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SIMPLIFIED DESIGN TECHNIQUES FOR DISTRIBUTED POWER AMPLIFIERS FOR SYNCHROTRON AND CYCLOTRON DRIVER APPLICATIONS*

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Summary

Design of distributed power amplifiers has been simplified through the use of a set of design charts and simplifying circuit techniques. A circuit in which the anode and grid lines are built from bridge-T constant-resistance networks seems to be best. These lines have an image impedance that is resistive and of constant value for all frequencies, and therefore do not require terminating half sections. For a given characteristic impedance and shunt capacitance, the bridge-T lines have about twice the bandwidth of the constant-K lines. The design of the ferrite transformers for matching the anode and grid lines to standard line impedances has been simplified by means of design charts and tables. These amplifiers are simple to build and very stable. A 3.5-kW amplifier which operates from 1.0 to 70 MHz is used to illustrate the technique.

Introduction

Distributed power amplifiers are convenient in variable frequency RF systems because they eliminate the need for the complex tracking equipment required for tuned amplifiers. We have also found them to be very useful for spin-echo research because they can produce bursts of RF with extremely short rise and fall times. For these advantages a price is paid in amplifier efficiency. For a given class of amplifier the efficiency is about half that for a tuned amplifier. However, in the accelerator applications, the power is small compared with the other components, so that the low efficiency can be tolerated. In the spin-resonance application the low duty factor makes the average power requirements modest.

Distributed amplifiers were first described by Percival in 1936,¹ but there was very little interest until 1948 when Ginzten et al. published the first complete general theory.² The first application of distributed amplifiers to synchrotrons was described by Bard in 1962,³ and to cyclotrons by W. P. Johnson in 1963.⁴

We have studied both single-ended and push-pull amplifiers. While the push-pull circuit provides even-ordered harmonic cancellation

which makes distortionless class-B amplification possible, the usual high efficiency obtained in this class of service is lost because the anode line of the nonconducting half of the amplifier provides a prohibitive load on the conducting half of the amplifier. In class-A amplifiers, the single-ended circuit usually provides a better impedance match between tubes and anode line than the push-pull circuit.

Of the various lump-constant lines for the anode and grid circuits studied by Ginzten et al., the bridge-T network provides the highest gain-bandwidth product. It can be designed to have a constant resistance characteristic, i. e., the image impedance is purely resistive and of constant value for all frequencies. This eliminates the need of terminating half sections and permits the line to be terminated with a resistor. For a given gain the bridge-T line provides about twice the bandwidth of the constant-K line. For frequencies less than about half the cutoff frequency, the time delay through the amplifier is almost constant. This is convenient in synchrotron applications where many resonators must operate in phase and where the phase of the RF must be related to the phase of the beam bunches.

Constant Resistance Bridge-T Networks

The bridge-T network can be used to transform the input capacitance of a vacuum tube to a pure resistance as shown in Fig. 1a, and as the basic building block of a lump-constant transmission line in distributed amplifiers as shown in Fig. 1b. The constant-resistance bridge-T network is derived in Ref. 2, the results of which are shown in Fig. 2a. The midshunt inductance can be obtained from mutual coupling between the two halves of the coil as shown in Fig. 2b. Bard has shown that if m is chosen to be 1.27, the maximum time-delay error for frequencies less than midfrequency is 1%. In this case, the inductance to the midpoint of the coil must be 40.3% of the total inductance of the coil. The low-frequency time delay per section is:

$$\tau = \frac{1}{2} (LC)^{1/2} \text{ sec.}$$

By using the equation

* Work done under the auspices of the U. S. Atomic Energy Commission.

$$L = \frac{r^2 n^2}{9r + 10l} \mu\text{H},$$

with n the number of turns and r and l the radius and length of the coil, it can be shown that the correct coefficient of coupling results if the length of the coil is 1.35 times its diameter. Figure 3a is a design chart for the bridge-T constant resistance network of Fig. 2c, and Fig. 3b is a design chart for the inductance of this network.

Transmission-Line Transformers

Transmission-line transformers are best described in terms of coaxial cables. The basic concept is that several equal length cables are connected with their output terminals in parallel and their input terminals in series. Sufficient ferrite encircles the cables to suppress current flow on the outside of the cables, thus isolating the input and output terminals. Figures 4a, b, and c illustrate the technique for voltage ratios of 2-, 3-, and 4 to 1. The amount of magnetic field which appears in each core corresponds to the voltage developed across a single cable. The ferrite cross section required is determined by the voltage per cable at the lowest frequency of the transformer pass band. In practice, several turns of cable are put around each core in order to use the ferrite as effectively as possible. The spacing between turns must not be too close or the turn-to-turn capacitance produces undesirable resonances.

