An 80-Megabit 15-Watt Transistor Pulse Amplifier

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A transistor pulse amplifier delivering a 1.1-ampere, 11-nanosecond pulse at an 80-mc rate into a 15-ohm load has been designed. This amplifier was developed for use with an optical modulator. This paper describes the performance and gives the design and construction details of the amplifier.

I. INTRODUCTION

High-power broadband amplifiers operating into impedances of less than 50 ohms have not been previously available. With the advent of optical modulators a need has arisen for such amplifiers. In particular an optical modulator[1] designed for experiments on optical communication systems requires a one-ampere peak signal to produce a one-radian phase shift. This modulator consists of a one meter long strip line partially loaded with a KDP (potassium dihydrogen phosphate) dielectric. The line impedance is 15 ohms. A PCM system was proposed requiring an 80-mc pulse rate and 10 to 12 nanosecond (ns) raised cosine pulses. The use of raised cosine pulses was dictated by the low-level vacuum tube pulse generator in the PCM system. A suitable driver for this optical modulator must deliver a one-ampere peak pulse into the 15-ohm line over a band extending from near dc to at least 100 mc.

The discussion of the amplifier that realizes these requirements is divided into several parts. Section II is a general description of the completed amplifier, including the configuration used and the results obtained. Section III gives the performance that was obtained, including photographs of the response to both square wave pulses and raised cosine pulses at rates up to 80 mc. The design considerations are taken up in detail in Section IV, and the mechanical construction is described in Section V. A discussion of the results is contained in Section VI.

II. GENERAL DESCRIPTION

The complete amplifier consists of a power amplifier and a preamplifier. The power amplifier consists of a pair of VHF silicon power transistors coupled at the input and output with broadband transformer hybrids. This amplifier has a current gain of 3.2 and delivers a 1.02-ampere pulse 11-ns wide at the base into an output impedance of 15 ohms.

A three-stage transformer-coupled transistor preamplifier is used to drive the power amplifier. The preamplifier has a current gain of 5.3 and has a bandwidth larger than that of the power amplifier so that it has negligible effect on the pulse response. An input pulse of 60 ma into 50 ohms is required to achieve the full peak output power of 15 watts.

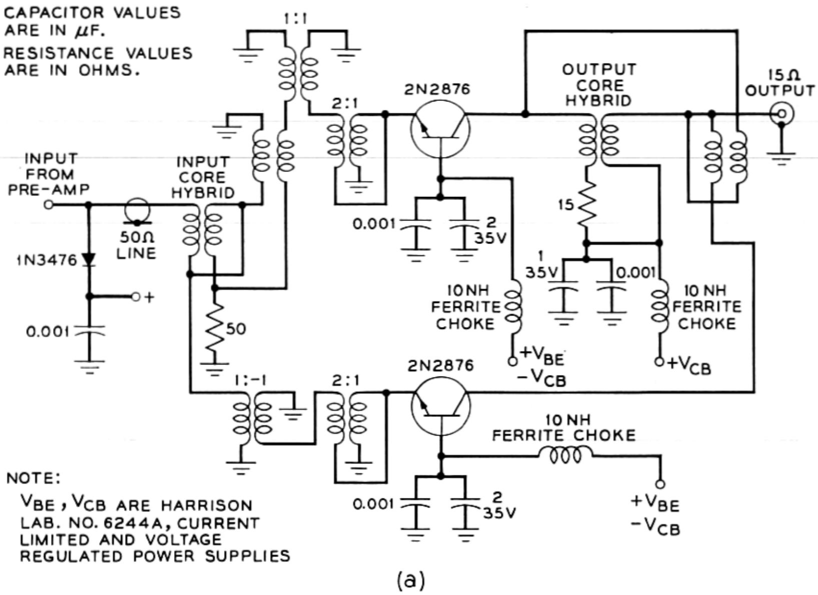
These amplifiers use the common-base configuration with current gain obtained by transformer coupling. The pulse response of the combined amplifier is limited by the parameters of the transistors used in the output stage. Under these circumstances, the common-base configuration is the only one which can be used to deliver the required power with the desired pulse response.

The complete circuit diagram is shown in Figs. 1(a) and 1(b). The transistors used in the output stage are RCA 2N2876; those in the preamplifier the RCA TA2307 (2N3375). The transformers and the hybrids are of the type described by Ruthroff.[2]

The total dc power required is 13 watts: 0.46 ampere at 28 volts. The power amplifier occupies a space of approximately 3-inches wide by 1-inch deep by $1\frac{3}{4}$-inches high. The preamplifier occupies a space of 1 by 1 by 4 inches. The total volume of the complete amplifier excluding the heat sinks is $9\frac{1}{4}$-cubic inches. A photograph of the complete amplifier is shown as Figs. 2(a), (b) and (c).

III. PEFORMANCE

The amplifier was tested with 8-ns raised cosine pulses recurring to rates varying from 10 mc to 80 mc and with 50-ns square wave pulses at 100 cps rate. The raised cosine pulses were generated by tube pulse generators supplied by A. F. Dietrich of Bell Telephone Laboratories. The pulse amplitude was variable up to a maximum of 15 volts into 50 ohms. All waveforms were observed on a Hewlett-Packard 187A sampling oscilloscope. The arrangement of the test circuit is shown in Fig. 3. A coaxial probe for one channel of the sampling scope was placed in the input line with a 3-db General Radio coaxial pad as iso-



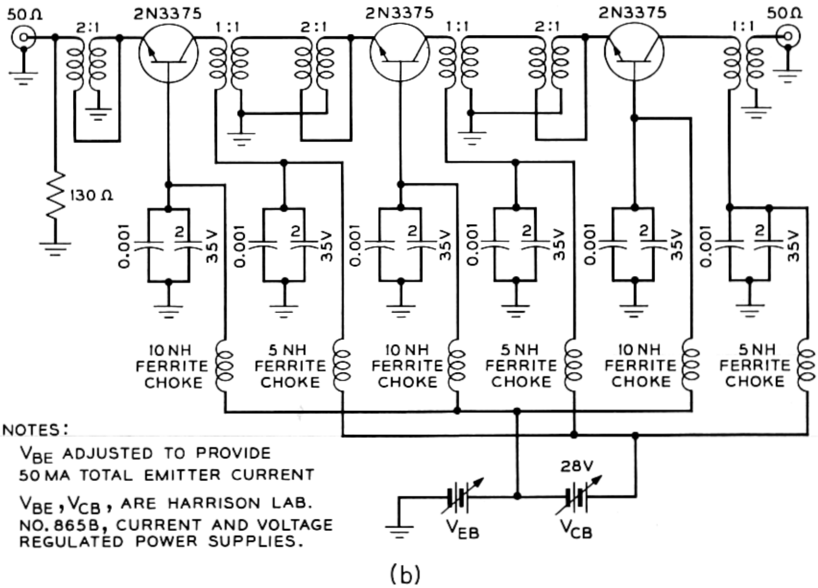


Fig. 1 — (a) Power amplifier schematic diagram; (b) preamplifier schematic diagram.

