Error-Controlled High Power Linear Amplifiers at VHF

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Two amplifiers, each covering a different band in the VHF range, were constructed for test set applications. These amplifiers were required to meet stringent specifications beyond the general scope of previous art. These specifications related to the simultaneous availability of relatively large output power, tight linearity, minimal delay distortion across a 40 per cent band, high input and output return loss, large dynamic range, and time stability.

The performance problems were solved by using:

(*i*) *Quadrature couplers in a corporate structure array to provide a high multiplicity of "paralleled" transistors, yielding both the power and return loss capabilities.*

(*ii*) *The emitter follower configuration as the basic amplifier element, providing broad band, high level performance owing to the large self degeneration of that configuration.*

(*iii*) *A "feed-forward," instead of a feedback, system for error control and noise cancellation.*

I. INTRODUCTION

There are times when problems arise in device implementation that appear to be the product of the sheer perversity of the system designer. Worst of all, he can justify his need—as in the need to develop a solid-state 10-watt amplifier between 60 and 90 Mc, with low intermodulation distortion, time delay distortion, and gain ripple, with a high return loss, and with excellent time stability. Subsequently, a similar need was expressed over a band of 25 to 35 Mc.

A format of amplifier design was devised meeting these requirements and subsequent constructions justified the design philosophy. While some of the techniques used are known, we believe that the nature of their use here emphasizes latent potentialities which had not been generally apparent. In particular, a "feed-forward" error compensation system is used which has parallels to McMillan's and Van Zelst's generalization of Black's error injection methods.[1–3] There is a very specific recognition here, however, of the physical role played by time, which dominates the form of the error network and which provides, among other advantages, a means of arbitrary error control of large time delay amplifiers which are intractable to feedback methods.*

We shall describe further the use of "corporate" amplifier binary arrays using quadrature couplers and emitter-follower amplifier stages. The word "corporate" is intended to suggest the process of a corporate coordination of many sources to deliver their powers in some desired collective mode, in much the same fashion as in a corporate antenna. With respect to the emitter-follower amplifier, much to the contrary usage of that device only as a unit gain isolation stage, there are many advantages highly germane to its use as an amplifier with certain basic advantages over the common emitter and common base modes of transistor operation.

The name "feed-forward" error control scheme, coined to distinguish it from feedback, is fully descriptive. Error is detected, amplified, and injected after proper time delay in the forward time stream of the amplifier, and error cancellation is accomplished over the entire band of interest. The error, it should be emphasized, is composed of ripples in the amplifier transfer function, distortions through nonlinearity, and amplifier noise. To some extent the low level error amplifier has noise; nevertheless, there is advantage gained in noise performance by exchanging that noise for the much greater noise of the higher power amplifier. Further, the feed-forward system makes demands on the active elements of the system only within the band of interest, and great advantage is gained in using more powerful, lower quality transistors without the concerns of high frequency dispersion instabilities attendant upon conventional feedback systems.

II. CORPORATE POWER STAGES

2.1 *Coupler Arrays*

High-frequency transistors generally have a smaller dynamic range than the lower frequency and power transistors. One may, however,

* During the late stages of preparing this manuscript, we learned of a patent issued by the German Federal Republic to Heinrich Schmidt-Brücken in which he recognizes time as a factor in error compensation. In Appendix 3 we compare his implementations with present implementations, and we believe the differences are significant.

resort to transistor arrays to multiply the power capability inherent in a single stage while retaining the high-frequency performance. The notion of an array of elements to provide power is well within the framework of conventional electronics as observed in the early ring oscillator, an antecedent of the existing distributed amplifiers and oscillators.

In the ring oscillator, as in the magnetron, there is a coordination of the various elements into a collective mode. These modes comport with the rotational symmetry of the ring and so are characterized by eigenvalues which are the various rational roots of unity. Often the modes are virtually degenerate and mode strapping and mode killing techniques are a necessity to isolate the desired mode of operation. This isolation is frequently marginal and mode-hopping is a well-known consequence of improper excitation or loading of the system.

The difficulty in mode suppression stems from the requirement that, for an n degree of freedom system, $n - 1$ modes must be suppressed without hampering the transmission qualities of the nth. Further, reactive strapping may operate, not so much to suppress, but simply to alter the dispersion characteristics of the network permitting the unwanted modes to pop out elsewhere and, often, with deleterious effects.

Seidel has examined the problems of active network synchronization,[4] and the resolution of the problems of stability would appear to exist through the use of hybrid or conjugate couplers with absorbing suppressors. Basically, through the use of couplers, the terminating loads on the ideally unexcited ports provide exactly the $n - 1$ independent absorbers needed; furthermore, these absorbers are precisely where they should be to provide the necessary orthogonality against absorption of the desired mode.

As opposed to the ring, the mode structure of the hybrid coupled amplifier is not characterized by a simple eigenvalue description. Further, the forms of building the coupler array may be quite variable and the choices reflect on considerations of some moment. There are two particular considerations which are of major consequence.

(*i*) Out-of-band stability of the harnessed system

(*ii*) Low level of multiple reflection in interconnection of amplifier stages.

While these two considerations are phrased independently, a small interdependence can exist.

With respect to out-of-band stability, assume the possible use of a transmission line hybrid such as, for example, a rat race or 3 dB in-phase coupler. While meeting proper in-band specifications on match or power, what are the consequences of performance of this harnessing structure upon the element whose activity persists well beyond this point? Further, with respect to the second consideration, what are the proper harnessing formats to permit arbitrary cascading of elements whose intrinsic isolations may, at best, be modest?

2.2 *Quadrature Couplers*

A solution which contains many practical advantages is the use of arrays of lumped quadrature couplers. A quadrature coupler, as an elementary device, is a matched reactive reciprocal four-port having two orthogonal axes of symmetry. This characterization suffices to endow the coupler with the following properties:

(*i*) Corresponding to an excitation port, labelled 1, there exists a conjugate, or hybrid, port labelled 4.

(*ii*) The entire incident power emerges in quadrature phase from the remaining ports 2 and 3.

An evident symmetry of these coupling characteristics obtains with respect to rotations about the two symmetry axes. The about results are demonstrated in Appendix 1.

When we apply the device symmetries to the above labelling scheme we find the following scattering matrix:

$$S = \begin{bmatrix} 0 & S_{12} & S_{13} & 0 \\ S_{12} & 0 & 0 & S_{13} \\ S_{13} & 0 & 0 & S_{12} \\ 0 & S_{13} & S_{12} & 0 \end{bmatrix} \tag{1}$$

Application of the unitary properties of the scattering matrix, namely the relationship $\tilde{S}S = 1$, yields

$$S_{12}S_{13}^* + S_{13}S_{12}^* = 0 \tag{2}$$

$$S_{12}S_{12}^* + S_{13}S_{13}^* = 1. \tag{3}$$

From (2) $S_{13}^* = -S_{13}(S_{12}^*/S_{12})$ which we substitute in (3) to provide

$$S_{12}^2 - S_{13}^2 = \frac{S_{12}}{S_{12}^*}. \tag{4}$$

Equations (2) and (4) may be put in a more useful form. Equation (2) provides that Re $S_{12}S_{13}^* = 0$ so that $S_{12}S_{13}^*$ is imaginary implying that S_{12} and S_{13} are in phase quadrature. Equation (4) leads to a result which we shall employ crucially:

$$|S_{12}^2 - S_{13}^2| = 1. \tag{5}$$

Equation (5) is of major importance, since it states that it is possible to use quadrature couplers in an all-pass arrangement, as shown in Fig. 1, without regard to the frequency sensitivity of S_{12} and S_{13}. Since this result is independent of the frequency sensitivities of the matrix elements of S, it is possible to use the structural elements of Fig. 1 to implement far more complex power divider arrays without sharpening the frequency characteristics of transmission. Figure 2 shows a four-way all-pass division and the extension of a 2^n division is clearly evident. Figure 3 shows the harnessing of two 4-way dividers to form the 8-way divider. Shown, to, for the first time, is the placement of the elementary amplifiers.

There is a key shown in Fig. 2 which is the algorithm for patterning the phase shifter arrays which introduce the relative 0° and 180° phase additions. Given any 2^n array of phase shifts, the pattern for 2^{n+1} is given by forming the complement array and adjoining the latter to the former. For example, if the 8-way array from top to bottom is $(+--+-++-)$, where $+$ and $-$ have substituted for 0° and 180°, respectively, the complement pattern is $(-++-+--+)$ so that the 16-way pattern is $(+--+-++-;\ -++-+--+)$, the 32-way is $(+--+-++--++-+--+;\ -++-+--++--+-++-)$, and so forth.

Why quadrature coupler arrays instead of more conventional in-phase divider arrays? Why introduce the complexity in phasing the array? The essential answer is inherent in the simplest of the structures shown in Fig. 1. The coupler has a 3 dB split with energy incident upon each of the amplifiers having 90° difference. Under

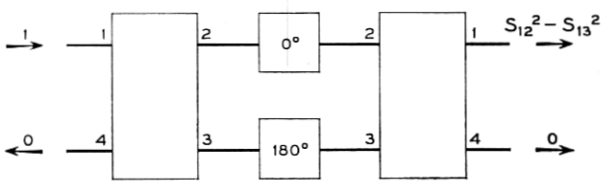

Fig. 1 — Quadrature couplers in all-pass arrangement.

reflection from identical amplifier inputs the phase difference mounts to 180°, as returned to the source, and a high degree of the reflection is cancelled. On the other hand, the reflected energies return in phase to the port conjugate to the input and are dissipated in a dummy load.

This technique, using 3-dB quadrature couplers to provide high input return loss to arrays of mismatched elements, was first used for short-slot couplers in radar transmit-receiver switches to prevent

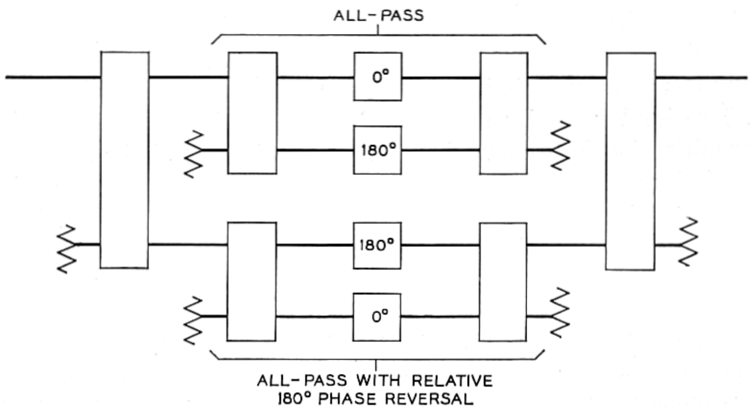

Fig. 2 — All-pass four way quadrature power divider.

mismatch back to the magnetron. While this technique has been extended to tunnel diode and transistor amplifiers, nevertheless the specific arrays shown here differ in two basic respects:

(*i*) The arrays can have arbitrarily large multiplicity.

(*ii*) All-pass transmission results from the use of 180° phase difference sections.*

The arrays shown in Figs. 1 through 3 were developed through a graphical induction process. Their properties are, however, more clearly phrased by returning to an analytic description. In terms of the scattering coefficients, the transmission T and input reflection K of the array are

$$T = (S_{12}^2 - S_{13}^2)^n t \tag{6}$$

* The use of 180° phase difference sections and their application to high multiplicity dividers was independently conceived. Original recitation of their use in two section dividers was made by E. A. Marcatilli, and D. H. Ring, IEEE Professional Groups Microwave Theory & Techniques *MTT-10* #4 (July 1962) pp. 251–8.

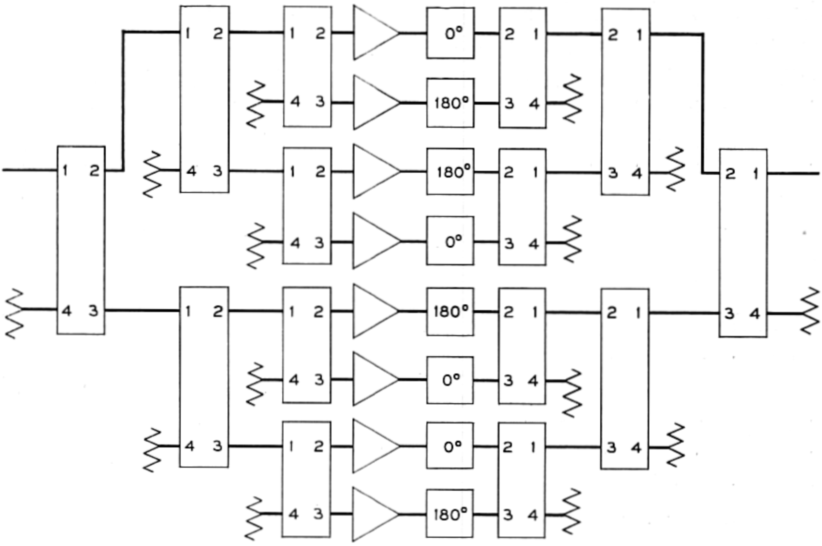

Fig. 3 — Eight way quadrature divider with amplifiers.

and

$$K = (S_{12}^2 + S_{13}^2)^n k \tag{7}$$

where t and k are the transmission and reflection, respectively, of each amplifier element and where 2^n is the number of amplifier elements. We observe from (5) and (6) that $|T| = |t|$, confirming the all-pass construction.

At the center frequency of each coupler, $|S_{12}| = |S_{13}| = 1/(2)^{\frac{1}{2}}$ and, via (2), there always exists phase quadrature. Therefore, $S_{12}^2 + S_{13}^2$ vanishes at centerband and, whatever the multiplicity of that zero, (7) shows the reflection zero to be raised to nth degree. Simple coupler synthesis techniques permit an arbitrary multiplicity of the degree of the zero and no major difficulty exists in obtaining octave performance and much greater.

2.3 *Coupler Economics*

It is instructive to consider the simplest couplers meeting the objectives of the amplifier to be constructed. The coupler enters into the technical design of the amplifier with respect to the return loss of the coupled amplifier array over the band together with a deterioration of the dynamic range relative to a single amplifier element. On the other hand, the choice of coupler enters into the economics as a greater cost for a more complex coupler.

We shall find that the return loss is not the essential problem of the array but that it is the dynamic range which is limiting. While dynamic range deterioration does dominate the coupler choice, we shall see that the degree of deterioration is not major and that the feed-forward compensation may, in many instances, make the problem negligible. Let us now investigate both the return loss and the dynamic range problems.

2.3.1 *Return Loss*

We shall consider two coupler types; one and two stage couplers. Corresponding to the number of stages is the multiplicity of zeros in the reflection factor, k, of (7) over the band. We intend to discuss these couplers in detail in another publication[5] and we shall consider it sufficient here to state simply that the couplers are compact and are composed primarily of lumped elements. They meet the characterizations of Section 2 to a high degree over extremely wide bandwidths and all the properties described are met substantially in practice.

We shall be concerned here only with the magnitude of k. Equation (7) yields

$$| K | = | \, | S_{12} |^2 - | S_{13} |^2 \, |^n \, | k |, \tag{8}$$

which produces the return loss L in db;

$$L = 20n \log_{10} | \, | S_{12} |^2 - | S_{13} |^2 \, | + 20 \log_{10} | k |. \tag{9}$$

Recognizing that $|S_{12}|^2 + |S_{13}|^3 = 1$, from (5), equation (9) yields

$$L = 20n \log_{10} | \, 2 \, | S_{12} |^2 - 1 \, | + 20 \log_{10} | k |. \tag{10}$$

Let us choose a normalized frequency variable, Ω, such that $\Omega = 1$ at center band. We then assert for one and two stage couplers, respectively, that

$$| S_{12} |^2 = 1/1 + \Omega^2 \quad \text{(single stage)} \tag{11}$$

and

$$| S_{12} |^2 = 1/1 + [1.545\Omega \cos (20.6°\Omega) - 1.195\Omega^2 \sin (20.6°\Omega)]^2 \quad \text{(double stage)} \tag{12}$$

where the coefficients in the two stage unit are chosen such that the power transmitted to port 2 and the power transmitted to port 3 differ by the same amount at centerband and at the edges of a 1.5:1 bandwidth. The behavior of (12) is shown in Fig. 4.

For a specific choice of a 1.5:1 bandwidth, as was required for the 60–90 Mc amplifier, we find

$$2 \mid S_{12}^2 \mid - 1 \leqq 0.22 \qquad \text{(single stage)}$$
$$\leqq 0.038 \qquad \text{(double stage).}$$

If we assume that our manner of use of the amplifier elements as emitter followers produces $|k|$ generally close to unity, we may neglect its inclusion in (10) and

$$L = -13.16n \text{ dB} \quad \text{(single stage)} \tag{13}$$
$$= -28.4n \text{ dB} \quad \text{(double stage).} \tag{14}$$

The major need for high return loss resides in the interaction between amplifiers. The dominant error produced by multiple reflections is the "third time around" incidence which is down from the initially incident signal by the sum of the return losses of the output and input of the first and second amplifiers, respectively. Since the gain per stage exceeds 6 dB for emitter follower amplifier elements, each stage is capable of feeding a next stage composed of four times as many amplifier elements. Therefore, if the first stage contains 2^n elements the second contains 2^{n+2}. If $L_{1+2} \equiv L_1 + L_2$ in dB. Then, from (13) and (14)

$$L_{1+2} = -26.32(n + 1) \text{ dB} \quad \text{(single stage couplers)} \tag{15}$$
$$= -56.8(n + 1) \text{ dB} \quad \text{(double stage couplers).} \tag{16}$$

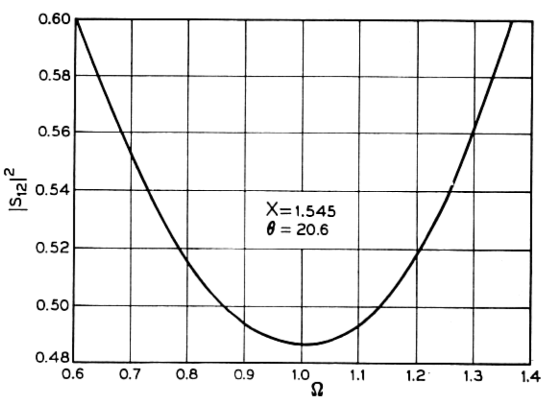

Fig. 4 — Two stage coupler designed for 1.5:1 bandwidth.

The fine structure in transmission ofr $n = 1$ or greater is very small and represents extremely small power variation. This presents a negligible taxation of the power handling capabilities of the feed-forward compensation system which will wash it out. Under these circumstances, the choice between single and double stage couplers does not rest upon return loss considerations.

