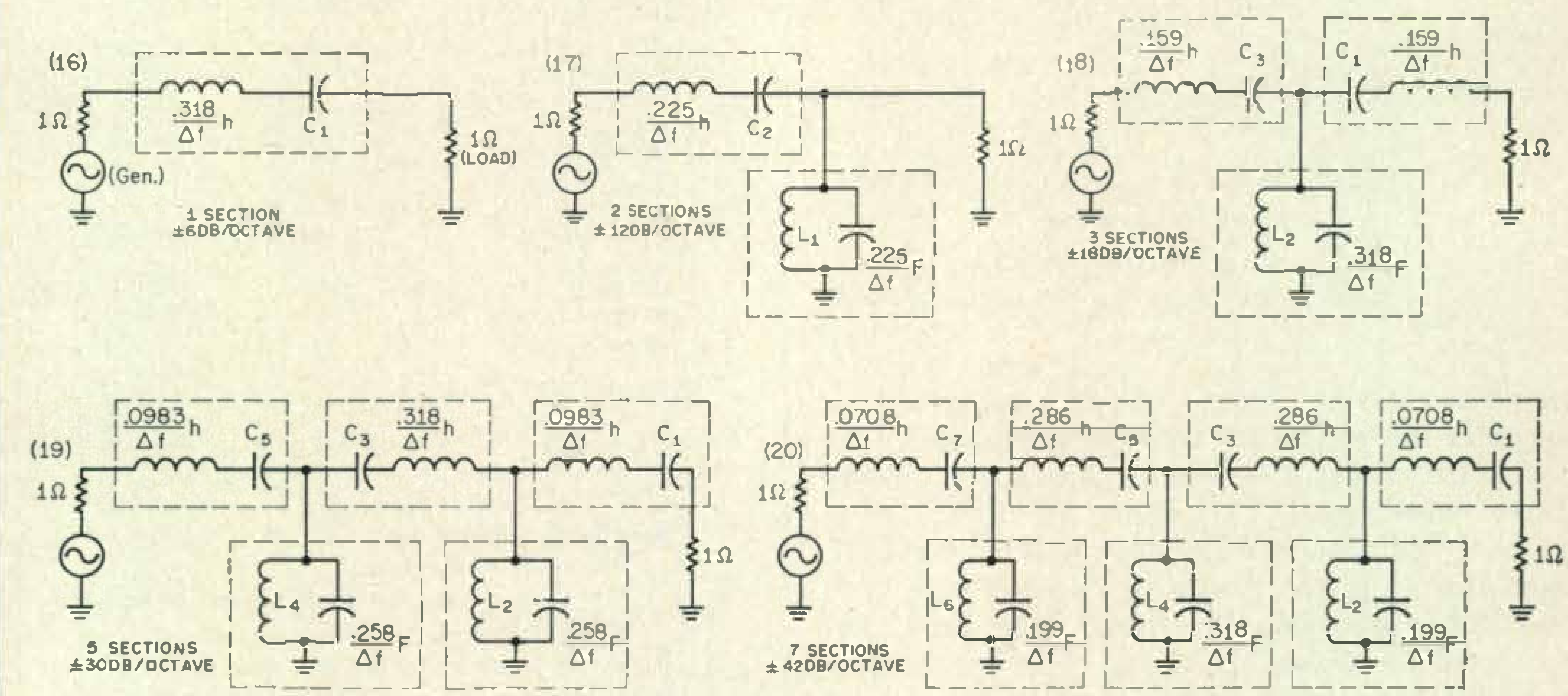
Table I—Basic circuits and the procedure for designing low pass filters.

Step 3 requires the determination of the steepness of the skirts. To determine the skirt steepness of the bandpass filter the low pass curves of fig. 2 are used. However, these curves must be reexpressed as so many *bandwiths* above cutoff. Since cutoff occurs at F_c, one bandwidth *above*

(A) BASIC BANDPASS VOLTAGE SOURCE FILTERS:



(B) BASIC BANDPASS MATCHED SOURCE FILTERS:



1. Specify the following:
   A. The load resistance.
   B. The type of source (voltage or matched).
   C. The upper and lower cutoff frequencies.