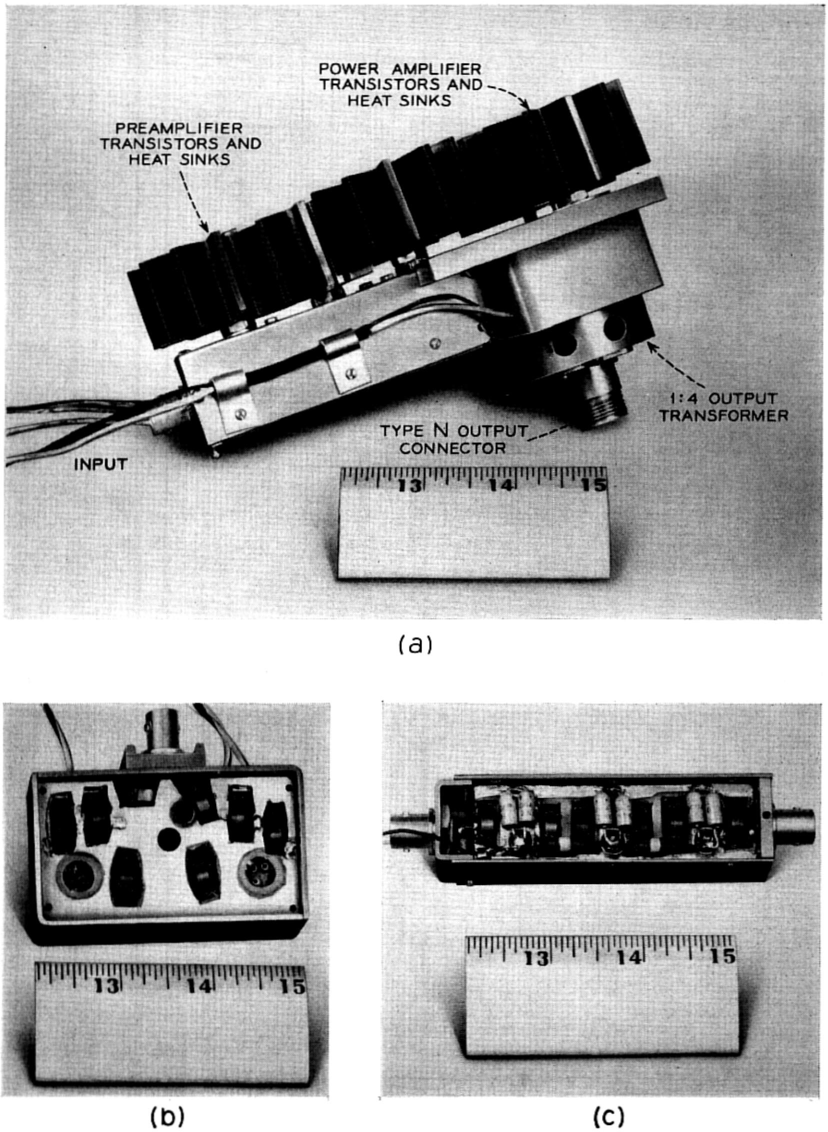


Fig. 2 — (a) Side view of complete amplifier; (b) top view of interior of power amplifier; (c) bottom view of interior of preamplifier.

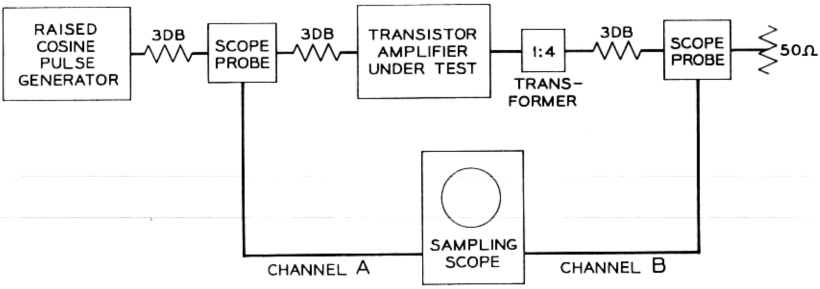


Fig. 3 — Test circuit for transistor amplifier.

lation between the pulse generator and the amplifier input. The coaxial probe of the second scope channel was placed in the amplifier output between a 3-db isolation pad and a 50-ohm, 10-watt termination. Coupling of the 15-ohm output of the amplifier to the 50-ohm test circuit was accomplished with a 4:1 core transformer and a length of tapered strip line. The 15-ohm side of this transformer was connected to the output hybrid of the amplifier through a 0.01-μf capacitor to provide dc isolation. The collector-base and base-emitter circuits of the power amplifier were supplied with dc bias from separate Harrison Lab 36-volt, 3-ampere current- and voltage-regulated power supplies. The same circuits in the preamplifier were separately supplied by Harrison Lab 40-volt, 0.5-ampere power supplies. This division of power supplies was used to allow a variety of bias conditions for test purposes.

The power amplifier was tested separately with an 8-ns pulse at a 10-mc rate. The pulse in the input line and the output pulse are shown in Fig. 4. The collector was biased with +28 volts and the zero pulse emitter bias was 20 ma. The output pulse amplitude corresponds to a 0.9-ampere pulse in 15 ohms. There is a large reflection evident in the input line as seen in Figs. 4(a) and (b). This is a consequence of designing the transformers to match the 50-ohm input line to the transistor input impedance of 6 ohms at the peak pulse amplitude. At lower amplitudes, the transistor input impedance is approximately $2\frac{1}{2}$ times greater and varies both with drive and frequency. The frequency variation of the input impedance is less at high drive than at low drive. The reflection in the input produces the small pulse following the main output pulse shown in Fig. 4(a).

To suppress the reflected pulse from the power amplifier, and prevent further reflection from the preamplifier with the attendant

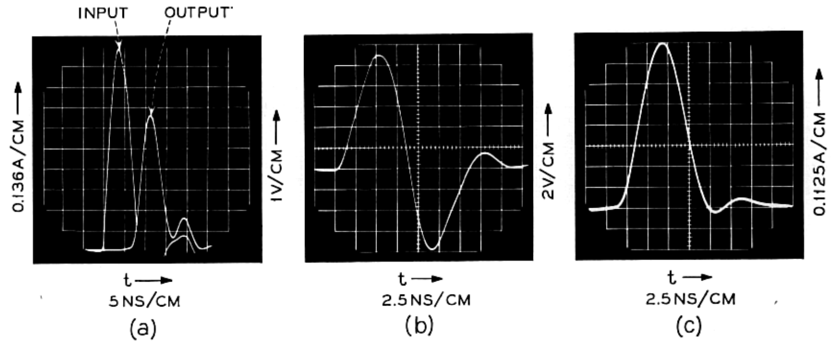


Fig. 4 — Power amplifier pulse response: (a) power amplifier input and output without compensating diode; (b) expanded view of amplifier input without diode; (c) amplifier output with diode compensation.

degradation in the output of the power amplifier, a Western Electric 1N3476 "pin head" diode was placed across the short section of 50-ohm line in the power amplifier input (Fig. 1). The diode anode was connected to the center conductor. The cathode was connected to a positive bias supply through a 0.001-μf feed-through capacitor. This diode has a 1-pf capacitance and will dissipate 200 mw. Thus the negative-going input pulse to the power amplifier will cause no diode conduction. The 1-pf capacitance will cause negligible loading. The reflected positive pulse will cause the diode to conduct. By adjusting the diode bias voltage, the amount of conduction and hence the impedance of the diode can be controlled. This arrangement damps out the reflected pulse. The effect of this damping is evident in the output pulse of Fig. 4(c). The pulse following the main pulse in Fig. 4(a) has been eliminated.

Fig. 5 shows a comparison between the measured wave shape and the calculated wave shape for this amplifier. The plot of the measured response was normalized so that the peak amplitudes of the measured and calculated responses were equal. The beginning of the output pulse was taken to coincide with the calculated pulse (thus eliminating the amplifier delay) so that a comparison of the pulse shapes could be readily seen.

The response of the power amplifier with the compensating diode to a 50-ns pulse with 0.5-ns rise time from the Spencer-Kennedy pulse generator is shown in Fig. 6. The input is shown in Fig. 4(a) with the amplifier output in Fig. 4(b). The rise time of the amplifier is approximately that shown in Fig. 5, and there is no indication of sag.