2.3.2. *Saturation Power*

We may find from (6) a means for characterizing the energy paths through the coupler array. In much the same way as that in which the Pascal triangle builds up the binomial expansion in terms of the number of paths connecting two points, we may use the expansion of (6) to enumerate the number of paths from input to output. As an example, consider $n = 3$ which corresponds to an eight element amplifier stage. We have

$$T = \{S_{12}^6 - 3S_{13}^4S_{13}^2 + 3S_{12}^2S_{13}^4 - S_{13}^6\}. \tag{17}$$

As may confirmed using Fig. 3 as a guide, we have

(*i*) One path making six transitions of an S_{12} form
(*ii*) Three paths making four transitions of an S_{12} form and two of an S_{13} form, together with a net 180° phase shift
(*iii*) Three paths making two transitions of an S_{12} form and four of an S_{13} form
(*iv*) One path making six transitions of an S_{13} form together with a net 180° phase shift.

As is evident from this enumeration, different paths contain somewhat different amplitude content because of the frequency sensitivities of the scattering parameters S_{12} and S_{13}. This suggests that the amplifiers are somewhat unevenly excited which must have deleterious consequences to the net dynamic range of the composite amplifier.

Equation (17) contains too much information in that it does not distinguish between paths prior to amplifier excitation, which would bear on saturation, and paths leading from the amplifiers which have only to do with collection, which do not bear on saturation. This dichotomy into exciting and collecting paths may be accomplished simply by factoring (6) into

$$T = (S_{12} + S_{13})^n(S_{12} - S_{13})^n t.$$

We then define two structure factors; an excitation function F_e and a collection function F_c. We have

$$F_e = (S_{12} + S_{13})^n \tag{18}$$

$$F_c = (S_{12} - S_{13})^n. \tag{19}$$

If in Fig. 3, the 180° phase structures had been placed prior to the individual amplifiers, F_e and F_c would have reversed in formulation.

From the point of view of saturation we are only concerned with the form of the structure factor F_e. Again considering an eight amplifier array which corresponds to $n = 3$, we have

$$F_e = S_{12}^3 + 3S_{12}^2S_{13} + 3S_{12}S_{13}^2 + S_{13}^3 \tag{20}$$

which provides the following excitation paths

(*i*) One threefold transition S_{12}
(*ii*) Three twofold S_{12} transitions with one S_{13} transition
(*iii*) Three single S_{12} transitions with a twofold S_{13} transition
(*iv*) One threefold transition S_{13}.

To form relative magnitudes, take the relative phase of S_{12} to be 0° and that of S_{13} to be 90°. We have the following values at the lower band-edge in a 1.5:1 frequency range.

	Single-Stage Coupler	Douple-Stage Coupler
S_{12}	0.78	0.720
S_{13}	i 0.625	i 0.694

In Table I are tabulated the terms of (20) where the quantity

$$\frac{1}{(2^{\frac{3}{2}})} = 0.354$$

would represent the equal term values if the coupler were an ideal 3 dB power splitter. It is evident that the amplitude variations shown in the F_e column for the single stage coupler may be quite significant and, while they are more modest for the double stage couplers, they are, nevertheless, still present. Let us propose a harsh model of saturation in which any one amplifier saturates completely at the 0.354 level. In the SAT column of Table I are listed the relative amplifier outputs after the application of saturation with the assumption of no substantial phase shift effects.

TABLE I — EXCITATION AND OUTPUT TERM VALUES FOR SATURATION MODEL

Term	Multiplicity	Value for Single Stage Coupler		Value for Double Stage Coupler	
		F_e	SAT	F_e	SAT
S_{12}^3	1	0.474	0.354	0.386	0.354
$S_{12}^2S_{13}$	3	$i\,0.381$	$i\,0.354$	$i\,0.3585$	$i\,0.354$
$S_{12}S_{13}^2$	3	0.3045	0.3045	0.347	0.347
S_{13}^3	1	$-i\,0.244$	$-i\,0.244$	$-i\,0.334$	$-i\,0.334$
$\frac{1}{2^{3/2}}$		0.354		0.354	

If we recognize that the individual amplifiers are symmetrically disposed between input and output, as shown typically in Fig. 3, then the term value corresponding to the transmission from the input terminal to the input of any one amplifier corresponds, within a minus sign, to the term value for the transmission from that amplifier to the output terminal. Therefore, to form the transmission magnitude under this model of saturation, we sum the magnitudes, respectively, of the product of multiplicity, the term value, and the value listed in the saturation column, corresponding to each of the term values. Thus, using the single couplers we have

$$\begin{aligned} T &= t\{0.474 \times 0.354 + 3 \times 0.381 \times 0.354 + 3 \times (0.3045)^2 + (0.244)^2\} \\ &= 0.9095t \quad \text{(single stage coupler, 8 element amplifier)}, \end{aligned} \tag{21}$$

while far the double stage couplers we have

$$\begin{aligned} T &= t\{0.386 \times 0.354 + 3 \times 0.358 \times 0.354 + 3 \times (0.347)^2 + (0.334)^4\} \\ &= 0.9894t \quad \text{(double stage coupler, 8 element amplifier)}. \end{aligned} \tag{22}$$

Corresponding to the single stage, eight element amplifier there is a 0.88 dB distortion at the lower band edge while, corresponding to the double stage coupler amplifier there is a 0.08 dB distortion, where these distortions occur relative to an amplifier output at eight times the saturation power of an individual elementary amplifier. The implications of these distortions must be further elaborated, however, in relating the choice of couplers on the economics of the amplifier.

The intent in design, from the beginning, was the use of a feed-forward compensation system able to "clean up" amplitude distortions of 1 dB at maximum power, namely, the use of a clean-up amplifier having about one-fourth the output power of the main amplifier. While the 0.88 dB, corresponding to the single stage coupler, is within the 1 dB clean-up capability, it does form a reasonably large proportion of it. For many applications, however, amplifier power covers the entire spectrum of the bandwidth and the proportion of the power associated with the band edge is very much less than full power. Under these circumstances, the clean-up capabilities are quite adequate. If it is required to provide single frequency performance anywhere within the band, then about 10 per cent power output reduction is advisable.

The use of double stage couplers forms more of a fail-safe situation and the feed-forward amplifier is barely exercised. This leaves the feed-forward amplifier with its capabilities intact to handle phase and gain distortions without the excess power demand introduced by premature saturation within the main amplifier. The cost, of course, is in the use of couplers having roughly double the expense.

A further basis of choice of the double coupler lies in the use of the amplifier without the feed-forward implementation. According to the tabulation for the eight element amplifiers the term with the highest degree of saturation is S_{12}^{3} which has an unsaturated value of 0.386 for the two-stage coupler as opposed to the saturation level of 0.354, corresponding to a 0.72 dB saturation. Without feed-forward, this loss of capacity of the full amplifier, namely 0.72 dB, would be required to avoid *any* of the aspects of saturation. The corresponding penalty paid using the single stage coupler is 2.52 dB, which might be too high.

2.4 *Hybrid Power Divider Arrays*

We have discussed power divider arrays using only quadrature hybrids. It might be argued that it is required, at most, to be concerned only with paired amplifiers since this is all that is required to form adequate in-band match. After having performed such a match then an arbitrary number of such pairs might be combined using in-phase, as opposed to quadrature-phase, power dividers.

There are several objections. We list and discuss them in order.

Stability: Quadrature couplers have excellent match characteristics from dc to, typically, greater than three times their center frequency of operation. Since match and isolation are directly related, the amplifier elements of a large quadrature coupler array stage generally do not interact over the active range of these elements, assuring stability. The control of parasitics in an in-phase divider, particularly in a high frequency ferrite core type of divider, is not of the same degree of excellence and interactions leading to oscillation are a more severe problem.

Match: Assume that the array were composed of amplifier element pairs coupled via quadrature couplers, with the remainder of the array being in-phase dividers. The return loss from (13) and (14) would be 13.16 dB at the lower portion of a 1.5:1 band for a single stage coupler and 28.4 dB for a double stage, assuming the amplifier to be a perfect mismatch. The economic advantage of the single stage coupler is so major that, assuming an insistence on in-phase dividers, and all other considerations aside, a minimum single-stage quadrature coupling of four elements would be most desirable. This would lead to a 26.32 dB return loss, which is generally acceptable.

Cost: By far the most major advantage of the lumped quadrature coupler is its extremely low cost, particularly as a single stage device. It is reasonable to anticipate a cost of a few cents, as opposed to several dollars for other types of hybrids. In the vhf range an in-phase divider is composed of a pair, or more, of multiply-wound ferrite cores (Ruthroff transformers).[6] Microwave distributed couplers are vastly more complex. A hybrid array of in-phase and lumped quadrature couplers is far more costly than an array of quadrature couplers only.

Power Handling: Amplifier arrays under development are yielding typical average powers between 50 and 100 watts. It takes no great leap of imagination to conjecture such amplifiers in the kilowatt to multikilowatt range for any of assorted operations. Should any severe imbalances occur within the array owing to the failure of a transistor, an in-phase divider in the high power portion of the amplifier would experience large internally circulating power. If the divider contains high frequency ferrite transformers, burnout is virtually assured because of ferrite heating. Lumped quadrature couplers are air-core devices capable of tolerating kilowatt power levels without burnout; the only limitation is the dielectric strength of the insulation, which can be great.

2.5 *A Two-Rail Array*

While the authors believe the exclusively lumped quadrature coupler array to have prominent advantages over other forms of organization, it is of interest, nevertheless, to show one specific situation in which a hybrid system of in-phase and quadrature couplers has certain advantages in construction.

Consider a situation in which ferrite core implementations are not realistic to provide the 180° phase shifts required for all-pass operation. Such a situation occurs in the upper portion of the uhf region, and beyond, where the ferrite magnetization can no longer follow a rotation mode. Both in-phase dividers and differential 180° sections become much more expensive and cumbersome, but let us assume that the expenditure is warranted in providing two each of these aforementioned devices.

Figure 5 uses the in-phase dividers in a fashion reminiscent of two rail logic. Two universes are formed 180° apart via the in-phase dividers and the differential 180° sections. Whenever a 180° phase shift is called for it is accomplished simply by a symmetric interchange or transposition between the two universes. Thus, the cost of a large multiplicity of 180° sections is reduced to that of providing only two, aside from the two in-phase dividers.

Another advantage is gained in respect to saturation. Figure 5 is, in actuality, a doubled four-way quadrature divider to form an eight element amplifier and, as such, has a lower degree of compression

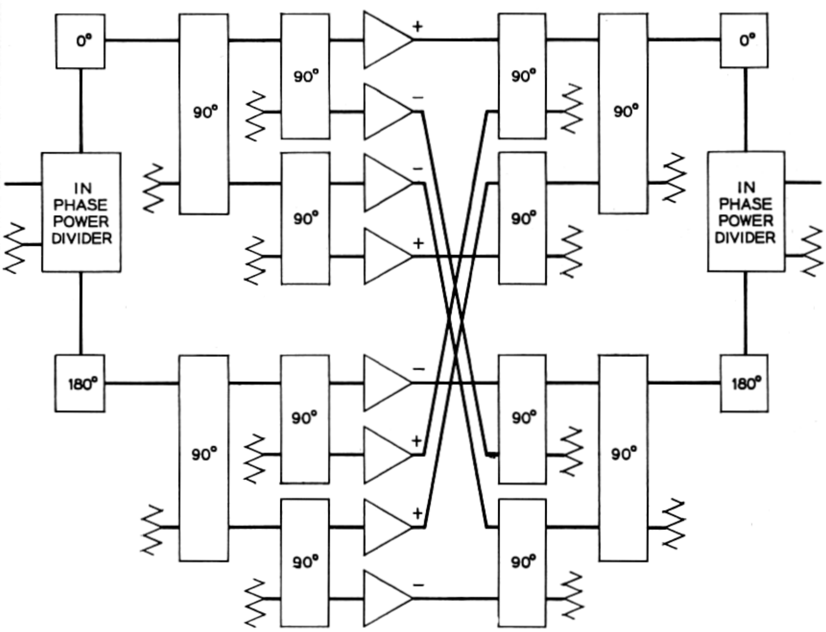

Fig. 5 — Hybrid array of in-phase and quadrature couplers: doubled for way divider to form eight element amplifier.

than does a pure quadrature coupler array. Corresponding to (18) we have for $n = 2$

$$F_e = (S_{12} + S_{13})^2 = S_{12}^2 + 2S_{12}S_{13} + S_{13}^2. \tag{23}$$

Let us now consider only the single stage coupler for which we have tabulation shown in Table II. Under the model of saturation proposed earlier we have

$$\begin{aligned} T &= t\{0.5 + 2 \times (0.487)^2 + (0.397)^2\} \\ &= t\{0.9362\}. \end{aligned} \tag{24}$$

The transmission value of $0.9362t$, or 0.48 dB, in the double-four amplifiers compares to the 0.88 dB compression relative to capacity of the full eight amplifier array. There is less demand on the feed-forward cleanup and power reduction is, therefore, unnecessary. Single stage couplers, therefore, produce acceptable performance and the resulting economy offsets the in-phase divider costs.

Having discussed the limitations of hybrid arrays, is Fig. 5 a workable system? The two major questions one may pose are those of power and stability. The experimental sections of this paper disclose that such a structure was successfully constructed for the 75 Mc amplifier. The ferrite loaded in-phase dividers were adequately broad, extending beyond the range of significant transistor activity, and stability was not the problem. However, the means to provide power handling capability within the ferrite transformers was a problem.

Assume that the amplifier were to be designed, not on a fail-safe basis, but on the assumption that all stages operate appropriately. Since the in-phase power divider (Ruthroff hybrid) has a conjugate antisymmetric port, can this port be used to save the need for the 180° differential phase shifters? The answer is negative in that the antisymmetry port would require power transformation through the ferrite. Since the core must be small to avoid circuit parasitics, the

TABLE II—SINGLE STAGE COUPLER

Term	Multiplicity	Value F_e	SAT
S_{12}^2	1	0.608	0.5
$S_{12}S_{13}$	2	$i\,0.487$	$i\,0.487$
S_{13}^2	1	−0.397	−0.397

handling of large continuous power might cause burnout. The symmetry port barely excites the ferrite and its use is indicated.

The circuitry of a simple 1:1 transformer is much less complex than that of a Ruthroff hybrid and the 180° section admits construction as a heavy core ferrite reversing transformer. In the event that the power might yet be too high to sustain in the reversing transformer, simple nonferrite circuits are available which produce 180° phase differentials over a 3:1 bandwidth.

As stated, this system does not contain the reliability feature discussed earlier. Transistor failure produces severe imbalance in the array and ferrite excitation in the transformers would result, with possibly deleterious effects at high average power. The specific amplifier design using the transformers with which the authors were concerned, related to pulsed operation, and this failure possibility was not significant.

2.6 *Out-of-Band Filters*

We have, up to now, considered in-band performance in which the return losses of the various stages were adequate to suppress multiple reflection effects. Out of band, the power split of the couplers becomes highly unequal and the input reflection into a quadrature-coupled pair of amplifiers approaches the amplifier reflection.

A simple fix for this unstable multiple interactions situation is obtained through the use of band-pass multiplex filters. In band, these filters provide excellent transmissions between various stages of the filter; out of band, all energy is transmitted reflectionlessly into dummy loads, with high isolation between the previously well coupled ports.

The multiplex, or out-of-band filters, are themselves quadrature couplers having a desired synthesis of $|S_{12}|$. The design through quadrature coupler techniques[5] is made to assure the wideband match required for stability. In the construction of the amplifier system, the use of these filters was determined on a practical basis. If stages directly feeding one another tended to oscillate, an out-of-band filter was injected between them and a large margin of stability was restored in each case.

2.7 *Comparison with Distributed Amplifiers*

The distributed amplifier is a chain of active elements in contrast to the parallel-like arrangement of the binary complex tree. Since the excitation varies along the chain one must reduce the gain of each successive stage to preserve equal power output per stage. The implicit assumption here is that each amplifier stage has been designed for power output to the limit of its intrinsic power capability. The gains of the various stages in sequence are clearly interdependent.

There are four major advantages of the coupler array over the distributed chain:

(*i*) The gain profile along the chain in a distributed amplifier is tailored to accommodate to a presumed performance of each stage, and the net energizing of any one stage in the chain is a function of all those preceding it. Fluctuations of performance in time, or simple tolerance effects, are cumulative. This is not true for the coupler array in which each element is orthogonally energized.

(*ii*) Imperfections along a chain produce multiple internal scattering with a necessary result of rapid gain variation with frequency. In a signal flow sense, each active element of the coupler array is "hit" only once with substantially no multiple reflection.

(*iii*) In the event of the failure of a single element of a distributed amplifier chain, the transmission properties are greatly modified, producing no fail-safe guarantees. Through the orthogonality properties of the coupler array construction, the failure of any one active element component simply causes its ordinarily contributed voltage to be subtracted from the common mode without any catastrophic interactions with the other properly functioning components. Thus, the coupler array provides a fail-safe redundancy.

(*iv*) The return loss properties of a quadrature coupled array improves with complexity. The return loss of a chain, contrariwise, deteriorates with complexity.

The intrinsic return loss property of a quadrature coupled array improves with increasing complexity through the presence of an ever greater number of parasitic-mode absorbers. A chain composed of imperfectly isolated stages becomes ever more difficult to match with an increasing number of stages.

2.8 *Dynamic Range*

It is evident that coupler distributed systems provide increased power in proportion to the number of active elements. Notice that this increased power output comes with little expense to the system noise temperature over that of any one of the active elements, aside from the copper loss of the couplers.

The noise emission of any one amplifier is spontaneous and, as such, is entirely uncorrelated with that of any other active element. The coupler system is a $2^n = N$ mode system and, with this lack of internal correlation, each active element as a δ function source equally stimulates each mode. Of the N modes, $N - 1$ are absorbed in the dummy loads, and only $1/N$ of this power finds egress. With N individual sources, the output noise emission is $N \times 1/N$ of the noise power of any one element, and so, is just the noise emission of one of the elements.

The dynamic range of the system is given by the ratio of the power at saturation to the noise power emitted over the amplifier bandwidth. Since the upper limit of the power has increased by a factor of N, and the noise level is unaffected, the dynamic range also increase by a factor of N. It should be recognized that the dummy loads of the coupler system occupy only ports orthogonal to the power transmission path, hence, none of these loads contributes to noise. It is concluded, therefore, that the coupler organized power multiplex increases dynamic range with insignificant noise figure deterioration.

III. EMITTER FOLLOWER AMPLIFIERS

3.1 *General Considerations*

The desire to deliver large power requires, in turn, a choice of an active element compatible with that power, together with a capability of supplying the circuit demands for proper frequency behavior. High power transistors always exhibit large internal capacitances and are limited in bandwidth.