2. Calculate the bandwidth and center frequency by

$$\Delta F = F_h - F_l$$

$$F_{cn} = \sqrt{F_h \cdot F_l}$$

3. Establish some criterion for the steepness of the skirts required in the following form:
   The signal must be "X" decibels attenuated at a frequency of "Y" *bandwiths above* the upper cutoff frequency.

4. From fig. 2 find the number of sections required for the filter. (Note that 2 bandwidths above the cutoff frequency occurs at $F = 3F_c$, etc. If the lower skirt is of interest, use geometrical symmetry to see what the upper skirt must do to satisfy the lower. Any two frequencies X and Y will have the same attenuation if $XY = F^2_{cn}$.)

5. From Table II, remove the basic filter with the proper source and the right number of sections.

6. Each box in the basic filter will have one inductor or capacitor value that is to be divided by the bandwidth. Perform this division.

7. Find the inductor or capacitor required to resonate each box at F_{cn}. Use a resonance nomogram or the formula:

$$LC = \frac{0.0253}{F_{cn}^2}$$

8. *Multiply* each R and L by the load resistance. *Divide* each C by the load resistance.

The design is now complete.

Table II—Basic circuits and the procedure for designing a band pass filter.

cutoff would be at $2F_c$; two bandwidths above cutoff would be at $3F_c$, etc.

For the problem at hand the bandpass cutoff frequency is 1200 c.p.s. and the —18 db point must be one octave up or 2400 c.p.s. Thus, the number of bandwidths is equal to:

$$\frac{2F_h - F_h}{\Delta F} = \frac{2400 - 1200}{800} = 1.5$$

As previously explained, 1.5 bandwidths above F_c occurs at $F = 2.5\ F_{cn}$ in fig. 2.

At this point, $n = 2$ will have 15.8 decibels loss, 2.2 shy of the required 18, while $n = 3$ will have a more than adequate 24 decibels. We might be tempted to "cheat" on the skirt requirements to save two parts, but if the problem is to be satisfied as stated, $n = 3$ must be used. We chose basic filter #13 in Table II. Dividing each L or C by the bandwith gives:

$$L_1 = \frac{0.0795\ h}{800} = 0.0000991\ h = 0.0991\ mh.$$

$$L_3 = \frac{0.238\ h}{800} = 0.000297\ h = 0.297\ mh.$$

$$C_2 = \frac{0.212}{800} = 0.000266\ f = 266\ mf.$$

Each of these sections must resonate at F_{cn} which, for this problem, is 695 c.p.s. As directed in step 7 of Table II, the value of the component required to bring the section to resonance is determined by:

$$LC = \frac{0.0253}{F_{cn}^2} = \frac{0.0253}{483{,}000}$$

$$LC = 0.0524 \times 10^{-6}$$

We substitute in this equation to get C_1, L_2 and C_3 in terms of L_1, C_2 and L_3, respectively. The results are:

$$C_1 = \frac{0.0524 \times 10^{-6}}{L_1}$$

$$= \frac{0.0524 \times 10^{-6}}{0.0991 \times 10^{-3}}$$

$$= 529\ mf.$$

$$L_2 = \frac{0.0524 \times 10^{-6}}{C_2}$$

$$= \frac{0.0524 \times 10^{-6}}{266 \times 10^{-6}}$$

$$= 0.197\ mh.$$

$$C_3 = \frac{0.0524 \times 10^{-6}}{L_3}$$

$$= \frac{0.0524 \times 10^{-6}}{0.297 \times 10^{-3}}$$

$$= 176\ mf$$

All six elements are then impedance scaled by multiplying the inductors by 16 and dividing the capacitors by 16. The results, rounded off to stock room values, are shown in fig. 6.

Problem #2—A 50 ohm r.f. mixer is to produce a 2 mc output signal with a local oscillator frequency of 8 mc and an input signal of 6 mc having a 200 kc bandwidth. Design a filter that attenuates input, local oscillator, and the sum frequency by a minimum of 30 decibels, but passes the difference signal with less than one decibel of loss.