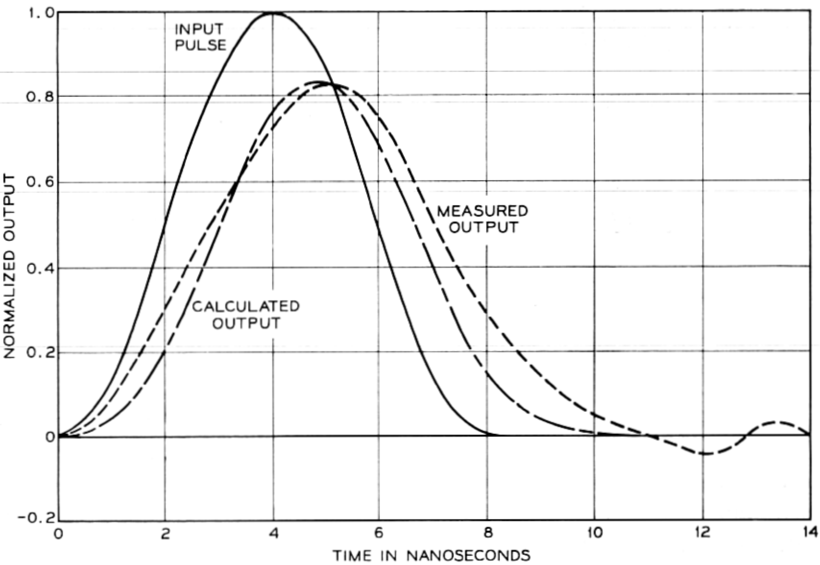


Fig. 5 — Comparison of calculated and measured pulse response of power amplifier.

The pulse response of the complete amplifier is shown in Fig. 7. Fig. 1(a) shows the amplifier output with 8-ns driving pulse at a 40-mc rate. The 35-ma input pulses (in 50 ohms) are 8-ns long. The 0.9-ampere output pulses measured in 15 ohms are approximately 11-ns long. Fig. 7(b) shows an expanded scale of Fig. 7(a). The output base line ripple is 0.076 ampere for a 0.008-ampere ripple in the input pulses.

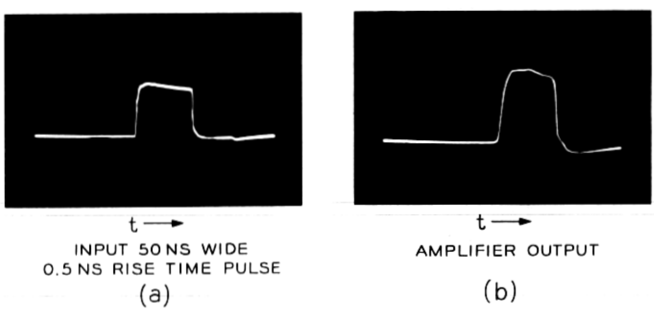


Fig. 6 — Amplifier step response.

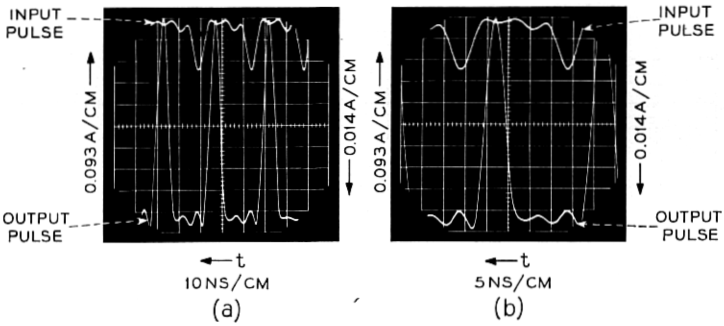


Fig. 7 — Amplifier pulse response, 40-mc pulse rate.

Fig. 8 shows the response to a group of pulses at an 80-mc rate. The output of the amplifier to a group of 4 pulses occurring at an 80-mc rate (12.5-ns pulses) is shown in Fig. 8(a). It was not possible to obtain sufficient output from the pulse generator to drive the amplifier with all 8 pulses. Fig. 8(b) shows an expanded view of the 4 pulses.

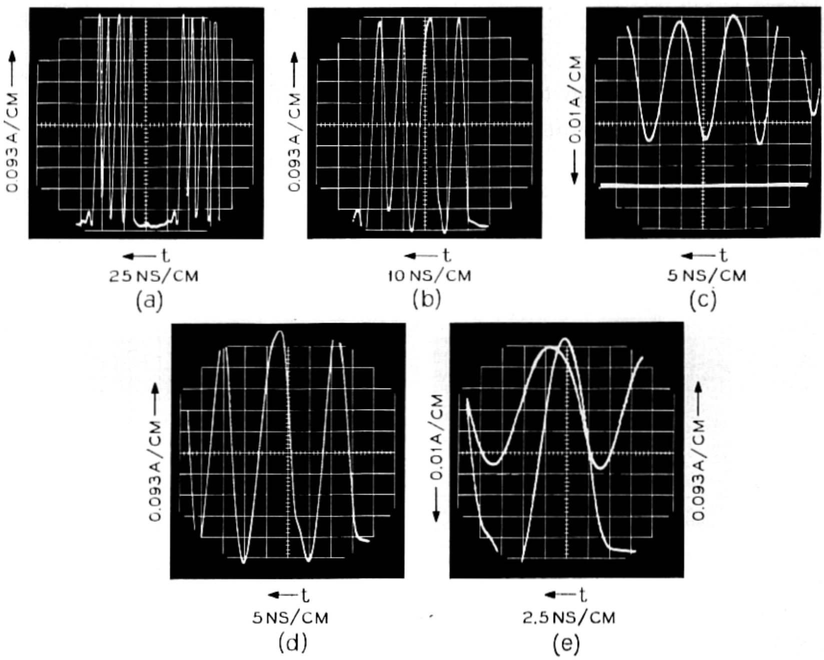


Fig. 8 — Amplifier pulse response, 80-mc rate.

These 0.93-ampere pulses measured in 15 ohms are approximately 12 ns wide. The 60-ma driving pulses and the 1.02-ampere output pulses measured in 15 ohms are shown in Fig. 8(c) and 8(d), respectively. An expanded view of these pulses is shown in Fig. 8(e). Here a 56-ma driving pulse produces a 0.93-ampere output pulse about 12-ns long. This corresponds to a total amplifier gain of 19.2 db.

IV. DESIGN CONSIDERATIONS

The design of the amplifier with the required specifications was divided into two stages. Since the power amplifier has the most difficult power and bandwidth requirements, it was considered first. With the power amplifier design completed, a preamplifier was designed so that the overall amplifier would have a gain of 20 to 23 db.

A number of low-level broadband transistor amplifiers have been built. These have been of the distributed type[3,4] and of the staggered tuned[5] variety. Both types have delivered only a few milliwatts of power into 50 ohms. These amplifiers have been operated in a linear region of the transistor characteristic. In addition to the linear range of operation, however, transistors can be operated in a pulsed mode. In this type of operation, the transistor is normally off, drawing no collector current until a driving pulse turns the transistor on.[6] In many applications, the input pulse is sufficient to drive the transistor into the saturation region where the collector current is dependent on the dc supply voltage and the load impedance only. This type of operation decreases the turn-on time of the output pulse, but increases the pulse length due to charge storage in the collector-base region. Such transistor switching amplifiers are usually operated in the common-emitter configuration with a passive network in the base lead to compensate for the fall-off in gain above the β cutoff frequency. Current gain can also be realized in the common-base configuration if suitable broadband transformers are available to match the transistor input and output impedances to the source and load.

Gartner[6] points out that the common-base power amplifier produces the same output power as the common-emitter amplifier but with less distortion. However, the power gain of the common-base amplifier is less than that of the common-emitter amplifier. Since this power amplifier must deliver a large peak pulse power with a minimum distortion of the pulse shape, and since sufficient over-all gain can be obtained in the lower-level preamplifier, the common-base configuration using broadband coupling transformers is best suited for this application.