The major problems arising in a general design that uses power transistors are the circuit complexities to realize the desired frequency characteristics, and the reproducibility from device to device for large variations of the internal parameters. The resolution of these problems may be carried out effectively by means of the emitter-follower amplifier.

The emitter follower, the analogue to the cathode follower, is generally used as a buffering device in transforming, at substantially a unit voltage gain, between a high impedance source and a low impedance load. While it is well recognized that power gain exists, a means for obtaining voltage gain between equal impedances has been viewed as demanding too complex a use of transformers in respect to parasitic elements. The eminent desirability of the emitter follower comes, of course, from its large self-degeneration which makes its response fairly impervious to internal parameter variations.

3.2 *Marriage of Emitter Followers and Quadrature Couplers*

Let us review some pertinent features of the ideal emitter follower, where the connotation of ideal implies an infinite input impedance, a unit voltage gain, and very low output impedance. Figure 6 shows the insertion of an ideal emitter follower between a normalized Thevenin generator and an equal impedance load. Since the open circuit voltage is twice that delivered to a matched load, the generator is characterized by a voltage of amplitude 2. The insertion of the emitter follower, therefore, doubles the voltage ordinarily delivered by the generator to the load, and a 6 dB gain results.

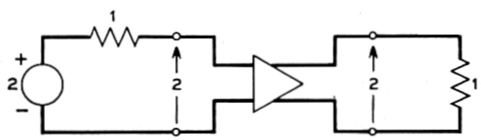

Fig. 6 — Thevenin equivalent of insertion of ideal emitter follower.

This gain, however, does not lend itself to a simple transformerless cascade of emitter followers as seen in Fig. 7.* It is evident that having achieved a voltage level of 2, this amplitude is transferred from section to section without further enhancement. The fallacy in forming this simplistic cascade is that match has not been restored from stage to stage, remembering that it was precisely through the interface of a matched source and open circuit that the 6 dB gain was achieved.

This situation is rectified by the use of quadrature couplers to pair emitter followers as shown in Fig. 8. The harnessing structure of the couplers forms an all-pass unity transmission network and the gain takes place via the mechanism of Fig. 6. Since the couplers do restore match, substantially, at the input as well as the output, this circuit is capable of iteration with return losses as given by (7). Specifically, incident waves of amplitude S_{12} and S_{13} impinge, respectively, on each of the emitter followers. Since each emitter follower is an open circuit, a voltage reflection factor of unity doubles each of these incident

* The lack of a transformer distinguishes this cascade from a Darlington network in which cumulative power gain occurs only because of impedance transformation.

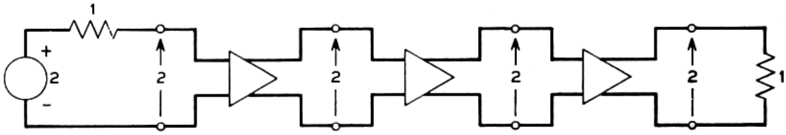

Fig. 7 — Transformerless cascade of emitter followers.

voltages, forming emergent waves of $2S_{12}$ and $2S_{13}$, respectively. The termination is excited by an amount $2(S_{12}^2 - S_{13}^2)$ which, from (5), demonstrates a 6 dB gain.

Since the circuit of Fig. 8 shows an inheent 6 dB gain, no transformers are required to achieve this modest, but highly useful, gain. However, a mild input or output voltage transformation might be used to raise it somewhat beyond this value. From the point of view of power generation, the stages shown in Figs. 1 though 3 might increase at a four-fold rate. Such proliferation of circuit complexity seems not nearly so modest.

3.3 *Real vs. Ideal Emitter Followers*

The ideal emitter follower is a most imperfect representation of the real device. However, it is not too difficult to patch things up to obtain a creditable performance within the frequency range of our interest. Figure 9 is a simplified high frequency equivalent from which we may deduce the more significant features of operation.

Define the following quantities:

$$Y = 1 + i\omega C_l = 1 + i\frac{\omega}{\omega_l}$$

$$r_b C_{cb} = \frac{1}{\omega_{cb}}$$

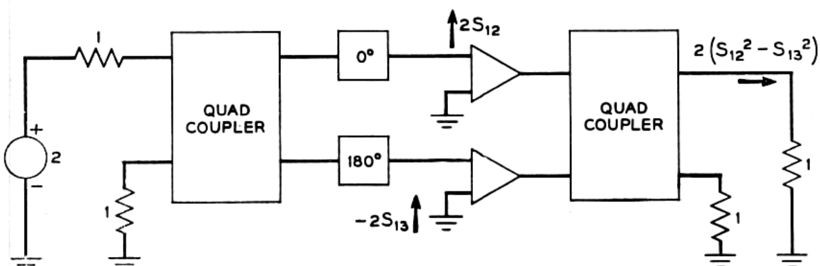

Fig. 8 — Transformerless pairing of emitter followers through quadrature couplers.

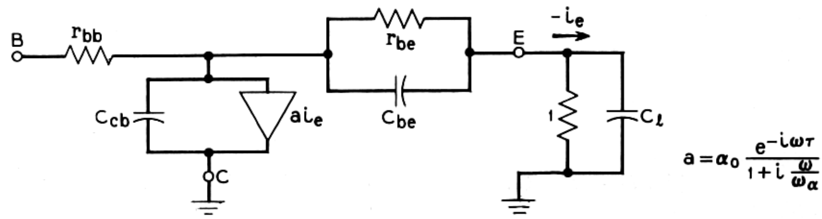

Fig. 9 — Simplified high frequency equivalent of terminated emitter follower.

$$E_1 = \text{Base signal voltage}$$

$$E_2 = \text{Output voltage at the emitter}$$

$$a = \frac{\alpha_0 \epsilon^{-i\omega\tau}}{1 + i\dfrac{\omega}{\omega_\alpha}}$$

τ = emitter to base drift time and is of order $0.6/\omega_\alpha$. Analysis yields the results

$$E_1 = \left\{1 + \frac{i\omega}{\omega_{cb}} + r_{bb}(1 - a)Y\right\} E_2 \tag{25}$$

$$Z_1 = \frac{E_1}{I_1} = \frac{\left(1 + \dfrac{i\omega}{\omega_{cb}}\right) + r_{bb}Y(1 - a)}{i\omega C_{cb} + Y(1 - a)}. \tag{26}$$

Since we are seeking nominally small gain we may, with good approximation, take $\alpha_0 = 1$. Expanding, and dropping terms beyond ω^2, (25) becomes

$$E_1 = \left\{1 + r_{bb}\left[1 - \cos\omega\tau - \frac{\omega^2}{\omega_l}\left(\frac{1}{\omega_\alpha} + \tau\right)\right] + i\omega\left[\frac{1}{\omega_{cb}} + r_{bb}\left(\frac{1}{\omega_2} + \tau\right)\right]\right\} E_2.$$

The quantity r_{bb} is small relative to unit loading and we may omit the term $r_{bb}(1/\omega_\alpha + \tau)$ in contrast to $1/\omega_{cb}$. We have then

$$E_1 = E_2\left\{\left[1 - \omega^2\left(\frac{1}{\omega_\alpha\omega_l} + \frac{\tau}{\omega_l} - \frac{\tau^2}{2}\right)\right] + \frac{i\omega}{\omega_{cb}}\right\}. \tag{27}$$

We perceive in (27) a negative real part resulting from the Miller effect action with respect to C_l. The tuning procedure demands flat power gain, and we may offset the decrease of the real part in (27) against the rising imaginary portion to maintain constant magnitude. Assuming small parasitics at the frequency of operation, the constant magnitude condition yields

$$\frac{1}{\omega_{cb}^2} = 2r_{bb}\left(\frac{1}{\omega_2\omega_l} + \frac{\tau}{\omega_l} - \frac{\tau^2}{2}\right). \tag{28}$$

A value of ω_l always exists to satisfy (28).

According to (27) and (28), the voltage appearing at the base emerges with unit gain from the emitter. The question, indeed, is what voltage appears at the base? To determine it, we examine the input impedance function given in (26) to within first order terms in ω. We find

$$Z_1 = r_{bb} + \frac{1}{i\omega\left(C_{cb} + \dfrac{1}{\omega_\alpha} + \tau\right)}. \tag{29}$$

While (29) has a disconcerting form with respect to dimensions, as does (28), that a unit terminating impedance is dimensionally concealed in these results. Under these circumstances, r_{bb} is small compared with unity and we momentarily neglect it in comparison with the series capacitance $C_{cb} + 1/\omega_\alpha + \tau$.

Assuming this capacity not to be excessively large, it may be incorporated as a terminating element of a bandpass filter that has a characteristic impedance equal to that of the generator. Over its bandwidth the generator incidence is maintained to within the time delay of the filter circuit and all the appurtenances of the ideal emitter follower are realized; an incident signal is delivered at matched impedance from the generator to the open circuit, and the voltage at the input is transmitted to the output in substantially frequency-independent fashion.

In general, this bandwidth is far larger than the 1.5:1 frequency width desired and, corresponding to this gain-bandwidth availability, tradeoff may be made with the bandwidth to enhance the gain by using appropriate input circuit design. This circuit may also compensate for r_{bb} by providing high frequency peaking to offset it.

3.4 *Linearity*

While much has been said here of the linear behavior of the emitter follower, the feedback which causes the emitter to follow the base extends well beyond the region of linear characterization. In large signal excitation the interelectrode capacitances vary throughout the cycle, causing nonlinear interactions with the circuit environment. A particularly significant term is C_{cb}, but its variation is greatly damped by the generator impedance. Aside from the considerations of frequency characteristics, it is the desire to maintain high generator damping that vitiates a greater use of impedance transformation to yield higher gain.

In class A operation the region of linearity is limited, in any event, by current cutoff. In class AB push-pull operation the distortion takes the form of a slightly different low and high signal gain. This last distortion may well be the more insidious of the two and it is only the use of the feed-forward system which makes it acceptable. This acceptability is extremely significant because the push-pull class AB operation, when operated near the class B state, has several advantages over class A:

(*i*) Power supplies are substantially reduced.

(*ii*) Since all of the transistor current is used to produce RF power, and since twice the number of transistors are used, about four times as much power is generated for the same amount of coupler hardware.

(*iii*) Lower average currents yield wider band response because of lower average internal capacitance. In actual measurement the limit of the linear range seems closely identified with avalanche breakdown, for which there is the danger of thermal damage to the transistor.

There is a notable disadvantage in class AB operation, but it is not too objectionable in a 50 per cent or less duty cycle amplifier. As the excitation varies, the average current varies, shifting bias points. Further, the operating temperature may shift the quiescent point in each transistor, significantly affecting low signal distortion. However, an active clamp may be used which samples the off-signal voltages and restores proper operating levels. This clamp is described in Section 6.8.

3.5 *Economics of Coupler Organized Emitter Follower Amplifiers*

While it is implicit throughout the technical descriptions, one general feature which must be stressed is the significant economy possible through emitter followers. Because of its large internal negative feedback, reproduction of the transistor becomes a problem of relatively negligible proportions.

In contrast to the situation in which refined transistor devices may be improved in yield, we have a situation in which frankly cheap devices are used to provide an adequate yield. The low potential costs of quadrature couplers and relatively inexpensive transistors, operating well beyond their traditional ranges, forms the prospect of much less expensive, high performance amplifiers.

Acceptable within this framework may be a moderate aging of semiconductor parameters, usually a bugaboo in device manufacture. Its importance is diminished not only through the degeneration offered by the emitter follower circuit, but by the feed-forward provisions as well. Thus, cheaper encapsultations might be permitted than otherwise would be, admitting possibly consequential cost reductions in transistor fabrication.

IV. FEED-FORWARD

4.1 *Feedback vs. Feed-Forward*

Much of the development of the high power amplifier stands or falls on the ability to devise a proper error correcting system. As stated in the introduction, the choice made in the design of this system, if indeed a choice was open, was a "feed-forward" construction.

In the late 1920's, H. S. Black conceived of feed-forward as well as feedback, and his patent disclosures show either grid or plate injection of the error signal.[3] The grid meant back injection whereas the plate meant forward injection and either was adequate to within 180° of phase. Clearly, the plate injection was at a higher level and required subsidiary amplification, while grid injection ingeniously reused the same amplifier.

With an occasional, but short lived, resurrection, feed-forward fell into a general disuse until independent revivals by McMillan and Van Zelst.[1,2] These were followed by further activities by Deighton, Cooke-Yarborough, Miller, and Golembeski.[6,7] The essence of the McMillan and Van Zelst disclosures was to combine both techniques of Black by providing frequency shaping networks simultaneously in feedback as well as in feed-forward. The aim was to provide channel redundancy as well as parameter desensitizing. The Deighton paper uses this technique to stabilize a particular circuit of a pulse amplifier while Golembeski has represented a generalized parallel channel configuration and developed a synthesis procedure and realizable conditions.

Even though the basic inadequacy of feedback techniques were evident, we developed the feed-forward compensation system because we needed to provide stable correction to an amplifier that had a relatively large time delay. Indeed, simple physical insights not only repudiate the use of feedback but force use of the alternative, for our case.

Feedback, a technique which has been used with much success, attempts a causal contradiction: after an event has occurred, reshape its cause. This violation may be resolved only by time smearing the event to blur the distinction between "before" and "after" to an adequate degree. This smearing requires a prescribed control of spectral response; a craft too well known to presume further comment.

Feed-forward evolves from entirely different premises, avoiding causal anomaly.[8] Instead of seeking to reverse time, it recognizes the passage of time. Error is determined in relation to a time-shifted reference and corrected in a time sequence compatible with the main signal. No obscuration of time is permitted and frequency shaping to attain stability is irrelevant.

The fact that frequency shaping (that is, control of the return ratio function) is irrelevant is of major importance. A feedback amplifier consumes a good portion of its gain-bandwidth figure of merit in controlling the return ratio well beyond the band of interest to prevent instability.

The power transistors demanded for economical construction of the power amplifier clearly are not geared to high frequency application. To obtain the needed bandwidths, gains must be kept modest. The amplifier, then, would be built of a sufficiently large number of stages to offset the gain consumed in a feedback correction. With this number of stages, however, the excess phase would mount to such degree as to be beyond control in a feedback system.

The feed-forward system has three strikingly important features:

(*i*) Feed-forward correction does not reduce amplifier gain.

(*ii*) Gain-bandwidth is entirely consumed within the band of interest.

(*iii*) Feed-forward realization is independent of the magnitude or shape of the amplifier time delay and it exists in general.

We demonstrate these features in the following section. The major point to be made here, however, is that it is not a question of feedback vs. feed-forward as a choice, but as a necessary engineering decision.

4.2 *Detailed Description of Feed-Forward Operation*

Figure 10 is the general block diagram of the feed-forward correction system. Input signal is divided into two portions; that going into the amplifier and that entering a reference path. The signal is amplified through the main amplifier A and a sampling is taken through a directional coupler with very flat coupling 1/G, where G approximates the in-band gain of A. This sample is compared with the reference path signal which has been delayed a time τ, a quantity approximately equal to the time delay of the main amplifier.

A time-shifted error is thus obtained and it is amplified by a subsidiary amplifier whose gain G′ is such that this error is restored to the appropriate level to cancel the error of the main amplifier. If the subsidiary amplifier consumes a time roughly equal to τ', a time delay of that value is inserted in the main amplifier path. The injection of the error back into the main line takes place through a reactive two-port, shown typically as a transformer.

The choice of a reactive two-port, as opposed to the use of a coupler four-port, is made irrespective of the ordinary desire of keeping the outputs of both amplifiers decoupled. While a coupler would isolate the subsidiary amplifier from the main amplifier, the cost of energy losses of the subsidiary amplifier into the dummy loading of a coupler would be unacceptable.

While we have emphasized the care to be taken with time delay, there is an implicit condition within the phase specification to produce error cancellation. Since time delay is the slope of phase with respect to frequency, assuming small phase curvature, it is adequate to adjust phase at but a single point in the band and have it track throughout. In the event that the main amplifier and the subsidiary amplifier had large phase curvatures, then τ and τ', respectively, would require more complex realization than simple delay lines; they would have to incorporate the same curvatures as well. Evidently, this may be extended to yet higher phase derivatives with respect to frequency.

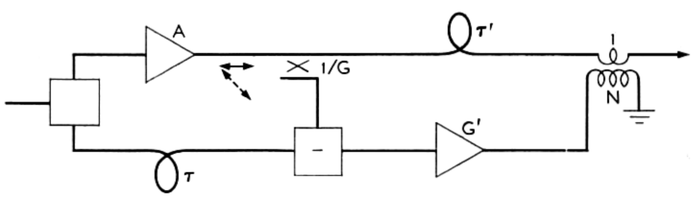

Fig. 10 — One loop feed-forward network.

As indicated, the gain and phase of A and the gain and phase of G′, are known only to within some first order tolerance. Via the coupler 1/G and the time delay τ, strict gain and phase standards are set with which to compare the amplifier A. These form a Procrustean bed such that anything within A that does not fit these standards is to be treated as error, and chopped off or stretched. In processing just this error portion through G′ and τ', first order errors of the latter two quantities only produce negligible second order quantities in error correction.

Having "feed-forward compensated" the amplifier A in this manner, it is quite proper to consider this circuit complex as a new amplifier, B, which is to be "feed-forward compensated." Since B has desired performance to within second order errors, this new compensation would reduce the errors to third order errors. This method of successive compensation may be iteratively compounded, controlling errors to a very high degree. A two-loop system is shown in Fig. 11.

4.3 *Transformer Design for Error Injection Circuit*

As already discussed, the use of hybrid couplers for error injection is avoided, since any such matched device leads to excessive power loss to the system. If, for example, the coupler provided equal power division, then a 3 dB power loss would be sustained by the main amplifier. If a division ratio more favorable to the main amplifier were used, the consequences to the subsidiary amplifier might well prove disastrous. It could be called on to deliver a greater power than the main amplifier, in some instances, to compensate moderate errors in that main amplifier.

We will try to design a feed-forward compensation system that can

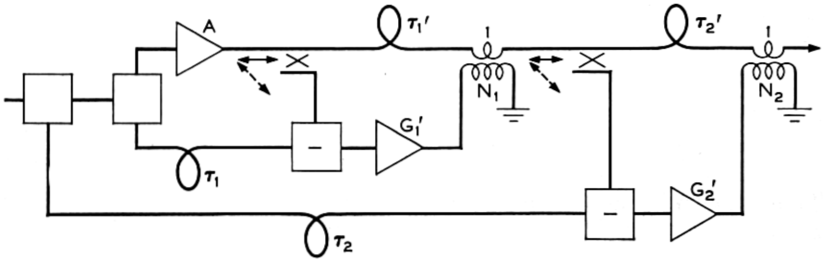

Fig. 11 — Two loop feed-forward network.

correct the main amplifier over a range of about 1 dB, using transformer injection.