Solution—A single matched source low pass filter will do the job. Since the Butterworth low pass attenuation increases with increasing frequency, only the lowest frequency to be blocked need be considered, for all others will have higher attenuation. The lowest blocked frequency has to be 5.9 mc (6 mc—half the 200 kc bandwidth). The highest pass frequency has to be 2.1 mc (2 mc + half the 200 kc bandwidth). We cannot use a cutoff frequency of 2.1 mc, since the response will be down *three* decibels at this point, and only one decibel is allowed. Let us try a 2.4 mc cutoff frequency. The pass signal *as a fraction of the cutoff frequency* will be $2.1/2.4 = 0.88\ F_c$. The lowest frequency signal to be blocked *as a fraction of the cutoff frequency* will be $5.9/2.4 = 2.46\ F_c$. Consulting fig. 2, we see that $n = 5$ will have slightly over 30 decibels of attenuation at 2.46 F_c and only 0.9 decibels of loss at 0.88 F_c, meeting both pass and stop requirements. We enter Table I for a matched source $n = 5$ filter, or basic filter #9. The L and C values are scaled to F_c # 2.4 mc, and the impedance is scaled to 50 ohms, just like the previous low pass design. The figures are shown below.

$$L_1 = L_3 = \frac{0.0983 \cdot 50}{2.4 \times 10^6}$$

$$= 2.05\ \mu h.$$

$$L_2 = \frac{0.318 \cdot 50}{2.4 \times 10^6}$$

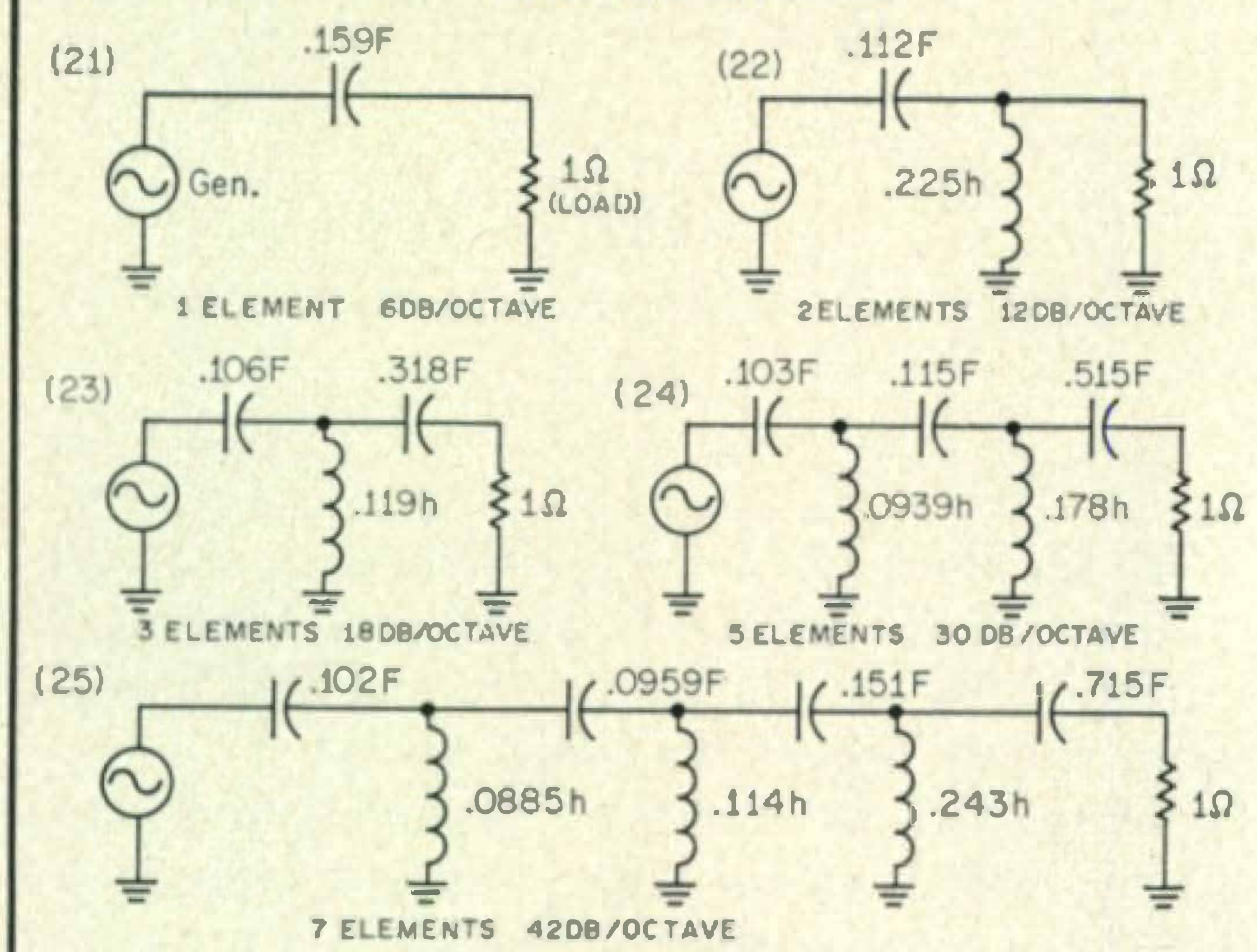
$$= 6.62\ \mu h.$$

$$C_1 = C_2 = \frac{0.258}{50 \cdot 2.4 \times 10^6}$$

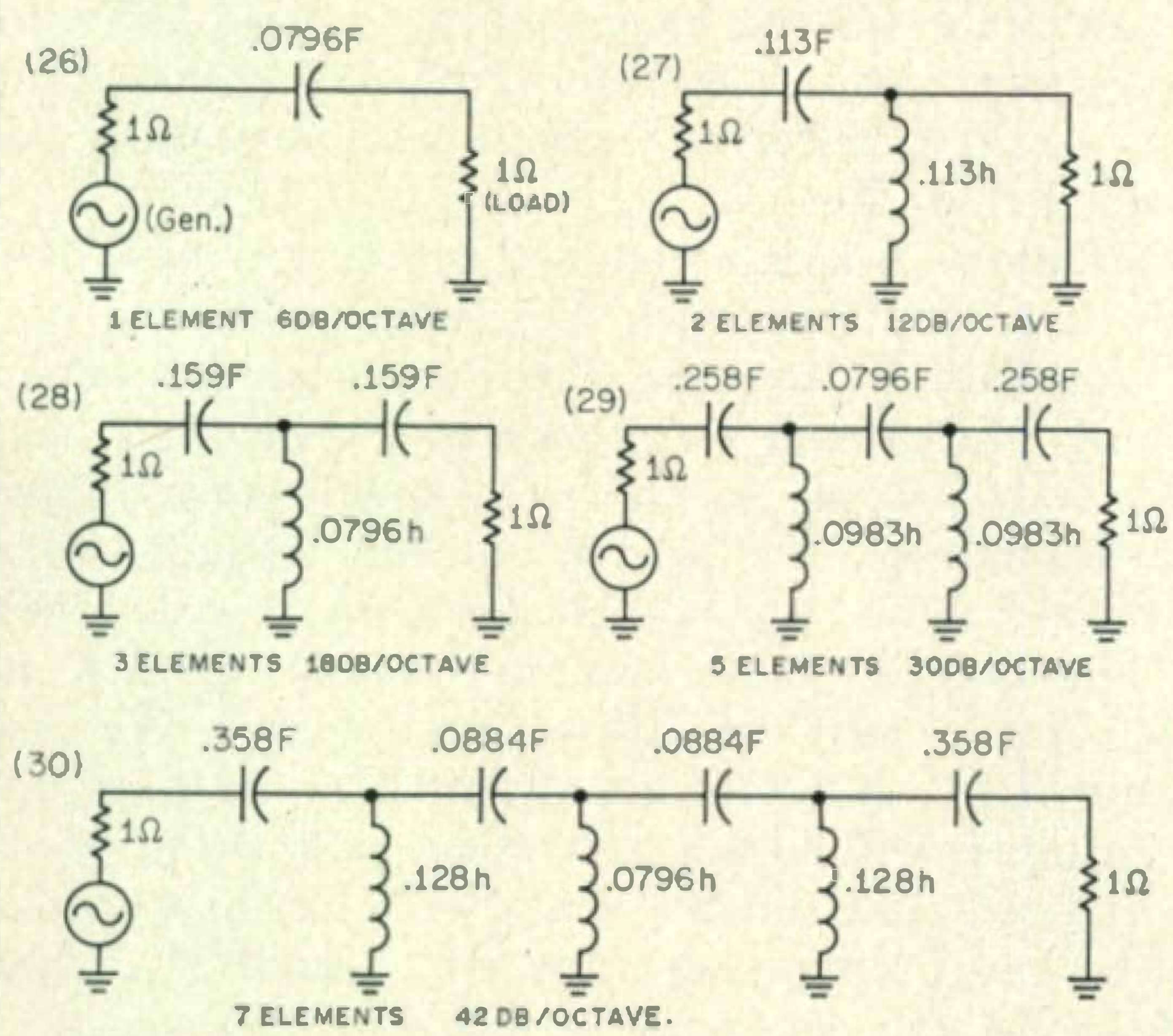
$$= 2150\ mmf.$$

The rounded off results are shown in fig. 6.

(A) DRIVEN FROM VOLTAGE SOURCE:



(B) DRIVEN FROM A SOURCE MATCHED TO THE LOAD:



1. Specify the following:
   A. The load resistance.