The required broadband transformers have been developed by Ruthroff.[2] These transformers consist of a bifilar winding about a toroidal core of ferrite material. Hybrids, 4:1 impedance transformers, and 1:1 reversing transformers can all be assembled by combinations of the basic core transformer with appropriate interconnection of windings. Schematic representations of these three coupling devices are shown in Fig. 9.

Of the commercially available VHF power transistors, the RCA 2N2876 and TA2307 provided the best compromise in power dissipation, α-cutoff frequency and output capacitance for broadband high-power operation. These NPN transistors are of the triple-diffused silicon planar construction using a novel overlay construction. Either transistor will handle the required 3 watts of average pulse power. The use of a single common-base transistor with input and output transformers to achieve current gain would result in peak voltages that are close to or exceed the breakdown conditions of the transistors. To prevent this possibility and to allow cooler operation, two transistors are used in a hybrid coupled power amplifier.

The operation of the hybrid coupled amplifier can be understood with reference to the schematic diagram in Fig. 1(a). We can consider a unit amplitude current pulse incident upon the 50-ohm terminal of the input hybrid. The current pulse out at the conjugate terminals of the hybrid will be of unit amplitude at a 25-ohm impedance level. The polarity of the pulse at the conjugate arms will be opposite. In order to convert the positive pulse in one conjugate arm to the required negative pulse, to drive the transistor, a 1:1 reversing transformer is used. Since we want both transistor input circuits to have the same time delay, a 1:1 nonreversing transformer was inserted in the opposite conjugate arm. The negative unit amplitude current pulses are coupled to the emitter of each transistor through 4:1 impedance transformers. This arrangement provides a 2-unit amplitude negative pulse at a $6\frac{1}{4}$-ohm impedance level to each emitter-base circuit. The collector current of each transistor will be a 1.8-unit amplitude pulse since the α's of these power transistors are approximately 0.9. The output of the two transistors are combined in the output hybrid to produce a 3.6-unit amplitude negative pulse into the 15-ohm load. With a 15-ohm load, the collector load impedance of each transistor is 30 ohms. The current gain of the amplifier from the 50-ohm input to the 15-ohm output is 3.6 (a power gain of 5.9 db). Thus a one-ampere peak pulse in the 15-ohm load requires a 0.28-ampere driving pulse in the 50-ohm input line.

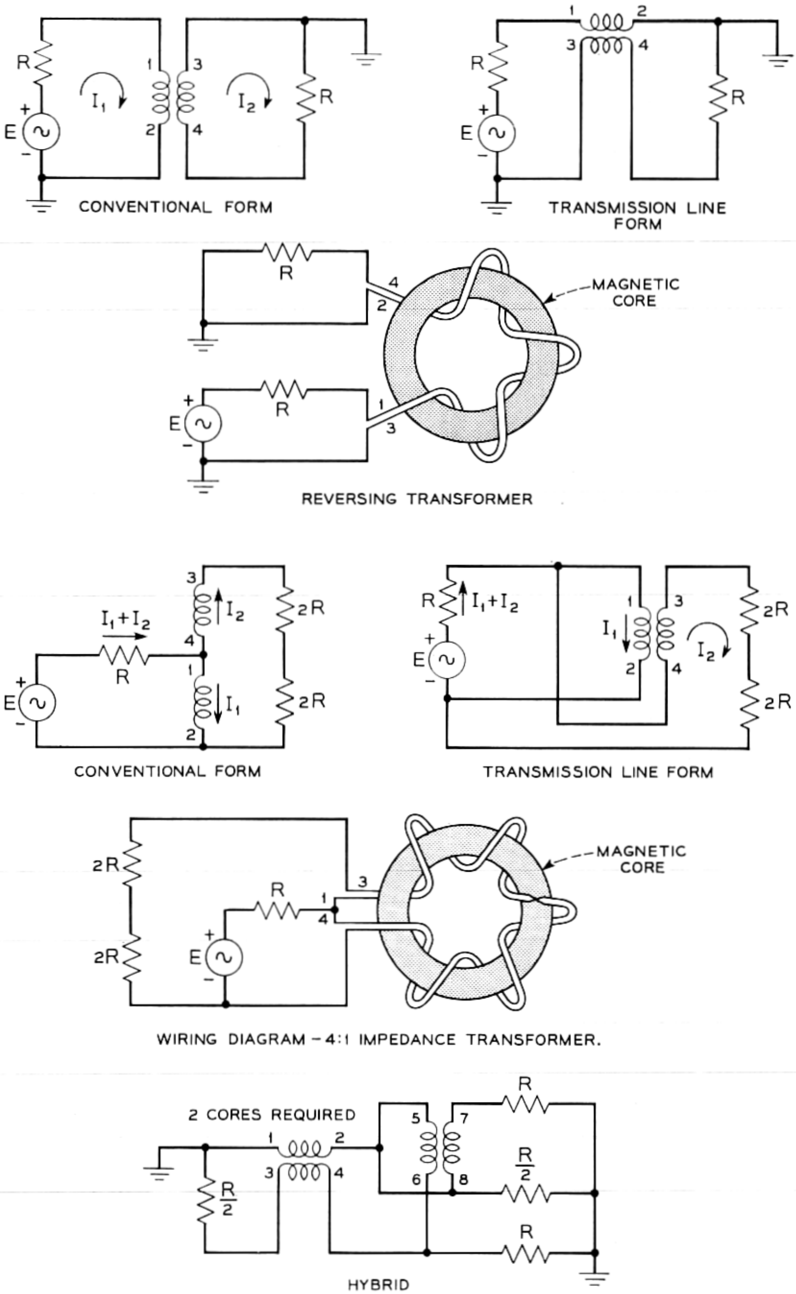


Fig. 9 — Typical core transformers used in amplifier.

The theoretical response of the hybrid coupled amplifier to a raised cosine pulse was determined by a combination of experimental and analytic investigation of the component parts. Ruthroff[2] has shown that the core transformers have a flat response from less than 1 mc to greater than 400 mc at low current levels. To determine what saturation effects might be present at peak currents of one ampere and greater, we tested a reversing transformer and a pair of 4:1 transformers placed back-to-back at various current levels from 0.1 ampere to 3 amperes. Transformers, bifilar wound with #20, #22, and #24 Formex wire, were tested with 0.5-ns rise time, 50-ns long pulses of varying amplitudes generated by a Spencer-Kennedy pulse generator operated at a 100-cps rate. Representative pictures taken from the Edgerton Germeshausen & Grier traveling-wave oscilloscope of the input and output pulses are shown in Fig. 10. It is evident from these pictures that there is no significant change in the pulse shape in going from a 0.1-ampere pulse to a 3-ampere pulse.

The amplitude and phase of the α of the 2N2876 transistor was measured as a function of frequency for two collector currents, 0.25 ampere and 0.5 ampere at 28 volts. The rather novel method of measurement is described in the Appendix A. Representative response curves are shown in Fig. 11. At 0.25 ampere, the α-cutoff frequency is approximately 200 mc while at 0.5 ampere it is 170 mc. These measurements confirm the manufacturer's nominal f_α of 200 mc.