Appendix 2 shows that the optimum transformer ratio is

$$N = \sqrt{\frac{1 - \delta_p}{\delta_p}}, \tag{30}$$

under the condition that the subsidiary amplifier sees a matched load when it is correcting the peak relative in-phase voltage error, δ_p. At this point the power being delivered by the subsidiary amplifier is

$$P_s = \frac{\delta_p}{1 - \delta_p} P_m, \tag{31}$$

where P_m is the main amplifier peak power. It is shown that the mismatch introduced by the transformer results in no loss to the power output of the system, only a small loss of gain, so there appears *none* of the loss that would be associated with a hybrid coupler instrumentation. In our case of 1 dB correction, $\delta_p = 1/8$, for which $N^2 = 7$.

The analysis shows another interesting feature. If, with no feed-forward circuitry present, a relative distortion occurs at peak power approximately equal to $2\delta_p$, then:

(*i*) In the presence of feed-forward circuitry the power gain is depressed by a relative amount δ_p, so that the power necessary to produce an undistorted output is only δ_p.

(*ii*) This residual distortion power of δ_p is contributed by the subsidiary amplifier.

(*iii*) In this sense, a subsidiary amplifier power capability of δ_p controls a distortion of $2\delta_p$.

The above design takes into account the in-phase errors, mainly in-phase compression in the amplifier at peak power. Another consideration is the ability of the above design to compensate for phase distortion. It is shown that, if the maximum relative quadrature voltage at peak main amplifier power is $i\delta_p'$, the relation between this and the maximum permissible in-phase voltage error is

$$\delta_p' = \tfrac{1}{2}\delta_p\,. \tag{32}$$

Then the same system which will correct a 1 dB amplitude error with $\delta_p = \frac{1}{8}$ will provide a phase correction of $\pm$ arctan $\delta_p' = \pm 3.6°$ at maximum main amplifier power.

The above might seem a small phase correction range but remember that the main amplifier consists of an array of emitter follower amplifiers. To within small variable transit time effects with high power, the emitter follower amplifier would tend to saturate as a function of instantaneous excitation. Hence, satuation effects would tend to be in phase and would not seriously tax the range of quadrature correction capability. Nevertheless, from the equalization point of view, it would be well to limit phase ripple peaks to below 2° before applying feed-forward correction.

We emphasize that the numerical results obtained above correspond to the worst case and should not be used inflexibly. For instance, there might be some best phase positioning of the subsidiary amplifier along its transmission path to optimize the phase correction range. Or suitable judgment might suggest an optimum trade-off between in-phase and quadrature error correction in a specific construction.

Appendix 2 also considers two questions concerning match within the system. First it shows that, even though the subsidiary amplifier is mismatched at less than maximum excitation, reflections within the error correction loop can never be phased so as to demand either more current or voltage from the subsidiary amplifier than is demanded at peak power. Second, it considers the output mismatch and the use of nonoptimal transformers.

The term nonoptimal is meant to suggest that the transformer is not optimally designed to minimize the power demand on the subsidiary amplifier. As shown, a non-optimal transformer yields the advantage of simplifying the problems in providing very high output return loss from the amplifier system, but at major cost to the subsidiary amplifier.

4.4 *Circuit Modification for Optimal Noise Performance*

As discussed, the dynamic range of a coupler organized amplifier is raised at the saturation power level in proportion to the number of active elements without deteriorating the noise level of the system. This promises a dynamic range increase of only a factor of N. It is possible to obtain much more spectacular ratios by a noise exchange with a low noise subsidiary amplifier. This improvement is of major significance since the active elements are individually high power devices emitting large noise powers and advantage is to be gained, in providing dynamic range, by working at the low end of that range.

We emphasize that the subsidiary amplifier in the feed-forward system has an entirely different function in noise reduction than does an amplifier used as a low noise preamplifier. In the latter use, the amplifier must sample the entire signal range and, intrinsically, limits dynamic range. The subsidiary amplifier, on the other hand, handles only the spontaneous error and, consequently, a negligible portion of the coherent signal range. Therefore, the dynamic range of the subsidiary amplifier may have virtually no relationship to the dynamic range of the system it controls.

4.4.1. *Major Considerations*

The feed-forward or subsidiary amplifier has a considerably lower dynamic range than does the main amplifier. With multiple feed-forward looping, as indicated in Section 4.2, the dynamic range of the ultimately controlling amplifier may be extremely small, yet adequate. The low power requirement of the subsidiary amplifier in such use may be turned to advantage by designing for relatively low noise figure performance.

The noise of the main amplifier is an erroneous output which the feed-forward system corrects. However, noise generated by this subsidiary amplifier is introduced in the process. The advantage of such an exchange would be lost if the signal-to-noise ratio of the subsidiary amplifier were to be lessened by the use of significant power division at the input to the reference path.

Figure 12 shows a modification of Fig. 10 to achieve optimal noise performance. Couplers C_1 and C_3 are made to direct most of their power to the subsidiary amplifier G′. While the gain of amplifier A must be made larger to offset the coupling loss of C_1, this increase is, in principle, irrelevant. The worsened noise figure of A is of no consequence since the error cancellation of the feed-forward system simply treats the additional noise as additional error.

4.4.2. *Nonideal Behavior*

Of course, what we have just stated is an ideal view; the cancellation is imperfect. The primary source of this imperfection lies in the varia-

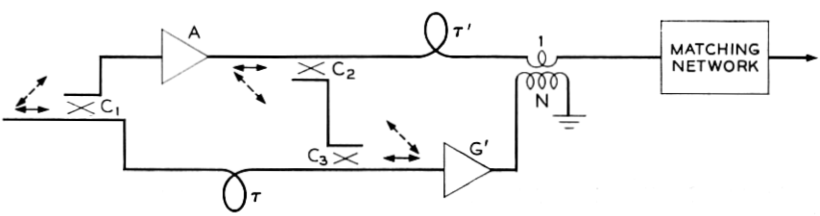

Fig. 12 — Modification of feed-forward network to improve noise performance.

tions of the subsidiary amplifier, which is uncontrolled. A realistic quantitative assumption might be a power transfer variation of ± 5 per cent over the frequency range or a peak voltage transfer variation of $2\frac{1}{2}$ percent. This latter number limits the cancellation to a power ratio of $(0.025)^2$ or 32 dB. Let us assume further that C_1 is a 12 dB coupler, increasing the excess noise of the main amplifier by that amount. Nevertheless, even with a reduction of the 32 dB cancellation possibility by 12 dB, the net noise of the feed-forward compensated system is the sum of the excess noise of the subsidiary amplifier plus a 20 dB reduced excess noise of the main amplifier. Since the subsidiary amplifier noise is presumed minor, relative to that of the main amplifier, this represents a major gain in performance. We have neglected the small transmission loss of C_1 which deteriorates the noise performance slightly.

In multilooping along the lines of Fig. 11, successive reductions of the above order may be accomplished until the noise of the lowest excess noise subsidiary amplifier stage dominates the process. Because of the rapidity of convergence it is questionable whether more than three subsidiary loops would ever be meaningful with respect to noise minimization.

4.4.3. *Subsidiary Amplifier Stabilization*

Since the limit of noise cancellation of the high power amplifier relates to gain variations of the subsidiary amplifier, we suggest another form of multilooping, differing from Fig. 9, in which the subsidiary amplifier itself is feed-forward stabilized. We believe that this variant, shown in Fig. 13 is inferior to that of Fig. 11 in that a much more restrictive tolerance is placed on couplers C_2 and C_3. This is so because the error determination is now only "one shot" with reference to the main amplifier as opposed to the progressive error sampling of the earlier system.

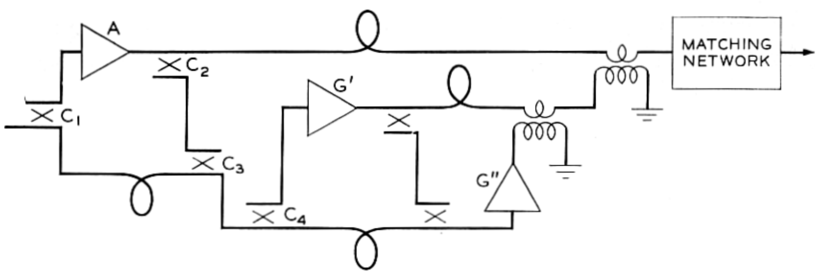

Fig. 13 — Modification of low noise system in which subsidiary amplifier is stabilized.

Nevertheless, the looping progression shown in Fig. 13 requires that some weak subsidiary amplifier only be confronted with a next stage merely one degree higher in power. In contrast, Fig. 11 requires that all subsidiary stages work into the highest power level of the system. That this might be a consequential difference is not yet clear.

4.5 *Subsidiary Amplifier Gain*

In the development to this point we have given the relationship between subsidiary amplifier gain and main amplifier gain in various implicit forms. Here we explicitly state this relationship in terms of the insertion power gain G_m of the main amplifier and G_s of the subsidiary amplifier.

Let us form a skeletal schematic of the levels of power flow in the full amplifier, which is shown in Fig. 14. We define the following quantities:

L_m—Loss through input power splitter to main amplifier
L_s—Loss through input power splitter to subsidiary amplifier
L_s'—Loss through the error coupler to subsidiary amplifier.

If we particularize the circuit between the subsidiary amplifier output and the load, $1 + 1/N^2$, we find the equivalent shown in Fig. 15, whereas the output of G_m to the load is shown in Fig. 16. The amplifiers G_m and G_s are each normalized to a unit impedance and, through Figs. 12 and 13, we define the transfer functions T_m and T_s to the load, $1 + 1/N^2$, which we determine to be

$$T_s = \frac{N}{1 + N^2} \tag{33}$$

and

$$T_m = \frac{N^2}{1 + N^2}. \tag{34}$$

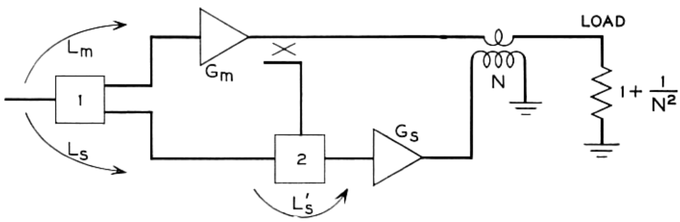

Fig. 14 — Feed forward amplifier schematic.

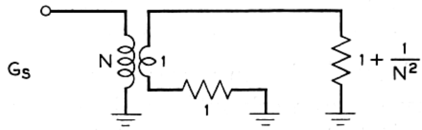

Fig. 15 — Equivalent circuit as seen by the subsidiary amplifier.

The nature of feed-forward correction is such that the overall transfer power from the input through the main amplifier path equals that through the subsidiary amplifier path. This provides

$$\frac{T_m^2}{L_m} G_m = \frac{T_s^2 G_s}{L_s L_s'} ,$$

which, through (33) and (34), yields

$$\frac{G_s}{G_m} = \frac{N^2 L_s L_s'}{L_m} . \tag{35}$$

Equation (35) determines the differential gain of the subsidiary amplifier over the main amplifier.

We may quantitatively evaluate the differential gain for a particular, but typical, low noise figure amplifier. For this amplifier we have

$$L_s = L_s' = 1.07 \rightarrow 0.3 \text{ dB},$$

$$L_m = 20 \rightarrow 13 \text{ dB},$$

and

$$N^2 = 9 \rightarrow 9.55 \text{ dB}.$$

The differential gain, in dB, of the subsidiary amplifier is, in this case,

$$9.55 + 0.3 + 0.3 - 13 = -2.85 \text{ dB}.$$

In this calculation we find, as a practical matter, that the subsidiary amplifier actually has a smaller gain than does the main amplifier in a low noise configuration.

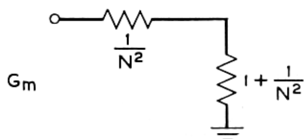

Fig. 16 — Equivalent circuit as seen by the main amplifier.

4.6 *Sampling Via Directional Coupling*

The causal arguments, which motivate the development of this form of amplifier, imply that all sampling be done in a strictly ordered time sense. This requires sampling devices which are well isolated from reflections within the system which appear as events in improper time sequence.

Figures 12 and 13 shows the use of directional couplers for sampling. Let us assume for the moment that the output of the main amplifier, A, is perfectly matched. Under the circumstances error injection by the subsidiary amplifier G into the final transformer has only backward component of wave travel through the coupler C_2 and no feedback occurs to that amplifier.

If the output port of A is mismatched, any reflections of the injected error signal create a new forward wave representing undesired signal. However, the undesired signal is sampled by C_2, treated as an error, and corrected via the correction loop.

Energy reflected back into the system from a mismatched load is similarly handled. Thus, in the steady state, the directional character of C_2, has acted to preserve the prescribed net emergent signal. To the extent that a coupler has finite directivity this ideal is blurred, but couplers with high directivities are available in practice and the idealization may be met to high degree.

Aside from its use to eliminate parasitic feedback paths, the sampling coupler stabilizes the main amplifier output impedance and provides for a high output return loss. As we have already discussed, the fixed amplifier system has an unusually high degree of linearity and, as such, is entirely describable by a Thevenin equivalent. Let us then, in gedanken fashion, measure its output impedance, which is a physically meaningful quantity.

To measure the system output impedance, a wave emanating from an external generator is made incident upon the output terminals, and its reflection measured. From Figs. 12 and 13 we can see there is a first reflection from the series impedance imaged through the $1:N$ transformer, and a second reflection from the main amplifier itself. However, as we have seen, this second reflection is negated via the coupling through C_2 back to the subsidiary amplifier.

By design, the effect of the transformer is small and any mismatch of the subsidiary amplifier is a thoroughly negligible second order effect. The net input to the output terminals therefore appears as a resistor of normalized value $1/N^2$ in series with a perfect absorber;

a total resistive impedance of $1 + 1/N^2$. Since $N^2 = 7$, the mismatch to unity is relatively small and a simple matching section may be used to produce a large return loss across the band of interest. Most importantly, this return loss relates only to the quality of passive circuits and is essentially independent of the active system.

A simple catechism helps us form an effective perspective in recognizing the role of directional coupler sampling in error correction in relation to the two other, better known, types.

(*i*) Voltage sensing produces an unvariable emergent voltage, making the source impedanceless.

(*ii*) Current sensing produces an unvariable emergent current, making the source admittanceless.

(*iii*) Incident wave sensing produces an unvariable emergent wave, making the source reflectionless.

Within this context, the use of directional couplers to provide incident wave sensing is not limited to only the feed-forward system described here to produce a highly matched generator.

4.7 *Parallel Literature on Feed-Forward Technique*

It seems clear that the presently described structure is closely related to the previously mentioned class of redundancy and desensitization networks.[1, 2, 7] Nevertheless, we feel that we emphasize certain important points in this paper that are contained, but not obvious in the general treatments:

(*i*) The significant role of matched delay is cited, from causal considerations, as the basic means of providing a degree of correction realization unavailable in feedback. The degree of correction is arbitrary.

(*ii*) The absence of stability considerations in feed-forward.

(*iii*) Realization of full gain-bandwidth, fully consumed within the band of operation, without restriction on the degree of correction.

(*iv*) The advantages of the use of unequal amplifiers with respect to noise figure.

(*v*) Use of coupler sampling both for stability and high output return loss.

(*vi*) Transformer injection of the error correction to minimize loss of power capacity.

V. SPECIAL COMPONENTS CONSIDERATIONS

It is intended to review the coupler art elsewhere; we are not concerned with it here. Indeed, the couplers that we used were commercially available items* and were not, themselves, the source of any general concerns.

There were, however, certain specific concerns in the experimental developments which did not materialize. While they did not manifest themselves for our particular equipment, they still might show up in other circumstances.

5.1 *Error Injection Transformer*

While not in exact correspondence with the specific values chosen for the mathematical model, a 3:1 transformer is close enough to the value needed and generally can be constructed over the necessary bandwidth range at the frequencies of interest. The major concerns were the practical one: can a ferrite loaded transformer sustain the power levels flowing through it?; and the hypothetical one: what if the frequency range were beyond the capabilities of ferrites?

In seeking other realizations of the transformer, one must appreciate that the transformer shown in Fig. 14 is little more than a schematic format of the circuit. Specifically, the transformer is a reactive reflection isolating the subsidiary and main amplifiers. Any other reactive two-port having the same insertion loss would, according to Weissfloch,[9] be equivalent to the transformer to within the insertion of phase corrections at each of the ports. Further, the error injection might take the form of a dual shunt current feed as opposed to the series voltage form and, as such, there are quite flexible options in design.

While there is hardly a dearth of networks to produce the desired circuit function, we shall develop one which seemed simply suited for synthesis. Figure 17(a) shows a tee of inductances. The power transmission $|T^2|$ vanishes at zero and at infinite frequency without any intervening singularity so that, as shown in Fig. 17(b), a stationary transmission value exists. If the circuit is modified as shown in Fig. 18(a), by introducing the shunt capacity C'', a high frequency propagating region is introduced as observed in Fig. 18(b). With appropriate proportioning of the capacitance an inflection region is introduced at the point of stationarity and a flat insertion loss region is obtained. This flat insertion loss may be corresponded to the trans-

* Merrimac R&D Co., West Caldwell, N.J.

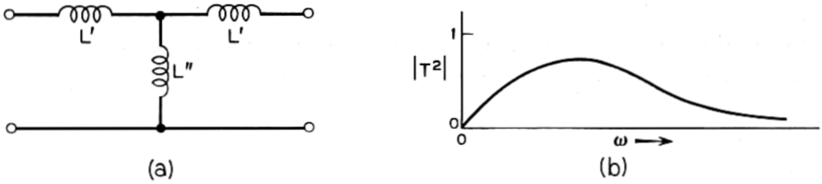

Fig. 17 — Network with stationary transmission.

former N within the Weissfloch equivalent and the various component values are shown in Fig. 19.