Usually, it proves to be more convenient to use twisted-pair transmission line rather than coaxial cables, because for a given impedance and given cross sectional dimension of the cable, the copper cross section is larger and will carry more current. The amount of line required is small and can be made by the technician building the amplifier.

The characteristic impedance, Z_0 , of a twisted-pair line can be measured conveniently with a grid dip meter in the following way. Take a piece of line a few feet long. Short circuit one end and measure the quarter-wave resonant frequency, f_1 . The velocity of propagation is:

$$v = \frac{2}{\pi} \omega_1 l.$$

Now foreshorten the line by connecting a capacitor, C , of about 100 pF across the open end. Measure the new quarter-wave resonance, f_2 . Z_0 can be calculated from

$$Z_0 = \frac{1}{\omega_2 C \tan \frac{\omega_2 l}{2}} = \frac{.159}{f_2 C \tan \left(90 \frac{f_2}{f_1} \right)}.$$

Table I shows some useful twisted-pair lines.

Distributed-Amplifier Construction Techniques

The combination of high power and large

bandwidths requires that particular attention and care be given to circuit layout. It is essential that tube lead inductances, such as to the grid, cathode, and screen, be kept low so that the resonances associated with these electrodes will lie well above the operating band of the amplifier. We have found that a number of small capacitors connected in parallel and mounted on a low-inductance etched circuit board serves well as a by-pass or coupling network. Low-value series resistors connecting supply leads to screen by-pass capacitors are an effective method of isolating the modes of the B supply wiring from the amplifier circuitry. This technique can be employed in grid bias and anode circuits also. For the terminating resistors of the anode and grid lines, 2-W carbon resistors mounted in a matrix are satisfactory up to about a 50-W rating. Above this we have found that the commercial coaxial 50- Ω water-cooled terminating resistors work out well. These can be connected to 200- or 400- Ω anode lines by means of ferrite transformers.

Frequently one finds the grid capacitance of the vacuum tubes inconveniently large, requiring an inconveniently low-impedance grid line. Two techniques are available for handling this problem--a series capacitor can be used to reduce the effective capacitance, or several parallel grid lines, each sharing a part of the grid capacitance, can be used. For tubes with multiple grid connections, the latter technique raises the self-resonant frequency of the grid: it is essential that this resonance be at least an octave above the high-frequency limit of the amplifier in order to prevent reflections on the grid line.

A design problem which occurs fairly frequently is that at the desired operating point the vacuum tubes are operating in the region of negative screen current. We have found that this condition can be tolerated if the positive current in the bleeder of the screen supply is made larger than the maximum negative screen current. In some cases a Zener diode bleeder has been convenient.

Design Example

Suppose one must design an amplifier which will supply 3-1/2 kW to a 50- Ω load over a frequency range from 1 to 70 MHz. Suppose that we choose the Eimac 4CW2000 tetrode, which has a measured input capacitance of 130 pF and an output capacitance of 20 pF when mounted in the Eimac socket. With a screen grid voltage of 325 V and 0 V grid bias, the plate current is 2A. If the amplifiers operate Class A-1, the quiescent plate current should then be 1 A. For the rates anode dissipation of 2 kW, the anode voltage is 2 kV. It is convenient to design the anode line for 200 Ω --for then by means of a 2-to-1-ratio output transformer, the load impedance is 50 Ω . For 20 pF and 200 Ω , Fig. 3a shows a cutoff frequency of 70 MHz. The required inductance is 0.8 μH . Figure 3b shows that this inductance can

be obtained with seven turns and a diameter of 1 in. The length of the coil should be 1.35 times the diameter of 1.35 in.

For an output power of 3.5 kW, the peak RF plate voltage will be 1200 V and the peak load current will be 6 A. Since there must be an equal amount of current supplied to the terminating resistor, the tubes will have to provide 12 A; thus 12 tubes will be required. The grid line must have the same cutoff frequency as the anode line, so that the velocity of propagation will be the same in both lines. Figure 3b shows that for a 130-pF grid capacitance and 70-MHz cutoff frequency, the line impedance should be about 33 Ω . It would be more convenient if the grid line were 50 Ω . This requires that the input capacitance be 80 pF. By connecting 200 pF in series with the grid, the input capacitance is 79 pF. From Fig. 3a the required inductance is 0.2 μ H. Figure 3b shows that five turns with a diameter of 0.6 inches will meet this requirement. The coil length should be 0.81 in. We have found from experience that it helps to connect these grid networks by means of 50- Ω coaxial cable. The time delay in a short piece of coaxial cable is small compared with the time delay through the bridge-T network--the latter is about 1.5 ns, whereas the former is about 0.25 ns. There is, of course, a similar time delay between sections in the anode circuit. Here, however, the characteristic impedance of the lead connecting the inductances is close enough to 200 Ω to eliminate the need of a coaxial cable.