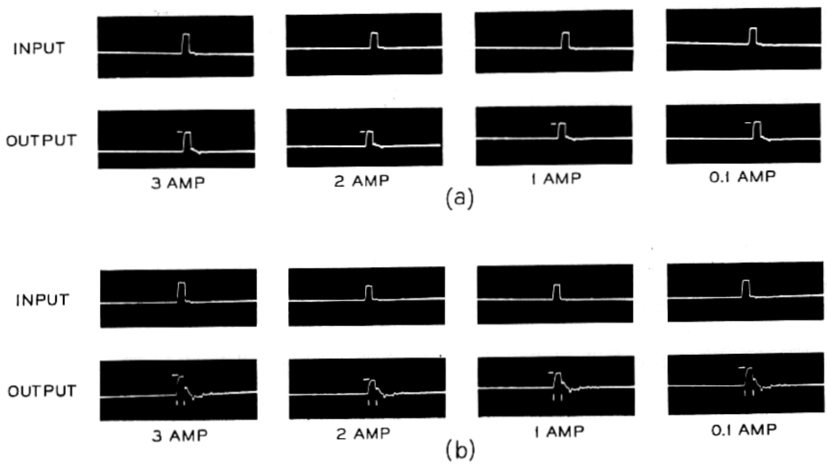


Fig. 10 — Top: reversing transformer, #20 wire; bottom: two 4:1 transformers back-to-back.

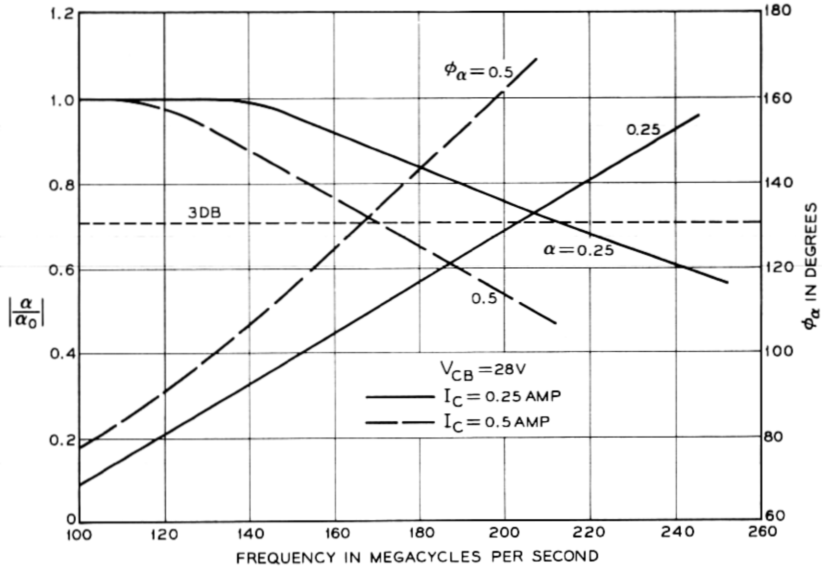


Fig. 11 — 2N2876 transistor measured magnitude and phase of short circuit current gain, alpha, as a function of frequency.

From these measurements it is evident that the response of the amplifier to a raised cosine pulse will be determined by the α of the transistor rather than the connecting circuitry. Gartner[6] has a general analysis of the step response of the common-base transistor operated in the nonsaturating switching mode. The response to a raised cosine pulse can be obtained by using a step approximation to the convolution integral.

The effect of the collector circuit RC cutoff frequency has been analyzed by Easley.[7] For the common-base configuration where ω_α RC_c is less than one, but of the order of one, the effect of RC_c is to alter the shape of the waveform and slow the response for times small compared to $1/\omega_\alpha$, i.e., in this case less than 1 ns. The details of this analysis are contained in Appendix B. Since some 400 calculations were required for each point of the response curve, the solution was programmed on an IBM 7090 computer. The response of a transistor with an α of 0.9, an ω_α of 200 mc, a collector capacity of 20 pf, and a load of 30 ohms to a raised cosine pulse 8 ns long is plotted in Fig. 12. This figure shows that the output pulse is broadened to 11 ns and decreased in amplitude by 17 per cent. Applying these results to the hybrid amplifier of Fig. 1, we see that the amplifier using the

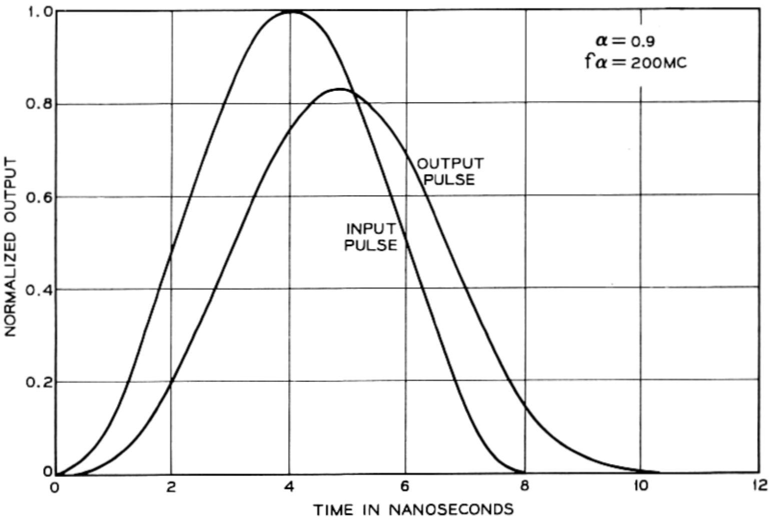


Fig. 12 — Calculated pulse response of 2N2876 transistor.

2N2876 transistor will have a current gain of 3.32 or 5.2 db operating from 50 ohms to 15 ohms. Thus for a 1-ampere output pulse we require a 0.3-ampere driving pulse. A preamplifier with 16-db gain will require a 48-ma peak current pulse to drive the power amplifier to full output.

The schematic diagram of a three-transistor preamplifier that will supply the required gain is shown in Fig. 1(b). This amplifier uses the common-base configuration with the ferrite core transformers to provide the current gain. To maintain a 200-mc bandwidth in the presence of the bandwidth reduction that results from cascading identical amplifiers, the TA2307 (2N3375) transistor with its α-cutoff frequency of 500 mc was used. Since the preamplifier operates at lower power levels than the power amplifier, the 11.5-watt dissipation of the TA2307 is adequate. As in the power amplifier, the transistors in the preamplifier are operated in a nonsaturating switching mode.

The pulse response of a single 2N3375 transistor was calculated in the same manner as above (see Appendix B). The width of a unit amplitude pulse 8 ns wide is increased to 8.3 ns and the amplitude is decreased by 12 per cent, yielding a current gain of 0.88 per transistor.

The operation of the preamplifier between 50-ohm impedances can be understood with reference to the schematic diagram, Fig. 1(b). A negative unit amplitude pulse incident upon the input of the first 4:1

impedance transformer produces a 2-unit pulse to the emitter of the first transistor at a 12.5-ohm impedance level. The real part of the input impedance of these transistors ranges from 12 to 20 ohms, depending on pulse amplitude. The collector current, using the effective α of 0.88, is then 1.76 units. A 1:1 transformer is used to isolate the collector dc bias from the 4:1 input transformer of the next stage. Since each succeeding stage is identical, the collector current of the output of the preamplifier is 5.43 units. The current gain is 5.43, or a power gain of 14.7 db. For the 0.3-ampere output pulse, a driving pulse of 55 ma in 50 ohms is required. This corresponds to a peak power of 150 mw per pulse which is available from most pulse generators. Since the transistors and the core transformers have bandwidths of at least 400 mc, the bandwidth of the amplifier will be limited to less than 400 mc by the effect of cascading. The reduction in bandwidth due to transformer coupling will be less than the reduction caused by cascading RC coupled amplifiers. Three RC coupled stages using these transistors would have a bandwidth of 204 mc.