For completeness we provide the relevant formulations. If L is the numeric power insertion loss, we have

$$L = 1 + \frac{1}{4}\left(N - \frac{1}{N}\right)^2; \tag{36}$$

$$L = 1 + \left[\omega L' + \frac{\sqrt{3L'^2}}{1 + 6L'^2 - 3L^4}\left(\frac{\omega}{\sqrt{3L'}} - \frac{\sqrt{3L'}}{\omega}\right)(1 + \omega^2 L'^2)\right]^2; \tag{37}$$

$$L'' = \frac{1 + 6L'^2 - 3L'^4}{6L'^3}; \tag{38}$$

and

$$C'' = \frac{2L'}{1 + 6L'^2 - 3L'^4}. \tag{39}$$

The net results of these machinations are shown in Fig. 20. Since the network has a relatively high inductive impedance at midband, a short transmission line θ is used to present a resistive input. Thus input impedance is a minor perturbation and frequency effects are correspondingly small but, if desired, a low Q resonator matching network in tandem with the load may be employed for higher degree matching.

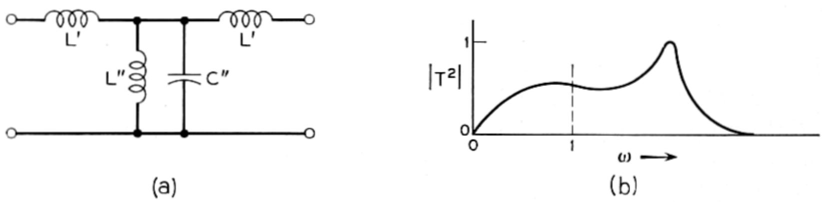

Fig. 18 — Modification to produce an inflection point.

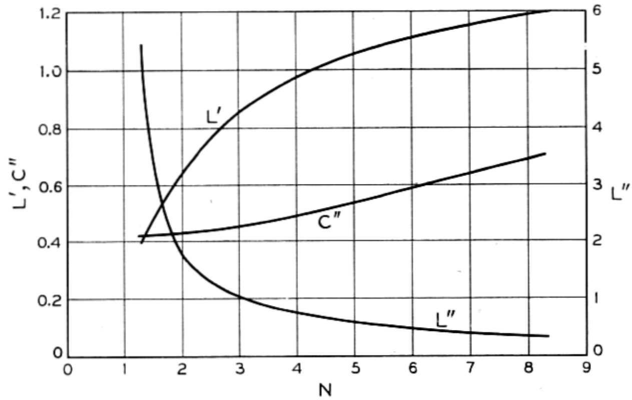

Fig. 19 — Component values for equivalent transformer.

The load conductance G_L is equal to $1 + 1/N^2$ in dual fashion to series injection form shown in Fig. 21.

5.2 *180° Phase Differential Networks*

The 180° phase differencing networks are staples of the interconnection formats we have described here. They can be realized to a good degree in the tens of magacycles range as a pair of mutually reversed ferrite transformers together with capacitive compensations to account for the differences in the way in which the winding capacitance is energized.[10] There was fear that the power handling capacity might prove excessive for the ferrite or that the parasitics might dominate. These

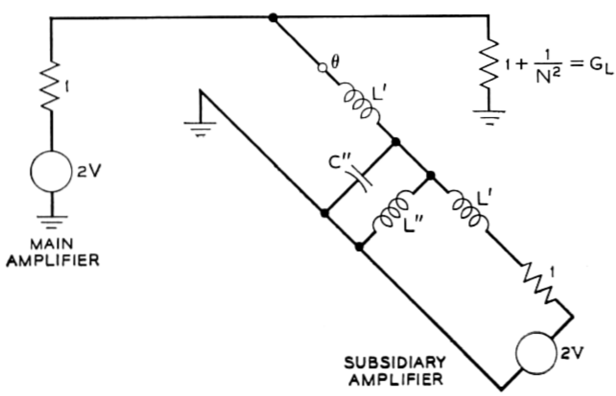

Fig. 20 — Shunt error injection using transformer network equivalent.

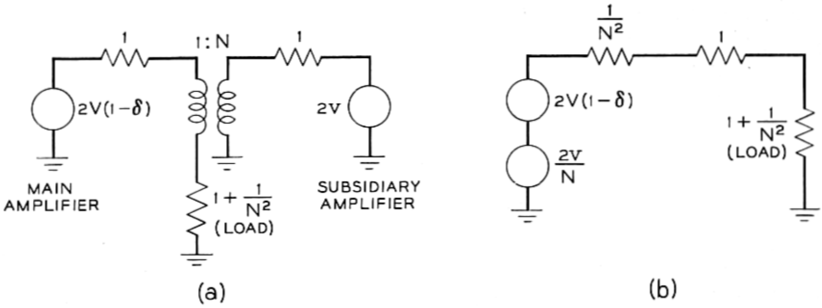

Fig. 21 — Thevenin equivalents of subsidiary amplifier feed into main amplifier output circuit.

were exactly the anticipatory fears with relation to the injection transformer and, in precisely like fashion, the fears proved groundless at the power level and the frequency range of operation.

Nevertheless, there is the question of what to do in the event that ferrite transformers are unrealistic. We use a single section quadrature coupler together with a technique credited to E. M. T. Jones to produce the desired realization. Jones's technique recognizes that the transmission between the input and its conjugate port, in a symmetrically mismatch terminated quadrature coupler, is directly proportional to the reflections terminating the coupled ports. The transmission in such a device whose coupled ports are short-circuited differs by 180° of phase from that same device coupled into open circuits.

Equations (10) and (11) show a single section coupler used in such application to have only a relatively narrow band for high return loss and that the transmission is more akin to that of a resonant section. Further analysis shows, indeed, that the identification with a resonator is a very good one and that the conventional technique of adding prior and subsequent resonators to form a filter chain is entirely applicable.

With the coupled ports open-circuited, the equivalent circuit is exactly a series resonator with both the inductance and capacitance corresponding perfectly to unit impedances at band center. A maximally flat three section filter chain is constructed by bounding this resonator on both sides with shunt resonators composed of elements possessing exactly double unit impedance.

If the coupled ports are short circuited, a dual situation results. The coupler equivalent is a shunt resonator composed of unit admittances and the bounding networks are series resonators composed of elements of twice unit admittance.

Since the open and short circuited couplers are perfectly dual structures, the 180° phase difference is perfectly maintained irrespective of frequency. The bandwidth of the structure pair is about 3:1 before the onset of any significant transmission loss; this results in an eminently practical component.

VI. REALIZATIONS

6.1 *Motivation*

The above philosophies and design considerations have been realized in the construction of two amplifiers intended for testing dispersive ultrasonic delay lines in the 60-90 Mc and 25-35 Mc ranges. The large loss inherent in these devices made necessary large power output while the dispersive nature of the delay imposed the requirements of linear flat gain with little intermodulation distortion. Since the delay line represents a nearly lossless capacitive load on the amplifier, the latter must be a well matched device in order to prevent large interaction effects between amplifier and delay line.

6.2 *60–90 Mc System Performance*

The first development undertaken was the 60–90 Mc amplifier. During the course of this project we learned much about the operation of a feed-forward system. This increased sophistication was applied to the subsequent 25–35 Mc system. The first amplifier is shown in Fig. 22, and it is seen to be a single stage feed-forward system with its input sampling loop and its output error correction loop which is capable

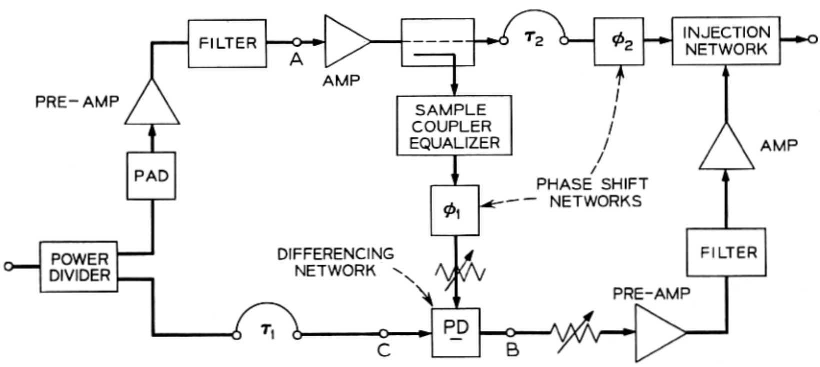

Fig. 22 — 75Mc feed forward system.

of supplying one quarter of the main amplifier power. This amplifier delivers 10 watts average power in CW operation (20 watts peak) to the 0.1 dB compression point.

At this power level the phase distortion is 1.25 degrees (see Fig. 23) at 75 Mc. The swept response across the band is shown in Fig. 24 both for small signals and at maximum power output. (The dashed curve shows the system response with the correction circuit disabled.) A sensitive measurement was made of the small signal gain and phase across the band (Figs. 25 and 26) which showed ±0.1 dB gain ripples superimposed on a 0.5 dB slope* and ±2° phase deviation from con-

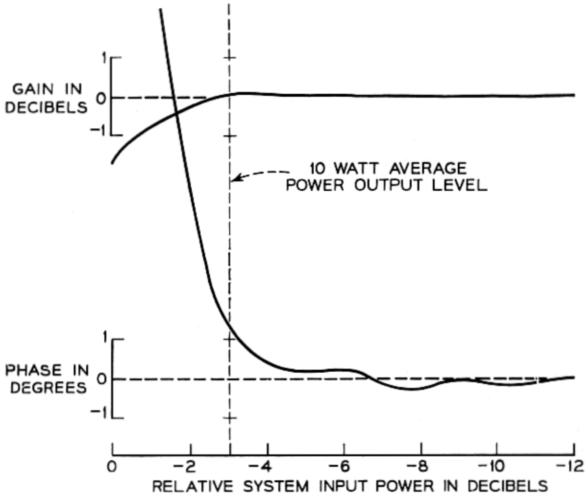

Fig. 23 — System gain and phase at 75Mc as a function of input power.

stant time delay (linear phase vs. frequency) out of a 1300° phase variation across the band.

6.3 *Couplers in 60–90 Mc System*

The couplers, used to pair amplifiers in the preamplifiers and in the more general harnessing in the main power amplifiers, were double stage quadrature hybrid couplers purchased from Merrimac Research and Development Co. (see Sections 1.2 and 1.3, and Fig. 27. The transfer characteristics of this coupler are shown in Fig. 28 and the input

* The ±0.1 dB ripples may have been partially due to a procedural error in the operation of the test set used for the measurement.

match with two identical total reflections in Fig. 29. This last figure points up one of the main advantages of the quadrature coupler type of organization; the fact that input match is maintained for arbitrary port termination as long as the two terminations are identical.

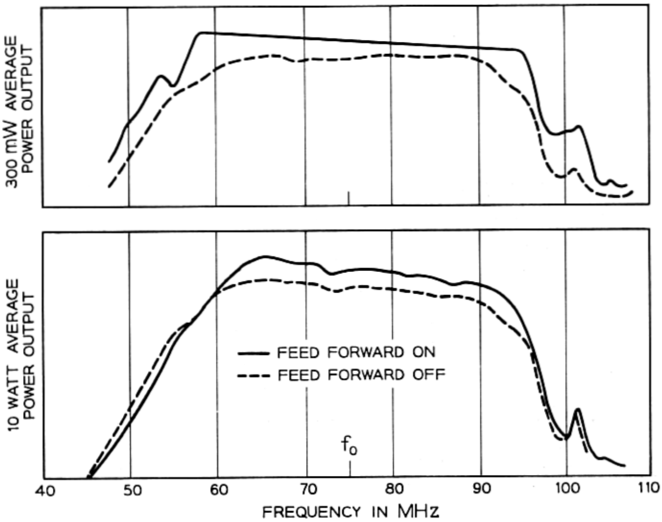

Fig. 24 — 75Mc system output as a function of frequency.

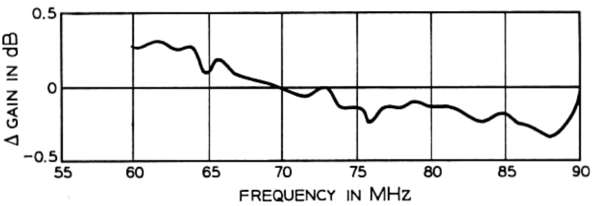

Fig. 25 — Gain variation with frequency at low power output.

6.4 *Amplifiers in 60–90 Mc System*

The preamplifiers shown in Fig. 22 are composed of an input pair of low-level high-gain amplifiers and enough stages of intermediate power emitter follower pairs emitter follower pairs (Fig. 30) to provide sufficient power to drive the main power array of emitter followers (Fig. 31).* All emitter followers are biased to class A operation.

* The amplifiers used were based on designs and models provided by R. V. Goordman and R. C. Petersen.

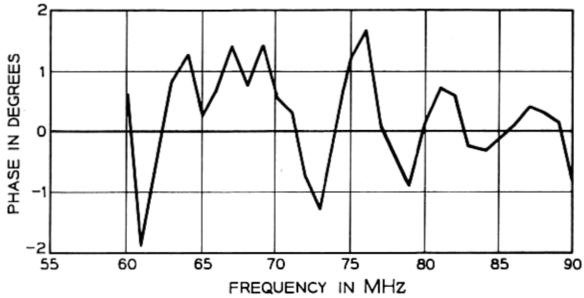

Fig. 26 — Phase deviation from linear phase vs. frequency behavior (Constant time delay). Average time delay = 2.5μsec.

The high gain amplifiers are of a broadband design with a flat 52 dB gain from 0.2 Mc to 100 Mc and capable of 100 milliwatts output before compression at the higher frequency (greater power at lower frequency).

The drive stage emitter followers in the preamplifiers used 2N3375 transistors they have sufficiently high base impedance at these frequencies to permit some modest input transformation in order to enhance the 6 dB theoretical gain made available by driving the high

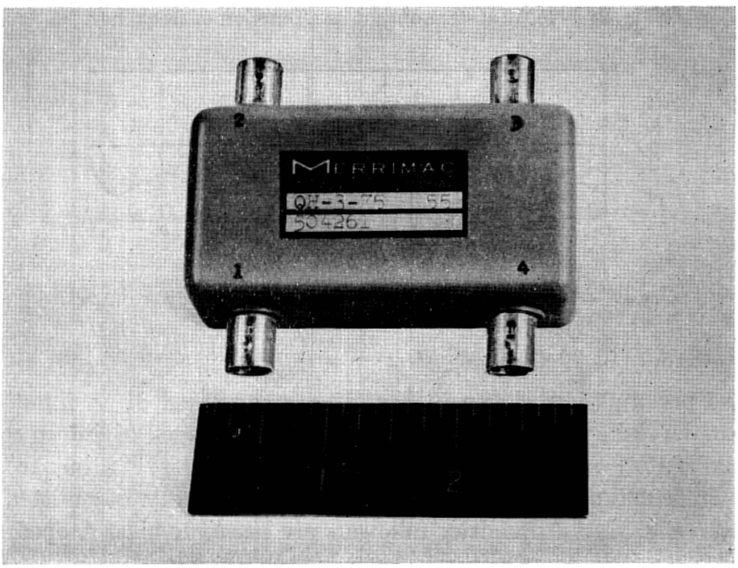

Fig. 27 — Quadrature hybrid coupler.

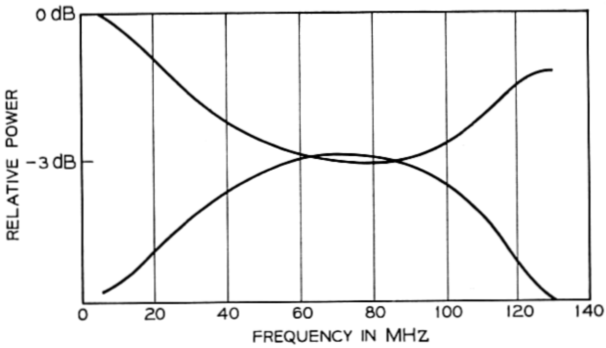

Fig. 28 — Coupling characteristics of 2 section coupler.

impedance emitter follower input from a matched source. The driver stage amplifiers had gains of 8.5 to 9 dB and were capable of 0.6 to 0.7 watts output. The 2N3632 transistors used in the power stage amplifiers, however, did not allow any transformation and, after some input tuning, had 6-7 dB gain but were capable of 1.25 watts output.

In order to realize the matching properties of the coupler organization it was found necessary to select matched pairs of transistors. The transistors were inserted in a test jig and tuned for best reponse with calibrated reactances. The values and final response were noted. Pairs of transistors were selected by comparison of these data. The amplifier pairs constructed using them showed typically > 25 dB

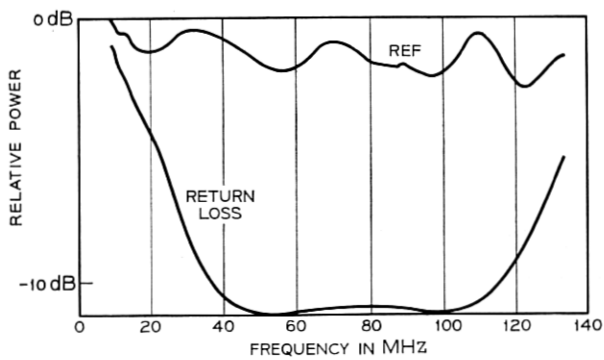

Fig. 29 — Return loss of 2 section coupler with open circuit terminations on coupled ports.

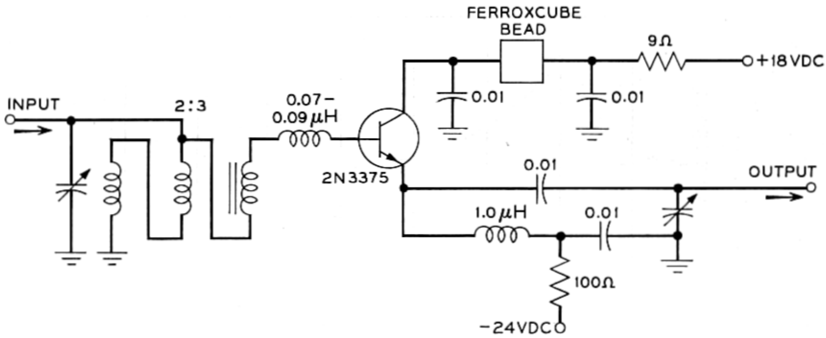

Fig. 30 — 60-90 Mc class A, 8.5-9.0 dB gain, 600-700 mW linear driver stage.

input and output return loss across the band. Interaction between the amplifiers outside of the band is prevented by using "out-of-band" filters as described in Section 2.6 (see Fig. 22).