   B. The type of source (voltage or matched).
   C. The cutoff frequency.
2. Establish some criterion for the steepness of the skirts required in the following form:
   The signal must be "X" decibels attenuated at a frequency of "Y" times the cutoff frequency.
3. From fig. 2, find the required number of elements. If the result lies between the curves, choose the higher number of elements.
4. From Table III, remove the basic filter having the right number of elements and the proper type of source.
5. *Multiply* each R and L by the load resistance. *Divide* each C by the load resistance.
6. *Divide* each new L and C by the cutoff frequency.

The design is now complete.

Table III—Basic circuits and the procedure for designing a high pass filter.

Performance

And how good do they work? The results are almost always in nearly perfect agreement with the design curves provided some rather obvious and invariably overlooked requirements are met. The circuit on the page and the one on the bench must be as nearly identical as possible. Inductors must have high Q (at the very least 15; preferably 75 or more). More important, *the Q must be specified and measured at the cutoff frequency.* The millihenry sized r.f. chokes in the distributor's catalog may have a very good Q at a few megacycles, but their d.c. resistance makes them ridiculously unsuitable for use as inductors in problem #1. (The Q will typically be less than 0.05 or so and the concept of Q is meaningless for $Q < 1$).

There must be no magnetic coupling between different inductors in the filter. This requires magnetic shields or else self-shielding inductors such as toroids and cup cores. Lead inductance must be allowed for on very small inductors. An eighth inch of wire has around ten nanohenries of inductance, enough to foul up a 0.1 microhenry coil if not allowed for. It is best to arrange the circuit so that the longest leads are in series with the largest inductors. Obviously, no inductor can be self-resonant anywhere near the frequencies of interest.

The same goes for capacitors. Some paper foil capacitors have considerable inductance and resonate at relatively low frequencies. Worse yet, some disc capacitors are radically temperature dependent. Worst of all are electolytics with their high dissipation and intrinsic polarity. The best choice in a small capacitor is the new monolithic glass ones, with ordinary silver mica capacitors being a close substitute.

The new polystyrene capacitors, (Mallory SX, etc.) are almost as good as silver micas for all filter applications, and at 21¢ each, are much cheaper. For large values of C, ordinary mylar capacitors will work fairly well, provided they are the type that have solid metallic ends on the foil wrap. For giant C values, back-to-back tantalums are the only economical answer.

In bandpass filters, each LC pair must be precisely resonant at the center frequency. Very narrow bandpass filters require extremely high Q inductors.

Finally, the circuit layout must be the same as the filter design. Short cut ground paths between input and output can drastically alter response. Ground and return paths should be heavy gauge or wide foil. The load should be placed directly at the output termination of the filter.

Any readers who want to look into the "why" of these filters might find the following synthesis texts of interest.