V. MECHANICAL CONSTRUCTION

The basic amplifier construction uses a quasi-stripline to interconnect the core transformers and transistors. The input and output circuits are placed on opposite sides of a center board in nonoverlapping areas. Ground plane boards are placed on top and bottom of the common board. This results in a 3-layer sandwich construction. The center board is $\frac{1}{8}$-inch thick glass-loaded Teflon with 2-oz copper on each side. The ground plane boards are $\frac{1}{16}$-inch thick glass-loaded teflon with 2-oz copper on each side. Transistor sockets were made by placing a spring contact for each pin on the transistor case into holes drilled in the center board. These contacts were made using the center conductor of BNC female coaxial connectors. After being placed into the center board, these contacts were soldered to the copper. Matching rectangular holes were cut in each board to accommodate the core transformers. The copper on the board surfaces was cut away so as to leave stripline connections between the cores and transistors. The boards were separated by placing $\frac{3}{16}$-inch thick brass spacers around the periphery of the boards. Where this spacing was not adequate for the proper stripline impedance, brass pieces were attached to the center board to decrease the spacing in a local region. The terminations for the input and output hybrids, 50 ohms and 15 ohms, respectively, are 1 watt 1 per cent metalized film microwave rod resistors made by

Film Ohm Corp. The base by-pass capacitors consist of a 0.001-μf feed-through capacitor mounted in the ground plane with the center connected to the base terminal of the center board. Two 2-μf miniature tantalum capacitors were used in parallel with the feed-through capacitor. The collector dc connection is by-passed with a 0.01-μf postage stamp type capacitor in parallel with a 2-μf miniature tantalum capacitor. Collector bias is applied between collector-base terminals. Emitter bias is applied between base terminals and ground.

The layout of the three circuit boards for the power amplifier is shown in Fig. 13. Both sides of each board are shown. The layout of the 3-transistor circuit board of the preamplifier is shown in Fig. 14. The positions of the parts and of the sections of stripline are evident upon examination of these figures.

The outside dimensions of the circuit boards were dictated by the space available inside the KDP modulator structure and by a need to keep all leads as short as possible. To keep the connection between the power amplifier and the KDP stripline to a minimum, the power amplifier had to be placed in a space 1-inch deep, 3-inches wide, and $1\frac{5}{8}$-inches high. A mounting box of these outside dimensions was made of $\frac{1}{16}$-inch brass and the circuit board sized so as to fit inside this box. The transistor heat sinks — 2-inch diameter, 1-inch high, 12-fin aluminum cylinders — were extended out through the end plate of the KDP mounting box. The heat sinks are mounted on the transistor package stud and are kept isolated from ground. Since the collector of the 2N2876 and TA2307 are isolated from the case (collector-to-case capacitance 6 pf), the mounting of the heat sinks in this manner reduced the collector circuit capacitance to a value close to the transistor collector capacitance of 20 pf for the 2N2876 and to 10 pf for the TA2307.

The cores in the output hybrid of the power amplifier are bifilar wound with 6 turns of #22 Formex wire. The transformer cores and the input hybrid cores are bifilar wound with 6 turns of #24 Formex wire. All the transformer cores in the preamplifier except the output transformer are bifilar wound with 6 turns of #24 Formex wire. The output transformer core is bifilar wound with 5 turns of #24 Formex wire.

The size of the preamplifier was also dictated by the dimension of the KDP modulator structure. A narrow 1-inch by 1-inch passage existed between the top of the power amplifier and the top wall of the modulator structure. The preamplifier was mounted in a $\frac{3}{64}$-inch thick brass box, 1 by 1 by 4 inches, mounted on top of the power

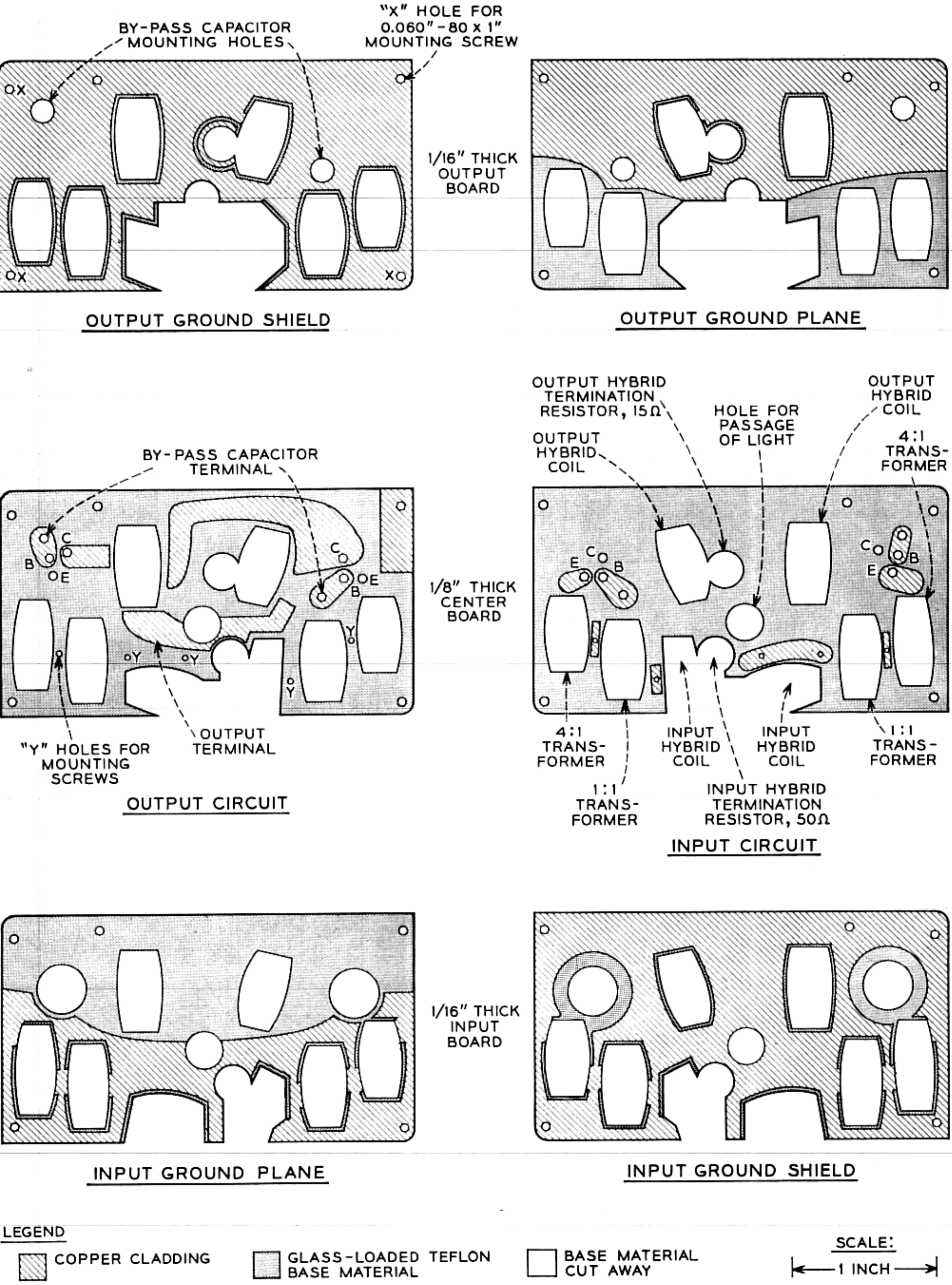


Fig. 13 — Power amplifier circuit boards.