6.5 *Signal Processing Components in 60–90 Mc System*

In the feed-forward system, as developed in the preceding text, there are six essential signal processing operations (see Fig. 22):

(*i*) Splitting off a reference signal from the input

(*ii*) Delaying this reference signal for a time equal to the time of passage through the main amplifier

(*iii*) Splitting off a sample of the amplified signal

(*iv*) Comparison of the sample with the reference

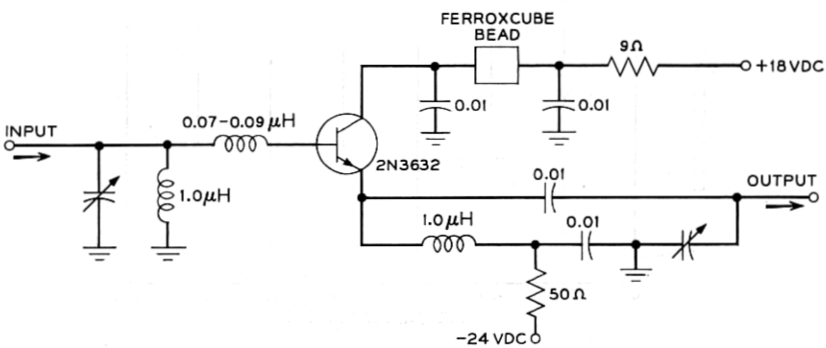

Fig. 31 — 60-90 Mc class A, 6.0-7.0 dB gain, 1.25 watt linear power stage.

(*v*) Delaying the amplified signal for a time equal to the time of passage of the error through the subsidiary amplifier

(*vi*) Injecting the correction into the output path.

In the 60-90 Mc system operations *i* and *ii* are carried out within in-phase power dividers (Merrimac Research and Development Co. PD-75). Operations *ii* and *v* are carried out in approximately 60- and 30-foot lengths of coaxial cable. (These lengths were first calculated from measurements of phase vs frequency within the system.) Operation *iii* is carried out in a two-section quadrature hybrid coupler similar to the ones used within the amplifiers (see Figs. 27, 28 and 29). However, instead of a 3 dB power split, this coupler was synthesized to couple out a signal 15 dB down from the input and to present only 0.27 dB loss to the main power path (see Fig. 32). The final operation, (*vi*), uses the reactive network described in Section 4.1 and Fig. 18.

The other components appearing in Fig. 22 are of secondary importance to the feed-forward concept but are quite essential to system operation. The fixed pad in the main amplifier input and the variable attenuator in the sample path insure that the power levels are appropriate to a proper differencing in the power divider. The variable attenuator in the error correction loop performs a similar function. These variable attenuators were constructed based on the transmission and match properties of the quadrature hybrid coupler. Figure 33 shows the transmission between the input and hybrid ports of a two-stage quadrature coupler both under open circuit termination of the coupled ports and with 10 Ω terminations. The use of ganged carbon rheostats as terminations (Fig. 34) produces a well-

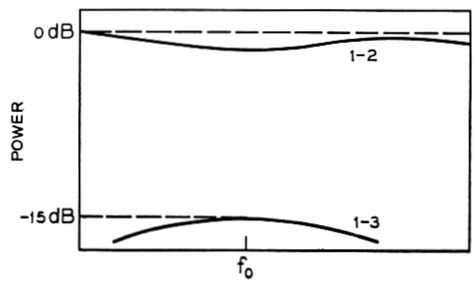

Fig. 32 — Coupling characteristics of hybrid directional coupler used as sampling coupler.

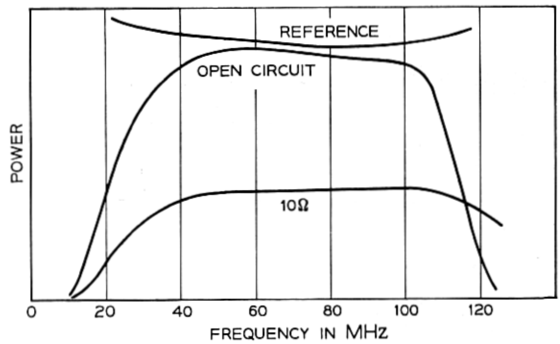

Fig. 33 — Transmission of two section coupler under open circuit and 10 Ω termination.

matched variable attenuator which is quite flat over the modest attenuation ranges needed.

The "sample coupler equalizer" is a low Q shunt resonant circuit which has a transmission characteristic that compensates for the bow in the sampled signal from the coupler, as shown in Fig. 35. The "phase shift networks" shown in Fig. 22 are also resonant equalizers. When the cable length corresponding to τ_1 was adjusted such that the time delay through the main amplifier equaled that through the reference path (equal phase vs frequency slopes), it was found that

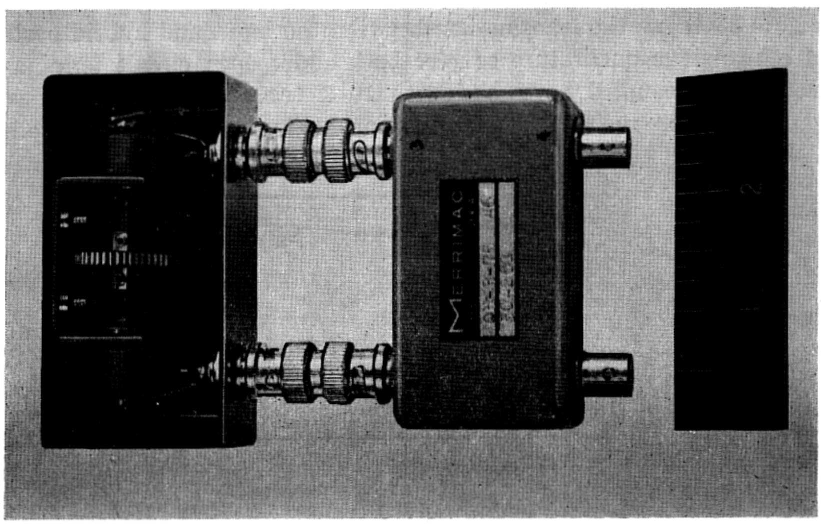

Fig. 34 — Variable attenuator.

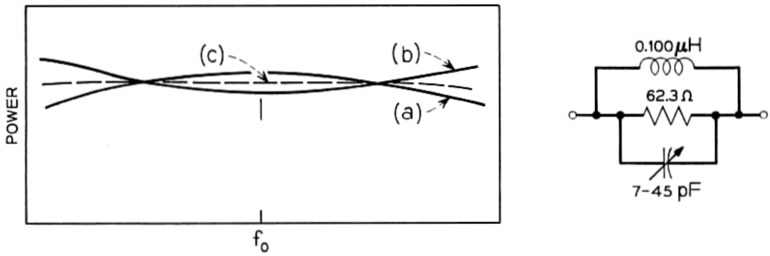

Fig. 35 — Sampling coupler equalization. (a) Sample coupler characteristics. (b) Transmission of equalizer. (c) Equalized characteristics.

there was a constant 70° excess phase shift through the amplifier out of a total variation of 1500° across the band (see Fig. 36). In order to equalize this difference, use was made of the fact that a resonant circiut has zero phase shift at the resonant frequency, at which point there is an inflection in the phase vs frequency characteristic (see Fig. 37). The latter produces constant time delay at the resonant frequency. An extra length of line was then added to the reference path to produce equal phase shift at center frequency and the two-resonator network shown in Fig. 38 added to the sample path to equalize the phase slope (see Fig. 36). It was found that a two-resonator net-

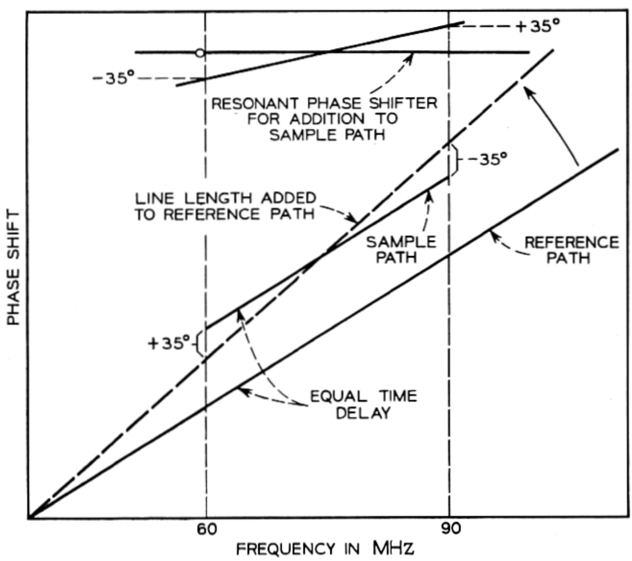

Fig. 36 — Phase equalization of reference loop.

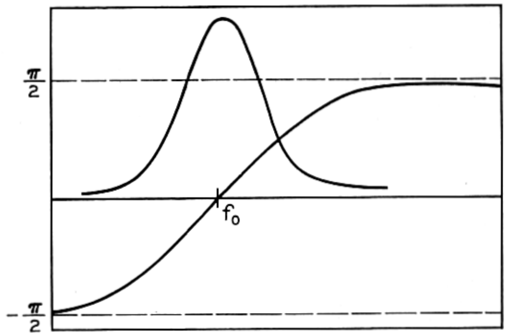

Fig. 37 — Amplitude and phase of transmission through a resonator.

work was needed, minimally, to produce sufficient flatness across the band. The insertion of the phase shift network, φ_2, in the correction loop relative to τ_2 is similar to the above.

6.6 *System Checks*

With the system assembled, each of the two loops was put through an operational check, and final adjustments were made. The reference loop is checked by applying a small signal to the input and observing the output of the comparison power divider at *B*. The variable attenuator in the power sampling line was adjusted to produce the best null across the band. Small adjustments of τ_1 were made and the best null was obtained by using a sweep generator and an oscilloscope.

The error correction loop was then adjusted by feeding a swept frequency signal into the main amplifier at A and observing the system output. Both the variable attenuator at the input to the subsidiary amplifier, and τ_2, were adjusted to produce a null output.

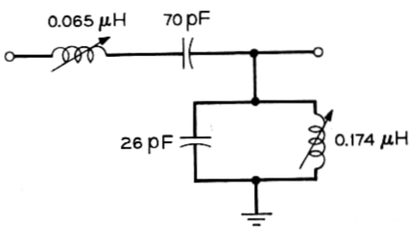

Fig. 38 — Phase shift section (values shown for 75 Mc, 70° correction) in 50 Ω system.

The final experimental proof of system performance lies in breaking (and terminating) the reference line at C in order to "fool" the error correction loop into operating as if there were no input signal, and observing the entire system between input and output ports. The excellence of the correction system may be judged by the degree of degeneration of the output. The results of this observation with the subsidiary amplifier turned off, and with the subsidiary amplifier restored are shown in Fig. 39 along with a simplified schematic indicating the measurement condition.

6.7 *25–35 Mc Amplifier—Differences from 75 Mc System*

After experience with the 75 Mc system, the 30 Mc amplifier was designed to increase the power output several fold over the 10 watts attained above, without increasing the complexity of the system (that is, the number of coordinated branches in the power amplifier). This was done by going to class B push-pull operation rather than to a single-ended class A operation as in the previous system. In addition to doubling the number of transistors, the dc-rf power conversion capability of each was doubled. The push-pull pair (see Fig. 40) delivered 7.5 watts, linearly.

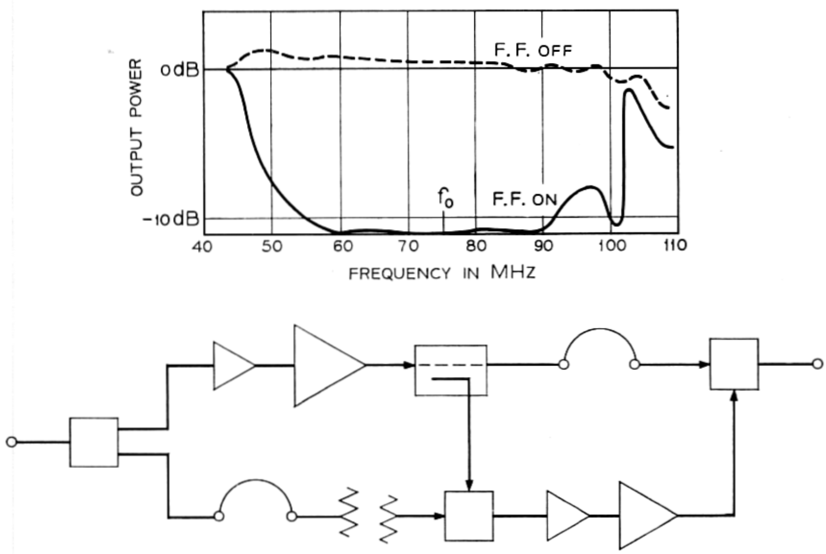

Fig. 39 — System output with reference line open, feed forward switched on and off.

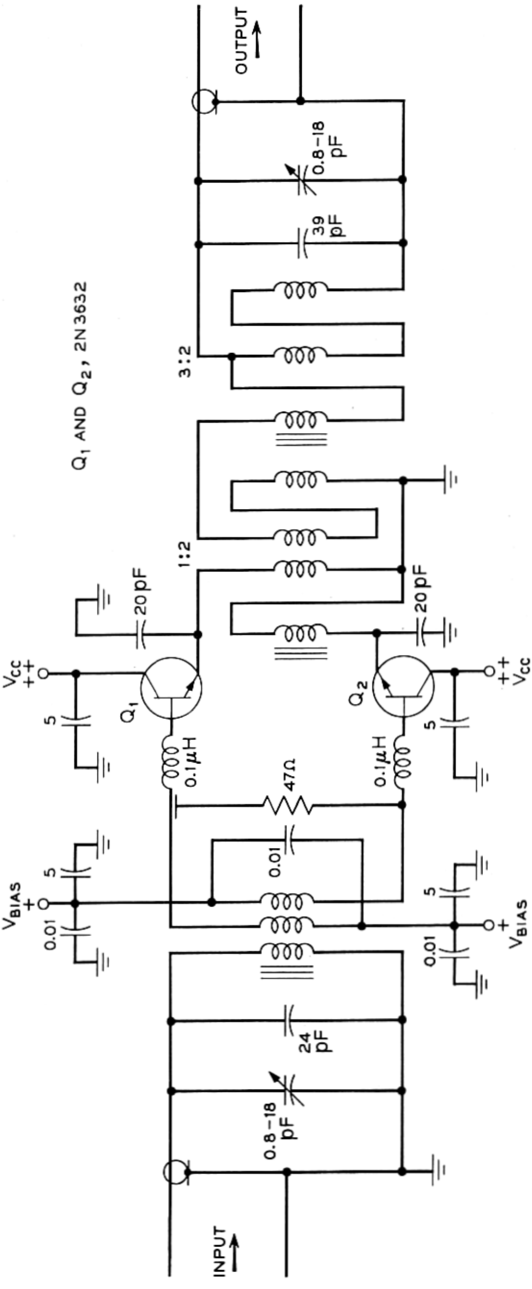

Fig. 40 — Class B push-pull amplifier for 30 Mc system (20-40 Mc, class A-B, push-pull, 7.5 watts linear).

This transition to class B operation made another design change possible. It was decided to design for low duty cycle pulsed operation and to use an "active clamp" biasing circuit which samples and restores bias voltages between pulses.

In addition to extending the dynamic range of the amplifier by increasing its power output capability, an attack was made on the system noise figure by the use of a directional coupler at the input favoring the reference path. As a further improvement the series input matching resistors ($\sim$43 Ω) of the low-level preamplifier in the subsidiary amplifier were removed. The input match was restored by the amplifier pairing operation using quadrature couplers, which produced a matched amplifier having approximately a 4 dB noise figure.

Other changes were in the use of a series 3:1 injection transformer instead of the previously used shunt injection network, and in the use of limiting in the low-level preamplifier sections of both amplifier chains to prevent destruction of the power transistors under accidental overdrive. A schematic diagram of this system appears in Fig. 41.

6.8 *"Active Clamp" Bias Circuit*

The active clamp biasing used in the 25-35 Mc amplifier (see Figs. 42 and 43 provides a means of stabilizing the amplifier against changes in bias condition with variations in the signal power level and duty cycle. At high signal levels an unrealistically large bypass capacitor (see Fig. 44) would otherwise be required to perform this function. The clamp circuit also serves to make the bias levels insensitive to

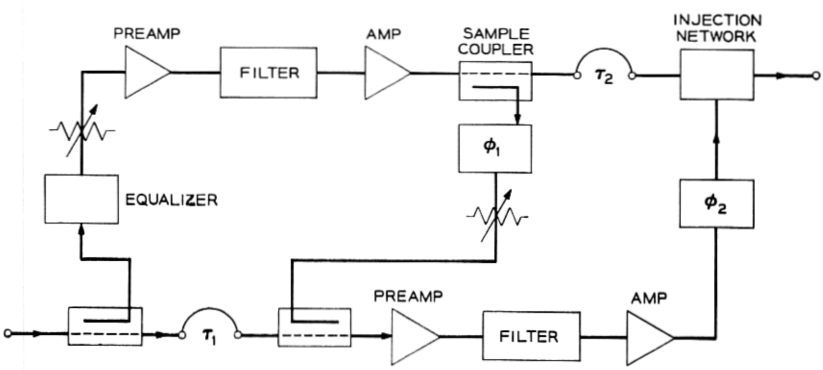

Fig. 41 — 30 Mc system.

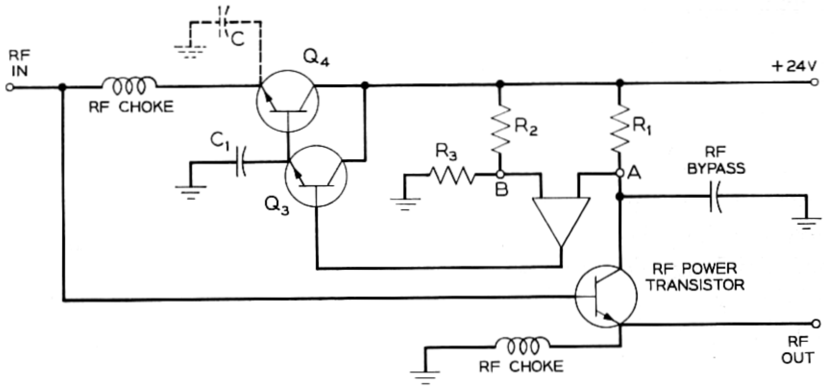

Fig. 42 — Active clamp schematic.