Next, consider the ferrite transformers in the anode circuit. The peak RF voltage is

1200 V. Since the transformer is essentially an autotransformer, the ferrite has to support only 600 V. There must be sufficient ferrite for the lowest operating frequency, which is 1 MHz. If we use a maximum flux density of 50 gauss, the design chart, Fig. 5, yields a value of 20 V per turn per square inch. Thus, the turn area product must be

$$NA = \frac{600}{20} = 30 \text{ square-inch turns.}$$

If too many turns are used, capacitive coupling from turn to turn will produce resonances within our pass band; it is best to use as few turns as is practical. A good solution would be a core with a cross section of 4 square inches and 8 turns. The twisted-pair line must have a Z_0 of 200 Ω and be able to carry 6A peak. The schematic diagram, frequency response, and photographs of this amplifier are shown in Fig. 6.

References

1. W. S. Percival, British Patent Specification No. 460562, applied for July 24, 1936.
2. E. L. Ginzton, W. R. Hewlett, J. H. Jasberg, and J. D. Noe, Distributed Amplification, Proc. IRE 36, 956 (1948).
3. D. A. Barge, Advances in High Power Distributed Amplifiers, IRE Transaction on Nuclear Science NS9 - No. 2, 19 (1962).
4. W. P. Johnson, "Radio Frequency System for the MSU Cyclotron," in Proceedings of the International Conference on Sector-Focused Cyclotrons and Meson Factories, CERN 63-19, May 1963, pp. 279-285.

Table I. Twisted pair transmission line*

Z_0	Wire size AWG No.	Insulation	Current rating at 50 MHz (peak amperes)
200	22	2 layers, 10-mil wall H. S. T. †	0.5
100	20	10-mil wall Teflon (hook-up wire)	0.7
90	18	10-mil wall Teflon (hook-up wire)	1.0
120	16	15-mil wall Teflon (hook-up wire)	1.5
110	14	1 layer, 15-mil wall H. S. T.	2.0
100	12	1 layer, 15-mil wall H. S. T.	3.0
120	12	2 layers, 15-mil wall H. S. T.	3.0
100	10	Formvar plus 3 layers, 10-mil wall, H. S. T.	4.0
65	10	Formvar plus 1 layer, 10-mil wall, H. S. T.	4.0
25	10	Formvar	4.0
100	1/8 in. O.D. tubing H ₂ O cooled	3 layers, 10-mil wall H. S. T.	50.0

* Twisted about 8 turns per ft.

† H. S. T. is an abbreviation for heat-shrinkable tubing.

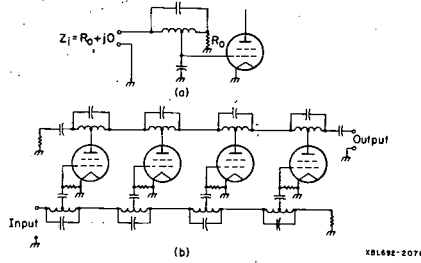


Fig. 1. Constant-resistance bridge-T network used (a) to convert the input capacitance of a vacuum tube to a pure resistance which is independent of frequency, and (b) to build the grid and anode lines of a distributed amplifier. Because the image impedance of such lines is purely resistive and frequency-independent terminating half sections are not needed. The bandwidth is about twice as great as for constant-K lines.

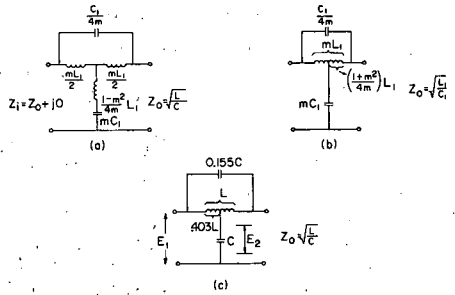


Fig. 2. (a) Constant-resistance bridge-T network. (b) The mid-shunt inductance may be obtained from the mutual inductance of the coil. (c) With $m = 1.27$ the time delay through the network varies by less than 1% up to half the cutoff frequency.