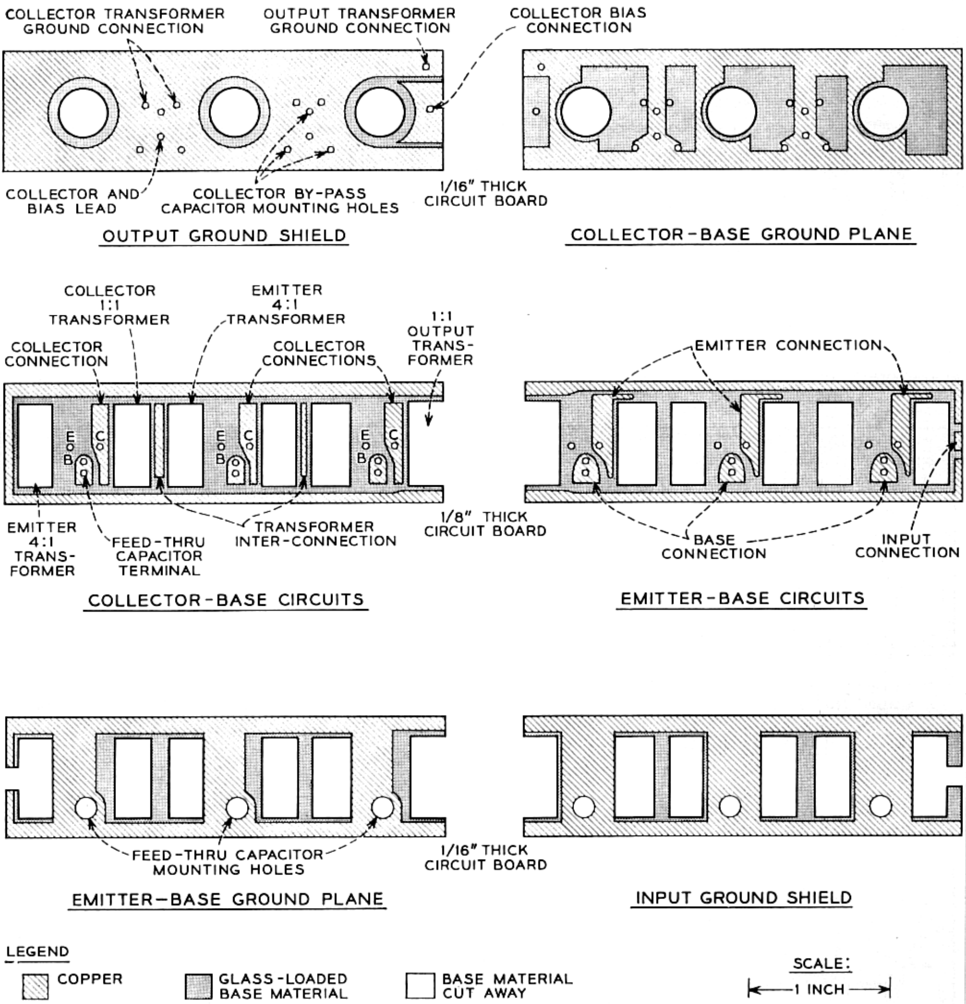


Fig. 14 — Preamplifier circuit boards.

amplifier and extended through the top of the modulator structure. Connection of the preamplifier with the coaxial output of the modulator source was made through a BNC coaxial connector. The circuit boards of the preamplifier were cut to dimensions to fit this box. The heat sinks for the preamplifier are of the same type as those of the power amplifier. Certain fins of each heat sink were cut away so that the heat sinks of successive preamplifier transistors could interleave. The details of the complete amplifier can be seen in Fig. 2.

VI. CONCLUSIONS

This experimental transistor amplifier has demonstrated that broadband high-power transistor pulse amplifiers operating into impedances less than 50 ohm can be designed and built with existing commercial transistors. In the common-base configuration with broadband transformer coupling, the amplifier design depends on the transistor switching analysis and the RC product in the output circuit. In addition, the experimental work with these bifilar wound core transformers shows that these transformers can handle short rise time pulses of several amperes without saturation effects.

Several amplifiers of this type have been constructed. Their performance is equivalent to the amplifier reported in this paper. One of these amplifiers has been in daily use for six months with no degradation in performance.

These results also support Easley theory on the effect of ω_α RC_c products that are of order unity but less than unity. Easley in his paper[7] discusses this effect theoretically but only examines experimentally the effect of ω_α RC_c on the common-emitter and common-collector configurations. For these latter transistor configurations, the effect of ω_α RC_c is to lengthen the rise time by $(1 + \omega_\alpha RC_c)$. If this theory were valid for the common-base configuration, the output pulse width would be doubled.

It is evident from the experimental results obtained with this common-base amplifier that the pulse width is not broadened appreciably from that due to the transistor alpha. Since the ω_α RC product for this amplifier is 0.75, the results are in accord with the theory. Further support of Easley's theory is evident in the small distortion present in the measured output.

The results obtained with this amplifier justify the use of the common-base configuration when the highest frequency of operation is of the same order of magnitude as the f_α of the transistor. The use of a common-emitter configuration would result in the pulse-broadening noted above and would require additional circuits in the base or emitter leads to compensate for the frequency variation of the transistor beta. While in theory such compensation is possible, the addition of lumped circuit elements in high-frequency circuits introduces parasitics which would adversely effect the amplifier performance.

A problem of conditional stability is associated with the use of the common-base amplifier. The possibility of oscillation can be avoided by careful isolation of the input and output circuits and by the proper choice of the bias conditions and the load resistance. The isolation of the input and output of each transistor in the amplifier was achieved with the three-layer sandwich construction. The choice of the load resistance and the bias eliminated any tendency to oscillate.

The use of a hybrid-coupled transistor amplifier for the power amplifier increases the maximum power by 3 db for the same transistor rating without paying a penalty in the RC cutoff frequency in the individual transistor output circuit. It is evident that 2N transistors could be coupled by tandem hybrids to obtain a 3N db increase in output power without a significant bandwidth reduction over the single amplifier circuit. As an example, consider the circuit for a power amplifier delivering a 2-ampere pulse into a 15-ohm load. Four of the present hybrid coupled power amplifiers would have their inputs and output coupled by core hybrids. Since the input and output currents are additive, we require a 0.15-ampere pulse to produce a 2-ampere, 11-ns pulse at an 80-mc rate.

The success of these amplifiers is due in part to the ability of J. W. Batton to accurately construct and assemble the strip line connections.

APPENDIX A

Measurements of Transistor Alpha Cutoff Frequency

To determine the response of a transistor pulse amplifier in the common-base configuration, it is necessary to know the variation of the short circuit current gain alpha, with frequency. The manufacturer lists the f_T, a gain bandwidth product, of the 2N2876. Since this transistor is of a new design, the relation between f_T and f_α was not known.

A coaxial test jig was designed to measure both the magnitude and phase of alpha as a function of frequency. The drawing of the test jig is in Fig. 15. The diameters of the inner and outer conductors are the same as the General Radio 50-ohm air coaxial line. A phenolic transistor socket was mounted in a hole cut in the outer conductor. The base terminal was grounded directly to the inner wall of the outer conductor. The emitter terminal was connected to the center conductor through a parallel RL circuit with a cutoff of 140 mc. The collector terminal was connected directly to the center conductor with a short

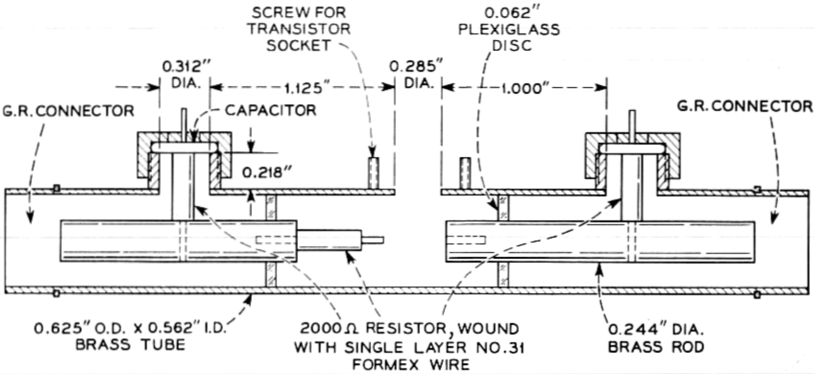


Fig. 15 — Transistor —test jig.

length of #22 wire. Bias connections to the center conductor were supplied by by-passed terminals in the outer wall of the test jig. Silver mica button capacitors, 0.001 μf, were soldered to the outer wall to provide RF by-passing. The bias terminal of these capacitors was connected to the center conductor through a parallel RL circuit with 140-mc cutoff frequency.