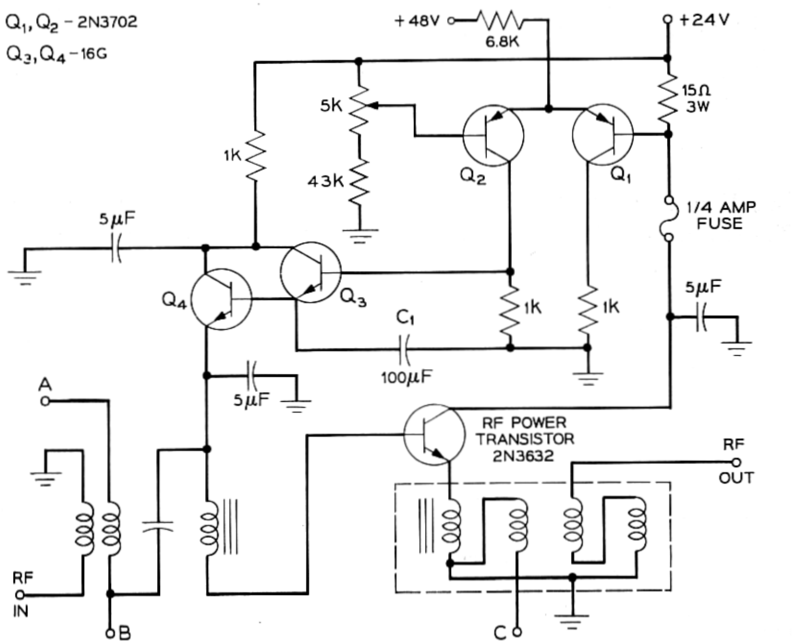

Fig. 43 — Active clamp biasing of power transistor. A, B, and C are connections to the other transistor of the push-pull pair.

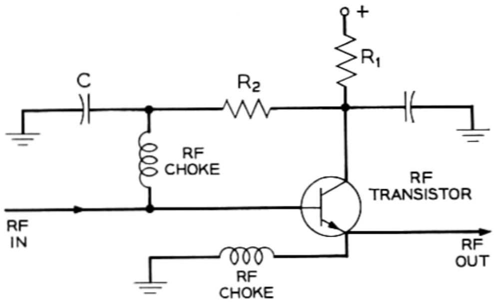

Fig. 44 — Conventional transistor biasing arrangement with DC stabilization.

thermal drift in the controlled transistor and to parameter variations between substituted transistors.

The high gain differential amplifier, shown in Fig. 42 provides the self-correcting bias condition since its output is fed into the base of the power transistor in such a way as to keep the potential at A of Fig. 3 equal to the reference potential at B. The gain of the differential amplifier also allows the use of a much smaller value of R_1 (15 Ω as in Fig. 43) than would be used in a conventional current feedback biasing scheme (shown in Fig. 44) to sense variations in quiescent current through the power transistor.

Ordinarily, as stated, a large bias bypass capacitor C, shown in Fig. 44 and dotted in Fig. 42, must be used to provide base current to the transistor during the conducting part of the cycle. It must be made adequately large to prevent substantial discharge, independently of duty cycle or drive amplitude. Transistor Q_4 of Fig. 44, allows the use of a much smaller capacitor, C_1, since this capacitor need only supply a control current to its base. The transistor, Q_3, prevents the discharge of C_1 into the differential amplifier during the pulse and allows a rapid recharge of C_1 after the pulse because of its low emitter follower output impedance.

6.9 *Performance of the 25–35 Mc Amplifier*

The characteristics of the completed amplifier at low power are shown in Figs. 45 through 48. These figures show that the gain of the amplifier slowly oscillates ± 0.1 dB about a 0.4 dB slope across the band, the total amplification being 39 dB. The phase varies $\pm 1°$ from a constant time delay phase characteristic. Further, the spot noise

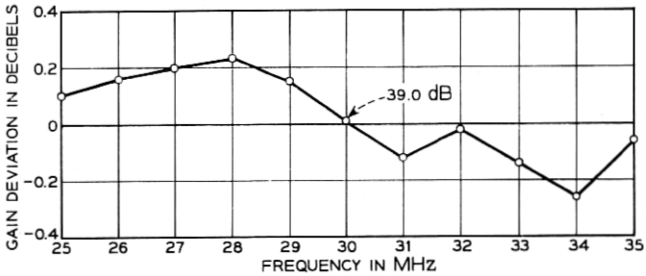

Fig. 45 — Deviation from flat gain at low power.

figure is better than 12.5 dB across the band and better than 9.5 dB over 85 percent of the band. The integrated noise figure is 8.6 dB.

Figures 49 through 51 show the gain and phase as functions of power level with the system fully functional and with the feed-forward correction loop disabled. These figures show the marked improvement obtained by the use of feed-forward correction. Notice that, as implied by the analysis in Section 3.3, when the subsidiary amplifier is called on to deliver large power to compensate for gain variation, it can no longer provide phase compensation. (It was shown that the compensation of only 3.6° at maximum power output requires all the power capability of the subsidiary amplifier.) The 40 watts indicated on the figures and the 8.6 dB noise figure shown above yield a dynamic range for the amplifier of 103.5 dB. Its output return loss was > 25 dB (see Fig. 52).

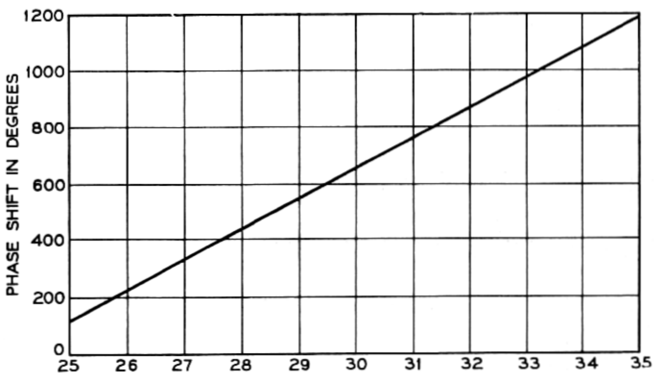

Fig. 46 — Phase shift through 30 Mc amplifier system as function of frequency (low power).

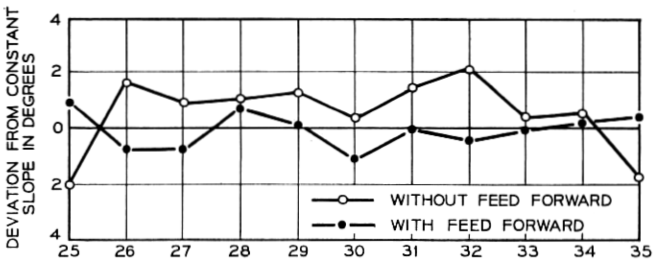

Fig. 47 — Deviation from constant time delay (low power).

VII. CONCLUSIONS

The amplifier systems that were constructed at 30 and 75 Mc had an excellent correlation with design and showed the engineering practicability of the methods of design construction. The primary features were the use of:

(*i*) Quadrature coupled corporate amplifier arrays to phase coordinate power emission of the individual active elements

(*ii*) Emitter follower amplifiers as the basic elements to provide broadband, low distortion gain with transistors, otherwise incapable of producing this type of performance with a high yield

(*iii*) A time-compatible or "feed-forward" error injection technique to equalize the transfer function of the amplifier, reduce distor-

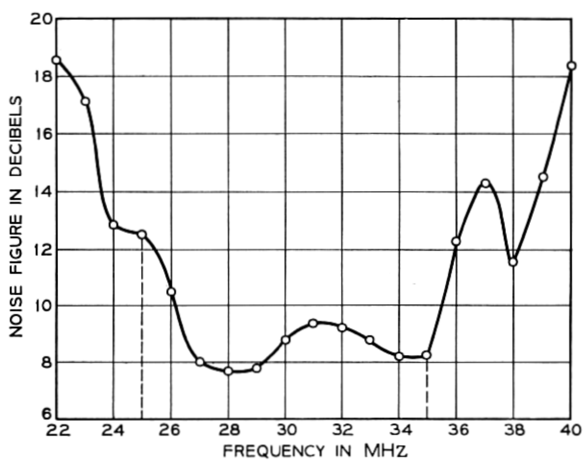

Fig. 48 — Amplifier noise figure.

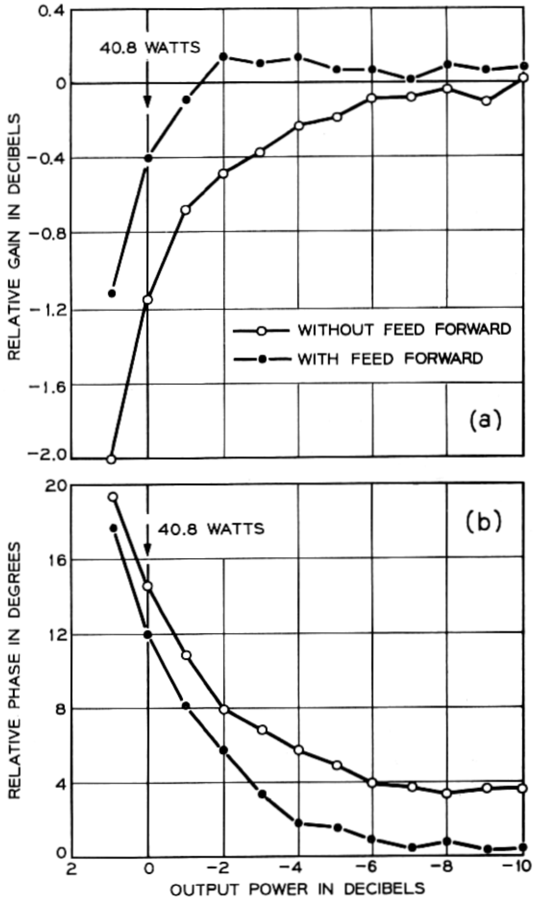

Fig. 49 — Gain (a) and Phase (b) Variation at 28 Mc.

tion and diminish noise in a system having relatively large time delay

(*iv*) Directionally coupled error sensing to produce a high return loss at the amplifier output

The 75 Mc amplifier was operated over a band of 60–90 Mc producing a 10 watt output average power with an amplitude compression there of about 0.2 dB. Its transfer function showed slow amplitude variations within 0.1 dB and phase deviations from linearity of about 1.5° out of a total phase spread of greater than 1500° across the band.

The 30 Mc amplifier had a band of operation between 25 and 35

Mc. It produced a 40 watt average power with an amplitude compression of 0.2 dB. Transfer characteristics showed slow amplitude excursion of about 0.2 dB with phase distortion about 1°. Output return loss was about 25 dB across the band and lent itself to simple equalization.

The 30 Mc amplifier had an 8.6 dB noise figure across the 10 Mc band of design interest which, together with its 40 watt average undistorted output capability, provided a dynamic range of 103.5 dB.

While correction did not extend beyond that of a single loop in these amplifiers, the textbook-like adherence of the results to the design elements makes it clear that further feed-forward looping would provide an arbitrary degree of correction over the prescribed bandwidth, probably with some further noise improvement. This is to be contrasted to the inabiilty to use a feedback control for these

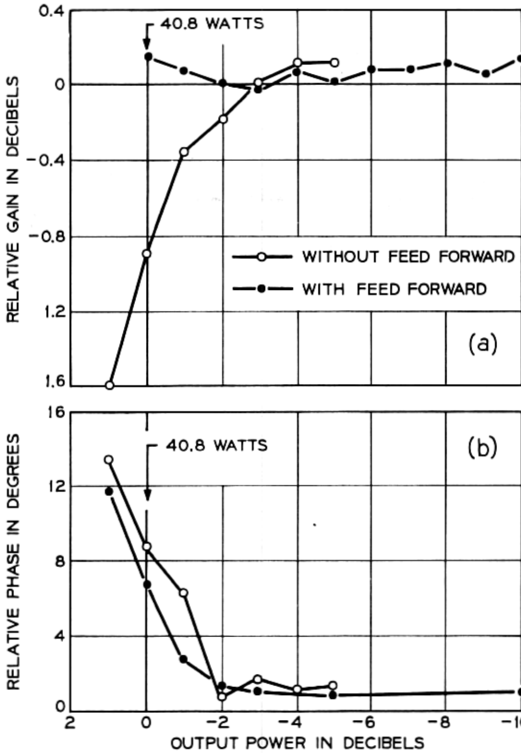

Fig. 50 — Gain (a) and Phase (b) Variation at 32 Mc.

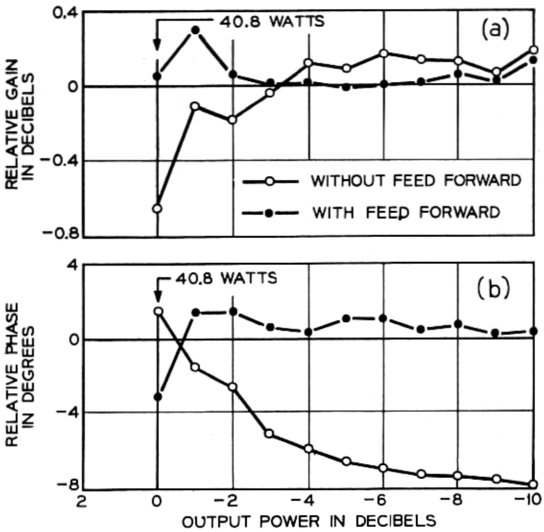

Fig. 51 — Gain (a) and Phase (b) Variation at 35 Mc.

amplifiers specifically because of excessive delay over the bandwidth, and to the limited ability of feedback, in general, to degenerate error cumulatively because of stability concerns.

The feed-forward correction scheme does not impose any concern for active element characteristics beyond the band of use, as does

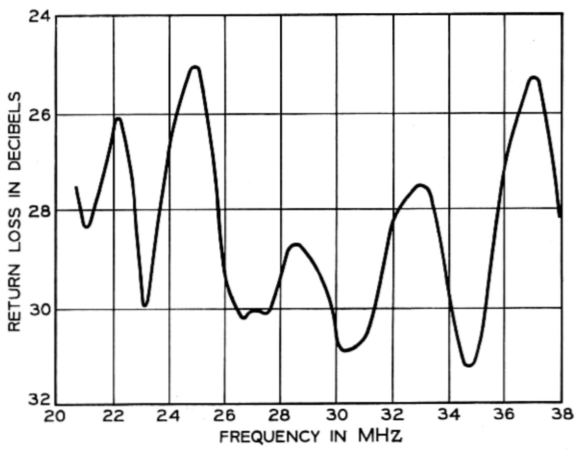

Fig. 52 — Output return loss of 30 Mc amplifier.

feedback, providing the simultaneous ability to use higher power and cheaper transistors than would ordinarily be acceptable at the frequency used. Using the self-degenerating emitter-follower mode of amplification, experience with the two amplifiers constructed showed a high degree of insensitivity to transistor selection aside from those failing catastrophically. Nothing more than a dc transistor checker was needed to construct the 30 Mc unit, while a rough rf sweep selection procedure was employed at 75 Mc to form slightly preferred transistor pairs.

The use of time-compatible error cancellation is not limited to the specific circuits described here either in frequency or in the specific active element circuitry used. It is generally applicable and, in conjunction with the corporate power accumulation available through coupler arrays, and smooth (no guard-band) frequency multiplex arrangements, there are available at least in principle, controlled, arbitrary (bandwidth, arbitrary power, capabilities virtually throughout the entire rf spectrum.

VIII. ACKNOWLEDGMENTS

We gratefully acknowledge the expert skills of others who helped this activity succeed. We are particularly indebted to R. V. Goordman whose teaching, consultation, and services we most often sought in the area of transistor electronics; to R. C. Petersen, for providing an initial feasibility demonstration of the linear power capabilities, at high frequency, of an overlay emitter follower transistor amplifier; to A. B. Wertz for the extensive and highly refined measurements of amplifier performance characteristics; and not least to J. J. Golembeski for his critical, and most informed, reading of the manuscript.

APPENDIX A

Conditions for Directional Coupling

In Section 2.2 we indicated that a sufficient condition that a passive, reciprocal, and reactive, fourport possess directional coupling properties was that it be matched and that it possess two axes of symmetry. We shall show the sufficient conditions to result from the application of simple time reversal arguments.

Figure 53 shows the fourport excited by a unit incidence into port 1 without reflection but with scatterings α, β, and γ, into ports 2, 3, and 4, respectively. If this figure is time-reversed, being reactive, the

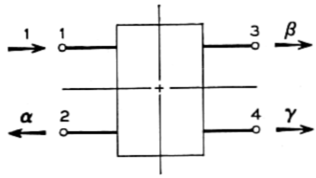

Fig. 53 — Prototype network.

fourport remains unchanged, but the scattered quantities are reversed in flow direction and their phases inverted. This result is shown in Fig. 54 which then must be an excitation format compatible and simultaneous with that of the first figure.

Use in turn incident amplitudes, $-\beta^*$, $-\gamma^*$, and α^*, into ports 3, 4, and 2, respectively, to cancel the various incident quantities of Fig. 54. The respective scattered results are shown in Fig. 55, where the two axes of symmetry and the scattering format of Fig. 53 were used. If the results of Fig. 55 are superposed, then the excitation scheme of Fig. 56 is yielded. This last figure shows only emergent waves. Since there are no incident waves, each of the emergent waves must vanish and we have

$$1 - \alpha\alpha^* - \beta\beta^* - \gamma\gamma^* = 0, \tag{40}$$

$$\alpha\gamma^* + \gamma\alpha^* = 2 \operatorname{Re} \alpha\gamma^* = 0, \tag{41}$$

$$\alpha\beta\alpha + \beta\alpha^* = 2 \operatorname{Re} \beta\alpha^* = 0, \tag{42}$$

and

$$\gamma\beta^* + \beta\gamma^* = 2 \operatorname{Re} \gamma\beta^* = 0. \tag{43}$$

Equation (40) simply confirms the conservation of energy in a reactive fourport. Equations (41) through (43) state that α, β, and γ are all mutually perpendicular phasors. Since these three phasors

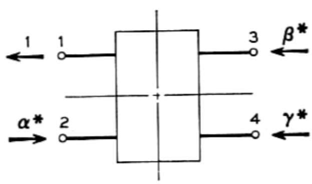

Fig. 54 — Time reversed description.

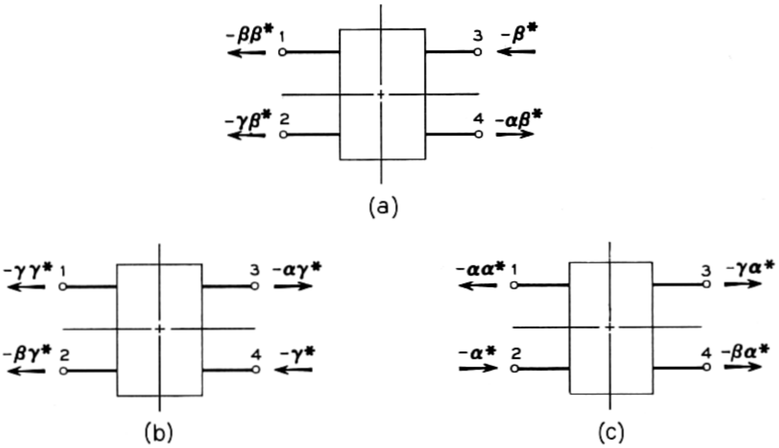

Fig. 55 — Excitation of 'β' Port (a), 'γ' Port (b) and 'α' Port (c).

are in a two-dimensional space this mutual perpendicularity is absurd unless one of the quantities vanishes. Therefore,

(*i*) There exists one port "hybrid" to the incident port
(*ii*) All the power is transmitted to the remaining two ports.