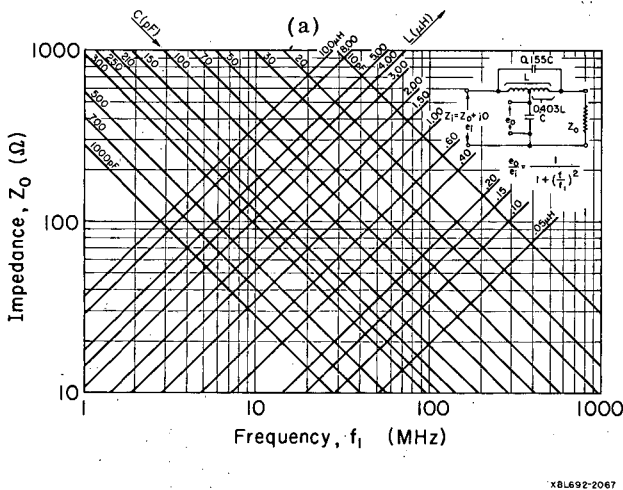


Fig. 3 (a) Design chart for bridge-T constant-resistance networks. (b) Design chart for the inductance of the constant-resistance network. If the ratio of the length to diameter of the coil is 1.35, the mutual inductance is correct for the networks of Fig. 3a.

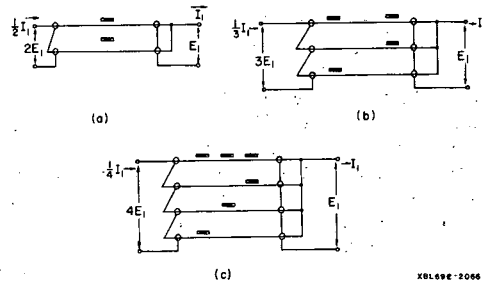


Fig. 4. Schematic diagram of ferrite transformers with 2-to-1, 3-to-1, and 4-to-1 voltage ratios. The input of the transmission lines are connected in series, whereas their outputs are connected in parallel. The ferrite core provides sufficient inductance to make the common-mode currents on the lines negligible.

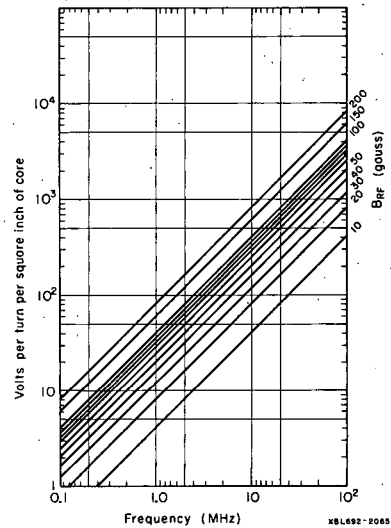
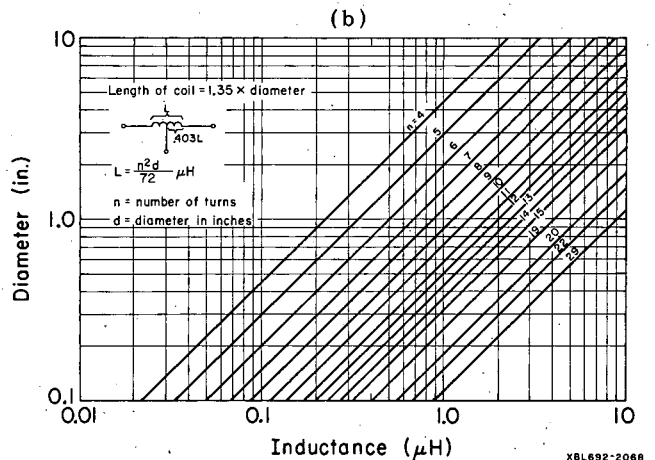
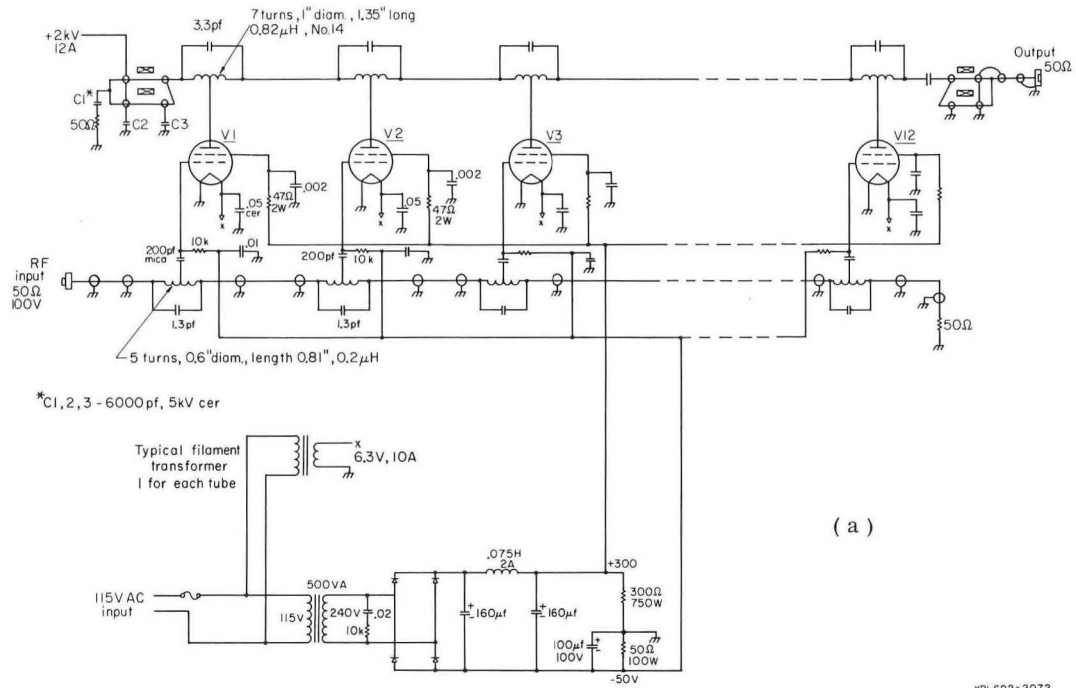