The arrangement of the test circuit is shown in Fig. 16. The emitter side of the test jig is connected through a 20-db pad to the coaxial probe of one channel of a Hewlett Packard 185/187A sampling oscilloscope. The coaxial probe was connected through a 3-db pad to a General Radio mixer rectifier. The low-frequency output of this mixer (dc to 30 mc) was connected to the sync terminal of the scope.

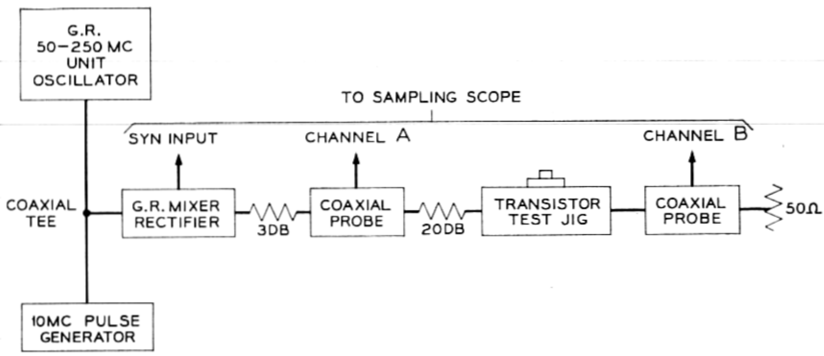


Fig. 16 — Test circuit for measurement of alpha,

The mixer input was fed from a 10-mc pulse generator and a General Radio 50- to 250-mc unit oscillator through a coaxial tee. The oscillator output was much greater than the pulse amplitude. This arrangement provides a 50- to 250-mc sine wave to the test jig and a sync signal in the 100-kc to 30-mc range that the scope will accept. The output of the test jig is connected to a 50-ohm termination through the coaxial probe for the other scope channel.

A measurement of the magnitude and phase of alpha is made in the following manner. The emitter-collector terminals are shorted together with a low inductance plate that covers the transistor socket except for the base terminal region. The input and output waveforms were displayed on the dual traces of the scope. The amplitudes of the two waves and the difference between the zero crossings were noted. The difference in zero crossing in centimeters was converted to degrees. The short circuit was removed and the transistor was inserted in the socket. The required bias conditions were set up, normally a collector-base voltage of 28 volts and a collector current of 250 ma or 500 ma. The RF input amplitude was adjusted to the value noted on the scope with the short present. The output amplitude and the difference in zero crossing between the input and output wave were determined from the scope. The magnitude of alpha is the ratio of the output amplitude to the input amplitude. The phase of alpha is the difference of the difference in zero crossings with the short and with the transistor. This procedure was repeated at a number of frequencies so that a complete curve was obtained.

The validity of this measurement procedure was checked by measuring Western Electric transistors whose alpha vs frequency curves were available. The measurements checked within 10 per cent of the nominal value curves of the transistors. The accuracy is determined by the impedance of the socket short circuit and the ability to read vertical and horizontal dimensions on the waveform displayed on the scope.

APPENDIX B

Analysis of Transistor Pulse Response

The analysis of the transistor pulse amplifier in the common-base configuration follows directly from the work of Gartner.[6] From the physics of the transistor, the short-circuit current gain is defined as

$$\alpha = \operatorname{sech} \frac{w}{L_{pB}} (1 + j\omega\tau_{pB})^{\frac{1}{2}} \tag{1}$$

where

$$\begin{aligned} \omega &= \text{frequency} \\ w &= \text{base width} \\ L_{pB} &= \text{base diffusion length} \\ \tau_{pB} &= \text{base lifetime.} \end{aligned}$$

We define

$$a = \frac{w}{L_{pB}}$$

$$b = \frac{w}{L_{pB}} \sqrt{\tau_{pB}}$$

$$s = j\omega$$

and rewrite (1) as

$$\alpha = \operatorname{sech} (a^2 + b^2 s)^{\frac{1}{2}}. \tag{2}$$

The low-frequency α is determined from (2) by letting $\omega = 0$.

$$\alpha_{N0} = \operatorname{sech} a. \tag{3}$$

The short-circuit current gain is also defined as

$$-i_c(t) = \alpha i_E(t).$$

Applying (2) and taking reverse Laplace transforms leads to

$$-i_c(t) = \mathfrak{L}^{-1}[(\operatorname{sech} [a^2 + b^2 s]^{\frac{1}{2}}) \ \mathfrak{L}(i_e[t])]. \tag{4}$$

If $i_e(t)$ is a step of emitter current, i_{E1}, the collector current becomes

$$\frac{-i_c(t)}{\alpha_{N0} i_{E1}} = \frac{1}{\alpha_{N0}} \mathfrak{L}^{-1} \left[\frac{\operatorname{sech} (a^2 + b^2 s)^{\frac{1}{2}}}{s} \right]. \tag{5}$$

To solve (5) we write the hyperbolic secant in a series expansion and apply the tables of integral transforms. This yields

$$\begin{aligned} \frac{-i_c(t)}{\alpha_{N0} i_{E1}} = \frac{1}{\alpha_{N0}} \sum_{n=0}^{\infty} (-1)^n \Bigg\{ & \exp [(2n + 1)a] \operatorname{erfc} \left[\frac{(2n+1)}{2} \frac{b}{\sqrt{t}} \right. \\ & \left. + \frac{a}{b} \sqrt{t} \right] + \exp [-(2n + 1)a] \operatorname{erfc} \\ & \cdot \left[\frac{(2n+1)}{2} \frac{b}{\sqrt{t}} - \frac{a}{b} \sqrt{t} \right] \Bigg\} \end{aligned} \tag{6}$$

where erfc is the complementary error function.

We define the alpha cutoff frequency ω_N as

$$\frac{|\alpha|}{a_{N0}} = \frac{1}{\sqrt{2}}.$$

Using (1) for α we obtain

$$\omega_N = K/b^2$$

where $K = 2.64$ for $\alpha_{N0} = 0.98$.

This step response of the transistor can be used to find the response to arbitrary waveforms. The arbitrary waveform can be approximated by a stair case function — i.e., a series of finite steps. This step-wise approximation can be expressed mathematically as

$$i_{e_N}(t) = \sum_{\alpha=0}^{n} \delta(i_\alpha)\, S_{-1}(t - \alpha\delta t) \tag{7}$$

where $S_{-1}(t)$ is a symbol for the unit step and n is the largest integer, so that

$$t - n\delta t > 0.$$

Equation (6) can be written as

$$-i_c(t) = A(t) i_{E1}(t).$$

The approximate collector current for an arbitrary emitter current becomes

$$-i_c(t) = \sum_{\alpha=0}^{\infty} \delta(i_{e\alpha}) A(t - \alpha\delta t) S_{-1}(t - \alpha\delta t). \tag{8}$$

For a raised cosine emitter pulse defined by

$$i_e(t) = I_0 \cos^2 \pi \left(t - \frac{t_C}{2}\right)$$

equation (8) becomes

$$-i_c(t) = \sum_{\alpha=0}^{t_0} \delta[i_{e\alpha}(t)] A(t - \alpha\delta t) S_{-1}(t - \alpha\delta t)$$

where $\delta[i_{e\alpha}(t)]$ is

$$\delta[i_{e1}(t)] = i_e(\delta t) - i_e(0)$$

$$\delta[i_{e2}(t)] = i_e(2\delta t) - i_e(\delta t), \text{ etc.,}$$

and $A(t - \alpha\delta t)$ is written in a like manner. The solution of (8) was obtained using an IBM 7090 computer with $\delta t = 0.1t_0$ and several values of α_{N0}.

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