These two statements define a directional coupler.

APPENDIX B

Transformer Injection

B.1 *First Considerations*

B.1.1. *In Phase Errors*

Let us first see what is required of the circuitry of Fig. 14 to accomplish in-phase error correction then consider the correction of

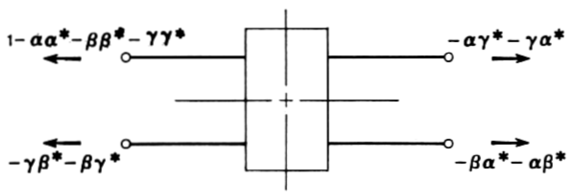

Fig. 56 — Superposed excitations.

quadrature phase errors. Our specific concern is the specification of the $1{:}N$ transformer shown as the means for final injection of the error cancellation signal.

Figure 21 shows a first Thevenin equivalent of the insertion of the subsidiary, error correction, amplifier into the output circuit of the main amplifier. Both amplifiers are shown with normalized internal impedances and the external load matched to the combined impedance. We assume proportional error δ, of the main amplifier open circuit voltage 2V, and we assume further that the open circuit voltage of the subsidiary amplifier is 2v.

As seen from Fig. 21(b) the reduced equivalent, the condition that the load be excited independently of δ is given by

$$v = NV\delta. \tag{44}$$

If we isolate the subsidiary amplifier, as in Fig. 57, the amplifier has a voltage v' at its load terminals corresponding to a load current $V/N(1 + 1/N^2)$. At the maximum voltage of V, defined as V_p, with corresponding subscript notation on all other quantities, we would like to see the subsidiary amplifier match terminated to deliver maximum power. Accordingly, v' is half the open circuit voltage and it is equal to the current at match in the impedance normalized description. Then

$$v_p = \frac{V_p}{N\left(1 + \frac{1}{N^2}\right)} \tag{45}$$

and, using (44), we have

$$N = \sqrt{\frac{1 - \delta_p}{\delta_p}} \tag{46}$$

and

$$v_p = V_p\sqrt{\delta_p(1 - \delta_p)}. \tag{47}$$

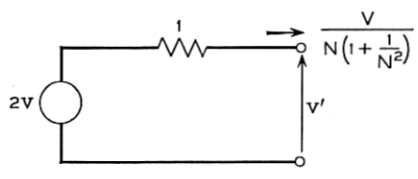

Fig. 57 — Thevenin equivalent of loaded subsidiary amplifier.

By the above construction, at peak power the subsidiary amplifier works into a match and furnishes a power P_s which compares to the main amplifier P_m, via (46), by the relationship

$$P_s = \delta_p(1 - \delta_p)V_p^2 = \frac{\delta_p}{1 - \delta_p} P_m \tag{48}$$

where $V_p^2(1-\delta_p)^2 = P_m$ is the peak available power of the main amplifier.

We may consolidate our results to this point under the approximation that δ_p is small.

(*i*) The main amplifier power deviation is approximately $[(1-\delta_p)^2 - 1]P_m \simeq -2\delta_p P_m$, at its peak power capability.

(*ii*) The subsidiary amplifier furnishes a correction power of approximately $\delta_p P_m$.

There seems a curious divergence of a factor of 2 in providing the necessary compensation. If, for example, $\delta_p = \frac{1}{8}$, implying a peak-power deviation of $\frac{1}{4} P_m$, a subsidiary amplifier having only a power capacity of $\frac{1}{8} P_m$ would appear to be adequate for compensation.

A second feature of transformer injection may also be discerned which at first would appear unrelated to the missing factor of 2. The error injection is not without some cost to the main amplifier. As Fig. 21 shows, the subsidiary amplifier images a resistance of $1/N^2$ into the main amplifier circuit, causing an effective generator characteristic impedance of $1 + 1/N^2$ to be presented to the load.

The subsidiary amplifier provides the power both to equalize the irregular main-amplifier transfer characteristic and to cancel distortions produced by nonlinearities. Assume the main-amplifier power to be small such that no nonlinear distortions are observed and assume further that equalization is unnecessary. There is, therefore, no error and, consequently, no contribution from the subsidiary amplifier, and the open circuit voltage is that of the main amplifier only. The available power is decreased by an amount equal to the changed impedance and an insertion loss of

$$L = 1 + \frac{1}{N^2} = \frac{1}{1 - \delta_p} \tag{49}$$

is sustained where the results of (46) have been used.

The question exists whether this is indeed just a loss in gain or a loss in power capability. We shall now show that it is only a loss in gain and that we add to the power capability of the main amplifier by the amount of the subsidiary amplifier peak power.

From Fig. 21 we see the delivered voltage of the main amplifier to be, at peak,

$$2V_p(1 - \delta_p) - \frac{V_p}{1 + \frac{1}{N^2}}.$$

Since $1 + 1/N^2 = 1/1-\delta_p$ by (46), the voltage delivered reduces to $V_p\ (1-\delta_p)$. Since the current drawn is $2V_p/2(1 + 1/N^2) = V_p\ (1-\delta_p)$, the total power delivered by the main amplifier is

$$P_m = V_p^2(1 - \delta_p)^2. \tag{50}$$

However, (50) is, identically, the available power from the main amplifier without the subsidiary circuit, so that there is no difference in the power actually delivered by the main amplifier into the load.

The power delivered by the subsidiary amplifier is the product of its output voltage referred to the main loop, $2v_p/N - [(1/N^2 \times V_p/(1 + 1/N^2)]$ and the main loop current $V_p/(1 + 1/N^2)$. The use of equations (46) and (47) shows this power to conform to (50) producing a power

$$P_s = V_p^2(1 - \delta_p)\delta_p .$$

The total power reaching the load is $P_m + P_s = P_L$ which is

$$P_L = V_p^2(1 - \delta_p) = \frac{P_m}{1 - \delta_p} \tag{51}$$

Equation (51) shows most succinctly that the power reaching the load exceeds the power capability of the main amplifier and that the transformer interconnection, as characterized earlier, is vastly superior to a power divider interconnection which would have produced a 3 dB loss in power capacity. Further, the insertion loss shown in (48) is not a power loss, but a small loss in gain only.

B.1.2. *Quadrature Errors*

Let us assume that the transformer N has been specified by (47) and the subsidiary amplifier by (48) in terms of some desired real peak voltage error δ_p. We now seek to determine the peak capability of the system to correct a quadrature error $i\delta_p'$.

Given this quadrature error, the subsidiary amplifier output voltage is $v = iNV_p\delta_p$ corresponding to its current $i = V_p/N$. The subsidiary amplifier perceives, therefore, a terminating reflection factor

$$\frac{v - i}{v + i} = \frac{iNV_p\,\delta_p' - \dfrac{V_p}{N}}{iNV_p\,\delta_p' + \dfrac{V_p}{N}}$$

which, obviously, has a unit magnitude.

Given a unit reflection, the subsidiary amplifier, depending on its electrical position relative to the load, might be called on to develop twice its rated voltage or twice its rated current. To safeguard against this possibility, the open circuit voltage must be down rated by a factor of 2, and corresponding to (47), we have

$$v_p' = \frac{iV_p}{2}\sqrt{\delta_p(1 - \delta_p)} = NV_p\,\delta_p' \,. \tag{52}$$

So that, by (46)

$$\delta_p' = \tfrac{1}{2}\delta_p \,. \tag{53}$$

Quantitatively, if $\delta_p = \frac{1}{8}$, corresponding roughly to a ± 1 dB correction, (53) yields $\delta_p' = 0.0625$. Since the phase correction has a magnitude of the arctangent of δ_p', the phase cleanup range is, minimally, 3.6° at high power.

B.2 *Further Considerations*

B.2.1. *Subsidiary Amplifier Mismatch*

At peak power, by design, the subsidiary amplifier is matched. Let us now go to the extreme of low power. Using Fig. 21(b) the main amplifier supplies a current $V/(1 + 1/N^2)$ which develops a voltage

$$\left[\frac{V}{1 + \dfrac{1}{N^2}}\right]\left(\frac{1}{N^2}\right)(N) = \frac{V}{N + \dfrac{1}{N}} \tag{54}$$

across the step-up side of the injection transforms via the subsidiary amplifier image impedance $1/N^2$. The main current reduces by $1/N$, and a current flow also equal to $V/(N + 1/N)$ flows through the voltage step-up side of the transformer.

As at high power the subsidiary amplifier output voltage is equal to its current, but with a major difference. The voltage sense is reversed and now, instead of finding a positive resistance match, the termination is a negative match. In transition between high and low power, the termination of the subsidiary amplifier has passed through violent shift and a question exists whether, as power decreases, some requirement existed on the subsidiary amplifier worse than that at peak main amplifier power.

Having aroused this concern, we shall now dismiss it. Equation (54) shows the negative voltage bound to be less than the positive voltage bound, since V, by hypothesis, is less than its peak voltage V_p. Similarly, the subsidiary amplifier current is always less than its peak value $V_p/(N + 1/N)$. The subsidiary amplifier, in actuality, is separated from the terminating transformer by a transmission line. Along this line, peak current and voltage amplitudes interchange. Since neither of these peaks ever exceeds the value supplied by the subsidiary amplifier at peak main-amplifier distortion, the initial design of the subsidiary amplifier covers the worst case.

B.2.2. *Nonoptimal Transformers*

While the transformer design chosen optimizes the powers delivered by the main and subsidiary amplifiers, nevertheless one may wish to forego this optimization somewhat to reduce the contribution of $1/N^2$ to the output matching problem. While this problem is not particularly serious, it is generally, a nuisance in the following sense. Very high return loss is guaranteed by the directional coupler sampling arrangement exclusive of the $1/N^2$ contribution, and the very insertion of any matching circuit at all is a severe perturbation on the goodness of the system. The less the mismatch to be corrected, the higher the degree of return loss realization across the band.

Let us now consider a nonoptimal transformer n replacing the optimum value N deduced in earlier expressions. If $2v'$ is the open circuit error voltage applied to the high side of the injection transformer then

$$\delta v_p' = 2nV_p\delta_p$$

and the current referred to the high turns side is

$$I_p' = \frac{V_p}{1 + \frac{1}{n^2}} \cdot \frac{1}{n} .$$

The terminating transformer, together with the correlated current flow in the low turns side, provides an output impedance Z for the subsidiary amplifier which must conform to the relationship

$$1 + Z = \frac{2v_p'}{I_p'} = 2n^2\left(1 + \frac{1}{n^2}\right)\delta_p$$

and

$$Z = 2(n^2 + 1)\delta_p - 1. \tag{55}$$

Equation (55) shows that for $(n^2+1) > 1/\delta_p$ the subsidiary amplifier, as expected, operates into a high impedance mismatch.

The subsidiary amplifier must be made to cope with the standing wave, to be capable of supplying either an excess open circuit voltage or an excess short circuit current. We assume a power capacity to be required from the subsidiary amplifier such that it can tolerate that peak through its internal impedance. Since this peak, relative to matched loading, is given by the ratio of $2Z/(Z + 1)$, namely $1 + |k|$, where k is the reflection factor, we have

$$\frac{v_p'}{v_p} = \frac{2(n^2 + 1)\ \delta_p - 1}{(n^2 + 1)\ \delta_p} \cdot \frac{n}{N} \tag{56}$$

with the corresponding peak subsidiary amplifier power ratios

$$\frac{P_s'}{P_s} = \left(\frac{2(n^2 + 1)\ \delta_p - 1}{(n^2 + 1)\ \delta_p}\right)^2 \frac{n^2}{N^2}. \tag{57}$$

Combining (57) with (46) we have, finally,

$$\frac{P_s'}{P_s} = \frac{(2(n^2 + 1)\ \delta_p - 1)^2 n^2}{(n^2 + 1)^2\ \delta_p(1 - \delta_p)}. \tag{58}$$

The formulation of (58) shows, of course, that when $n^2 = N^2 = 1 - \delta_p/\delta_p$, this ratio reduces to unity. As a quantitative determination, let us increase n such that it introduces roughly one-half the mismatch it inserts under optimal conditions for $\delta_p = \frac{1}{8}$, our 1 dB design center. Since $N^2 = 7$, let us choose $n^2 = 16$, where the latter number is chosen to form an integral turns ratio. Equation (58) demands 7.3 times as much power as that for the matched case, a level of power for the subsidiary amplifier virtually equal to that of the main amplifier.

The quantitative example chosen, of course, represents a fairly undesirable situation; nevertheless, it shows a tradeoff possibility which is not to be ignored. In multiple looping feed-forward systems, each loop is called on to handle progressively smaller and smaller deviations. In using these progressively smaller values of δ_p, we form a sequence of progressively higher values of N, corresponding in turn, according to (48) to lower values of P_s. Below some value, the value of P_s is no longer limiting and we may wish to use correction power as a tradeoff against mismatch.

APPENDIX C

Schmidt-Brücken Patent

During the final stages of preparing this article we found a German Federal Republic patent, issued July 14, 1960 (document 1085194), to Heinrich Schmidt-Brücken, entitled "A Method For Compensating Distortion In Amplifiers."

Clearly this patent, in spirit, precedes our own activities in attempting to provide time compatible error cancellation. While it suggests a time flow correction system seemingly much similar to our own, its actual embodiment severely compromises the intent of the patent. We have reproduced Figures 1 and 3 of the Schmidt-Brücken patent as our own Figures 58 and 59. We claim the disclosure shown in Fig. 58 to be operative only in principle, and that of Fig. 59 to be generally inoperative.

We feel it relevant to pursue the dissection of these devices since they illustrate, negatively, important considerations in the format of implementation of feed-forward amplifiers. It is not, however, intended by so doing to deprive Mr. Schmidt-Brücken of a well earned recognition of his priority in recognizing time as a controllable factor in error cancellation.

A major consideration lacking in both the configurations of Figures 58 and 59 is that of directional wave sampling. Let us observe the consequences of a simple voltage sampling as used in Fig. 58 where such sampling follows the usual practice in feedback amplifiers.

If all components in Fig. 59 are matched, that system will operate. In general, however, there is no disclosure of the degree of match required in the patent, and the effect of multiple reflections are

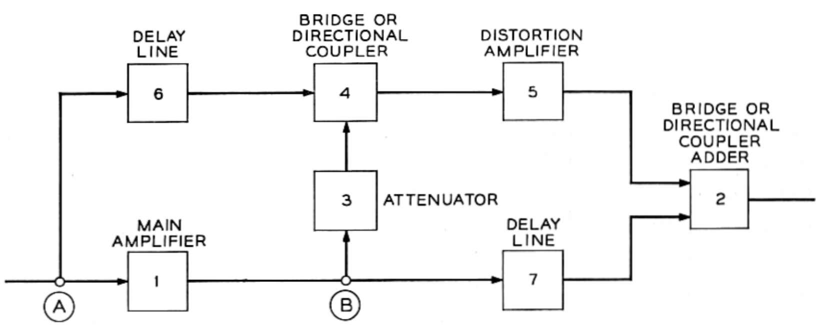

Fig. 58 — Schmidt-Brücken Fig. 1.

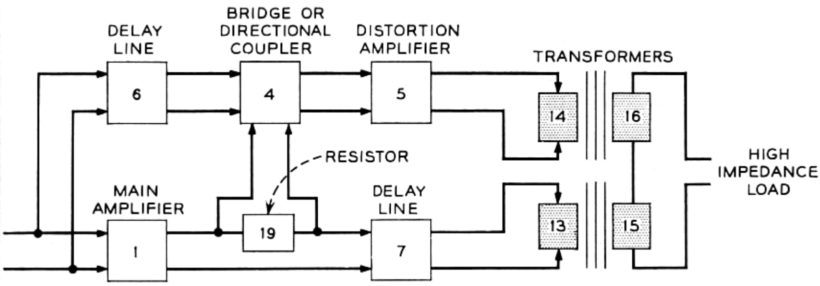

Fig. 59 — Schmidt-Brücken Fig. 3.

ignored. Let us assume, however, that all devices are matched only to within reasonable specification of return loss, and that the amplifiers, particularly, might have return losses of the order of 20 dB.

The error reference is established at point A in Fig. 58 through the use of a voltage divider network, and, consequently, the error voltage is a function of the input impedance seen looking into the delay-line labeled 6. The phase of the 10 per cent reflection factor of the distortion amplifier 5 may rotate through more than one cycle as a function of the delay line length, producing a peak to peak impedance variation of 50 per cent. The reference sampling is modulated by this order of magnitude, making a shambles of the compensation system.

Inherent in this system, therefore, is an inclusion of an active device, namely the amplifier, as part of the reference standard, greatly limiting the general practicability of this class of device. In our own embodiments, we have carefully employed directional coupler sampling, stressing the incident wave, and not the voltage, as the reference. With properly designed match, third time around incidence is not greatly significant since it is down by twice the return loss, the order of 40 dB or greater, making amplifier mismatch a secondary consideration.

The same form of argument applies to point B with respect to the adder 2. Notice that both boxes 4 and 2 are bridges or directional couplers, and that they are used to prevent spurious coupling between chains. Awareness is manifested, therefore, within the patent itself of the existence of directional coupling devices, but nowhere are these couplers applied to the crucial problem of error sampling.

Figure 59 shows a device containing not only the above fault, but adds fallacy as well. The intention is to show alternative means for providing chain isolations by the use of constant current sources, such as pentodes, operating in parallel or, as actually shown, by the use of series connected secondaries of transformers 13 and 14 working into a relatively infinite output impedance.

The implications to multiple reflections are obvious and the consequence to the current sensing resistor 19 is disastrous. If, for example, delay line 7 passes through a half wavelength, 19 is open circuited and develops no sense voltage. The paralleling of pentode chains is equally disastrous since one of the pentodes, if viewed looking back into the delay line, may prove to be anything but a current source.

There are no considerations by Schmidt-Brücken to use his device to transfer power. All considerations of the relevant requirement of the output combiner are missing, as are considerations of output impedance. Pointedly, we differ with Schmidt-Brücken in calling for an output bridge or directional coupler, since these must be power wasting four-ports. We have expressly demanded amplifier interactions in an output three-port to conserve power. In like fashion we disagree with Schmidt-Brücken in demanding an open circuit output to avoid interaction when he does, indeed, chance to use an output three-port.

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