Fig. 5. Design chart for the ferrite cores of transmission-line-type transformers. The core area is selected on the basis of the maximum permissible B_{RF} at the low-frequency cutoff of the transformer.





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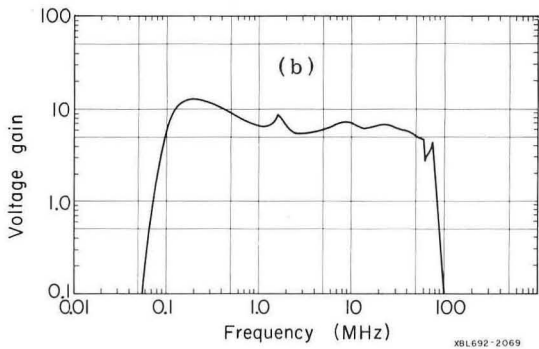
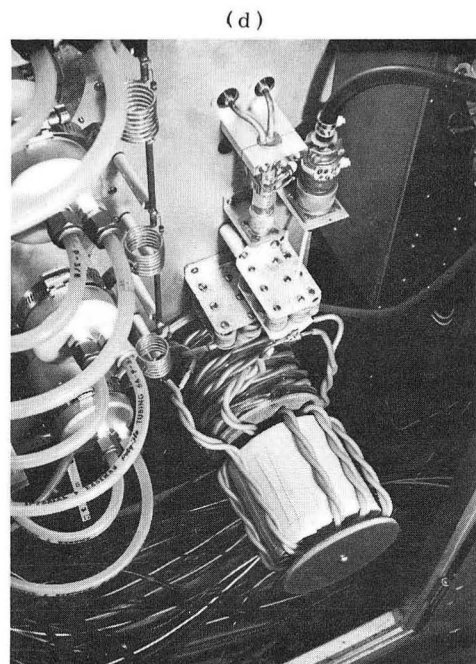
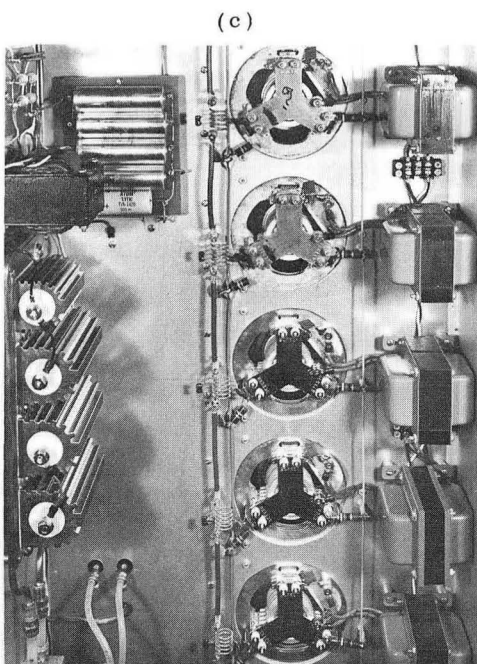


Fig. 6 (a) Schematic of the 3.5-kW amplifier used in the example. (b) Frequency response. (c) Grid-line construction technique. (d) Anode line and ferrite transformer.



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