Accuracy

The author is quite confident of all element values given in Tables I, II, and III as they were verified twice by others using independent checking methods. The element values were obtained in normalized form from Table 2.1 a and 2.1 b of reference 3, above. These element values are renormalized to the 1 c.p.s. cutoff frequency instead of the 1 r.p.s. cutoff frequency given in Reference 3. For the high pass networks, the

[*continued on page 92*]

Announcements [*from page 8*]

[from page 8]

Stolen Equipment

On Aug. 24, 1966 a Lafayette HA-650 6 meter walkie-talkie was stolen from the car of Reid S. Edles, WA2TBT, in Queens, N.Y. Anyone having information about this please contact Reid S. Edles, 31-21 54th Street, Woodside, N.Y., 11377.

A prototype linear amplifier bearing the trade mark BTI, model LK-2000 was stolen from a parked trailer at the Western Single Sideband Convention at the Newporter Inn, Newport, California. It has a serial number of 1743 on it. A reward of $200.00 has been offered for information leading to its return. Contact Brad Thompson, P.O. Box CCCC, Indio, California, 92201.

Correction

In the article Medical Aspects of Radiation which appeared in the October 1966 issue of *CQ* on page 62, one megacycle should be considered 10,000 centimeters not 10,000 meters as stated.

The FD-30 [*from page 20*]

should be at least 6 henries. The female plug on the end of the power cable automatically switches the filament and relay wiring for 6 volt a.c. operation.

Figure 2 (B) shows the plug wiring used for a mobile power supply. The 300 volts must also be developed from the 12 v.d.c. source. The most efficient method is a transistorized d.c. to d.c. converter. This type of supply has been used with the FD-30 for several years with no problems. A Triad #TY79 transformer (rated for 200 ma at 300 volts) is suitable and Triad includes the circuit of a d.c. to d.c. converter. If you build this unit make sure that the power transistors are heat sinked properly.

Conclusion

At the time of this writing there are 3 of these transmitters in the Chicago area and they have been used for mobile, Field day and fixed station use with no problems.

The ease of duplication makes it a project that can be built by the inexperienced amateur, yet its versatility is such that it will not be obsolete. The writer wishes to thank W9DZM without whose help this article couldn't have been written. ■

A companion receiver for the FD-30 is in the works by the author and should be ready for publication very shortly. ed.

Butterworth Filters [*from page 26*]

elemental values are inverted and then scaled to 1 c.p.s. All necessary transformations and characteristics are given in any and all of the references. ■

Bibliography

[1] Balbanian, Norman, *Network Synthesis*, Prentice Hall, Englewood Cliffs, N.J., 1958, chapter 9.

[2] Weinberg, Louis, *Network Analysis and Synthesis*, McGraw Hill, New York, N.Y., 1962. Chapters 11 and 13.

[3] Weinberg, Louis, *Network Design by use of Modern Synthesis Techniques*, Hughes Research Laboratories Technical Memorandum #427, Culver City, California, April 1956, Chapters 1, 2, and 5.

DX [*from page 74*]

ZL4 Chatham Island: ZL4CH has been very active on Friday and Saturday evenings (US time). Has no set operating frequency and can be found all over the phone band. *(Tnx NEDXA).*
5R8 Malag Sy Republic: Chet, 5R8AS, is a new station active from here. He was on 15 meter SSB with a big signal and asked QSLs go via W6ZPX. *(Tnx LIDXA).*

EA8 Contest: This contest starts 0000 GMT on December 21, 1966 and ends on March 20, 1966. A certificate will be awarded to those who submit confirmation of two-way QSOs, any band, any mode that include the exchange of previous day min/max local temperature. The quantity of confirmations depends upon location, as follows: Spain, Portugal, North Africa—250 QSO's; Europe—10 QSO's; Rest of World—5 QSO's. Submit to: Tenerife Eterna Primavera Award, POB 215, Tenerife, Canary Islands. It is planned that at the end of the contest period, the sponsoring Canary Island civic groups will hold a drawing to select the winners from those who have earned certificates. First prize will be a trip to the Canary Islands.

QTHs and QSL Managers

For those of you who have not seen it, the W6GVS *QSL Managers and QTH Directory* is a welcome addition to any operating position. It is most complete and is brought up to date every three months. The cost is $3.00 per year in the U. S. and Canada, and $4.00 elsewhere. Send directly to W6GVS, Box 54222, Terminal Annex, Los Angeles, Calif. 90054.

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